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FM-UWB as a Low-Power, Robust Modulation Scheme

3.1 Introduction

Ultra-wideband (UWB) systems were originally intended to provide robust, low-cost, low-complexity and low power wireless solutions for localization and communication. The first UWB systems were based on a time domain approach, they used a very short pulse to carry the information. Initially, they were used in radar systems, where pulse duration translated into spatial resolution. When used for communications, these pulses could be modulated using one of the standard approaches, such as OOK, PPM, PSK or FSK. This was impulse radio (IR) UWB, and although it was able to provide robust, high-speed communication, it came at the price of circuit complexity and relatively high peak power consumption. The frequency-modulated (FM) UWB was developed as an easy to implement, complementary solution, preserving robustness and offering low to medium data rates. This analog spread spectrum technique is intended for short to medium range applications that require a reliable communication link, low cost and high degree of integration and miniaturization, and therefore perfectly fits the IoT requirements.

This chapter begins by introducing the fundamentals of FM-UWB, explaining the modulation and demodulation principles and basic transmitter and receiver architectures. Then, the Gerrits' BER approximation is presented and extended to cases with multiple FM-UWB users and narrowband interferers. Finally, possible extensions of standard FM-UWB modulation are briefly discussed, highlighting its potential evolution. In the second part of this chapter, state of the art FM-UWB receivers and transmitters are discussed and analyzed, and a brief summary of their key characteristics is provided. They are also compared to narrowband and IR-UWB radios to point out the advantages and disadvantages of the FM-UWB modulation scheme.

3.2 Principles of FM-UWB

3.2.1 FM-UWB Modulation

The FM-UWB can be seen as an analog spread-spectrum technique. In its basic form it is a double FM modulation. A low modulation index FSK, called a sub-carrier, is followed by a high modulation index FM ($\beta \gg 1$) to achieve large bandwidth. The principle of FM-UWB modulation is shown in Figure 3.1. The resulting FM-UWB signal can be represented as [1]:

$$s_{UWB}(t) = A\cos\left(\omega_c t + \Delta\omega \int_{-\infty}^{t} m(t)\mathrm{d}t\right) = A\cos\left(\omega_c t + \phi(t)\right), \quad (3.1)$$

where ω_c is the center frequency, $\Delta\omega = 2\pi\Delta f$ is the frequency deviation and $m(t)$ is the normalized, FSK modulated sub-carrier. According to definition, to be considered UWB the signal must either exceed 500 MHz or 20% of its center frequency. The bandwidth of the FM signal can be approximated using the Carson's rule [1]:

$$B_{FM} = 2f_m(\beta + 1) \;=\; 2(\Delta f + f_m). \quad (3.2)$$

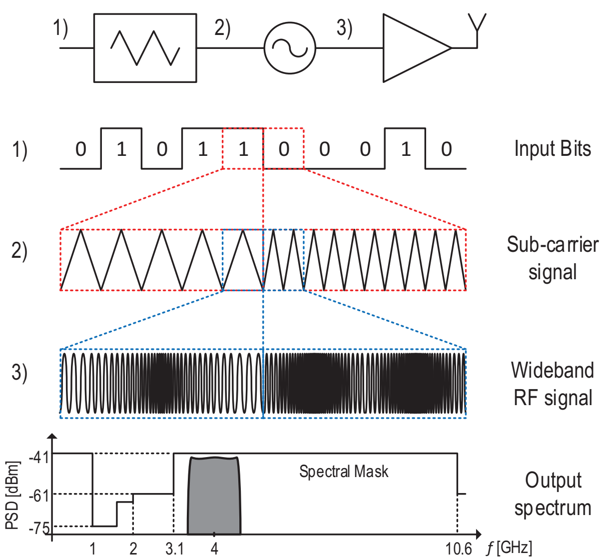


Figure 3.1 Principle of FM-UWB signal modulation.

In the above equation f_m is the maximum frequency in the FSK signal spectrum which depends on the sub-carrier center frequency f_{SC} and the data rate R, according to $f_m = f_{SC} + R$. Spectral properties of the FM-UWB signal depend on the sub-carrier waveform. For an FM signal with modulation index much larger than unity, quasi-stationary approximation is valid and the FM-UWB signal power spectral density (PSD) will be a function of the probability density function (PDF) p_m of $m(t)$ [2]:

$$S_{FM-UWB}(\omega) = \frac{\pi A^2}{2}\left[p_m\left(\frac{\omega - \omega_c}{\Delta\omega}\right) + p_m\left(\frac{\omega + \omega_c}{\Delta\omega}\right)\right]. \qquad (3.3)$$

As long as the sub-carrier frequency is reasonably low (keeping the second FM modulation index high, $\beta \gg 1$), the FM-UWB spectrum will be largely determined by the sub-carrier waveform. For an ideal triangular sub-carrier the FM-UWB spectrum will be flat with a relatively steep roll-off. A steeper roll-off can be achieved by using a sinusoidal sub-carrier, but this results in curved spectrum shape, with peaking at the edges of the band [3]. As a result the maximum transmit power must be lowered in order to comply with the spectral mask. At higher sub-carrier frequencies, or equivalently lower modulation index (practically $\beta < 20$) Equation (3.3) is no longer valid, and good spectral properties of the FM-UWB signal are lost.

Performance of the FM-UWB modulation can be studied using a wide-band FM demodulator presented in Figure 3.2. After multiplying the signal s_{UWB} with its delayed version and disregarding the high-frequency components, signal at the output of the demodulator will be given by [1]:

$$s_{dem}(t) = \frac{A^2}{2}\cos(\omega_c\tau + \phi(t) - \phi(t-\tau)). \qquad (3.4)$$

By choosing the time delay equal to an odd multiple N of the quarter period of the carrier center frequency $\tau = NT/4 = N\pi/2\omega_c$

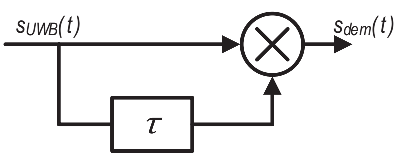


Figure 3.2 Wideband FM demodulator.

($N = 1,\ 3,\ 5, \ldots$) Equation (3.4) can be written in the following form:

$$s_{dem}(t) = (-1)^{(N+1)/2} \frac{A^2}{2} \sin(\phi(t) - \phi(t-\tau)) \tag{3.5}$$

$$= (-1)^{(N+1)/2} \frac{A^2}{2} \sin\left(\tau \frac{d\phi(t)}{dt}\right) \tag{3.6}$$

$$= (-1)^{(N+1)/2} \frac{A^2}{2} \sin\left(N \frac{\pi \Delta\omega}{2\omega_c} m(t)\right), \tag{3.7}$$

under the assumption that the delay τ is much smaller than the period of the modulating frequency f_m. The bandwidth of the demodulator, herein defined as the frequency range over which the demodulator characteristic is monotonic, depends on N and is given by

$$B_{dem} = f_c \frac{2}{N}. \tag{3.8}$$

A small delay deviation results in offset between the demodulator center frequency and the FM-UWB signal center frequency. This offset will lead to a distortion of the output signal that is dependent on the bandwidth of the signal and the demodulator. It should be noted that the demodulated signal is proportional to the square of the input amplitude (as seen from Equation 3.7). This results in expanded dynamic range of the demodulated signal, for example a 10 dB variation in the input amplitude causes 20 dB variation in the demodulated signal amplitude. Furthermore, the signal to noise ratio (SNR) at the demodulator output will be a non-linear function of the input SNR. Based on simplified analysis provided in [1] the SNR at the demodulator output is given by

$$\mathrm{SNR_{out}} = \frac{B_{RF}}{B_{SC}} \frac{\mathrm{SNR_{in}^2}}{1 + 4\mathrm{SNR_{in}}}, \tag{3.9}$$

where $\mathrm{SNR_{in}}$ and $\mathrm{SNR_{out}}$ represent the signal to noise ratio at the input and the output of the demodulator, respectively. The ratio B_{RF}/B_{SC} is the ratio of the FM-UWB signal bandwidth and sub-carrier bandwidth, and can be seen as a kind of analog processing gain. At the demodulator output the ratio of the energy per bit and the noise PSD is given by

$$(E_b/N_0)_{\mathrm{dem}} = \mathrm{SNR_{out}} \frac{B_{SC}}{R}, \tag{3.10}$$

where R is the data rate. As shown in [1], assuming a coherent, optimal demodulator and an orthogonal FSK sub-carrier modulation, the BER can be calculated as:

$$P_b = \frac{1}{2}\mathrm{erfc}\left(\sqrt{\frac{(E_b/N_0)_{\mathrm{dem}}}{2}}\right). \tag{3.11}$$

The erfc function is defined as:

$$\mathrm{erfc}(x) = \frac{2}{\sqrt{\pi}}\int_x^{\infty} e^{(-t^2)}dt. \tag{3.12}$$

In most cases, however, a practically implemented FSK demodulator will be non-coherent. Even though there is a small performance penalty, a non-coherent demodulator is simpler, easier to implement and consumes less power. For a non-coherent demodulator the probability of error is given by

$$P_b = \frac{1}{2}e^{\frac{(E_b/N_0)_{\mathrm{dem}}}{2}}. \tag{3.13}$$

The penalty for using a non-coherent demodulator is below 1.5 dB.

A comparison between coherent FM-UWB and FSK signals having equal power is given in Figure 3.3. The ratio of energy per bit and noise power spectral density at the input E_b/N_0 is used instead of $\mathrm{SNR}_{\mathrm{in}}$ in order to provide a fair comparison. This ratio is defined as

$$E_b/N_0 = \mathrm{SNR}_{\mathrm{in}}\frac{B_{RF}}{R}. \tag{3.14}$$

In the given example $B_{RF} = 500\,\mathrm{MHz}$, the sub-carrier modulation index is $\beta_{sub} = 0.5$ (the same modulation index is used for FSK) and R is the

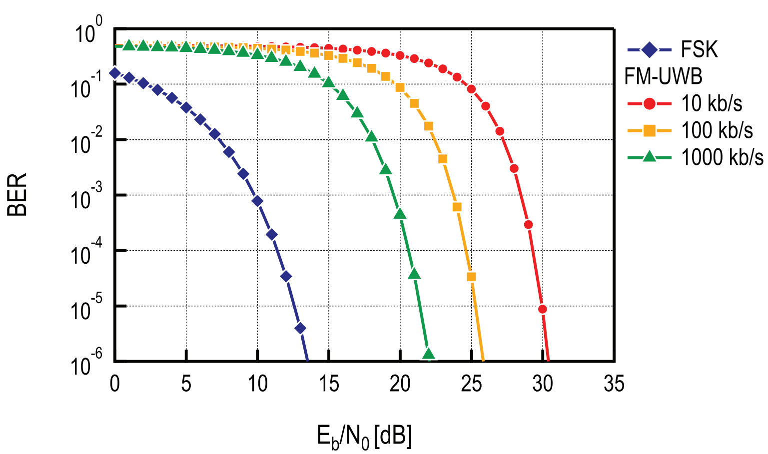


Figure 3.3 Comparison of standard orthogonal FSK and FM-UWB modulation.

data rate. In terms of BER, the FM-UWB is clearly suboptimal compared to standard FSK modulation. This is not surprising considering that the larger signal bandwidth results in higher noise power, that ultimately lowers the input SNR. Part of the lost SNR will be recovered in the process of demodulation owing to the processing gain, however the BER degradation compared to the FSK remains notable. The gap between the two modulations decreases with increasing the FM-UWB data rate, and hence higher data rates should yield better performance. However, higher data rates will also require higher sub-carrier frequencies, which results in loss of the spectral properties of the FM-UWB signal.

Compared to narrowband modulations, FM-UWB is clearly suboptimal in terms of sensitivity. There are however some benefits that are perhaps not apparent at a first glance. The FM-UWB offers robustness against frequency selective fading and interferers. The behavior of FM-UWB signal in a multipath environment has been studied in [4]. It has been shown that even in severe environments performance degradation of FM-UWB is only minor. Owing to the fact that the signal is spread over a very large band, frequency selectivity is not as harmful as it is for narrowband signals (this can be seen as a kind of frequency diversity). The second benefit of using FM-UWB is its inherent robustness to narrowband interferers. Unlike narrowband systems that rely purely on filtering, the FM-UWB provides some inherent interferer rejection. This further implies that it does not require an increase of the receiver complexity or external filters to provide good performance, hence providing higher potential for miniaturization.

3.2.2 Multi-User Communication and Narrowband Interference

In a wireless sensor network, multiple nodes may need to communicate at the same time. One way to resolve this is the time-division multiple access (TDMA), that allocates time slots in which certain nodes can transmit or receive. This approach requires precise synchronization between the nodes, and as the number of nodes in the network grows, the latency increases quickly. Use of other techniques, such as frequency-division multiple-access (FDMA), where different frequencies are allocated to different users, may reduce the overall latency and synchronization requirements. This section studies the behavior of an FM-UWB system in the presence of multiple input signals and is mainly based on the approach presented in [1].

Suppose there are two signals present at the input of the wideband FM demodulator (Figure 3.2) $s_1(t)$ and $s_2(t)$. At the demodulator output the signal will be given by:

$$s_{dem} = s_1(t)s_1(t-\tau) + s_2(t)s_2(t-\tau) \\ + s_1(t)s_2(t-\tau) + s_1(t-\tau)s_2(t). \quad (3.15)$$

Let us assume that the $s_1(t)$ is the FM-UWB signal and the $s_2(t)$ is a narrowband interferer. The component $s_1(t)s_1(t-\tau)$ corresponds to the demodulated sub-carrier. The component $s_2(t)s_2(t-\tau)$ is the FM demodulated narrowband signal. Since its bandwidth is rather small compared to the FM-UWB bandwidth, this component will be located close to dc and can easily be filtered out. It will therefore not influence the sensitivity of the receiver (at least in the ideal case). The last two terms in Equation (3.15) constitute the residual signal that will pollute the useful signal [1]:

$$W(t) = s_1(t)s_2(t-\tau) + s_1(t-\tau)s_2(t). \quad (3.16)$$

The low-frequency terms of the residual signal $W(t)$ will fall within the sub-carrier band, effectively increasing the noise floor of the receiver and lowering sensitivity. Assuming the narrowband signal is located close to the FM-UWB signal center frequency, residual signal will be located at baseband frequencies from 0 to $B_{RF}/2$. If flat spectrum of the residual signal is further assumed, then the signal to interference ratio can be estimated as [1]:

$$\mathrm{SIR} = 20\log\left(\frac{A_1}{2A_2}\right) - 10\log\left(\frac{B_{RF}}{2B_{SC}}\right), \quad (3.17)$$

where A_1 and A_2 are the amplitudes of the two input signals. Factor $B_{RF}/2B_{SC}$ is a result of sub-carrier filtering. Interestingly, the amount of interference rejection is proportional to the FM-UWB processing gain.

Multiple FM-UWB signals can be distinguished by assigning different sub-carrier frequencies to different users. This technique will be referred to as the sub-carrier FDMA (SC-FDMA). Assuming that the signals $s_1(t)$ and $s_2(t)$, from Equation (3.15), are the two FM-UWB signals it is clear that the simultaneous demodulation of different FM-UWB signals is possible. The component $s_2(t)s_2(t-\tau)$ will in this case correspond to the second demodulated FM-UWB signal. Provided that the sub-carrier frequency of the second signal is separated from the first, the two can be distinguished and demodulated separately. The principle of multi-user communications using the SC-FDMA is illustrated in Figure 3.4. Multiple signals transmitted from different nodes can be demodulated either by a single node (for example

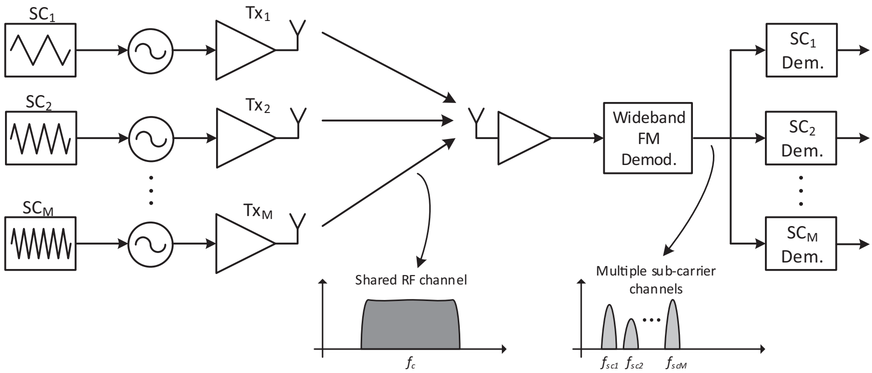


Figure 3.4 FM-UWB multi-user communication.

gathering data from multiple sensors simultaneously), or by different nodes (e.g. to allow isolation of different parts of the network).

Just like in the case of the narrowband interferer, the residual signal will cause sensitivity degradation. Assuming two FM-UWB signals at the input, with aligned center frequencies, Equation (3.16) can be written as:

$$\begin{aligned} W(t) &= \frac{A_1 A_2}{2} \cos\left(\omega_c \tau + \phi_1(t) - \phi_2(t-\tau)\right) \\ &\quad + \frac{A_1 A_2}{2} \cos\left(\omega_c \tau + \phi_2(t) - \phi_1(t-\tau)\right) \qquad (3.18) \\ &\approx (-1)^{(N+1)/2} A_1 A_2 \sin\left(\frac{\tau}{2}\left(\frac{d\phi_1(t)}{dt} - \frac{d\phi_2(t)}{dt}\right)\right) \\ &\quad \times \cos\left(\phi_1(t) + \phi_2(t)\right) \qquad (3.19) \\ &= (-1)^{(N+1)/2} A_1 A_2 \sin\left(\frac{\Delta\omega\tau}{2}\left(m_1(t) - m_2(t)\right)\right) \\ &\quad \times \cos\left(\phi_1(t) + \phi_2(t)\right). \qquad (3.20) \end{aligned}$$

The residual signal is proportional to the difference of the two modulating signals multiplied by the factor $\cos(\phi_1(t) + \phi_2(t))$, which is a signal that occupies a bandwidth of $B_{RF} = 2\Delta f$. Again, assuming the spectrum of the residual signal is flat, signal-to-interference ratio can be estimated as [1]:

$$\mathrm{SIR} = 20\log\left(\frac{A_1}{2A_2}\right) - 10\log\left(\frac{B_{RF}}{B_{SC}}\right). \qquad (3.21)$$

The achievable BER is limited by the SIR. Increasing the number of users, or increasing the difference in power levels between the two users, reduce the SIR and could eventually prevent correct demodulation of the useful signal. The maximum required BER will ultimately limit the tolerable SIR, and subsequently, the number of users or the maximum acceptable power difference.

The analysis conducted by Gerrits in [1, 5] can be extended to the case of multiple FM-UWB users in the presence of noise. Assuming that the delay can be considered relatively small and that the noise autocorrelation function is $R_n(\tau) = 1$ for values of the delay $\tau = N\pi/2\omega_c$, the noise analysis can be simplified while maintaining good accuracy. The result reported in [6] can be generalized for the case of M FM-UWB users. Under the above assumptions the demodulator output signal is given by

$$s_{dem} = (s_1 + s_2 + \cdots + s_M + n)^2 \tag{3.22}$$

$$= \sum_{i=1}^{M} s_i^2 + 2\sum_{i=1}^{M}\sum_{j=i+1}^{M} s_i s_j + \sum_{i=1}^{M} s_i n + n^2. \tag{3.23}$$

Terms of the form s_i^2 correspond to the demodulated sub-channel i. Terms of the form $2s_is_j$, $i \neq j$, correspond to the interference among different FM-UWB signals, the number of these terms is $M(M-1)/2$. Finally, following the same reasoning as in [1, 5], and assuming that all the noise and interference terms are independent, the output signal to noise and interference ratio (SNIR) is given by

$$\mathrm{SNIR}_{\mathrm{k,out}} = \frac{B_{RF}}{B_{SC}} \frac{S_k^2}{N^2 + 4\sum_{i=1}^{M} S_i N + 4\sum_{i=1}^{M}\sum_{j=i+1}^{M} S_i S_j}, \tag{3.24}$$

where S_i corresponds to the input power of signal s_i and N is the input noise power. For a multi-user environment two cases are of particular importance:

1. Two FM-UWB users of different input power levels
2. M FM-UWB users of equal power levels

For the case of two users, Equation (3.24) reduces to

$$\mathrm{SNIR}_{1,\mathrm{out}} = \frac{B_{RF}}{B_{SC}} \frac{S_1^2}{N^2 + 4S_1N + 4S_2N + 4S_1S_2} \tag{3.25}$$

$$= \frac{B_{RF}}{B_{SC}} \frac{\mathrm{SNR}_{1,\mathrm{in}}^2}{1 + 4\mathrm{SNR}_{1,\mathrm{in}}(1 + \mathrm{SIR}_{\mathrm{in}}^{-1}) + 4\mathrm{SNR}_{1,\mathrm{in}}^2 \mathrm{SIR}_{\mathrm{in}}^{-1}}, \tag{3.26}$$

where $\mathrm{SIR}_{\mathrm{in}} = S_1/S_2$. Compared to Equation (3.9) two additional terms exist that depend on the input signal to interferer ratio $\mathrm{SIR}_{\mathrm{in}}$. Furthermore, for increasing values of $\mathrm{SNR}_{1,\mathrm{in}}$ the output signal to noise and interference ratio $\mathrm{SNIR}_{1,\mathrm{out}}$, is no longer limited by noise, but solely by the interference and approaches

$$\mathrm{SNIR}_{1,\mathrm{out}} = \frac{B_{RF}}{B_{SC}} \frac{\mathrm{SIR}_{\mathrm{in}}}{4}, \quad \text{for } \mathrm{SIR}_{\mathrm{in}} \gg 1, \tag{3.27}$$

which is the limit from Equation (3.21). As an example, consider that the FM-UWB signal is used with RF bandwidth $B_{RF} = 500\,\mathrm{MHz}$, using a 100 kb/s sub-carrier, with orthogonal FSK and a modulation index of 1 ($B_{SC} = 200\,\mathrm{kHz}$). The required SNIR for orthogonal FSK to achieve a BER of 10^{-3} is approximately 13 dB. The maximum difference in power levels between the two FM-UWB signals is then 21 dB.

For the case of M users of equal input power, $S_1 = S_2 = \cdots = S_M = S$, the Equation (3.24) reduces to

$$\mathrm{SNIR}_{\mathrm{out}} = \frac{B_{RF}}{B_{SC}} \frac{S^2}{N^2 + 4MSN + 2M(M-1)S^2} \tag{3.28}$$

$$= \frac{B_{RF}}{B_{SC}} \frac{\mathrm{SNR}_{\mathrm{in}}^2}{1 + 4M\mathrm{SNR}_{\mathrm{in}} + 2M(M-1)\mathrm{SNR}_{\mathrm{in}}^2}. \tag{3.29}$$

Again, if the signal power is sufficiently higher than the noise power, the output $\mathrm{SNIR}_{\mathrm{out}}$ becomes a function of FM-UWB signal bandwidth and the number of users:

$$\mathrm{SNIR}_{\mathrm{out}} = \frac{1}{2M(M-1)} \frac{B_{RF}}{B_{SC}}, \quad \text{for } \mathrm{SIR}_{\mathrm{in}} \gg 1. \tag{3.30}$$

The above equation can be used to determine the maximum achievable number of users, for a given minimum required signal to noise and distortion ratio. For example, assuming the same system parameters as above ($B_{RF} = 500\,\mathrm{MHz}$, $R = 100\,\mathrm{kb/s}$, $B_{SC} = 200\,\mathrm{kHz}$), the maximum number of equal power users is 16. In both described cases FM-UWB signal bandwidth can be increased in order to increase the achievable $\mathrm{SNIR}_{\mathrm{out}}$.

The limits predicted by Equations (3.27) and (3.30) assume an ideal system since they only take into account a limited number of effects. In practical systems, these limits are upper bounds and will be difficult to achieve. The above analysis only considers an ideal AWGN channel, with a perfectly flat frequency characteristic. In reality, this will never be the case. Part of the channel transfer function will come from the transmitter and receiver, and part will come from the wireless channel (e.g. due to multi-path propagation). Intuitively, one can see the FM-UWB signal as a carrier that slowly

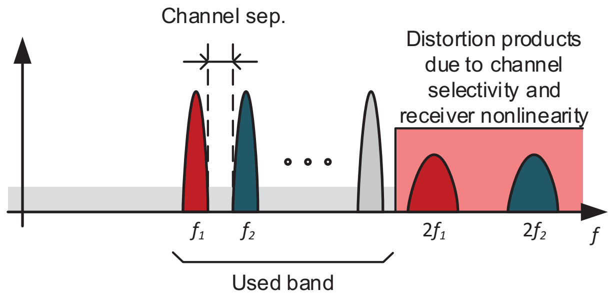


Figure 3.5 FSK sub-channel frequency allocation and limits due to distortion.

moves across a broad frequency range. Since the equivalent channel transfer function is not constant, the amplitude will vary with the instantaneous carrier frequency. Assuming that the channel does not change with time, amplitude will be a periodic function, with the period of the sub-carrier. Even if the wideband FM demodulator is perfect, these amplitude variations will result in the appearance of harmonics. Aside from the channel transfer function, the harmonics will also appear as a product of non-linearities in the receiver chain. Finally, these harmonics will limit the useful sub-carrier band to one octave. If $f_{SC,min}$ is the minimum sub-carrier frequency, then spectrum above $2f_{SC,min}$ will be corrupted by the second and higher order components. The quality of any useful signal at frequencies above $2f_{SC,min}$ would, therefore, be degraded by the harmonics of other sub-carriers, effectively preventing correct demodulation (Figure 3.5). With one octave limit for the sub-carrier band, the lowest sub-carrier frequency $f_{SC,min} = 1\,\text{MHz}$, and 200 kHz wide FSK channels, the number of sub-carrier channels that can be accommodated is 5. This number can be increased by increasing $f_{SC,min}$. In principle, the effect of channel transfer function can be canceled out by equalization of the input FM-UWB signal. However, equalization techniques are complex, they would drastically increase power consumption of the receiver, and as such are not suited for low power systems.

In the previous example, it was assumed that the FSK channels can be placed adjacent to each other. This is impractical for two reasons. The first reason is that the undesired channels must be filtered out before the final FSK demodulation. Because of the finite quality factor of the filter, some spacing must be introduced between the channels. The second reason is interference among adjacent FSK channels. Theoretically the spectrum of the FSK signal is infinitely wide. Although largest portion of the channel power is located inside the band defined by Carson's rule, part of the spectrum will leak

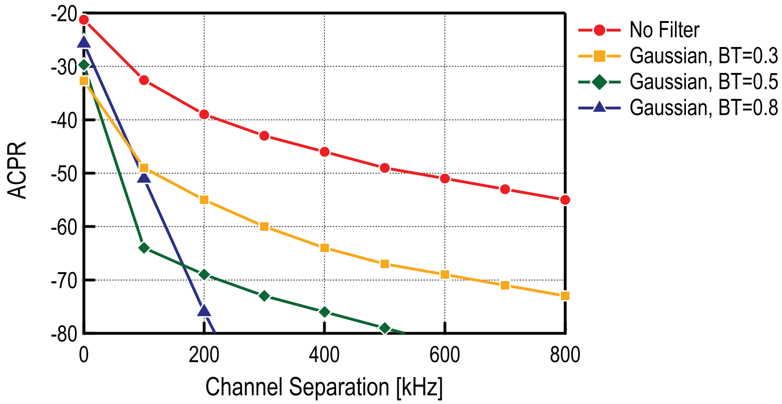


Figure 3.6 ACPR for filtered and non-filtered FSK signal, as a function of channel separation (100 kb/s data rate, modulation index 1).

to side channels and interfere with adjacent users. This effect is quantified by the adjacent channel power ratio (ACPR), and is defined as the ratio of the power inside the channel to the power in the adjacent channel. The ACPR generally depends on the type of modulation, pulse shaping filter and transmitter non-linearity. In the case of FSK modulation, typically Gaussian pulse shaping is used. The shape of the Gaussian pulse is determined by the bandwidth-time (BT) parameter, defined as the ratio between the 3 dB filter bandwidth and data rate. Decreasing the BT parameter results in more compact spectrum, but increases the inter-symbol interference as the pulse duration increases (over several bit periods). ACPR as a function of channel separation, for different values of the BT parameter, is given in Figure 3.6. Although filtering can be used to reduce interference, this was rarely done in reported FM-UWB implementations. The reason is that it adds complexity on both transmitter and receiver sides, and since multi-user communication with FM-UWB has rarely been explored it was not needed. Interference among channels can always be decreased by increasing the channel separation, but this also reduces the number of available FSK channels. For a system with $B_{RF} = 500\,\text{MHz}$, $R = 100\,\text{kb/s}$, $B_{SC} = 200\,\text{kHz}$, the required SNIR of the FSK signal to achieve a BER of 10^{-3} is 13 dB. If a channel separation of 100 kHz is used, with no filtering, then the adjacent channel power can be at most 20 dB above the desired channel power. This will correspond to 10 dB difference in power between the two FM-UWB signals. For this particular case, it is the ACPR that will limit the maximum tolerable power difference between the two users and not the interference from the residual signal (Equation (3.27)).

Additional constraints may come from the receiver non-linearity and limited dynamic range. Due to the quadratic demodulator characteristic, the dynamic range requirements are higher for the circuits following the wideband FM demodulator. If one of the FSK signals is sufficiently strong it may saturate the circuits causing suppression of weaker FSK signals (FM capture effect). Since there is typically a trade-off between power and dynamic range in amplifiers, a larger acceptable power difference between the received signals will come at the cost of increased power consumption.

Different choice with respect to the system parameters leads to different performance in terms of complexity, sensitivity, data rate, number of channels and power consumption. By modifying the RF bandwidth, sub-carrier frequencies, dynamic range etc., it is possible to perform various trade-offs and to optimize the FM-UWB transceiver according to the specific needs of the system.

3.2.3 Beyond Standard FM-UWB

The FM-UWB modulation was originally intended as double FM modulation, where a low modulation index FSK is followed by a large modulation index FM. It is an optional mode in the IEEE 802.15.6 standard for wireless body area networks [7]. According to the UWB PHY specifications, two modulations are supported; IR-UWB as mandatory and FM-UWB as an optional mode. For FM-UWB, the data rate is set to 250 kb/s, using a continuous phase (CP) FSK modulation, centered at 1.5 MHz, with a frequency deviation of 250 kHz. A Gaussian filter is used for pulse shaping with the BT parameter set to 0.8. For the sub-carrier waveform, either a triangular, a sawtooth, or a sine waveforms are allowed.

Strict standard definitions do not allow different sub-carrier frequencies, higher or lower data rates, or multi-user communication. The lack of flexibility limits the use of FM-UWB in WBAN applications, and does not allow FM-UWB to reach its full potential. In general, the sub-carrier modulation does not need to be limited to 250 kb/s 2-FSK. Speed and modulation order could be modified according to the channel conditions (a less frequency selective channel allows higher data rates). A transmitter implementing a data rate of 1 Mb/s has been reported in [8], that demonstrates the feasibility of moving to higher data rates. Furthermore, higher order FSK can be explored, such as 4-FSK and 8-FSK, allowing to further boost communication speed.

One such transmitter is reported in [9]. Finally, it would also be possible to use PSK modulations without affecting the good spectral properties of FM-UWB (note that standard PSK modulation requires a coherent SC demodulator). Other variations are possible, and one example is the Chirp-UWB (C-UWB) modulation [10], that is a trade-off between FM-UWB and IR-UWB. Instead of continuous frequency sweep, a single up or down chirp is transmitted depending on the input bit. The duration of the chirp is much lower than the symbol duration and allows duty cycling of the transceiver at a symbol level, thus saving power. At the same time the duration of the pulse is much longer than in the case of IR-UWB and does not require precise synchronization. One downside of C-UWB is that the good spectral properties of the FM-UWB signal are lost.

A minor modification of a standard FM-UWB signal can be used to enable simultaneous transmission on multiple sub-channels. Instead of using a single FSK sub-channel, multiple sub-channels can be summed, and the resulting signal used to modulate the RF carrier. This would allow a single transmitter to transmit different messages to multiple receivers at the same time. The concept is shown in Figure 3.7. In order to preserve the same frequency deviation, if M sub-channels are used, sub-carrier signals are scaled by a factor $1/M$. The example for two sub-carriers is shown in Figure 3.8. Unfortunately, the flat spectrum of the transmitted signal is lost and, as a consequence, transmit power will have to be decreased in order to maintain the signal below the spectral mask defined for the UWB band. The spectrum will take the shape of the PDF of the modulating signal (as shown by Equation (3.3)) which is in this case an average of the two sub-carrier signals, and is no longer a triangular waveform. The exact shape of the resulting sum of sub-carrier signals will

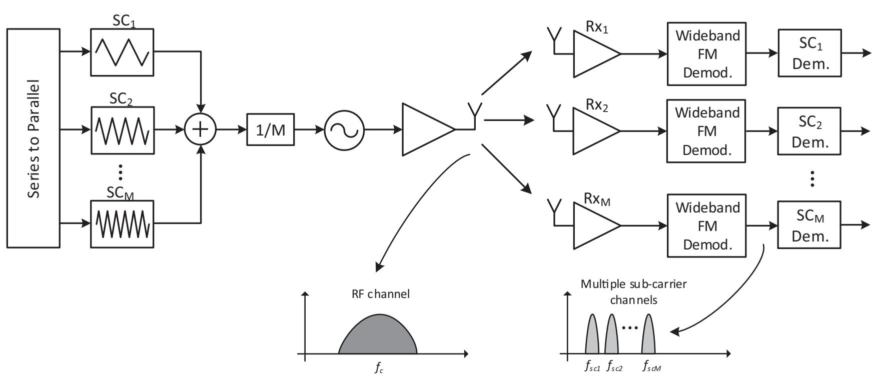


Figure 3.7 FM-UWB multi-channel broadcast.

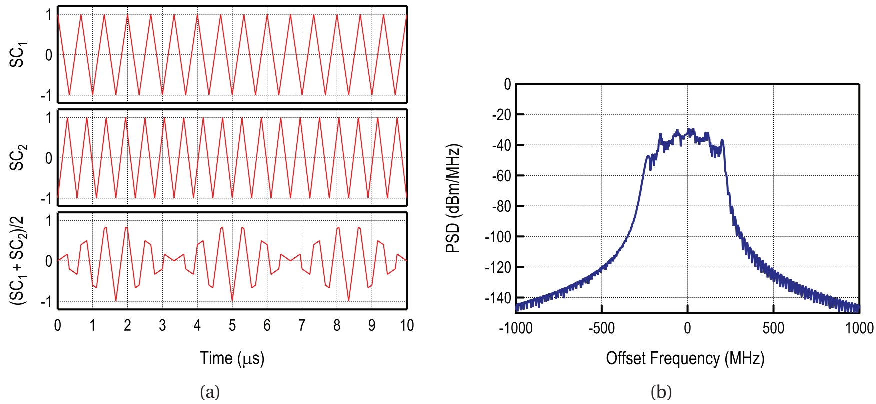


Figure 3.8 Example of transmission on two channels, time domain sub-carrier signal (a) and transmitted signal spectrum (b).

depend on the number of sub-channels, their frequencies and initial phases. The BER calculation can be extended to the case of M sub-channels. The only difference compared to standard FM-UWB is that the power of each channel is scaled by M. This is equivalent to reducing the RF bandwidth by the same factor and hence the $\mathrm{SNR_{in}}$ will be scaled as well. Equation (3.9) can then be modified accordingly to estimate the output SNR:

$$\mathrm{SNR_{out}} = \frac{B_{RF}}{B_{SC}} \frac{\mathrm{SNR_{in}^2}/M^2}{1 + 4\mathrm{SNR_{in}}/M}. \tag{3.31}$$

The probability of error is then calculated in the same way as for the single user case. One advantage of the proposed modification compared to the described multi-user scheme is that a larger number of channels can be used in the same bandwidth. If orthogonal sub-carrier frequencies are used, there will be no interference between the channels on the receiver side (in that sense the proposed scheme resembles OFDM). In the multi-user case, transmitters would have to be perfectly synchronized to preserve orthogonality, which is practically impossible, and as a result produces interference among different users. The only way to solve this is to separate and filter out the unwanted channels.

An existing degree of freedom in the proposed modulation technique is the sub-channels scaling. If different receivers are located at different distances from the transmitting node, the received power, and subsequently the BER, may vary. This can be circumvented by using a different scaling factor for each of the channels. Smaller scaling factor could be assigned to more distant receivers, in order to improve the BER on their sub-channels. As long as the sum off all scaling factors is 1, the maximum frequency deviation will remain the same, maintaining the signal spectrum within the defined limits.

3.3 State-of-the-Art FM-UWB Transceivers

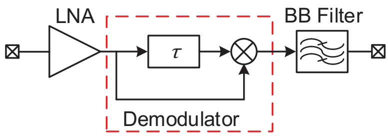
One of the main advantages of the FM-UWB is the simplicity of the transceiver architecture, which offers a low power consumption and a high degree of integration. Different transmitter and receiver implementations have been presented in the literature. They will be discussed in the following paragraphs, with a focus on both architecture and circuit level techniques. Finally, FM-UWB will be compared to state of the art narrow-band and IR-UWB receivers, to gain insight into some of the advantages and drawbacks of the chosen modulation scheme.

3.3.1 FM-UWB Receivers

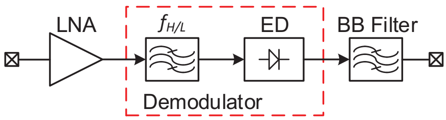
Different FM-UWB receiver architectures found in the literature are presented in Figure 3.9. The originally proposed wideband FM demodulator based on a delay line demodulator is depicted in Figure 3.9(a). Two other implementations are based on an FM discriminator, they rely on filtering to convert the input FM signal into an amplitude modulated (AM) signal. Conversion characteristics of all the demodulators are shown in Figure 3.10.

The FM-AM characteristic of the delay line demodulator was studied in the previous section (Equation (3.7)). The output AM signal will be a sine function of the input frequency. It can be seen that the choice of delay is a trade-off between the conversion gain and the bandwidth of the demodulator. Decreasing delay leads to lower conversion gain, but also increases the useful frequency range. In addition, this delay is constrained to a discrete set of values and must be equal to an odd multiple of the quarter period of the carrier frequency. It must be determined precisely in order to avoid frequency offset. In practice, a small offset will always be present as a result of process variation, however since the transmitted signal is at least 500 MHz wide, this offset should not have a major impact on the receiver performance. The first fully integrated FM-UWB receiver based on a DL demodulator was described in [11]. It achieves a sensitivity of −88 dBm while consuming 9.4 mW. The demodulator itself consumes around 5.8 mW, and the additional 3.6 mW are used by the LNA.

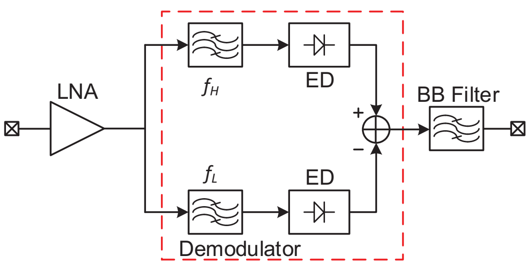
An LNA that provides high gain across a large bandwidth inevitably requires more power compared to a narrowband LNA. In order to reduce



(a) Delay-Line demodulator



(b) Regenerative demodulator



(c) Dual-Band-Pass-Filter demodulator (balanced frequency discriminator)

Figure 3.9 FM-UWB receiver architectures reported in the literature.

the power consumption, a narrow-band regenerative receiver was proposed in [12]. This approach allows for preservation of high gain and relatively good noise figure, while minimizing the power consumption. The high-Q filtering is in fact implemented in the LNA and its center frequency corresponds to either the highest or the lowest frequency of the FM-UWB signal. The band-pass filter behaves as a frequency discriminator that converts the input FM signal into an AM signal, that is then converted to IF using an envelope detector. Due to the high-Q factor of the filter that results in a very nonlinear FM-AM conversion characteristic, the demodulated signal will be a train of pulses whose frequency corresponds to the sub-carrier frequency (Figure 3.10). The receiver from [12] consumes 2.2 mW while achieving −84 dBm sensitivity. A later implementation presented in [13] introduced several improvements at the circuit level (most notably current

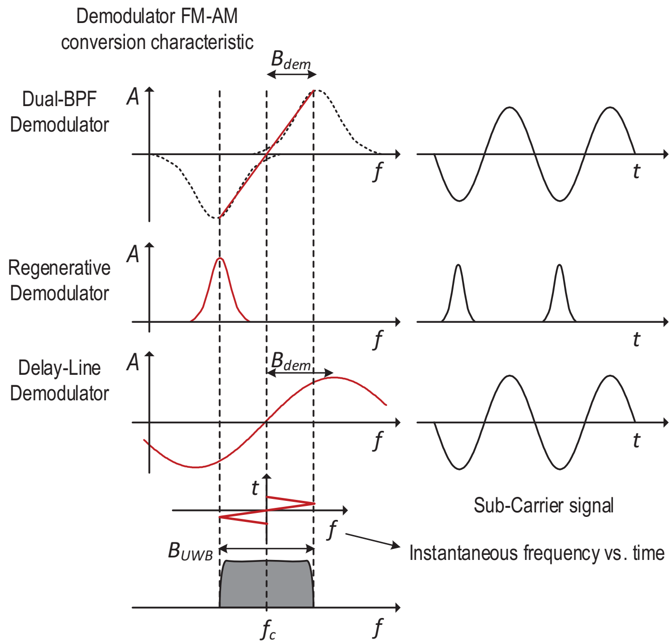


Figure 3.10 Frequency-to-amplitude conversion characteristic of reported FM demodulators.

reuse among several blocks) which resulted in power consumption of only 560 μW and only a slight reduction of sensitivity. Although the regenerative receiver achieved significant power savings, there are some downsides to this architecture. Narrow-band interferer rejection mostly relies on the high-Q input filtering. If the interferer falls inside the pass-band it could easily saturate the stages following the LNA and prevent reception. Indeed, such a scenario could be avoided by introducing the on-chip tuning circuit that could shift the filter center frequency, but this adds complexity to the system. The second downside comes from the nonlinear FM-AM conversion. If several FM-UWB signals were to occupy the same RF band, the weaker signals would be attenuated in the nonlinear conversion process, which would prevent correct demodulation. This is known as the capture effect [14], and limits the regenerative receiver to cases where only one FM-UWB signal is transmitted in the given RF band.

In an attempt to improve the linearity of the regenerative demodulator, a modified architecture was proposed in [15]. Instead of using just one band-pass filter, a second branch was added (Figure 3.9(c)), resulting in a

Dual Band-Pass Filter (DBPF) demodulator, otherwise known as a balanced frequency discriminator. The two filters are tuned to the highest and the lowest frequency in the FM-UWB signal spectrum, they are followed by the two envelope detectors that remove the RF carrier from the signal, and the difference of the two IF signals finally yields the demodulated sub-carrier. The equivalent linearized characteristic is shown in Figure 3.10. Compared to the original regenerative receiver, the Q-factor of the two filters can be lowered, which allows some power savings per filter, but the two still consume more than the single filter from [13]. The dominant source of power consumption remains the wideband LNA, that must provide equal gain over the entire band in the DBPF receiver. The two architectures perfectly illustrate the trade-off between linearity and power consumption in FM-UWB receivers. The implementation from [15] consumed 3.8 mW, and achieved −78 dBm of sensitivity. The same architecture was reused in [10] for demodulation of a Chirp-UWB signal, where symbol-level duty-cycling of the receiver was used to bring down the average power consumption to 0.6 mW. The DBPF receiver exhibits better narrow-band interferer rejection compared to a standard regenerative receiver and should perform better in scenarios with multiple FM-UWB users, although such capability was not confirmed by measurements.

A performance summary of different FM-UWB receivers is given in Table 3.1. Each of the proposed architectures has its own advantages and disadvantages. Receiver from [11] generally has the best performance but is

Table 3.1 Performance summary of state-of-the-art FM-UWB receivers

Reference	[11]	[12, 16]	[17]	[15]	[10]	[13, 18]
Year	2009	2010	2012	2013	2014	2014
Demodulator	RF-DL	Reg	RF-DL	DBPF	DBPF	Regen.
Frequency [GHz]	7.5	3.75	3.8	3.75	8	4
Power cons. [mW]	9.4	2.2	7.2	3.8	0.6/4*	0.58
Supply [V]	1.8	1	1.6	1	1	1
Data rate [kb/s]	50	100	50	100	1000	100
Sensitivity [dBm]	−88	−84	−70	−78	−76	−80.5
NB SIR [dB]	−25	−30	–	−23	–	−18
SC-FDMA	Yes	No	–	No	No	No
Efficiency [nJ/b]	188	2.2	144	38	1	5.8
Tech. node [nm]	250	90	180	65	65	90

*Power consumption is 0.6 mW with duty-cycling and 4 mW without duty-cycling.

also the most power hungry. The regenerative receiver can provide a very low power consumption while maintaining good sensitivity, but at the cost of linearity. A trade-off between linearity and interference rejection on one side, and power consumption on the other, is demonstrated with the balanced frequency discriminator from [15]. One thing that is common for all the architectures is that the largest contributors to the power consumption are the RF blocks, mainly the LNA. Therefore, one approach to decreasing consumption would be to minimize the number of RF blocks, or to completely remove them if possible. This approach will be studied in the following chapters.

3.3.2 FM-UWB Transmitters

Unlike the FM-UWB receivers, the architecture of FM-UWB transmitters has remained unchanged over the past several years. Considering its simplicity (Figure 3.1) it is clear that there is not a lot of potential for improvement at the architectural level. In fact, the reduction of power on the transmitter side is mainly a result of improvements at the circuit level. Every FM-UWB transmitter consists of three blocks, the sub-carrier generator, the VCO (sometimes as a part of a PLL or an FLL) and a power amplifier (PA).

The sub-carrier generator synthesizes the triangular waveform that is used to drive the VCO. As the sub-carrier frequencies are rather low (typically 1–2 MHz) this block does not contribute significantly to the overall transmitter consumption. One way to implement it is a Direct Digital Synthesis (DDS) as described in [19]. The advantages of digital implementation are the simple and precise frequency control without the need for calibration. The drawback of the fully digital approach becomes apparent at higher data rates, where higher sub-carrier frequencies are needed. In [8] 51 MHz sub-carrier frequency is used. Since roughly 20 points per period are needed to generate a reliable sub-carrier waveform, a DDS would need to operate at a clock speed of more than 1 GHz, which would be difficult to implement and would consume a significant amount of power. Instead, a relaxation oscillator is used within a PLL, a simpler and lower power solution in this case. Another interesting approach that leads to a very low power consumption is a free-running relaxation oscillator that is periodically calibrated using an FLL [20]. In this case, a digital frequency control is provided through a capacitor bank, however this approach is usually not precise enough if multiple sub-carrier channels are to be used. Additionally, it might occupy a larger area due to the size of capacitors needed at the frequency of interest.

The two main parts of the FM-UWB transmitter that essentially determine its power consumption are the VCO and the PA. In cases where the transmitted power is 10 dBm or more, the transmitter efficiency is dominated by the PA, however at lower output powers, such as -10 dBm the contribution of the VCO becomes quite significant. In some of the earlier implementations, the RF carrier was synthesized using an LC VCO within a PLL [8, 21]. To decrease power, the frequency synthesizer is duty cycled, making the frequency dividers active for only 10% of the time. Although this allowed some savings, the power consumption was still on the order of 10 mW. A significant improvement was made when the LC oscillator was replaced with a ring oscillator [9, 20]. This was possible owing to the loose phase noise constraints of the FM-UWB modulation. Additionally, instead of the quasi-continuous PLL, an FLL calibration loop was used [20]. Since the FM-UWB spectrum is very wide, the center frequency can deviate slightly without a major impact on performance and it does not need to be monitored continuously. Therefore, once calibrated, the VCO can operate in a free running manner until temperature or some other external factor causes a significant frequency shift. Since these external processes are usually slow, calibration only needs to be done once in a few hours or days, which makes the average power consumption of such an FLL practically negligible. The described approach led to the first sub-milliwatt FM-UWB transmitter [20]. The next step in reducing the VCO consumption was reducing the frequency of oscillation. Since an N-stage ring oscillator produces N equally spaced phases, these phases can be combined to produce a frequency that is N times higher [22]. It is then possible to use a ring oscillator that works at a frequency that is N times lower than the carrier center frequency. The approach was demonstrated in [22] and used for the FM-UWB transmitter in [13] to reduce the power consumption down to 0.63 mW. A three-stage ring was used that oscillated at one third of the carrier frequency, which resulted in the VCO power consumption of less than 90 μW.

Even though the VCO cannot be neglected, the PA remains the most power-hungry block in the system. The key to further reducing the power consumption of an FM-UWB transmitter is an efficient power amplifier. However, design of an integrated PA for such a low power and wide band poses a number of challenges. In standard narrow-band applications targeting 10 dBm output power or more, the most efficient approach is to use a switching PA such as class D or E. The first problem with class E is that the output matching network is set to a very narrow range of frequencies and achieving good efficiency over a large band would be impossible. The second problem with switching amplifiers is that their efficiency is directly related to the on-resistance of the switch, which dictates the minimum size of the output transistor. In addition, the PA must be driven by a square wave with very sharp transitions to minimize the turn-on time of the switch. The two requirements impose very hard constraints, resulting in power dissipation in the driving circuits that is comparable to that of the PA. Therefore, when the driving circuit is also accounted for, switching PAs seem not to be the best solution. The linear power amplifiers, classes A, AB, B and C, do not achieve as high efficiency, but their driving requirements are also lower. Moving from class A to class C operation, the maximum attainable efficiency increases, but the power gain decreases and larger driving signal is necessary, thus again shifting the burden from the PA to the driver. A good compromise is the class AB that attains decent efficiency and does not need a rail-to-rail input signal. In fact, all of the transmitters reported in [9, 13, 20], which achieve the lowest power consumption reported so far, use a complementary class AB power amplifier.

For linear PAs in general, optimal efficiency is obtained when the output voltage swing is maximized. In case of a complementary class AB or B amplifier, maximum output voltage swing is equal to the supply voltage. The load resistance seen from the power amplifier, must therefore be chosen such as to provide the desired output power. The problem with low power transmission is that the value of the optimal load resistance is relatively high. As a consequence a large transformation ratio of the matching network is needed, which then increases the losses in the network. One way to solve this problem would be to reduce the supply voltage. However, such an approach would require another circuit that would lower the voltage to the desired level, such as a DC-DC converter. This would not only increase complexity but also introduce its own losses and possibly require off-chip components. A better and simpler way is to apply current reuse technique demonstrated in [13], where the PA and the driver share the same current. Since the effective PA supply voltage is lower, there is no need for such a high transformation ratio of the matching network and, at the same time, the PA bias current is used to supply the driver. The efficiency of the transmitter can therefore be improved without any increase in complexity, which led to the current lowest power FM-UWB transmitter [13, 18], as shown in Table 3.2.

3.3.3 FM-UWB Against IR-UWB and Narrowband Receivers

So far the main characteristics of FM-UWB have been discussed, the potential of FM-WUB has been shown and the existing implementations of FM-UWB

Table 3.2 Performance summary of state-of-the-art FM-UWB transmitters

Reference	[21]	[8]	[20]	[17]	[9]	[10]	[13, 18]
Year	2010	2011	2011	2012	2013	2014	2015
SC Modulation	2-FSK	2-FSK	2-FSK	2-FSK	8-FSK	2-FSK	2-FSK
Frequency [GHz]	3.8	3.8	4	3.8	3.75	8	4
Bandwidth [MHz]	600	700	500	560	500	500	500
Power cons. [mW]	9.6	18.2*	0.9	8.7	1.14	3.5**	0.63
Supply [V]	1.6	1.6	1	1.6	1	1	1
Data rate [kb/s]	10	1000	100	50	750	1000	100
Out. power [dBm]	−14.5***	−12.8	−10.2	−13.7	−14	−11***	−10.1
Efficiency [nJ/b]	960	18.2*	9	174	1.5	0.39	3.1
Tech. node [nm]	180	180	90	180	65	65	90

*Excluding the output PA.
**In continuous mode, 0.39 mW with duty cycling.
***Estimated from figure.

transmitters and receivers have been presented. The question now is how does the FM-UWB compare to other low power modulation schemes? In Figure 3.11 FM-UWB receivers are placed together with low power receivers from Chapter 2, showing the data rate against power consumption. They consume a somewhat lower power than BLE receivers, but also target lower speed. In terms of power consumption they cannot achieve nanowatt levels of wake-up receivers.

As explained previously, if FM-UWB is compared to a standard FSK modulation, there is an inherent loss in sensitivity. Unfortunately, this is an unavoidable drawback of FM-UWB. If a narrowband receiver and an FM-UWB receiver using the same data rate perform similarly in terms of noise figure, the narrowband receiver will provide better sensitivity. This can be observed in Figure 3.12, where efficiency of receivers is plotted against sensitivity. The FM-UWB receivers cannot achieve the same sensitivity at comparable efficiency levels as the narrowband receivers. However, FM-UWB provides other benefits that may not be apparent at first. It is inherently robust against interferers, unlike NB radios that need to rely on filtering. Owing to the spread spectrum, FM-UWB is also robust against frequency selective fading. Narrowband radios might be unable to establish a link due to a notch in the channel frequency characteristic, whereas the FM-UWB only suffers a minor performance degradation. Also, FM-UWB

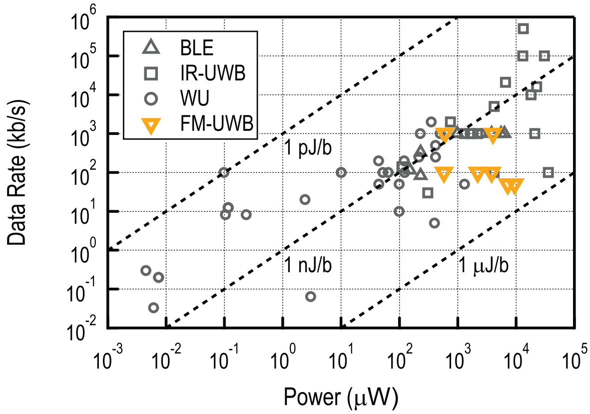


Figure 3.11 Comparison of FM-UWB receivers and other low power receivers from Chapter 2, data-rate against power consumption.

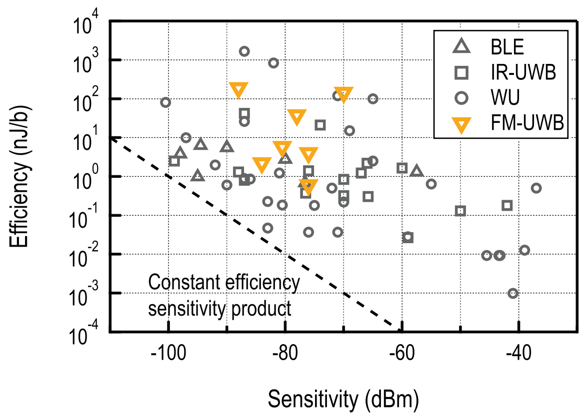


Figure 3.12 Comparison of FM-UWB receivers and other low power receivers from Chapter 2, efficiency against sensitivity.

could provide support for multi-user communication at almost no increase in power consumption. Finally, FM-UWB has better potential for miniaturization, enabling implementations with no off-chip components. Every narrowband radio needs a crystal oscillator to provide a precise frequency reference, and in most cases other off-chip components are needed to provide additional filtering, or output matching. Thanks to robustness to reference frequency offset, that partially comes from the large signal bandwidth, FM-UWB is capable of using an imprecise, on-chip reference oscillator, while still providing reliable communication. The combination of robustness, architecture simplicity and high degree of integration are, finally, the main arguments in favor of FM-UWB when compared to narrowband radios.

IR-UWB receivers cover a very wide range of data rates, from kb/s to almost Gb/s, while maintaining an almost constant efficiency. This is possible assuming symbol-level duty cycling can be applied. In order to benefit from symbol-level duty cycling the IR-UWB receivers must have a good timing reference, and require initial synchronization, both of which add complexity and cost to the system. The FM-UWB generally requires a fairly simple receiver architecture and has a lower peak power consumption making it cheaper and more appealing for battery powered systems. One other advantage of FM-UWB compared to IR-UWB is the multi-user capability. FM-UWB devices can transmit in the same RF band at the same time. A similar TDMA based scheme at the symbol level would be possible with IR-UWB wherein each transmitter has a time slot in which it can transmit a pulse during one symbol period. However, this would require a nanosecond level synchronization among the nodes, adding a prohibitively high level of complexity to the system.

3.4 Summary

The first part of this above chapter describes the main principles of the FM-UWB modulation. Basic calculations related to the modulation technique are presented and extended to the cases with multiple users. The described techniques, such as multi-user communication and multi-channel transmission, can be used to optimize the system performance according to the specific needs. Different sub-channels can be used, trading data-rate per channel with the number of available sub-channels, depending on the number of nodes in the network and their purpose.

In the second part of the chapter, the state of the art FM-UWB transceivers are discussed along with the the most important power reduction techniques reported in the literature. These techniques, combined with technology scaling, led to sub-milliwatt power consumption levels in today's implementations. The evolution of power consumption over the past 8 years is illustrated in Figure 3.13 for both transmitters and receivers, from which a decrease by a factor of 20 can be observed. However, the narrow-band receivers still have the edge, at least with respect to power consumption. The proposed wake-up receivers found in literature consume from 100 μW [23] all the way down to 100 nW [24]. FM-UWB can hardly compete with such low

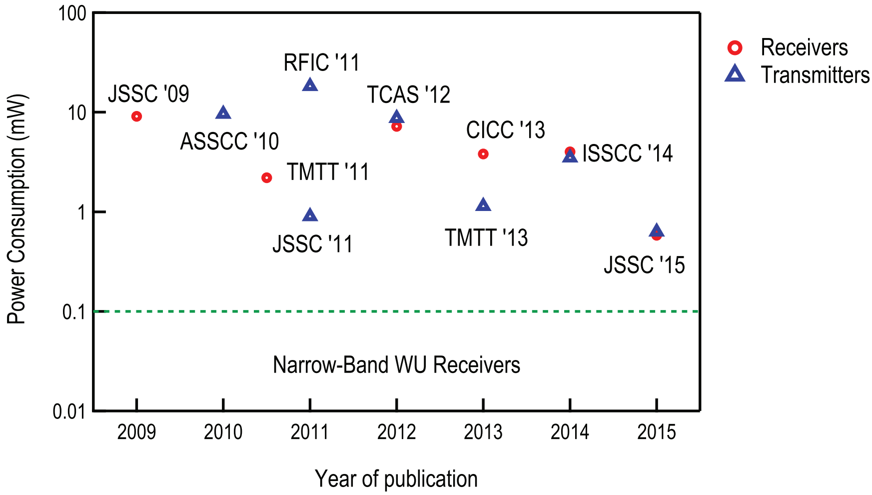


Figure 3.13 FM-UWB transmitters and receivers, evolution of power consumption. Type of demodulator used in each receiver is indicated on the graph.

levels, a simple consequence of the fact that wide-band circuits require more power to achieve the same performance in terms of gain and noise figure. On the other hand, the FM-UWB brings higher resilience to interferers, without off-chip components such as SAW filters, better performance in frequency selective channels and higher potential for miniaturization. All of these are very favorable capabilities that could assure a place for FM-UWB in short-range applications such as wireless body area networks.

References

[1] J. F. M. Gerrits, M. H. L. Kouwenhoven, P. R. van der Meer, J. R. Farserotu, and J. R. Long, "Principles and Limitations of Ultra-wideband FM Communications Systems," *EURASIP J. Appl. Signal Process.*, vol. 2005, pp. 382–396, Jan. 2005.

[2] M. Kouwenhoven, *High-Performance Frequency-Demodulation Systems*. Delft University Press, 1998.

[3] N. Saputra and J. Long, *FM-UWB Transceivers for Autonomous Wireless Systems:*, ser. River Publishers Series in Circuits and Systems. River Publishers, 2017.

[4] J. F. M. Gerrits, J. Farserotu, and J. Long, "Multipath Behavior of FM-UWB Signals," in *IEEE International Conference on Ultra-Wideband, 2007. ICUWB 2007*, Sep. 2007, pp. 162–167.

[5] ——, "Multi-user capabilities of UWBFM communications systems," in *IEEE International Conference on Ultra-Wideband*, Sep. 2005, pp. 1–6.

[6] V. Kopta, J. Farserotu, and C. Enz, "FM-UWB: Towards a robust, low-power radio for body area networks," *Sensors*, vol. 17, no. 5, 2017.

[7] *IEEE Standard for Local and Metropolitan Area Networks – Part 15.6: Body Area Networks*, 2012.

[8] B. Zhou, H. Lv, M. Wang, J. Liu, W. Rhee, Y. Li, D. Kim, and Z. Wang, "A 1 mb/s 3.2–4.4 ghz reconfigurable FM-UWB transmitter in 0.18 μm CMOS," in *2011 IEEE Radio Frequency Integrated Circuits Symposium (RFIC)*, June 2011, pp. 1–4.

[9] F. Chen, Y. Li, D. Lin, H. Zhuo, W. Rhee, J. Kim, D. Kim, and Z. Wang, "A 1.14 mw 750 kb/s FM-UWB transmitter with 8-FSK subcarrier modulation," in *2013 IEEE Custom Integrated Circuits Conference (CICC)*, Sep. 2013, pp. 1–4.

[10] F. Chen, Y. Li, D. Liu, W. Rhee, J. Kim, D. Kim, and Z. Wang, "9.3 A 1 mw 1 mb/s 7.75-to-8.25 ghz chirp-UWB transceiver with low peak-power transmission and fast synchronization capability," in *Solid-State Circuits Conference Digest of Technical Papers (ISSCC), 2014 IEEE International*, Feb. 2014, pp. 162–163.

[11] Y. Zhao, Y. Dong, J. F. M. Gerrits, G. van Veenendaal, J. Long, and J. Farserotu, "A Short Range, Low Data Rate, 7.2 GHz-7.7 GHz FM-UWB Receiver Front-End," *IEEE Journal of Solid-State Circuits*, vol. 44, no. 7, pp. 1872–1882, July 2009.

[12] N. Saputra and J. Long, "A Short-Range Low Data-Rate Regenerative FM-UWB Receiver," *IEEE Transactions on Microwave Theory and Techniques*, vol. 59, no. 4, pp. 1131–1140, Apr. 2011.

[13] N. Saputra and J. R. Long, "A Fully Integrated Wideband FM Transceiver for Low Data Rate Autonomous Systems," *IEEE Journal of Solid-State Circuits*, vol. 50, no. 5, pp. 1165–1175, May 2015.

[14] K. Leentvaar and J. Flint, "The capture effect in FM receivers," *IEEE Transactions on Communications*, vol. 24, no. 5, pp. 531–539, May 1976.

[15] F. Chen, W. Zhang, W. Rhee, J. Kim, D. Kim, and Z. Wang, "A 3.8-mW 3.5-4-GHz Regenerative FM-UWB Receiver With Enhanced Linearity by Utilizing a Wideband LNA and Dual Bandpass Filters,"

IEEE Transactions on Microwave Theory and Techniques, vol. 61, no. 9, pp. 3350–3359, Sep. 2013.

[16] N. Saputra, J. R. Long, and J. J. Pekarik, "A 2.2 mW regenerative FM-UWB receiver in 65 nm CMOS," in *2010 IEEE Radio Frequency Integrated Circuits Symposium*, May 2010, pp. 193–196.

[17] B. Zhou, J. Qiao, R. He, J. Liu, W. Zhang, H. Lv, W. Rhee, Y. Li, and Z. Wang, "A Gated FM-UWB System With Data-Driven Front-End Power Control," *IEEE Transactions on Circuits and Systems I: Regular Papers*, vol. 59, no. 6, pp. 1348–1358, June 2012.

[18] N. Saputra, J. Long, and J. Pekarik, "A low-power digitally controlled wideband FM transceiver," in *2014 IEEE Radio Frequency Integrated Circuits Symposium*, June 2014, pp. 21–24.

[19] P. Nilsson, J. F. M. Gerrits, and J. Yuan, "A Low Complexity DDS IC for FM-UWB Applications," in *2007 16th IST Mobile and Wireless Communications Summit*, July 2007, pp. 1–5.

[20] N. Saputra and J. Long, "A Fully-Integrated, Short-Range, Low Data Rate FM-UWB Transmitter in 90 nm CMOS," *IEEE Journal of Solid-State Circuits*, vol. 46, no. 7, pp. 1627–1635, July 2011.

[21] B. Zhou, R. He, J. Qiao, J. Liu, W. Rhee, and Z. Wang, "A low data rate FM-UWB transmitter with-based sub-carrier modulation and quasi-continuous frequency-locked loop," in *2010 IEEE Asian Solid-State Circuits Conference*, Nov. 2010, pp. 1–4.

[22] J. Pandey and B. P. Otis, "A sub-100 μW MICS/ISM band transmitter based on injection-locking and frequency multiplication," *IEEE Journal of Solid-State Circuits*, vol. 46, no. 5, pp. 1049–1058, May 2011.

[23] C. Salazar, A. Cathelin, A. Kaiser, and J. Rabaey, "A 2.4 ghz interferer-resilient wake-up receiver using a dual-if multi-stage n-path architecture," *IEEE Journal of Solid-State Circuits*, vol. 51, no. 9, pp. 2091–2105, Sep. 2016.

[24] N. E. Roberts and D. D. Wentzloff, "A 98 nw wake-up radio for wireless body area networks," in *2012 IEEE Radio Frequency Integrated Circuits Symposium*, June 2012, pp. 373–376.