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# Design of Radio-Frequency Transceivers for Wireless Sensor Networks

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## 1. Introduction

The SoC (System-on-Chip) design for the WSN (Wireless Sensor Networks) nodes is the most significant technology of modern WSN design. There are a large amount of nodes in a WSN system, the nodes are densely deployed either inside the environment or very close to it. Each node is equipped with a sensor, an ADC (Analog-to-Digital Converter), a MCU (Micro Controller Unit), a storage unit, a power management unit, and a RF (Radio-Frequency) transceiver, as shown in Fig. 1, so that it can sense, store, process, and communicate with other sensors using multi-hop packet transmissions. The basic specifications of WSN are reliability, accuracy, flexibility, expenses, the difficulty of development and power consumption. Because all the nodes are battery-powered, power consumption is the most important specification of WSN.

The core part of a WSN node is the RF transceiver, which is used to realize the wireless communication among the nodes. For a common used commercial chip, the distribution of power consumption is shown in Fig. 2, where *TX* and *RX* represent the transmitting mode and receiving mode of the transceiver. We can see that the RF part consumes the most power. Besides, modern RF design is composed of so many subjects that it requires IC designers to have sufficient knowledge. As a result, the IC design of RF transceivers becomes the most challenging research topic in the WSN field.

As the applications of WSN become more and more widespread, many companies have developed highly-integrated chips for RF transceivers. The main specifications of several commercial chips are shown in Table. 1. The common characters of these chips can be summarized into several aspects: 1) a low data rate, 2) low power consumption, 3) high sensitivity, 4) relatively low output power, and 5) a simple modulation scheme.

There are several ways to achieve low power consumption in WSN: 1) reduce the radiated power by using ad-hoc networks and multi-hop communication, 2) optimize the trade-off between communication and local computing, 3) design more power-efficient RF transceivers, and 4) develop more energy-efficient protocols and routing algorithms. And the third one is what we will talk about in this chapter.

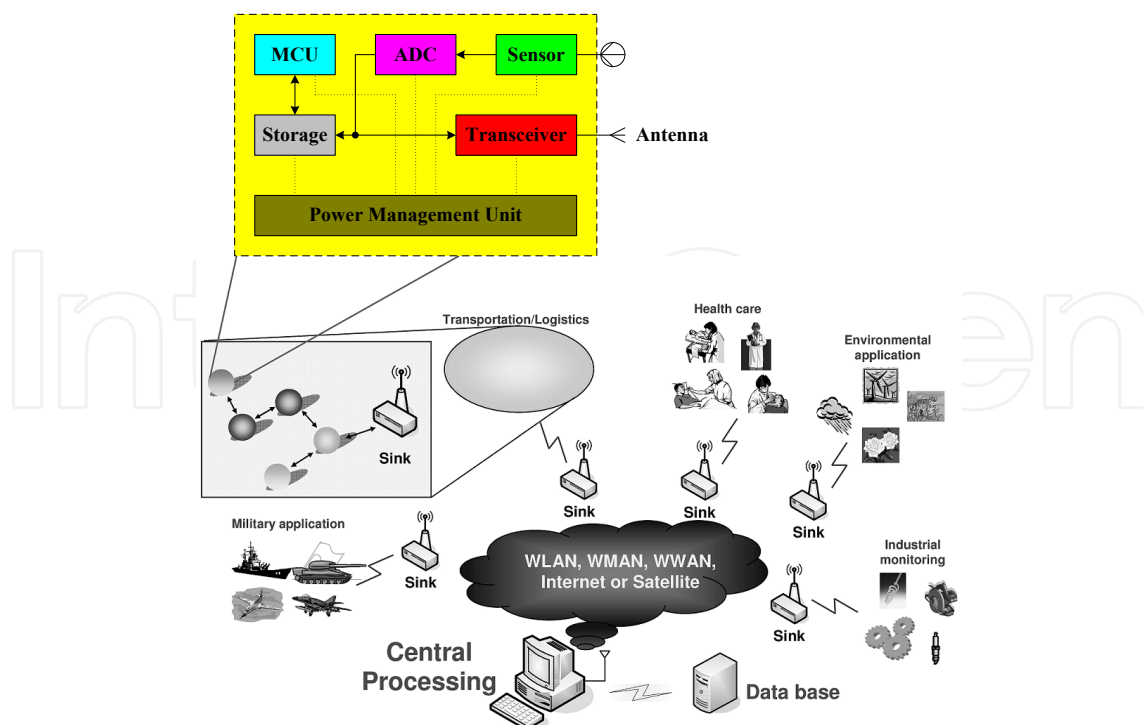


Fig. 1. The Wireless Sensor Networks.

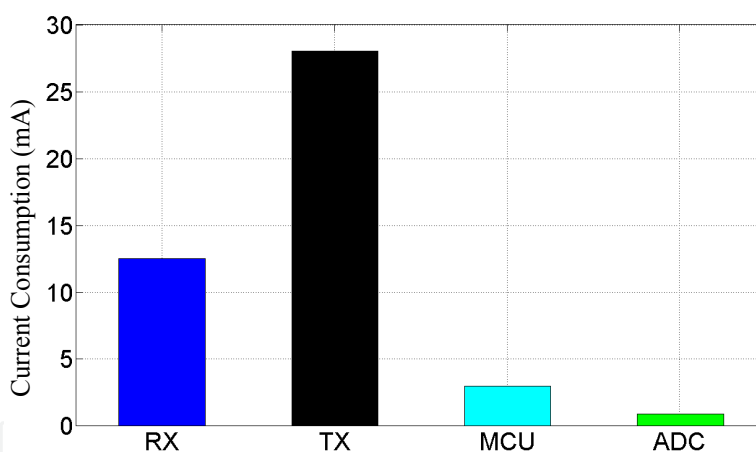


Fig. 2. The power distribution in a commercial WSN chip.

## 2. The classification of WSN transceivers

In this section, we make a classification of WSN transceivers, which are sorted by both modulation schemes and system architectures. Although some complex schemes such as OFDM (Orthogonal Frequency Division Multiplexing) can be adopted for the prospect of spectrum effective usage, RF designers are still inclined to simple schemes such as OOK (On-Off Keying), FSK (Frequency Shift Keying), UWB (Ultra-Wide Band), MSK (Minimum Shift Keying), BPSK (Binary Phase Shift Keying), QPSK (Quadrature Phase Shift Keying), and so on for low power consideration. From Table. 1, we can see that these simple modulation schemes are often adopted in the common-used commercial chips. In this section, we will analyze the advantages and disadvantages of the transceivers with these modulation schemes.

Chips	FB <sup>a</sup> (MHz)	DR <sup>b</sup> (kbps)	Power <sup>c</sup> (mW)	Sensitivity (dBm)	OTP <sup>d</sup> (dBm)	MS <sup>e</sup>
TR1000	916.5	115.2	14.4/36	-98	-1.2	OOK/ASK
TRF6903	300~1000	19.2	60/111	-103	-12~8	FSK/OOK
CC1000	300~1000	76.8	30/87.8	-107	-20~10	FSK
CC2420	2400	250	33.8/31.3	-95	0	O-QPSK
nRF905	433~915	100	37.5/90	-100	-10~10	GFSK
nRF2401	2400	0~1000	75/39	-80	-20~0	GFSK

<sup>a</sup>FB: The working Frequency Band of the chips.

<sup>b</sup>DR: The Data Rate of the transceivers.

<sup>c</sup>Power: The Power consumption for receiving/transmitting mode.

<sup>d</sup>OTP: The Output Transmitting Power.

<sup>e</sup>MS: The Modulation Scheme of the transceiver.

Table 1. The main specifications of several commercial RF chips common used in WSN.

2.1 The OOK transceiver

The OOK transceiver can often be realized by a simple architecture, since both the modulator and demodulator are easy to implement. As a result, the power consumption can be reduced. Another advantage is that a high data rate can be obtained by OOK. However, an AGC (Automatic Gain Control) with a wide dynamic range is often needed; a special coding is needed to avoid the saturation caused by long series of 0 or 1 in the receiver; it is spectrally inefficient; the most serious defect of OOK is that it is strongly susceptible to interferers, then the maximum communication distance of OOK transceivers is usually not long.

As a high data rate can be obtained by OOK transceivers, there are some works focused on multi-gigabit short-range wireless communications (Jri et al., 2010). The transceivers work at several GHz with low energy per bit. Besides, since the power consumption of OOK transceivers can be made very small, some recently works design OOK receivers for wake-up usages (Pletcher et al., 2009; Seungkee et al., 2010). The power consumption of these wake-up receivers can be as low as a few tens of  $\mu W$ .

One typical architecture of OOK transceivers is shown in Fig. 3. In the transmitter, the base-band data is modulated by a mixer with a carrier generated by an oscillator or a PLL (Phase-Locked Loop), and then the modulated signals are amplified by a PA (Power Amplifier) and emitted through an antenna. In the receiver, the signals received from an antenna is amplified by a LNA (Low-Noise Amplifier) firstly, and then detected by an envelop detector. At last, a simple DEM (DEModulator) can be used to demodulate the signals into 1-0 series. Therefore, in short-range and high-speed WSN applications, the OOK transceiver is a reasonable choice.

2.2 The FSK transceiver

As only the zero-crossing points of the signal contain useful information, FSK transceivers can work without an AGC, and special coding is not necessary. The modulator and demodulator are also easy to realize. As a result, the FSK transceiver has a simple architecture that ensures low power consumption and low cost. The spectrum efficient and ability to avoid interferences of FSK are both higher than that of OOK, so the communication distance of FSK transceivers can be longer. Besides, frequency hopping can be realized for FSK schemes. However, the complexity of FSK transceivers is relatively higher than that of OOK transceivers.

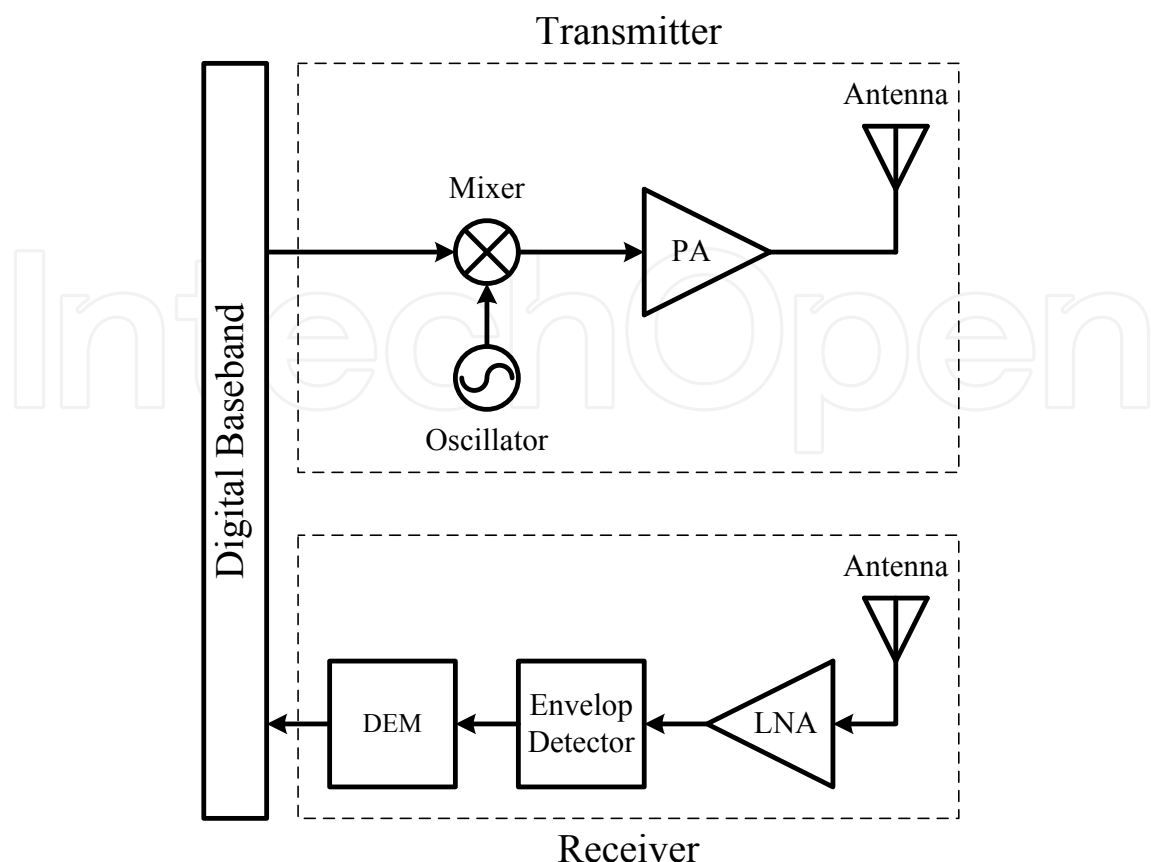


Fig. 3. A typical OOK transceiver.

A typical FSK transceiver is shown in Fig. 4. For the transmitter, the PLL directly digital modulation can be adopted (Perrott et al., 1997). This technology will be described in detail in section 4.4. The data stream from baseband is shaped firstly, and then input into a PLL to generate the FSK signals. Then the FSK signals are amplified by a PA and emitted by an antenna. In the receiver, the received signals are amplified by a LNA, and down-converted by a mixer that is followed by a filter to depress the interferences and the high-frequency component. The FSK demodulator is easy to realize since only the zero-cross points contain data information, and sometimes a zero-cross detector is enough. However, in many applications, the influence caused by frequency offset needs to be avoided. The power consumption of such transceiver is relatively low because the modulation and demodulation of FSK signals is easy to realize. As we analyzed above, OOK can realize a high data rate, but the communication distance is short; FSK transceivers can be used for long-distance communication, but the data rate is often not high. Therefore, a multi-mode transceiver in which OOK and FSK can be compatible with each other may be a good choice. The OOK modulation can be implemented by directly modulating the output power of the PA in Fig. 4.

### 2.3 The UWB transceiver

UWB is defined as a signal that occupies a bandwidth wider than 500 MHz or a fractional bandwidth larger than 20%. Although UWB is not a modulation scheme, we list the UWB transceivers individually here because the advantages of UWB are remarkable: 1) a high data rate, 2) low cost and low power consumption, and 3) high accuracy in distance measuring due to

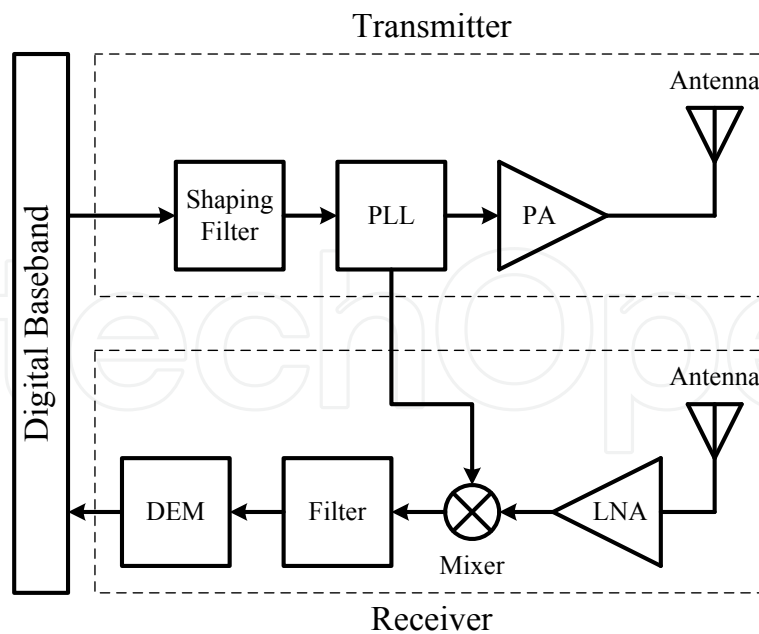


Fig. 4. An typical FSK transceiver.

the narrow pulse width of a few nanoseconds. However, the disadvantage is also serious: the spectrum is too wide and the emitting power is limited, so the single-hop transmission distance is very short (usually less than 10 m).

The IR (Impulse-Radio) UWB transceiver can be implemented with ultra-low power consumption. A typical example is shown in Fig. 5, where the architecture is similar to that of the typical OOK transceiver. The transmitter has a very simple structure, which is composed of an impulse generator and a PA. The impulse generator can be realized by a few logical gates. The receiver includes a LNA, a mixer as a squarer, an integrator and a comparator. In some applications, the architecture is even simpler than that of common OOK transceivers. Therefore, ultra-low power can be obtained with UWB transceivers, which fits the short-range WSN requirement very well.

## 2.4 Other Modulation Schemes

The spectrum efficiency of MSK is higher than FSK, but the demodulation is more complex and frequency hopping can not be realized for MSK. In 2003, IEEE released 802.15.4, which supports BPSK and O-QPSK (Offset-QPSK). The ability of BPSK to avoid interferences is better than that of FSK; and O-QPSK is more spectrally efficient. But the demodulation circuits of both BPSK and O-QPSK are complex and ADCs are usually needed. For general WSN applications that have a low data rate at several hundreds of Hz, the usage of these complex modulations is not recommended. However, in a few applications that have a high data rate and a strict spectrum limitation, MSK, BPSK, O-QPSK and even OFDM can be adopted.

## 2.5 The classification based on system architectures

In this section, we sort the transceivers according to the system architecture, and the types can be divided into superheterodyne, zero-IF (Intermediate Frequency), low-IF, slide-IF, super-regenerative, amplifier-sequenced hybrid architectures, and so on. These kinds of transceivers are summarized into Table 2. In WSN applications, low-IF transceivers are often adopted for high integration and low power consumption.

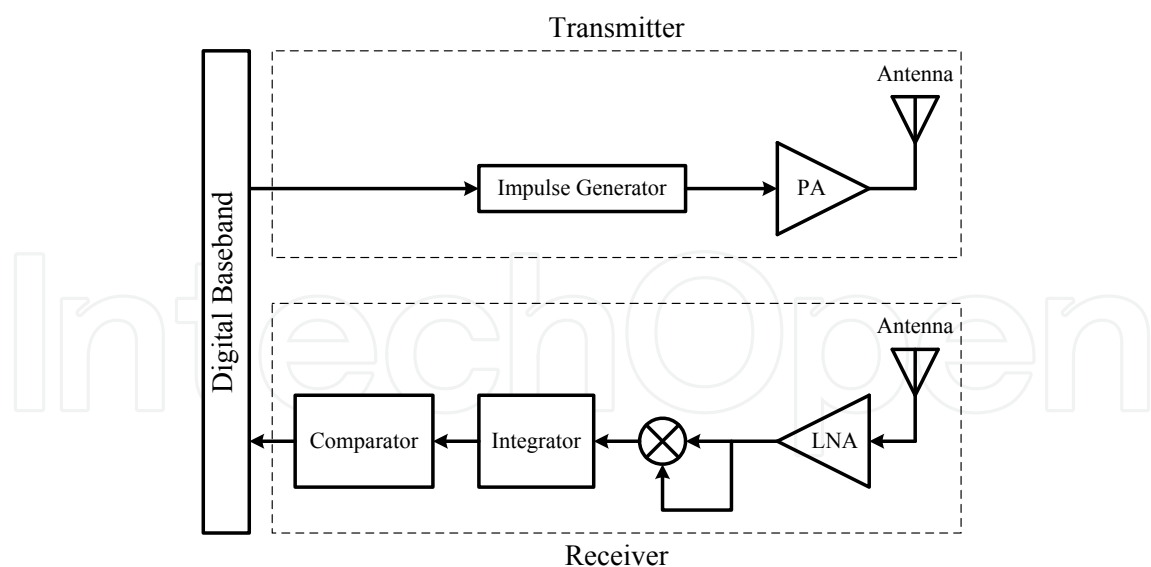


Fig. 5. A typical UWB transceiver.

A common transceiver architecture is shown in Fig. 6. In the transmitter, there is I/Q dual-path chains to realize the vector-signal transmitting. The coding, spreading, shaping, and so on are all finished in the digital domain. The data stream from digital baseband is converted to analog signals by two DACs (Digital-to-Analog Converters), and then the interferences and harmonics are filtered out by two filters. Two mixers are adopted for I/Q dual-path up-conversion and modulation. The phase difference of the two LO (Local Oscillation) signals input to the two mixers is  $90^\circ$ . At last, the I/Q signals are summed by an adder, amplified by a PA, and then emitted through an antenna. Assume the signal in the I path is  $a$ , the signal in the Q path is  $b$ , and then the transmission signal can be expressed as  $a + jb$ . Therefore, all the QAM (Quadrature Amplitude Modulation) signals in the constellation diagram can be generated by this transmitter. In the receiver, the received signals are amplified by a LNA firstly, and then down-converted by two mixers into I/Q dual-path signals. After being filtered by a filter, the analog signals are converted to digital domain by two ADCs. At last, the demodulating, de-spreading, synchronizing, decoding and so on are all done in digital domain. Such is the common architecture of SDR (Software Defined Radio), all the modulation schemes can be realized in this transceiver. Generally, the WSN usages have the characters of low cost, low power and a low data rate, so such complex architecture of SDR is not necessary in many WSN applications since simple analog circuits can be adequate. However, for specially WSN applications that need a complex modulation scheme, the SDR transceiver is indispensable.

### 3. The system design of transceivers

In this section, we will describe how to make system-level plans for the transceivers. The system-level specifications of a receiver are composed of sensitivity, maximum received signal, co-channel interferences, ACS (Adjacent/Alternate Channel Selectivity), image rejection, desensitization or blocking, and so on. And the specifications of a transmitter include ACPR (Adjacent Channel Power Ratio), EVM (Error Vector Magnitude), noise emission, and so on. These system-level specifications are decided by the protocols and network formation of WSN, and are affected by every submodule's performance, such as noise figure, gain, third intercept point, and so on. The main task for the system designers is to assign the system-



Architecture	Advantages	Disadvantages
Superheterodyne	Simple structure High sensitivity High selectivity	Poor integration level Large power consumption
Zero-IF <sup>a</sup>	Easy to be integrated Low power	DC offset Flicker noise
Low-IF	Easy to be integrated Immune to DC offset	Image rejection Difficult baseband design
Slide-IF	Small image interference Easy to be integrated	Difficult for LO design
super-regenerative	High sensitivity Low power	Low data rate Seriously affected by PVT <sup>b</sup> Poor stability
Amplifier-sequenced hybrid	Good stability Low power High Sensitivity	Low data rate

<sup>a</sup>IF: Intermediate Frequency  
<sup>b</sup>PVT: Process, Voltage, and Temperature

Table 2. The classification based on system architecture.

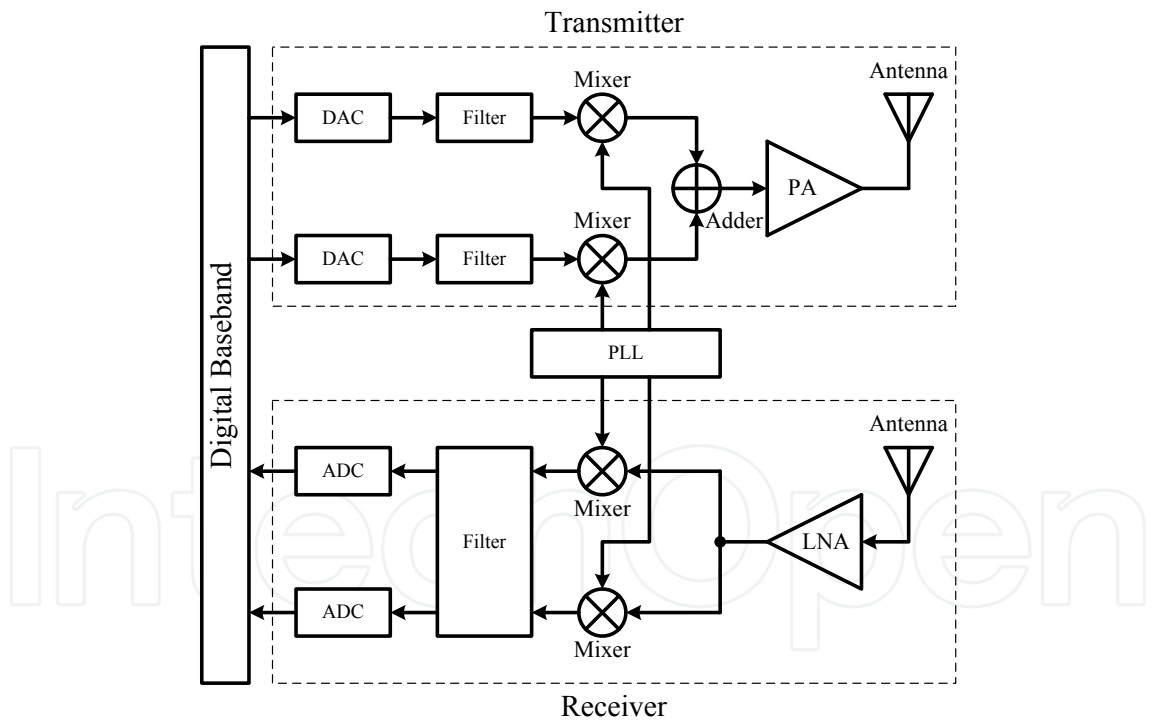


Fig. 6. A common architecture of transceivers.

level specifications to every individual submodules. Some of the conclusions in this section are referenced to Gu’s book (Gu, 2005).



For the AWGN (Additive White Gaussian Noise) channel, the noise at the input of a receiver chain can be equivalent to thermal noise, and the sensitivity of a receiver can be expressed as:

$$R_{min}(dBm) = -174 + 10\log(BW) + NF + CNR_{min} \quad (1)$$

where  $BW$  is the receiver noise bandwidth in Hz,  $NF$  is the overall noise figure of the receiver in dB, and  $CNR_{min}$  is the minimum CNR (Carrier-to-Noise Ratio) required for obtaining the required error rate.

As two different radio transmitters may use the same frequency in actual applications, RF receivers must have a certain ability to avoid the influence of co-channel interferences. Another important specification of receiver is the ACS, which measures the ability to receive a desired signal at the operating channel frequency in the presence of adjacent/alternate channel signals at a frequency offset from the assigned channel. The blocking characteristic is similar to the ACS, but it measures the anti-interference ability of an unwanted interferer at frequencies other than those of adjacent channels. The adjacent/alternate channel interference signal is usually modulated, while the blocking interferer is often defined as a continuous waveform tone. Both the adjacent/alternate channel selectivity and the blocking characteristics of a receiver are determined mainly by the characters of the channel filter, and the phase noise and spurs of the LOs in the adjacent/alternate channel bandwidth or around the unwanted interferer. For most wireless mobile systems, the desired signal level at receiver input for testing is defined as 3 dB above the reference sensitivity level  $R_{min}$  (Gu, 2005):

$$R_{d,i} = R_{min} + 3dB \quad (2)$$

For the single-conversion receiver, as shown in Fig. 6, it can be derived the ACS or the blocking characteristic  $ACS_{adj/alt/block}$  to have the following form:

$$ACS_{adj/alt/block} = I_{adj/alt/block} - R_{d,i} \quad (3)$$

$$I_{adj/alt/block} = 10\lg \left( \frac{10^{\frac{R_{d,i}-CNR}{10}} - 10^{\frac{-174+10\lg BW+NF}{10}}}{10^{\frac{PN+10\lg BW-\Delta R_{IF}}{10}} + 10^{\frac{SP-\Delta R_{IF}}{10}}} \right) \quad (4)$$

where  $CNR$  is the carrier-to-noise ratio for a given bit error rate,  $BW$  is the receiver noise bandwidth,  $NF$  is the noise figure of the RF front end,  $PN$  and  $SP$  are the phase noise and spurs of the LO signals, and  $\Delta R_{IF}$  is the relative rejection of the IF channel filter to the adjacent/alternate channel signal or the blocking interferer.

The maximum received signal is decided by both the 1-dB compression point and the intermodulation behavior of the receiver. For a RF receiver chain with multi-stages, the input signal strength needs to be smaller than the 1-dB compression point of the first stage, and the output signal strength of every previous stage must be smaller than the 1-dB composition point of the following stage, otherwise the receiver will be saturated. The maximum received signal is also limited by the intermodulation because the nonlinearity will become more serious under larger signal strength. The intermodulation characteristics reflect the linearity of the receiver, and it is limited by the noise figure, the receiver noise bandwidth, the phase noise and spurs of the LOs, and the cross-modulation. For the single-conversion receiver, as shown in Fig. 6, the expression of the allowed maximum degradation of the desired signal caused by noise and interferences at the receiver input is:

$$D_{max} = R_{d,i} - CNR_{min} \quad (5)$$

The minimum third intercept point of a receiver can be expressed as:

$$IIP3_{min} = \frac{1}{2} \left[ 3I_{min} - 10 \lg \left( 10^{\frac{D_{max}}{10}} - 10^{\frac{N_{nf}}{10}} - P_{pn} - P_{sp} - 10^{\frac{N_{cm}}{10}} \right) \right] \quad (6)$$

where  $I_{min}$  is the minimum intermodulation interference tone level at receiver input.  $P_{pn}$  and  $P_{sp}$  are the phase noise and spurs contributions of LO, and they can be expressed as:

$$P_{pn} = 10^{\frac{N_{pn} + 10 \lg BW + I_{in} - \Delta R}{10}} \quad (7)$$

$$P_{sp} = 10^{\frac{N_{sp} + I_{in} - \Delta R}{10}} \quad (8)$$

where  $N_{pn}$  is the average phase noise over the receiver bandwidth at a frequency offset equal to the space between the interference and the carrier,  $N_{sp}$  is the magnitude of the spurs at a frequency offset equal to or nearby the space between the interference and the carrier,  $I_{in}$  is the intermodulation interference tone, and  $\Delta R$  is the rejection to the interference tone of a filter.  $N_{nf}$  denotes the receiver inherent noise converted to its input port:

$$N_{nf} = -174 + NF + 10 \lg BW \quad (9)$$

$N_{cm}$  is the cross-modulation product, which can be approximately expressed as:

$$N_{cm} = I_{in} - 2IIP3_{lna} + 2(P_{Tx} + IL_{dRx} - R_{dTx}) + C \quad (10)$$

where  $IIP3_{lna}$  is the LNA input third intercept point,  $P_{Tx}$  is the transmitter output power at the antenna port in dBm,  $IL_{dRx}$  is the receiver side insertion loss of the duplexer in dB,  $R_{dTx}$  is the duplexer receiver side filter rejection to the transmission in dB, and  $C$  is a correction factor associated with waveform magnitude fluctuation and interference tone offset frequency, which is approximately -3.8 dB and -5.8 dB for the cellular and PCS band CDMA mobile station receivers, respectively (Gu, 2005).

Then another specification can be introduced, that is the dynamic range, which is generally defined as the ratio of the maximum input level to the minimum input level at which the circuit satisfied the required bit error rate. A common used specification called SFDR (Spurious-Free Dynamic Range) bases the definition of the upper end of the dynamic range on the intermodulation behavior and the lower end on the sensitivity. The SFDR can be expressed as:

$$SFDR = \frac{2(IIP3 - NF)}{3} - SNR_{min} \quad (11)$$

where  $IIP3$  is the input third intercept point of the RF front end,  $SNR_{min}$  is the required minimum SNR (Signal-to-Noise Ratio) of the IF circuits. The required  $IIP3$  can be calculated by equation (6), then the  $SFDR$  of a receiver can be obtained by equation (11).

Two strong in-band interference tones may cause a zero-IF or low-IF receiver to be completely jammed if the second-order intercept point of the receiver is not high enough. The two strong tones may generate in-channel interferences due to the second-order distortion of the receiver. The maximum level of the two equal blocking interference tones can be expressed as (Gu, 2005):

$$I_{block} = \frac{1}{2} \left[ 10 \lg \left( 10^{\frac{R_{di} - CNR_{min}}{10}} - 10^{\frac{N_{nf}}{10}} \right) + IIP2 \right] \quad (12)$$

where  $IIP2$  is the input second intercept point of the receiver.

The main problem of the low-IF receiver architecture is the image rejection since the IF is so low that it is difficult to separate the image from the desired signal by filters. The imbalance between I and Q channel signals in the low-IF receiver determines the possible maximum image rejection. The image rejection  $IR$  and the imbalances of the I/Q amplitude and the phase have the following relationship:

$$IR = 10lg \frac{1 + 2(1 + \Delta)\cos\gamma + (1 + \Delta)^2}{1 - 2(1 + \Delta)\cos\gamma + (1 + \Delta)^2} \quad (13)$$

where  $\gamma$  is the I and Q phase imbalance from the nominal  $90^\circ$  offset in degree,  $\Delta$  is the I and Q amplitude imbalance, which is usually expressed in dB by using the formula  $10lg(1 + \Delta)$ . For a full-duplex system, the receiver suffers from transmission leakage interferences particularly in the LNA. If a strong interference tone appears near the desired signal of the receiver, the amplitude modulation of the transmission leakage will cross-modulate the interference tone in the LNA. The spectrum of the cross-modulated tone may partially spread into the receiver channel bandwidth when the single-tone interferer is enough close to the desired signal. The receiver will be desensitized if the cross-modulation product getting into the receiver channel band is high enough. The expression of the allowed single-tone interferer is (Gu, 2005):

$$I_{st} = 10lg \left[ \frac{10^{\frac{D_{max}}{10}} - 10^{\frac{N_{nf}}{10}}}{10^{\frac{-[2(IIP3_{lna} - IL_{dRx}) - 2(P_{Tx} - R_{dTx}) - C]}{10}} + 10^{\frac{N_{pn} + 10lgBW}{10}} + 10^{\frac{N_{sp}}{10}}} \right] \quad (14)$$

where  $D_{max}$  can be got by equation (5),  $N_{nf}$  can be obtained by equation (9),  $P_{Tx}$  is the transmitter output power at the antenna in dBm, and  $C$  is the correction factor approximately equal to (Gu, 2005):

$$C = M_A + 6 + 10lg \frac{1.5 \times BW - \Delta f}{2 \times BW} \quad (15)$$

where  $\Delta f$  is the space between the interference tone and the carrier frequency of the desired receiving signal.  $M_A$  can be calculated by the probability density function  $p(x)$  of the transmission signals and the normalized low-frequency product of the second-order distortion:

$$M_A = 10lg \left\{ \int_{-\infty}^{\infty} \left[ \left( 10^{\frac{x}{10}} \right)^2 - 10^{\frac{IM2_{dc}}{10}} \right]^2 p(x) dx \right\} \quad (16)$$

$$IM2_{dc} = 10lg \left[ \int_{-\infty}^{\infty} \left( 10^{\frac{x}{10}} \right)^2 p(x) dx \right] \quad (17)$$

For the transmitter, the modulation accuracy is represented by EVM, which is defined as the mean square error between the samples of the actual and the ideal signals, normalized by the average power of the idea signal. The EVM of transmitters is influenced by the inter-symbol or inter-chip interference, the close-in phase noise of synthesized LO, the carrier leakage, the I and Q imbalance, the nonlinearity, the in-channel bandwidth noise, the reverse modulation of LO, and so on.

The influence caused by inter-symbol or inter-chip interferences can be obtained by:

$$EVM_{isi} = \sqrt{\sum_{k=-\infty}^{+\infty} \Delta I_{isi}^2(k)} \quad (18)$$

$$\Delta I_{isi}(k) = \left| \frac{h_{ir}(t_0 + kT_s)}{h_{ir}(t_0)} \right| = \frac{\int_{t_0 - \Delta t}^{t_0 + \Delta t} |h_{ir}(t + kT_s)| dt}{\int_{t_0 - \Delta t}^{t_0 + \Delta t} |h_{ir}(t)| dt} \quad (19)$$

where  $h_{ir}(t)$  is the impulse response of the pulse-shaping filter,  $k$  is equal to  $\pm 1, \pm 2, \pm 3, \dots$ ,  $T_s$  is the period of a symbol, and  $2\Delta t$  is the duration of a sampling pulse.

The influence caused by close-in phase noise of synthesized LO can be obtained by:

$$EVM_{pn} = \sqrt{2 \times 10^{\frac{N_{phase}}{10}} \times BW_{lf,pll}} \quad (20)$$

where  $N_{phase}$  is the average phase noise in dBc/Hz within the PLL loop bandwidth, and  $BW_{lf,pll}$  is the bandwidth of the PLL loop filter in Hz.

The carrier leakage is mainly caused by the DC offset of the baseband, the LO-to-RF leakage and the IF-to-RF leakage. The influence on EVM caused by carrier leakage can be obtained by:

$$EVM_{cl} = \sqrt{\frac{CL_{offset}}{10} + \frac{CL_{lo}}{10} + \frac{CL_{if}}{10}} \quad (21)$$

where  $CL_{offset}$ ,  $CL_{lo}$  and  $CL_{if}$  represents the leakage results from the DC offset, the LO-to-RF leakage and the IF-to-RF leakage, respectively.

The EVM caused by the I and Q imbalance can be expressed as:

$$EVM_{iq} = \sqrt{10^{\frac{IR}{10}}} \quad (22)$$

where  $IR$  is the image suppression, which can be calculated by equation (13).

It's assumed that only the signal amplitude equal to and greater than the output 1-dB compression of the power amplifier  $P_{-1}$  will affect the modulation accuracy, then the EVM caused by nonlinearity of the transmitter chain can be expressed as:

$$EVM_{nonlin} = \int_0^\infty P(\delta) \times \left(10^{\frac{\delta+1}{20}} - 1\right) d\delta \quad (23)$$

$$\delta = P_{Tx} - P_{-1} \quad (24)$$

where  $P_{Tx}$  is the output power level, and  $P_\delta$  is the amplitude probability density function of the signal.

The EVM caused by in-channel bandwidth noise can be expressed by:

$$EVM_{ibn} = \sqrt{10^{\frac{N_{ibn} - P_{Tx}}{10}}} \quad (25)$$

where  $N_{ibn}$  is the integrated noise over the channel bandwidth, and  $P_{Tx}$  is the transmission power in dBm.

The transmission signals may be reflected from the load of the modulator, then the reflected signals and their harmonics may modulate the LO if the frequency of the carrier or the harmonics of the reflected signals is equal to LO frequency. Then reverse modulation occurs. The EVM caused by reverse modulation is:

$$EVM_{rm} = \sqrt{10^{\frac{N_{rm}}{10}}} \quad (26)$$

where  $N_{rm}$  is the integrated reverse modulation noise of the synthesized LO over the transmission signal bandwidth below the LO level.

The overall EVM of the transmission signal can be expressed as:

$$\begin{aligned} EVM_{total} = & EVM_{isi}^2 + EVM_{pn}^2 + EVM_{cl}^2 + EVM_{iq}^2 \\ & + EVM_{nonlin}^2 + EVM_{ibn}^2 + EVM_{rm}^2 + \dots \end{aligned} \quad (27)$$

The ACPR specification is generally defined as ratio of the power integrated over an assigned bandwidth in the adjacent/alternate channel to the total desired transmission power. The ACPR can be expressed as:

$$ACPR = \frac{\int_{f_a}^{f_a + \Delta B} SPD(f) df}{\int_{f_o - BW/2}^{f_o + BW/2} SPD(f) df} \quad (28)$$

where  $f_a$  is the start frequency of the adjacent/alternate channel,  $\Delta B$  is the bandwidth of measuring adjacent/alternate channel power, which varies with different mobile systems. The general formula for the ACPR of a transmission signal at the output of the power amplifier with an output third intercept point  $OIP3$  can be expressed as:

$$ACPR \approx 2(P_{Tx} - OIP3) - 9 + C_0 + 10lg\left(\frac{\Delta B}{BW}\right) \quad (29)$$

$$C_0 \approx 0.85 \times (PAR - 3) \quad (30)$$

where  $P_{Tx}$  is the transmission signal power at the output of the power amplifier,  $PAR$  is the peak-to-average ratio of the random noise. In the mobile communication systems, the adjacent/alternate power may be tested in a bandwidth  $\Delta B$  that is different from the desired transmission signal bandwidth  $BW$ .

We only discuss the noise emissions that are those located outside of alternate channels here. In general, we like to have lower gain power amplifier for achieving low noise emission, but this is completely opposite to the gain setting of the power amplifier to obtain a good ACPR performance. The noise emission in mW/Hz of the transmitter has an expression as:

$$P_{nm} = G_{Tx} \times P_{n,in} + kT_0 G_{Tx} (NF_{Tx} - 1) \quad (31)$$

where  $P_{n,in}$  in mW/Hz is the noise at the transmitter input,  $G_{Tx}$  is the overall transmitter gain,  $NF_{Tx}$  is the overall noise factor of the transmitter, and  $kT_0 = 10^{-174/10}$  mW/Hz.

## 4. The design of key modules in the transceiver

A common used transceiver is composed of a LNA, Mixers, filters, IF circuits, a PA, ADCs and DACs, a PLL and so on. The individual performance and the matching among these modules determine the performance of the whole transceiver system. A general description of these modules will be given in this section.

### 4.1 Low-Noise Amplifier

The specifications of a LNA can be summarized into:

- The working frequency;
- The noise figure;
- The third intercept point;
- The voltage or power gain;

- The reflection coefficient at the input port and the isolation between the output port and the input port;
- The power consumption;

The most important specification of the LNA is the noise figure since the first stage of a receiver chain decides the noise performance of the whole system. One of the common used LNAs is the inductively source degenerated type, and a typical design example is shown in Fig. 7(a).  $M1$  is a common-source amplifier transistor,  $L_S$  is the source degenerated inductor,  $L_G$  is the gate inductor,  $V_{IN}$  is the input Port, and  $V_{OUT}$  is the output port;  $M2$  is a common-gate transistor that is used for isolation and gain enhancement; the load  $Z_L$  can be a resistor, a inductor or a inductor-capacitor tank. This structure has a large gain and a low noise figure, but the input reflection is a problem. There is a trade-off between the noise figure and the input impedance matching.

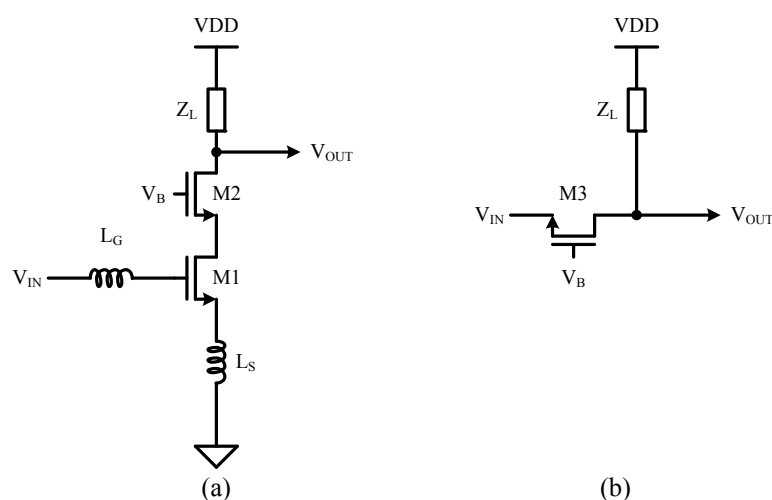


Fig. 7. (a)An inductively source degenerated LNA. (b)The common-gate input type.

Another common used LNA is the common-gate input type, as shown in Fig. 7(b). If the transconductance of the common-gate transistor  $M3$  is  $g_m$ , then the input resistance is equal to  $1/g_m$ . Therefore, the input matching of common-gate LNA is easier to realize compared to the source degenerated LNA. However, the noise performance is poor, since the common-gate amplifier has a low gain.

The WSN receivers usually have a high sensitivity, as shown in Table 1. According to equation (1), the sensitivity is proportional to the noise figure of the RF front end. Therefore, in WSN applications, we often adopt the inductively source degenerated LNA assisted by some low-noise technologies.

#### 4.2 Mixer

The mixers in transceivers can be divided into two types: 1)the up-converting mixers and 2)the down-converting mixers. The up-converting mixers are used in the transmitter, while the down-converting mixers are used for the receiver. The specifications of a mixer can be summarized into:

- The working frequency including RF frequency, LO frequency, and IF frequency;
- The noise figure;
- The third intercept point;



- The second intercept point (for zero-IF or low-IF receivers);
- The voltage or power conversion gain;
- The isolation between the RF port and the LO port, the RF port and the IF port, and the LO port and the IF port;
- The magnitude and phase imbalance between I and Q channel down converters (for the receivers and transmitters that use I and Q dual-path converters);
- The power consumption;

A classical mixer is known as the Gilbert cell, as shown in Fig. 8.  $I_B$  is a current source,  $RF+$  and  $RF-$  are the differential RF input ports,  $LO+$  and  $LO-$  are the differential LO input ports, and the output differential currents are  $I_{OUT+}$  and  $I_{OUT-}$ ;  $M1$  and  $M2$  convert the input RF voltage into current, and  $M3$ - $M6$  are used as switches for mixing.

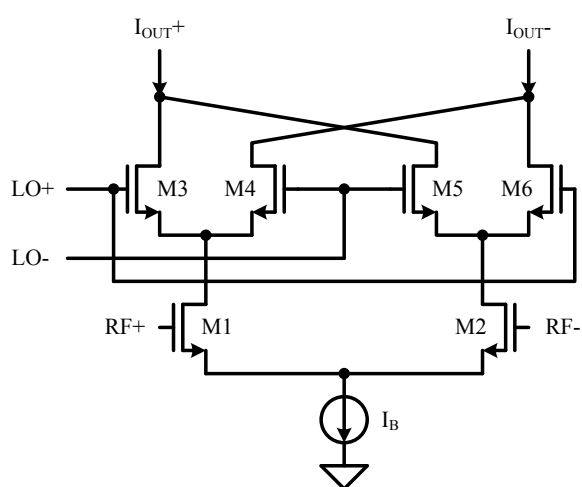


Fig. 8. The Gilbert cell.

The Gilbert mixer is a typical example of active mixers, which have high gain, low noise, but poor linearity. Another type of mixers are called passive mixers, which usually have low gain, large noise and high linearity. The passive mixers can be divided into two types: 1) voltage-mode mixers: the MOS switches are used for voltage switches, and loaded with high resistance. Because of the nonlinearity of the switches, distortions will be enlarged when the amplitudes of RF and IF signals are increased and the switches are modulated. 2) Current-mode mixers: the MOS switches are used for current switches, and loaded with low resistance. So the amplitudes of RF and IF signals are relatively low in current-mode mixers, then the linearity is improved. A typical example of current-mode passive mixers is shown in Fig. 9 (Valla et al., 2005). The input  $I_{IN}$  is a differential current signal, the output  $V_{OUT}$  is a voltage signal. Transistors  $M1$ - $M4$  are used for switches that are controlled by the LO signals  $LO+$  and  $LO-$ . An OTA (Operational Amplifier) together with resistors  $R1$ ,  $R2$  and capacitors  $C1$ ,  $C2$  is adopted to amplify and filter the signals. Two additional capacitors  $C3$  and  $C4$  are used to generate an extra pole to suppress the amplitude. The LNA and mixers are often designed and tested together to realize an optimized trade-off among gain, noise figure and linearity for different applications.

The distribution of WSN nodes is random, and the distance between two nodes may be very short or very long, so the dynamic range of the transceivers is very important. According to equation (11), in order to improve the SFDR performance, the noise figure needs to be



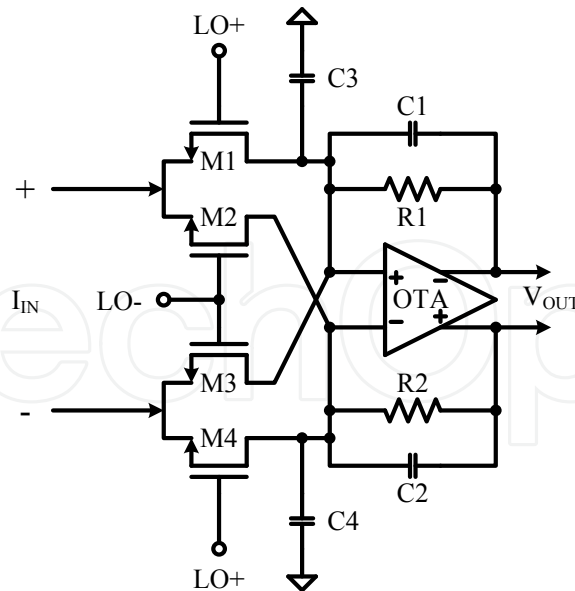


Fig. 9. A typical current-mode passive mixer.

decreased and the IIP3 needs to be enhanced. Therefore, a combination of a low-noise inductively source degenerated LNA and a current-mode passive mixer can be adopted for WSN usages.

#### 4.3 Active Filter

According to the pattern of implement, the active filters usually used for transceivers can be summarized into three types: 1)switched-capacitor filters, in which the resistors are replaced by switched capacitors; 2)active-RC filters, which is composed of OTAs and resistor-capacitor networks; 3)gm-C filters, in which the resistors and inductors are replaced by transconductors. For switched-capacitor filters, the advantages can be summarized here: 1)high precision without tuning, 2)small chip area and low power, and 3)insensitive to parasitics. However, there are several disadvantages: 1)affected by sampling, 2)requirement for extra clock generation circuit, and 3)not suitable for high-frequency applications.

For active-RC filters, the advantages can be summarized into: 1)high precision with tuning, 2)easy to design with classical RC structures, 3)insensitive to parasitics, 4)no sampling effect, and 5)large dynamic range. The disadvantages can be summarized into: 1)requirement for tuning circuits and 2)limited working frequency caused by OTAs.

For gm-C filters, the advantages can be summarized into: 1)high precision with tuning, 2)able to be realized based on simple open-loop OTAs, 3)lower power consumption than active-RC filters, 4)no sampling effect, and 5)good frequency performance. The disadvantages can be summarized into: 1)requirement for complex on-chip tuning circuits, 2)poor dynamic range, and 3)sensitive to parasitics.

According to the transfer characters, the filters can also be divided into four types: 1)Butterworth filters, which has the maximum flat amplitude in the pass band; 2)Chebyshev filters, which has the minimum ripples in the pass band; 3)Bessel filters, which has the maximum flat of group delay; 4)Ellipse filters, which has the minimum transition band. The other characters of these filters are summarized into Table 3.

Type	Amplitude-Frequency Characteristic			Phase-Frequency characteristic
	Pass Band	Stop Band	Transition Band	
Butterworth	Flat	Monotonic Decreasing	Gentle Monotonic Decreasing	Moderate
Chebyshev	Fluctuant	Monotonic Decreasing	Steep Monotonic Decreasing	Poor
Bessel	Flat	Monotonic Decreasing	Slowly Monotonic Decreasing	Excellent
Ellipse	Fluctuant	Fluctuant	Steep Monotonic Decreasing	Poor

Table 3. The characteristics of different filters.

In wireless transceivers, there is a special kind of filter named complex filter, which is usually used in low-IF receivers for image rejection. A classical complex filter is designed in 1995 (Crols & Steyaert, 1995). Fig. 10 shows the block diagram, which has I/Q dual-path inputs and I/Q dual-path outputs. The input of the Q path  $Q_{IN}$  is  $90^\circ$  delay of the input of the I path  $I_{IN}$ , and the output of Q path  $Q_{OUT}$  is also  $90^\circ$  delay of the output of the I path  $I_{OUT}$ . That is  $Q_{IN} = -jI_{IN}$  and  $Q_{OUT} = -jI_{OUT}$ . The transfer function of this complex filter can be

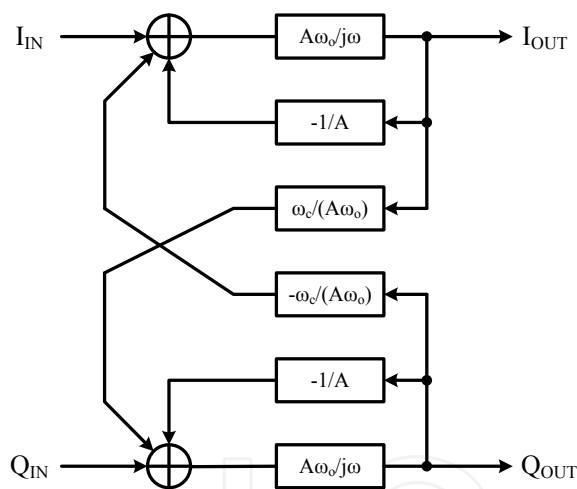


Fig. 10. The block diagram of a first-order complex filter.

expressed as:

$$H_{cf}(j\omega) = \frac{A}{1 + j(\omega - \omega_c)/\omega_o}$$

(32)

where  $\omega_c$  is the central frequency, and  $2\omega_o$  is the double-sideband bandwidth. It is equivalent to a low-pass filter’s pass band moved by  $\omega_c$ , and then the transfer curves of positive and negative frequency become asymmetric. As a result, the image can be rejected. As shown in Table 2, low-IF transceivers has the advantages of easy to be integrated and immune to DC offset, so the low-IF SDR transceiver is adopted in many WSN applications. Besides, the data rate of WSN is usually not very high, then the IF can be relatively low and gm-C filters are not necessary. Therefore, an active-RC complex filter is suitable for such WSN receivers because of their low power, large dynamic range and image rejection function.

#### 4.4 Phase-Locked Loop

The PLL is the core part of a transceiver system, as it's used for both the down-converting in receiver and the up-converting in transmitter. A typical sigma-delta charge-pump PLL is shown in Fig. 11 (Zhao et al., 2009), which is composed of a PFD (Phase-Frequency Detector), a charge pump, a loop filter, a VCO (Voltage-Controlled Oscillator), a multi-modulus frequency divider, and a sigma-delta modulator. Although the all-digital PLL has appeared in recent years, the classical charge-pump PLL is still widely used in the industrial community.

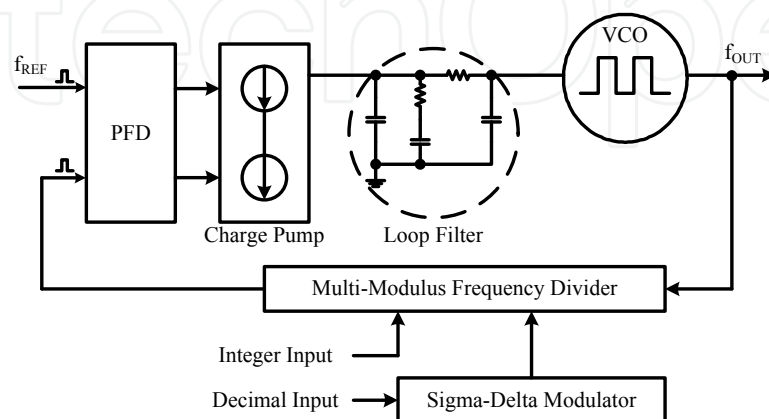


Fig. 11. A typical sigma-delta charge-pump PLL (Zhao et al., 2009).

For WSN transceivers, we tend to design low-power, full-integrated and fast-settling PLL.

The power of the PLL is mainly limited by these modules: 1)VCO (Voltage-Controlled Oscillator), 2)prescaler, and 3)the buffer connected at the output of VCO. Therefore, the power reduction of these modules is significant to the low-power design of PLL.

For full-integrated design, the chip area needs to be decreased. In a typical PLL, the LF (Loop Filter) and the inductors in LC-tank VCO take up the largest chip area. In order to decrease the area of LF, some one has proposed a discrete-time architecture (Zhang et al., 2003). For the applications in low-frequency bands, the inductors will be large if the resonant frequency of the VCO is low, so the VCO is required to be designed at a high frequency with a frequency divider connected after it.

The settling speed of PLL is decided by the loop bandwidth. Too Large bandwidth brings not only fast settling, but also large in-band noise and spurs. As a result, there is a trade-off between the settling speed, and the phase noise and the spur performance. How to set the trade-off depends on the requirement of the transceiver system.

As we referenced in section 2.2, the PLL can be adopted for directly digital modulation, and such method is proposed by Perrott in 1997 (Perrott et al., 1997). The architecture of such PLL based transmitter is shown in Fig. 12. The data stream can be shaped by a filter firstly, then the shaped data are input into the sigma-delta modulator in order to change the dividing ratio. As a result, the output frequency can be modulated by the variation of dividing ratio according to the input data, and the FSK signals can be generated in this way. A PA is connected at the output of the PLL so that the FSK signals can be emitted through an antenna. Generally, the data rate can not be larger than the bandwidth of the PLL. Although some one has proposed a compensation technology with a digital filter whose transfer function is the reciprocal of PLL's (Perrott et al., 1997), mismatch and inaccuracy depress the performance in actual designs. As a result, we would rather enlarge the bandwidth of PLL to obtain a relatively high data rate.

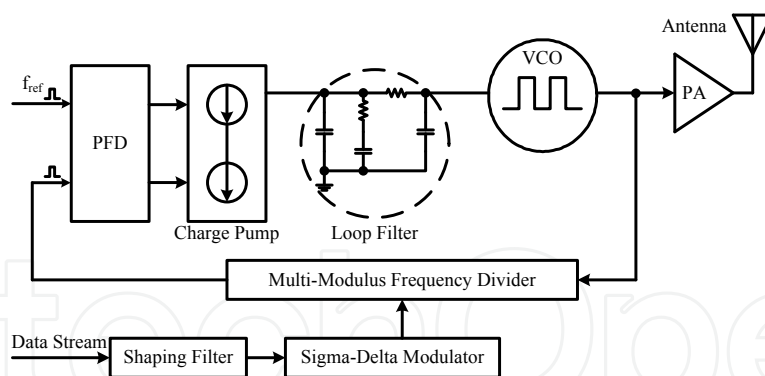


Fig. 12. A transmitter based on PLL directly digital modulation.

For WSN usages, the data rate is often not high, so such PLL directly digital modulation is a reasonable choice for common frequency and phase modulation schemes, such as FSK, MSK, and so on. As a result, mixers can be moved away, the cost and power consumption of the WSN transmitter can be saved a lot.

#### 4.5 Power Amplifier

Generally, PA is the most power hungry module of a transceiver. Therefore, the output power of PA is usually relatively small for WSN usages, as shown in Table 1. The types of PA can be divided into class A, B, C, AB, D, E, F and  $F^{-1}$ .

For class-A PAs, the amplifier MOSFET is kept in the saturation region. The transistor always dissipates power because the product of drain current and drain voltage is always positive. It should be noticed that the maximum theoretical drain efficiency of class-A PAs is just 50%. However, drain efficiencies of 30~50% are common for practical class-A PA designs. The normalized power output capability is about 1/8. The class-A amplifier provides high linearity at the cost of low efficiency and relatively large device stresses.

In a class-B PA, the device is shut off in half of every cycle. It should be mentioned that most practical class-B PAs are push-pull configurations of two MOSFETs. The peak drain current and maximum output voltage are the same as for the class-A PAs. The maximum drain efficiency for a class-B PA is 78.5%. The normalized power capability of the class-B PAs is 1/8, the same as for class-A PAs, since the output power, maximum drain voltage, and maximum drain current are the same.

In a class-C PA, the transistor conducts less than half the time. As the conduction angle shrinks toward zero, the efficiency approaches 100%, but the gain and output power unfortunately also tend toward zero at the same time. Furthermore, the normalized power capability of class-C PAs approaches zero as the conduction angle approaches zero. In one word, the efficiency can be large, but at the cost of normalized power capability, gain, and linearity.

The class-AB PAs conducts between 50% and 100% of a cycle. Both the conduction angle and efficiency of class-AB PA vary between that of class-A PA and class-B PA.

In a class-D PA, only one transistor is driven on at a given time, and one transistor handles the positive half-cycles and the other handles the negative half-cycles, just as a push-pull class-B PA. The difference between class-D and class-B is that the transistors are driven hard enough to make them act like switches for class-D PA, rather than as linear amplifiers. The normalized power capability of class-D PAs is about 0.32, which is better than a class-B push-pull and much better than a class-A PA. The MOS switches in class-D PAs function well only

at frequencies substantially below  $f_T$ , which is the cut-off frequency. Usually, one transistor fails to turn completely off before the other turns on, then the efficiency is deteriorated.

The class-E PA uses a high-order reactive network that provides enough space to shape the switch voltage to have both zero value and zero slope at switch turn-on, then the switch loss is reduced. The efficiency can approach theoretically 100% with ideal switches. The normalized power capability is about 0.098, which is worse than class-A PA. The class-E PA is more demanding of its switch performance than even class-A PAs because of the poor power capability and the reduced efficiency due to switch turn-off losses.

The termination of a class-F PA appears as an open circuit at odd harmonics of the carrier beyond the fundamental and as a short circuit at even harmonics, while the class-F<sup>-1</sup> employs a termination that appears as an open circuit at even harmonics and as a short circuit at the odd harmonics. The class-F PA is capable of 100% efficiency in principle. The normalized power capability of class-F PAs is about 0.16, which is half that of the class-D PAs.

In summary, there is a trade-off between the efficiency and the linearity. For receivers with constant-envelope modulation, such as FSK, high-efficiency PAs can be adopted; for linear operation such as ASK (Amplitude Shift Keying), or systems with high ACPR requirement, high-linearity PAs can be adopted.

The high-efficiency PAs such as class-E are usually used in WSN transceivers, as the power consumption is the most significant specifications of WSN system.

#### 4.6 IF circuits

The function of IF circuits includes demodulation, data decision, and clock recovery. There are two main kinds of IF circuits: 1) the digital scheme. In common receivers, an ADC is connected after the RF front end, and the frequency detecting, data decision and received signal strength indicating are all realized in digital domain, then the performance of the circuit can be easily improved in digital domain. Such is the general architecture for SDR as we described in section 2.5. However, the ADC usually consumes a large amount of power, and a high-linearity AGC circuit is required before the ADC. 2) The analog scheme. For low-power applications such as WSN, the CMOS analog resolution is sometimes a reasonable choice since the power consumption can be saved a lot.

#### 4.7 ADC and DAC

There are two kinds of ADCs in a WSN node. One is connected after the sensor and used for data sampling, as shown in Fig. 1; and another is used in the transceiver, as shown in Fig. 6.

The main parameters of an ADC can be summarized into several aspects:

- Resolution: The minimum voltage level that can be discriminated by the ADC is  $V_{ref}/2^N$  for a N-bit ADC with an input range from 0 to  $V_{ref}$ .
- DNL (Differential Non-Linearity): The maximum deviation between the actual conversion step and the ideal conversion step.
- INL (Integrated Non-Linearity): The maximum deviation of actual center of bin from its ideal location.
- Offset: The non-zero voltage or current at the output of the ADC when the input is zero since the OTAs or comparators have offset voltages and offset currents.
- Gain Error: The deviation of actual input voltage from the ideal value when the ADC outputs the full-scale bits.

- SNR (Signal-to-Noise Ratio): The theory equation of SNR for a N-bit ADC can be expressed as:

$$SNR = 6.02N + 1.76 \quad (33)$$

- SNDR (Signal-to-Noise and Distortion Ratio): The power of the noise and harmonics divided by the power of signal.
- SFDR (Spurious Free Dynamic Range): The ratio of the signal's power to the maximum harmonic's power.
- ENOB (Effective Number of Bits): The ENOB can be calculated by:

$$ENOB = (SNDR - 1.76)/6.02 \quad (34)$$

- THD (Total Harmonic Distortion): The ratio of all the harmonics' power to the signal's power.

Besides the specifications above, another very important parameter is power consumption. In the aspect of architecture, the ADCs can be divided into flash, SAR (Successive Approximation), folding, pipeline, sigma-delta, and so on.

For the ADC connected after the sensor in a WSN node, SAR ADCs may be the best choice since the data collection is executed at most of the time. There are two reasons: 1)The only power hungry module in a SAR ADC is the comparator, so the overall power consumption is low; 2)the circuit structure of a SAR ADC is simple, so the cost of chip area is small. In the WSN transceivers, the data rate is not high and the modulation scheme is simple, so a low bandwidth and low precision ADC is usually enough. Therefore, a SAR ADC may also be a good choice for WSN transceivers.

The specifications of DAC also includes DNL, INL, SNR, SNDR, SFDR, power consumption, and so on. The DACs are often used in transmitters, as shown in Fig. 6, the shaped digital waves are converted to analog signals that are sent to up-converting mixers for modulation. As a result, the performance of the DACs will affect the EVM and ACPR of the transmitters.

## 5. Conclusion

The goal of this chapter is to give a brief manual for WSN transceiver design. Section 1 gives an introduction of WSN and the RF transceivers for WSN. The WSN transceivers are classified by both modulation schemes and architectures in section 2. How to calculate and assign the system specifications are described in section 3. The design of key modules is analyzed briefly in section 4. The readers is expected to master a top-to-down design method for WSN transceivers through the chapter.

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## **Wireless Sensor Networks: Application-Centric Design**

Edited by Yen Kheng Tan

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Over the past decade, there has been a prolific increase in the research, development and commercialisation of Wireless Sensor Networks (WSNs) and their associated technologies. WSNs have found application in a vast range of different domains, scenarios and disciplines. These have included healthcare, defence and security, environmental monitoring and building/structural health monitoring. However, as a result of the broad array of pertinent applications, WSN researchers have also realised the application specificity of the domain; it is incredibly difficult, if not impossible, to find an application-independent solution to most WSN problems. Hence, research into WSNs dictates the adoption of an application-centric design process. This book is not intended to be a comprehensive review of all WSN applications and deployments to date. Instead, it is a collection of state-of-the-art research papers discussing current applications and deployment experiences, but also the communication and data processing technologies that are fundamental in further developing solutions to applications. Whilst a common foundation is retained through all chapters, this book contains a broad array of often differing interpretations, configurations and limitations of WSNs, and this highlights the diversity of this ever-changing research area. The chapters have been categorised into three distinct sections: applications and case studies, communication and networking, and information and data processing. The readership of this book is intended to be postgraduate/postdoctoral researchers and professional engineers, though some of the chapters may be of relevance to interested master's level students.

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