15

CMOS RF Transceiver Considerations for DSA

Mark S. Oude Alink, Eric A.M. Klumperink, André B.J. Kokkeler, Gerard J.M. Smit and Bram Nauta

Centre for Telematics and Information Technology, University of Twente, Enschede, Netherlands

15.1 Introduction

Cognitive Radio (CR), and in particular dynamic spectrum access (DSA), promises a much more efficient use of the spectrum by opportunistically using available frequencies. This asks for specific functionality, like spectrum sensing and frequency-agile transmission and reception. We will show that this functionality poses challenging hardware requirements, which go far beyond what is currently possible with an analogue-to-digital converter (ADC) and digital-to-analogue converter (DAC). Instead, a transceiver (transmitter + receiver) with filtering and frequency conversion is required. By starting from a mathematical abstraction for the description of transceivers and an overview on transceiver implementation, we will show that the flexibility required by CR calls for changes in the architecture, putting severe constraints on linearity and spurious emission performance. We will discuss several existing and proposed solutions to alleviate the design of CR transceivers and spectrum sensing functionality, with a special emphasis on CMOS as it is low-cost and enables the integration of both analogue and digital on one integrated circuit (IC).

Communication systems transmit and receive data, and therefore require a transceiver. In this chapter we focus on the physical layer: how can a transceiver for CR be implemented in hardware? In addition, we look at the spectrum sensing functionality that is

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required for DSA. We focus mainly on the challenges for the analogue part of the hardware, except for the antenna, which we assume here as an ideal wideband source with 50 Ω impedance, as is commonly done during transceiver design and characterization.

15.1.1 Terminology

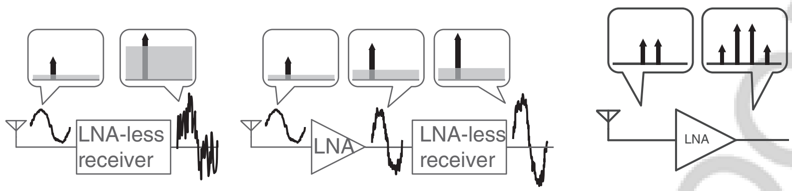
This section introduces terminology for the reader who is not very familiar with transceiver and/or system design.

The digital data transmitted over the air is *modulated*, that is (groups of) bits are transformed into an analogue waveform, with its properties (amplitude, phase, frequency, etc.) denoting the transmitted bits. This waveform is transmitted at a certain *power*, often denoted in dBm, which is a dB-scale with 1 mW as a reference. Typical transmitted power levels of mobile devices are in the range of 0–20 dBm, or 1 to 100 mW. Only a small fraction of the transmitted power is received by the receiver, because the transmitter generally transmits in many or all directions. The transmitted waveform is susceptible to *noise*, which is added in the wireless link (the *channel*) as well as in the receiver. This noise causes errors in the detection process, because it changes the waveform. The transmitted waveform can also be changed by *fading*, where the channel introduces frequency- and time-dependent changes to the waveform, but we will not discuss that in this chapter.

The *noise figure* (NF) quantifies the noise performance of a receiver as a deterioration of *signal-to-noise ratio* (SNR) from input to output. The SNR is the ratio of the useful signal power divided by the noise power in the signal band. The higher the SNR, the less bit errors are introduced by noise. The fraction of bits that are incorrectly received is denoted by the *bit error rate* (BER). Depending on the type of modulation, the typically required SNR to obtain the desired BER ranges from 8 dB to 25 dB [1]. Since the demodulation is done at the output of the ADC, the SNR should be high enough for the desired BER. This means that the NF of the receiver should be low enough. Receivers typically have a NF of 2 to 10 dB.

The received signal is often so weak that it would be completely obscured by the receiver noise if it is not first amplified. To obtain a low NF, almost all receivers start with a low-noise amplifier (LNA), which amplifies the input signal without adding much noise, such that the noise added by successive stages becomes relatively less important. The basic principle is shown in Figure 15.1(a), where the received signal is a sine wave plus a little bit of noise.

Ideally, such an amplifier can be described mathematically as $y(t) = ax(t) + n(t)$, where $x(t)$ is in the input signal (which can be a voltage, current, or charge), a is the amplification, $n(t)$ is the noise added, and $y(t)$ is the output signal. Unfortunately (neglecting $n(t)$), amplifiers are better described by $y \approx a_1x(t) + a_2x^2(t) + a_3x^3(t)$, that is, as a nonlinear device approximated using a Taylor-series. The values $|a_1/a_2|$ and $|a_1/a_3|$ are a measure for the *linearity* of the system. An elaborate discussion can be found in [2], but we will be brief here.



(a) **The LNA mitigates the effect of noise added by the following stages of the receiver**

(b) Non-linearity in the LNA distorts the spectrum, and hence increases BER

Figure 15.1 The LNA is a crucial component of receivers, as it should provide gain and have a low NF to keep receiver NF low enough, while at the same time it should be very linear. The spectra are drawn on a dB-scale, while the time-signals are drawn on a linear scale. (a) The LNA mitigates the effect of noise added by the following stages of the receiver. (b) Nonlinearity in the LNA distorts the spectrum, and hence increases BER.

Suppose two sine waves of equal power P at frequencies f_1 and f_2 are applied to the input of such a system. Using Fourier theory, it can be shown that the nonlinear behaviour will introduce frequency components at $2f_1 - f_2$ and $2f_2 - f_1$, which is schematically depicted in Figure 15.1(b). The input-referred third-order intermodulation intercept point (IIP3) is the (extrapolated) input power for which these undesired *intermodulation components* have equal magnitude as the desired components, and is usually expressed in dBm. Extrapolation is often necessary, because for such high input powers the Taylor-series approximation with only a few terms is usually not valid anymore. Typical receiver IIP3 is in the range of -30 dBm to 0 dBm. The input-referred second-order intermodulation intercept point (IIP2) is defined in a similar way, where the intermodulation components at $f_2 - f_1$ and $f_1 - f_2$ are considered. Values for IIP2 typically range from 20 to 70 dBm. For narrowband systems, the products due to second-order distortion are out-of-band, often making IIP3 the dominant distortion mechanism. The *dynamic range* (DR) is defined as the ratio, usually expressed in dB, of the maximum to the minimum signal input power levels over which a device can operate; the minimum level is usually determined by the noise, and the maximum level by nonlinearity.

For a transmitter, the nonlinear behaviour results in in-band distortion and *spectral regrowth*, that is a widening of the spectrum in which significant power is transmitted. The combination of in-band distortion and noise in the transmitter gives rise to an *error vector magnitude* (EVM), which must be kept small for low BER at the receiver.

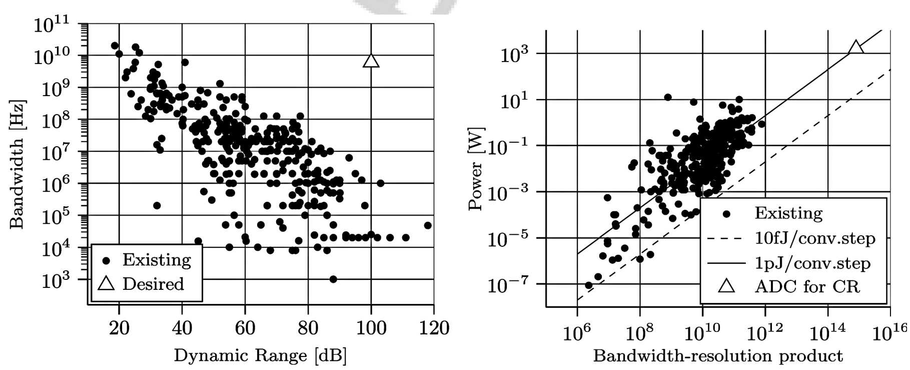
The range of frequencies over which a receiver is able to receive information is its *bandwidth*, sometimes called radio frequency (RF)-bandwidth. It is usually defined as the band over which the receiver is matched to the antenna (e.g. the power received by the antenna is (almost) completely transferred to the receiver) and performance of the receiver is not degraded too much (e.g. NF increased by 2 dB or gain decreased by 3 dB). The bandwidth that the ADC eventually has to convert to the digital domain is often referred to as intermediate frequency (IF)-bandwidth, and in general is much smaller than the RF-bandwidth, because the signal to be demodulated only occupies a fraction of the RF-bandwidth. Similar considerations hold for the transmitter.

15.1.2 Transceivers for DSA: More than an ADC and DAC

With the tremendous performance improvement of digital circuits over the years, one would like to do as much as possible in the digital domain. In principle, the analogue waveforms to be generated can be very well approximated in the digital domain when the sample rate is high enough and enough bits per sample are available. Assuming it can provide enough output power, a DAC near the antenna, possibly with just a low pass filter (LPF) in between, would then be the complete analogue transmitter. Similarly, an ADC placed directly behind the antenna can convert the received analogue signal to the digital domain, after which the transmitted bits can be retrieved digitally. This is only possible when the DAC and ADC can handle the required sample rate at the required resolution.

Unfortunately, this is not feasible at this moment, see Figure 15.2(a). In general, DACs are easier to make than ADCs (internally, an ADC often requires a DAC with the same resolution and speed), and therefore ADCs will form the ultimate bottleneck. Let us make the reasonable assumption that the received signals have a power between -100 and 0 dBm, so that DR = 100 dB, and that the CR has to be able to cover the RF-spectrum up to 6 GHz. The ADC would then require 17 bits at a sample rate of 12 GS/s, which is far from ADCs that exist already. If we extrapolate historical trends, for example the observed doubling of bandwidth-resolution product of ADCs every four years [3], such ADCs will become available only as early as the year 2055.

Even if such an ADC would exist, the power consumption is likely to be prohibitive. If we assume an energy of 1 pJ per conversion step, the power consumption would be $10^{12} \times 2^{17} \times 12{\cdot}10^{9} = 1.6\,\text{kW}$, see Figure 15.2(b). The battery of a mobile device would be



(a) **Currently, no ADC achieves a DR of 100dB and a BW of 6GHz.**

(b) **A 2 times higher bandwidth-resolution product requires roughly twice the power.**

Figure 15.2 State-of-the-art ADC-performance (a) Currently, no ADC achieves a DR of 100 dB and a BW of 6 GHz. (b) A 2 times higher bandwidth-resolution product requires roughly twice the power. (from Reference [4] which is regularly updated).

quickly drained and would require significant cooling. Most standards have a bandwidth below 20 MHz and do not need such a high DR, so it would be a waste of power to capture the whole RF-bandwidth. Hence, we need a way to alleviate the requirements on the ADC.

15.1.3 Flexible Software-Defined Transceiver

Radios that perform as many functions as they can in the digital domain, such that new standards and options can be accommodated by changing the software, are called software-defined radios (SDRs). A SDR can be seen as an all-in-one radio, for example to receive FM-radio, to send data to the WLAN-router and to make a phone call over GSM.

Although a CR does not necessarily have to be implemented on top of a SDR (in its basic form, it is just a standard radio operating with a flexible carrier frequency f_c), it certainly makes sense: a CR needs to perform many other functions, such as described elsewhere in this book, which heavily depend on digital signal processing (DSP) and software. Moreover, it would enable the CR to adapt its modulation type to the environment, which makes it more flexible and allows it to use the spectrum more efficiently. In addition, a CR needs to perform spectrum sensing, which also requires a flexible receiver (not necessarily the same as used for demodulation). Therefore, one could consider a CR as a flexible SDR [5,6].

15.1.4 Why CMOS Transceivers?

With the large amounts of DSP and software required for all the different functionality of a CR, it requires a power-efficient and fast technology for the digital circuitry. Since CR has a potential to be used in many different consumer devices, it should also be cheap to manufacture. These are precisely the strengths of CMOS.

CMOS can also be used for analogue circuits, such that the CR can potentially exist of one IC, which makes it very attractive for mobile applications due to the low cost, low power consumption, small form factor, and the simplified printed circuit board (PCB)-design. Unfortunately, the analogue performance does not scale nearly as well as the digital performance, partly due to the ever decreasing supply voltages. Although not all implementation issues we discuss in this chapter are CMOS-specific, the presented solutions and directions certainly are considering the specifics of CMOS analogue circuitry.

15.2 DSA Transceiver Requirements

In Section 15.1.1 we introduced the main performance metrics of receivers and transmitters. In this section, we will summarize the requirements of DSA transceivers in terms of these metrics, using requirements from the Federal Communications Commission (FCC), Office of Communications (Ofcom), and IEEE 802.22 as the references. The FCC is the regulatory commission on spectrum use in the USA, while Ofcom is its UK counterpart. IEEE 802.22 is currently in the draft phase, with the goal to become a CR

Table 15.1 CR requirements set by different authorities assuming mobile devices that rely on spectrum sensing

	FCC [7]	Ofcom [8]	802.22 [9]	Unit
Maximum transmit power	50	50		mW
Maximum adjacent-to-TV channel	40	2.5		mW
Adjacent-channel emission	−72.8 dB[1]	−46 dBm	−49 dBm	
Modulation	free	free	OFDM	
DTV sensing limit	−114	−120	−116	dBm
DTV bandwidth	6	8	6, 7, 8	MHz
WM sensing limit	−107	−126	−107	dBm
WM bandwidth	0.2	0.2	0.2	MHz
Monitoring interval	≤60	1		s
P_{FA}			10	%
P_{MD}			10	%
Back off time	2	<1	2	s

standard in the TV-bands (from roughly 50 to 900 MHz). The exact numbers may change from draft to draft, but we will use the ones shown in Table 15.1. Of course, the concept of CR applies to other frequency bands as well, and especially the emission and sensing parameters may be quite different in other frequency bands. Nevertheless, these numbers provide us with a good starting point.

The maximum transmission power and maximum out-of-band emissions are a strict requirement. Adjacent-channel emissions limit the maximum power a transmitter is allowed to transmit in the TV-channels *adjacent to* the TV-channel it is actually transmitting in. The FCC specifies these limits as the power in a 100 kHz bandwidth, relative to the power transmitted in the 6 MHz channel. When the CR transmits at full power (17 dBm), the adjacent-channel emission limit is −55.8 dBm in 100 kHz, so maximally −38 dBm in the total adjacent channel. Note that when the transmitter is not transmitting at its maximum power level, Ofcom and 802.22 regulations become easier to satisfy because the emission limit is absolute, while the FCCs relative requirement remains equally strict.

Since 802.22 is a communication standard, it defines the type of modulation, while the FCC and Ofcom only care about the protection of other users. With no modulation requirements set by the FCC and Ofcom, and no regulations on how to divide the white space, it is very well possible to use multiple TV-channels for data transmission, or only fractions of TV-channels. These channels do not even have to be adjacent in frequency, although there are no requirements to accommodate these possibilities.

The most important primary user in the TV-bands is digital TV (DTV), which is implemented in different ways around the world. The DTV-bandwidth ranges from 6 to 8 MHz, as discussed in Chapter 12, with the modulation in the USA being 8VSB and in Europe OFDM. Another important primary user is the wireless microphone (WM), of

[1] Measured in 100 kHz out-of-band with reference total power in 6 MHz.

which 0, 1 or more can be present in a single otherwise unoccupied TV-channel. Each WM occupies a bandwidth up to roughly 200 kHz, but neither the bandwidth nor the modulation scheme is standardized. Many WMs use analogue FM, but analogue AM and several digital modulation schemes are also in existence. Other primary users in the TV-bands are analogue TV and some medical equipment, but for the sake of brevity we do not consider them here.

Although spectrum databases are now the primary source of information to find white space, spectrum-sensing-only devices are still allowed, and may become more important in the future if it becomes more mature and reliable. A false alarm is the situation where a CR wrongfully concludes that a channel is occupied by a primary user, whereas a missed detection occurs when a CR wrongfully concludes that a channel is free for secondary use. A missed detection causes harmful interference, while a false alarm wastes an opportunity. P_{FA} and P_{MD} denote the respective probabilities that these events happen, and they are upper bounded in the 802.22 standard, which places constraints on the sensing performance.

The back off time is the time a CR may take to move to another band when a primary user returns; Ofcom requires a CR to sense at a minimum interval of 1 s, hence we infer that the back off time should be shorter than that. The spectrum sensing device should be able to cope with these different bandwidths. With the Ofcom regulations being the strictest for a sensing interval of 1 s, spectrum sensing of a single channel should not take more than 100 ms in order not to reduce the overall data rate by more than 10%.

With a received noise power spectral density (PSD) equal to -174 dBm/Hz, and assuming the receiver has a NF of 5 dB, but is otherwise ideal, the noise power in 200 kHz will be $-174 + 5 + 10\log_{10}(200 \cdot 10^3) \approx -116$ dBm. So, the SNR for detecting a WM in 200 kHz BW is around 9 dB for FCC and 802.22, and around -10 dB for Ofcom. For DTV, the detection SNR is -13 dB (FCC for 6 MHz BW), -20 dB (Ofcom for 8 MHz BW) and -16 dB (802.22 for 6 MHz BW).

15.3 Mathematical Abstraction

To gain insight in the operation of transceivers without directly assuming a certain implementation, we describe the process of data communication at a more abstract level as shown in Figure 15.3. Starting from this abstraction, we will then look at implementation aspects in later sections. This also allows us to include or exclude certain

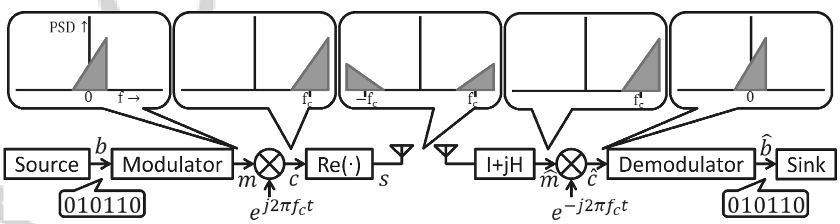


Figure 15.3 Our mathematical abstraction of a transmitter and receiver.

non-idealities and assess their impact on the communication quality. The use of complex math (i.e. complex numbers, not complicated math) allows us to use a more compact description of several aspects. A more elaborate discussion on the use of complex math in the context of transceivers and its advantages is given in [10].

Usually, the bits to be transmitted are coded and interleaved to achieve a better BER. Such a block takes in a sequence of bits, and produces another (possibly longer) sequence of bits. We define this output sequence of bits as b, and we neglect the coding at the transmitter and the decoding at the receiver. Therefore, the ultimate goal is that the received sequence of bits $\hat{b}$ resembles b as good as possible. $\hat{b}$ is modulated to a sequence of symbols $m(t)$, which are usually one or more sine waves of which the amplitude, phase and frequency are set according to b. It is customary practice to describe the symbols as complex samples or as a complex time-domain signal. We can draw its spectrum $|F(m(t))|^2$, with $F(\cdot)$ denoting the Fourier-transform operator. Note that, because $m(t)$ is a complex signal, its spectrum is only symmetric around 0 in a some special cases, so for generality we draw it as being asymmetric. It is also customary practice to centre the spectrum around DC and then refer to $m(t)$ as the *baseband signal*, with

$$m(t) \equiv m_R(t) + jm_I(t), \tag{15.1}$$

such that we can individually address its real and imaginary components. They can be completely independent, which means that each carries part of the information of $m(t)$.

The communication takes place around a specified centre frequency f_c, which can be defined by the standard, or chosen because it is available or has desirable characteristics, such as the ability to penetrate buildings. Therefore, we can generate

$$c(t) \equiv m(t)e^{j2\pi f_c t}, \tag{15.2}$$

which is a frequency-shifted version of $m(t)$ (and still complex). Because it has to be physically transmitted, it must be *real*. Taking the real part of a complex number z, $R(z)$, is equivalent to $R(z) = {}^1\!/_2)z + z^*)$, with z^* the complex conjugate of z. Taking the real part of $c(t)$ results in

$$s(t) \equiv \Re(c(t)) = m_R(t)\cos(2\pi f_c t) - m_I(t)\sin(2\pi f_c t). \tag{15.3}$$

As shown in Figure 15.3, $s(t)$ has a symmetric spectrum. Note that *no information is lost* by taking the real part of $c(t)$.

At the receiver side, the operations are reversed. The received signal $r(t)$ is (neglecting the channel) equal to $r(t) = d(t) + u(t)$, where $d(t) = s(t)$ is the desired signal, and $u(t)$ contains all other undesired signals (assumed 0 for now). How to get to $\hat{c}(t)$, which should be equal to $c(t)$, is somewhat more involved mathematically and is most easily described using the Hilbert transform. The Hilbert transform can be thought of as a linear time-invariant (LTI) filter with impulse response $h(t) = 1/\pi t$ and frequency response

$$\mathrm{H}(f) = \begin{cases} j, & f < 0 \\ -j, & f > 0 \\ 0, & f = 0. \end{cases} \tag{15.4}$$

In other words, the Hilbert transform gives a $-90°$ phase shift for positive frequencies, and $90°$ phase shift for negative frequencies. From this it immediately follows that $H^{-1}(x(t)) = -H(x(t))$. As a basic example

$$H(\sin(x)) = H\left(\frac{1}{2j}\left(e^{jx} - e^{-jx}\right)\right) = \frac{1}{2j}\left(-je^{jx} - je^{-jx}\right) = -\frac{1}{2}\left(e^{jx} + e^{-jx}\right) = -\cos(x). \tag{15.5}$$

A useful theorem is Bedrosian's theorem [11]

$$H(x_{\mathrm{LPF}}(t)x_{\mathrm{HPF}}(t)) = x_{\mathrm{LPF}}(t)H(x_{\mathrm{HPF}}(t)), \tag{15.6}$$

when $x_{\mathrm{LPF}}(t)$ is a low-pass signal and $x_{\mathrm{HPF}}(t)$ is a high-pass signal, with non-overlapping components (more general requirements can be given, but this definition suffices here).

With $r(t) \equiv s(t)$, and using Bedrosian's theorem, we can write

$$\breve{r}(t) \equiv H(r(t)) = m_R(t)\sin(2\pi f_c t) + m_I(t)\cos(2\pi f_c t) \tag{15.7}$$

With $\hat{c}(t) \equiv r(t) + j\breve{r}(t)$ we find[2]

$$\begin{aligned} \hat{c}(t) &= m_R(t)(\cos(2\pi f_c t) + j\sin(2\pi f_c t)) + m_I(t)(j\cos(2\pi f_c t) - j\sin(2\pi f_c t)) \\ &= m_R(t)e^{j2\pi f_c t} + jm_I(t)e^{j2\pi f_c t} = m(t)e^{j2\pi f_c t} = c(t) \end{aligned} \tag{15.8}$$

So, when we define a block with a transfer $I + jH$, where I denotes the identity operator, and H the Hilbert transform, and $r(t)$ is the input, the desired $\hat{c}(t) = r(t) + j\breve{r}(t)$ is the output.

The baseband signal $\hat{m}(t)$ can be obtained via

$$\hat{m}(t) = \hat{c}(t)e^{-j2\pi f_c t}, \tag{15.9}$$

which can then be demodulated to $\hat{b}$. So, without any impairments, $\hat{b} = b$, and data communication has taken place.

[2] This approach assumes that the signal bandwidth is much smaller than the centre frequency, such that the phase difference between the antennas remains roughly the same over the band, which is a good approximation for most modern wireless standards (with UWB the notable exception).

15.4 Filters

As we will see later, filters play a very important role in transceivers. A filter in the context of transceivers is a building block that ideally passes some frequencies (transfer $H = 1$), the *pass band*, and blocks other frequencies (transfer $H = 0$), the stop band. In practice, these filters have *insertion loss* (IL), which means that the signals it should pass are somewhat attenuated, and *finite rejection*, which means that the signals it should block are only attenuated by a finite amount. The rejection often gets better further away from the pass band. The pass band is often defined as the frequency range over which the attenuation is less than 3 dB as compared to the optimal point. The pass band of a band pass filter ranges from some f_1 to some f_2, with f_c as the centre frequency between these points. The bandwidth of the filter then is $BW = f_2 - f_1$. The Q (quality factor) of a filter is defined as the centre frequency divided by the bandwidth of the pass band, or $Q = f_c/BW = (f_1 + f_2)/2(f_2 - f_1)$, and is a measure for the *selectivity* of the filter. The order of the filter determines how fast the attenuation increases away from the pass band. Most terms are graphically depicted in Figure 15.4.

Filters can be *active* or *passive*. Passive filters do not require a power supply. The IL of such a filter directly contributes to the NF of the system. These filters are very linear, a general characteristic of passive components. Active filters are usually built using amplifiers with feedback, such that besides filtering, the signal can be amplified as well. The feedback improves linearity, but requires high loop gain, which is only available at low frequencies [2]. Moreover, active components tend to be more noisy and less linear than passive filters.

Important differences exist between *integrated* and *discrete* filters, which will be discussed next.

15.4.1 Integrated Filters

When integrated in a CMOS-IC, passive filters are made up of resistors, capacitors and inductors. To implement a filter for a pass band of 2.401 GHz to 2.483 GHz (802.11 g,

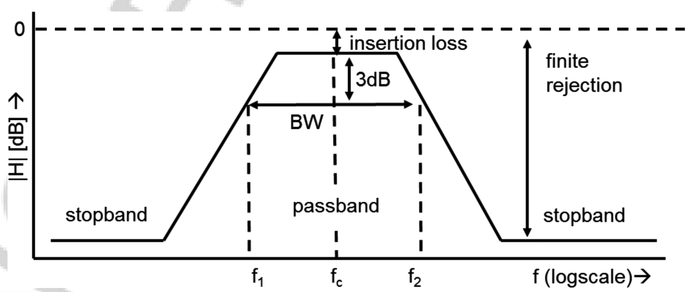


Figure 15.4 A BPF transfer characteristic and terminology.

WLAN), and to suppress at least 60 dB at 1.8 GHz (GSM-band), Q must be at least 30 and the order of the filter needs to be over 20. There are three important issues with these specifications:

- The maximum Q that can be obtained with integrated inductors is about 10 to 20 due their parasitic resistance and induced eddy currents (currents that are induced in other conductors, such as the substrate, by the changing magnetic field of the inductor);
- Filters with high order are more sensitive to variations in the value of the components (which is not well controlled in ICs) and their Q, which also varies over frequency [12];
- Filters with high order require a lot of components, and especially inductors take up a lot of expensive chip area.

Therefore, transceivers such as used in mobile phones always have discrete band select filters, although a lot of effort is undertaken to relax the requirements of the filters for transceivers to enable the use of integrated filters. We will discuss some of this in Sections 15.5 and 15.7.

Active filters are often used in the baseband section of receivers, where the signals have been amplified enough to make sure that the additional noise is not a problem, and the strongest interferers have already been significantly attenuated. The filters often are LPFs, where the bandwidth can be easily tuned by changing feedback resistance and/or capacitance. Cascading multiple stages enables the use of higher orders.

15.4.2 External Filters

External filters can be fabricated without the constraints of being process- or material-compatible with integrated circuitry. Several types exist, such as surface acoustic wave (SAW), bulk acoustic wave (BAW), resonant cavity, waveguide and yttrium iron garnet (YIG)-filters. The SAW-filters are most often found in mobile devices because many of the other filter types do not operate below several GHz, become too bulky because they use structures relying on wavelength, or are simply too expensive for the consumer market.

SAW-filters are electromechanical devices that use piezoelectric material to convert the incident RF-frequency to an acoustic wave that travels across the surface of the material. At the other end, the acoustic wave is converted back to the electrical domain via the reverse operation. The typical IL of a SAW-filter is 2 dB [13], with a package size of 1×1 mm to 7×7 mm (roughly equivalent to a typical IC-package) and a price tag of €0.10 to €1, which is not much lower than the price of a CMOS-receiver. An example transfer is shown in Figure 15.5.

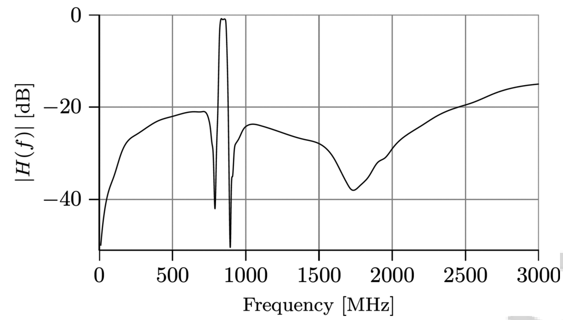


Figure 15.5 Example transfer of a SAW-filter for the 850 MHz GSM-band.

15.5 Receiver Considerations and Implementation

In this section, we will provide a basic overview of receiver considerations and implementations in order to provide the context for the more CR-specific challenges. We refer the interested reader to [14] for more information on general receiver implementation considerations.

The goal of the receiver is to transform the received waveform $r(t)$ into $\hat{b}$. One of the most important challenges for the receiver is that $s(t)$ is not the only signal present in a wireless medium. In fact, it is often the case that the other signals are much stronger than $s(t)$, for example because other transmitters transmit at higher power levels and/or are closer to the receiver. Because $s(t)$ in general is so weak (if it were not, a lot of power would be wasted in the communication), it needs to be amplified to overcome the noise limitations of some receiver components. Schematically this is depicted in Figure 15.6. As we already saw in Section 15.1.2, we cannot simply do this by directly sampling all RF-frequencies.

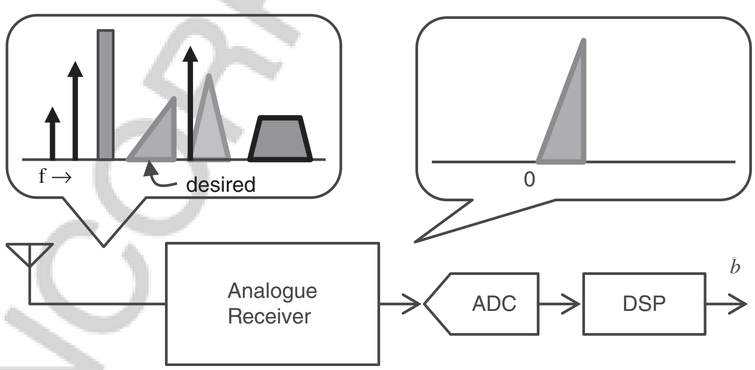


Figure 15.6 The goal of a receiver is to amplify the weak signal to be demodulated and to suppress other signals.

15.5.1 Sub-sampling Receiver

A possible solution is to use *sub-sampling*, which is schematically depicted in Figure 15.7. Variable gain is required to properly interface the (possibly) weak signal with the ADC. Let us assume that the ADC has infinite resolution, so that we can focus only on the sampling process, denoted as $q(t)$, with

$$q(t) = \sum_{k=-\infty}^{\infty} \delta\left(t - \frac{k}{f_s}\right), \tag{15.10}$$

with $\delta(t)$ the Dirac-function and f_s the sample rate. Then $r(t)\cdot q(t)$ equals (in the frequency domain)

$$R(f) \otimes Q(f) = \sum_{k=-\infty}^{\infty} R(f - kf_s) \tag{15.11}$$

where $\otimes$ denotes convolution.

Note that after sampling we only need to be concerned with the frequency range from 0 to $f_s/2$, because the spectrum is repetitive. When we assume that the desired signal $d(t)$ occupies the band from nf_s to $(n + {}^1/_2)f_s$, then (using $r(t) = d(t) + u(t)$, with $u(t)$ the undesired signals)

$$R(f) \otimes Q(f) = D(f - nf_s) + \sum_{k=-\infty}^{\infty} U(f - kf_s). \tag{15.12}$$

Clearly, the received signal is polluted with contributions from $u(t)$ from many different frequencies. Thus, a very steep filter with a bandwidth of $f_s/2$ is required, centred around $D(f)$. Such a filter is required for each band of interest. As we saw in Section 15.4, this is not possible with integrated filters, so it will be extremely bulky and costly to have such a dedicated filter for a wideband receiver. One can imagine that even when such filtering is possible, the other bands will still contain noise, for example, added by the LNA. In (15.12), $U(f)$ then represents only the noise, resulting in the translation of noise from all the bands on top of the desired signal, a process that is called *noise folding*.

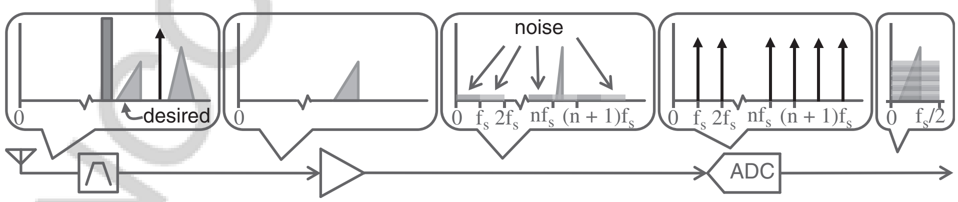


Figure 15.7 A sub-sampling receiver performs frequency conversion and sampling in one step, but requires a dedicated high-Q filter for each band. It suffers severely from noise folding.

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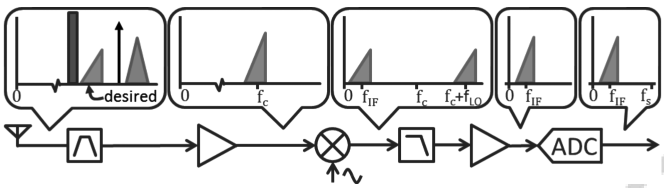


Figure 15.8 A heterodyne receiver performs a frequency conversion on the signal to be demodulated in order to facilitate further processing.

15.5.2 Heterodyne Receivers

To alleviate the aforementioned problems, receivers use *mixers* to perform frequency translation. The frequency translation is performed by multiplying with a sinusoid (convolution in the frequency domain)

$$F(r(t)\cos(2\pi f_{lo}t)) = \frac{1}{2}R(f - f_{lo}) + \frac{1}{2}R(f + f_{lo}) \tag{15.13}$$

and then filtering away the undesired component, such that the resulting spectrum can be converted to the digital domain without a lot of noise folding. Such receivers are called *heterodyne* receivers. So, when a signal at f_c is multiplied with $cos(2\pi f_{lo}t)$, it appears at $f_{if} \equiv f_c - f_{lo}$ and at $f_c + f_{lo}$. When f_{if} is a low frequency, a LPF can be used to remove the component at $f_c + f_{lo}$. This process is schematically depicted in Figure 15.8.

Once in the digital domain, filters can be made that perform as well as required (although keeping processing power consumption to a minimum is a challenge [15]), and the operations depicted in Figure 15.3 can directly be applied, as shown in Figure 15.9.

Two important issues with this architecture are the *image problem* and the *channel selection problem*, both of which are illustrated in Figure 15.10.

Let us assume that $r(t) = d(t) + u(t)$, with $d(t) = m(t)cos(2\pi(f_{lo} + f_{if})t)$ the desired component and $u(t) = v_{img}(t)cos(2\pi(f_{lo} - f_{if})t) + v_{int}(t)cos(2\pi(f_{lo} + f_{if} + \Delta f)t)$ the undesired image plus an undesired interferer that is close in frequency to the desired signal. Then

$$\begin{aligned} r(t)\cos(2\pi f_{lo}t) = {} & \big(m(t) + v_{img}(t)\big)\cos(2\pi f_{if}t) + v_{int}(t)\cos\big(2\pi\big(f_{if} + \Delta f\big)t\big) \\ & + \text{other components to be filtered.} \end{aligned} \tag{15.14}$$

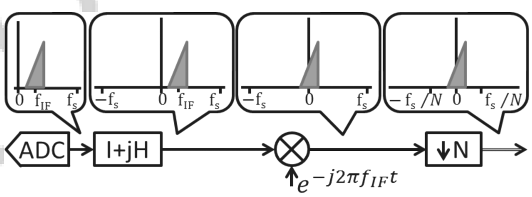


Figure 15.9 A block schematic showing the possible DSP-steps in a heterodyne receiver to obtain $\hat{m}(t)$.

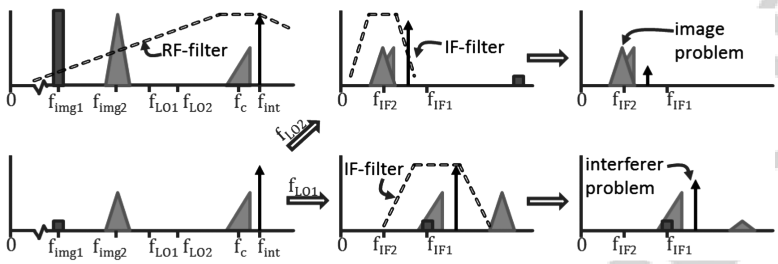


Figure 15.10 The position of f_{lo} with respect to f_c determines how well the image and interference close to the desired signal can be suppressed.

Clearly, $m(t)$ may be obscured by $v_{img}(t)$ and/or $v_{int}(t)$ if they are not properly suppressed.

The image component $v_{img}(t)$ is originally at a distance of $2f_{if}$ from f_c, so the higher f_{if}, the more the image can be suppressed by a filter before the frequency conversion, as is illustrated in Figure 15.10. The interfering signal $v_{int}(t)$ needs to be filtered out *after* frequency conversion, because at RF the filters do not have enough selectivity. This is referred to as channel selection, because the situation often occurs in standards where this strong interferer is another transmitter of the same standard using another frequency channel (in GSM these channels are only 200 kHz apart). In order to be able to provide sufficient gain without running into linearity and voltage headroom problems, and to lower the requirements on the ADC, the interferer should be suppressed sufficiently by filtering. A lower f_{if} makes the filter requirements easier.

Clearly, the image problem and channel selection problem have contradictory requirements for f_{if}. The *superheterodyne* receiver solves this dilemma by using multiple frequency conversions, while the *direct-conversion* receivers resort to the Hilbert transform in order to reject the image. These solutions are discussed next.

15.5.2.1 Superheterodyne Receiver

By using multiple frequency translations, the superheterodyne receiver can first suppress the image frequency using a sufficiently high f_{if}, while a second frequency translation to a low f_{if} makes sufficient channel selection possible. Given the fact that a signal can be as much as 100 dB stronger than the desired signal, a very strong suppression of the image is required.

A big problem can be *feedthrough*, which is the fact that some part of the signal power shows up at the output of the mixer without frequency translation. This requires either a very good control of feedthrough, or f_{if} should be above the highest or below the lowest useful input frequency.

Using multiple frequency translation steps requires very good *frequency planning*, because every frequency translation must have its own image and feedthrough suppression, as well as suppression of components that show up as images due to the previous frequency translations.

A big downside of this architecture is the use of multiple local oscillators (LOs), which makes the design more complicated and power-hungry.

15.5.3 Direct-Conversion Receivers

Instead of trying to suppress the image by filtering, it can also be rejected with the aid of the Hilbert-transform. Two main types exist: the *zero-IF* receiver, where $f_{if} = 0$, and the *low-IF* receiver, where $f_{if} > 0$ and large enough to have the whole channel to be demodulated above 0 Hz.

The block schematic of a zero-IF receiver is shown in Figure 15.11. The mixer is different from that of the heterodyne receiver. If it would be implemented in the same way as shown in Figure 15.8, the information at xHz above f_c falls on top of the information at xHz below f_c. This is not a problem for double sideband (DSB)-modulated information (equivalent to $m(t)$ being Hermitian, which means $M(f) = M^*(-f)$), because both frequencies contain exactly the same information. However, DSB-modulation is a waste of spectrum, and hence most digital communication takes place using single sideband (SSB)-modulation. In essence, the mixer should perform a multiplication with $exp(-j2\pi f_c t)$

$$
\begin{aligned}
r(t)e^{-\mathrm{j}2\pi \mathrm{f_c t}} &= \cdot\frac{1}{2}e^{-\mathrm{j}2\pi \mathrm{f_c}t}(c(t)+c^*(t)) = \frac{1}{2}e^{-\mathrm{j}2\pi \mathrm{f_c t}}\left(m(t)e^{\mathrm{j}2\pi \mathrm{f_c}t} + m^*(t)e^{-\mathrm{j}2\pi \mathrm{f_c}t}\right) \\
&= \frac{1}{2}m(t) + \frac{1}{2}m(t)e^{-\mathrm{j}2\pi 2\mathrm{f_c}t}
\end{aligned}
\tag{15.15}
$$

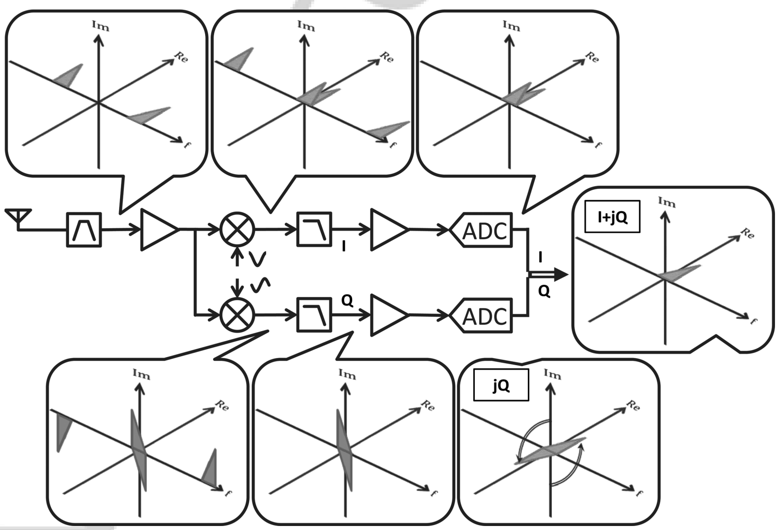


Figure 15.11 The zero-IF receiver rejects the image by using a complex frequency translation. For zero-IF, $f_{if} = 0$ and the image is the signal itself.

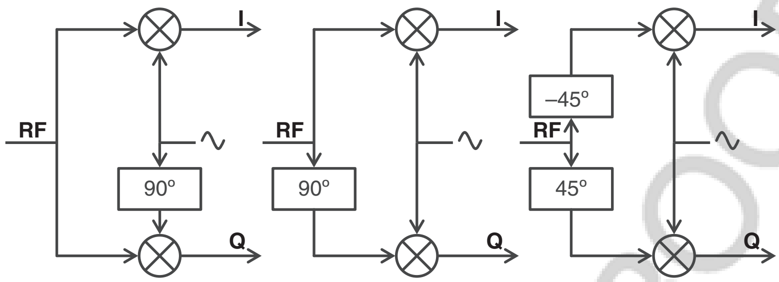


Figure 15.12 Some possible implementations for creating *I* and *Q* baseband signals.

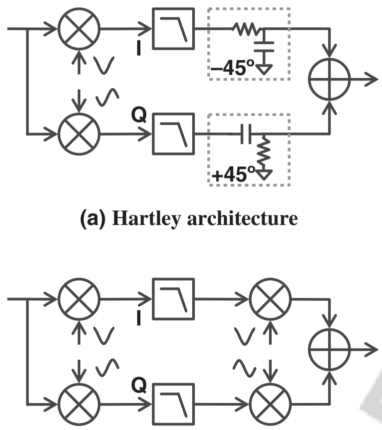
after which a LPF removes the undesired component at $2f_c$. Therefore, the analogue mixer should output a complex signal. Because the implementation must be physically realizable, the mixer has two real outputs, where output *I* represents the in-phase part (the real part), and output *Q* represents the quadrature part (the imaginary part) of the complex signal $I + jQ$. By noting that $exp(-j2\pi f_c t) = cos(2\pi f_c t)\text{-}jsin(2\pi f_c t)$, we see that we can simply duplicate the mixer of Figure 15.8, but using $cos(2\pi f_c t)$ for the LO of the *I*-mixer and -$sin(2\pi f_c t)$ for the LO of the *Q*-mixer, as is illustrated in Figure 15.11.

There are more ways to generate the *I* and *Q* baseband signals; a few variants are shown in Figure 15.12. Note that the 90° phase shift is a block with one real input signal and one real output signal. Given the Hermitian properties of real signals, the block gives 90° phase shift for positive frequencies, and necessarily −90° phase shift for negative frequencies; therefore, the 90° block can be interpreted as a Hilbert filter.

After filtering and amplification, the *I* and *Q* signals need to be combined. This can be done in the digital domain, using two separate ADCs, as shown in Figure 15.11. As the signal is centred around 0, occupying the band from $-BW/2$ to $BW/2$, each ADC now only has to sample at a rate *equal* to the bandwidth of the signal.

In a low-IF-implementation, the image frequency belongs to another channel, but the image rejection can be obtained in the same way as shown in Figure 15.11. For low-IF, in principle only 1 ADC is needed, sampling at a rate of twice the bandwidth. However, it then requires the I and Q signals to be combined in the analogue domain. The two most popular variants to realise this are the Hartley architecture and the Weaver architecture, as shown in Figure 15.13.

The Hartley architecture directly implements the 90° phase shift by two opposite 45° phase shifts in the *I* and *Q* signals. The simplest implementation of this phase shift is with two *RC*-filters. Perfect cancellation (infinite image rejection) only occurs when the phase difference between both paths is exactly 90° *and* the gains in both paths are exactly equal. The Hartley-architecture is fundamentally limited here because the two *RC*-filters will only have the same amplitude transfer at exactly one frequency. It is easier to make the phase shift in the LO-path, because it is ideally just a single frequency. The Weaver-architecture makes use of this by using a second frequency conversion, at the cost of extra images that have to be taken care of in some way.



(a) Hartley architecture

(b) **Weaver architecture**

Figure 15.13 Two main architectures exist to combine the *I* and *Q* signals to a single real analogue output signal where the image is rejected. (a) Hartley architecture (b) Weaver architecture.

In general, both the phase and the gain will not be exactly as desired due to *mismatch* between the components. The Hartley and Weaver architecture suffer from this problem, as well as the separate analogue paths in the zero-IF receiver. It is instructive to calculate the effect of phase and gain mismatch, and this is most easily done for a receiver with a single LO.

Since we multiply the LO with the input signal, it does not matter where this mismatch occurs, so for convenience we define the mismatch to be at the LO. The ideal LO (written as the combination of the I and Q-paths) looks like

$$LO_{\text{ideal}}(t) = \cos(\omega_{lo} t) + j\sin(\omega_{lo} t) = e^{j\omega_{lo} t} \tag{15.16}$$

and the LO in practice like

$$LO_{\text{practice}}(t) = \cos(\omega_{lo} t) + j(1+\varepsilon)\sin(\omega_{lo} t + \varphi) = LO_{\text{ideal}}(t) + \frac{1-(1+\varepsilon)e^{-j\varphi}}{1+(1+\varepsilon)e^{j\varphi}}\, e^{-j\omega_{lo} t} \tag{15.17}$$

This means that instead of just the desired signal, we also get a fraction of the image. Figure 15.14 shows the image rejection ratio (IRR) as a function of gain and phase mismatch (both together are known as *IQ-mismatch*), where IRR is defined as

$$\text{IRR} \equiv 20\log_{10}\left|\frac{1+(1+\varepsilon)e^{j\varphi}}{1-(1+\varepsilon)e^{-j\varphi}}\right| \tag{15.18}$$

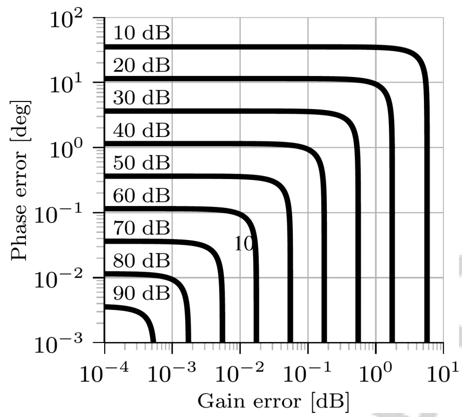


Figure 15.14 Image frequency suppression as a function of *IQ*-mismatch. The phase error is φ and the gain error is $10\log_{10}(1+\varepsilon)$.

The IRR in ICs is typically limited to 40 dB due to the analogue mismatches [14,16], but several digital compensation algorithms exist that can improve IRR by tens of dBs [16]. This can be enough, because typically the total suppression required is 60 dB to 70 dB [14].

Direct-conversion architectures, however, have their own issues [14]:

- Part of the LO-signal leaks to the RF-port of the mixer and then mixes with itself, introducing a DC-component at the output of the mixer. The same happens for leakage from the RF-signal to the LO-port. The DC-offset makes frequencies at or very close to 0 Hz unusable for demodulation, and/or can saturate subsequent amplifiers, unless DC-offset cancellation techniques are used;
- Assuming a narrowband receiver, the high frequency second-order distortion products fall out-of-band, and are filtered away. However, the low frequency second-order distortion products can leak through the mixer (feedthrough), appearing as distortion in the signal band;
- Flicker noise of especially transistors, caused by a variety of physical effects, can significantly degrade the SNR at low frequencies (kHz to MHz-range), and becomes more significant for smaller devices, but can be mitigated by chopping [17];
- The LO is at or close to the frequency of the signal, and may leak to the antenna, such that the receiver is radiating power and may violate emission regulations.

In ICs, nevertheless, direct-conversion architectures are very popular. The reason is mainly the compactness of the design due to increased simplicity and more integration (fewer external components), thus reducing cost and power consumption [18]. Many of the drawbacks have been solved to such an extent that it is feasible to integrate and comply with the standards.

15.6 Cognitive Radio Receivers

Receivers for CR face more stringent requirements than conventional receivers, since they have to be able to receive signals over a wide frequency band in a highly dynamic and a priori unknown environment, whereas many standards have definitions and limitations on what interferer power levels to expect at certain frequencies and frequency offsets. This gives the following requirements and constraints for a CR receiver:

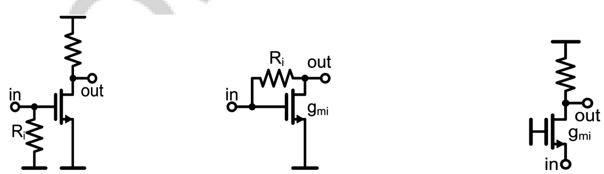
1. The RF-section of the receiver should be wideband;
2. An external RF filter bank cannot be used, as it becomes too expensive and bulky;
3. The internal frequency generation needs to be tuneable over several decades of frequency.

We will discuss these points and their consequences next.

15.6.1 Wideband RF-Section

Input matching in receivers is desirable, as it allows components to be cascaded with a known effect on system parameters. Moreover, many components, such as SAW-filters, only function properly when terminated by a matched load. Many narrowband receivers use a tuning circuit to obtain impedance matching without compromising noise performance.

By definition, this tuning is narrowband, which is unacceptable for wideband receivers. Therefore, matching should be obtained over a wide band, or at least around a tuneable frequency, such that it is matched in the desired frequency band. The simplest solution is by matching with a resistor, as shown in Figure 15.15(a), but this already gives 3 dB NF without any gain. To obtain sufficient sensitivity, a low NF is desired. Feedback, as shown Figure 15.15(b), can obtain a low NF with matching as long as the amplifier provides enough gain, and the feedback resistance can be tuned to follow the amplifiers gain profile for different settings, and the system can be kept stable. Especially high gain and stability become a challenge in the GHz-range due to the parasitic capacitances. Another well-known method is by using the small-signal input impedance of a common-gate amplifier, as shown in Figure 15.15c, where $R_{in} = 1/g_m$. Again the NF will be at least 3 dB, because R_{in} is fixed by the matching requirement.



(a) **Using a resistor** (b) **Using feedback** (c) **Using a common-gate amplifier**

Figure 15.15 Wideband matching can be obtained with different methods. (a) Using a resistor. (b) Using feedback. (c) Using a common-gate amplifier.

A leap forward is obtained by the *noise-cancelling LNA*, which was first described in [19]. For example, it uses Figure 15.15(b) or (c) for matching, and at the same time another amplifier (with high input impedance) in parallel that cancels the noise of the matching amplifier. In this way, matching and NF can be decoupled. The principle is shown in Figure 15.16, using Figure 15.15(b) as the matching technique.

The noise of the second amplifier is not cancelled, but it can be lowered at the cost of larger transistors and more power consumption, since the second amplifier does not have an input-matching constraint. Eventually, the NF is limited by the required bandwidth of the system, as the parasitic capacitance of the larger transistors limit input-matching, but wideband LNAs operating on this principle have been demonstrated with a NF around or below 2 dB.

Several improvements and variants of the principle have been demonstrated since then. The broadband CMOS LNA presented in [20] achieves a NF of 1.4 dB to 1.7 dB over a frequency range of 100 MHz to 2.3 GHz with 20 dB gain, at a power consumption of 18 mW and with an IIP3 of -1 dBm. An interesting LNA designed with CR in mind is presented in [1], where a noise-cancelling LNA in combination with feedback is used to (partially) cancel the capacitive part of the input impedance of the LNA, hence extending its bandwidth to roughly 10 GHz, providing 19 dB gain with a NF between 2.9 and 5.9 dB at a power consumption of 22 mW (IIP3 not available). For an overview of the performance of other broadband LNAs, we refer the interested reader to [20].

15.6.2 No External RF-Filterbank

As discussed in Section 15.5, traditional receivers have an external high-quality RF filter that already filters out or strongly attenuates many of the interferers. With this concept, in-band signals are under control (power limits are specified by the standard), while the out-of-band interference is strongly reduced, which lowers nonlinearity problems in the receiver.

For a CR, the distinction between 'in-band' and 'out-of-band' becomes somewhat vague, and the allowed power levels are not restricted per se. When many interferers are present, the nonlinearity of the receiver causes the distortion components to act as an increased noise level (hence the term 'interference temperature' is introduced by

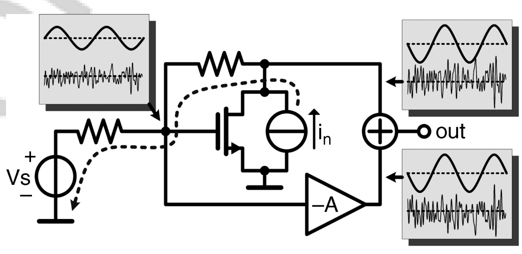


Figure 15.16 The noise-cancelling LNA of [19]. The signal is amplified, and the noise from the transistor (modelled as a current source) is cancelled by proper choice of the parallel amplifier gain A.

Haykin [5]). Marshall [21] argues that these distortion products can cause an overload in the receiver, or at least severely increase the noise level, and therefore a CR should use a large bank of RF-filters. This may be feasible for military applications, but we believe that it is too bulky and expensive for the consumer market.

It may be feasible to use a relatively small set of RF-filters, but in that case only a limited number of interferers can be suppressed. We can think of several other, not necessarily orthogonal, ideas to (partially) offset the lack of RF-filtering.

1. Exploit the DSA-capability of a CR;
2. Increase the linearity of the receiver;
3. Improve harmonic rejection (HR);
4. Use spatial filtering;
5. Make a tuneable integrated RF-filter.

We will discuss these options next.

15.6.2.1 Exploit the DSA-Capability

The distortion components generated by the nonlinearity in the receiver have a deterministic relation with the input frequencies. A CR with DSA has some form of spectrum analyser (SA) on board, and this can be put to good use. In [21] the receiver itself is the SA in the sense that it simply measures the total power that is received in the pass band of an external filter. The system then selects the filter from the available filter bank that receives the least amount of power.

When such a bank of filters is not available, the DSA capabilities can still be exploited. Under the assumption that the SA is ideal (we will see in Section 15.9.3 that it can be made more linear than a normal receiver), the input spectrum can be determined, and the location and power of the distortion components can be estimated by using the linearity specifications of the receiver. Based on this calculation, it is easy to find the white spaces with the lowest linearity requirements [22].

The linearity requirements of a receiver are derived in [22], under the reasonable assumptions that the signal to be demodulated has a power of −60 dBm, several interferers are present, each with a power of −10 dBm, the required signal-to-distortion ratio is 10 dB for proper demodulation, several external octave BPFs are present to make sure that IIP3 (and not IIP2) is the limiting factor, and all channels are 6 MHz wide (e.g. TV-channels). A scenario is shown in Figure 15.17 with an external filter with a pass band that is 120 MHz (20 channels) wide.

In this scenario, it can be shown that the IIP3 requirement for channel 10 is 33 dBm and for channel 11 36 dBm (both channels suffer from third-order distortion products from the signals present in channels 1, 6 and 19). As we will see shortly, this is far from what is currently possible. In the channels next to such a strong interferer, such as channel 2, IIP3 requirements are even higher at 40 dBm. So, even when only a few strong

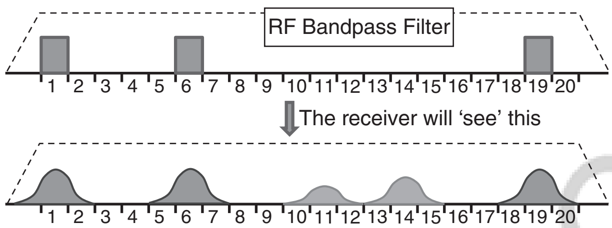


Figure 15.17 Using a good SA, the effect of the receiver linearity on each vacant channel can be calculated, allowing the selection of a channel with achievable requirements. In the scenario shown here, with three large primary signals, only channels 3, 4, 8, 9, 16, and 17 will be usable.

signals are present, large portions of white space become unusable due to the lack of RF-filtering in combination with the limited linearity of the receiver.

It turns out that even in the remaining white space, linearity requirements imposed by the strong interferers are raised significantly due to what is called *cross-modulation*: via third-order non-linearity, the interfering signal modulates the desired signal [2]. Fortunately, the IIP3 requirement in such a channel turns out to be a more reasonable +4 dBm (a similar number is derived in [23]), which is tough, but not impossible to achieve in a low-voltage integrated CMOS solution. Moreover, DSP can aid in reducing the effect of cross-modulation. In [24] a scheme is proposed that uses a second receiver to receive the blocker signal, such that its effect on the desired channel can be estimated and corrected for. Simulations in [24] indicate that for reasonable assumptions, the effective IIP3 for crossmodulation can be increased by 10 dB.

15.6.2.2 Increasing Receiver Linearity

The NF of a receiver is usually dominated by the LNA, because the main purpose of the LNA is to amplify the signal such that the noise of subsequent stages is much less important.

Designing a very linear LNA is far from trivial. The wideband LNA described in [25] is one of the most linear published in or before 2011, with an IIP3 varying from 0 dBm for two closely spaced frequencies to 16 dBm for a tone separation of 160 MHz. The NF is 2.6 dB for a frequency range of 800 MHz to 2.1 GHz. It provides a gain of 14.5 dB and consumes 17 mW. The high linearity is achieved by distortion cancellation via different biasing regimes of transistors, which makes it quite sensitive to bias voltages: 50 mV of change can drop IIP3 by 10 dB. Moreover, one cannot guarantee a large frequency separation between two strong signals, so this high linearity of 16 dBm IIP3 cannot be relied on for CR.

Even when the LNA can be made linear enough, the gain of the LNA (more gain is better for NF) causes distortion of subsequent stages to become the linearity bottleneck. One can think of a receiver where the gain is reduced when the demodulation is limited by distortion, and where the gain is increased when the demodulation is limited by noise. Unfortunately, variable-gain LNAs often have relatively poor linearity. An extreme solution is to bypass the LNA entirely, which allows very linear receiver implementations.

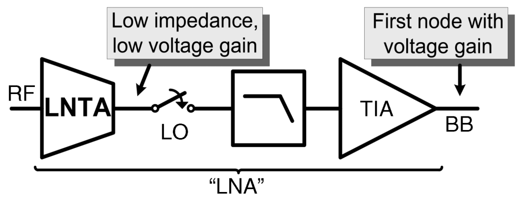


Figure 15.18 High linearity can be obtained by keeping voltage swings low as long as possible.

This latter solution is demonstrated in [26], which proposes a receiver starting with a passive mixer, followed by IF-amplifiers. At IF, feedback is exploited to achieve gain with high linearity. The receiver achieves IIP3 = +11 dBm (limited by the IF-amplifiers) at a NF of around 6 dB and a gain of 20 dB with a RF frequency range of 0.2–2.0 GHz. The upper frequency is solely limited by the on-chip switching frequency, which will improve in newer CMOS-technologies. The passive mixer acts as a frequency converter and an RF band pass filter at the same time (as we will explain in a bit more detail shortly), which significantly improves the linearity for strong interferers that are present outside of the desired band. The mixer itself achieves an IIP3 of +26 dBm.

Another architectural change is proposed in [27]. Whereas traditional LNAs often provide a voltage output, the key idea in [27] is that non-linearity is mainly caused by voltage swings, so the architecture keeps voltage swings low as long as possible, see Figure 15.18. Additionally, high voltage swings are obtained only *after* out-of-band interferers have been attenuated, relaxing RF-filter requirements even more. To implement this idea, the amplifier is a low-noise transconductance amplifier (LNTA), which can be made more linear than a LNA and provides an output *current*. This output current passes through a passive mixer, after which it arrives at the virtual ground node of an opamp, which makes sure that the voltage swing remains low. The current is converted to a voltage using a transimpedance amplifier (TIA), which is an opamp with a feedback resistor. By placing a capacitor in parallel with the feedback resistor, a filter is obtained, simultaneously removing interferers. In total this scheme obtains an in-band IIP3 of +4 dBm, while the out-of-band IIP3 is +16 dBm, where out-of-band is defined as outside of the *RC*-filter bandwidth. This receiver manages to do this with only 4 dB NF.

In [28], the two prior ideas are combined to get the best of both worlds. The complete receiver (up to the ADCs) operates over 0.4–6 GHz, providing up to 70 dB gain, while consuming less than 100 mW. By using digital calibration, IIP2 = 70 dBm. The LNA has a noise-cancelling structure and uses a higher supply voltage to increase linearity. A high linearity is obtained partly because a noise-cancelling LNA turns out to also *cancel distortion* of the matching device. At maximum gain, IIP3 = 10 dBm for signals 20 MHz away from the desired signal at a NF of 3–7 dB. In-band IIP3 is +6 dBm when the LNA is bypassed, at a higher NF of 6–8 dB.

To improve the IIP3 of the entire receiver without compromising gain [29], proposes to use a second 'receiver' (with much more relaxed requirements as compared to the real receiver) which generates the third-order power of the received signal. Via a least mean squares (LMS)-algorithm, it can be used to cancel third-order distortion components (including IM3) in the main receiver. The technique is activated only when necessary, and when active consumes (including DSP) an additional 25 mW. It improves the receiver IIP3 from −9 dBm to +4 dBm.

15.6.2.3 Improved Harmonic Rejection

An additional problem caused by the lack of adequate RF-filtering before downconversion is *harmonic downmixing*. Although ideally the desired signal is downconverted by multiplication with an exponential signal (the LO), in practice the LO will also contain higher harmonics. These higher harmonics may originate in the mixing device, because the mixing is performed with nonlinear components, but also from the increasing trend to use a digital LO. This is shown in Figure 15.19.

One reason to use a digital LO is that a digital frequency synthesizer can generate many different frequencies, but as a square wave rather than a sinusoid. A second reason is the good performance of hard-switching passive mixers, which are simply switches in the signal path that are turned on and off by a digital LO:

- CMOS is very good at switching, and will only get better as technology improves;
- A digital LO is much easier to generate and handle over a wide frequency range (see Section 15.6.3);
- Passive mixers do not carry DC-current, and therefore suffer far less from flicker noise;
- The fundamental is a factor $4/\pi$ larger in amplitude than a full-scale sinusoidal LO.

Mathematically, for a square wave with amplitude ± 1

$$\mathrm{LO}_{\mathrm{sq}}(t) = \frac{4}{\pi}\sum_{k\in Z}\frac{1}{4k+1}e^{j(4k+1)\omega_{lo}t} = \frac{4}{\pi}\mathrm{LO}_{\mathrm{ideal}}(t) + \frac{4}{\pi}\sum_{k\in Z\setminus\{0\}}\frac{1}{4k+1}e^{j(4k+1)\omega_{lo}t} \qquad (15.19)$$

such that the LO has essentially become a parallel combination of multiple LOs (note that all odd harmonics are present, but each at either a negative or a positive frequency).

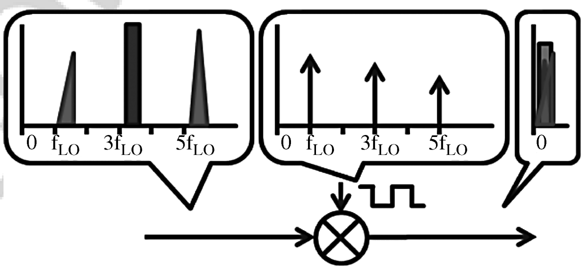


Figure 15.19 Harmonic downmixing is a fundamental problem when RF-filtering is lacking.

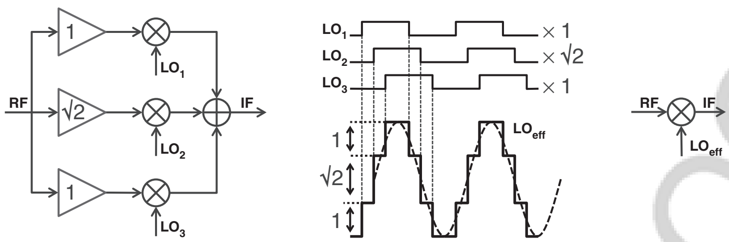


Figure 15.20 Appropriate weighting of different square wave LO-phases yields a closer approximation to a sine wave, effectively removing the third, fifth, eleventh, thirteenth, (and so on) harmonics, leaving the seventh and ninth harmonics as the first uncancelled ones.

For example, the fifth harmonic is only 14 dB weaker than the fundamental. So, when a strong interferer is present at $5f_{lo}$, it will be mixed down to the same frequency as the desired signal, obscuring it and making demodulation impossible. Therefore, HR is desired, that is the suppression of downconversion of the signals present at those harmonic frequencies.

Weldon [30] tackles this problem by combining several mixers in such a way that the total conversion wave looks more like a sine wave. The basic idea is depicted in Figure 15.20. By separately amplifying the (differential) input signal with three (differential) amplifiers, with a relative gain of 1, $\sqrt{2}$, and 1, and then mixing them down with an LO consisting of multiple phases separated by 1/8 of a period, simple Fourier analysis shows that effectively the (even harmonics and the) 3rd and 5th harmonics are cancelled, leaving the seventh harmonic as the first uncancelled one. Mismatch limits the rejection of the third and fifth harmonics to roughly 40 dB. Nevertheless, the suppression strongly relaxes RF filter requirements, allowing a few external wideband filters, or lower-Q integrated filters, to remove the signals at these high-order harmonic frequencies.

Even better HR can be obtained by cascading two HR-stages [27]. Both stages provide approximately the $1:\sqrt{2}:1$ ratio, but by clever design the amplitude mismatches within the two stages are multiplied; that is when the mismatch in each stage is 1%, the total mismatch is 0.01%. The receiver robustly obtains more than 60 dB of HR for the 2nd, 3rd, 4th, 5th, and 6th harmonic. More stages can be used for better amplitude accuracy, but phase accuracy then dominates, similar as in Figure 15.14. For 60 dB of HR, the phase error between adjacent LO-phases must be less than 0.1°, so more improvement becomes extremely difficult.

More HR can be obtained by DSP, such as adaptive interference cancellation (AIC). For each interferer to be removed, an additional observation of the received signal is required, because each interferer has to be estimated before it can be subtracted from the received signal, complicating the design. Therefore, [27] uses one additional observation. The additional suppression of only one harmonic component will often be good enough, because the probability of having multiple strong interferers simultaneously present, each at a different harmonic frequency, is small. AIC works better when the interferer is stronger, because then it can be estimated more reliably. Therefore, the

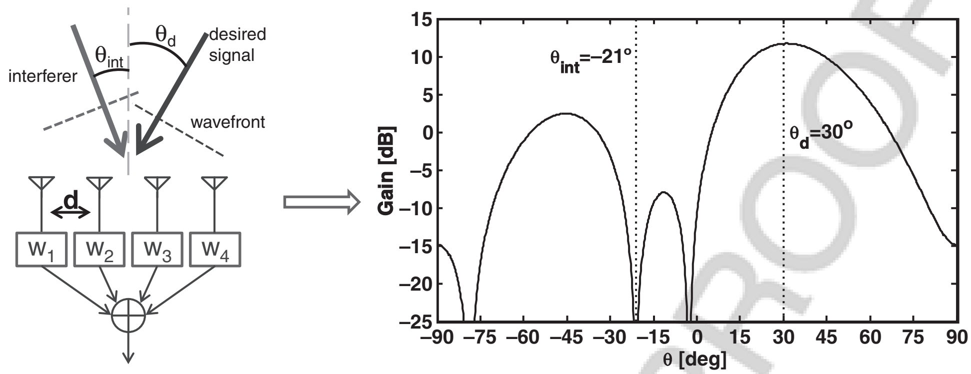


Figure 15.21 Beamforming provides a means for spatial filtering to suppress interferers and lowers the NF by providing passive gain.

interferer will always have roughly the same power after the AIC algorithm. A HR of 80 dB has been reported in [27] by using this technique.

15.6.2.4 Spatial Filtering

The trend for modern wireless standards is to use multiple-input multiple-output (MIMO)-architectures, that is to use multiple receive and/or transmit antennas. A special case of MIMO is receiver beamforming, where the receiver combine(s) the inputs at the N different antennas via weighted addition, as is shown in Figure 15.21. Every weight w_i is complex, with a phase φ_i and amplitude A_i, that is $w_i = A_i\, exp(j\varphi_i)$.[3]

Assume the received signal at antenna i consists of the desired signal and K interferers at the same frequency:

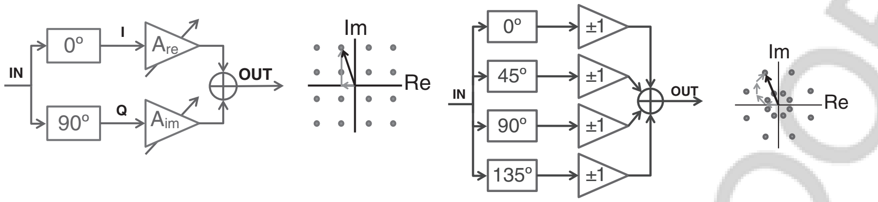
$$\hat{c}_i(t) = m(t)e^{j(\omega_{lo}t+\alpha_i)} + \sum_{k=1}^{K} v_{i,k}(t)e^{j(\omega_{lo}t+\beta_{i,k})} \tag{15.20}$$

where the phase shifts α_i and $\beta_{i,k}$ occur due to the physical distance between the antennas and the location of the desired signal and the interferers. By properly setting the phases φ_i and gains A_i, the gain is increased and NF is lowered, while simultaneously one or multiple strong interferers cancel. Mathematically, the goal is to get as close as possible (with some optimization criterion) to

$$\sum_{i=1}^{N} w_i\hat{c}_i(t) = Nm(t)e^{j\omega_{lo}t}. \tag{15.21}$$

Larger N gives more degrees of freedom, and hence better ability to suppress interferers. The antennas should be placed at a distance of approximately half a wavelength for optimal spatial filtering and main beam width, which limits the applicability for mobile devices to GHz-range frequencies.

Q1 [3]



(a) **Cartesian combining** (b) **Phase oversampling**

Figure 15.22 Applying complex weight to signals can be implemented in several ways. (a) Cartesian combining. (b) Phase oversampling.

To lower the DR- and linearity requirements of the analogue circuitry, including ADCs, the summation should be performed in the analogue domain as early in the chain as possible. The complex weights required for addition of the desired signal and cancellation of the interferers then need to be implemented in the analogue domain. Several techniques exist to do this, each with its advantages and disadvantages. Two of these techniques are shown in Figure 15.22.

The use of Cartesian combining is often proposed. Here, the basic idea is to separately amplify the *I* and *Q* part of the received signals, and then add them together, see Figure 15.22a). Per antenna, one then has to calculate

$$\hat{c}(t) \cdot w = \hat{c}(t) A e^{j\varphi} = A(r(t)\cos\varphi - \breve{r}(t)\sin\varphi) + jA(r(t)\sin\varphi + \breve{r}(t)\cos\varphi), \quad (15.22)$$

where r(t) corresponds to the I-part, and $\breve{r}(t)$ to the Q-part.

The weighted addition of antenna outputs thus mathematically requires the generation of sine- and cosine weights, which is not easy to do in the analogue domain.

In [31] the sine and cosine function are approximated with simple rational functions, see Figure 15.23. Such rational functions can be efficiently implemented using a few switches and capacitors, which are both very linear in CMOS, via charge sharing. The summation of the weighted antenna signals is then performed in the current domain at IF. Measurements show 6 dB improvement in SNR due to the use of four antennas, while null depth (equivalent to the maximum achievable interferer rejection, limited by achievable gain and phase accuracy and resolution) is more than 25 dB.

In [32] it is proposed to use phase oversampling. The basic idea, shown in Figure 15.22(b), is that the variable gain consists now of only +1 or −1, which in a differential implementation means just interchanging the wires. In other words, the (nonlinear) amplification can be removed. By combining multiple non-orthogonal phase shifts (hence the name ‘phase-oversampling’), each weighted with this +1 or −1, both the phase and gain can be set. The obtained interferer rejection is measured to be more than 24 dB, similar to [31].

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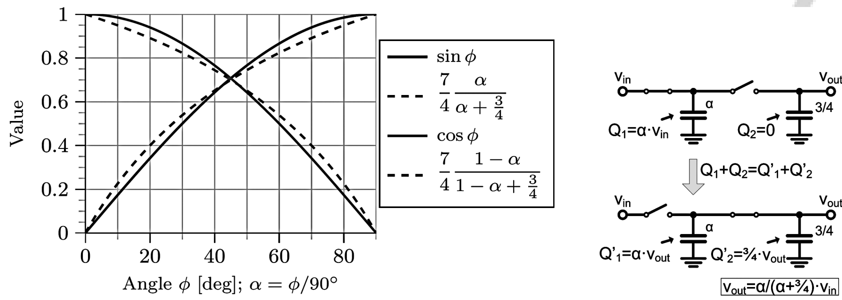


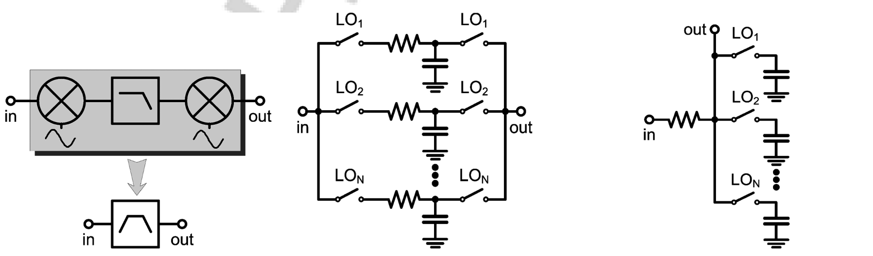
(a) **Approximation** (b) **Implementation**

Figure 15.23 The use of a rational function to approximate the sine-function allows complex weights to be easily generated in the analogue domain, thus reducing DR-requirements further on in the analogue receiver [31]. (a) Approximation. (b) Implementation.

15.6.2.5 Tuneable RF-Filter

To cover a wide frequency range without additional linearity requirements, a tuneable RF-filter would be very convenient. Such a filter could also be used to enable frequency division duplexing (FDD): simultaneous transmission and reception using the same antenna, where the tuneable filter protects the receiver from the transmitted signal. A recent development makes this closer to reality [33,34]. Although the concept is at least 50 years old, it was used in an entirely different context, and has only recently been rediscovered.

The basic idea is that ideal downconversion, low-pass filtering, and upconversion effectively provides band pass filtering, as shown in Figure 15.24(a). A (tuneable) LPF is easily integrated on-chip, and flexible downconversion and upconversion can be implemented with mixers controlled by a digital LO. For simplicity, let us implement the LPF with an *RC*-filter and the mixers with switches, and forget the problem of harmonic downmixing, see Figure 15.24(b). At least four paths are required for decent



(a) **BPF implemented as LPF with down/upconversion** (b) **Straightforward implementation** **(c) Using a shared resistor and removing redundant switches**

Figure 15.24 A tuneable BPF can be implemented as the cascade of a downconversion mixer, LPF, and upconversion mixer, with a surprisingly simple circuit implementation. (a) BPF implemented as LPF with down/upconversion (b) Straightforward implementation (c) Using a shared resistor and removing redundant switches.

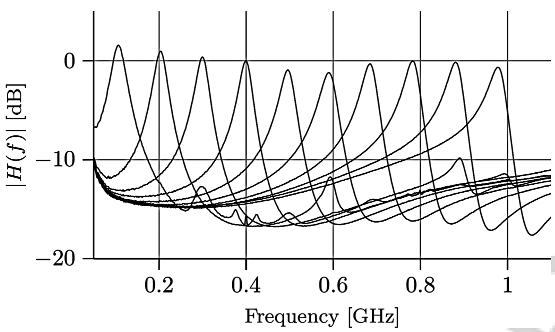


Figure 15.25 Measurements of the 65 nm CMOS implementation of [34] (the circuit shown in Figure 15.24c).

operation, because the zero-IF mixing requires image rejection, and a differential implementation is needed to suppress even-order harmonics, common-mode disturbances, and LO-radiation. The mathematics of these circuits are quite complicated, see for example [35], but the operating principle is relatively easy to understand.

To reduce the loss, it is beneficial to always have exactly one switch conducting. In this example with four paths, each switch is therefore controlled by an LO with 25% duty cycle. This means that the resistor can also be moved to the input, or effectively, that the resistor is simply the output resistance of the driving source. With the downconversion and upconversion switch operating simultaneously, one might as well remove the second set of switches, and take the output before the first set of switches. The resulting circuit becomes like shown in Figure 15.24(c).

Measurements of such a filter, designed in 65 nm CMOS [34], are shown in Figure 15.25. The filter can be tuned from 100 MHz to 1 GHz. This filter has an excellent linearity of IIP3 = 14 dBm, but a limited out-of-band attenuation of 20 dBm due to the on-resistance of the switches (in the order of 5 Ω). Apart from the loss of 2 dB, the filter has poor harmonic rejection: the third harmonic is mixed to the same frequency with only 10–15 dB of attenuation. This means that an additional filter needs to be placed in front of this tuneable filter, increasing the loss even more. Nevertheless, this architecture has promising characteristics, because it uses only switches and capacitors, both of which will always be available in CMOS, and especially switches will only get better when feature sizes decrease.

Broadcom takes this approach a step further to implement a quad-band SAW-less receiver [13], see Figure 15.26. An on-chip transformer takes care of the matching, while two tuneable RF-filters are used in the LNA to improve the stop band attenuation. In combination with a similar third filter, but now used as a mixer (similar to [26]), out-of-band interferers are sufficiently attenuated. In total, this receiver obtains 3.1 dB NF at an IIP3 of −12 dBm. As rightly remarked in [13], this NF is about 1dB better than that of ‘ordinary’ receivers when taking the IL of external SAW-filters into account.

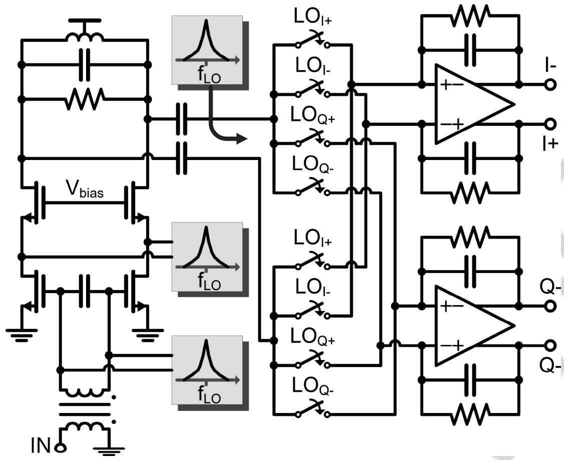
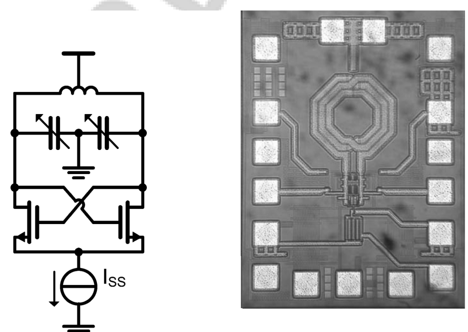


Figure 15.26 The quad-band receiver of Broadcom [13] extensively uses tuneable BPFs to implement a SAW-less receiver.

15.6.3 Wideband Frequency Generation

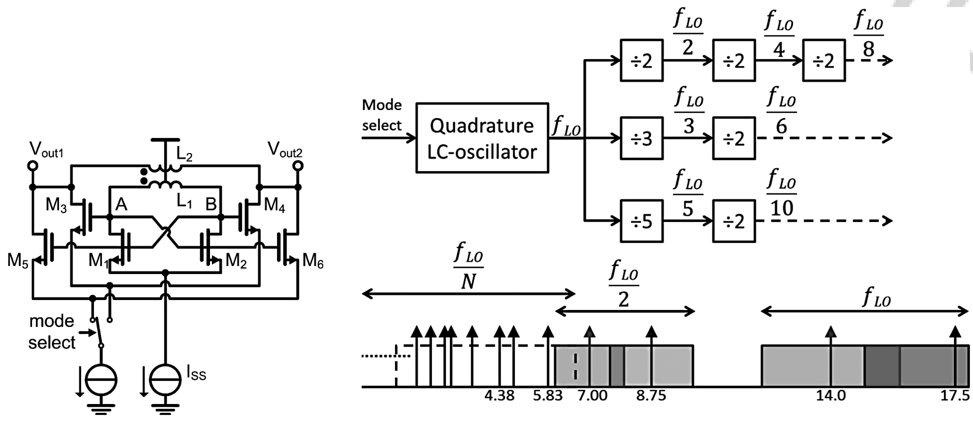
Most standards, such as GSM and WLAN (but with the notable exception of UWB), have an RF-bandwidth less than 10% of their centre frequency. It is then possible to implement a low-power *LC*-oscillator with good phase noise characteristics, and make it tuneable with some variable or switchable capacitors to directly generate any frequency within this RF-bandwidth. A typical circuit schematic and chip layout of an *LC*-oscillator is shown in Figure 15.27. As can be observed, integrated inductors are relatively large. As an example, an inductor of 3nH with Q > 10 easily occupies 250 μm by 250 μm, an area which can also house a complete microprocessor in modern CMOS-technology.



(a) **Typical circuit schematic**

(b) **Circuit layout (Reproduced by Permission from IEEE [36])**

Figure 15.27 An LC-oscillator occupies a significant portion of chip area (a) Typical circuit schematic. (b) Circuit layout. (Reproduced by Permission from IEEE [36]. © 2006 IEEE. Reprinted, with permission, from More on the 1/f2 Phase Noise Performance of CMOS Differential-Pair LC-Tank Oscillators, Andreani, P. and Fard, A., Solid-State Circuits, IEEE Journal of, vol 41, no 12, nov 2006).



(a) **Schematic** (b) **Frequency coverage**

Figure 15.28 The bimodal *LC*-oscillator of [1] and frequency coverage. (a) Schematic. (b) Frequency coverage.

More than 15% tuning of these oscillators is possible, but there is a trade-off between tuning range and phase noise. A wideband receiver with many tuneable *LC*-oscillators in parallel would become too bulky. Moreover, integrated inductors do not work well below 1 GHz. A possibility is to make a very high frequency oscillator, and divide this frequency by 2, 3, 4, 5, and so on to cover all the desired frequency bands. The oscillator must then be running in the order of 100 GHz to cover all frequencies below 10 GHz with 10% tuning, which in itself is possible. However, the dividers have to operate at similar speeds, which is far from trivial, and very power-hungry.

A more scalable solution is to use two *LC*-oscillators, each tuned to such a frequency that any desired frequency can be obtained by selecting the appropriate oscillator in combination with an appropriate division ratio, such as used in [23]. In [1], this idea is taken one step further by using *nested inductors*, further reducing the area requirements. The schematic is shown in Figure 15.28(a). The combination of L_1, M_1, M_2 and I_{SS} is essentially the same as shown in Figure 15.27(a), where the capacitance is formed by the transistors and the load (not shown). The pair (M_3, M_4) or (M_5, M_6), in combination with L_2, creates an additional circuit like Figure 15.27(a). Only one of the pairs is activated by the mode select, with the only difference that nodes A and B are interchanged in this second oscillator. This effectively changes the sign of the mutual coupling between L_1 and L_2, and thus the resonance frequency. Depending on this mutual coupling, the *LC*-oscillator oscillates at either 14 or 17.5 GHz.

By employing variable capacitors for 14% tuning range (not implemented in [1]), and integer frequency division (which is well feasible at these frequencies), this bimodal oscillator can cover the whole range from 50 MHz to 10 GHz, as is shown in Figure 15.28b). The power consumption is 31 mW with a phase noise ranging from −91 dBc/Hz to −120 dBc/Hz at 1 MHz offset, which is close to or in the range of requirements for most current wireless standards (referred back to 1 MHz offset: −100 dBc/Hz for DECT, −110 dBc/Hz for WLAN, WCDMA and Bluetooth, −122 dBc/Hz for UMTS, and −130 dBc/Hz for GSM [37,38]).

For lower frequency ranges, such as used for 802.22, a direct digital synthesizer (DDS) can be used. A DDS consists of a frequency reference and digital logic that outputs digital words based on the desired output frequency. Followed by a DAC and reconstruction filter, any frequency (or arbitrary periodic waveform) can be generated with a resolution set by the reference frequency. Such an approach is used in [39]. It has the additional advantage that it can change its frequency almost instantaneously. The downside is the relatively high power consumption and the presence of spurious frequency content due to the digital implementation. The latter is mitigated in [39] by dithering (i.e. intentionally adding noise) to reduce the spurs to below −35 dBc. In total, the synthesizer in [39] can be tuned from 100 MHz to 2.5 GHz with 15 Hz resolution and with a phase noise of around −130 dBc/Hz at 1 MHz offset at a power consumption of 120 mW.

15.7 Transmitter Considerations and Implementation

In this section, we will provide a basic overview of transmitter considerations and implementations in order to provide the context for the more CR-specific challenges. Compared to receivers (as discussed in Section 15.5), transmitters do not suffer from undesired interferers, which lowers the DR-requirements of components. On the other hand, transmitters face strict regulations on what is allowed to be radiated by the antenna.

Most modern transmitters generate the zero-IF or low-IF versions of the baseband signal in the digital domain, as shown in Figure 15.29. Then, via a DAC and LPF to filter out the harmonics caused by the zero-order hold (ZOH) of the DAC and reduce the far-out quantization noise, it is brought to the analogue domain. A mixer then brings the signal to RF, where it is followed by a power amplifier (PA) to amplify the signal before it is transmitted by the antenna. An external filter is used to filter out undesired components and noise outside of the desired band, in order to comply with emission regulations. Like receivers, transmitters can use direct-conversion or (super)heterodyne conversion, and can include filters and/or Hilbert-transforms for image rejection.

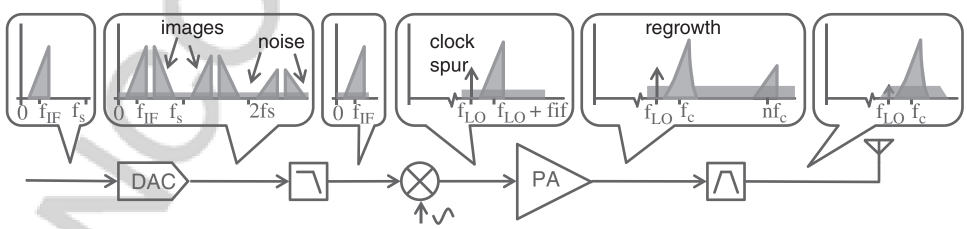


Figure 15.29 Block diagram of a standard transmitter.

Because of the high output power of the PA combined with limited on-chip isolation, it can disturb the LO-generator when it is close in frequency, an effect known as *injection pulling* [18], which results in noise and distortion of the transmitted signal. To make sure that the LO frequency of the last mixer is not close to the output frequency, many transmitters first generate a low-IF signal, either in the analogue or digital domain. An alternative approach is to use a DDS, which is immune for injection pulling [39].

Transmitters in mobile applications typically transmit at a (maximum) power level of 0–33 dBm. If the modulation scheme has a constant envelope, nonlinear PAs can be used, which are highly efficient [40]. Multicarrier schemes, such as OFDM (as often proposed for CR), or higher-order constellations, such as 64-QAM, need a linear PA, which is much less efficient. Moreover, the high peak-to-average power ratio (PAPR) of such signals require the PA to operate far below the maximum output power *on average* (sometimes as much as 20 dB), further lowering the efficiency.

A high output power is often achieved by cascading a driver amplifier and a PA, the latter of which may be external due to the limited supply voltage of modern CMOS. Many standards require the transmitter to have transmission power control (TPC), that is they have to minimize the output power to obtain a certain SNR in order to limit interference to other users. To improve efficiency, the PA itself may be bypassed (and turned off) at low transmission power [40].

The PA and the driver amplifier have to generate a large output power, such that beside the power bottleneck they are also often the linearity bottleneck of the transmitter. Some in-band distortion can be tolerated, as long as the EVM induced by the distortion stays well below the EVM-requirement of the communication standard. The trade-off between linearity and efficiency is alleviated by the *Kahn transmitter* architecture, where the amplitude and phase of the signal are separated. The phase signal, which has a constant envelope, is fed to the PA as the regular input, while the envelope is used to modulate the power supply of the PA. The basic principle is shown in Figure 15.30. This technique is also referred to as *polar modulation* or *envelope elimination and restoration*.

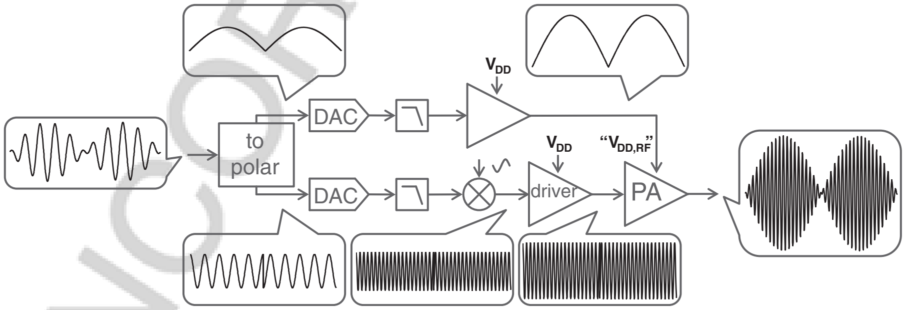


Figure 15.30 The Kahn transmitter separates the phase and envelope of the baseband signal to allow the use of a high-efficiency nonlinear PA.

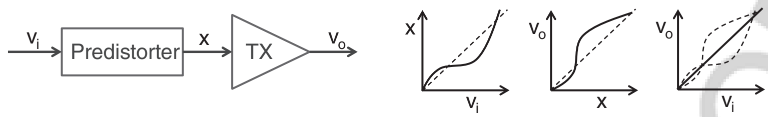


Figure 15.31 Predistortion is a widely applied technique to linearize transmitters.

Another direction to deal with the nonlinear PA is *predistortion*, and is widely employed in transmitters. Since it is exactly known what should be radiated by the antenna, it is possible to distort the baseband signal in such a way that one gets the desired signal at the output of the nonlinear PA (or complete transmitter). The basic principle is shown in Figure 15.31. The achievable improvement is ultimately limited by a number of factors:

- The digitally generated baseband signal has limited accuracy due to quantization;
- The distortion often depends on the history of the signal, which means that exact calculations become very complex and the inverse of the distortion can only be approximated;
- The distortion characteristics change over time due to environmental factors (e.g. temperature) and ageing, which means the output of the PA needs to be monitored to track these changes. Monitoring suffers from measurement errors and noise.

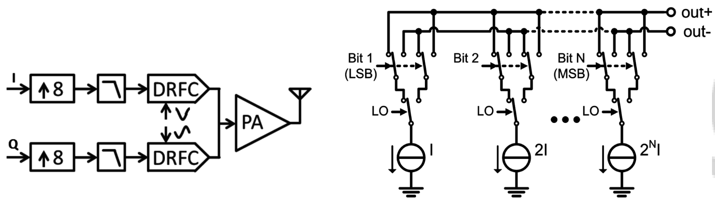
For more information and some other linearization techniques we refer to [40].

15.8 Cognitive Radio Transmitters

A complication in the context of CR is that efficient PA-architectures generally require a tuned high-Q filter at the output. For a narrowband application, this can be solved with an integrated *LC*-filter or a single external filter. For CR, the wideband operation would require many of such filters, or tuneable low-loss high-Q filters. To alleviate the design of such filters, CR-transmitters should be designed with low noise levels at the output of the transmitter, and with low spurious components, such as harmonics of the desired signal, LO-feedthrough and its harmonics, and IM-components. In other words, for CR out-of-band emission is the major bottleneck. Several different approaches are highlighted in the next subsections, which address some of the above mentioned issues.

15.8.1 Improving Transmitter Linearity

One way of reducing distortion components is to make the transmitter more linear. To obtain high integration in CMOS, [41] proposes a direct-digital RF (DDRF)- architecture, where the different analogue functionalities of a direct-conversion transmitter (such as a DAC, reconstruction filter and upconverter) are all integrated into one functional block, the digital-to-RF converter (DRFC), as is shown in Figure 15.32(a). In order not to suffer from DAC images, 8-times oversampling plus filtering is used in the



(a) **Architecture** (b) **Implementation of DRFC-block**

Figure 15.32 The DDRF-architecture as proposed by [41] combines most of the analogue functionality of a direct-conversion transmitter in a single block. (a) Architecture. (b) Implementation of DRFC-block.

digital domain. The digital output words in combination with a digital LO are directly used to convert the signal to RF, as is shown in Figure 15.32(b). This has a number of important advantages:

- No (nonlinear) variable-gain amplifier (VGA) is needed, as the DAC (or DRFC) directly determines the current, and thus the gain;
- Because the digital baseband signal is brought directly to the mixer switches, it is immune to analogue DC-offset in the baseband signal, and IQ-imbalance is reduced;
- The extensive use of switches gives a high linearity.

15.8.2 Reducing Harmonic Components

As discussed in Section 15.6.2, a HR-mixer can be used in the analogue downconversion process to implement HR. The same can be done in the upconversion process. When the range of centre frequencies of the transmitter is limited, such as is the case for 802.22, suppression of the third and fifth harmonic may be enough.

In [42], the 54–862 MHz band is split into two parts, where the lower frequency range uses HR-mixers. For the upper frequency range, image-reject mixers suffice, as the third harmonic is above 900 MHz. This allows lower-power operation at these higher frequencies. For the lowest frequencies in the lower frequency range, the seventh and ninth harmonics are also inside the bandwidth of the transmitter. These harmonics are tuned out by two notches at RF, implemented via active filter stages with a fixed inductor and variable resistors and capacitors. In total, harmonic distortion components are kept below −42 dBc.

A set of two quadrature square wave signals can be modelled as a complex signal, as was given in Equation 15.19. The first, fifth, ninth, . . . harmonic have a positive frequency, while the third, seventh, eleventh, . . . have a negative frequency. This is exploited in [43], where the third harmonic is removed using a polyphase filter. The LO-generation is done with a ring oscillator in such a way that the fifth and higher order harmonics of the LO are already very low, that is it looks more like a sine wave than a square wave. It achieves an output spectrum where all distortion products are below -40 dBc, because the higher harmonics are already quite low.

15.8.3 The Polyphase Multipath Technique

The polyphase multipath (PMP)-technique is a combination of improving transmitter linearity and harmonic rejection by suppressing most (but not all) harmonic and distortion components [44]. The basic idea is to split the nonlinear circuit into N identical circuits, and place (real) phase shifters before and after these N circuits, as shown in Figure 15.33.

By choosing the phase shifts as $\varphi_n = 360° \cdot n/N$, it can be shown that all harmonics are cancelled, except at $mN+1$, with $m \in N$. The simplest example of a PMP-circuit is a well-known differential circuit driven with balanced (anti-phase) input signals. It cancels all even harmonics, but not the odd harmonics.

If the nonlinear system is excited by a two-tone input signal $x(t) = A_1\,cos 2\pi f_1 t + A_2\,cos 2\pi f_2 t$, besides harmonics the output will also contain intermodulation products at new frequencies $pf_1 + qf_2$, with $p, q \in Z$. It can be shown that most intermodulation products are cancelled, except if $p+q = mN+1$, with $m \in N$[44]. The most important intermodulation products $2f_1 - f_2$ and $2f_2 - f_1$ are thus not cancelled.

One of the difficulties of the PMP-technique as presented in Figure 15.33 is the second set of phase shifts. For a transmitter, the first set of phase shifts can be implemented in the digital domain, and only needs to be accurate over the bandwidth of the baseband signal. At the output of the nonlinear circuits, the higher-order terms of the transfer have generated terms across a very wide band. For proper suppression, the second set of phase shifts needs to be accurate over this very wide band, and this is very difficult. However,

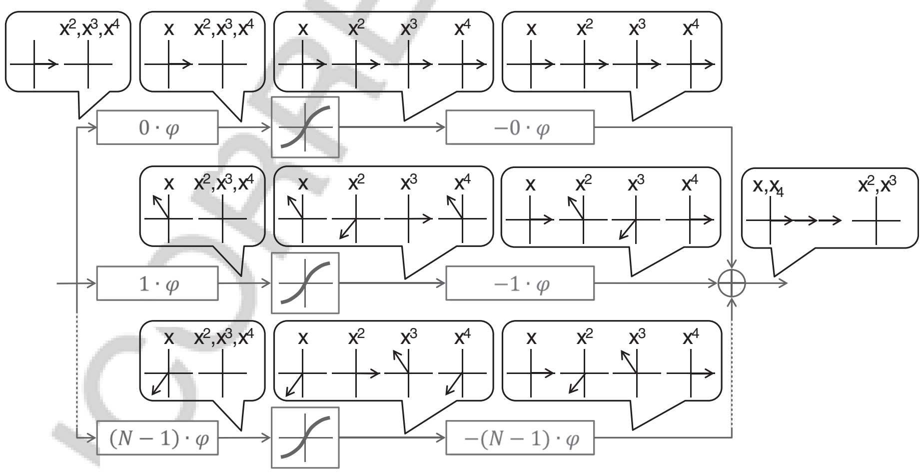


Figure 15.33 The PMP-technique allows cancellation of many harmonics and IM-products. Here, the cancellation of the second and third harmonic are illustrated for a 3-path system.

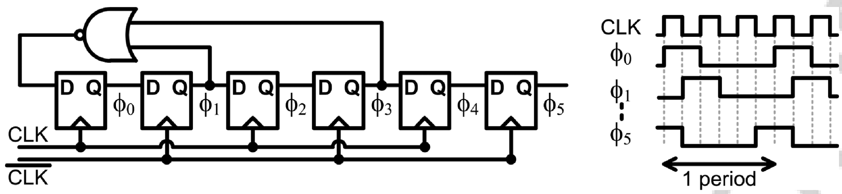


Figure 15.34 Example circuitry for a 1/3 duty cycle 6-phase LO-signal.

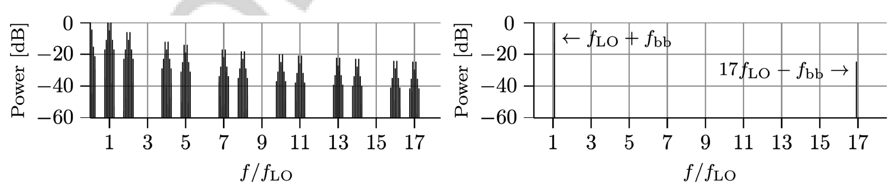
when the second phase shift can be combined with a frequency conversion, the desired phase shift can be applied to the LO instead of the modulated signal (similar as in Figure 15.12), which makes it again feasible.

The *N*-phase LO-signals can be generated relatively easily using a chain or ring of flip-flops, and the duty cycle can be set with simple combinatorial feedback. An example is shown in Figure 15.34.

When the second set of phase shifts is implemented by a hard-switching mixer, the output spectrum of an individual path becomes even more crowded due to the mixing of the output of the nonlinear circuit with the many LO-harmonics, see Figure 15.35(a). Again, relatively simple calculations can be used to find out which ones are cancelled and which ones are not with the PMP-technique: Spectral components may pop up at frequencies pf_{lo}-qf_{bb}, due to the multiplication of the square wave LO with the baseband input signal at f_{bb}. In this scheme, all products are cancelled except when $p = q + mN$, with $m \in Z$. This means that, besides other terms, all terms where $p = q$ are not cancelled, such as $3f_{lo} + 3f_{bb}$.

One can remove the $3f_{lo}$ components from the output by removing the $3f_{lo}$ component from the mixer, for example by using HR-mixers. An alternative implementation is to change the duty cycle *d* of the LO-signal to 1/3, as shown in Figure 15.34. Using a differential 18-path transmitter and a duty cycle of 1/3, the strongest remaining harmonic component would be $17f_{lo} - f_{bb}$ (see Figure 15.35b), and this is indeed what is measured in [45], at a power of −31 dBc. Since it is so far from the desired signal, a simple LPF would be sufficient to suppress this component to negligible levels.

Interestingly, the use of more paths makes the scheme less sensitive to mismatch [44]. On the other hand, the generation of a large number of LO-phases can become



(a) **Output spectrum when *N=1, d=1/3*** (b) **Output spectrum when *N=18, d=1/3***

Figure 15.35 With the PMP-technique, many harmonics, intermodulation products and LO-terms can be cancelled. (a) Output spectrum when $N = 1$, $d = 1/3$ (b) Output spectrum when $N = 18$, $d = 1/3$.

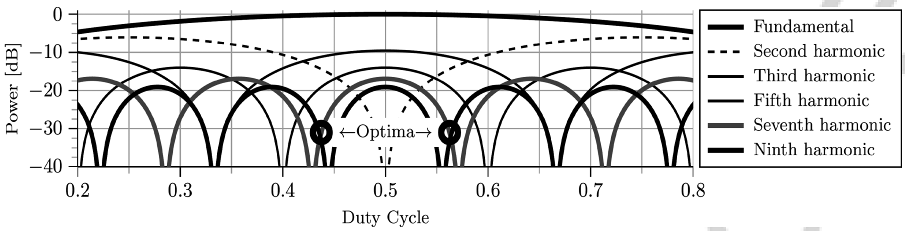


Figure 15.36 By properly choosing the duty cycle, the seventh and ninth harmonic in an 8-path PMP-system can also be partly suppressed.

troublesome, as well as the requirement to have so many DACs. Therefore, [46] proposes an 8-path PMP-transmitter with suppression of the seventh and ninth harmonic by tuning of the LO duty cycle. The strengths of the LO-harmonics as a function of duty cycle are plotted in Figure 15.36, with the fundamental at 50% duty cycle as the reference. With a duty cycle of 7/16 or 9/16, the seventh and ninth harmonic in the LO-waveform are suppressed by an *additional* 14 dB. Together with a simple tuneable LPF, this technique can reduce all harmonic components to below −40 dBc.

The PMP-technique requires a mixer after the nonlinear circuit, which eliminates the use of certain transmitter architectures. In [47] an alternative is presented, depicted in Figure 15.37, where the outputs of the 2 N nonlinear circuits are directly added without additional phase shifts, such that an arbitrary transmitter architecture can be used. The negative phase shift is used as the de-rotation operation that was present after the nonlinear circuit in the PMP-architecture. The phase shifts can now be chosen rather freely because they can all be generated in the digital domain.

The advantage is that all phase shifts can now be generated in the digital domain and thus does not require a high-frequency N-phase LO, but this system does not cancel

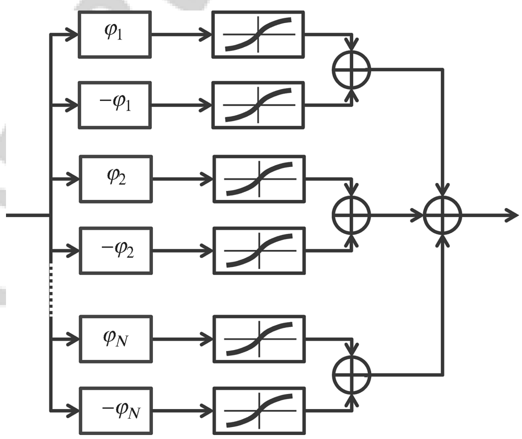


Figure 15.37 A variant of the PMP-technique does not require phase shifts at the outputs of the nonlinear circuits, thereby allowing arbitrary transmitter architectures.

LO-harmonics. One of the LO-harmonics is the −1st; as a result, it does not have image rejection. Moreover, it has less gain, as the desired signal outputs do not line up perfectly in-phase.

The original PMP-technique is combined with poly-harmonic predistortion linearization (PHPL) in [48] to provide suppression of remaining components such as the IM3-components. The predistortion takes into account both mismatches and memory effects, and is based on optimizing the output spectrum of the transmitter via spectral analysis of this output, which may be provided by the spectrum sensing functionality of the CR. With a two-tone test, a clean spectrum down to −70 dBc is obtained.

15.9 Spectrum Sensing

Spectrum sensing is an essential functionality of a CR to find available white space if no database infrastructure is available. Most work on spectrum sensing focuses on the DSP-part, where the analogue frontend is assumed an 'ideal' device that only adds some white Gaussian noise. As we saw in Section 15.5, such an ideal receiver does not exist in reality; noise is not white in general, and the signal that is obtained at the output of the ADC consists of (a more or less distorted version of) the desired signal plus images, aliased frequency components, spurious tones originating from analogue and digital circuitry, and so on.

These nonidealities hamper detection performance. For example, a receiver with poor HR trying to perform energy detection at 300 MHz will mistakenly identify it to be occupied when a strong signal is present at 900 MHz, leading to a false alarm. It is likely that each sensing technique has a different robustness against each nonideality, but in this section, we will focus our attention on energy detection, where the energy (or power) detected in a channel determines whether it is deemed occupied or not.

From Section 15.2, the regulations impose detection of DTV-signals in highly negative SNR. In [49], it is experimentally shown that due to imprecise knowledge of the noise level (e.g. caused by temperature fluctuations in the receiver), signals below a certain SNR cannot be detected anymore. This minimum SNR is now commonly referred to as the *SNR-wall*, because the required number of samples rapidly goes to infinity. This is shown in Figure 15.38.

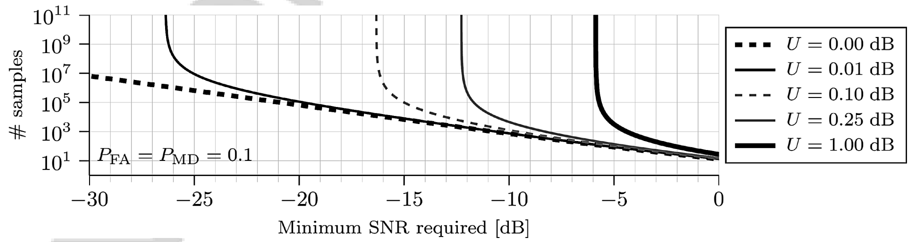


Figure 15.38 The SNR-wall is the minimum SNR required to detect a signal, regardless of the number of samples. For a noise uncertainty U of 1 dB, the SNR-wall is at −6 dB.

In order to provide adequate sensing performance, the receiver should have low noise and high linearity, which together can be captured in the term spurious-free dynamic range (SFDR). The SFDR is defined as the difference in power between the strongest and weakest signal that can be detected at the same time, and is equal to

$$\mathrm{SFDR} = \frac{2}{3}(174 - \mathrm{NF} + \mathrm{IIP3} - 10\log_{10}\mathrm{RBW})[\mathrm{dBm}] \tag{15.23}$$

when only NF and IIP3 are taken into account, neglecting factors such as IIP2, oscillator phase noise, clock spurs, and so on. The resolution bandwidth (RBW) is the frequency resolution: a lower RBW means that less noise power will be present in such a band, and hence the SFDR increases. The useful increase in SFDR by lowering RBW is limited by the bandwidth of the signals to be detected; at some point, the signal power will also drop, such that the SNR is not further increased when RBW is lowered.

To visualize the impact of a limited SFDR, we have simulated the output spectrum of a SA with $\mathrm{IIP3} = +1$ dBm and $\mathrm{NF} = 5$ dB with a $\mathrm{RBW} = 100$ kHz ($\mathrm{SFDR} = 80$ dB). At the input (Figure 15.39a), some sine waves are present, where circles indicate their

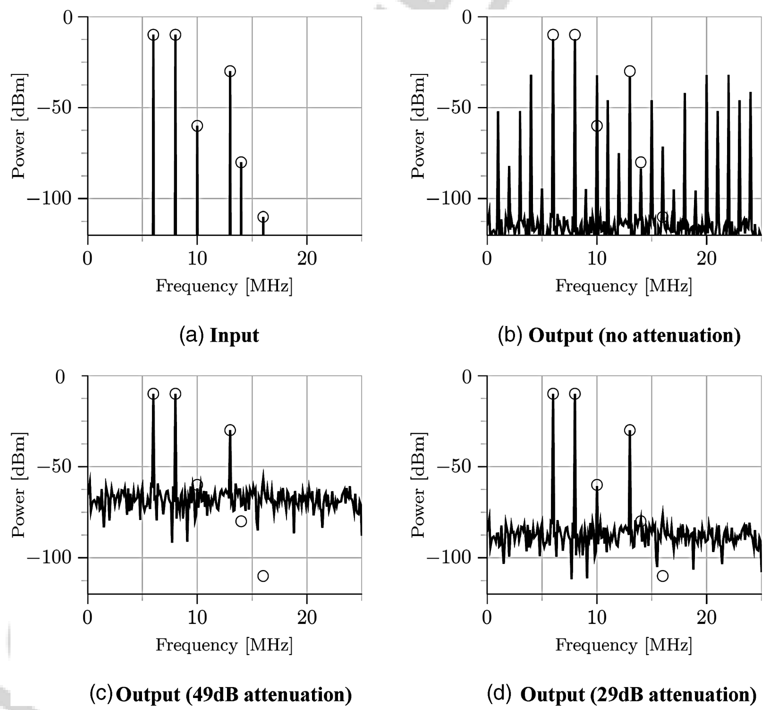


Figure 15.39 Simulation of energy detection using a receiver with $\mathrm{NF} = 5$ dB, $\mathrm{IIP3} = +1$ dBm ($\mathrm{SFDR} = 80$ dB in $\mathrm{RBW} = 100$ kHz) which is equipped with an ideal attenuator at the input. (a) Input. (b) Output (no attenuation). (c) Output (49 dB attenuation). (d) Output (29 dB attenuation).

power levels for easy reference. At the output, the spectrum looks quite different and depends on the attenuation (the power levels are referred to the antenna), see Figure 15.39(b)–(d). When the linearity is limited by the LNA, it can only be improved by attenuating the signal in front of the LNA. Assuming a matched system and an ideal attenuator, xdB of attenuation raises both NF and IIP3 by xdB. At low attenuation, the strongest signals generate many intermodulation products, which may generate false alarms. At high attenuation, the increased noise obscures weak signals, which may generate missed detections. Even at the optimum attenuation (in this case, 29 dB for a frequency resolution of 100 kHz) where the noise and distortion products are at the same level, some signals cannot be detected, see Figure 15.39(d). Note that this balancing of distortion peaks and noise can be performed quickly and easily in the digital domain, as the SA roughly knows its NF and IIP3, and the strong input signals are readily detected.

The sensing device is allowed to occasionally make mistakes (see Section 15.2), so the SFDR requirements can be reduced by ignoring situations that rarely occur. For example, input signals stronger than 0 dBm are almost never received [50]. Nevertheless, it is not uncommon that received signals have powers in the range of -10 dBm, while the requirements state that signals with a power of -114 dBm should be detected. This still calls for a very high SFDR.

Two extreme situations can occur:

1. No strong signal is present. Then only NF plays a role, and the problem is to detect signals in highly negative SNR;
2. Strong signals are present. The SFDR is key to the ability to detect relatively weak signals in the presence of these strong signals.

The next few sections contain solutions to improve the performance for one or both of these situations.

15.9.1 Analogue Windowing

Ideally, the ADC captures a large chunk of spectrum, after which for example a fast Fourier transform (FFT) is used to parallel process all the channels inside this chunk, minimizing measurement time. As is well known, an FFT suffers from spectral leakage due to the truncation in the time domain. Therefore, the digital samples are often windowed to reduce this effect. Mathematically, with $x(t)$ the (relatively wideband) signal, and $w(t)$ the window, one obtains

$$\int_{0}^{T_w} w(t)x(t)dt = \int_{-\infty}^{\infty} w(t)x(t)dt = \int_{-\infty}^{\infty} W(f)X(f)df, \tag{15.24}$$

where T_w denotes the length of the window. In other words, windowing in combination with an integration operation has a filtering effect.

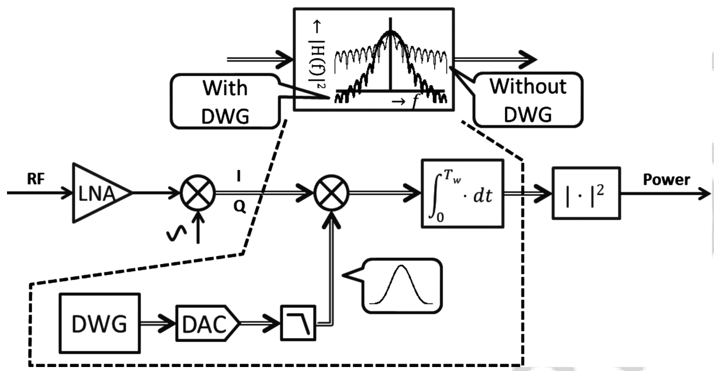


Figure 15.40 Spectral leakage can be reduced by time-windowing the signal, as is done in the analogue domain in [51].

An analogue implementation of this idea is proposed and implemented in [51]. The analogue window is generated in a digital window generator (DWG), which is a digital memory plus DAC and LPF. This digital generation allows the characteristics, such as bandwidth, window length, and out-of-band suppression to be very flexible. This reduces the requirements on bulky analogue filters. As an addition to an integrated CMOS UHF-band receiver, the SA shares the RF-part, but uses a separate baseband path including the analogue windowing. The SA-part is shown in Figure 15.40. The measured suppression of adjacent channels is 35 dB.

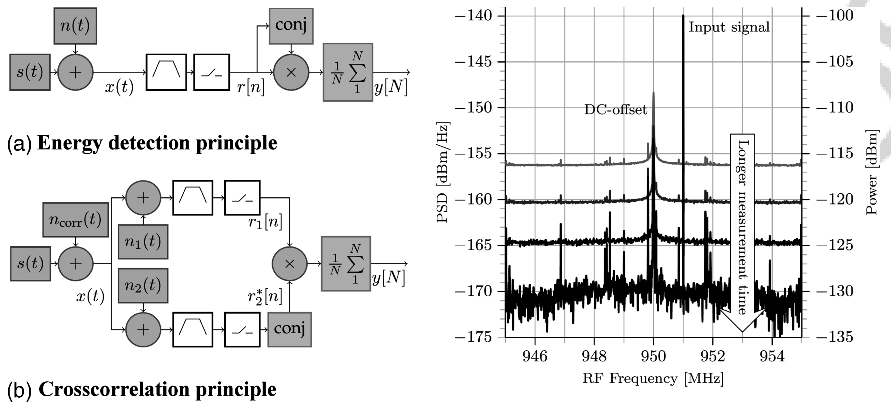
15.9.2 Channelized Receiver

In the context of UWB, where strong interferers may be present in the same band as the desired signal [52], suggests the use of a channelized receiver to handle several separate frequency channels in parallel in the analogue domain, such that the requirements per path and per ADC are relaxed. The different bands are combined in the digital domain.

A similar solution could be used for sensing. When each path is used for single-channel sensing, image rejection problems can be removed by using a zero-IF architecture. The performance improvement, however, will be limited due to the poor scalability of this approach: each path requires a different LO-frequency, additional filters and additional chip area.

15.9.3 Crosscorrelation Spectrum Sensing

The existence of the SNR-wall suggests that one should try to lower the receiver noise to improve the SNR. As we saw in Section 15.5, lower NF often means lower linearity. A combined analogue/digital spectrum sensing solution to break this trade-off is proposed in [53]. It takes advantage of the additional freedom that spectrum sensing does not require signals to be demodulated.



(a) **Energy detection principle**

(b) **Crosscorrelation principle**

(c) **Crosscorrelation measurement showing the reduction in noise level**

Figure 15.41 Crosscorrelation spectrum sensing lowers the effective receiver noise, allowing the receiver to be designed for higher linearity. (a) Energy detection principle. (b) Crosscorrelation principle. (c) Crosscorrelation measurement showing the reduction in noise level.

A single receiver with output $r[n]$ calculates $|r[n]|^2$ for energy detection, see Figure 15.41(a). The power of the noise $n(t)$ added by the receiver shows up at the output as well. By connecting two receivers in parallel to the antenna, (part of) the noise in one receiver will be independent of the noise in the other receiver, see Figure 15.41(b). This independent noise of each receiver is denoted by $n_1(t)$ and $n_2(t)$, respectively. The correlated noise that remains, for example due to shared components, is denoted by $n_{corr}(t)$.

When performing energy detection by crosscorrelation of the outputs of both receivers, the noise added by each receiver is reduced at the cost of measurement time. The crosscorrelation solution estimates the cross-spectrum, which ideally ($n_{corr}(t) = 0$) converges to the input spectrum:

$$\begin{aligned} E[y] = E\left[r_1 r_2^*\right] &= E[(x + n_1)(x + n_2)^*] = E\left[|x|^2\right] + E\left[x n_2^*\right] + E[x^* n_1] + E\left[n_1 n_2^*\right] \\ &= E\left[|x|^2\right] = E\left[|s|^2\right] + E\left[|n_{corr}|^2\right] \end{aligned} \tag{15.25}$$

So, eventually it is equal to Figure 15.41(a), but with less noise, because $E[|n_{corr}(t)|^2] < E[|n(t)|^2]$, that is the correlated noise power is (much) less than the total noise power generated in a receiver.

Using attenuators [53], manages to improve the linearity at only the cost of higher $n_1(t)$ and $n_2(t)$, such that the final noise floor after crosscorrelation remains at the same level. In effect, the SFDR is improved. For high linearity, the implementation uses the highly linear mixer-first architecture of [26], while the rest of the receiver is built with off-the-shelf components. An attenuator is placed in front of the mixer to provide matching and to increase linearity at the cost of noise. In [53], multiple channels are sensed at once by employing an FFT and correlating the corresponding frequency bins. This approach allows a flexible RBW by simply changing the number of points per FFT. In total, a residual NF of 4 dB is obtained with an IIP3 of 24 dBm, which for 1 MHz RBW results in a SFDR of 89 dB. The maximum improvement in SFDR is limited by the allowed measurement time, set by regulatory restrictions (see Table 15.1) or throughput constraints, while the residual NF is limited by noise correlation in both receivers.

By employing separate LOs for the two receivers, phase noise of these LOs can also be significantly reduced via the same principle, which can save additional power and/or make detection of weak signals very close to a strong signal possible.

15.9.4 Improved Image and Harmonic Rejection using Crosscorrelation

As mentioned in Section 15.9, limited HR can also be a problem. Apart from implementing the analogue frontend with good HR, crosscorrelation in combination with a frequency offset between the receivers can be used to further improve the HR, as proposed in [54].

The basic concept is shown in Figure 15.42. In one receiver, the RF-signal at f_c is downconverted to DC, while in the second receiver, the RF-signal at $f_c + \Delta f$ is downconverted to DC. With respect to harmonic downmixing, the *n*-th harmonic is located at $n{\cdot}f_c$ for receiver 1, and at $n{\cdot}fc + n{\cdot}\Delta f$ for receiver 2. By digitally shifting the output of receiver 2 back in frequency by Δf, the desired signal is now fully correlated with that of receiver 1, while the harmonic mixing products are not. Hence, by crosscorrelating the resulting two outputs, the harmonic images are suppressed at the cost of measurement time. A similar argument holds for the image frequencies that are not perfectly cancelled due to mismatch, so this algorithm also improves the image rejection.

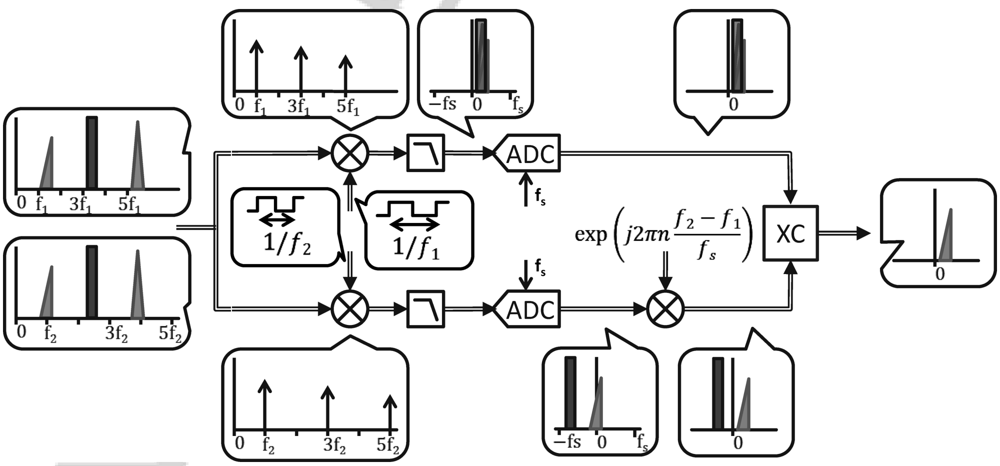


Figure 15.42 Crosscorrelation with an analogue frequency offset and digital correction provides improved HR (as well as image rejection, but this is not shown).

The overlapping frequency range of the two receivers is the IF-bandwidth minus $2\Delta f$, so Δf should not be chosen too large. On the other hand, Δf should be chosen large enough to allow sufficiently fast decorrelation of the harmonic images. For efficient implementation, the crosscorrelation can be implemented with FFTs. When Δf is chosen to be equal to an integer number of RBW, the frequency shift is simply a shift in the FFT-bins.

15.10 Summary and Conclusions

Traditional (narrowband) transceivers heavily rely on high-quality discrete filters that significantly attenuate interfering and out-of-band signals, such that transceiver linearity requirements are relaxed. For high integration, low power and low cost, the analogue and digital parts of a transceiver can be combined using CMOS. For wide-band CR-transceivers, discrete filters are preferably not used, as they are inflexible, bulky and expensive.

In this chapter we have identified the specific challenges the wideband operation and the lack of high-quality filtering pose on a CMOS-implementation of CR, and we have given an overview of existing and proposed solutions towards a fully integrated solution.

For receivers, the main bottleneck is the trade-off between noise and linearity, as well as a good performance over a wide band. For transmitters, PA-efficiency and linearity are the most stringent problems. For spectrum analysis, problems are similar as for receivers, although additional ways of improvement are available because demodulation is not required.

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