# Chapter 2
# Radar Systems

Automotive safety systems require information about the objects in the vicinity of the vehicle. These data are usually obtained by sensing the surroundings. A typical sensor system usually transmits a signal and estimates the attributes of the available targets, such as velocity or distance from the sensor, based on the measurement of the scattered signal. The signal used for this purpose in radar (radio detection and ranging) systems is an electromagnetic (EM) wave at microwave frequencies. The main advantage of radar systems compared to other alternatives such as sonar or lidar is the immunity to weather conditions and potential for lower cost realization.

Section 2.1 describes the principle of radar. There are two main operation principles, continuous wave (CW) and pulsed. The latter is not treated within this scope, since the frequency regulations in the ISM band result in a limitation on the absolute transmitter power. Thus, pulsed radar would result in a lower SNR due to a lower duty cycle compared to the CW operation. Radar operation is discussed in sections 2.2 - 2.4. Frequency regulations around 24 GHz are described in section 2.5. Typical radar architectures and circuit related challenges are presented in section 2.6. Section 2.7 provides an overview of the automotive radar systems and their application for car safety. Finally, section 2.8 concludes this chapter with considerations on technology features needed for radar realization.

## 2.1 Radar Principle

Radar systems are composed of a transmitter that radiates electromagnetic waves of a particular waveform and a receiver that detects the echo returned from the target. Only a small portion of the transmitted energy is re-radiated back to the radar, which is then amplified, down-converted and processed. The range to the target is evaluated from the travelling time of the wave. The direction of the target is determined by the arrival angle of the echoed wave. The relative velocity of the target is determined from the doppler shift of the returned signal.

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For automotive radar applications the separation between the transmitter and receiver is negligible compared to the distance to a target. Thus, these systems are monostatic in a classical sense. However, the automotive radar systems are usually referred to as *bistatic* when two separate antennas are used for transmit and receive and *monostatic* when the same antenna is used for these functions, as depicted in Fig. 2.1. The latter configuration requires a duplexer component to provide isolation between transmitter and receiver. This is usually realized using expensive external bulky transmit/receive (T/R) switch or circulator components. The solution of using hybrid ring coupler [1] offers a cost advantage at the expense of lower performance due to higher losses and increased noise figure.

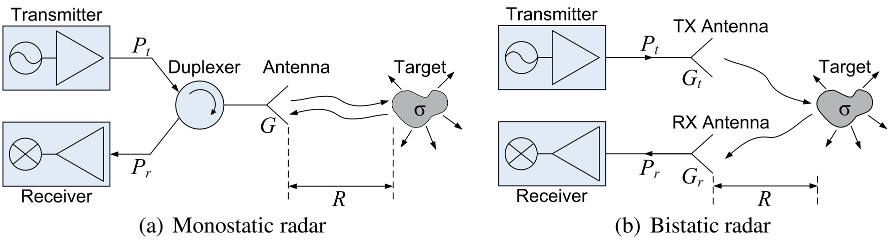


**Fig. 2.1** Radar configurations.

## 2.2 Radar Equation and System Considerations

The radar equation provides the received power level as function of the characteristics of the system, the target and the environment. The well-known bistatic radar equation [2] for the system in Fig. 2.1(b) is given by

$$P_r = \frac{P_t A_{\mathrm{er}} A_{\mathrm{et}} \sigma}{4\pi R^4 \lambda^2 L_{\mathrm{sys}}}, \tag{2.1}$$

where $P_r$ is the received power, $P_t$ is the transmitted power, $A_{\mathrm{er}}$ and $A_{\mathrm{et}}$ are the effective area of the receive and transmit antennas, respectively, $R$ is the distance to the target, $\sigma$ is the radar cross-section (RCS), defined as the ratio of the scattered power in a given direction to the incident power density and $L_{\mathrm{sys}}$ is the system loss due to misalignment, antenna pattern loss, polarization mismatch, atmospheric loss [3], but also due to analog to digital conversion and fast Fourier transform (FFT) windowing. Taking into consideration that the effective area of the receive and transmit antenna is related to the wavelength $\lambda$ and to the antenna gain $G_r$ and $G_t$, as $A_{\mathrm{er}} = G_r \lambda^2/4\pi$ and $A_{\mathrm{et}} = G_t \lambda^2/4\pi$, respectively, the radar equation can be rewritten as

$$P_r = \frac{P_t G_r G_t \lambda^2 \sigma}{(4\pi)^3 R^4 L_{\text{sys}}}. \tag{2.2}$$

Based on the system characteristics and the noise floor of the receiver a certain minimal signal power level $P_{\text{r,min}}$ is required in order to detect the target. Thus, from (2.2) the maximum achievable radar range can be calculated as follows

$$R_{\max} = \left( \frac{P_t G_r G_t \lambda^2 \sigma}{(4\pi)^3 P_{\text{r,min}} L_{\text{sys}}} \right)^{1/4}. \tag{2.3}$$

Furthermore, in most practical designs a minimal signal to noise ratio (SNR) at the output of the receiver $\text{SNR}_{\text{o,min}}$ is considered in order to ensure high probability of detection and low false-alarm rate. Typically, SNR values of higher than 12 dB are required. The noise factor of a receiver is defined as

$$F = \frac{S_i/N_i}{S_o/N_o}, \tag{2.4}$$

where $S_i$ and $S_o$ are the input and output signal levels, respectively, $N_o$ is the noise level at the receiver output and $N_i$ is the input noise level, given by

$$N_i = k_{\text{B}} T B, \tag{2.5}$$

where $B$ is the system bandwidth, $k_{\text{B}}$ is the Boltzmann constant and $T$ is the temperature in Kelvin. Taking into consideration that there is an additional processing gain due to the integration over several pulses, approximately given by $G_{\text{int}} = T_{\text{CPI}} \cdot B$, where $T_{\text{CPI}}$ is the coherent processing interval (CPI), the maximum radar range in (2.3) can be rewritten as a function of $\text{SNR}_{\text{o,min}}$ as follows

$$R_{\max} = \left( \frac{P_t G_r G_t \lambda^2 \sigma T_{\text{CPI}}}{(4\pi)^3 \cdot kTF \cdot \text{SNR}_{\text{o,min}} \cdot L_{\text{sys}}} \right)^{1/4}. \tag{2.6}$$

The attenuation for the propagation of the electromagnetic waves at 24 GHz is about 0.15 dB/km [4]. Taking into consideration that the typical range for automotive radar sensors is up to 200 m, the contribution of the atmospheric attenuation to $L_{\text{sys}}$ is negligible. Even under heavy rain or fog conditions the attenuation over these distances is in the range of few decibels.

The RCS of typical targets in automotive applications ranges from 0.1 to 200 m$^2$. The antenna gain is usually in the range of 15 - 25 dBi. Antennas are typically realized as patch antenna arrays for beam shaping. Their large size at 24 GHz limits the dimensions of radar modules.

Equation (2.6) can be rearranged for the noise factor $F$. Plugging in the smallest RCS and the largest required distance of operation results in the required receiver noise figure. For example, for an object with a $\sigma$ of 0.1 that has to be detected at a maximal distance of 100 m with transmit and receive antenna gains of 20 dB, transmitter power of 0 dBm, the system losses of 3 dB, the CPI time of 2 ms and minimum required SNR after the FFT of 12 dB, the required receiver front-end noise figure is 10.75 dB. For a typical narrow-band 24 GHz system a single side-band (SSB) noise figure (NF) of less than 10 dB is needed. The NF is related to the noise factor in (2.4) as $\mathrm{NF} = 10 \cdot \log(F)$. The gain of a receiver front-end is less crucial, since it can be compensated in the baseband stage. However, it still has to be above 10 dB for a low receiver NF, due to noise figure cascading.

Another limiting case, referred to as the *blocker* case, is the scenario of a large target with maximum RCS being present very close to a radar at a minimal distance of operation. This sets the requirement on the front-end linearity in terms of input-referred 1dB compression point (IP1dB), which should be typically above $-15$ dBm. Combination of both mentioned limiting cases results in a requirement on the receiver's dynamic range (DR), which usually should be above 70 dB.

## 2.3 CW and Frequency-Modulated Radar

### *2.3.1 Doppler Radar*

A classical continuous wave (CW) or Doppler radar implementation uses a fixed transmit frequency to detect a moving target and its velocity. It is based on the Doppler frequency shift. If there is a non-zero relative velocity $v_r$ between a radar transmitter sending a signal at frequency $f_0$, and a moving target, the returned signal has frequency $f_0 + f_d$, where $f_d$ is the Doppler frequency shift given by

$$f_d = \frac{2v_r}{c} f_0, \tag{2.7}$$

where $c$ is the speed of light. The relative velocity $v_r$ of a target is determined by the velocity component along the line-of-sight of the radar and is given by

$$v_r = v_a \cos\theta, \tag{2.8}$$

where $v_a$ is the actual velocity of a target and $\theta$ is the angle between the target trajectory and the line-of-sight, as depicted in Fig. 2.2.

It can be observed from (2.8) that for an acute angle $\theta < 90°$, corresponding to an approaching target, the Doppler shift is positive $f_d > 0$ and for an obtuse angle $\theta > 90°$, corresponding to a receding target, the Doppler shift is negative $f_d < 0$. Furthermore, for $\theta = 90°$ the Doppler shift is zero. Thus, the velocity component perpendicular to the line-of-sight cannot be determined.

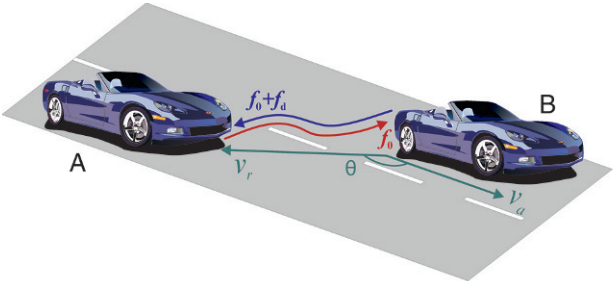


**Fig. 2.2** Doppler effect.

### *2.3.2 Frequency-Modulated Radar*

A simple CW radar allows determination of target velocity, but not the distance to target. Therefore, frequency-modulated continuous-wave (FMCW) systems have been developed to resolve this drawback. There are several possibilities to modulate a carrier frequency in time, such as linear frequency modulation (LFM), frequency shift-keying (FSK) or frequency-stepped continuous-wave (FSCW) modulation. This section describes the commonly implemented LFM and FSCW modulation schemes.

#### 2.3.2.1 Linear FM Continuous-Wave Radar

The most common modulation technique is LFM that modulates the transmit frequency with a triangular waveform [5]. The principle is exemplified in Fig. 2.3 showing the waveforms for a target at a distance $R$, approaching with a relative velocity $v_r$.

The transmit signal varies between the minimum frequency $f_0$ and the maximum frequency $f_0 + B$ with a period $T_m$, where $B$ is the bandwidth. At time $t_1$, the transmitter sends a signal with frequency $f_1$. This signal is received at time $t_2$, after a round-trip delay of $\tau = 2R/c$, with a frequency shifted by $f_d$. Meanwhile, the transmitter frequency is $f_2$. A mixer produces a base band signal at the instantaneous difference frequency between the transmit $f_{\mathrm{TX}}$ and the receive $f_{\mathrm{RX}}$ signals, referred to as the beat signal $f_b = |f_{\mathrm{TX}} - f_{\mathrm{RX}}|$. When the target is stationary, the beat frequency level is only related to the range and is given by

$$f_R = f_{\mathrm{b,stationary}} = \frac{4B}{T_m} \cdot \frac{R}{c}. \tag{2.9}$$

For a moving target the Doppler effect shifts the absolute values of the received frequencies. Thus, the down-converted frequencies are

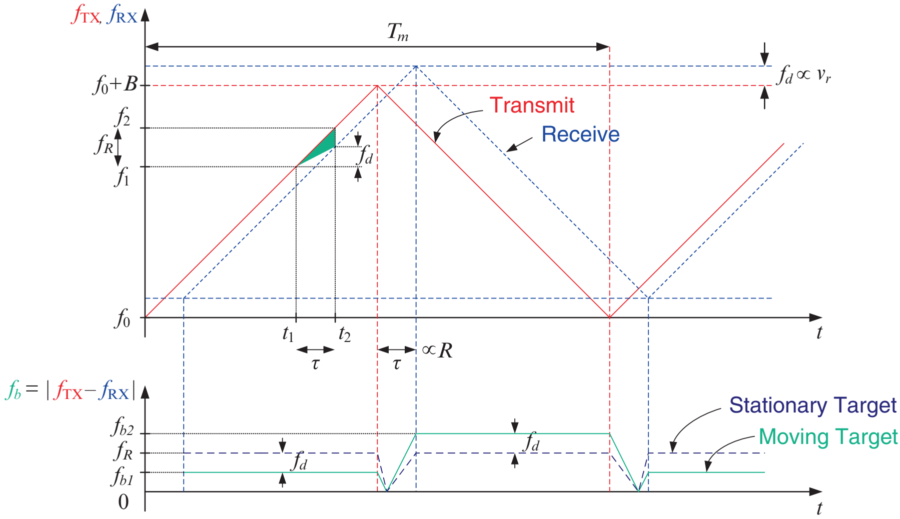


**Fig. 2.3** Range and relative velocity detection.

$$f_{b1} = f_R - f_d, \tag{2.10}$$

$$f_{b2} = f_R + f_d, \tag{2.11}$$

for the rising and for the falling slope of the transmit signal, respectively. Once the beat frequency levels $f_{b1}, f_{b2}$ have been measured in the baseband, the range and the relative speed can be calculated by [2]

$$R = \frac{c\,(f_{b1} + f_{b2})\,T_m}{8B}, \tag{2.12}$$

$$v_r = \frac{c\,(f_{b2} - f_{b1})}{4 f_0}, \tag{2.13}$$

where $f_0$ is the frequency of the transmitted signal.

Since the beat frequency signal is a rectangular signal with a period $T_m/2$, the corresponding spectrum is a sinc function centered at $f_b$ and the first zero crossing occurs at $2/T_m$. Thus, the smallest resolvable frequency $\Delta f$ is the reciprocal of the measurement time

$$\Delta f = \frac{2}{T_m}. \tag{2.14}$$

Substituting (2.14) into (2.7) the minimal resolvable velocity is obtained

$$\Delta v_r = \frac{c}{2 f_0} \cdot \Delta f = \frac{c}{f_0} \frac{1}{T_m}. \tag{2.15}$$

Thus, for larger $T_m$ or lower modulation frequency, higher velocity resolution can be achieved. Additionally, substituting (2.14) into (2.9) one obtains expression for the range resolution

$$\Delta R = \frac{cT_m}{4B} \cdot \Delta f = \frac{c}{2B}. \quad (2.16)$$

As can be observed, larger bandwidth offers higher range resolution. Thus, the range resolution of a narrow-band radar is limited.

Further insight on the FMCW radar using linear frequency modulation is presented in Appendix A. Additionally, a more advanced modulation algorithm FSCW that combines LFM and FSK techniques is briefly described in Appendix B.

## 2.4 Angle Detection

The monopulse principle can be described on two antennas having complex receive patterns $G_1(\alpha)$ and $G_2(\alpha)$. The distance between the antennas is $d$, as shown in Fig. 2.4. The phase difference $\Delta\varphi$ for an incident plane wave is

$$\Delta\varphi = d\sin(\alpha)\frac{2\pi}{\lambda}, \quad (2.17)$$

where $\lambda = c/f_0$ is the wavelength, related to the carrier frequency $f_0$.

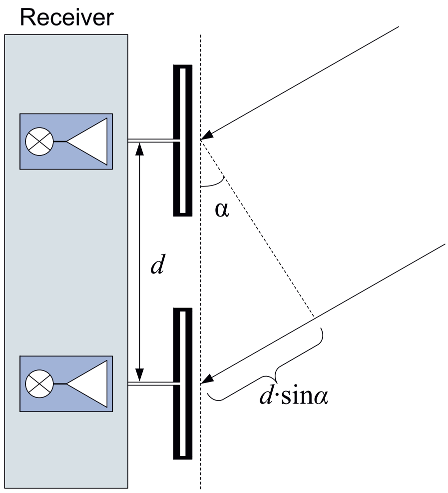


**Fig. 2.4** Antenna array in receive mode.

The difference and sum of the received signals of both antennas are given by

$$\Delta(\alpha) = G_1(\alpha) - e^{-i\Delta\varphi} \cdot G_2(\alpha), \tag{2.18}$$

$$\Sigma(\alpha) = G_1(\alpha) + e^{-i\Delta\varphi} \cdot G_2(\alpha). \tag{2.19}$$

For the monopulse angle detection the ratio $R_{\text{mono}} = \Delta/\Sigma$ is considered.

If the detection is based only on the amplitude of $R_{\text{mono}}$, it is called *amplitude-comparison monopulse*. This approach uses two overlapping antenna beams, so that the radiation patterns have slightly different look directions. This technique is preferred for long-range radars (LRR) with a small coverage angle, as e.g. for the 24 GHz LRR in [6].

If the antenna patterns $G_1$ and $G_2$ are identical, only the phase difference can be used for angle detection. This is referred to as the *phase-comparison monopulse* and or as the *phase interferometry*. In this case the ratio $R_{\text{mono}}$ becomes

$$R_{\text{mono}} = \frac{\Delta}{\Sigma} = \frac{1 - e^{i\Delta\varphi}}{1 + e^{i\Delta\varphi}}. \tag{2.20}$$

In the phase monopulse technique the phase difference $\Delta\varphi$ is evaluated in order to avoid the amplitude calibration, required in the amplitude monopulse technique. The angle of arrival is then easily obtained by rearranging equation (2.17)

$$\alpha = \sin^{-1}\left(\frac{\lambda\Delta\varphi}{2\pi d}\right). \tag{2.21}$$

The phase monopulse technique is preferred for 24 GHz systems, because the antennas are implemented as patch antennas orientated in the same look direction, as e.g. in [7].

The unambiguous angular range depends on the distance $d$ between two receive elements

$$\Delta\alpha = 2 \cdot \sin^{-1}\left(\frac{\lambda}{2 \cdot d}\right). \tag{2.22}$$

For short-range applications the unambiguous angular range should be very close to $+/-90°$. Therefore, the spatial sampling theorem has to be fulfilled and the antenna separation has to be half-wavelength. For mid-range and long-range systems the spacing has to be chosen according to the beamwidth of the transmitter antenna. An increased spacing between the receiver antennas allows to increase the size of the antennas and thus the gain. Furthermore, this results in a direct improvement of the angle measurement accuracy. However, this results in an increased radar module size.

## 2.5 Frequency Regulations

The performance of radar systems and the applied waveform principles are strongly influenced by the frequency regulations. The maximum allowable power limits and the corresponding measurement procedures for 24 GHz radar systems are defined in the ETSI standard EN 302 288-1 [8]. This document defines the spectral mask of the maximum allowed transmitter power in the ISM and UWB frequency bands around 24 GHz, as shown in Fig. 2.5.

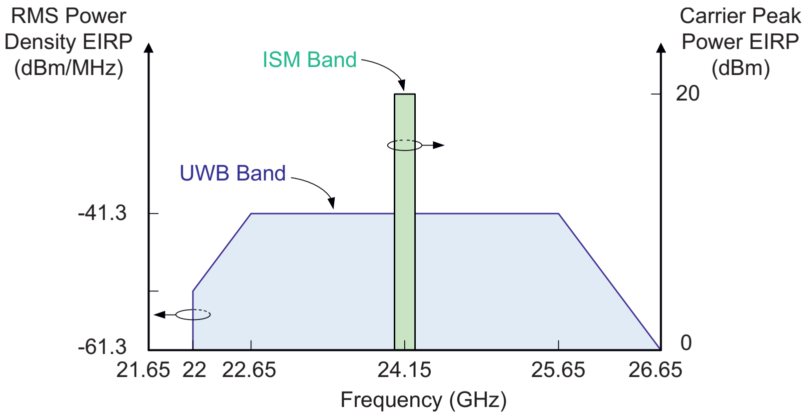


**Fig. 2.5** Transmitter maximum radiated power spectral mask.

The limit for the transmitted power is given as equivalent isotropic radiated power (EIRP). The EIRP value is given in dBm by adding the gain of the transmitter antenna to the actual transmitter power

$$P_{\mathrm{EIRP}}\,(\mathrm{dBm}) = P_{\mathrm{TX}}\,(\mathrm{dBm}) + G_{\mathrm{TX}}\,(\mathrm{dB}). \tag{2.23}$$

In the ISM band from 24.05 GHz to 24.25 GHz the maximum power is constrained to 20 dBm. For the ultra-wide band from 22.65 GHz to 25.65 GHz a maximum power spectral density of only $-41.3\,\mathrm{dBm/MHz}$ is allowed. This spectral density is very low and can only be used by pulsed systems with high bandwidth.

The ISM band of 200 MHz is applicable for automotive mid-range to long-range applications, since these are typically implemented as continuous wave (CW) systems and require higher power to achieve the necessary maximum range. CW systems provide a higher SNR compared to pulsed systems for the same transmitter power. Furthermore, the narrow available bandwidth of the ISM band is practical, since the maximum bandwidth is limited by the required signal to noise ratio at the maximum range [2]. Apart from the automotive radar systems, the ISM band is used for less demanding applications, such as door openers or surveillance. These systems are implemented using Doppler sensors without any additional frequency modulation.

## 2.6 Receiver Architectures

A receiver is used to amplify and down-convert a radio frequency (RF) signal with minimal added distortion. Therefore, the requirements for the receiver performance are usually very demanding. It should offer low noise, high dynamic range, and high local oscillator (LO) isolation so as to avoid radiation emission. The choice of receiver architecture is usually determined by complexity, power dissipation and system considerations. Receiver architectures can be classified with respect to the down-conversion topology.

### *2.6.1 Homodyne*

A homodyne or direct down-conversion receiver translates an RF signal directly to zero-IF. The frequency of the local oscillator (LO) is equal to the carrier frequency of the received RF signal. This architecture offers the advantage of simplicity. Avoiding an additional down-conversion to an intermediate frequency (IF) saves chip area, current consumption, complexity and avoids the image rejection problem inherent to heterodyne systems. A simplistic block diagram of a continuous-wave radar system implementing a homodyne receiver is presented in Fig. 2.6. The spectrum diagram of a homodyne receiver is depicted in Fig. 2.7. Both upper and lower side bands are down-converted to zero-IF.

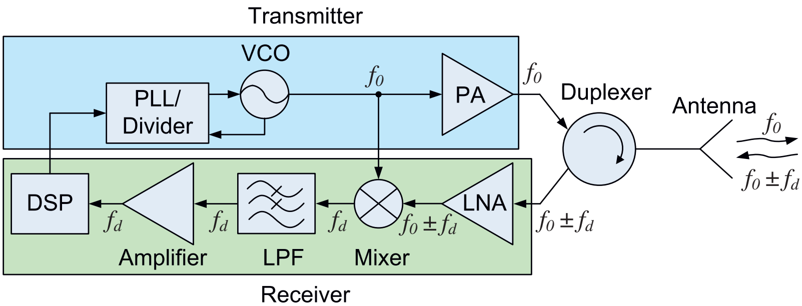


**Fig. 2.6** Block diagram of a homodyne CW radar system.

However, it has a serious disadvantage when implemented in CMOS technology. The expected baseband frequencies for a 24 GHz FMCW radar are in the range from 1 kHz to 100 kHz, whilst the flicker noise corner frequency of the CMOS circuits is around 10 MHz. Thus, for an active mixer implementation very high noise figures of above 40 dB at 1 kHz can occur. Therefore, either implementation of advanced circuit techniques [9] or of passive mixers [10] is required to resolve this issue. This problem will be addressed in detail in chapter 6. Bipolar transistors have a

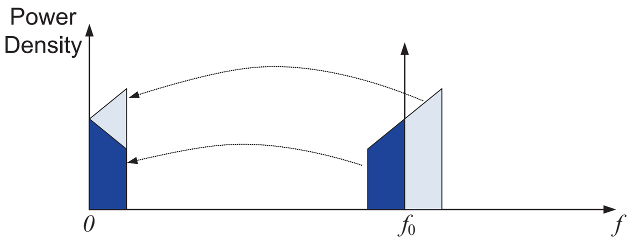


**Fig. 2.7** Frequency translation of a homodyne receiver.

much lower flicker noise corner frequency and therefore may be suitable for zero-IF receivers. However, there is an additional disadvantage. Parasitic DC signals appear due to mismatch, LO self-mixing and RF crosstalk [11]. The DC offset can be suppressed by DC blocking capacitors. The high-pass characteristics offered by these capacitors also act as a sensitivity time control (STC), which suppresses low-frequencies generated by nearby targets.

### *2.6.2 Heterodyne*

A heterodyne receiver down-converts an RF signal to an intermediate frequency, which is typically in the range of few gigahertz. Implementation of the IF frequency mitigates the flicker noise problem and allows for better selectivity due to an easier bandpass filter realization at the IF. However, it requires more circuit blocks and an additional IF reference frequency. A conceptual simplified block diagram of a CW radar using heterodyne architecture is presented in Fig. 2.8. The spectrum diagram of a heterodyne receiver is depicted in Fig. 2.9.

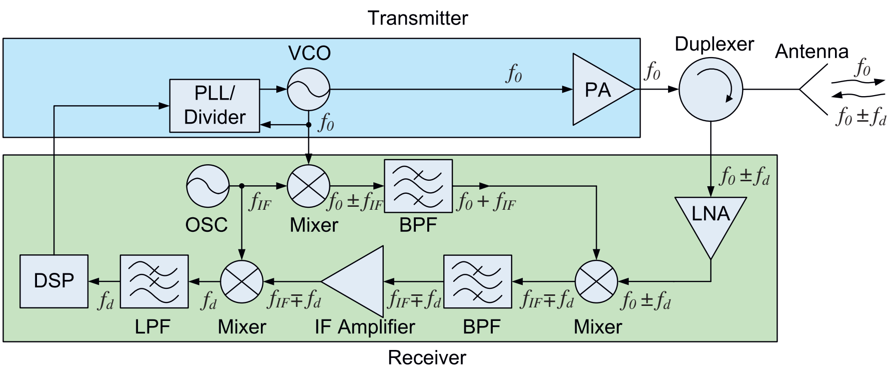


**Fig. 2.8** Block diagram of a heterodyne CW radar system [2].

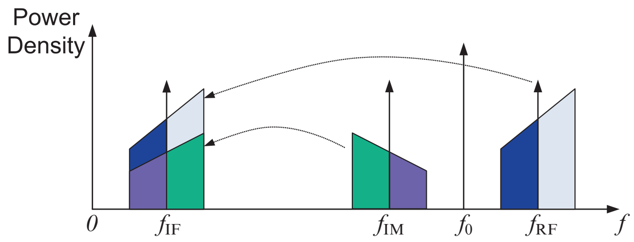


**Fig. 2.9** Frequency translation of a heterodyne receiver.

As can be seen, both the wanted and the image bands are down-converted to the IF frequency. Thus, the image frequency has to be suppressed before it is mixed down to the IF. This requires a bandpass filter. For a practical filter quality factor, the IF frequency should be sufficiently high, so that the RF is far from the LO frequency.

If the IF is low, image suppression at the RF becomes impossible, but signals could be processed directly at low frequencies of few megahertz and the image suppression performed at the IF [11]. For automotive radar systems in SiGe a direct down-conversion is the preferred option due to lower power consumption and complexity, whilst for CMOS a low-IF solution is preferred.

## 2.7 Status of Automotive Radar Systems

There are various ways of grouping the commercially available radar systems for automotive applications, e.g. by their bandwidth (narrow-band or wide-band), by their operation principle (pulsed or continuous wave) or by the covered area (short-range, mid-range and long-range radar). This section presents a brief overview of the current automotive radar implementations and classifies them with respect to their operating range and typical applications.

Short-range radar (SRR) systems are typically operated in a pulsed mode, have a maximum range of up to 30 m and a wide horizontal angular coverage of about $\pm 65°$ to $\pm 80°$ [12], [13]. Usually, several SRR sensors are equipped to fully cover the nearest surroundings of the vehicle. Pulsed SRR systems require a wide bandwidth of about $3-5$ GHz and are realized at the temporarily allocated UWB band around 24 GHz for cost reasons. The primary targeted safety features are blind-spot surveillance, parking aid and ACC support [12].

Mid-range radar (MRR) systems have a maximum range of 70 m and an angular coverage of $\pm 40°$ to $\pm 50°$ [7], [14]. These systems use the 200 MHz narrow ISM band around 24 GHz and operate in continuous wave mode using linear frequency modulation (LFM) or advanced modulation techniques such as e.g. frequency shift keying (FSK) or frequency-stepped continuous wave (FSCW) [15]. Due to the low available bandwidth the range resolution is limited to 0.6 m. Therefore, the primary targeted application for these sensors is the lane-change assistant.

Long-range radar (LRR) systems have a maximum range of up to 200 m and an angular coverage of $\pm 4°$ to $\pm 8°$ [16]. Most of the commercially available systems use the allocated $76 - 77$ GHz frequency range and operate in continuous wave mode using FMCW. However, there are also systems available that offer similar functionality using a narrow-band 24 GHz radar [14]. The long-range sensors are implemented typically for ACC.

## 2.8 Technology Requirements for Radar Chipset

Based on the radar system considerations presented in section 2.1, the integrated circuits for a radar front-end should fulfill demanding performance requirements around the frequency of 24 GHz. This poses a challenge on circuit design, which can be relaxed by implementation of high performance technologies, based on III-V semiconductor compounds such as e.g. gallium-arsenide (GaAs) or indium-phosphide (InP) [17]. These technologies can provide transistors with very high output power, very low noise, high gain and good linearity. However, they have a low integration level, which results in increased bill of materials (BOM) and module assembly costs. Implementation of circuits in cheaper silicon-based technologies offers the advantage of high integration, but at the cost of lower performance. Thus, in order to achieve sufficient performance, implementation of advanced circuit techniques is required.

Until now the commercial radar sensors have used front-end chip sets realized mainly in the III-V semiconductor technologies [18], [19]. The implementations usually comprise many discrete components. Therefore, the new generation of radar sensors uses integrated circuits based on SiGe technology [20]. Future generations of radar circuit implementation would further take advantage of high integration capabilities of CMOS. Transceiver front-ends for automotive applications have been already demonstrated at 24 GHz and at 77 GHz in BiCMOS technology [21].

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