Ultra Wideband

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Recent advances in wireless communication technologies have had a transformative impact on society and have directly contributed to several economic and social aspects of daily life. Increasingly, the untethered exchange of information between devices is becoming a prime requirement for further progress, which is placing an ever greater demand on wireless bandwidth. The ultra wideband (UWB) system marks a major milestone in this progress. Since 2002, when the FCC allowed the unlicensed use of low-power, UWB radio signals in the 3.1–10.6 GHz frequency band, there has been significant synergistic advance in this technology at the circuits, architectural and communication systems levels. This technology allows for devices to communicate wirelessly, while coexisting with other users by ensuring that its power density is sufficiently low so that it is perceived as noise to other users.

UWB is expected to address existing needs for high data rate short-range communication applications between devices, such as computers and peripherals or consumer electronic devices. In the long term, it makes available spectrum to experiment with new signaling formats such as those based on very short pulses of radio-frequency (RF) energy. As such it represents an opportunity to design fundamentally different wireless systems which rely on the bandwidth of the signals to enhance the data rate or which use the available bandwidth for diverse applications such as ranging and biomedical instrumentation.

This book offers its readers a comprehensive overview of the state of the art of the physical implementation of ultra wideband transceivers. It addresses system level aspects, architectural design issues, circuit level implementation challenges as well as emerging challenges in the field. The material assumes the reader has a basic familiarity with wireless communication systems and RF integrated circuit design.

The editors thank the chapter authors for their excellent contributions and help in coordinating this book into a cohesive treatment of the subject. Many thanks go to the Springer editorial staff, in particular Katelyn Stanne and Carl Harris. We also
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Chapter 1
Ultra Wideband: Circuits, Transceivers and Systems

R. Gharpurey and P. Kinget

Abstract This chapter discusses circuit-level issues related to the design of transceivers for ultra wideband systems. Several techniques for achieving broadband gain, and their trade-offs with respect to power, performance and area are presented. An overview of circuit approaches for front-end and variable gain amplification, frequency translation, filtering, data conversion and frequency synthesis is provided. The problem of interference and coexistence in UWB systems is introduced.

1.1 Introduction

The field of wireless communications has recently witnessed the emergence of technologies characterized by channel bandwidths that are of the same order as the carrier frequencies. For example, the ultra wideband (UWB) system employs a frequency spectrum spanning 3.1–10.6 GHz, with a minimum channel bandwidth of 500 MHz. UWB is a low-power system that utilizes a power level for transmission that is below the FCC limit on spurious emissions (\(<\)–41.3 dBm/MHz) [1]. The small power density is necessary to ensure that UWB can coexist with other systems, without causing performance degradation. As a consequence the system is also relatively short distance, especially when used for high-data rate applications. It is intended for a diverse set of applications such as high-speed communications, biomedical applications and short-distance radar.

UWB represents a fundamentally different way of designing wireless systems in comparison to most current wireless communication systems that are predominantly narrowband, that is the carrier frequency employed is significantly larger than the channel bandwidth, such as, e.g., in cellular telephony. Current narrowband systems rely primarily on increasing channel SNR to enhance capacity, since they
operate in a highly spectrum-constrained environment, while UWB systems rely primarily on bandwidth. Broadband wireless redefines circuit design techniques and requirements, transceiver and synthesizer architectures and system considerations compared to narrowband systems. Additionally, given that narrowband front-end filters cannot be employed, in-band interference and coexistence with other systems become a major consideration.

This text is meant to provide the reader with an overview of the state of the art in various aspects of ultra wideband technology. The book includes description of circuit techniques, architectures and system considerations, while addressing emerging challenges in the field. System-level issues are discussed in Chapters 2 and 3, while Chapters 4–6 present implementations of various types of UWB transceivers for pulse-based and OFDM-based systems.

Chapter 2 by Foerster et al. describes system implementations that have been proposed for UWB communications. It covers issues fundamental to UWB system design, such as multipath performance, channel response, processing gain, multiuser access, implementation and link budgets, initial acquisition and narrowband interference. An overview of pulse-based and OFDM-based techniques for UWB communication systems is presented. System-level enhancements such as detect and avoid for interference mitigation are also described. The chapter relates UWB to another emerging development in the field of broadband wireless, namely cognitive radios.

Chapter 3 by Stephan ten Brink et al. discusses baseband architectures for ultra wideband communication systems based on the multiband OFDM approach. Aspects from preamble processing such as packet detection, frame synchronization and frequency offset estimation illustrate the challenges posed to reliable detection and synchronization over wideband channels. Algorithms, performance benefits and implementation costs of several next-generation high rate extensions are described in detail, including higher-order modulation as well as different multiple antenna techniques.

Chapter 4 by Razavi et al. presents an implementation of a direct-conversion UWB transceiver for MB-OFDM using the 3–5 GHz band. Three resonant networks are used at the input along with three phase-locked loops for carrier generation. Typical specifications for the analog section of an MB-OFDM transceiver are presented in this chapter.

Chapter 5 by Lee et al. describes a pulse-based UWB transceiver. The signaling is based on a 500 MHz sub-band approach utilizing the full bandwidth from 3.1 to 10.6 GHz. The chapter includes a description of the RF front-end and transmitter sections, as well as the baseband used in the design. A description of the antennas used in the test setup is also provided. Synchronization requirements and the design of a RAKE receiver for addressing multipath are presented.

The final chapter by Zheng et al. discusses the implementation of pulse-radio transceivers that use pulses with full-band coverage instead of a sub-band approach. This chapter describes the design of the RF front-ends and baseband sections used in the design and implementation of three types of impulse radios developed by the
authors. Architectural issues in such systems such as timing and synchronization are addressed in detail.

The discussion in this chapter focuses on some of the critical bottlenecks in circuits for ultra wideband systems, with an emphasis on the problem of achieving broadband gain. Most of the challenges in UWB circuit design in fact arise from the broadband nature of the designs, which necessitates much greater gain-bandwidth products than have been required of narrowband radio front-ends. As we will discuss briefly, this aspect of the system also leads to a key bottleneck arising from the potential for interference-related degradation of the system. Other challenges include the requirement for fast-hopping signal sources in multiband schemes capable of spanning the entire UWB band. Several other design issues relevant to the analog section of these transceivers are also addressed. Some of these issues are relevant to all UWB implementations, while other challenges are more system specific.

1.2 Front-End Designs for UWB Systems

The front-end of UWB transceivers is similar across different standards, such as pulse-based (e.g. [2]) and multiband approaches, and depends primarily on the full band covered by the system. In the case of pulse-based systems, the signal may be down-converted to baseband through a mixer or else a correlator-based approach may be used for detection as discussed in Chapters 5 and 6. In the multiband OFDM approach [3] a mixer is used to down-convert the incoming spectrum to the desired IF frequency or baseband in the case of direct-conversion implementations, and the signal is then filtered and quantized. The receiver chain in this case looks very similar to that employed in a narrowband system.

Regardless of the down-conversion approach used, the front-end amplifier has to have the ability to process the entire desired bandwidth. The design specifications are similar to narrowband amplifiers and include gain, noise figure, input matching, measures for linearity such as the 1 dB input compression point and intermodulation intercept points. The key difference is that these metrics have to be achieved over a broad signal bandwidth, which is of the same order as the center frequency of operation.

Depending on the implementation of the system, the approximate band covered by the LNA can vary from 3.1 to 5 GHz (low band), 6 to 10.6 GHz (high band) or 3.1 to 10.6 GHz (full band). Several approaches have been presented in literature for these designs. These can be broadly categorized into three types: designs utilizing resistive feedback and loads for broadband performance, designs using broadband input and output matching, and distributed amplifier techniques. Achieving broadband gain is a fundamental requirement in a UWB receiver; thus much of the discussion provided here also applies to the input stage of mixers as well as broadband variable gain amplifiers used in these systems. The UWB system has a strict limit
on the transmitted power density of $-41.3 \text{ dBm/MHz}$. This limits the output power requirement of the transmit amplifier to be of the order of approximately $0–3 \text{ dBm}$. The amplifier topologies discussed below are all capable of providing this level of output power. Thus much of the discussion below is relevant to the output stage at the transmitter as well.

### 1.2.1 Resistive Matching and Noise Cancellation Techniques

The use of resistively loaded amplifiers is motivated primarily by the requirement for area efficiency in short-channel CMOS technologies. While other techniques discussed later can offer much higher power efficiency and dynamic-range performance, the use of integrated inductors for tuning and matching may lead to unacceptably high area requirements.

The input amplifier in addition to providing gain also needs to be matched to the external source impedance. Two approaches that can be employed for this purpose include common-gate designs that have input impedance proportional to the inverse of the device transconductance and resistive-feedback-based designs, such as a shunt–shunt feedback topology. Negative feedback is a classical technique for increasing amplifier bandwidth [4, 5]. Since the UWB band extends up to 10.6 GHz, in order to achieve adequate gain in a single stage the gain-bandwidth product of the devices needs to be of the order of 100 GHz. It is only recently that CMOS devices with such performance have become available in the commercial space. Alternatively cascaded sections can be used to enhance the equivalent gain-bandwidth product beyond that of a single stage. However, this can lead to degradation in linearity and a loss of power efficiency.

A simple shunt–shunt resistive-feedback circuit is shown in Fig. 1.1a. This design has an input impedance of $(R_F + R_L)/(1 + g_m R_L)$, a voltage gain of $-g_m (R_L||R_F) / 2$, assuming input matching, and can be designed to provide the

![Fig. 1.1](image)

**Fig. 1.1** Basic input-matched broadband amplifiers. (a) Shunt–shunt feedback and (b) a common-gate design
desired output impedance level, by appropriate choice of the feedback resistance $R_F$. $C_C$ is a large ac-coupling capacitor.

One of the key issues in this design, and similarly in a common-gate design (Fig. 1.1b), is that the input impedance looking into the amplifier is restricted to the impedance of the source $R_s$, typically 50 Ohms. This severely restricts the flexibility in choosing the value of the transconductance of the input device. In the limit that $R_L$ tends to infinity, the input impedance of a shunt–shunt amplifier equals the inverse of the device transconductance which is thus constrained to be equal to the conductance of the source. Similarly the transconductance of the input device of a common-gate design also has to equal the conductance of the source.

A consequence of the fixed value of the input transconductance is that the noise figure of the amplifier is also determined by the input power matching requirement. The voltage gain for the case when $R_L$ tends to infinity is given by

$$A_V = \frac{1 - g_m R_F}{2} \quad (1.1)$$

It can be shown through noise analysis that the noise factor for the amplifier at low frequencies is given by

$$F = 1 + \frac{R_F}{R_s} \left( \frac{4}{(1 - g_m R_F)^2} + \gamma \frac{(1 + g_m R_F)^2}{(1 - g_m R_F)^2} \right) \quad (1.2)$$

Thus as the gain is increased by increasing the value of $R_F$, the noise factor asymptotically approaches $1 + \gamma$, where $\gamma$ is $2/3$ for long-channel devices and higher for short-channel length devices. It should be noted that this is a best-case result and in practice the noise factor will be higher, especially as frequency increases, and becomes a significant fraction of the device cut-off frequency.

The noise factor of a common-gate device at low frequencies, with its input impedance matched to the source, is given by

$$F = 1 + \gamma + 4\frac{R_s}{R_L} \quad (1.3)$$

and the gain by $g_m R_L/2$. Thus as the gain is increased by increasing the value of $R_L$, the noise factor similarly asymptotically assumes a value of $1 + \gamma$. This result also assumes that the common-gate amplifier utilizes an RF choke to connect the source to the ground. If a resistor or current source is used instead for biasing the device, the noise factor will increase above this ideal value.

Thus the noise and gain performance of a shunt–shunt feedback stage is virtually identical to that of a common-gate amplifier for high-gain conditions. To the first-order the linearity is similar as well, especially if degeneration in the source path of the shunt–shunt device is ignored. This can be appreciated by observing that the small-signal gate-to-source voltage for both amplifiers is identical. A key difference that arises at high frequencies is that the load capacitance has a very significant
impact on the input impedance in the case of shunt–shunt amplifier, while this is not so in the common-gate case.

The basic shunt–shunt feedback and common-gate amplifier topologies cannot typically be used directly in UWB front-ends, primarily due to inadequate noise performance over the desired bandwidth, as well as potentially inadequate gain-bandwidth product of the input device. The bandwidth of the amplifiers may be severely limited at high gain due to capacitive loading at the output and the resulting pole. The single-transistor topologies thus need to be enhanced to achieve the desired noise, gain, bandwidth and linearity specifications.

The gain-bandwidth product of amplifiers can be significantly increased by using inductively tuned loads, through the use of appropriate design techniques. For example, one design approach applicable in the broadband case is the use of multiple stagger-tuned stages. While well suited for enhancing electrical performance, the added area penalty may not be acceptable in short-channel processes.

A compromise between the conflicting requirements of bandwidth and area is offered by applying the shunt-peaking technique, by adding an inductor in series with the load resistor [5]. At higher frequencies, as the impedance of the load capacitance decreases, that of the series combination of resistance and inductance increases. By properly controlling the relative values of the load resistance and inductance in relation to the parasitic capacitance, a flat gain can be achieved over wider bandwidth. In fact, a bandwidth extension of as much as 70% can be achieved by use of a single inductor, in comparison to a simple shunt R–C load. The inductor does not require a high-quality factor, since it is in series with a relatively large resistor. Thus, in integrated applications, the interconnect trace used to implement the inductor can be kept relatively thin, thereby further minimizing area penalty [6]. Shunt-peaking can be used in both shunt–shunt and common-gate designs to increase the bandwidth, without leading to excessive area penalty, thus retaining the motivation behind the single-stage design.

Another effective technique for increasing the gain-bandwidth product of a single-stage amplifier is to cascade multiple stages [5]. If an amplifier has a constant gain-bandwidth product, then by using many of these stages in cascade, where each stage provides a low level of gain, an overall gain-bandwidth that is much greater than that of the single-stage amplifier can be achieved.

The design in [6] combined the cascade approach with shunt-peaking to implement a front-end LNA with a flat-gain bandwidth from 2 to 5.2 GHz, gain of 16 dB, a noise figure of 4.7–5.7 dB in the UWB band. The power dissipation in the design was 38 mW and the design was implemented in a 0.13 μm CMOS technology and measured in a low-cost BGA package. The design also provided single-ended to differential conversion. A combination of cascading and shunt-peaking was also reported in [7]. The design was employed as the front-end LNA of a 3.1–9.5 GHz UWB transceiver and provided a cascaded gain of 27 dB. It was implemented in a 90 nm CMOS process. A 5GHz CMOS LNA that employed a $g_m$-boosted cascode topology with each device in the cascode contributing almost identically to the overall voltage gain of 25 dB was reported in a 90 nm technology in [8]. The design had a 3 dB corner frequency of 8.2 GHz and a noise figure of 2 dB at 5 GHz.
The topology was inductor-less resulting in a very low area requirement and utilized dual-feedback loops. The power required for this amplifier was 42 mW.

Besides limited gain-bandwidth product, the other major design limitation of single-stage amplifiers of Fig. 1.1 is that the best-case noise figure is directly determined by the input matching requirement. We observe that the noise figure is ultimately limited by the noise generated by the input device for large values of gain. A solution for decoupling the noise and matching performance was presented in [9]. The key elements of the idea are shown in Fig. 1.2. This design utilizes a unique property of the shunt–shunt amplifier: while the amplifier provides a phase inversion for the signal path, the drain noise of the input device appears in phase at the gate node. Thus if the signal at the gate node is inverted and combined with the signal at the drain node, it is theoretically possible to cancel the drain noise arising from the input device by using the appropriate ratio in the combiner. The noise at the output is dominated by the noise of the signal combiner, which can be minimized by increasing the gain of the combiner at the cost of higher power dissipation. The input matching is still set by the transconductance of the input device, but this device does not contribute to the output noise. The implementation shown in Fig. 1.2 shows a combiner that uses the upper and bias devices of a source follower stage for generating signals with opposite polarity. This technique results in a noise factor given by [9]

\[ F = 1 + \frac{R_S}{R_F} + \frac{\gamma}{g_{m,2}} \left( \frac{1}{R_S} + \frac{3}{R_F} + 2 \frac{R_S}{R_F^2} \right) \]  \hspace{1cm} (1.4)

For large values of \( g_{m,2} \), and for \( R_F \gg R_S \), this expression asymptotically approaches

\[ F \approx 1 + \frac{R_S}{R_F} \]  \hspace{1cm} (1.5)

In effect, this technique allows the designer the flexibility to trade off power dissipation with noise performance, while not affecting the input match. An amplifier

![Fig. 1.2 Shunt–shunt amplifier with noise cancellation ([9])](attachment:image_url)
using a common-gate input can also be similarly modified to decouple the noise performance and the input matching requirement, if the source point of a common-gate device is applied to a second common-source amplifier. Such a technique was presented in [10]. The key elements of this approach are shown in Fig. 1.3a, where we assume that the source resistance is matched to the input transconductance of the common-gate device. For this condition, it can be shown that the noise contribution of the common-gate device can be nulled at the differential output observed across the drains of the common-gate and the common-source devices, $M_1$ and $M_2$, respectively, if the voltage gain of the common-gate stage from its source to its output is made equal to that of the common-source stage from its gate to its output [9]. In packaged amplifiers, the transconductance of the common-source device becomes frequency dependent due to the parasitic source impedance presented by the bond-wires of the package, which may lead to non-ideal cancellation.

The dominant noise source at the differential output is that of the common-source device. If the transconductance of the common-source device is made equal to that of the common-gate device, then the noise figure is effectively equal to that of the common-gate stage which is not a useful result. On the other hand, the requirement for noise cancellation is merely that the voltage gains of the common-gate and common-source stages be made equal, which can also be achieved by using a larger transconductance of the common-source device, with a smaller load impedance, such that the product of the two equals that within the common-gate stage. The output noise contribution of the common-source stage is given by $4kT\gamma \frac{g_{m,2} R_{L2}}{\gamma}$ or equivalently $4kT\gamma \left( \frac{g_{m,2} R_{L2}}{g_{m,2}} \right)^2$. Under the constraint that $g_{m,2} R_{L2}$ is constant, increasing $g_{m,2}$ can be seen to decrease the output noise voltage of the common-source device. As before, this implies a trade-off between noise performance and power dissipation.

An added benefit observed in this implementation is that it is an inherent single-ended to differential converter and obviates the need for an external broadband

![Fig. 1.3](image.png) Noise cancellation in common-gate stages. (a) Differential output ([10]) and (b) single-ended output ([11])
balun. The front-end amplifier in [10] combines the above approach with shunt-peaking, to achieve broadband gain of 19 dB over a bandwidth of 0.1–6.5 GHz, with a noise figure of 3 dB and a power dissipation of 12 mW in a 0.13 μm CMOS process.

The design in [11] also employs noise cancellation utilizing a common-gate input device. However, the output is combined in a single-ended manner, after a second inverting amplifier is used at the load of the common-gate device. A conceptual view of this design is shown in Fig. 1.3b, where Cc is a large coupling capacitor. The amplifier also employed shunt-peaking and achieved broadband performance from 1.2 to 11.9 GHz with an in-band gain of 9.7 dB, a noise figure in the range of 4.5–5.1 dB, an IIP3 of –6.2 dBm and was implemented in a 0.18 μm CMOS process.

A key consideration in the design of UWB front-ends is interference robustness. This is a very important issue, since given the broadband nature of UWB, the system is inherently susceptible to jammers that can arise from a multiplicity of sources, including intentional transmitters such as cellular phones and WLAN systems. A particularly challenging interferer is the UNII-band WLAN system at 5 GHz that appears in the center of the UWB bandwidth. This issue is introduced in Section 1.5 and a detailed treatment is provided in Chapter 2.

1.2.2 Broadband Input and Output-Matched Amplifiers

Another approach to broadband designs for UWB front-end amplifiers is to use high-order reactive matching networks [12, 13]. This type of design has also been employed by Zheng et al. in Chapter 6. The design of passive networks for broadband matching is a subject of classical network theory [14, 15]. It can be treated as an impedance matching or a filter-synthesis problem. In contrast to the techniques discussed in the previous section, the above techniques provide ideally lossless matching by the use of passive reactive elements. Such techniques are thus capable of providing significantly better gain and dynamic-range performance normalized to power dissipation than those discussed previously at similar channel lengths, but at the expense of requiring multiple integrated inductors, with the associated area penalty.

The elements of the matching network for various transfer functions, such as Tchebycheff and Butterworth type, can be easily determined from tables, [14, 15], or by determining the roots of the polynomials through numerical simulation. The matching networks are typically tabulated for low-pass prototypes, but can easily be transformed to implement bandpass, band-reject and high-pass type responses. For example, to implement a bandpass response each inductor in the low-pass network is replaced by a series LC, while each capacitor by a shunt LC network. Impedance transformation, that is matching a source resistance to a lower or higher value of the input resistance, is also possible through the use of certain matching networks, for example, by use of even-order Tchebycheff polynomials.
A 3rd-order bandpass network for broadband input matching of the gate of a common-source device is shown in Fig. 1.4 [12]. The band-pass nature of the network can be deduced by observing that the network acts as an attenuator near DC as well as in the high-frequency limit. The matching network transforms the impedance seen looking into the input of the device to the source resistance. Thus the real part of the input impedance is transformed to this value of resistance, and the reactive part is ideally canceled over a broad spectrum. Parasitic inductance of packages can be easily absorbed in such designs. The technique of synthesizing a lossless real part in the input impedance by utilizing the bond-wire inductance of the package can be employed in these matching networks as well, since to the first order the real part to the input impedance is given by \( \omega_0 L_s \), where \( L_s \) is the degeneration inductance. This is independent of frequency, and thus can be used as the load resistance of the input matching network [12, 13].

The design of [13] employed a SiGe BiCMOS technology and provided a power gain of 21 dB, with a noise figure in the range of 2.5–4.2 dB from 3 to 10 GHz, and an IIP3 of –1 dBm with a power dissipation of 30 mW. The design presented in [12] was implemented in a 0.18 \( \mu \)m CMOS process, and reported a gain of 9.3 dB, a noise figure of approximately 4–8 dB across the frequency band from 3 to 10 GHz and an IIP3 of –6.7 dBm with a power dissipation of 9 mW.

Simultaneous noise and power matching can be challenging in such networks. It can be shown through analysis that the optimal noise reactance at the input of an inductively degenerated device is approximately \( -1/j \omega C_{in} \). Thus a broadband matching network that provides a conjugate match to the source impedance will provide the optimal noise reactance to the first order. The real part of the optimal noise impedance, however, is also frequency dependent. Thus the matching network, which provides a constant real part looking into the source, may not match the noise resistance over all desired frequency bands.

These networks provide voltage gain from the source to the gate of the MOSFET, which is a consequence of the typically higher impedance of the MOS input compared to the source resistance. In fact a substantial portion of the gain may be achieved through the passive matching network. This can degrade the compression point of the active device, as in the narrowband case.

Ultimately the limits on matching are set by the Bode–Fano criterion [15] which places a cumulative limit on the quality of matching over all frequencies as a function of the quality factor of the load. Consider an inductively degenerated common-source device. The input impedance in this case is represented by the series
combination of the input capacitance and a series resistance of value $\omega_0 L_q$. For this impedance, if $\Gamma(\omega)$ is the frequency-dependent reflection coefficient looking into the series $R$–$C$ load, then the Bode–Fano criterion places the following bound on the input match:

$$\int_0^\infty \ln \left( \frac{1}{|\Gamma(\omega)|} \right) d\omega \leq \pi \omega_0^2 RC$$

(1.6)

In the above expression, $\omega_0$ is the center frequency of operation. In an ideal bandpass match, $\Gamma(\omega)$ is set to 1 outside the bandwidth of interest and is less than 1 within the band of interest, implying power delivery to the load. If the desired band is $B$ and we design for a constant in-band reflection constant $\Gamma_0$, then the above expression places a lower bound on the value of $\Gamma_0$ to be

$$\Gamma_0 \geq e^{-\left(\frac{-\pi \omega_0^2 RC}{B}\right)}$$

(1.7)

For a constant RC product and center frequency, the Bode–Fano criterion places a limit on the best possible constant match that can be achieved in practice, assuming an ideal passive matching network. Alternatively, instead of achieving a constant match over all frequencies, it is possible to achieve a very good match over narrow frequency bands, at the expense of a worse match at other frequencies within the band of interest [15].

A key challenge in the use of such designs, especially in comparison to the broadband resistively matched techniques of the prior section, is their area requirement which is typically in the range of a square millimeter or more. This area requirement can be expensive in short-channel CMOS processes. Many designs have reported measurements in an on-wafer probe environment, rather than in a package, which can have significant impact on the performance. On the other hand, the input and ground path package inductance can be absorbed into the matching network with relative ease using these techniques.

### 1.2.3 Distributed Amplification

Distributed amplifiers also provide broadband input matching; however, the approach taken is different compared to the broadband matching technique considered earlier. In this type of amplifier (Fig. 1.5), a single large device is divided into multiple smaller sections, each with smaller unit input and output capacitance. The capacitance of the unit devices is absorbed into a lumped approximation of a broadband transmission line, at both the input and the output, by using discrete inductors. The input transmission line is terminated in a matched resistive load $R_0$, and similarly the output transmission line is also terminated by a resistive matched load $R_0$ on one end (node D) and the output load $R_L$ on the other (node C). In
this way, the input source and the output load see a broadband matched termination looking into the input and the output of the amplifier, respectively. Following the discussion from [16], we assume that the signal is shifted in phase by $\phi$ at the input and $\phi$ on the output lines by each subsequent device. For identical phase delays on the input and output transmission lines, the broadband power gain from the input node A to the output node C can be shown to be

$$G_F = \frac{n^2 g_m^2 R_L R_S}{4}$$  \hspace{1cm} (1.8)$$

where $n$ is the total number of stages in the amplifier, $g_m$ is the transconductance of each stage, $R_L$ is the load resistance and $R_S$ is the source resistance. The voltage gain of the amplifier is limited by the impedance of the loads employed at the input and output terminations.

The input termination at node B is defined by the source impedance, but the terminations at C and D are design parameters in integrated applications. Since the output termination needs to be broadband, the upper limit on the output load is typically set by the capacitance that appears in shunt with the load, for example the capacitance at the input of the down-conversion mixer driven by the amplifier. For an operational bandwidth of 10 GHz, for a 100 $\Omega$ output load, the maximum capacitance at the output is of the order of 100 fF or less, which can be easily exceeded for typical input devices of the down-conversion mixers and interconnect parasitics. In
such cases, the output resistance may need to be kept relatively small, to extend the bandwidth. Thus the voltage gain of these amplifiers is typically limited. Enhancement of voltage gain can be achieved through the use of matrix amplifiers, which are the distributed analogues of cascaded single-stage amplifiers, but these are likely to be too power-hungry for UWB applications, in addition to requiring excessively large die area.

The power delivered to the output transmission line can flow towards both the output load and the termination at node D. The gain from the input to the drain termination is referred to as the reverse gain of the amplifier and is given by [16]

\[
G_R = \frac{g_m^2 R_L R_S}{4} \left( \frac{\sin (n (\varphi + \Phi)/2)}{\sin (\varphi + \Phi)/2} \right)^2
\] (1.9)

The above gain is significantly reduced within the band of interest due to cancellation of the drain current towards the termination, as a consequence of destructive interference. Proper phase combining ensures that the desired signal is forced to propagate towards the output, where the drain currents add constructively. The reduction of reverse gain within the band of interest has very interesting implications for noise. Since the noise of the input line termination resistor is scaled by the reverse gain when it reaches the output load at C, we see that this noise is reduced significantly. Thus the input line termination resistor provides an input power match, but does not provide substantial noise at the output. This is a key advantage provided by the distributed topology, similar to the decoupling of noise and impedance matching in the noise-cancellation stages discussed earlier.

The noise figure of the distributed amplifier can approach that of a narrowband noise-matched common-source amplifier, although over a much broader bandwidth. This behavior is seen in recent examples of distributed amplifiers that show a smaller variation in noise figure across the frequency bandwidth compared to those employing multi-section broadband matching.

The gain-bandwidth product of distributed amplifiers is limited primarily by the input and output transmission lines. Since these lines are discrete, their characteristic impedance changes with frequency. As shown in [17] this cut-off frequency is given by \(1/\pi R C_g\), where \(R\) is the input resistance and \(C_g\) is the gate capacitance of an individual device. For an \(n\)-stage distributed amplifier, the gain-bandwidth product to the first order is then given by \(n g_m/2\pi C_g\), where \(g_m\) is the transconductance of a single device.

The above expression succinctly captures the advantage provided by distributed amplification. A single device with transconductance \(g_m\) and capacitance \(C_g\) has a gain-bandwidth product of \(g_m/2\pi C_g (\sim f_t)\). In a distributed amplifier, the gain-bandwidth product scales linearly with the number of unit devices \(n\).

The above is a first-order approximation, since it does not take into account losses in the input and output lines [18]. As the number of sections increases, losses on the input line progressively attenuate the signal level at the devices further away from the input. Similarly, the current from the devices closest to the source suffers
increasing attenuation before it reaches the output load. Consequently there exists an optimal number of stages at which the gain-bandwidth is maximized.

Ignoring the losses in the input and output lines and using the ideal expression for the gain-bandwidth product, we can derive interesting insights into the optimal biasing point of the device used in a distributed amplifier, utilizing a figure of merit given by the ratio of the gain-bandwidth product to the total bias current used in the amplifier. For a MOS device, the $g_m/I$ ratio is high in weak- to moderate inversion, that is for sub-threshold operation. However, the cut-off frequency is also lower than the strong-inversion case. By using $n$ devices in weak inversion, the effective gain-bandwidth can be enhanced over that of a single device, while retaining the higher overall $g_m/I$ ratio of the amplifier. The above reasoning was described in [19] and used to implement a CMOS distributed amplifier with moderate inversion device operation. The amplifier was implemented in a 0.18 $\mu$m CMOS process, demonstrated a gain of 8 dB from 0.04 to 6.2 GHz, with a noise figure of 4.2–6.2 dB and an IIP3 of 3 dBm at a power level of 9 mW.

An interesting feature of distributed amplifiers is that since there are no high-impedance nodes within the amplifier, the voltage levels at the input and the output are relatively small, for example in comparison to approaches such as those using multi-section LC matching. Consequently, distributed amplifiers also happen to exhibit high output 1 dB compression point, compared to any of the topologies discussed to this point. Thus these amplifiers are also very well suited for the output buffers used in UWB transmitters. The transmitter output stages in UWB need relatively modest output 1 dB compression points of the order of 2–3 dBm. This has been demonstrated in DAs even for low-voltage CMOS technologies, e.g., [19] and [20], which reported a DA with 10.6 dB gain from 0.5 to 14 GHz bandwidth, a noise figure of 3.4–5.4 dB and an output 1 dB gain-compression point of 10 dBm at a power dissipation of 52 mW, in a 0.18 $\mu$m CMOS technology.

Distributed amplifiers thus offer an effective combination of broadband gain, excellent broadband noise performance as well as very good output compression point. There are two major issues to be considered, however, before choosing a distributed amplifier topology, the first being area. Since both the input and the output require several inductors, the area requirement can be high, especially compared to some of the earlier inductor-less approaches, such as those utilizing broadband feedback and noise cancellation. A second major issue with distributed amplifiers is that the input impedance of the devices needs to have relatively high Q. Source degeneration inductance, such as that arising from package inductance translates into an input resistance in series with the gate capacitance. This can significantly degrade performance of the distributed amplifier. In a similar manner, any bond-wire inductance in series with the input and the output can cause significant deviations from the broadband characteristic. Thus if a DA is to be utilized in a practical application, the bond-wire inductance associated with the input and ground paths has to be minimized. On the ground node, this can be accomplished in principle by using several bond-wires in parallel or advanced packaging technologies with low series inductance such as flip chip. While this can prove to be uneconomical in terms of utilization of bond-pads, the desired electrical performance can be achieved. Another approach to reduce
the impact of the ground inductance is to use a differential amplifier topology, although this requires the use of a broadband balun externally, the design of which is non-trivial.

The bond-wire inductance on the input can be absorbed into the input line if the interstage inductance in the amplifier is larger than the bond-wire inductance; however, this may not always be the case, especially in high-frequency designs. The bond-wire inductance can be decreased by using multiple bond-wires in parallel but this has the undesirable side effect of simultaneously increasing the bond-pad capacitance that loads the input node.

It is perhaps due to the above practical considerations that most reported UWB transceivers have in fact not utilized DA topologies. On the other hand, the topology continues to be of significant interest, and if a low-cost solution is found to the packaging parasitics, will doubtless be well utilized, owing to its excellent dynamic-range performance per unit current.

1.3 UWB IF: Mixers, Variable Gain Amplifiers, Filters and A–D Converters

This section outlines the circuits that comprise the IF section of a UWB transceiver. Much of this portion of the transceiver is system specific, unlike the earlier designs for front-end amplification techniques that were fairly independent of the system itself. As such the discussion is brief, and for detailed insights references are provided to the appropriate chapters in the text. It should be mentioned that the broadband design issues discussed earlier are significant in some instances in the IF section as well, for example for broadband variable gain amplification, and at the input stage of mixers. In other cases, the designs can borrow directly from techniques employed in traditional narrowband receivers while redesigning for higher bandwidths although with relatively limited dynamic range. Since the baseband in the system extends to 250 MHz in the multiband OFDM approach and over 2 GHz in many of the pulse-based schemes, significant modification is required, as they are especially sensitive to parasitics.

1.3.1 Mixer Design for UWB

Mixers are required primarily for multiband systems, such as the MB-OFDM approach, since the system requires down-conversion to baseband or an IF. For the MB-OFDM system, the channel bandwidth at baseband is of the order of 250 MHz, which is much smaller than the RF input frequency that can range from 3.1 to 10.6 GHz. Therefore mixer techniques typically used in narrowband wireless receivers can be employed without significant modification at the IF output. The input transconductor needs to be capable of operating over the entire frequency range, and therefore any of the earlier mentioned broadband amplifiers can in principle be used
to drive a passive switch-based down-converter or a current-commutating switching stage. A switch-based down-converter requires large swing and does not have internal nodes with low impedance. It is thus preferable to use a current-commutator design instead. Given the broad bandwidth of the channels, even in the case of multiband systems, a usually important disadvantage of MOS current-commutating mixers, namely in-band flicker noise, is not a significant consideration in this case.

The mixers used in pulse-based UWB systems are usually employed as correlators, where they can be used as part of the receiver-matched filter. The circuit operation is similar to the above case and therefore not repeated here. The reader is referred to Chapters 5 and 6 for implementations of such designs.

1.3.2 Variable Gain Amplifiers

In a typical pulse-based approach to UWB, the UWB band is down-converted to baseband for an aggregate baseband channel of approximately 1 GHz, for the lower band of 3–5 GHz, and 2 GHz for the upper band of 6–10 GHz. Several pulse-based approaches limit the input signal and therefore do not require significant levels of gain control. In the MB-OFDM approach to UWB, the channel bandwidth is relatively much narrower, of the order of 250 MHz. The gain variation requirement can be of the order of 40–50 dB, with a peak-gain requirement of the same order, depending on the specific architecture. A peak gain of 50 dB with a bandwidth of 250 MHz corresponds to a gain-bandwidth of 75 GHz, which is similar to that required in the front-end LNA. Thus VGA designs for MB-OFDM UWB can be challenging.

It is typically not possible to achieve the high peak-gain requirement in one stage of amplification, while simultaneously achieving the desired bandwidth. An effective technique for broadband variable gain implementation is to use cascaded stages of amplifiers [5] to achieve the desired bandwidth. This also improves the attenuation of out-of-band interferers, since cascaded stages have a sharper high-frequency roll-off for the same effective gain-bandwidth product, compared to a single-stage amplifier.

VGA functionality has been demonstrated in recently reported UWB transceivers using multiple techniques. Some degree of gain switching is usually required in the front-end LNA (see Chapter 4), especially to accommodate large interferers. If broadband VGA functionality is required over the full UWB band, or a significant fraction of it, then distributed amplification offers an excellent approach to cascaded stages. As discussed earlier, these stages can maintain low input- and output impedance and can be used to provide broadband matching. The design from [19] was demonstrated to have variable gain from –10 dB to 8 dB, for a bandwidth from DC to 6.2 GHz, where the gain flatness and matching performance were maintained over the gain tuning range.

The majority of VGA functionality is usually implemented in baseband, where gain steps can be controlled more accurately than at RF, and the gain of the VGA can
be distributed between filter stages, to maximize the overall dynamic range of the baseband stage. In [21], the authors employ a passive switched attenuator between filter stages. In [22], the VGA is implemented as a switched transconductor and a variable gain buffer, the first circuit before the filter, and the second after the filter. A VGA based on a Gilbert-cell multiplier is employed in [23]. As discussed in Chapter 4 a key challenge in the MB-OFDM system is handling DC offsets within the VGA chain, since the DC offset in response to a gain step needs to settle very rapidly, in fact within nanoseconds.

1.3.3 Filter Design for UWB

Pulse-based UWB systems employ correlators and matched filters for spectral shaping, band limiting and detecting signal energy. Such implementations are presented in detail in Chapters 5 and 6.

Multiband OFDM implementations require strict bandwidth controls. Typically 4–5th order filter topologies have been employed for channel selection and filtering of adjacent channel interference in these implementations. For example in [22], the authors present a 5th-order Gm-C filter with a corner frequency of 240 MHz in a 0.13-μm CMOS technology with a power dissipation of 24 mW and input referred noise of 7.7 nV/sqrt(Hz). The gain of the filter is 48 dB which includes the gain of the VGA, embedded within the design. A 5th-order Tchebycheff design is employed in [21] with a gain range from 16 to 46 dB and a tunable bandwidth from 232 to 254 MHz. A 4th-order Tchebycheff design is described in [24] with a nominal corner frequency of 259 MHz. Two cascaded 3rd-order elliptic filters are employed in [7]. A passive LC 5th-order elliptic low-pass channel select filter is employed in [23].

1.3.4 Data Conversion for UWB

In OFDM-based approaches, the baseband ADC in the receiver can be a challenging design, more so than the DAC in the transmitter. Given that for the MB-OFDM approach, the baseband bandwidth is of the order of 264 MHz, the sampling rate required in the ADC needs to be at least 528 MHz. On the other hand, the resolution requirement is limited [3], of the order of 4–5 bits.

Given the requirement for a modest resolution and high sampling rate, flash ADCs are a suitable topology for this implementation. A recent example includes a 6-bit flash converter in a 0.13 μm technology that was demonstrated in [25] with a power consumption of 160 mW. A 5-bit flash at 1 GHz sampling rate also in a 0.13 μm CMOS technology with a power dissipation of 46 mW is reported in [26], while a 4-bit flash at 1.2 GHz sampling rate was reported in [27] with a power dissipation of 86 mW.
Substantially lower power requirement has been reported by the use of successive approximation (SAR) approaches. A 5-bit time-interleaved SAR requiring a power consumption of 6 mW was demonstrated in [28] using a 65 nm CMOS technology. Another 5-bit SAR was also demonstrated in a 0.18 μm CMOS technology, with power requirement of 7.5 mW [29].

DAC designs for MB-OFDM UWB are relatively easier, owing to the limited resolution. Current-steering DACs can easily be employed for this purpose and are not detailed here. DAC designs for the transmitter sections of pulse-based UWB systems on the other hand pose much more design challenge, owing to their broadband nature. Additionally these are usually tailored to specific system implementations and often are integral to the final spectrum shaping functionality. Examples and design details can be found in Chapters 5 and 6.

1.4 Frequency Synthesis in UWB

Pulse-based schemes in UWB can employ correlators for detecting the energy in the transmitted signals and, thus in some implementations, may not require an explicit frequency synthesizer. For utilization of the upper band (6–10 GHz), it may be necessary to down-convert the band to baseband and thus a fixed carrier at mid-band can be employed [30].

Multiband solutions including MB-OFDM employ relatively narrower channels and cycle through these bands in order to utilize the entire UWB spectrum. The time required to hop and settle the frequency between successive steps needs to be minimized and is of the order of 2 ns in the MB-OFDM system [3]. Phase-locked loop-based synthesizers, such as those utilizing the integer-N architecture, would be challenging for this purpose due to settling time limitations, as well as other issues such as an excessively high oscillator center frequency, of the order of the lowest common multiple of all center frequencies, and the need for extremely fast dividers running at several decades of GHz. A solution proposed for this problem employed an oscillator with a fixed center frequency, where the output frequency was divided and combined through single-sideband mixing to generate the desired channel frequency [3, 31]. The original proposal assumed a center frequency of 4224 MHz, which was divided by 4 to provide 1056 MHz and further divided by 4 to provide 264 MHz. By using single-sideband mixing, these bands could be combined to provide center frequencies at 3432, 3960 and 4488 MHz. Band-selection in single-sideband mixing is accomplished by merely changing the polarity of the single-sideband combiner. Since this is an open-loop operation, it is inherently fast, as no feedback-related settling dynamics are involved and the only limitations on speed arise from the load parasitics at the output of the combiner.

Various solutions have been demonstrated in IC implementations that use an open-loop approach. An abstraction of the basic technique is shown in Fig. 1.6, where the output frequency is derived by linearly combining the outputs of two cascaded frequency dividers. Examples include a 14-band synthesizer [32] demon-
strated in a 0.18 μm CMOS process that employed two PLLs with outputs at 6336 and 3960 MHz and their divided outputs; a two PLL solution for the lower band [21] that was demonstrated in a RF-BiCMOS process; and a 7-band solution [33] spanning 3–8 GHz that was implemented in a 0.18 μm CMOS process. All the above solutions demonstrate band switching speeds of the order of nanoseconds. An approach by [34] utilized direct digital synthesis (DDS) for implementation of one of the bands for a 3–5 GHz UWB implementation. DDS technologies promise to be an exciting alternative to mixer-based approaches, especially with short-channel technologies below 90 nm, as they can be used to provide outputs with low spurious levels.

1.5 Interference and UWB

Isolation between users of different systems in narrowband wireless systems is achieved largely by spectral separation. Cellular standards such as 3G-WCDMA, for example, employ an exclusive band for operation. Broadband wireless schemes such as UWB on the other hand receive and transmit signals over vast portions of the spectrum. Thus coexistence and interference of broadband wireless systems with other systems become critically important. The wide bandwidth also substantially increases the scope for interference in the front-end due to substrate- and package-coupling. The potential for in-band interference can have a significant impact on the design of the RF and analog section of the transceiver. Without the use of specialized techniques, the front-end is exposed to all large jammers that can exist in the system, and this can place correspondingly large linearity requirement in the front-end. A brief overview of interference and coexistence issues is presented below. The reader is referred to Chapter 2 for detailed analyses and insights into this issue.

1.5.1 External Interference and Coexistence

UWB transmissions can act as interferers to other systems. For example, the allowed UWB power-spectral density in the 3G-WCDMA band is –51.3 dBm/MHz. For a single channel of WCDMA, which has a double-sided signal bandwidth of 3.84 MHz, this corresponds to –45 dBm of integrated power. The sensitivity of a WCDMA receiver is approximately –99 dBm, including the processing gain allowed in the signal. Thus the allowed UWB transmission in this band needs to be attenuated by a factor of approximately 60 dB in order to minimize the impact
on WCDMA [31]. With an omni-directional antenna, a combination of transmit filtering and frequency separation is required in order to ensure coexistence. Similar considerations apply for co-existence with systems such as IEEE 802.11b, 802.11a and 802.15.4.

UWB receivers themselves can be the victims of large in-band radiators from other systems. UNII-band transmitters in the 5.1–5.85 GHz band are major sources of interference for UWB. Many UWB frequency plans, including the multiband OFDM approach and certain pulse-based implementations have an option to exclude the UNII-band altogether. It is interesting to note that if two UNII-band systems are used simultaneously for transmission (e.g., at 5.3 and 5.8 GHz) in-band 3rd-order intermodulation can be a design issue even in UWB systems that avoid the UNII bands. Interferers that can lead to in-band IM3 products and the UWB bands that they impact are identified in [35].

Spurious radiation introduced by other commercial systems can also limit the performance of UWB. For example, the specifications for WCDMA mobile transmitters allow for a spur level as high as –30 dBm in a 1 MHz bandwidth from 1 to 12.75 GHz, which can be within the channel bandwidth of a UWB system [31]. An emerging system that can have significant coexistence and interference issues with UWB is WiMAX. A detailed analysis of these issues, with simulation results indicating the potential degradation in UWB, is described in Chapter 2.

To minimize degradation caused by such an in-band spur, the analog dynamic range will need to be sufficiently large. One technique to alleviate this requirement is to increase the resolution in the baseband ADCs. This can be power hungry, since the converters typically run at high sampling rate (> 500 Ms/s). Further the bit-width of at least a part of the baseband digital path would need to be increased, which will also increase overall system power requirement. If the dynamic range of the receiver is insufficient, for example if the analog section saturates in the presence of an in-band spur, it will result in the loss of useful information in part of the bandwidth in a sub-band approach. If the modulation uses the entire band, a large in-band spur can pose a significant challenge. Saturation of the signal chain in this case will impact the entire band.

Regardless of the modulation type and bandwidth, sophisticated interference mitigation techniques at the circuit and system level are required to ensure the robustness of UWB systems. Circuit techniques are reported in [35] and [36], which address the issue of mitigation of interference, caused by UNII-band WLAN systems, through introduction of a notch response in the transfer function of the front-end LNA. A multi-antenna receiver is employed by [24] to use linear cancellation of interferers by combining the outputs of two receive-paths with the required relative phase difference. While these schemes rely on attenuation of interference, another approach is to employ techniques based on interference detection. These are especially useful in multiband schemes, where if an interferer is sensed within a specific band the use of that band can be avoided altogether. This requires a scheme to detect interferers across the entire UWB band [37]. This approach is based on the observation that an auxiliary receiver that is specifically designed for detecting interferers has a relaxed sensitivity requirement, since it is not intended to detect small signals.
In theory, therefore, the dynamic range can be split between the auxiliary receiver, which detects the large signals, and the main path receiver, which can avoid the large signals, and thus has a reduced linearity requirement. The dynamic range is thus relaxed in both the main and auxiliary receiver paths.

1.5.2 Interference Due to Circuit Activity

Low-cost, highly integrated solutions are critical to the adoption of broadband wireless as a viable technology for commercial high-speed communication applications. It can be expected that UWB solutions featuring increasingly higher levels of integration will appear in the near future. Amongst the most significant unknowns at this time in the development of highly integrated UWB systems is the potential coupling of interfering signals generated by on-chip circuit activity, through substrate and package parasitics. The scope for such coupling is significant due to the wide bandwidth and can prove to be a bottleneck for highly integrated implementations. UWB systems can require large digital circuits of the order of hundred thousand gates or more, such as FFT cores that operate at several hundreds of MHz. It is possible that in a highly integrated UWB system, the sensitivity of the system is limited not by the thermal noise at the input, but by the above signal coupling. Spectral domain isolation, e.g., spur planning, is very often exploited in narrowband systems to maximize integration level, but will be difficult to employ in UWB. The ability to estimate the level of self-induced noise in potential implementations will be a key requirement for future broadband standards and integrated architectural solutions. A discussion of this can be found in [38].

1.6 Summary

Ultra wideband represents a fundamentally different way of implementing wireless communication systems. The two key attributes that distinguish this system include the broadband nature of the system, where the channel bandwidth is a significant fraction of the carrier frequency; and the approach used for coexistence with other systems, which relies on low power-spectral density for transmission and reception, rather than strict isolation in the frequency band. These characteristics have necessitated the development of new techniques at the circuit, architecture and system levels. Thus it promises to be an exciting space for innovation in wireless technologies. The following chapters in the text are representative of various aspects of this innovation.

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Chapter 2
High-Rate UWB System Design Considerations

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Abstract The ability to optimally exploit the 7.5 GHz of newly created unlicensed spectrum for UWB technology depends on addressing a number of challenging system design issues. This chapter provides an overview of many of these issues and some technical trade-offs and comparisons with different system designs. Some of the challenges include dealing with multipath propagation, energy capture, narrowband interference, rapid synchronization, and varying regulatory rules throughout the world, just to name a few.

2.1 Introduction

According to the rules set forth by the FCC on February 14, 2002, ultrawideband (UWB) systems are defined as systems that occupy more than 20% of a center...
frequency or more than 500 MHz bandwidth. For communications systems, the available spectrum is 7.5 GHz, from 3.1 to 10.6 GHz, with slight differences in the spectral mask for indoor and handheld devices. So, from a high-level perspective, this looks like a tremendous opportunity if one can figure out how to best, and in a cost effective manner, exploit this newly available bandwidth. In order to optimally exploit this available bandwidth, it is important to understand the various system design and implementation trade-offs when it comes to dealing with multipath, energy capture, narrowband interference, implementation complexities, and different regulatory constraints. This chapter primarily investigates the potential for UWB technology to be used for very high-throughput, short-range applications like high-speed cable replacement (wireless USB), video distribution within the room, and fast image downloads from a camera to a wireless kiosk, for example. However, there are also a number of other uses of the technology that are currently being developed. These include low-rate, low-power sensors; inventory tracking and cataloging devices; building material analysis; and radar and position location-based applications, just to name a few. Many of these functions would also be beneficial to high-rate devices as well, but are not covered here. The ability for a single UWB physical layer solution to exploit high-rate, low-power, and accurate positioning capabilities of the technology could result in some interesting future capabilities.

This chapter is organized as follows. First, a brief introduction to UWB technology and the trends which have led to the development of the first industry standard is presented in Section 2.2. Section 2.3 covers a number of system design considerations and trade-offs to be taken into account when developing a high-rate UWB system, including issues related to multipath, energy capture, processing gain and spectral flatness, multi-user access, implementation, link budgets, initial acquisition, and narrowband interference. Section 2.4 provides an update on the current regulatory status for UWB both inside and outside the United States and introduces a relatively new concept of “detect and avoid” (DAA) which will likely be needed in order to more efficiently share the available spectrum with other users. Finally, future possibilities including a link with cognitive radios and conclusions are provided in Sections 2.5 and 2.6, respectively.

2.2 Brief History

2.2.1 The Link to Early Wireless

UWB technology is as old as radio itself. The earliest transmitters of Bose\textsuperscript{1} and Marconi,\textsuperscript{2} circa 1895, used spark gap technology that generated radio waves across a multi-GHz spectrum in a largely uncontrolled manner.

\textsuperscript{2} IEEE History Center, URL http://www.ieee.org/organizations/history_center/milestones_photos/swiss_marconi.html.
Over the next 25 years, radio technologists sought methods to allow more systems to share spectrum on a non-interfering basis. Motorized spark generators and LC tank circuits limited the bandwidth of spark-based signals and helped control center frequencies. With DeForest’s invention of the vacuum tube triode, circa 1906, it became possible to transmit very narrowband signals at a frequency of one’s choosing. As a result, spark technology largely vanished by the 1920s.

Thanks to vacuum tube technology, it also became possible to regulate wireless on a spectrum-allocation basis. In the United States, the Federal Communications Commission was chartered to do just that by the Communications Act of 1934.

For over 70 years, reserving portions of the spectrum for specific purposes has been an effective way to limit interference. But that approach suffers from the disadvantage that at any given time, large portions of the radio spectrum often go unused by anyone. Over the years, the FCC has used various sharing mechanisms to mitigate this inefficiency, including such concepts as primary, secondary, coequal, and other licensing constructs. UWB is just the latest attempt by the FCC to improve spectrum use.

2.2.2 Ultrawideband Reemerges

In the mid-1960s, interest in pulse-based broad-spectrum radio waveforms reemerged, growing largely out of radar technology. Most of the work was carried out under classified U.S. Government programs. However, beginning in 1994, much of the work became non-classified, and a group of pioneers began exploring commercial as well as government/defense applications of pulse-based UWB. In 1998, working with those pioneers, farsighted engineers at the FCC Office of Engineering and Technology launched a formal FCC Notice of Inquiry, ET Docket No. 98–153, proposing the use of UWB systems on an unlicensed basis.

The Notice of Inquiry was controversial, to say the least. The FCC was proposing to allow unlicensed operation across broad swaths of the radio spectrum, most of which had already been licensed to others. Rather than relying on now-traditional frequency segregation to avoid interference, the FCC proposed an average power spectral density limit of $-41.3 \text{ dBm/Hz}$. This was and is the same limit already assigned in Part 15 of the FCC rules to devices such as hair dryers, electric drills, laptop computers, and other unintentional radiators. Despite these low emission

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4 Fontana, Robert, “A Brief History of UWB Communications”, www.multispectral.com, see link “UWB History”, “Papers on UWB” and other links.
6 http://www.fcc.gov/oet/info/rules/.
limits, the wide spectrum made it possible to transmit hundreds of megabits, or even gigabits per second over short distances with very low power.

After more than 900 filed comments on ET 98-153, the FCC released a formal Report and Order in April of 2002\(^7\) specifying the rules under which UWB systems could operate. Among those rules is a frequency mask (see Table 2.1) specifying the maximum power spectral density for UWB transmitters as a function of frequency. Interest focused quickly on the 3–5 GHz portion of the spectrum where propagation losses were lowest, CMOS silicon performance was best, and interference with WLAN UNII systems (5150–5350 MHz) could be avoided.

### 2.2.3 Ultrawideband and Standards: Challenges and Eventual Path(s) to Convergence

With UWB approved for use in the United States, IEEE Task Group 802.15.3a began working on an alternate physical layer (PHY) technology to support the 802.15.3 Personal Area Network (PAN) protocol. A list of requirements for the PHY was developed\(^8\), among them was the requirement to support at least 110 Mbps at a distance of 10 m. Among the applications foreseen were high-speed “synch-and-go” and “download-and-go” file transfers among PCs, laptops, digital cameras, portable media players, cellphones, and other portable devices. Also foreseen were high-definition video streaming from portable devices to nearby displays. Virtually, all the 26 proposals appearing before the IEEE group were focused on UWB because of its high-rate and low-power possibilities.

Most of the initial UWB proposals were traditional impulse-based designs. That is, the signals consisted of a stream of short-duration (approximately 1 μs) pulses. Depending on the proposal, these pulses were modulated by polarity, amplitude, time position, or other characteristics. Impulse-based proposals offered simple, efficient transmitter designs with pulses formed by pin diodes or other

<table>
<thead>
<tr>
<th>Table 2.1 FCC spectral mask for indoor devices</th>
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</thead>
<tbody>
<tr>
<td>Frequency (MHz)</td>
</tr>
<tr>
<td>-----------------</td>
</tr>
<tr>
<td>960–1,610</td>
</tr>
<tr>
<td>1,610–1,990</td>
</tr>
<tr>
<td>1,990–3,100</td>
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<tr>
<td>3,100–10,600</td>
</tr>
<tr>
<td>Above 10,600</td>
</tr>
</tbody>
</table>

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devices – a modern-day equivalent of the old spark systems. Furthermore, some of these proposals used simple 1-bit ADCs in the receiver with chip rates proportional to the data bandwidth being transmitted. One of the proposals, later known as Direct-Sequence UWB (DS-UWB), combined a high-chip-rate impulse-based approach with orthogonal coding.

A second cluster of UWB proposals sought to strike a balance between the reduced implementation complexity of lower instantaneous occupied bandwidth and the performance advantages of increased overall bandwidth, based on the “sweet spot” of CMOS process technology in the near future. These proposals also foresaw a future need to allow finer control over the transmitted UWB spectrum, so emissions could be suppressed over selected frequencies in order to protect other narrowband systems or to adapt to local regulations in countries outside the United States. One such approach, called “Multiband UWB” was based on pulses with longer duration (3–5 μs), resulting in occupied bandwidths of approximately 500 MHz, as shown in Fig. 2.1. By “hopping” the center frequencies of three or four segments, most of the 3–5 GHz spectrum could be occupied, allowing an average transmitted power of approximately \(-41.3 \text{ dBm/MHz} + 10 \log_{10}(1700 \text{ MHz}) = -9 \text{ dBm}\). To obtain finer control over the transmit spectrum, one or more of these bands could be dropped.

Multiband UWB turned out to be short lived. A third proposal, dubbed Multiband OFDM (MB-OFDM),\(^9\) combined multiband hopping with OFDM. This turned out to be a superior solution, offering the best combination of implementation complexity and performance, and the various multiband UWB proponents converged to this proposal. After multiple rounds of comments and updates, the MB-OFDM proposal became what is today WiMedia/ECMA-368. The details of this specification are given in the following sections.

---

\(^{9}\) IEEE P802.15-03/268r1, October, 2003.
Another proposal, described above as Direct Sequence Ultrawideband\(^{10}\) (DS-UWB), also survived down-selection in the IEEE process. DS-UWB is based on traditional, narrow pulses that can themselves occupy 1.5 GHz bandwidth or more. Two modes were described for DS-UWB – one using BPSK modulation and variable length spreading codes to obtain various data rates, and another employing orthogonal modulation with 4-BOK (binary orthogonal keying) sequences of different length. These modes were defined for use both in the lower frequency band from 3–5 GHz, and in a band of twice the width above 6 GHz. It is interesting to note that for the BPSK modes with data rates of 110 Mbps and higher, the DS-USB proposal is actually to be regarded as an impulse radio system [IEEE P802.15-04/137r4, Table 2.7].

MB-OFDM and DS-UWB each had technical pros and cons and attracted their adherents in the IEEE 802.15.3a Task Group, but neither was able to achieve the 75% majority of voters needed for confirmation. As a result, in January 2006, the Task Group voted to recommend termination of the effort. However, adherents of both approaches had already been pursuing alternate paths to industry alliance, resulting in the WiMedia Alliance (MB-OFDM) and the UWB Forum (DS-UWB). Products using both technologies were planned for 2006. In January 2006, the MB-OFDM approach was accepted by the European standards organization ECMA as ECMA-368, and in March 2006, the Bluetooth Special Interest Group chose the WiMedia/ECMA-368 standard as the technology to be used in the next-generation, higher speed version of Bluetooth.

\section{2.3 System Design Considerations}

\subsection{2.3.1 UWB Channel Models}

In order to implement an efficient UWB system for high-rate communications, it’s critical to understand the characteristics of the propagation channel. Intel and other companies performed several channel measurements spanning the frequency spectrum from 2 to 8 GHz (see [2, 3, 3] and related references) which were contributed to the development of a channel model in the IEEE 802.15.3a study group. An example channel realization is shown in Fig. 2.2, which points out two important characteristics of a very wideband, indoor channel. First, as can be seen in the figure, the multipath spans several nanoseconds in time which could result in inter-symbol interference (ISI) if UWB pulses are closely spaced in time, even when the separation distance is relatively short (less than 10 m). This ISI would need to be mitigated through proper waveform design, signal processing, and equalization algorithms at the cost of additional complexity. Second, the very wide bandwidth of the transmitted pulse permits resolution of several multipath components, which has its pros and cons. On the one hand, the multipath

\footnote{IEEE P802.15-04/137r4, January, 2005.}
arrivals will undergo less amplitude fluctuations (fading) since there will be fewer reflections that cause destructive/constructive interference within the resolution time of the received impulse. On the other hand, the average total received energy is distributed between a large number of multipath arrivals. In order to take advantage of that energy, unique systems and receivers need to be designed with multipath energy capture in mind. For a traditional impulse based UWB waveform, this may consist of a rake receiver with multiple arms, one for each resolvable multipath component. However, as the bandwidth of the UWB waveform increases, the complexity of the rake receiver could become limiting in order to capture the same energy. As a result, careful bandwidth selection of the UWB waveform can help balance the receiver complexity for capturing multipath energy while still benefiting from the reduced fading of the short duration of the pulses.

For proper system design, and to understand and quantify the impact of multipath propagation, it is important to have a reliable channel model that captures the important characteristics of the channel. Towards this end, a number of popular indoor channel models were evaluated in the IEEE to determine which model best fits the important characteristics that were measured and documented in [1]. The analysis and the results of this channel modeling work are contained in the IEEE 802.15.3a channel modeling sub-committee final report [3] and are briefly summarized here for completeness. Three indoor channel models were considered: the tap-delay line Rayleigh fading model [4], the Saleh–Valenzuela (S–V) model [5], and the Δ-K model described in [6]. Each channel model was parameterized in order to best fit the important channel characteristics, which included the mean excess delay, mean RMS delay, and mean number of significant paths defined as paths within 10 dB
of the peak path power. The results found that the S–V model was able to best fit the measurements and observed characteristics of the channel. In particular, the channel measurements showed a clustering of the multipath arrivals, which is also found in [7] and captured by the S–V model. In addition, the amplitude statistics of the measurements was found to best fit the log-normal distribution rather than the Rayleigh, which was part of the original S–V model. We have since compared the amplitude distribution to the Nakagami distribution and found that both the log-normal and Nakagami distributions fit the data equally well. So, the S–V model was modified slightly in order to take the log-normal fading distribution into account. The final proposed model is described in [3], and the reader is encouraged to visit these references for more detail.

Since completion of the work of the IEEE 802.15.3a channel modeling subcommittee, a number of additional measurements and models have been investigated which give even greater insight into the UWB propagation channel. An excellent tutorial of the latest theory and modeling of UWB channels is provided in [8], which also provides a thorough list of references addressing further UWB propagation characteristics including angular dispersion needed for MIMO implementations, temporal variation, and considerations for low-rate and localization applications.

2.3.2 Multipath Energy Capture and ISI

2.3.2.1 Impulse Radios’ Design Impact in Multipath Channels

There are different ways to design a UWB impulse radio system in order to mitigate the impact of multipath and ISI. For example, if the impulses are sufficiently far apart in time (further than the maximum delay spread of the channel), then the ISI is reduced. The throughput is also significantly reduced, but this could be recovered to some extent by using higher order modulation on the impulses (of course, at the expense of range and increased linearity requirements of the RF front end). In order to illustrate this, the following results compare various impulse $M$-PAM systems based on the same throughput (see [9] for additional details). As $M$ increases, the pulse repetition period increases for a given throughput. Figure 2.3 shows the performance of a UWB system without a rake receiver (i.e., $L_p = 1$). For simplicity, a so-called $\Delta$-K model [10] is used to generate the following results which uses a two-state Markov model for the path arrival probabilities and is based on a number of previously reported indoor channel measurements for narrowband channels, but suffices here for explanation.

In this case, the 2-PAM system experiences considerable ISI, which causes it to have an error floor at high SNR regions. As the modulation level is increased, the error floor is lowered due to the reduced ISI. The various cross-over points between the $M$-PAM systems occur as the benefits of reduced ISI for the higher order
Figure 2.3 shows the performance of $M$-PAM systems with maximal ratio combining RAKE (MRC-RAKE). Interestingly, the ISI seems to have less of an effect when an MRC-RAKE is used. This is due to the effective diversity that is obtained by capturing the multiple paths (i.e., the ISI experienced on each path is different and therefore results in averaging the overall ISI caused by the channel at the output of the RAKE). Averaging the results over more channel realizations could show the existence of an error floor, but the general observation that the effect of ISI is lessened by the use of a rake receiver should still hold.

Figure 2.4 shows how the performance of the MRC-RAKE changes with the number of combined paths. In order to remove the effects of ISI, the symbol rate is reduced to 5 Mbps in this case. This figure shows that the majority of the energy is
Fig. 2.5 Performance of 2-PAM with an MRC-RAKE and a pulse repetition period of 200 μs (5 Mbps and thus no ISI)

captured with $L_p=50$ where $L_p$ is the number of arms in the rake. Clearly or there is a trade-off between receiver complexity and performance, but this figure shows that there are significant gains possible with the use of a RAKE combiner. This type of system design may not be appropriate for high-rate communications since it requires the impulses to be spread apart in time in order to minimize the impact of ISI. However, it might be very appropriate for low-rate designs.

Since rake reception both significantly improves energy capture and mitigates ISI due to the averaging effect of the rake, another impulse system design option is to use a direct sequence (DS) spread spectrum approach, with an impulse radio and a rake receiver. In this case, actual channel measurements are used to generate the following results (see [11] for additional details and analysis). Figure 2.6 shows the performance of the DS UWB system in a multipath channel at a bit rate of 100 Mbps (i.e., for a non-spread system, pulses are separated by 10 μs, and for a spread system with $N_r$ chips per bit, pulse are separated by $10/N_r$ μs). These results show that the performance is actually worse for the DS systems compared to a system with no DS spreading due to the inter-chip interference caused by the multipath and the non-zero autocorrelation of the spreading sequences. This figure also shows that there can be significant benefits for increasing the number of RAKE arms from 3 to 6 for these indoor channels.

One of the main reasons for using direct sequence spreading is to enable multiple users to share the same spectrum simultaneously [12]. Figure 2.7 shows results with multiple access interference (MAI) and multipath present for a data rate of 16.7 Mbps (corresponding to a bit period of 60, 62, and 63 μs for the length 15, 31, and 63 codes, respectively). In AWGN, there is a clear advantage for using DS spreading and the benefits of the longer codes which have lower cross-correlation values. For the non-spread system, the probability of making an error is dependent on the probability of a collision, which can be relatively high with even a single user present. The correlations of the DS codes alleviate this problem when a direct collision occurs. However, the benefit of spreading is reduced in a multipath
channel, as shown in Fig. 2.7 due to the inter-chip interference. It is interesting to note that the length 31 code has similar performance compared to length 63 code with multipath and MAI. This is due to the improved cross-correlation properties of the longer code which helps reduce the impact of other users even in the presence of severe multipath. The results clearly show that the performance of a DS system is highly dependent on the auto- and cross-correlation properties of the spreading sequences and highlights the importance of taking into account the propagation model when designing the system. The combination of careful sequence design along with equalization techniques should further improve the systems performance.
Finally, it is worth noting other impulse radios, designs which are not covered here. In particular, impulse radios using time-hopping and pulse position modulation (PPM) have been heavily studied in the research literature (see [13] and related references). These systems will have similar design trade-offs when it comes to optimizing performance in a multipath and multiple access environment. In particular, energy capture may also require a rake receiver to capture the energy from the reflections in the channel, and the combination of multipath and multiple access interference would need to be carefully considered when designing the PPM receiver and hopping patterns.

### 2.3.2.2 MB-OFDM Radio Design for Multipath Channels

Multiband OFDM (MB-OFDM) combines the advantages of multibanding for frequency diversity and implementation complexity, with the well-known robustness of OFDM in dense multipath channels. OFDM systems can mitigate multipath using either a cyclic prefix or a zero pad [14]. In the former, a portion of the OFDM symbol is copied from the end and pre-appended to the beginning of the symbol before transmission. Unfortunately for UWB, the cyclic prefix causes repetitive structure in the signal which shows up as spectral ripple in the OFDM spectrum, as shown in Fig 2.8. Since the FCC specifies allowed PSD based on the peak spectrum, such ripple is not desirable.

However, the spectral ripple can be eliminated by using a zero-padded prefix [14]. The zero padding at the transmitter requires a receiver using overlap-add FFT

![Spectral flatness comparison between OFDM with a cyclic prefix vs. zero padding](image)

**Fig. 2.8** Spectral flatness comparison between OFDM with a cyclic prefix vs. zero padding
processing – this implies that the effect of the zero padding can be regarded as similar to employing a noisy cyclic prefix. This typically results in a small SNR loss in performance when compared with the cyclic prefix OFDM case, but the reduced spectral ripple for the zero-padded signal permits enough of an increase in average TX power to compensate for this loss.

In general, the design of the MB-OFDM signal follows that of a conventional OFDM system, with the major difference being the frequency hopping induced by the time-frequency codes (TFC). The choices for the numerical values of the OFDM symbol duration follow from the multipath channel characteristics – considering the delay spread over all the different environments, a zero pad of ~60 μs was found to yield a good compromise. Based on this, keeping the overhead of the zero pad at roughly 25% led to an OFDM symbol length of ~240 μs (for a total length of ~300 μs). This OFDM symbol length implies a sub-carrier spacing of ~4 MHz, and given a minimum occupied bandwidth of 500 MHz, leads to a 128 point FFT/IFFT being necessary. An additional guard interval of about 9.5 μs is inserted between the successive OFDM symbols, in order to simplify the design of the frequency-hopping circuits. The precise values for these parameters lead to a total duration of 312.5 μs on each frequency band (242.4 μs OFDM symbol, 60.6 μs zero pad, and 9.5 μs additional guard interval).

The WiMedia MB-OFDM symbol is formed from a 128 point FFT, which means that potentially there are 128 sub-carriers with a spacing of 4.125 MHz in the OFDM symbol, of which only 122 are actually used, for both data and reference signals. A time frequency code (TFC) is used to modulate each successive OFDM symbol on multiple frequency bands, each band being 500 MHz wide as required by the FCC part 15.500 rules. Thus, if we represent each OFDM symbol as $Z(t)$, then the transmitted RF signal would be written in terms of the complex baseband signal as

$$s_{RF} = \Re \left\{ \sum_{n=0}^{N-1} Z_n(t - T_{SYM}) \exp(j2\pi f_c(q(n))t) \right\}$$

where $\Re(\cdot)$ represents the real part of the signal, $N$ is the number of symbols in the packet, $f_c(m)$ is the centre frequency for the $mth$ frequency band, $q(n)$ is a function that maps the $n$th symbol to the appropriate frequency band, and $Z_n(t)$ is the complex baseband signal representation for the $n$th symbol.

Modulation is applied to each of the OFDM sub-carriers using traditional techniques such as BPSK, QPSK, or 16-QAM. This allows the signal to carry multiple bits per symbol and permits the system to change the bit rate while holding the OFDM symbol rate constant. An example signal spectrum is shown in Fig. 2.9 (see ECMA-368 standard).

The MB-OFDM signal, especially for rates up to 200 Mbps, offers considerable protection against Rayleigh fading on the individual sub-carriers as well as MAI through a combination of time, frequency, and FEC code diversity techniques. At higher data rates, however, with the diversity gain weakening, an additional
Fig. 2.9 Example spectral plot of the MB-OFDM waveform in a single band

This technique is used for frequency selective fading mitigation. This is called dual carrier modulation (DCM), in which four-coded and interleaved bits are mapped onto modulated symbols on a pair of widely separated sub-carriers. This orthogonal mapping is derived from a modification of the $2 \times 2$ Hadamard matrix. It can be equivalently represented as mapping four interleaved bits onto two 16-point symbols using two fixed but different mappings as shown in Fig. 2.10. The two resulting 16-point symbols are mapped onto two different tones separated by 50 ton, as shown in Fig. 2.10.

Fig. 2.10 Dual carrier modulation (DCM) block diagram and signal mapping
2 High-Rate UWB System Design Considerations

This technique has the advantage of having the same 4 bits of information on 2 tones that are separated by at least 200 MHz. The assumption here is that the probability that there is a deep fade on both tones is small. And if there is a deep fade on one of the 2 tones, the 4 bits of information can be recovered using a simple detection scheme. This approach offers a large degree of frequency diversity and enhanced performance for the higher data rates where the FEC code is weak (320–480 Mbps).

2.3.3 Processing Gain and Spectral Flatness

Processing gain is a measure of the increase in signal-to-noise ratio achieved by integration of the spread spectrum code word over the duration of the data symbol period. In continuous chip systems, the processing gain is related to the number of chips per symbol. This definition is difficult to apply to an UWB based system; instead, we use an alternate definition of processing gain as the ratio of the UWB bandwidth to the symbol rate.

\[ PG = \frac{BW_{UWB}}{R} \]

The reason we utilize this definition has to do with the fact that in UWB, transmitter power is typically allocated on a per MHz basis. For example, in the United States, the FCC has set the power spectral density (PSD) for UWB to be –41.3 dBm/MHz. This means that the larger the occupied bandwidth the more available transmitter power. To calculate the total average transmitted power, we integrate the power over the UWB bandwidth. Then, the total power that will be allocated to each data symbol is the average power times the processing. As a result, the slower the data symbol rate, the more energy allocated to each bit when calculating \( E_b/N_0 \).

Implied in the way that transmitter power is allocated to UWB is the need for spectral flatness to achieve the maximum transmit power. For example, the FCC has indicated that the spectral peak must not exceed the –41.3 dB/MHz limit. The total transmitted power will be determined by the average PSD but the maximum value in the spectrum can not exceed regulatory limits. Hence, keeping the spectral peak-to-average ratio as small as possible is the key to maximizing the total transmit power. OFDM achieves spectral flatness by transmitting a spectral series of RF tones, where each RF tone’s amplitude is controlled to facilitate a flat spectrum.

2.3.4 Multi-User Access

Multi-user access in the WPAN environment is determined by how many uncoordinated simultaneous operating piconets (SOP) can be accommodated in the same vicinity at the same time. In IEEE802.15.3a and in ECMA-368, it was desired to accommodate an SOP of at least four.
Multiband OFDM makes use of several mechanisms to facilitate multi-user access via a combination of frequency channels, hopping codes, and spreading codes. All this comes under the general heading of a time-frequency coding.

The time-frequency codes are generated by first subdividing the UWB band into fourteen 500 MHz wide sub-bands, band 1 through band 14. These 14 sub-bands are collected into band groups as shown in Fig. 2.11 below.

Figure 2.12 shows an example time-frequency code using the frequencies in band group 1. This particular code sequentially steps through the available bands.

Other codes can be generated as shown in Table 2.2. Each time-frequency code gives some degree of separation between multi-users simultaneously using band group 1. In addition, codes 5, 6, and 7 are used for frequency division multiple access, giving complete separation for up to three users.11

The use of these codes decreases the overall probability of two transmissions interfering with each other. Still, under a heavily loaded scenario, it is possible for interference from other users to occur. In these cases, we rely on the OFDM processing gain and on the forward error correction to allow successful transmission of the data. In addition, decoding techniques which apply erasures to interfered symbols can also help mitigate the impact of strong interferers.

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11 TFI = time-frequency interleaving; FFI = fixed-frequency interleaving.
2.3.5 Implementation Issues

2.3.5.1 Antenna Dispersion and Aperture

Antenna aperture relates to the concept of a propagating sphere of radiated energy emanating from an isotropic radiator, as shown in Fig. 2.13.

We can model the antenna as a patch of fixed area ingesting energy off the surface of an energy sphere. As the surface area of the sphere expands with radius distance, the percentage amount of power in the antenna patch area proportionally decreases. Since the surface area is proportional to the square of the radius \( S = 4\pi \cdot r^2 \), the amount of available power decreases inversely to the radius squared, or in decibels, as \( 20 \times \log_{10}(r) \) where \( r \) is the sphere radius.

The antenna patch square itself is modeled as being proportional to its center frequency on each side; that is, the dimensions of the antenna patch can be modeled as being some fraction of a wavelength on each side, which then gives an antenna patch area that is proportional to the operating wavelength squared multiplied by some scaling constant. This is called the antenna effective aperture. (Fig. 2.14).

\[
A_{\text{aperture}} = \kappa_y \kappa_x \lambda_c^2
\]

Since the wavelength is inversely proportional to frequency, as the center frequency increases the effective aperture decreases (our conceptual patch antenna gets...
Fig. 2.14 An antenna aperture is proportional to the operating wavelength smaller). In terms of decibels, the decrease in ingested signal power is proportional to $-20 \times \log(F_c)$, where $F_c$ is the center operating frequency.

The impact that antenna aperture has on UWB can be assessed by considering the wide operating frequency of UWB: from 3.1 to 10.6 GHz in the United States. Optimizing the antenna for this wide bandwidth can be problematic. An antenna optimized for the low end of the UWB band (i.e., low VSWR, etc.) will have good aperture characteristics at low frequencies but poor VSWR and efficiency at high frequencies, and visa versa for an antenna optimized for the high end of the UWB band. In general, the philosophy followed is to apply the $20 \times \log 10(F)$ aperture shrinkage to the operating center frequency in an attempt to capture the varied performance characteristics in some summary manner. However, systems need to be able to adapt to non-flat antenna patterns due to the very wide occupied bandwidth and expected variability in patterns resulting from different antenna designs as well as different m factors.

2.3.5.2 Analog Front-End Challenges

OFDM-Based Transceivers

Figure 2.15 shows a generic analog front end (AFE) for an OFDM-based UWB transceiver. In this section, we describe challenges associated with the key elements of the AFE, emphasizing those attributes that are unique to UWB. Based on analysis and simulation, nominal values are offered for some parameters, but the reader should understand that based on designer choices, these values may differ substantially across different implementations.

Fig. 2.15 OFDM-based receiver block diagram
A UWB antenna (Ant) must be broadbanded, covering anywhere from a single 500 MHz segment to a span of multiple GHz with low loss and near omnidirectional coverage. Still better is an antenna that also offers some rejection for frequencies where other devices are known in advance to be active, such as IEEE 802.11a or 802.11n devices. Early results have been encouraging, and UWB antenna design remains an area of active research.

Band definition filters (BDF) and transmit/receive switches (T/R) must be designed with very low insertion loss since every dB of loss ahead of the low-noise amplifier (LNA) contributes directly to the overall noise figure of the receiver. In some designs, the front-end filtering may be split between two filters, with one ahead of the LNA as shown, and another just after the LNA. This may allow a filter with lower insertion loss to be placed ahead of the LNA. Nominal insertion losses for BDFs vary from 0.5 to 1.5 dB.

The LNA must offer very low noise figure since here too, every dB of noise figure adds directly to overall receiver-chain noise figure. The designer must balance the need for high gain against the increased noise figure that typically comes with higher gain. Some designs implement the LNA function in two or more stages, allowing for an initial stage with modest gain and very low noise figure. Subsequent stages and the remainder of the receiver chain can then have relaxed requirements on noise figure. Overall receiver chain noise figures vary with implementation, but a nominal value of 6–7 dB has been assumed in analyses and simulations.

Besides low noise figure, an LNA section must offer good linearity since nonlinearities will generate new spectral content that can damage many tones in an OFDM-based system. Depending on its design, pulse-based system may be somewhat less sensitive to non-linear distortion. As an example, in OFDM-based systems, nominal values for input third-order distortion (IIP3) are –10 to –15 dBm in an LNA that offers a nominal 25 dB of gain.

In an ECMA-368 OFDM-based UWB system, the mixer and local oscillator (LO) have additional challenges not found in traditional AFE designs. Specifically, the LO must “hop” among the different sub-band center frequencies and stabilize in as little as 9 μs. Furthermore, in a direct-conversion design like that in Figure 2.15, DC-offset in the mixer is likely to be different for each sub-band, requiring special compensation measures in either the analog circuitry or in the digital computations that follow the analog-to-digital converter (ADC). As with any LO, phase noise is of concern, although with sub-carriers located 4.125 MHz apart, phase noise is not as critical as it is in OFDM-based systems like IEEE 802.11a/g/n, where sub-carriers are spaced much more closely.

As in traditional designs, the variable gain amplifier (VGA) readies the signal for optimal use of the dynamic range of the analog-to-digital converter (ADC).

---

an ECMA-368 receiver, the ADC sampling rate is at least 528 MHz, and in some
designs may reach 1056 MHz. At these speeds, the ADC can consume considerable
power. However, bit precision in this ADC is typically 5 or 6 bits, which helps
hold down power consumption. As usual, there is a trade-off between total dynamic
range and quantizing noise. One study has shown that for a 6-bit ADC, the optimum
dynamic range is approximately $3.5 \sigma$, where $\sigma$ is the root-mean-square voltage at
the input to the ADC.\textsuperscript{14}

On the transmit side, the digital-to-analog converter (DAC) is typically 6–8 bits
precision. The baseband filtering section (BB Filtering) may vary widely among
implementations. In its simplest form it is a zero-order “hold” filter, although such
a choice can suppress the energy in the outer tones of the UWB spectrum. Many
implementers will choose to use some form of interpolation in order to flatten the
resulting spectrum and reduce sidelobe energy. This also eases the constraints on
baseband filtering that precedes the up-conversion mixer. The power amplifier per-
forms a traditional role, but in an ECMA-368 transmitter, the maximum radiated
power is only about 100 mW, making the design for linearity somewhat easier than
in larger power amplifiers. OFDM signal levels occasionally do reach large peak
levels, but based on simulations, “clipping” the signal at 7–9 dB above rms value
causes little degradation to the system performance.

Pulse-Based Transceivers

Pulse-based UWB receiver AFE designs look similar to those of other receivers,
consisting of the same antenna, BDF, LNA, Mixer, LPF, VGA, and ADC chain.\textsuperscript{15}
Some of the challenges, such as low-noise figure designs are similar to those in
OFDM-based counterparts. Some other attributes, like LNA and mixer linearity,
may be less critical in a pulse-based design. The largest differences from OFDM-
based designs appear after the ADC, where rake filters and decision feedback equal-
izers are used.\textsuperscript{16}

Pulse-based UWB transmitter AFE designs may look quite different from tra-
ditional ones. While mixers, power amplifiers, and filters may still be present, the
preceding baseband digital circuitry is quite different, consisting in some cases of
hybrid digital and analog circuitry for pulse formation. Many of the implementation
details are still proprietary or experimental.

\textsuperscript{14} Stein, Bart, Selecting Resolution and Dynamic Range in an OFDM UWB System, IEEE P802.15-
04/0007r0, January, 2004.
\textsuperscript{15} McCorkle, John, “Ultra Wide Bandwidth (WUB): Gigabit Wireless Communications for Bat-
tery Operated Consumer Applications”, 2005 Symposium on VLSI Circuits Digest of Technical
Papers, June 16, 2005, pp. 6–9.
\textsuperscript{16} Parhar, Ambuj et al., “Analysis of Equalization for DS-UWB Systems”, ICU 2005 International
Conference on Ultra-Wideband, Session 6, September 6, 2005.
2.3.5.3 Digital Processing Challenges

The primary challenges of baseband digital signal processing in a UWB transceiver arise from the extremely large bandwidths and associated sampling rates, as well as the need to process very high data rates. Partly related to this is also the challenge of combating the effects of the highly frequency selective dense multipath channel. Different approaches to UWB system design have been already described previously, and here we shall summarize some of the key features of the baseband digital processing in the transceivers for these systems.

For a pulse-based UWB system, the most demanding processing is at the “front end” since the receiver will typically sample the UWB pulses at the Nyquist rate – this also corresponds to chip rate sampling for a DS-UWB system. Thus, for example, a 1.5 GHz bandwidth UWB system will have a 1.5 Gsps complex input sample stream from the ADCs. Typically, the ADC will need to have some excess dynamic range to contend with in-band narrowband interferers and so a 2–3 bit ADC would be employed. Based on studies of the typical indoor UWB channel models, especially the NLOS channels, a rake receiver with up to 16 arms may be needed to capture adequate energy from the multipath in order to maintain a good link budget margin [see IEEE802.15-03/388r2]. Further, for good performance at the higher data rates with either impulse or DS-based UWB systems, a symbol rate equalizer is needed in order to suppress the ISI.

The multiband OFDM UWB system represents a very efficient realization of the multiband approach to UWB system design. The “dwell” time on each band is increased through use of a 242.4 μs long OFDM symbol, which offers good energy capture, and the zero pad plus overlap-add FFT processing offers an efficient equalizer implementation. The WiMedia MB-OFDM transceiver reduces the ADC sample rate requirements compared with impulse-based UWB systems – a 528 MHz complex input sample stream is required. However, in most cases, a little higher dynamic range is required in the ADC and DAC, and 6 bits of precision is typically employed. The signal processing blocks in the baseband chain are similar to those in conventional OFDM modems – including functions such as timing synchronization, channel estimation, pilot symbol tracking, soft-bit mapping, etc. The FFT engine needs to perform a 128 point FFT every 312.5 μs. The other major signal processing block is the Viterbi decoder, which is required to decode the rate 1/3, constraint length \( K=7 \) convolutional code at a peak source data rate of 480 Mbps (when the code is punctured to rate \( 3/4 \)).

2.3.6 Link Budget

Link budgets for UWB radios seem to cause undue confusion with system engineers trying to convert from peak power to average power, taking into account processing gain, etc. Actually, to calculate \( E_b/N_0 \) – in an AWGN environment due to the receiver front-end noise – we basically only need to know four things: the transmit spectrum bandwidth, the bit rate, the path loss, and the receiver noise figure...
(assuming isotropic antennas). An example link budget used by IEEE802.15.3a [3] is shown in Table 2.3.

From a simplistic point of view, $E_b$ is determined by integrating the regulatory limited PSD over the wavelet noise bandwidth, scaling by the path loss and dividing by the bit rate. The noise density $N_0$ is then determined by the noise figure of the receiver. An example link budget is shown in Table 2.3 for a generic UWB system using the table defined in [3].

Table 2.3 only yields average expected performance of the system, while more accurate performance and reliability needs to be validated with detailed simulations in realistic channels to determine how much additional margin is sufficient. The range at which the ECMA-368 multiband OFDM system, operating in band group 1, achieves a PER of 8% with a link success probability of 90% in example channels with mutlipath and shadowing is listed in Table 2.4 for various AWGN and multipath channel environments as defined by the IEEE802.15.3a committee in document [3] (see IEEE document 15-05-0648-00-003a-mb-ofdm-updates.pdf from November, 2005).

2.3.7 Initial Acquisition

Modern UWB systems are designed for high data rate packet-based wireless communications. Fast timing acquisition and synchronization is an especially important requirement for these systems since the preamble overhead (which is normally transmitted using a rate among the lowest data rates supported) needs to be minimized in order that overall throughput does not suffer. For example, for a 1,024 byte packet being transmitted at 100 Mbps, a 10 $\mu$s PHY preamble amounts to an overhead of approximately 11%. At 500 Mbps, the same length for the PHY preamble amounts to about a 38% overhead, while at 1 Gbps, the preamble will result in a 54% overhead. This is a simplistic calculation, and certainly the PHY layer preamble requirements can be affected by other aspects of the overall system design; however, the point remains that short preamble length/fast acquisition is an important factor in high data rate system design.

For a pulse-based UWB system design, the approach to timing synchronization can be based on using a serial search strategy with an analog or digital correlator, or using a digital matched filter. The serial-search-based approach trades lower implementation complexity for a longer average acquisition time due to the necessity of dwelling for a fixed integration time at any given timing epoch. The matched filter approach relies on a greater implementation complexity to obtain a faster acquisition time. The UWB channel characteristics have a significant impact on the acquisition time because the dispersion in the signal energy among the multiple reflections that define the dense multipath channel causes the per path signal energy to be a small fraction of the total energy in the transmitted UWB pulse. In [15], the effect of the multipath channel profile on impulse-based UWB and multiband UWB pulse timing acquisition was studied. As shown in Fig. 2.16, the multiband pulse system
Table 2.3 Example link budget calculation

<table>
<thead>
<tr>
<th>Parameter: Mode 1 DEV</th>
<th>Value</th>
<th>Value</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Throughput ((R_b))</td>
<td>100 Mb/s</td>
<td>200 Mb/s</td>
<td>500 Mb/s</td>
</tr>
<tr>
<td>Average Tx power ((P_T)) (~1.5 GHz bandwidth)</td>
<td>−10 dBm</td>
<td>−10 dBm</td>
<td>−10 dBm</td>
</tr>
<tr>
<td>Tx antenna gain ((G_T))</td>
<td>0 dBi</td>
<td>0 dBi</td>
<td>0 dBi</td>
</tr>
<tr>
<td>(f_c = \sqrt{f_{\text{min}} f_{\text{max}}}: \text{geometric center frequency of waveform} (f_{\text{min}} \text{ and } f_{\text{max}} \text{ are the} -10 \text{dB edges}))</td>
<td>3.850 MHz</td>
<td>3.850 MHz</td>
<td>3.850 MHz</td>
</tr>
<tr>
<td>(f_{\text{min}} = 3.1 \text{ GHz}; f_{\text{max}} = 4.8 \text{ GHz})</td>
<td>(f_{\text{min}} = 3.1 \text{ GHz}; f_{\text{max}} = 4.8 \text{ GHz})</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Path loss at 1 m ((L_1 = 20 \log_{10}(4\pi f_c/c))) (c = 3 \times 10^8 \text{ m/s})</td>
<td>44.2 dB</td>
<td>44.2 dB</td>
<td>44.2 dB</td>
</tr>
<tr>
<td>Path loss at (d) m ((L_2 = 20 \log_{10}(d)))</td>
<td>20 dB ((d = 10 \text{ m}))</td>
<td>12 dB ((d = 4 \text{ m}))</td>
<td>6 dB ((d = 2 \text{ m}))</td>
</tr>
<tr>
<td>Rx antenna gain ((G_R))</td>
<td>0 dBi</td>
<td>0 dBi</td>
<td>0 dBi</td>
</tr>
<tr>
<td>Rx power ((P_R = P_T + G_T + G_R - L_1 - L_2))</td>
<td>−74.2 dBm</td>
<td>−66.2 dBm</td>
<td>−60.2 dBm</td>
</tr>
<tr>
<td>Average noise power per bit ((N = -174 + 10 \times \log_{10}(R_b)))</td>
<td>−94 dBm</td>
<td>−91 dBm</td>
<td>−87 dBm</td>
</tr>
<tr>
<td>Rx Noise Figure ((N_F))</td>
<td>7 dB</td>
<td>7 dB</td>
<td>7 dB</td>
</tr>
<tr>
<td>Average noise power per bit ((P_N = N + N_F))</td>
<td>−87 dBm</td>
<td>−84 dBm</td>
<td>−80 dBm</td>
</tr>
<tr>
<td>Required (E_b/N_0) ((S)) (assuming strong coding)</td>
<td>4.0 dB</td>
<td>5.0 dB</td>
<td>6.0 dB</td>
</tr>
<tr>
<td>Implementation loss ((I))</td>
<td>3.0 dB</td>
<td>3.0 dB</td>
<td>3.0 dB</td>
</tr>
<tr>
<td><strong>Link margin available for multipath and shadowing</strong> ((M = P_R - P_N - S - I))</td>
<td>5.8 dB</td>
<td>9.8 dB</td>
<td>10.8 dB</td>
</tr>
</tbody>
</table>
Table 2.4 Range to achieve a PER fo 8% with 90% link success probability for band group 1 devices

<table>
<thead>
<tr>
<th>Rate(Mb/s)</th>
<th>AWGN(M)</th>
<th>CM1(M)</th>
<th>CM2(M)</th>
<th>CM3(M)</th>
<th>CM4(M)</th>
</tr>
</thead>
<tbody>
<tr>
<td>110</td>
<td>21.5</td>
<td>12.0</td>
<td>11.4</td>
<td>12.3</td>
<td>11.3</td>
</tr>
<tr>
<td>200</td>
<td>14.8</td>
<td>7.4</td>
<td>7.1</td>
<td>7.5</td>
<td>6.6</td>
</tr>
<tr>
<td>480</td>
<td>9.1</td>
<td>3.8</td>
<td>3.5</td>
<td>N/A</td>
<td>N/A</td>
</tr>
</tbody>
</table>

offers the potential for lower search times due to the larger average energy in the strongest path compared to a pulse-based UWB system occupying the same overall bandwidth.

As mentioned earlier, modern high data rate UWB systems extensively employ digital baseband processing with Nyquist sampling front ends. In these systems, the requirement for fast, accurate timing acquisition/synchronization demands matched filter-based processing. As an example, we describe the Wimedia MB-OFDM UWB system and its preamble structure, which forms the initial part of every transmitted packet and is designed to enable efficient matched filter processing in the receiver. The preamble is designed to keep the overall length <10 μs while yielding a robust acquisition/synchronization architecture with moderate complexity in the receiver. The standard preamble sequence consists of 30 symbols (of 312.5 μs duration) and is divided into two segments:

1. The packet/frame synchronization sequence: This segment consists of 24 symbols and is used in the receiver to detect the start of a packet, and to perform timing and coarse frequency error estimation. This segment is formed by concatenating 24 repetitions of a base time domain preamble symbol (and associated 70.1 μs zero pad), and overlaying this with a 24 length cover sequence that depends on the particular time-frequency code (TFC) being used. The base time domain symbol is based on a 128 sample length hierarchical binary sequence which is constructed as the Kronecker product of sequences of length 16 and 8. At the receiver, the digital matched filter for this base sequence can be efficiently implemented as an 8 sample long matched filter followed by a 16 sample long

Fig. 2.16 CDF of fractional energy in strongest multipath component for an example impulse-based UWB system, and a multiband UWB system
matched filter. The cover sequence serves to delimit the end of the first segment, among other functions.

2. The channel estimation sequence: This segment consists of six repetitions of a base channel estimation symbol which is formed by computing the IFFT of a suitably defined random QPSK sequence (and appending the 70.1 \( \mu s \) zero pad). This ensures that two channel estimation symbols are provided for each of the three bands spanned by the TFC, and the receiver can employ suitable channel estimation algorithms to exploit these symbols.

Given the above definition, the length of the standard preamble is equal to 9.375 \( \mu s \) (30 \( \times \) 0.3125 \( \mu s \)). In addition, a streaming mode preamble is also defined for improved efficiency when multiple packets are being transferred in burst mode. In this case, the packets after the initial one in the burst can have a shorter packet/frame synchronization sequence of only 12 symbols because the timing offset of this packet is fixed with respect to the first one. Thus, the total preamble length is reduced to 5.625 \( \mu s \) (18 \( \times \) 0.3125 \( \mu s \)) and reduces the overhead while preserving good performance.

### 2.3.8 Narrowband Interference

Because of their very wide bandwidth, UWB receivers are likely to encounter interference from narrowband systems, especially those in the 3–5 GHz band. Some may be quite powerful, operating in an uncontrolled manner in close proximity to a UWB receiver. Other systems may be outside the passband of the UWB receiver, but if they are close enough, can still degrade UWB performance to some extent.

Both DS-UWB and MB-OFDM have some inherent resistance to narrowband interferers. The code spreading in DS-UWB helps suppress interferers relative to the desired signal, and a narrowband interferer is unlikely to affect more than a few of MB-OFDM’s 100–300 data carriers. Of course if the interferer is strong enough, the analog front end for either system could saturate, but these are special situations that must be dealt with by other means (such as time sharing the use of UWB and the interferer). In the remainder of this section, one important example is examined via simulation: WiMax-like interference into a WiMedia PHY 1.1 receiver.

As shown in Fig. 2.17, the simulation models a UWB Transmitter (Tx) located \( Du \) meters from a UWB Receiver (Rx). The Rx is a “victim” because of interference from a nearby WiMax Tx located \( Dw \) meters away. Some particulars of the UWB Rx and WiMax Tx are indicated in Fig. 2.17. WiMax systems are expected to appear in the 3.4–3.8 GHz band in Europe and other parts of the world.

Two UWB signal modes are modeled in this simulation. In TFI mode, all three bands are transmitted sequentially, and in FFI mode, only band 3 is transmitted, centered at 4.488 MHz. The goal of the simulation is to compare UWB link performance for these TFI and FFI modes in the presence of a WiMax interferer.

Figure 2.18 displays the spectrum of the received signal at the UWB Rx in FFI mode when \( Du = 10 \) m and \( Dw = 2 \) m. As shown, the spectrum is dominated by
WiMax → UWB Interference Simulation Setup

**Key UWB Rx Characteristics**
- **LNA**
  - Gain = 25 dB fixed
  - OIP3 = +10 dBm
  - OP1 = 0 dBm
- **NF = 6.6 dB**
- **BB LPF = 7-pole 270 MHz Chebyshev**
- **ADC**
  - 6-bit quantization
  - AGC sets ADC input to 113.5 mV RMS
  - ADC range = +/-320 mV RMS + 9 dB

**UWB Signal Spectra for**
- **Du = 10 meters, Dw = 2 meters**

**Fig. 2.17** Narrowband interference simulation layout

**Fig. 2.18** UWB and interferer spectra

the powerful WiMax interferer at 3.8 GHz. Figure 2.18 also shows the baseband spectrum in the UWB Rx after down-conversion and low-pass filtering. As shown, the strong WiMax interferer is still present, but it has been suppressed by the low-pass filtering.

Figure 2.19 shows simulated Bit Error Rate (BER) versus WiMax-Tx to UWB-Rx separation (Dw) for FFI and TFI modes and for three values of UWB-Tx to
UWB-Rx separation. In each of the six curves, a BER of $10^{-5}$ is assumed to represent acceptable UWB performance. For example, in TFI mode with $Du = 4$ m, a minimum WiMax separation of approximately $Dw = 32$ m is required. In FFI mode, a minimum separation of only 1 m is required. By repeating the simulation for various values of $Dw$ and $Du$, the data in the tables of Fig. 2.20 were generated.

As shown in the Fig. 2.20, operating a UWB link in FFI mode provides considerable protection against nearby WiMax transmitters even though the WiMax transmitter may be within the passband of the UWB Rx front end. Regulations require

**Fig. 2.19** Deriving separation distances for successful operation at 106.7 Mbps

**Fig. 2.20** Required separation distances for successful operation
that FFI-mode UWB transmit at $10 \log(3) = 4.7$ dB lower power than TFI-mode UWB, but FFI mode still outperforms TFI mode when a WiMax Tx is nearby.

Imperfections in implementation may further limit performance. For example, if the local oscillator associated with the WiMax down-conversion mixer has side tones present, some of these tones may down-convert WiMax interferer energy into the UWB baseband. Depending on its strength, this energy could increase BER to unacceptable levels.

Even with the imperfections above, optional techniques (not required by the standard) may be used to improve performance still further in the presence of WiMax interference. In particular, a switchable bandpass filter sub-system that protects a UWB transceiver from WiMax signals could be used when their presence is detected. This filtering can dramatically reduce the impact of WiMax interferers. Sub-systems like these might be used in conjunction with “detect and avoid” mechanisms under discussion to protect WiMax receivers from possible interference by UWB and to meet emerging international regulations, as will be discussed in the next section.

2.4 Regulatory Environment

2.4.1 Inside and Outside the United States

The regulatory process for making UWB systems commercially legal has taken a long road both inside and outside the United States. The FCC first initiated a Notice of Inquiry (NOI) in September of 1998, which solicited feedback from the industry regarding the possibility of allowing UWB emissions on an unlicensed basis following the same power restrictions for unintentional emitters described in the FCC Part 15 rules. Since this would result in UWB signals overlaying other wireless systems, it raised many concerns about the potential for interference to existing systems. In May 2000, the FCC issued a Notice of Proposed Rule Making (NPRM), which solicited feedback from the industry on specific rule changes that could allow UWB emitters under the Part 15 rules. The interference concerns, especially related to safety critical systems like GPS, resulted in very large industry participation in the comment process to the FCC. More than 900 comments were filed during this period, including several very detailed interference studies (see FCC docket 98–153). Finally, on February 14, 2002, the FCC issued its final ruling allowing UWB systems to operate on an unlicensed basis under the Part 15 rules allowing UWB operation from 3.1 to 10.6 GHz at a power spectral density (PSD) of $-41.3$ dBm/MHz. According to statements by the FCC, these rules reflected a very conservative approach in order to protect existing wireless systems while allowing the technology to be further developed and proven. As a result, the opportunity was created to exploit this newly allocated spectrum, but many people in the industry and government are still watching carefully to ensure that the technology does not disrupt current services.
Although the United States successfully completed an initial round of regulations for UWB emissions which has been followed by a few key updates, including the issuing of a waiver allowing multiband OFDM technology to operate at normal operating power levels, the technology is still not legal anywhere else in the world. Regulatory bodies outside the United States are also in detailed discussions and performing comprehensive interference studies of their own to determine possible rules to allow the use of UWB technology in their geographic region. The International Telecommunications Union (ITU) TG1/8 concluded its work in 2005 on a detailed impact study and possible regulatory framework. However, the results of the studies were very mixed and highly dependent on the assumptions used which resulted in a framework coming out of the ITU which provided a range of results based on the ITU studies and flexibility for regulatory bodies to adopt different emissions limits based on their regulatory philosophies and specific spectrum usage. In Europe, a key difference exists from what was considered during the FCC proceedings which is the allocation of fixed services/WiMax systems above 3.1 GHz. This spectrum was not allocated for such systems in the United States (although the United States recently opened up the spectrum from 3.65 to 3.7 GHz on a quasi-licensed basis for similar types of service), and since the usage models for these fixed services/WiMax systems put them in close proximity to UWB enabled devices, there was a clear interference concern if UWB was able to operate at the FCC levels of –41.3 dBm/MHz. As a result, in order to harmonize as closely with the United States limits while not posing harmful interference into these systems when using the lower frequency bands (3–5 GHz), a “detect and avoid” (DAA) approach was proposed to allow UWB devices to share the available spectrum when there is no nearby narrowband system using the spectrum. This approach has also been promoted in Japan, Korea, and elsewhere as a possible solution to allow existing and future services to share the available spectrum below 5 GHz with UWB devices. There is clearly a desire to have spectrum both below 5 GHz and above 6 GHz to allow for more available channels as applications using UWB technology become more pervasive. In addition, in the near term, current technology capability will allow below 5 GHz silicon solutions sooner than solutions operating above 6 GHz, which would allow for devices to be introduced into these other geographies sooner.

The spectral mask in Fig. 2.21 shows the current draft emissions mask being considered for Europe and Japan, which includes part of the band which may be available immediately after the decision without requiring DAA (including 6–8.5 GHz, and hopefully 4.2–4.8 GHz), while part of the band will have to wait until DAA is approved (mainly impacting the bands below 4.8 GHz). The current proposed Japanese mask is also shown, which differs from the European mask in a couple of important ways: (1) DAA may be required all devices starting in January, 2009 compared with “phase-in” dates of 2010 being considered in Korea and 2010–2012 being considered in Europe, and (2) the upper limit starts a 7.25 GHz rather than 6 GHz which will make having a single harmonized radio able to operate worldwide difficult. The regulations for DAA are still being defined and must be proved to work effectively before regulators will adopt the mitigation technique. Possible architectures and issues for DAA are discussed in a following section. Other geographies
Comparison between Possible Spectral Masks

Fig. 2.21 Proposed UWB emission masks being considered in Europe and Japan

are also in the process of developing rules, including Canada, Korea, China, and others. From a UWB industry perspective, having some harmonized spectrum to allow single implementations to be used around the world with a part of the spectrum which could be used as a control channel to help identify the location would be a useful goal. It remains to be seen if this will happen.

2.4.2 Interference Modeling

Clearly, proper interference analysis and modeling is critical for determining viable UWB regulations both to ensure existing services are protected from harmful interference and to ensure fair access to that spectrum from UWB devices under realistic operating conditions. The first question which usually arises is how to model the UWB interference given the range of UWB waveform options. The simplest model is to just assume the UWB interference looks like additive white Gaussian noise (AWGN). However, this model is only valid for UWB systems with a pulse repetition rate greater than or equal to the bandwidth of the victim narrowband system [16]. In order to illustrate this, Fig. 2.22 shows the uncoded probability of error for a victim narrowband system as a function of SIR for different ratios of the pulse repetition frequency to the victim bandwidth (denoted as \( N_p \) in the figures) and bipolar modulation for the UWB pulses. In this case, the AWGN approximation to the interference is fairly accurate until the pulse repetition frequency becomes much lower than the victim bandwidth.

As another example to illustrate the impact of a different modulation scheme, Fig. 2.23 shows the interference comparison with AWGN for a binary PPM system. In this case, since the modulation is a non-zero mean modulation scheme, it will have
spikes in the spectral domain, and so the interference depends on $N_p$ as well as the relationship between the center frequency of the narrowband waveform ($f_0$) and the pulse repetition period ($T_p$). When the product $f_0 \times T_p$ is an integer, then there will exist a spectral line in the spectrum of the narrowband waveform which may cause the interference to look different than AWGN. This is illustrated in Fig. 2.23 which shows that the interference could actually increase significantly relative to AWGN when the pulse repetition rate is high.

However, this effect can be minimized by proper randomization of the position of the pulse, as illustrated in Fig. 2.24. In this case, the random location of the pulse eliminates the spectral line, and the AWGN approximation again becomes a valid model.
One of the other significant concerns from victim services is the impact of multiple UWB devices in the same area (referred to as the aggregate interference scenario). In order to properly analyze this case, it is critical to make realistic assumptions in the analysis. To illustrate this, the following discussion summarizes a detailed analysis of the impact of the aggregate interference scenario for a fixed service (FS) application (representing either a point-to-point or point-to-multipoint system). A number of assumptions need to be made in order to accurately model the inference. For illustrative purposes, the urban environment is described here, and the results will be presented for both suburban and urban environments (see [17] for additional details). The aggregate interference in urban and suburban scenarios is evaluated via Monte Carlo simulations of large areas of UWB deployment with the following topologies: circular area with 10,000 UWB devices/km² and 6 km radius; building density is 1,200 km⁻²; building height probabilities according to Milan, Italy statistics; device height probabilities are obtained explicitly from the building height probabilities; victim antenna is located at the center of the area; and for provision of Fresnel zone clearance, paths below main beam of the victim antenna has only UWB devices more than 25 m lower than the main beam axis, which corresponds to 7 floors maximum building height for a 45 m antenna and 12 floors for the 60 m antenna and path width is 25 m on both sides of the main beam.

The path loss model (from the UWB emitter to the FS antenna) is also critical to model realistically. The following model was adopted here. For each device, it is randomly chosen if it is line-of-sight (LoS) or non-LoS (NLoS) to the victim, according to probabilities based on floor layout and building heights. Then, the outdoor path loss is calculated as follows:

- Indoor to outdoor path
  - Random attenuation with a given probability distribution (see below)
- Outdoor path for LoS – free space
2 High-Rate UWB System Design Considerations

- Outdoor path for NLoS:
  - Low height floors (<10 m): IEEE 802.16 Category B – NLOS [18, 19]
  - Medium and high height floors (>10 m): IEEE 802.16 Category C: NLOS [18, 19]

In addition, the following probability distributions of indoor-to-outdoor attenuations are considered:

- 15% devices with 12 dB attenuation
- 15% – 40 dB attenuation
- 70% – linear from 12 to 40 dB.

Finally, for aggregate interference analysis, it is important to have a realistic activity factor for when the UWB devices are actually transmitting. A detailed evaluation of the expected usage models for UWB devices, it is expected that, when considering a very large population of UWB devices, fewer than 1% would be active at any particular time. Table 2.5 shows the results of the final analysis with different assumptions for the activity factor, consideration of an additional shadowing term in the path loss model, and power control which will likely be used by devices in very close proximity.

The results of this analysis show that the UWB interference impact is highly dependent on the assumptions of the analysis. However, when using realistic assumptions, it is likely to be well below the noise floor of the narrowband receiver. Similar analysis exists for fixed satellite services and other services, which is necessary to justify the appropriate power spectral density levels allowed for UWB devices.

### 2.4.3 Possible Detect and Avoid Architecture

Here, we briefly describe how the “detect and avoid” (DAA) procedure could be defined in UWB regulations and how it could be specified for implementation in

<table>
<thead>
<tr>
<th># item</th>
<th>Shadowing (9 dB)</th>
<th>Power control 0–8 dB</th>
<th>Activity factor %</th>
<th>( I/N ) value (dB) urban/suburban</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>No</td>
<td>No</td>
<td>1</td>
<td>–21/–23</td>
</tr>
<tr>
<td>2</td>
<td>No</td>
<td>No</td>
<td>5</td>
<td>–16/–18</td>
</tr>
<tr>
<td>3</td>
<td>No</td>
<td>No</td>
<td>17</td>
<td>–12/–14</td>
</tr>
<tr>
<td>4</td>
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<td>No</td>
<td>100</td>
<td>–5/–7.5</td>
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<tr>
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<td>No</td>
<td>5</td>
<td>–15/–18</td>
</tr>
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<td>17</td>
<td>–11/–14</td>
</tr>
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<td>8</td>
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<td>No</td>
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</tr>
<tr>
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<td>Yes</td>
<td>1</td>
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</tr>
<tr>
<td>10</td>
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<td>Yes</td>
<td>5</td>
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</tr>
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</tr>
<tr>
<td>12</td>
<td>Yes</td>
<td>Yes</td>
<td>100</td>
<td>–7/na</td>
</tr>
</tbody>
</table>
appropriate standards. This is still a current area of research and we need to continue developing on these ideas in order to address the full range of interference scenarios.

In developing the DAA approach, the definition of the problem scenarios, and the expected performance in these situations is quite important. To start with, in the regulatory domain, the DAA approach is intended as an interference mitigation solution in the case when a single UWB device is in close proximity to a particular kind of narrowband communications system, namely broadband wireless access (BWA) systems such as indoor WiMAX subscriber stations (SS). It is important to note that, in the end, the DAA approach must be demonstrably generic, in that it should not rely on very detailed features or parameters of the victim system – rather, it should exploit general behavior of the class of BWA systems. That having been said, the initial focus is on WiMAX systems in the 3.5 GHz band, since these are the systems that are currently specified and being considered for commercial deployment. In what follows, we refer to WiMAX but keep in mind the requirements of more general BWA systems.

The WiMAX SS can be in a number of different states: (i) active bidirectional communication with the WiMAX base station (BS), with the link carrying a variety of different traffic – VOIP, video streaming on the downlink, or TCP/IP data traffic; (ii) idle or sleep state, in which the SS is mostly dormant, and wakes up at periodic intervals to listen for messages on the downlink, and responds with messages on the uplink frequencies; (iii) being powered “on”, in which case the SS is said to be in a network entry state and is executing a defined series of steps – this starts with listening to, and acquiring, downlink synchronization signals before initiating uplink transmissions. The detailed behavior of the SS in different kinds of BWA systems may vary, but the general characteristics in terms of these different states are expected to be similar, particularly since the system requirements are mostly similar.

The UWB DAA device can also be in different states – (i) starting up following “power on” and (ii) “active” state, in which the UWB device is in either active communication with other UWB devices, or in idle mode.

Different interference scenarios can be defined for DAA to address based on the interaction between the states of the WiMAX SS and the UWB device(s).

The main tools in the DAA “toolkit” are listed below:

1. A narrowband signal detection function: whenever the WiMAX BS or SS are transmitting, the UWB device potentially has the opportunity to detect these transmissions. In order to preserve as much generality as possible, it is assumed the detection only relies on features such as power in a given bandwidth. Associated with this function is a detection threshold. In general, detection with a low-enough threshold in order to reliably detect the BS downlink transmissions even at the edges of the cell may not be feasible in practice. Further, even if it were practical to do so, this would result in higher false alarms caused by any number of spurious signals, and would result in unneeded performance degradation in the UWB system. Since the interference of concern to the WiMAX SS occurs in close proximity to the UWB devices, by reciprocity the UWB device
should receive the WiMAX SS uplink transmissions with relatively high signal levels. This suggests an approach that sets the detection threshold based on the assumption that uplink transmissions are to be detected (of course, over some fraction of the cell this would also mean that downlink transmissions from the BS are also detectable). A number of factors should be considered in determining the appropriate detection threshold. These include (i) lack of certainty regarding whether the WiMAX system is FDD or TDD (ii) whether transmission power control (TPC) is in effect on the uplink/downlink, (iii) the amount of margin to be provided to account for uncorrelated fading between the detection path (WiMAX uplink transmission detected at the UWB device) and the interference path (UWB interference experienced at the WiMAX SS in the downlink frequencies) (iv) typical levels of spurious emissions in the neighborhood of UWB devices performing detection. These factors will all need to be considered, and the detection threshold will need to be determined through analysis and extensive testing.

2. A means of ensuring that adequate silent periods are created in a synchronized manner in transmissions within a given UWB device cluster. As a consequence of the DAA system’s reliance on uplink detection as described above, some interference scenarios require provision of a mechanism through which an as yet undetected WiMAX SS can be enabled to come into a state in which it can be detected by the UWB DAA system. The most promising current approach for this is to ensure that the UWB transmissions have adequate periods of no transmission – both in terms of duration and frequency of occurrence – on the potentially affected WiMAX frequencies. This can help a WiMAX SS that is in network entry state in the vicinity of a UWB device cluster. For instance, the WiMAX SS will be provided a reasonable probability of establishing the downlink synchronization, and when it subsequently commences uplink transmissions, the UWB DAA devices will be able to detect it and thus triggering interference avoidance mechanisms. The optimum range of values for the parameters such as the duty cycle, duration, and average frequency of occurrence of these silent periods needs to be determined through extensive testing with a range of systems.

3. The third component of DAA toolkit is the avoidance mechanisms. UWB systems may be able to avoid interference to the detected WiMAX SS in a number of ways. The key is that once the detection mechanism is triggered, it also provides an estimate of the frequency of the WiMAX system (UL/DL). The UWB device can then employ a narrowband notch filter, or a scheme such as avoidance of transmission on a broad band of frequencies (e.g., the MB-OFDM system can suppress transmission on an entire 528 MHz wide subband which spans the estimated location of the WiMAX DL transmission), or it can employ time-domain avoidance techniques based on either a priori or estimated information about the WiMAX transmissions. It is important to take into consideration various factors in framing the avoidance mechanism, such as the uncertainty about the location of the WiMAX downlink transmissions, if the detected signal is the uplink signal in a FDD WiMAX system, for example.
Table 2.6 Parameters for detect operation ([XX] refers to values that are yet to be defined)

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
<th>Remarks</th>
</tr>
</thead>
<tbody>
<tr>
<td>Narrowband signal detection threshold</td>
<td>$P_{th} = [xx]$ dBm/MHz</td>
<td>The UWB device shall detect the presence of a “victim” when arriving at the UWB Rx with this minimum PSD level</td>
</tr>
<tr>
<td>Initial channel availability check time</td>
<td>$T_{init} = [xx]$ $\mu$s</td>
<td>Upon initial power-on, or when the UWB device has been inactive for more than [XX] seconds, the UWB devices shall observe the channel for this time before using the channel</td>
</tr>
<tr>
<td>In-service channel availability check time</td>
<td>$T_{in} = [xx]$ $\mu$s</td>
<td>The UWB device must detect the presence of a “victim” within this time when a victim moves into detection range. This may vary with “victim” traffic and operational profiles</td>
</tr>
<tr>
<td>Continuous in-service minimum silent interval</td>
<td>$T_{silent}[xx]$ $\mu$s per $T_{sper}[yy]$ seconds in minimum blocks of $T_{sblk}[zz]$ $\mu$s</td>
<td>The local UWB system shall be able to allocate this minimum continuous silent interval periodically to allow new “victim” devices to begin operation. The UWB device shall also be able to detect a “victim” transmission during this period</td>
</tr>
<tr>
<td>Detection reliability</td>
<td>$R = [99]$%</td>
<td>This shall be measured via test procedures to be defined in ETSI and shall be met for the different coexistence scenarios described here</td>
</tr>
</tbody>
</table>

Table 2.7 Parameters for avoid operation

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
<th>Remarks</th>
</tr>
</thead>
<tbody>
<tr>
<td>Maximum emission level in avoid operation</td>
<td>$xx$ dBm/MHz</td>
<td>The UWB device, after it has detected a “victim” service, shall reduce the emission level below this value in the “victim” channel</td>
</tr>
<tr>
<td>Mitigation time period</td>
<td>$T_{mit} = [xx]$ min</td>
<td>The UWB device, after it has detected a “victim” system, shall leave the “victim” channel for a minimum of this time</td>
</tr>
<tr>
<td>Response time</td>
<td>$T_{resp} = [xx]$ $\mu$s</td>
<td>The UWB device shall reduce emissions in the detected bands within this time of successfully detecting a “victim” system</td>
</tr>
</tbody>
</table>

Tables 2.6 and 2.7 show examples of DAA parameters which could be specified in the regulatory domain, but the final numbers for these parameters require further research. This would allow for UWB standards and devices to be developed that can be verifiably demonstrated to meet the intent of the regulations and provide interference mitigation.

2.5 Cognitive Radios And Other Future Areas For Research

In 2003, under ET Docket 03-108, the FCC began an investigation of cognitive or “smart” radios that can [operate on an] opportunistic basis, finding idle spectrum, using it as they need, then vacating the band for others to use, all without human intervention. This model presumes
UWB designs, especially those based on OFDM, are a significant step toward the cognitive concept. ECMA-368 (WiMedia, ECMA-368) transceivers already feature wide-bandwidth/low-noise front ends, low cost, broadbanded antennas, and especially transmitted spectra that are modifiable “on the fly” under software control. When combined with detect-and-avoid, many of the essential elements of a cognitive design will already be present. OFDM-based UWB radios may well be the progenitors of the cognitive designs called for by the FCC. In addition, because they are based on OFDM, WiMedia designs are technical “cousins” of OFDM-based 802.11, 802.16, and some cellular technologies, raising the possibility of common architectural approaches and shared functions among wireless personal-area, local-area, and wide-area wireless communications systems.

While there has been considerable progress in UWB, there are still many opportunities for future research that are especially important to UWB. Some of these are outlined below.

1. Antennas – broadband antennas, including those with frequency response to help reduce interference to and from the UWB device.
2. Tunable band-pass filters – wide-dynamic range or passive (non-saturating) designs that can protect UWB receiver front ends and reduce UWB interference to other systems.
3. Low-noise all-CMOS analog front ends – designs that can accommodate higher frequencies (up to 10.6 GHz) with good sensitivity and low noise figure.
4. High-speed, low power ADC/DAC designs – low-power-consumption designs that can over-sample analog signals with 6-bit or higher precision.
5. Dynamic spectrum control – “on the fly” sculpting of transmitted spectra to accommodate local conditions and regulations.

2.6 Summary and Conclusions

The first radios, dating back to the late nineteenth century, were based on “spark” technology and were ultrawideband out of necessity rather than choice. Following the invention of the vacuum tube triode, it became possible to control bandwidth very precisely, and ultrawideband systems disappeared by the 1920s, emerging again for defense purposes in the 1960s. With the declassification of that work in the 1990s, research groups and small entrepreneurial companies began investigating the potential of UWB for commercial purposes. After a lengthy deliberation at the FCC, with more than 900 submitted comments, rules for unlicensed, Part-15-based operation of UWB were issued in February, 2002. After that date, activity in UWB

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increased dramatically, and a wide variety of UWB technologies emerged. Many were, and still are, based on “traditional” narrow-pulse techniques. One in particular, known as DS-UWB, combines pulse-based transmission with spread-spectrum techniques. Others, in particular the ECMA-368 approach largely described in this chapter, combine UWB with modern digital techniques like OFDM. Each approach offers a different set of trade-offs among performance, cost, energy efficiency, and spectral flexibility.

As described in Section 2.3, the very wide bandwidth of a UWB channel presents both opportunities and challenges. On the plus side, a UWB system should be much less vulnerable to Rayleigh fading than a traditional narrowband system. Multipath inevitably suppresses some narrow spectrum segments, but the rest will still get through, and some will actually be enhanced by multipath. Typical indoor channels will contain several hundred paths, spread over tens of nanoseconds. Impulse-based designs attempt to resolve individual paths, combining the energy from a few of the paths via RAKE-based techniques. OFDM-based systems typically spread their information over hundreds of sub-carrier tones, capturing most all of the signal energy that the channel has to offer. Both systems use error correction, commonly convolutional codes and Viterbi-based decoders, to recover transmitted data.

As described in Section 2.3, UWB antenna and analog-front-end designs present new challenges relative to traditional narrowband antennas. UWB antennas not only need to be efficient radiators over a wide spectrum but also, in the case of pulse-based UWB, need to cause as little pulse dispersion as possible in order to limit inter-symbol and inter-channel interference. Broad-band analog front ends need low noise figures for any UWB technology, and in the case of OFDM-based UWB, linearity and dynamic range are especially important.

While the low permitted power of all UWB-based systems provides some inherent protection for potential nearby victim systems, some degree of (preferably dynamic) spectral control is a desirable feature of a practical UWB system. As described in Section 2.4, UWB systems also need to detect the presence of potential nearby victims and estimate what frequencies need protection. For both the detect and avoid functions, the FFT function in OFDM-based UWB systems can be used to spot both the presence and frequency of potential victims. Furthermore, the digitally created waveforms and multiband design in MB-OFDM can avoid interfering with other systems by using selectable sub-bands or by suppressing individual sub-carriers under software control.

Because of their wideband front-end response, all UWB systems are themselves potential victims of any narrowband interferers within the receiver passband. DS-UWB-based systems mitigate this interference by spreading and decorrelating the interferer energy over multiple pulses. OFDM-based systems mitigate the interference by sacrificing a few of the affected sub-carrier tones. Both types of systems are vulnerable to very strong interferers that could saturate the receiver front end.

Generally speaking, pulse-based UWB systems are simpler than OFDM-based systems, depending on the multipath channel, because the former do not need
IFFT/FFT engines operating at very high rates. However, when the need to detect and avoid interference is included in a pulse-based UWB system, complexity comparisons are not as clear.

Finally, as described in Section 2.5, OFDM-based systems appear to be a first step toward future cognitive designs, as called for by the FCC. This and other areas, like broadband antennas, filters, analog front ends, and ADC/DAC technologies remain areas for continued research.

References

Chapter 3
Integrated Multiple Antenna Ultra-Wideband Transceiver

Stephan ten Brink and Ravishankar Mahadevappa

Abstract Ultra-wideband (UWB) communication systems make efficient use of frequency spectrum from 3.1 to 10.6 GHz by transmitting at low output power, distributed over a large bandwidth. Communication systems operating in the same frequency band experience the presence of a UWB system merely as a small increase of the noise floor, resulting in hardly any link degradation. Two main system designs emerged from the standardization efforts over the past few years. In Impulse Radio (e.g., [1]), the narrowband information is spread over a bandwidth much greater than 500 MHz by means of a high-rate chipping sequence — a method familiar from CDMA cellular radio technology. In Multiband OFDM (MB-OFDM) [2], the bandwidth expansion is achieved by hopping a relatively narrowbanded 528 MHz-OFDM signal over a bandwidth of 1.584 GHz. OFDM is a proven technology in wireless local area network (WLAN) applications [3], and thus, a rather pragmatic approach to UWB. After first single chip implementations have shown the feasibility of physical layer data rates up to 480 Mbps [4], the trend goes to multiple antenna techniques to increase both reliability and data rate of the communication link. The key for cost-effective implementation of multiple antenna techniques is the integration of analog RF and digital baseband on a single chip.

In this chapter, we outline some RF/baseband basics of an integrated MB-OFDM based UWB transceiver and discuss the performance benefits of different multiple antenna techniques.
3.1 Preliminaries of UWB Communication

3.1.1 Power and Bandwidth Efficiency

A communication channel can be characterized by a single parameter, the capacity $C$ [5]. For the complex-valued additive white Gaussian noise (AWGN) channel $y = s + n$, the capacity (in bits/s/Hz) is given by

$$C = \log_2(1 + E_s/N_0).$$  \hspace{1cm} (3.1)

Many useful insights can be gained from this basic model. The term $E_s/N_0$ denotes the signal-to-noise ratio (SNR) per channel symbol, with $E_s$ being the power of the signal $s$ and $N_0 = 2\sigma^2$ being the power of the Gaussian distributed noise $n$. To operate close to capacity, we need to apply channel coding, which adds redundancy to the transmitted message. As long as the information rate $RM$, is smaller than the capacity $C$ of the channel, arbitrarily reliable (i.e., error-free) communication is possible. $R$ stands for the code rate and $M$ for the number of coded bits conveyed per channel use; for example in QPSK, we have $M = 2$. Figure 3.1 shows the Gaussian capacity limit (3.1), which assumes Gaussian distributed signals $s$, together with several information limits for modulation schemes with a finite number

![Fig. 3.1 Ultra-wideband communication in the spectral efficiency chart](image-url)
of amplitude levels, such as 16-QAM and 64-QAM. Note that the rate-normalized SNR is used, i.e.,

\[ \frac{E_b}{N_0} = \frac{(E_s/N_0)}{(RM)}. \] (3.2)

\(E_b\) is the energy needed to reliably convey one bit of information. Using repetition coding (RC) as compared to uncoded transmission allows to operate at lower SNR, \(E_s/N_0\), but there is no coding gain, that is, there is no improvement in power efficiency with respect to an uncoded system. In the figure, a marker is plotted at that \(E_b/N_0\) value when a bit error rate (BER) of \(10^{-5}\) is achieved; indeed, in terms of \(E_b/N_0\), uncoded transmission and a repetition code of rate \(R_{RC} = 1/2\) (i.e., repeating each information bit twice) perform the same over an AWGN channel, while, however, the repetition code reduces spectral efficiency. Convolutional codes (CC) can achieve a coding gain of several decibels (dB) over uncoded transmission, depending on the code rate \(R_{CC}\). The memory 6 convolutional code has become a quasi-industry standard and is in widespread use in WLAN, WPAN/UWB and cellular radio applications. Further coding gain could be obtained by using a higher code memory, which, however, would result in an exponential increase in decoder complexity. The figure shows all data rates of the WiMedia UWB system [2] as well as those data rates of WLAN 802.11 a/g [3] which operate at higher spectral efficiency using 16-QAM and 64-QAM. Observe that data rates of WPAN/UWB below 320 Mbps apply a combination of convolutional and repetition coding, to allow operation at very low SNR. As pointed out above, in terms of the rate-normalized SNR \(E_b/N_0\), there is no coding gain achieved using repetition coding, provided an AWGN channel model is assumed. However, over wireless fading channels, repetition coding can achieve diversity gains, as we will see later in Section 3.5.2, and thus is still a useful extension of the designer’s toolset to expand the rate table.

3.1.1.1 Power-Limited Regime

We are operating at low SNR, or low spectral efficiency (lower left area of chart), when the transmit power is the limited resource. The ultra-wideband channel (3.1–10.6 GHz) falls into this category. Obviously, binary antipodal modulation \(\pm \sqrt{E_s}\) (binary phase shift keying, BPSK, per real-dimension, i.e., quadrature phase shift keying, QPSK, over bandpass channels) is sufficient to get close to the limit. This simplifies the design of an UWB system, putting only mild constraints on a wideband RF front-end with regard to linearity and phase noise requirements, as well as ADC/DAC resolution. For the same spectral efficiency, a higher data rate \(W \cdot C\) can be achieved by just expanding the bandwidth \(W\). A linear increase in bandwidth \(W\) (i.e., a linear increase in transmit power, as the noise power increases linearly with bandwidth as well) provides a linear increase in data rate.
3.1.1.2 Bandwidth-Limited Regime

When not power, but bandwidth is the limited resource, we need to operate at high SNR, or high spectral efficiency (upper mid area of chart), which is the case for, e.g., WLAN operation. To make good use of the available spectrum, we have to apply multi-amplitude modulation, like PAM for real and QAM for complex (bandpass) signals. The RF design is much more demanding, with tight phase noise and linearity constraints; additionally, the ADC/DAC resolution needs to be higher than for QPSK, which results in increased power consumption. Moreover, digital baseband processing (detection, demapping) tends to be more complex. From (3.1), we can see that a linear increase of the transmit power only provides a logarithmic increase in the data rate.

3.1.1.3 The Multiple-Antenna Capacity Promise

We assume a configuration with $N_T$ transmit and $N_R$ receive antennas. The transmitted symbols are $N_T \times 1$ vectors $s$ with entries taken from some complex constellation of size $2^M$ signal points. Each vector symbol carries $N_T M$ bits. The total transmitted power is $E_s$, with energy constraint $E[|s|^2] = E_s/N_T$ per transmit antenna. The received $N_R \times 1$ vector is $y = Hs + n$, where $H$ is the $N_R \times N_T$ complex channel matrix, assumed to be known to the receiver, and $n$ is an $N_R \times 1$ vector of independent zero-mean complex Gaussian noise entries with variance $\sigma^2 = N_0/2$ per real component. As the channel has multiple inputs and multiple outputs, multiple antenna communication is also simply referred to as “MIMO”. Based on (3.2), we define

$$\frac{E_b}{N_0}_{dB} = \frac{E_s}{N_0}_{dB} + 10 \log_{10} \frac{N_R}{RN_T M}.$$  \hspace{1cm} (3.3)

We consider a Rayleigh fading channel, with entries of $H$ being independent complex zero-mean Gaussian random variables with independent real and imaginary parts each having variance $1/2$. We assume an ergodic channel where the channel matrix $H$ changes for every symbol $s$. This is approximated in practice by coding and interleaving over many realizations of $H$, like in multicarrier modulation (OFDM) employing time/frequency interleaving, as done in [2]. The capacity is (e.g. [6])

$$C = E \left[ \log_2 \det \left( I + \frac{E_s}{N_0} \frac{1}{N_T} HH^* \right) \right],$$  \hspace{1cm} (3.4)

where $I$ is the identity matrix and $H^*$ is the complex-conjugate transpose of $H$. A linear increase of the data rate can be achieved by linearly increasing the number of transmit and receive antennas — without increasing the transmit power. Figure 3.1 plots examples for $N_T \times N_R = 2 \times 2$ and $4 \times 4$-channels.
3.1.2 Frequency Hopping, Multiband OFDM

The Federal Communications Commission (FCC) in the United States opened up the unlicensed spectrum from 3.1 to 10.6 GHz for ultra-wideband communication, provided that the following conditions are met:

- The transmitted signal has to occupy either a bandwidth of more than 20 percent of the center frequency or more than 500 MHz to be classified as a UWB signal.
- The power radiated from an isotropic transmit antenna measured in a 1 MHz bandwidth must not exceed $-41.3$ dBm, corresponding to about 0.5 mW if the entire range of 7 GHz was used at the same time.

Owing to the low power, UWB communications is limited to short range, but with its wide bandwidth, high data rates of several Gbps are possible using simple binary antipodal modulation per dimension (QPSK). A multiband approach can keep the implementation complexity manageable: the UWB spectrum is divided into bandgroups (BG) of 1.584 GHz bandwidth, which themselves are further divided into subbands (SB) of 528 MHz bandwidth, as depicted in Fig. 3.2. Rather than using the entire spectrum at the same time, only 528 MHz is used during a symbol period, and thus, the instantaneous bandwidth for ADC/DAC sampling is limited to 528 MHz. To benefit from the large bandwidth available, and to bargain on the fact that not the effective isotropic radiated power (EIRP) but the power spectral density (PSD) is subject to regulatory limitations, frequency hopping among $N_{SB}$ subbands is applied, with $N_{SB}$ ranging from 1 to 3. This allows transmission at $10 \log_{10} N_{SB} \text{ dB}$ higher power as compared to the case without hopping, which directly translates to better range. From Fig. 3.2, we can see that only subbands 3, 9 and 10 are currently available for worldwide operation. Some subbands in Japan and Europe can only be used if a Detect-and-Avoid (DAA) strategy is applied: the UWB device has to detect potential victim systems such as WiMAX and avoid them by nulling

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Fig. 3.2 MB-OFDM spectrum allocation; worldwide compliance
out the respective spectral components in its transmissions. DAA is still in its early development stage and not part of the initial UWB device features. Standardization is ongoing.

3.2 Transmitter for Multiband OFDM

We start with describing the transmitter of an MB-OFDM system. While most of the concepts and effects are of a general nature, we keep to the nomenclature and system parameters of the WiMedia UWB system as outlined in [2], which is based on a single transmit antenna. We later show the advantages of using multiple transmit and receive antennas, either as compatible extensions to the existing standard or as viable options for a possible next-generation system with higher data rates and increased range.

3.2.1 Frame Format

The frame (or packet) format is shown in Fig. 3.3. The first 24 OFDM symbols are packet synchronization (PS) symbols which are used for packet detection, frame synchronization, frequency offset estimation and other synchronization tasks at the receiver (Section 3.4.2). The 6 channel estimation (CE) symbols are used as training sequences to estimate the channel impulse response, facilitating coherent detection. PS and CE symbols are commonly referred to as “packet preamble”. The 12 header symbols carry information, such as data rate and length of the payload part of the frame, while the payload carries the data content to be conveyed to the higher protocol layers. The center frequency is changed after each OFDM symbol, according to a “time-frequency code” (TFC), specifying the hopping pattern. The figure shows TFC 1, which just cycles through the three 528 MHz subbands per bandgroup. Each OFDM symbol is composed of 128 complex time-domain samples generated by a 128-point inverse fast Fourier transform (iFFT), 32 zero samples as a null-suffix and 5 zero samples as guard time to accommodate hopping transients. The sampling frequency assumed is 528 MHz, so each OFDM symbol has a period of $T_{sym} = 312.5 \text{ ns}$. Note that in most wireless systems based on OFDM, a cyclic suffix (or prefix) is used to cope with delay spread: the beginning of the iFFT output
at the transmitter is copied to the end, thus generating a periodic extension of the symbol; the receiver can then freely choose the position of the FFT window. The main reason for using a null-suffix in the UWB system at hand is that the spectral ripple of the transmitted signal is smaller compared to systems with cyclic suffix: the cyclic suffix introduces periodicity, which results in a ripple in the transmit spectrum. Since UWB is limited by FCC with regard to power-spectral density (PSD), no portion of the transmit spectrum must exceed $-41.3 \, \text{dBm/MHz}$, and thus, any spectral ripple would lead to reduced output power, i.e., less range. It is interesting to note that spectral ripple is of less concern in WLAN transmission, as the regulatory limitation is EIRP-based there.

### 3.2.2 Digital Baseband

The bit stream from the MAC (media access control) is encoded using a convolutional code of memory $\nu$ (i.e., $\nu$ registers), adding redundancy to facilitate the detection and correction of errors at the receiver. In the WiMedia UWB system, the memory is $\nu = 6$, and the generator polynomials of the rate $R_{CC} = K/N = 1/3$ mother code are $G_1(D) = 1 + D^2 + D^3 + D^5 + D^6$, $G_2(D) = 1 + D + D^2 + D^4 + D^6$ and $G_3(D) = 1 + D + D^2 + D^3 + D^6$, or, in the conventional short octal notation, $g_1 = 133_8$, $g_2 = 165_8$ and $g_3 = 171_8$, respectively. All other coding rates ($R_{CC} = 1/2$, $5/8$ and $3/4$) are obtained by appropriate puncturing, i.e., skipping some of the encoder output bits according to a predefined puncturing pattern. The encoded bit stream is interleaved by a symbol and a tone interleaver. The symbol interleaver permutes the coded bits over 6 OFDM symbols, while the tone interleaver performs a cyclic shift of the coded bits within a single OFDM symbol. The symbol interleaver greatly contributes to robustness over multipath fading channels, particularly in frequency hopping mode.

Depending on the data rate, the permuted bits from the interleaver are mapped onto a QPSK or a dual-carrier modulation (DCM) constellation, as summarized in Table 3.1. The possible high-rate extensions 512–1024 Mbps are not specified in [2]. To achieve lower coding rates than provided by the convolutional code, repetition coding over time and frequency is used. This is implemented as “time spreading”, i.e., repetition of OFDM symbols in time, and “complex conjugate symmetrical spreading”, which copies the QPSK symbols of the lower 50 data subcarriers as complex conjugate into the upper 50 data subcarriers within each OFDM symbol.

In DCM, the four coded bits $c_i \in \{0,1\}, i = 0,\ldots,3$ are mapped onto two 16-QAM symbols according to

$$
\begin{bmatrix}
s_0 \\
s_1
\end{bmatrix} = \frac{1}{\sqrt{10}} \begin{bmatrix} 2 & 1 \\ 1 & -2 \end{bmatrix} \begin{bmatrix} x_0 + jx_1 \\ x_2 + jx_3 \end{bmatrix},
$$

(3.5)

with $x_i = 2c_i - 1; x_i \in \{-1,+1\}; i = 0,\ldots,3$. This results in the two complex 16-QAM symbols $s_0, s_1$ per four bits $c_0, \ldots, c_3$, corresponding to a rate 1/2 repetition code over constellation points. Thus, DCM has the same spectral


Table 3.1 Modulation and coding rate table with 16-QAM extension

<table>
<thead>
<tr>
<th>Entry</th>
<th>Data rate</th>
<th>Modulation</th>
<th>Conj. sym. spread.</th>
<th>Time spread.</th>
<th>Code rate</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>53.3</td>
<td>QPSK</td>
<td>Yes</td>
<td>Yes</td>
<td>1/3</td>
</tr>
<tr>
<td>1</td>
<td>80</td>
<td>QPSK</td>
<td>Yes</td>
<td>Yes</td>
<td>1/2</td>
</tr>
<tr>
<td>2</td>
<td>106.7</td>
<td>QPSK</td>
<td>No</td>
<td>Yes</td>
<td>1/3</td>
</tr>
<tr>
<td>3</td>
<td>160</td>
<td>QPSK</td>
<td>No</td>
<td>Yes</td>
<td>1/2</td>
</tr>
<tr>
<td>4</td>
<td>200</td>
<td>QPSK</td>
<td>No</td>
<td>Yes</td>
<td>5/8</td>
</tr>
<tr>
<td>5</td>
<td>320</td>
<td>DCM</td>
<td>No</td>
<td>No</td>
<td>1/2</td>
</tr>
<tr>
<td>6</td>
<td>400</td>
<td>DCM</td>
<td>No</td>
<td>No</td>
<td>5/8</td>
</tr>
<tr>
<td>7</td>
<td>480</td>
<td>DCM</td>
<td>No</td>
<td>No</td>
<td>3/4</td>
</tr>
<tr>
<td>8</td>
<td>512</td>
<td>DCM</td>
<td>No</td>
<td>No</td>
<td>4/5</td>
</tr>
<tr>
<td>9</td>
<td>640</td>
<td>16-QAM</td>
<td>No</td>
<td>No</td>
<td>1/2</td>
</tr>
<tr>
<td>10</td>
<td>800</td>
<td>16-QAM</td>
<td>No</td>
<td>No</td>
<td>5/8</td>
</tr>
<tr>
<td>11</td>
<td>960</td>
<td>16-QAM</td>
<td>No</td>
<td>No</td>
<td>3/4</td>
</tr>
<tr>
<td>12</td>
<td>1024</td>
<td>16-QAM</td>
<td>No</td>
<td>No</td>
<td>4/5</td>
</tr>
</tbody>
</table>

Efficiency as QPSK, which is 2 bits/s/Hz, but can exploit frequency diversity over fading channels. As a possible high-rate extension, we can further use a true 16-QAM modulation, to increase the spectral efficiency to 4 bits/s/Hz per subcarrier, which doubles the data rate up to 960 Mbps, as indicated in Fig. 3.4 (dashed lines).

After mapping, the symbol stream is multiplexed into groups of 128 symbols $s_k$ (or subcarriers $-64$ to $63$ around DC), with expected value normalized to $E[|s_k|^2] = 1$, forming one OFDM symbol. Out of the 128 subcarriers,

- 12 subcarriers $k \in \{-55, -45, -35, \ldots, 45, 55\}$ are used as dedicated pilots to allow phase tracking at the receiver; they are QPSK-modulated by a pseudo-random sequence;
- 10 guard subcarriers $k \in \{-61, \ldots, -57, 57, \ldots, 61\}$, that carry redundant information, and 6 null subcarriers (including DC) $k \in \{-64, -63, -62, 0, 62, 63\}$, at the band-edges that carry no information, provide some frequency guard band between subbands;
- the remaining 100 subcarriers convey the interleaved and QAM-modulated coded bits coming from the convolutional encoder.

For each OFDM symbol, an inverse discrete Fourier transform provides the discrete-time signal $x_n, n = 0, \ldots, 127$,

$$
    x_n = \frac{1}{\sqrt{N_{\text{FFT}}}} \sum_{k=0}^{N_{\text{FFT}}-1} s_k \cdot e^{j2\pi kn/N_{\text{FFT}}}
$$

typically implemented using an inverse fast Fourier transform (iFFT), with $N_{\text{FFT}} = 128$. After adding a 37 sample null-suffix, the sequence $x_n$ is upsampled by 2, i.e.,
\((x_{2n}', x_{2n+1}') = (x_n, 0)\). Figure 3.4 further illustrates these steps. After FIR-filtering using, e.g., a raised cosine filter of the form

\[
h_{TX,n} = \frac{\sin (\pi n)}{\pi n} \cdot \frac{\cos (\pi \alpha n)}{1 - 4\alpha^2 n^2}
\]

with roll-off \(\alpha = 0.1\), the discrete-time complex baseband signal is

\[
x_{BB,n} = x'_n \times h_{TX,n}.
\]

The baseband signal is converted to analog domain using a digital-to-analog converter (DAC) with sampling period \(T_{s,DAC} = 1/1.056 \text{GHz} \approx 0.947 \text{ ns}\). The DAC outputs a continuous-time signal \(x_{BB}(t) = x_I(t) + jx_Q(t)\) with bandwidth 1.056 GHz. The upsampling by 2 (from 528MHz to 1.056GHz) is needed to relax the RF filter constraints.
The up-conversion to passband with carrier frequency $f_c$ can be written as

$$x_{PB}(t) = \text{Re} \left\{ x_{BB}(t) \cdot e^{j2\pi f_c t} \right\} = x_I(t) \cdot \cos(2\pi f_c t) - x_Q(t) \cdot \sin(2\pi f_c t)$$

with in-phase baseband component $x_I(t)$ and quadrature component $x_Q(t)$, respectively. Frequency hopping according to a TFC pattern is done on a per-OFDM symbol basis, changing $f_c$ every $T_{\text{sym}} = 312.5$ ns. For example, for BG 1 and TFC 1, the hopping is between frequencies $f_c = 3.432, 3.960, \text{and } 4.488 \text{ GHz}$.

### 3.2.3 Analog RF

The signal is up-converted to passband, amplified and radiated over the transmit antenna. Only 9 ns out of the total of 312.5 ns symbol duration are reserved for frequency hopping transients, making direct frequency synthesis by phase locking impractical. Rather, three separate RF blocks are implemented, one for each sub-band per bandgroup, to allow for fast frequency hopping [7]. A band-select signal controls the subband selection from the digital baseband, to follow the hopping pattern of the respective time-frequency code. Each RF block has its own tuned amplifier stage, I/Q mixers and integer-$N$ frequency synthesizer. A block diagram of the direct-conversion RF transceiver is shown in Fig. 3.7.

### 3.2.4 Multiple Transmit Antennas

Possible high-rate extensions are shown in Fig. 3.4 as dashed lines. Data rate and link quality can be further improved by using multiple antennas. In orthogonal space–time block coding, commonly referred to as transmit diversity, QAM symbols are repeated over transmit antennas (“space”) and in time using specifically designed orthonormal (unitary) matrices. Optimal detection at the receiver can be done by applying simple linear combination on each receive antenna. This method is mainly for increasing range and robustness of the transmission. In spatial multiplexing, different data is simultaneously transmitted over different transmit antennas, thus increasing spectral efficiency (data rate). Optimal detection is more complex in this case. Several combinations of both methods are possible, as indicated in the figure. Details on these modes are discussed separately, in Section 3.5, which is dedicated to multiple antenna techniques. For all high-rate extensions (16-QAM, multiple antennas), it is suggested to break down the single high-speed convolutional encoder into several low-rate ones, in particular to reduce the implementation burden of the channel decoder at the receiver.

### 3.3 Channel Model

A channel model is used to evaluate system performance, in particular, the receiver algorithms for packet detection, demodulation, deinterleaving and channel decoding. For the ultra-wideband channel ranging from 3 to 10 GHz, a modified version
of the Saleh–Valenzuela multipath model for indoor channels [8] was adopted. The multipath arrivals are grouped into \( L \) clusters, with each cluster being composed of \( M \) multipath components. The continuous-time channel impulse response is modeled as

\[
h(t) = \sum_{l=0}^{L-1} \sum_{m=0}^{M-1} \alpha_{l,m} \delta(t - T_l - \tau_{l,m}).
\] (3.6)

The arrival times \( T_l \) and \( \tau_{l,m} \) are exponentially distributed random variables, and the magnitudes of the complex gains \( \alpha_{l,m} \) are log-normally distributed. More details can be obtained from [9], which further specifies four typical channel models: the line-of-sight (LOS) channel CM1 and non-line-of-sight (NLOS) channel CM2 describe distances from 0 to 4 m, while the two NLOS channels CM3 and CM4 represent propagation environments as encountered at distances from 4 to 10 m. The duration of a packet ranges from 13.125 (zero-length payload) to 628.2 \( \mu \)s (53.3 Mbps with 4095 bytes payload); as variations of indoor propagation conditions are typically on the order of several milliseconds, the channel is assumed to stay constant over the duration of an entire frame.

For convenience, we stay in the discrete-time complex baseband domain. From (3.6), we infer that the multipath channel can be represented by a tapped delay line \( h_n \) (FIR filter) of sufficient length. For each model CM1–4, 100 time-domain impulse responses were selected [9], representing typical examples which are used for system simulation and optimization of baseband algorithms. The channel output \( s_{CH, n} \) is a convolution of transmitted signal and channel impulse response \( h_n \),

\[
s_{CH, n} = s_{BB, n} \times h_n.
\]

Figure 3.5 shows examples of such time-domain impulse responses \( h_n \) for three different channels (left), with the two multipath channels CM2, CM4, and, as a reference, the channel CM0 with delay profile \( h_n = \delta_n \), representing a non-multipath transmission over, e.g., cable-links. The function \( \delta_n \) is 1 for \( n = 0 \), and 0 for \( n \neq 0 \). The right chart shows the corresponding power spectral density (PSD, simply referred to as spectrum) of the transmitted signal after convolution with \( h_n \). The PSD corresponding to CM0 is, essentially, the transmit spectrum, normalized to 0 dB maximum magnitude. The three subbands and respective null- and DC-subcarriers can clearly be identified, spanning over a bandwidth of 1.584 GHz. For CM2, the delay spread due to multipath propagation is around 30 ns, causing fluctuations and deep notches in the PSD. The receiver has to cope with these effects by appropriate channel estimation and deinterleaving of coded bits across subcarriers and subbands, to avoid bursts of errors entering the Viterbi decoder. Compared to single carrier systems which would require complex multi-tap equalization, or several rake fingers (direct sequence spread spectrum systems), OFDM receivers can efficiently capture the multipath energy. The delay spread for CM2 is still below the duration of the null-suffix of 32 samples (60.6 ns), and there is hardly any inter-symbol interference; the OFDM receiver can easily compensate for the channel fluctuations. For CM4,
Fig. 3.5 Channel impulse responses CM0, CM2 and CM4 in time-domain (magnitude, left) and after convolution with transmit signal in frequency-domain (right)

The duration of the impulse response (around 90 ns) significantly exceeds the length of the null-suffix, and inter-symbol interference further degrades the signal quality. The spectrum has many more nulls compared to CM2.

### 3.3.1 Multiple Antenna Channel

For communication using $N_T$ transmit and $N_R$ receive antennas, the channel can be modeled by an $N_R \times N_T$ matrix $H$, with each entry now by itself being a multipath channel $h_{j,i}(t)$ from antenna $i$ to antenna $j$ according to (3.6), with different realizations of the random variables for delay and gain values, distributed as outlined before.

### 3.4 Receiver

In this section, we discuss an ultra-wideband receiver architecture for MB-OFDM. There are several distinct features in the RF and digital baseband section that take care of the wideband nature of the signal, and which are different from conventional narrowband receivers as used for, e.g., WiMAX or WLAN services. We first focus on the single antenna case, before going into multiple antenna extensions in Section 3.5.

With reference to Fig. 3.6, the receiver is composed of an analog RF section, a “preamble processing” part and a “data processing” part. In the RF section, the
received signal from the antenna is amplified and down-converted from passband to baseband. The signal processing part performs initial time- and frequency-synchronization tasks such as packet detection, frame synchronization and frequency offset estimation using the preamble of the received packet. The data processing part detects and decodes the bit information in the header and payload of the packet.

3.4.1 Analog RF

The wideband signal from the antenna is amplified using a low noise amplifier (LNA), with the gain controlled in three coarse levels of 10, 21 and 27 dB (Fig. 3.7). The mixer performs a direct down-conversion of the three subbands of bandwidth 528 MHz from RF to baseband. The mixer gain can be controlled in four coarse
levels given as $-13$ (actually, an attenuation), $-7$, $-1$ and $+5$ dB (amplification). All three subbands of the same bandgroup are amplified using the same LNA and mixer gains. Before analog-to-digital conversion, a programmable gain amplifier (PGA) further amplifies the signal. As different frequency bands may experience different RF path loss and fading characteristics on the wireless channel, the PGA gains can be controlled separately for each subband to account for these variations. The possible PGA gains range from 6 to 30 dB, in steps of 2 dB. All initial gains are set to maximal, i.e., LNA gain 27 dB, mixer gain 5 dB and PGA gain 30 dB to allow the detection of very weak signals (maximal sensitivity). These values are then refined iteratively by the automatic gain control (AGC) algorithm in digital baseband, using coarse and fine AGC steps, to regulate the ADC output to about 5% clipping rate (see also Section 3.4.2.2). The band-select signal from the digital baseband controls the frequency hopping by selecting different subbands according to the respective hopping pattern.

### 3.4.2 Preamble Processing

In this section, we discuss processes occurring at the receiver during the preamble, prior to demodulation and decoding of a packet. At the start of a typical receive operation, the analog front-end is tuned to the required band, and raw complex samples from the analog-to-digital converter are delivered to the digital baseband processor at Nyquist rate (528 MHz). The packet detection process is triggered first to identify the start timing of a valid packet. Further processing consisting of
automatic gain control (AGC), frame synchronization, frequency offset estimation and channel estimation are triggered at appropriate times after a packet is detected. The estimates obtained in the process are used in demodulating and decoding the header and payload information from the received samples. Thus, an error in packet detection or inaccuracies in these estimates affect all subsequent processes, causing significant degradation in link quality. Hence a robust preamble processing unit is essential. We start this section with a brief review of the channel model and the notation, following which the various processes involved are described, each under a separate heading.

The physical wireless channel can be modeled as a linear time-invariant (LTI) multipath channel and the noise process modeled as AWGN, as outlined in Section 3.3. Typically, UWB transmissions are short (the longest being \( \sim 600 \mu s \)), and the transmitter and receiver are generally static, so a LTI model serves adequately. The received signal can then be expressed in discrete-time domain as

\[
y_n = \sum_{p=0}^{L} h_p x_{n-n_0-p} + w_n, \quad n = 0, 1, \ldots, N - 1, \quad (3.7)
\]

where the values of \( x_n \) form the transmitted signal, \( w_n \) are i.i.d. noise samples with real and imaginary parts following the distribution \( \mathcal{N}(0, \sigma^2) \) and \( n_0 \) denotes the starting point of the transmission. In this section, we will use \( x_n \) to denote a complex value, \( x(m) \) to denote the \( m \)th vector of length \( N_{\text{sym}} = 165 \) which is one symbol duration, and \( x \) to denote the entire vector including all symbols,

\[
x(m) = [x_k x_{k+1} \cdots x_{k+N_{\text{sym}}-1}]^T, \quad k = mN_{\text{sym}} - n_0 \quad (3.8)
\]

\[
x = [x^T(0) x^T(1) \cdots x^T(N_{\text{pkt}} - 1)]^T \quad (3.9)
\]

assuming the transmission has \( N_{\text{pkt}} \) symbols.

As described in Section 3.2, the transmitted vector \( x \) consists of a preamble, header and payload. The preamble has two parts, as shown in Fig. 3.8, the first part consisting of PS sequences modulated by a cover sequence used generally for packet detection, timing acquisition, AGC, frequency offset estimation and frame synchronization, while the second part consists of 6 symbols used typically for channel estimation.

Let \( u \) denote the PS vector of length \( N_{\text{FFT}} \) and \( u_{\text{ext}} \) the extended PS sequence of length \( N_{\text{sym}} \) including the null-suffix. If the cover sequence is denoted by \( c_m \), then the preamble is given by

\[
x(m) = \begin{cases} 
  c_m u_{\text{ext}} & m = 0, 1, \ldots, N_{\text{PS}} - 1 \\
  v_{\text{ext}} & m = N_{\text{PS}}, \ldots, N_{\text{PS}} + 5,
\end{cases} \quad (3.10)
\]

where \( v_{\text{ext}} \) is the extended vector used for channel estimation, specified in [2]. The WiMedia MBOA specification supports two types of preambles: a standard (or long) preamble where \( N_{\text{PS}} \) is 24, and a streaming (or short) preamble where \( N_{\text{PS}} \) is 12. The
latter is used to reduce overhead due to long preamble when the link is known to be stable and thus increase throughput. In streaming mode, some of the parameters estimated from a previous packet could be used as coarse estimates and some refinements may be made using the short preamble.

Figure 3.8 also depicts an example partitioning of time available between the various processes. In practice, the partitioning could vary from one packet to another depending on which symbol in the preamble actually crosses the threshold set in the packet detection algorithm. Generally, this type of variance is seen more in low SNR conditions. The number of symbols used for AGC, frame synchronization, etc. could be further dependent on the TFC used. For example, when a band-hopping TFC is used, the receiver requires more symbols for AGC since there could be signal strength variations across bands, and the gain for each band has to be set appropriately. When a streaming preamble is expected, the receiver could use a timing reference obtained from the previous packet and just confirm the presence of a new preamble at the expected time. It could also skip coarse AGC process and use the gain settings from the previous packet as initial values.

### 3.4.2.1 Packet Detection

The receiver typically uses the packet synchronization (PS) sequences in the preamble for packet detection. The WiMedia UWB standard allows 10 logical channels (TFCs) to operate within a bandgroup and in order to effectively distinguish between channels and avoid false detections, each channel is assigned a specific PS sequence. These sequences are designed to have good auto-correlation and cross-correlation properties to enable reliable detection. We shall next describe a packet detection algorithm and quantify its performance in terms of the correlation properties of the PS sequences. The reader may find several such algorithms and more detailed analyses in texts such as [10].

Let $y_n$ denote the vector $[y_{n−N_{FFT}−L+1} \ y_{n−N_{FFT}−L+2} \ \cdots \ y_n]^T$ consisting of the most recent $(N_{FFT}+L)$ received samples, $h$ denote the vector of channel coefficients $[h_0 \ h_1 \ \cdots \ h_L]^T$ and $U$ a matrix of order $(N_{FFT} + L) \times (L + 1)$, constructed by arranging shifted versions of the expected PS sequence as
\[
U = \begin{bmatrix}
    u_0 & 0 & \cdots & 0 \\
    u_1 & u_0 & \cdots & 0 \\
    \vdots & \vdots & \ddots & \vdots \\
    u_L & u_{L-1} & \cdots & u_0 \\
    \vdots & \vdots & \vdots & \vdots \\
    0 & 0 & \cdots & u_{N_{\text{FFT}}-1}
\end{bmatrix}.
\] (3.11)

Then, the detection problem is essentially one of identifying which of the two hypotheses
\[
\mathcal{H}_0 : y_n = w_n \quad \text{and} \quad \mathcal{H}_1 : y_n = Uh + w_n,
\] (3.12)
is valid for every new sample \( y_n \) received, until a decision is made in favor of \( \mathcal{H}_1 \). When there is no multipath (\( L = 0 \)) and the SNR is high, this should happen at \( n = n_0 + (N_{\text{FFT}} - 1) \), when the first PS symbol in the preamble is fully received. However, when there is significant multipath and the channel coefficients are not known, the start time estimate at the receiver may not be precisely determined and could have a small offset with respect to \( n_0 \). Further, when the SNR is low, the first symbol may not always trigger the decision for \( \mathcal{H}_1 \), in which case the symbol counter will have an offset. This offset could be corrected by the frame synchronization process which is discussed later in this section.

In mathematical terms, we need to determine the region, say \( R_1 \), where the vector \( y_n \) should lie, to make a decision in favor of \( \mathcal{H}_1 \). The metrics often considered for optimization in determining \( R_1 \) are Probability of False Detection \( P_{\text{false}} \), Probability of Missed Detection \( P_{\text{miss}} \) or the Probability of Error \( P_{\text{error}} \) defined as
\[
P_{\text{false}} = \Pr(y_n \in R_1 | \mathcal{H}_0) \quad \text{(3.13)}
\]
\[
P_{\text{miss}} = \Pr(y_n \in R_0 | \mathcal{H}_1) \quad \text{(3.14)}
\]
\[
P_{\text{error}} = \Pr(\mathcal{H}_0)P_{\text{false}} + \Pr(\mathcal{H}_1)P_{\text{miss}}. \quad \text{(3.15)}
\]

The optimal decision rule, which minimizes \( P_{\text{miss}} \) for a given \( P_{\text{false}} \), is obtained by applying the Neyman–Pearson method [10], and is given by
\[
R_1 = \left\{ y_n : \frac{p(y_n, \mathcal{H}_1)}{p(y_n, \mathcal{H}_0)} \geq \lambda \right\}, \quad \text{(3.16)}
\]
where \( \lambda \) is determined by
\[
P_{\text{false}} = \int_{R_1} p(y_n | \mathcal{H}_0) \, dy_n. \quad \text{(3.17)}
\]

However, due to practical issues such as unknown channel coefficients and noise variance, we cannot use the above rule directly. Therefore, the unknown parameters
are replaced by corresponding Maximum Likelihood Estimates (MLE), and a Generalized Likelihood Ratio Test (GLRT) [10, Thm. 9.1] is formulated. Given \( y_n \), the MLE \( \hat{h} \) and \( \hat{\sigma}^2 \) under hypotheses \( \mathcal{H}_0 \) and \( \mathcal{H}_1 \) are found to be

\[
\mathcal{H}_0 : \hat{h}_0 = 0 \quad \hat{\sigma}_0^2 = \frac{1}{N_{\text{FFT}} + L} \| y_n \|^2 \tag{3.18}
\]

\[
\mathcal{H}_1 : \hat{h}_1 = (U^*U)^{-1}U^*y_n \quad \hat{\sigma}_1^2 = \frac{1}{N_{\text{FFT}} + L} \| y_n - U\hat{h}_1 \|^2. \tag{3.19}
\]

Substituting for \( \hat{h} \) and \( \sigma^2 \) in the likelihood ratio, we get

\[
L_G(y_n) = \frac{p(y_n, \mathcal{H}_1; \hat{h}_1, \hat{\sigma}_1^2)}{p(y_n, \mathcal{H}_0; \hat{h}_0, \hat{\sigma}_0^2)} = \left( \frac{y_n^*y_n}{y_n^* (I - U(U^*U)^{-1}U^*) y_n} \right)^{(N_{\text{FFT}} + L)/2}. \tag{3.20}
\]

We can alternatively use a simplified decision statistic which is a monotonically increasing function of the likelihood ratio,

\[
T(y_n) = 1 - (L_G(y_n))^{-2/(N_{\text{FFT}} + L)} = \frac{y_n^*U(U^*U)^{-1}U^*y_n}{y_n^*y_n}, \tag{3.21}
\]

and express the decision rule as

\[
\text{GLRT} : \mathbb{R}_1 = \left\{ y_n : \frac{y_n^*U(U^*U)^{-1}U^*y_n}{y_n^*y_n} \geq \lambda \right\}, \tag{3.22}
\]

where \( \lambda \) is determined by fixing the false alarm rate \( P_{\text{false}} \). The start time estimate, \( \hat{n}_0 \), is given by

\[
\hat{n}_0 = n |_{T(y_n) > \lambda} - N_{\text{FFT}} - L, \tag{3.23}
\]

where \( n |_{T(y_n) > \lambda} \) denotes the sample index where \( T(y_n) \) crosses the threshold for the first time.

The term \( U^*y_n \) in (3.22) essentially consists of cross-correlation values between \( y_n \) and the PS sequence \( s \) at different shifts of \( y_n \). If the PS sequence is well designed, in the sense that the auto-correlation function is close to an impulse, then the matrix \( U^*U \) can be approximated by an identity matrix with possibly a scalar multiplier, and the numerator reduces to a moving sum of correlation energy

\[
T(y_n) \approx \frac{\sum_{p=0}^{L} |r_{n-p}|^2}{y_n^*y_n}; \quad r_n = \frac{1}{\| u \|} \sum_{k=0}^{N_{\text{FFT}}-1} u_k^* y_n - N_{\text{FFT}} + 1 + k. \tag{3.24}
\]
The denominator is an estimate of the received sequence energy and may be implemented using an accumulator. Note that the decision statistic is invariant to scaling of $y_n$, resulting in a false alarm probability which is independent of the SNR.

One other unknown in the model is the parameter $L$, the delay spread of the channel. Theoretically, it can be estimated using the correlator output $r_n$, for instance, by applying a threshold over $|r_{n-p}|$ as follows,

$$
\hat{L} = \max \left\{ p : |r_{n-p}| \geq \lambda_L, \ p < L_{\text{max}} \right\}.
$$

$L_{\text{max}}$ here is a maximum limit on the delay spread, determined by the expected range of the system. In practice, $\hat{L}$ could be set to $L_{\text{max}}$.

The decision statistic of (3.22) is a ratio of two $\chi^2$ random variables which are central under hypothesis $H_0$ and non-central under hypothesis $H_1$. While it is possible to derive an expression for $P_{\text{false}}$ with some effort, $P_{\text{miss}}$ has to be computed numerically or from simulations. If we assume $L + 1$ taps for the decision statistic, we have

$$
P_{\text{false}} = (1 - \lambda)^{N_{\text{FFT}} - 1} \sum_{p=0}^{L} \left( \frac{N_{\text{FFT}} - 2 + p}{p} \right) \lambda^p,
$$

which is plotted in Fig. 3.9 for the case $N_{\text{FFT}} = 128$. Simulation results for $P_{\text{miss}}$ are plotted in Fig. 3.10, using the PS sequence of TFC 1 for a fixed SNR.

The parameter $L$ can be used along with $\lambda$ to configure the system in order to achieve the desired performance. Increasing $L$ could result in a lower $P_{\text{miss}}$ but a higher $P_{\text{false}}$.

The decision statistic in (3.22) can be extended to the case where multiple RF receiver modules provide independent observations which are used to improve

![Fig. 3.9 Probability of false alarm vs threshold $\lambda$ for various values of $L$](image)
Fig. 3.10 Simulation results for $P_{\text{miss}}$ vs SNR for various values of $L$ and fixed $\lambda$.

performance. If $y_{n,1}$ and $y_{n,2}$ are the observations from two receive modules, it can be shown that the GLRT decision statistic is of the form

$$T(y_{n,1}, y_{n,2}) = \frac{y_{n,1}^* U(U^* U)^{-1} U^* y_{n,1} + y_{n,2}^* U(U^* U)^{-1} U^* y_{n,2}}{y_{n,1}^* y_{n,1} + y_{n,2}^* y_{n,2}}.$$ \hspace{1cm} (3.27)

This decision statistic, however, is more stringent than desired. If one of the RF receivers has more noise or interference which results in low correlation values, then the packet is not detected, although the receiver could decode the payload if the SNR on the second RF receiver was sufficient. A less stringent rule which alleviates this problem is

$$R_1 = \{(y_{n,1}, y_{n,2}) : (T(y_{n,1}) \geq \lambda) \text{ OR } (T(y_{n,2}) \geq \lambda)\},$$ \hspace{1cm} (3.28)

where $T(y_n)$ is as defined in (3.21). For the general case of $N_R$ receive antennas, the rule can be extended easily by using the OR operation to combine the individual decisions. The performance of such a detector can be derived using $P_{\text{false}}$ and $P_{\text{miss}}$ for the single antenna case. The relevant probabilities are given by

$$P_{\text{miss}}(N_R) = \left(1 - (1 - P_{\text{miss}})^{N_R}\right),$$ \hspace{1cm} (3.29)

$$P_{\text{false}}(N_R) = 1 - (1 - P_{\text{false}})^{N_R}.$$ \hspace{1cm} (3.30)

Adding a second antenna and RF receiver module can improve the probability of detection; the gains that can be realized depend on the threshold used and the operating SNR. Note, however, that $P_{\text{false}}$ also increases with $N_R$, so the threshold $\lambda$ may need to be increased to retain the same $P_{\text{false}}$ which will, in turn, reduce the gain. In Fig. 3.11, the probability of detection for $N_R = 2$ is compared with that for a single antenna, after fixing $\lambda$ such that $P_{\text{false}}$ is approximately the same in both cases. We can see that the second antenna improves performance provided the SNR is sufficiently high. Since the minimum operating SNR for a WiMedia
Fig. 3.11 $P_{\text{miss}}$ vs SNR for 1 and 2 receive antennas

system is approximately $-2.5$ dB, which reduces to about $-4.5$ dB with MRC (see Section 3.5.2), we see that the improvement is about 1.5 dB at $P_{\text{miss}} = 10^{-3}$ and increases with SNR in the range of interest, for the channel considered.

3.4.2.2 Automatic Gain Control

The second process that is triggered in the receiver, immediately after a preamble is detected, is AGC. The objective of AGC is to set the gain of the RF module appropriately such that the input to the ADC exercises its full dynamic range, thus maximizing the signal to quantization noise ratio (SQNR). Since the transmitted signal can be significantly attenuated by the time it reaches the receive antenna, the RF module requires programmable gain amplifiers providing gains in the range of 20 to 80 dB. This gain is often distributed across various stages as shown in Fig. 3.7, with the bulk of it provided by the LNA which also acts as a wideband filter, some of it provided by the mixer and the remaining provided by PGAs in the baseband stage. Once a packet is detected and the symbol boundary identified, the receiver could use the methods outlined below to determine whether the gain should be increased or decreased.

Two useful parameters which indicate whether the gain is appropriate are energy estimates and a histogram of ADC values. Energy estimates are used in packet detection as well, so there is no additional complexity in implementing it. However, the linear range of energy estimates obtained from ADC values could be limited, requiring multiple iterations to converge to the correct gain setting. In some RF modules, a Receive Signal Strength Indicator (RSSI) module which measures energy at the baseband input provides a more accurate estimate with a wider linear range. In either case, the energy estimate is converted to an input power estimate by subtracting the current gain, which is then used to find a suitable new gain setting from a look-up table.

The histogram approach can be used to further fine-tune the gain settings, so as to achieve a particular clipping rate. In this method, depicted in Fig. 3.12, the receiver
counts the number of ADC samples crossing certain thresholds over a symbol duration and increments or decrements the gain depending on the counter values. The desired count values for ideal gain and the step size for the gain are predetermined from simulations based on the PS sequences.

AGC for each RF module can be done independently, once the packet timing is acquired.

### 3.4.2.3 Frame Synchronization

The frame synchronization process determines the correct OFDM symbol numbering within the packet, setting the symbol counter appropriately to identify the start of the channel estimation symbols, the header symbols and the payload. When the operating SNR is low or the packet detection threshold is set high, some of the PS symbols may not trigger a packet-detected decision. In such a case, the symbol counter set by the packet detection module has an offset with respect to the correct counter. This offset, however, can be corrected using the cover sequence information embedded in the preamble.

The cover sequence, as described earlier, is a binary (±1) sequence which modulates the PS symbols in the preamble. To match this sequence at the receiver, we need to extract the phase of each symbol. The received signal will have a random phase offset due to the physical channel, random phase noise added in the RF module and frequency offset (within ±40 ppm, see Section 3.4.2.5) with respect to the transmitted signal, in addition to AWGN, so the phase of a received symbol will not exactly match with that of the transmitted symbol. A constant phase offset can be easily handled by taking the phase change between symbols. Frequency offset introduces additional phase offset which can be quite significant when computing phase change between two symbols that are well separated in time,

\[
\Delta \phi_f = 2\pi(\Delta f_{\text{center}})D/f_s,
\]
where $D$ is the time delay between the two symbols in number of samples, and $f_s$ is the sampling frequency. For example, the phase offset induced by a 40 ppm frequency offset over 3 symbols duration is $\sim 46^\circ$ which is quite significant and can cause a false synchronization when coupled with low SNR and high phase noise conditions. However, frequency offset is systematic and generally constant, so the receiver can estimate and correct the phase offsets as described in Section 3.4.2.5.

The block diagram in Fig. 3.13 depicts an algorithm which is found to be quite robust against these effects. Complex correlator outputs $r_n$ from the packet detection process are used here to obtain the phase changes since they have relatively higher SNR compared to the raw ADC samples. A second correlator operates on $r_n$ to accumulate energy which may be spread over several samples by the channel to give

$$z_n = \sum_{n \in W_{pk}} r_n r_n^* - D,$$

(3.31)

where $D$ is a suitable integer number of symbol periods (see Table 3.2), and $W_{pk}$ denotes a window of appropriate length around the peak. The result is sampled at the end of every symbol and the phase of such samples is used to compute phase changes. Note that the phase of $z_n$ is already a difference between phase of correlator outputs at time $n$ and $n - D$, so the change in phase of $z_n$ (given by phase of $z_n z_n^* - D$) is a second-order phase difference. This is done to cancel out phase offsets due to frequency offset as follows:

First-order phase difference : $\Delta \phi(m) \approx \Delta \phi_f + \Delta \phi_{\text{cover}}(m)$

(3.32)

Second-order phase difference : $\Delta^2 \phi(m) \approx \Delta^2 \phi_{\text{cover}}(m)$,

(3.33)

where $\Delta \phi_{\text{cover}}$ and $\Delta^2 \phi_{\text{cover}}$ denote the first- and second-order phase differences in the cover sequence. If the inputs $y_n$ are already frequency corrected, then $\Delta \phi_f \approx 0$, so the first-order phase difference is sufficient. To check for synchronization, the phase differences are compared with the corresponding phase differences of the

---

**Fig. 3.13** Block diagram of a frame synchronization algorithm
Table 3.2 Delay for frame synchronization in number of samples

<table>
<thead>
<tr>
<th>TFC</th>
<th>1</th>
<th>2</th>
<th>3</th>
<th>4</th>
<th>5</th>
<th>6</th>
<th>7</th>
<th>8</th>
<th>9</th>
<th>10</th>
</tr>
</thead>
<tbody>
<tr>
<td>D</td>
<td>495</td>
<td>495</td>
<td>165</td>
<td>165</td>
<td>165</td>
<td>165</td>
<td>330</td>
<td>330</td>
<td>330</td>
<td>330</td>
</tr>
</tbody>
</table>

cover sequence, and when the patterns match the frame synchronization, indicator is asserted. The symbol counter is then set appropriately, either to 24 or 12, depending on whether a long or short preamble was expected.

When multiple RF modules are used, the quantities $z_n z_n^*$ from the independent modules can be added together before computing the phase change in Fig. 3.13.

### 3.4.2.4 Overlap-Add Process

Once frame synchronization is achieved, the sample and symbol counter are maintained without further corrections until the end of the packet. For each symbol period, samples from the ADC corresponding to one FFT window length are grouped together and delivered to the FFT module. During the process, the tail of a symbol due to channel delay spread is added to the head of the symbol as shown in Fig. 3.14. This is done to make the channel coefficient matrix a circulant one so that its eigenvector matrix is the DFT matrix.

### 3.4.2.5 Frequency Offset Estimation and Compensation

Owing to manufacturing tolerances, the crystal oscillators used to generate various clocks and carrier frequencies at the two ends of a communication link operate at slightly different frequencies. This offset relative to the actual frequency is generally on the order of $10^{-6}$ and is specified in parts per million, abbreviated as ppm. The WiMedia PHY standard [2] specifies the maximum offset to be 20 ppm, so the offset seen at the receiver can range from +40 to −40 ppm. For a carrier frequency around 5 GHz, this corresponds to a frequency offset of −200 to +200 kHz.

A frequency offset manifests as a continuous phase rotation of the complex baseband samples at the receiver. If left uncompensated, the phase offset accumulates over time and eventually causes errors in the decoded packet if the packet is long enough. Although a symbol-by-symbol phase correction process can compensate for this to some extent, a frequency offset compensation mechanism using estimates

![Fig. 3.14 Overlap–add process](image-url)
obtained over multiple symbols is a more robust solution. Besides, since UWB signals have bandwidths of 500 MHz or more, phase offsets accumulated are not the same over all subcarriers. They grow faster as we move from the center to the band-edges, so phase correction using the same phase estimate for all subcarriers is not adequate. The phase offset accumulated for subcarrier \( k \) in symbol \( m \) is

\[
\phi(k, m) = -2\pi \Delta(f_{\text{center}} + kf_{\text{sub}}) m T_{\text{sym}}, \quad k = -64, \ldots, 63,
\]

where \( T_{\text{sym}} \) is the symbol period and \( f_{\text{sub}} \) is the subcarrier spacing, set to 4.125 MHz in the WiMedia PHY standard.

An estimate of the frequency offset can be obtained from the last 6 PS symbols of the preamble as shown in Fig. 3.8. A first-order phase difference is computed along the lines of the frame synchronization process, except in this case, the second correlator output \( z_n \) is compensated for phase difference due to cover sequence. The phase differences are averaged over 6 symbols, thus covering all the bands in a hopping channel, and the frequency offset is obtained by simply scaling the average appropriately:

\[
\Delta\phi(m) = \angle z_n - \Delta\phi_{\text{cover}}(m) \quad \text{sampled at the end of symbol } m
\]

\[
\widehat{\Delta f_{\text{est}}} = -\frac{\Delta\phi(m) f_s}{2\pi D},
\]

where \( (\cdot) \) denotes the average and \( \angle z \) the phase of a complex variable \( z \). In this case, we set the delay \( D \) to 6 symbols (990 samples) for all TFCs since it provides a more robust phase estimate. Note that \( \Delta f_{\text{est}} \) is the offset with respect to \( f_{\text{center}} \). This is later translated to different subcarrier frequencies during the compensation process.

For the case of multiple RF modules, a common estimate is obtained by adding vectors \( z_n \), thus averaging \( \Delta\phi_m \) over different RX paths, in addition to averaging over symbols.

Although the estimation process is robust, it is sometimes possible that the preamble of a transmission may have some transients due to long settling times of the transmitter RF module. In such cases, the preamble based estimate will not be accurate enough, resulting in a loss in performance. To prevent this loss, the estimate may be further refined by tracking any residual phase rotation using the dedicated pilot subcarriers,

\[
\widehat{\Delta f_{\text{est,new}}} = \widehat{\Delta f_{\text{est,old}}} + \widehat{\Delta f_{\text{est,cor}}}.
\]

The estimate thus obtained is used by the frequency offset compensation module to continuously derotate the complex samples at the receiver. There are two ways of implementing the compensation module: in time domain with a complex multiplier and a sample clock phase correction mechanism, or in frequency domain with phase corrections applied to every subcarrier. Here we outline the frequency-domain method.
The compensation process is initiated at the start of channel estimation symbols (see Fig. 3.8) and is applied to every symbol until the end of the packet. For each subcarrier $k$ in symbol $m$, the phase offset is computed as follows:

$$f_k = f_{\text{center}} + kf_{\text{sub}}, \quad k = -64, \ldots, 63 \quad (3.37)$$

$$\triangle \theta_{k,m} = 2\pi \left( \frac{f_k\hat{\triangle f}_{\text{est}}}{f_{\text{center}}} \right) (mT_{\text{sym}}). \quad (3.38)$$

and a complex multiplier $e^{j\triangle \theta_{k,m}}$ is applied at the FFT output.

### 3.4.2.6 Channel Estimation

The last 6 symbols (denoted CE) in the preamble are typically reserved for channel estimation. Theoretically, it is possible to derive channel estimates using the cross-correlator of the packet detection process. However, while the PS sequences are designed for good correlation property and reduced implementation complexity, the channel estimation sequence $v$ is designed to have lower peak-to-average ratio than the preamble and better auto-correlation property than the PS sequence. Therefore, it is quite appropriate to use just the 6 CE symbols for channel estimation.

Channel coefficients $h$ can be obtained in two ways: either in time-domain or in frequency-domain. Further, there are several methods one could use in the time-domain, based on different optimization criteria like least squares or MMSE. The advantage of time-domain estimation is that the number of parameters estimated is smaller, limited to just the maximum length of the channel $L_{\text{max}}$. The least squares solution is essentially the Maximum Likelihood Estimate we have seen in Section 3.4.2.1 on packet detection, except that here the matrix $U$ is replaced with a matrix $V$ composed of vectors $v$,

$$\hat{h} = \text{avg}((V^*V)^{-1}V^*y_n), \quad n = \hat{n}_0 + mN_{\text{sym}}, \quad (3.39)$$

where $\hat{n}_0$ is the start time estimate at the receiver and $m$ is such that $y_n$ has the received samples corresponding to the CE symbols. The average operation here needs special consideration since the WiMedia PHY standard supports hopping. For example, transmission on TFC 1 hops over three bands. When hopping is used, the CE symbols are split evenly over all the bands used, so the averaging operation has to be done only over those symbols which were transmitted on the same band. For each band used, we thus have a separate set of channel coefficients.

The time-domain methods require a correlator which is computationally intensive, and more multipliers for implementing the matrix multiplication. On the other hand, the frequency-domain approach is much easier to implement, since it just uses the FFT, which is in any case required for an OFDM receiver, and a complex filter.
Let \( s_{m,k} \) denote the FFT output for subcarrier \( k \) and symbol index \( m \). The estimate of subcarrier \( k \)th channel coefficient, \( \tilde{h}_k \), is given by

\[
\tilde{h}_k = \text{avg}(s_{k,m} \tilde{v}_k^*), \quad \text{assuming } |\tilde{v}|^2 = 1,
\]

where \( \tilde{v}_k \) is the channel estimation sequence subcarrier \( k \) element, specified in [2]. As mentioned earlier, the channel coefficients for each subband used in hopping modes have to be estimated separately using appropriate CE symbols in the preamble. An FIR filter of appropriate order may be used to smoothen the coefficients if necessary.

### 3.4.3 Data Processing

The data processing block recovers the bit information conveyed in the header and payload part of the packet. A 128-point FFT transforms the signal from time-domain to frequency-domain. Channel estimates and phase estimates are used to compensate the effects of multipath fading and phase/frequency offset. Soft bit information on the coded bits is computed in QPSK/DCM demapping, and deinterleaved, depunctured and fed to the Viterbi decoder.

#### 3.4.3.1 Phase Tracking

There are 12 embedded pilot tones per OFDM symbol at subcarriers \( k \in \{-55, -45, \ldots, 55\} \), which are used for continuous tracking of phase variations due to phase noise and residual frequency offset. With received symbol \( \tilde{y}_{k,m} \) on subcarrier \( k \) in OFDM symbol \( m \), the common phase error can be computed as

\[
\varphi_{\text{CPE}} = \angle \left\{ \sum_{k=-55,-45,\ldots,55} y_{k,m} \cdot \left( \tilde{h}_{k,\text{sb}(m)} p_{k,m} \right)^* \right\}.
\]

\( \tilde{h}_{k,\text{sb}(m)} \) denotes the channel estimate (from the preamble) of subcarrier \( k \) and the subband used \( \text{sb}(m) \) in OFDM symbol \( m \). The expected pilot symbol, denoted as \( p_{k,m} \), is derived from a pseudo-random sequence known to the receiver. The phase-corrected symbol forwarded to the demapper is computed as

\[
\tilde{y}'_{k,m} = \tilde{y}_{k,m} \exp (-j 2\pi \varphi_{\text{CPE}}).
\]

#### 3.4.3.2 Soft Bit Demapping

After phase correction, we can write the received signal on subcarrier \( k \) and time \( m \) as

\[
\tilde{y}'_{k,m} = \tilde{h}_k s_{k,m} + \tilde{n}_{k,m}
\]
with $s_{k,m}$ being the transmitted constellation symbol, $\hat{h}_k$ being the channel coefficient and $\tilde{n}_{k,m}$ is assumed to be additive white Gaussian noise (AWGN) with variance $\sigma^2$ per real dimension. For convenience, the indices $k$, $m$ and $\tilde{\cdot}$ are dropped in the following derivations.

The a posteriori log-likelihood ratio value on the $i$th constellation bit $c_i$ is computed as

$$L(c_i | y) = \ln \frac{\sum_{\forall \hat{s} : c_i = 1} \exp \left[ - \frac{|y - h\hat{s}|^2}{2\sigma^2} \right]}{\sum_{\forall \hat{s} : c_i = 0} \exp \left[ - \frac{|y - h\hat{s}|^2}{2\sigma^2} \right]}.$$  

The sign of the soft value represents the binary decision whether a 0 or 1 was transmitted. Its absolute value indicates how reliable the decision is; values close to zero are regarded as unreliable. For practical purposes, it is desirable to have a simpler soft estimate which does not involve logarithmic or exponential functions.

The number of terms in the summations can be reduced by only considering terms with maximal contribution in numerator and denominator, i.e., those terms with minimal $|y - h\hat{s}|^2$. This results in the expression

$$L(c_i | y) \approx \frac{1}{2\sigma^2} \left( \min_{\forall \hat{s} : c_i = 1} |y - h\hat{s}|^2 - \min_{\forall \hat{s} : c_i = 0} |y - h\hat{s}|^2 \right)$$

which serves as the soft input metric to the Viterbi decoder.

Before demapping the QPSK symbols with (3.43), multiple observations due to complex conjugate spreading (for data rates 53.3–80 Mbps) and time-repetition (for data rates 53.3–200 Mbps) are combined coherently. Note that the QPSK demapping is separable into real and imaginary part, as bit $c_0$ is mapped to amplitude $s_{re} = (2c_0 - 1)/\sqrt{2}$, and bit $c_1$ to $s_{im} = (2c_1 - 1)/\sqrt{2}$. For data rates 320–480 Mbps, dual-carrier modulation (DCM) is used. For each DCM symbol, there are two received noisy 16-QAM constellation symbols $y_0 = h_0s_0 + n_0$ and $y_1 = h_1s_1 + n_1$, which are separated by 50 subcarriers within the same OFDM symbol. With mapping rule (3.5), and a joint metric over the two symbols $y_0, y_1$, the soft demapping of the four bits $c_0, \ldots, c_3$ writes as

$$L(c_i | y_0, y_1) = \ln \frac{\sum_{\forall (\hat{s}_1, \hat{s}_2) : c_i = 1} \exp \left[ - \left( |y_0 - h_0\hat{s}_0|^2 + |y_1 - h_1\hat{s}_1|^2 \right) / 2\sigma^2 \right]}{\sum_{\forall (\hat{s}_1, \hat{s}_2) : c_i = 0} \exp \left[ - \left( |y_0 - h_0\hat{s}_0|^2 + |y_1 - h_1\hat{s}_1|^2 \right) / 2\sigma^2 \right]}$$

which can be simplified in a similar way as (3.53) to obtain expressions without exponential and logarithmic functions. Prior to Viterbi decoding, the soft values are deinterleaved to break up possible bursts of errors that were caused by fading effects on the communication channel.
3.4.3.4 Viterbi Decoding

At the transmitter, a binary sequence \( \mathbf{b} \) of length \( K \) was mapped to a coded sequence \( \mathbf{c} \) of length \( N \). In the following, let us assume that \( \mathbf{c} \) has binary entries \( \{ \pm 1 \} \), rather than \( \{ 0, 1 \} \). The decoder’s task is to recover \( \mathbf{b} \) from a sequence of log-likelihood ratio values, denoted as \( \mathbf{z} = \mathbf{L} \), which can be regarded as the noisy received codeword sequence. All possible transmitted sequences \( \mathbf{b} \) are assumed to be equally likely, and decoding reduces to maximizing the likelihood function

\[
\max_{\mathbf{v}_b} p \left( \mathbf{z} \left| \mathbf{v}_b \right. \right).
\]

Assuming that \( \mathbf{z} \) is affected by Gaussian noise only, minimizing the Euclidean distance \( ||\mathbf{z} - \hat{\mathbf{c}}||^2 \) between received sequence \( \mathbf{z} \) and hypothesized codeword \( \hat{\mathbf{c}} \) suffices, which can be simplified to

\[
\tilde{p} \left( \mathbf{z} \left| \mathbf{v}_b \right. \right) = \hat{\mathbf{c}}^T \cdot \mathbf{z},
\]

i.e., a simple correlation is sufficient to find the best matching hypothesis \( \hat{\mathbf{c}} \) and corresponding information bit sequence \( \hat{\mathbf{b}} \). For convolutional codes, this can be done efficiently using the Viterbi algorithm, exploiting the trellis structure of the code [11].

3.5 Multiple Antenna Techniques

The robustness and spectral efficiency (data rate) of a wireless communication link can be improved by using multiple antennas. In the following, we discuss some open loop techniques, such as antenna selection (SEL), maximum ratio combining (MRC), space–time block coding (STBC) and spatial multiplexing (SMX). In open loop transmission, the transmitter does not require knowledge of the current channel conditions; only the receiver performs a channel estimation for coherent detection, using dedicated training symbols embedded in the transmitted packet.

3.5.1 Selection Diversity

A simple way of using multiple antennas at the receiver is selection diversity. No change in the transmission standard is needed. The power of all \( N_R \) receive antennas is measured during preamble transmission, and that antenna which has the maximal received power is selected for the header and payload part of the packet. Only a simple RF switch with one rather than \( N_R \) down-conversion chains (LNA, mixer, ADC stage) is required. However, the gains over fading channels are limited, and typically range from 0.5 to 2 dB.
3.5.2 Maximum Ratio Combining

More powerful, but also more complex to implement is maximum ratio combining (MRC), which makes efficient use of multiple receive antennas without requiring a change in the transmission standard. To perform a coherent superposition of the $N_R$ received signals, MRC requires full RF chains for each antenna. AGC is performed for each RF block separately. In digital baseband, the signals from each receive antenna are processed in separate overlap-add units, FFT units, channel estimation and phase estimation units up to the demapper. Prior to demapping, the signals are coherently combined. This is referred to as maximum ratio combining, as channel/phase estimation and correction is performed separately on each signal before combination. The remaining processing from demapper onward remains unchanged to the single antenna case. Combination per subcarrier $k$ and time $m$ is done by

$$y_{MRC} = \sum_{i=1}^{N_R} h_i^* y_i.$$  

Assuming perfect channel estimates $h_i$ at receive antenna $i$, this can be written as

$$y_{MRC} = \left( \sum_{i=1}^{N_R} |h_i|^2 \right) \cdot s + \sum_{i=1}^{N_R} h_i^* n_i,$$  \hspace{1cm} \text{(3.45)}

resulting in coherent superposition of wanted signal, and non-coherent superposition of the unwanted noise components. For $N_R = 2$, the gain on an AWGN test channel is 3 dB, going up to about 5 to 8 dB on fading channels.

3.5.3 Space–Time Block Coding

In space–time block coding [12, 13], the same data is transmitted simultaneously over different antennas. Space–time block codes are particularly useful when there are more transmit than receive antennas. They can be viewed as repetition codes over space and time, providing increased robustness and range extension. A change in the communication standard of [2] is required to account for this new transmission format. A quite attractive variant are orthogonal space–time block codes, which can be detected optimally at the receiver with simple linear operations. The Alamouti orthogonal space–time block code for the 2 × 1-channel [12] is given by

$$S = \begin{bmatrix} s_1 & s_2 \\ -s_2^* & s_1^* \end{bmatrix}.$$  

At time instance 1, constellation symbol $s_1$ is transmitted from antenna 1 and $s_2$ from antenna 2. At time instance 2, symbol $-s_2^*$ is transmitted from antenna 1 and $s_1^*$ from antenna 2. This pattern is repeated with new symbols $s_1, s_2$. There are two information symbols transmitted per two channel uses, commonly referred to as “spatial rate 1”. It can be proved [13] that no full rate (i.e. spatial rate 1) orthogonal STBC exists for $N_T > 2$.

### 3.5.3.1 Detection of the Alamouti Space–Time Block Code

Assuming that the channel remains unchanged over two consecutive symbol intervals in time (slow fading), the detection turns out to be very simple. The received symbols at time instances $m = 1, 2$ are

$$y_1 = h_1 s_1 + h_2 s_2 + n_1$$
$$y_2 = -h_1 s_2^* + h_2 s_1^* + n_2.$$ 

Estimates of the symbols $s_1, s_2$ are obtained by computing

$$\hat{s}_1 = h_1^* y_1 + h_2^* y_2 = (|h_1|^2 + |h_2|^2) s_1 + h_1^* n_1 + h_2^* n_2^*$$
$$\hat{s}_2 = h_2^* y_1 - h_1^* y_2 = (|h_1|^2 + |h_2|^2) s_2 + h_2^* n_1 + h_1^* n_2^*.$$ 

(3.46)

Note that the effective channel of (3.46) resembles the structure of maximum-ratio combining (3.45). The Alamouti code is unique: only for a $2 \times 1$-channel such a simple and elegant detection scheme exists. It does not scale to cases with more transmit antennas, $N_T > 2$. For $N_R > 1$, however, we can always combine the Alamouti scheme with MRC to benefit from the added spatial diversity.

### 3.5.4 Spatial Multiplexing

Spatial multiplexing [6, 14, 15] increases the data rate, providing high spectral efficiency, similar to going to higher order QAM; however, there is no power penalty. As different data is transmitted simultaneously over $N_T$ transmit antennas, a standard change (or evolution) of [2] is required to provide for this new transmission format. As with space–time block codes, the preamble needs to be extended to allow estimation of the entire channel matrix per subcarrier. The receiver has to decouple the $N_T$ spatial streams, to recover the transmitted information. Optimal detection is computationally intensive, and there exist several suboptimal, low-complexity detectors with different performance/complexity trade-off, ranging from simple zero-forcing to complex soft a posteriori probability (APP) detection. Spatial multiplexing can be considered as a vector-QAM, forming a compound constellation symbol of $2^{MN_T}$ signal points, conveying information of $MN_T$ bits per channel use over a matrix channel.
3.5.4.1 Spatial Soft Bit Demapping

For each subcarrier, the received signal can be written as

\[ y = Hs + n \]  

with \( s \) being the transmitted \( N_T \times 1 \) constellation vector of QAM symbols \( s_i, \ i = 1, \ldots, N_T \). \( H \) is the \( N_R \times N_T \) matrix of channel coefficients, \( n \) is the \( N_R \times 1 \) vector of additive Gaussian noise, with complex entries, the real and imaginary components of each having variance \( \sigma^2 \).

In a first step, we use the zero-forcing (ZF) technique to spatially separate the MIMO signal into subchannels,

\[ s_{\text{est}} = Wy. \]  

\( s_{\text{est}} \) is the \( N_T \times 1 \) vector estimate of the transmitted constellation vector used for APP post-processing to extract the soft bit metrics. The pseudo-inverse \( W \) (an \( N_T \times N_R \) matrix) is computed as

\[ W = (H^*H)^{-1}H^*. \]  

The a posteriori log-likelihood ratio value on the \( i \)th constellation bit \( c_{i,j} \) of TX antenna \( j \) can be computed as

\[ L(c_{i,j} | y) = \ln \frac{\sum_{\forall k, c_{i,j} = 1} \exp \left[ -\|y - Hs\|^2 / 2\sigma^2 \right]}{\sum_{\forall k, c_{i,j} = 0} \exp \left[ -\|y - Hs\|^2 / 2\sigma^2 \right]} . \]  

The Euclidean distance can be expressed as a sum of two positive quantities

\[ \|y - H\hat{s}\|^2 \]

\[ = y^*(I - H(H^*H)^{-1}H^*)(I - H(H^*H)^{-1}H^*)y \]

\[ + (\hat{s} - s_{\text{est}})^* (H^*H) (\hat{s} - s_{\text{est}}) . \]  

The first quantity is common to all terms in both numerator and denominator; omitting this term simplifies (3.50) to

\[ L(c_{i,j} | s_{\text{est}}) = \ln \frac{\sum_{\forall k, c_{i,j} = 1} \exp \left[ - (\hat{s} - s_{\text{est}})^* (H^*H) (\hat{s} - s_{\text{est}}) / 2\sigma^2 \right]}{\sum_{\forall k, c_{i,j} = 0} \exp \left[ - (\hat{s} - s_{\text{est}})^* (H^*H) (\hat{s} - s_{\text{est}}) / 2\sigma^2 \right]} . \]
Reducing the number of terms in the summations by ignoring off-diagonal elements in the matrix \((H^*H)^{-1}\), and considering only terms with maximal contribution in numerator and denominator, i.e., those terms with minimal \(|s_{est,j} - \hat{s}|^2\), results in the expression

\[
L(c_{i,j} | y) = L(c_{i,j} | s_{est,j}) \approx L(c_{i,j} | s_{est,j}) = \ln \frac{\sum_{\forall \hat{s} : c_{i,j} = 1} \exp \left( -|s_{est,j} - \hat{s}|^2 / 2\sigma^2_{jj} \right)}{\sum_{\forall \hat{s} : c_{i,j} = 0} \exp \left( -|s_{est,j} - \hat{s}|^2 / 2\sigma^2_{jj} \right)} \approx \frac{1}{2\sigma^2_{jj}} \left( \min_{\forall \hat{s} : c_{i,j} = 1} |s_{est,j} - \hat{s}|^2 - \min_{\forall \hat{s} : c_{i,j} = 0} |s_{est,j} - \hat{s}|^2 \right),
\] (3.53)

with \(\sigma^2_{jj} = \sigma^2 \cdot [(H^*H)^{-1}]_{jj}\). The soft values computed in (3.53) are the inputs to the Viterbi decoder.

### 3.6 Performance Results

In this section, we evaluate some of the methods by means of system simulation, to obtain the minimum SNR required for obtaining an 8% FER (“sensitivity”), and compute range estimates based on a simple free-space path loss model.

#### 3.6.1 Receiver Sensitivity

The receiver sensitivity is an important parameter used for characterizing the quality of both the RF as well as baseband processing. We next define the relevant quantities required for computing sensitivity, such as signal-to-noise ratio and transmit power in the ultra-wideband context.

##### 3.6.1.1 Signal-to-noise Ratio

In OFDM communications, it may be important to distinguish between the time-domain signal-to-noise ratio \(\text{SNR}_t = E_s / N_0\), based on the time-domain samples from the ADC, and a frequency-domain signal-to-noise ratio \(\text{SNR}_f\), describing the SNR of the QAM symbol constellations on the individual subcarriers. After Fourier transformation at the receiver, the noise power \(N_0\) is spread over \(N_{sc}\) subcarriers, while the signal power \(E_s\) is concentrated on \(N_{su}\) actually used subcarriers (omitting null subcarriers such as guard subcarriers and DC). Thus, \(\text{SNR}_t\) and \(\text{SNR}_f\) are related by

\[
\text{SNR}_f |_{\text{dB}} = \text{SNR}_t |_{\text{dB}} + 10 \log_{10}(N_{sc} / N_{su})
\]
with $N_{sc}$ being the total number of subcarriers, and $N_{su}$ being the number of actually used subcarriers. In WLAN 802.11 a/g, the numbers are $N_{sc} = 64$, $N_{su} = 52$, and thus the difference is $10 \log_{10}(64/52) \approx 0.9$ dB which is non-negligible. In MB-OFDM, however, we have $N_{sc} = 128$, $N_{su} = 122$, giving rise to a difference of 0.2 dB, which is irrelevant. For sensitivity computations, SNR, is assumed.

### 3.6.1.2 Transmit Power

In UWB communications, the transmit power $P_{TX}$ is limited to a spectral density of $-41.3$ dBm/MHz (FCC, 3.1–10.6 GHz). With no frequency hopping (TFC 5–7), the transmit power for one subband is

$$P_{TX} = -41.3 \text{ dBm} + 10 \log_{10}(122 \cdot 4.125 \text{ MHz/1 MHz}) \approx -14.28 \text{ dBm}. \quad (3.54)$$

For frequency hopping among three subbands (TFC 1–4), the transmit power can be increased by a factor of 3, or, equivalently, 4.77 dB,

$$P_{TX_3} = P_{TX} + 10 \log_{10} 3 \approx -9.51 \text{ dBm}. \quad (3.55)$$

Correspondingly, for frequency hopping among two subbands (TFC 8–10), we have

$$P_{TX_2} = P_{TX} + 10 \log_{10} 2 \approx -11.27 \text{ dBm}. \quad (3.56)$$

### 3.6.1.3 Sensitivity

The signal-to-noise ratio at the input of a receive antenna is

$$\frac{S_{in}}{N_{in} \mid_{dB}} = NF \mid_{dB} + \frac{S_{out}}{N_{out} \mid_{dB}}$$

with NF denoting the effective noise figure of the RF part. The signal-to-noise ratio at the output of the ADC is given by

$$\frac{S_{out}}{N_{out}} = \frac{E_s}{N_0} \mid_{\text{required}}$$

where the notion of “required” stands for the minimal $E_s/N_0$-value when a desired packet error rate is achieved (e.g., 8% as defined in [2]). The receiver sensitivity at the antenna input port is given by

$$S_{in \mid_{dBm}} = N_{in \mid_{dBm}} + NF \mid_{dB} + \frac{E_s}{N_0} \mid_{\text{dB,required}}.$$
For an input bandwidth $B$, the noise power writes as

$$N_{in} = 10 \log_{10} \left( \frac{kT B}{1 \text{ mW}} \right) = 10 \log_{10} \left[ kT \cdot 1 \text{ Hz} \right] -174 \text{ dBm} + 10 \log_{10} \left[ \frac{B}{1 \text{ Hz}} \right]$$

with Boltzmann constant $k = 1.38 \cdot 10^{-23} \text{ J/K}$ and system at room temperature, $T = 290 \text{ K}$. For example, for an instantaneous bandwidth of $(122 \cdot 4.125) \text{ MHz}$, corresponding to a single subband in MB-OFDM, we obtain

$$N_{in} = (-174 + 10 \log_{10} (122 \cdot 4.125 \cdot 10^6)) \text{ dBm} \approx -87 \text{ dBm}.$$

### 3.6.2 Expected Range

We can estimate the range of the transmission system using the free-space path loss model. It is a good approximation when there is a direct line of sight (LOS) between transmitter and receiver. The path loss is

$$\text{PL}_{fs} (f_c, d) \mid \text{dB} = 10 \log_{10} \left( \frac{f_c \cdot 4\pi d}{c} \right)^\gamma, \gamma = 2$$

(3.57)

with the speed of light $c = 3 \cdot 10^8 \text{ m/s}$. The center frequency is computed using the geometric mean of upper and lower band edge, i.e., $f_c = \sqrt{f_{\text{lower}} \cdot f_{\text{upper}}}$. We can parameterize (3.57) with respect to a path loss exponent $\gamma$ (which is 2 for free space) and apply it to the indoor environment. Curve fitting of indoor measurement data shows that typical values for $\gamma$ range from 1.7 for LOS to 4 for NLOS situations. Note that values $\gamma < 2$ are possible owing to constructive multipath combining at the receiver.

Assuming zero antenna gain, the link budget writes as

$$P_{RX} \mid \text{dBm} = P_{TX} \mid \text{dBm} - \text{PL}_{fs} (f_c, d) \mid \text{dB} - \text{LM} \mid \text{dB}.$$  

(3.58)

Setting $P_{RX} = S_{in}$, i.e., using the received power at the sensitivity limit, accounting for frequency hopping (3.54)–(3.56), and solving (3.58) for the distance $d$ provides an estimate on the achievable range. We can use the link margin $\text{LM}$ to account for further losses anticipated in the TX/RX chain. For example, in [2], the minimum receiver sensitivity numbers are given for a payload length of 1024 bytes, AWGN channel (cable, CM0), and a typical noise figure of 6.6 dB (referenced at the antenna), an implementation loss of 2.5 dB, and an additional margin of 3 dB.
3.6.3 Simulation Results

The SNR at which an FER of 8% is achieved was determined by simulation. For each SNR value, 500 packets were simulated, with a packet length set to 1024 bytes, using channel models CM0, CM2 and CM4.

Figure 3.15 (left) shows frame error rate (FER) curves for data rates from 53.3 to 1024 Mbps over an idealized additive white Gaussian noise channel (CM0). There is no delay spread. Extracting only that point per data rate when an FER of 8% is achieved gives the compact curve shown in the right chart. Connecting the data points by lines proves to be useful when several groups of different simulation results are presented in the same chart. Obviously, the higher the data rate, the higher the SNR required, spanning from –2.5 dB for 53.3 to more than 13 dB for 1024 Mbps.

The effect of delay spread is illustrated in Fig. 3.16. As we go from zero to around 100 ns delay spread, the higher data rates in particular lose in performance. For CM2, the losses range from 1.5 dB at 53.3, over 4 dB at 480, to 6.5 dB at 1024 Mbps. For CM4, no rates above 480 Mbps are possible: the delay spread exceeds the duration of the null-suffix, and inter-symbol interference severely degrades the signal quality. The benefits of MRC can clearly be seen, with gains around 2–3 dB for CM0, up to 6 dB for CM2 and CM4. Note that MRC does not help much for the high rates over CM4, as neither thermal noise nor fading, but self-interference due to inter-symbol interference is the dominating factor there.

With the required SNR values for 8% FER of Figs. 3.15 and 3.16, we can compute receiver sensitivity numbers, as well as obtain an estimate on the achievable range for the different data rates, plotted in the lower right corner of Fig. 3.16. For the distance estimates, we assumed a transmit power of $P_{\text{TX}} = -14.28$ dBm (TFC 6), a noise figure of $NF = 6$ dB, a path loss coefficient of $\gamma = 2$ and an additional link margin of $LM = 2$ dB.

![Fig. 3.15 FER of 53.3–1024 Mbps over CM0 (AWGN) channel](image-url)
3.7 Complexity Discussion and Conclusions

Table 3.3 summarizes the complexity increase and performance gains of the different methods discussed. Selection diversity (SEL), maximum ratio combining (MRC) and Alamouti space–time block coding (STBC) provide SNR gains ranging from 0.5 to 6 dB, thus increasing coverage and robustness of the transmission.

<table>
<thead>
<tr>
<th>Method</th>
<th>Analog RF TX (%)</th>
<th>Analog RF RX (%)</th>
<th>Digital baseband TX (%)</th>
<th>Digital baseband RX (%)</th>
<th>Max. rate (Mbps)</th>
<th>SNR gain (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>SEL $N_R = 2$</td>
<td>0</td>
<td>100</td>
<td>0</td>
<td>2</td>
<td>480</td>
<td>0.5–2</td>
</tr>
<tr>
<td>MRC $N_R = 2$</td>
<td>0</td>
<td>100</td>
<td>0</td>
<td>40</td>
<td>480</td>
<td>1–6</td>
</tr>
<tr>
<td>STBC $N_T = 2$</td>
<td>100</td>
<td>0</td>
<td>5</td>
<td>10</td>
<td>480</td>
<td>1–3</td>
</tr>
<tr>
<td>SMX $N_T = 2$, $N_R = 2$</td>
<td>100</td>
<td>100</td>
<td>10</td>
<td>50</td>
<td>1024</td>
<td>–2</td>
</tr>
<tr>
<td>16-QAM</td>
<td>0</td>
<td>0</td>
<td>5</td>
<td>10</td>
<td>1024</td>
<td>–7</td>
</tr>
</tbody>
</table>
at a moderate increase in complexity. Spatial multiplexing and 16-QAM double the data rate without bandwidth expansion, by increasing spectral efficiency. SMX only incurs a link margin penalty of about 2 dB, while implementation costs are high. 16-QAM, on the contrary, is simple to implement, but causes a significant reduction in link margin; SEL, MRC and STBC can be combined with 16-QAM to recover some of the losses. A 4-times increase in data rate can be obtained by combining SMX and 16-QAM at, however, a penalty of about 9 dB in link margin.

References

Chapter 4
Design of CMOS Transceivers for MB-OFDM UWB Applications

Behzad Razavi, Turgut Aytur, Christopher Lam, Fei-Ran Yang, Kuang-Yu Li, Ran-Hong Yan, Han-Chang Kang, Cheng-Chung Hsu, and Chao-Cheng Lee

Abstract This chapter describes the design of transceiver architectures and circuits suited to frequency-hopping UWB applications. A direct-conversion UWB transceiver for Mode 1 OFDM applications employs three resonant networks and three phase-locked loops. Using a common-gate input stage, the receiver allows direct sharing of the antenna with the transmitter. Designed in 0.13-μm CMOS technology, the transceiver provides a total gain of 69–73 dB and a noise figure of 6.5–8.4 dB across three bands, and a TX 1-dB compression point of −10 dBm. The circuit consumes 105 mW from a 1.5-V supply. A frequency plan for a full-band transceiver is also described.

4.1 Introduction

Ultra-wideband (UWB) communication by means of short, “carrier-free” pulses was first conceived in “time-domain electromagnetics” in the 1960s [1, 2]. At the time, the low interceptibility and fine-ranging resolution of UWB pulses made this type of signaling attractive to military and radar applications, but today the potential for high data rates has ignited commercial interest in UWB systems. Both direct-sequence impulse communication and multiband orthogonal frequency division multiplexing (OFDM) are presently under consideration for the UWB standard.

This chapter describes the design of the first UWB CMOS transceiver for Mode 1 multiband OFDM applications. Section 4.2 gives a system overview, and Section 4.3 summarizes the receiver (RX) and transmitter (TX) specifications. Sections 4.4

1 The term “ultra-wideband” was evidently coined by the U.S. Department of Defense in the 1980s.

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and 4.5 present the transceiver architecture and building blocks, respectively. Section 4.6 introduces a full-band architecture, and Section 4.7 deals with the experimental results.

4.2 System Overview

4.2.1 MBOA Standard

The Multiband OFDM Alliance (MBOA) standard for UWB communications draws heavily upon prior research in wireless local area network (WLAN) systems [3]. In a manner similar to IEEE 802.11a/g, MBOA partitions the spectrum from 3 to 10 GHz into 528-MHz bands and employs OFDM in each band to transmit data rates as high as 480 Mb/s. A significant departure from the original principle of “carrier-free” signaling, the multiband operation is chosen to both simplify the generation and detection of signals and leverage well-established OFDM solutions from WLAN systems. To ensure negligible interference with existing standards, the FCC has limited the output power level of UWB transmitters to $-41 \text{ dBm}/\text{MHz}$.

Figure 4.1 shows the structure of the MBOA bands and the channelization within each band. The 14 bands span the range of 3,168–10,560 MHz, with their center frequencies given by $m \times (264 \text{ MHz})$ for odd values of $m$ from 13 to 39. Each band consists of 128 subchannels of 4.125 MHz. In contrast to IEEE 802.11a/g, MBOA employs only QPSK modulation in each subchannel to allow low resolution in the baseband analog-to-digital (A/D) and digital-to-analog (D/A) converters (4–5 bits). Bands 1–3 constitute “Mode 1” and are mandatory for operation, whereas the remaining bands are envisioned for high-end products.

In order to improve the robustness of the system with respect to multipath effects and interference, the standard complements OFDM with band hopping. In Mode 1, for example, the information bits are interleaved across all three bands and, as illustrated in Fig. 4.2, the system hops at the end of each OFDM symbol (every 312.5 ns). The band switching must occur in less than 9.47 ns, thereby posing difficult challenges in the design of the transceiver.
Table 4.1 compares the receiver specifications of IEEE 802.11a and MBOA UWB systems for their respective maximum data rates. The latter demands both a much more stringent noise figure (NF) and a much greater overall bandwidth in the RX and TX paths.

It is interesting to examine the noise figure requirements if MOBA had retained the IEEE 802.11a 64-QAM format. To raise the data rate from 54 to 480 Mb/s, the channel bandwidth, $B$, would need to increase from 20 to 178 MHz. Writing

\[
\text{Sensitivity} = -174 \text{ dBm} + 10 \log B + \text{NF} + \text{SNR},
\]

where SNR denotes the required signal-to-noise ratio, and assuming SNR = 23 dB for a bit error rate (BER) of $10^{-5}$, we have

\[
\text{Sensitivity} = -68.5 \text{ dBm} + \text{NF}.  \tag{4.2}
\]

That is, it would have been impossible to achieve a sensitivity of $-73$ dBm. On the other hand, if SNR is relaxed by reducing the order of the modulation, the sensitivity can be improved even though $10 \log B$ increases to some extent. In particular, with SNR = 8 dB for QPSK modulation and $B = 528$ MHz,

\[
\text{Sensitivity} = -79 \text{ dBm} + \text{NF}, \tag{4.3}
\]

indicating that an NF of about 6 dB yields the required sensitivity.

---

**Table 4.1** Comparison of IEEE 802.11a and MBOA specifications

<table>
<thead>
<tr>
<th></th>
<th>IEEE 802.11a</th>
<th>MBOA UWB</th>
</tr>
</thead>
<tbody>
<tr>
<td>Sensitivity</td>
<td>−65 nBm</td>
<td>−73 dBm</td>
</tr>
<tr>
<td>Data rate</td>
<td>54 Mb/s</td>
<td>480 Mb/s</td>
</tr>
<tr>
<td>channel BW</td>
<td>20 MHz</td>
<td>528 MHz</td>
</tr>
<tr>
<td>Modulation</td>
<td>64–QAM</td>
<td>QPSK</td>
</tr>
<tr>
<td>SNR(BER=10$^{-5}$)</td>
<td>23 dB</td>
<td>8 dB</td>
</tr>
<tr>
<td>RX NF</td>
<td>12–14 dB</td>
<td>6–7 dB</td>
</tr>
</tbody>
</table>

---

2 The MBOA modulation for maximum bit rate has recently changed to “dual-carrier modulation” (DCM), but the required SNR remains the same.
3 Owing to the coding gain in the system, the NF can be a few decibels higher.
One can attribute the foregoing significant improvement in the sensitivity to Shannon’s theorem:

\[ C = B \log_2(1 + \text{SNR}), \]  

(4.4)

where \( C \) denotes the capacity. Since \( C \) exhibits a stronger dependence on \( B \) than on SNR, higher data rates are more efficiently afforded by raising the symbol rate (with low-order modulation) than by requiring a high-order modulation (with low symbol rate).

### 4.2.2 Problem of Band Hopping

As mentioned above, the MBOA standard exploits frequency diversity through band hopping while demanding a settling time of only 9.47 ns. Since typical phase-locked loops (PLLs) take several hundred input cycles to settle, it is not possible to accommodate such fast band switching in a phase-locked synthesizer. This issue dominates the choice of frequency planning.

Another difficulty arising from band hopping relates to offset cancellation in the baseband. On the one hand, the dc offsets change in different bands and, on the other hand, analog offset cancellation circuits cannot settle in 9.47 ns if they must not attenuate the lowest OFDM subchannels (a few megahertz away from zero). For this reason, (coarse) offset cancellation can be performed by measuring and digitally storing the offsets for each band during the preamble and applying the results through D/A converters during payload.

### 4.2.3 Frequency Synthesis by Single-Sideband Mixing

In order to generate “agile” local oscillator (LO) signals, two frequencies can be added or subtracted by means of single-sideband (SSB) mixers [5]. However, SSB mixing, particularly in CMOS technology, presents a number of difficult spur issues. First, at least one port of each “submixer” must be linear to avoid mixing harmonics of that input with those of the other. The required linearization translates to a low conversion gain, small output swings, and hence the need for power-hungry (and perhaps inductor-hungry) buffers. Second, the waveforms applied to the linear ports must themselves exhibit low distortion, a difficult problem at gigahertz frequencies. Third, phase and gain mismatches in the quadrature paths and within the mixers introduce additional spurs. Fourth, dc offsets lead to leakage of the input components to the output. Of particular concern here are spurious components that fall in the IEEE 802.11a/g bands as they can corrupt the downconverted signal in the presence of large WLAN interferers.
4.3 Transceiver Specifications

The design of UWB transceivers faces the following issues: (1) the need for broadband circuits and matching; (2) gain switch in the LNA without degrading the input match; (3) broadband transmit/receive switch at the antenna; (4) desensitization due to WLAN interferers; (5) fast band hopping.

With a 528-MHz channel bandwidth, the RX and TX paths of UWB systems may naturally employ direct conversion. Typical direct-conversion issues plague the receive path, except that flicker noise negligibly affects the signal. Also, the TX side is free from injection pulling of the oscillator by the output stage because the transmitted level falls below $-41$ dBm/Hz. (The wide PLL bandwidth also suppresses the pulling [4].) This section describes the transceiver specifications, derived from the MBOA requirements and extensive system simulations.

Particularly important here is the maximum tolerable synthesizer phase noise as it determines the choice between ring and LC oscillators. In order to quantify the effect on the constellation, system-level simulations are performed, wherein a QPSK-modulated OFDM signal carrying a data rate of 480 MB/s is subjected to phase noise and the resulting bit error rate is measured. Figure 4.3(a) shows the phase noise profile assumed in simulations, and Fig. 4.3(b) plots the BER as a function of the SNR for various profiles. Each profile is characterized with a “plateau” level, $S_{\phi 0}$, and a corner frequency, $f_c$. Since the rotation of the signal constellation is given by the total integrated phase noise, the degradation depends on both the magnitude of the plateau and the corner frequency. It is observed that a plateau phase noise of $-100$ dBc/Hz with $f_c \approx 5$ MHz affects the performance negligibly, making ring oscillators (along with wideband synthesizers) a viable solution.

4.3.1 Receiver

Depending on the bit rate, MBOA specifies receiver sensitivities ranging from $-84$ dBm (for 55 Mb/s) to $-73$ dBm (for 480 Mb/s). With a required SNR of about 8 dB, these specifications translate to a noise figure of 6–7 dB. The RX must provide a maximum voltage gain of approximately 84 dB so as to raise the minimum signal level to the full scale of the baseband A/D converter. Also, based on the interference expected from IEEE 802.11a/g transmitters, a 1-dB compression point ($P_{1dB}$) of $-23$ dBm (in the high-gain mode) is necessary.

Table 4.2 summarizes the required performance. The phase noise specification is tightened by 5 dB with respect to the value obtained above to allow similar corruption in the transmitter. Note that the LNA must accommodate a gain switch of 16–20 dB to avoid excessive nonlinearity (due to the mixer and subsequent stages) in the OFDM signal as the received level exceeds $-40$ dBm. The total range for automatic gain control (AGC) is 60 dB.

The baseband channel-select filter must be designed in conjunction with the A/D converter. Greater stopband rejection provided by the former relaxes the sampling
4.3.2 Transmitter

The transmitter performance follows corresponding observations in the receiver and is summarized in Table 4.3. The maximum carrier leakage is chosen so that the rate of the latter. For example, a third-order Butterworth response necessitates a sampling rate of about $4 \times 264 = 1056$ MHz. The A/D converter resolution is determined by the tolerable quantization noise, the AGC resolution, and the level of WLAN interferers that are only partially attenuated by the filter.

![Graph showing phase noise profile and BER vs. SNR](image)

**Table 4.2 Required receiver performance**

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Requirement</th>
</tr>
</thead>
<tbody>
<tr>
<td>Sensitivity</td>
<td>–84 to –73 dBm</td>
</tr>
<tr>
<td>NF</td>
<td>6–7 dB</td>
</tr>
<tr>
<td>$P_{1dB}$</td>
<td>–23 dBm</td>
</tr>
<tr>
<td>I/Q mismatch</td>
<td>6° and 0.6 dB</td>
</tr>
<tr>
<td>Phase noise</td>
<td>–105 dBc/Hz (plateau)</td>
</tr>
<tr>
<td>Voltage gain</td>
<td>84 dB</td>
</tr>
<tr>
<td>LNA Gain switch</td>
<td>16–20 dB</td>
</tr>
<tr>
<td>Total AGC range</td>
<td>60 dB</td>
</tr>
<tr>
<td>ADC</td>
<td>5–Bit, 528–1056 MHz</td>
</tr>
</tbody>
</table>
TX incurs negligible degradation in the error vector magnitude (EVM). Note that, despite the large peak-to-average ratio of OFDM signals, the TX can operate with only a 4-dB backoff from the 1-dB compression point. This is because the large peaks appear infrequently and, more importantly, QPSK signals degrade very gradually by the intermodulation of OFDM subchannels when compression occurs. (By contrast, 64-QAM modulation typically requires an additional 5 dB of backoff for proper operation.)

### 4.4 Transceiver Architecture

Figure 4.4 shows the transceiver architecture. (The circuitry in the dashed box contains quadrature components but is drawn with only one phase for clarity.) The receive path consists of an LNA having three resonant loads corresponding to the three bands, with each load driving selectable quadrature mixers. The downconverted signal is applied to a fourth-order Sallen-and-Key (SK) filter and a first-order low-pass stage. This amount of filtering allows ADC sampling rates slightly greater than 512 MHz. An AGC range of 60 dB is distributed as 16 dB in the LNA, 30 dB at the output of the mixers, and 14 dB in the baseband. The transmit path

---

**Table 4.3** Required transmitter performance

<table>
<thead>
<tr>
<th>DAC</th>
<th>5–Bit, 528–1056 MHz</th>
</tr>
</thead>
<tbody>
<tr>
<td>I/Q mismatch</td>
<td>6° and 0.6 dB</td>
</tr>
<tr>
<td>Output power</td>
<td>–10 dBm</td>
</tr>
<tr>
<td>Output $P_{1\text{dB}}$</td>
<td>–6 dBm</td>
</tr>
<tr>
<td>Carrier leakage</td>
<td>–30 dBc</td>
</tr>
<tr>
<td>Phase noise</td>
<td>–105 dBc/Hz (Plateau)</td>
</tr>
</tbody>
</table>
similarly employs fourth-order SK filters, upconversion mixers, and an output stage that shares the antenna with the LNA.

The LO frequencies are synthesized using three independent PLLs to avoid SSB mixing. Running with a reference frequency of 66 MHz, each PLL generates quadrature phases and consumes 15 mW, less than the power required by two sets of SSB mixers that would produce comparable output swings.

Three aspects of the above approach to frequency synthesis should be noted. First, the use of a 66-MHz reference allows a loop bandwidth of about 5 MHz, thus suppressing the close-in phase noise of the oscillators considerably. Second, the reference spurs at \( n \times 66 \)-MHz offset fall within the desired channel for \( n \leq 4 \); i.e., only spurs resulting from the fifth- and high-order harmonics of the reference must be sufficiently small, a point that eases the trade-off between the loop bandwidth and the spur levels in the design of the PLLs. Third, since the three PLL frequencies are far from each other and not related by integer multiples, injection pulling is negligible.

4.5 Building Blocks

4.5.1 Low-Noise Amplifier

In order to achieve both a broadband input match and a broadband transfer, inductively degenerated CMOS cascode LNAs can incorporate (1) an additional input bandpass network that cancels the reactive part of the input impedance across a wide frequency range [6] and (2) an inductively peaked resistive load to broaden the output bandwidth [6]. The principal issue here is that the inductors used in the input network introduce loss, raising the noise figure. Moreover, with a 1.2-V supply, it becomes increasingly difficult to accommodate a large dc drop across the load resistor and hence obtain a reasonable gain.

Figure 4.5 shows the LNA topology used in this work. A common-gate stage provides an input resistance of 50Ω, and the large (20-nH) inductor \( L_1 \) resonates with the total capacitance at the input at about 4 GHz, thereby yielding adequate return loss across the Mode 1 frequency range without degrading the noise figure. The required 16-dB gain switch is realized by turning \( M_1 \) off (\( W_2 = W_1/8 \)), and the resulting increase in the input resistance is compensated by turning \( M_6 \) on (in the triode region). The on-resistance of \( M_6 \) varies with process and temperature, but the correction still guarantees \( S_{11} > 10 \text{ dB} \) under all conditions.

Transistors \( M_3-M_5 \) serve as switched cascode devices with tanks resonating at the center frequency of each band. The \( Q \) of the tanks is reduced to about 3 by addition of a parallel resistance to ensure a small droop near the band edges. With no series resistance necessary in the loads, the circuit can achieve a high gain at low supply voltages. Each output drives a set of quadrature mixers.

Simulations indicate that the LNA displays a noise figure of 3.3 dB and a voltage gain of 22 dB while drawing 2.5 mA from the supply.
4.5.2 Mixer

Figure 4.6 depicts the downconversion mixer circuit. The single-balanced topology incorporates resistor $R_H$ to halve the bias current commutated by $M_2$ and $M_3$, thereby allowing these transistors to switch more abruptly and hence inject less noise to the output. Furthermore, for a given voltage headroom, the mixer load resistance can be doubled, raising the conversion gain by 6 dB.

In order to ensure accurate current splitting between $R_H$ and the switching pair, the common-mode level of the LO port is defined by means of a tracking circuit. As illustrated in Fig. 4.6, with $M_5$ carrying $I_2 = 0.25I_B$, a current of $0.25I_B$ must flow through $2R_H$, yielding $V_X = V_{DD} - 0.5R_HI_B$ and thus $V_Y = V_{DD} - 0.5R_HI_B + V_{GS5}$. Consequently, $V_P = V_{DD} - 0.5R_HI_B + V_{GS5} - V_{GS2,3}$, reducing to $V_{DD} - 0.5R_HI_B$ if $V_{GS5} \approx V_{GS2,3}$ and, therefore, establishing a current of $0.5I_B$ through $R_H$.

The mixer must provide 30 dB of gain variation in steps of 6 dB. To this end, each load resistor is decomposed into six binary-weighted segments, and the output
current of the mixer is routed to one of the nodes according to the gain setting. The six PMOS switches required here must be wide enough to consume minimal voltage headroom while allowing a bandwidth of greater than 300 MHz at the output.

In addition to high linearity, the gain switching scheme of Fig. 4.6 also provides a constant output impedance. This property proves essential to the design of the subsequent baseband filter.

Simulations predict a noise figure of 16 dB and a voltage conversion gain of 10 dB for the mixer with a supply current of 2.5 mA.

4.5.3 Baseband Filter

With a mixer gain of 10 dB, the LNA/mixer cascade tends to experience compression at the output of the mixer. To alleviate this issue, the baseband filter can create a low impedance at the mixer output (at the interferer frequency) and hence reduce the voltage swing at these nodes considerably.

Figure 4.7(a) shows the SK filter design and its interface with the mixer. The (binary-weighted) load resistors of the mixer serve as part of the filter, and the core amplifier employs a gain of 2. Despite the low open-loop gain, the filter lowers the interferer voltage swings\(^4\) at nodes \(X\) and \(Y\) by about 3 dB, moving the compression bottleneck to the input of the mixer.

Figure 4.7(b) shows the core amplifier. To obtain an open-loop bandwidth of greater than 1 GHz, the circuit incorporates a simple (resistively-degenerated) linearized differential pair and source followers.

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\(^4\) The worst-case interferer lies above the third band, namely, at 5.15 GHz, and is downconverted to 670 MHz along with the desired signal.
4.5.4 Transmitter Front End

Unlike the receiver, the transmitter must accommodate the Mode 1 frequency range in only a single signal path to avoid complicating the interface with the antenna. The transmitter comprises double-balanced quadrature upconversion mixers followed by the front end shown in Fig. 4.8. The differential signal produced by the mixers is applied to the differential to single-ended (D/S) converter consisting of $M_1$–$M_3$ and $L_1$. Here, $L_1$ serves as a shunt-peaking element for the cascode and as a series-peaking element for the source follower. Owing to the finite output impedance of the follower and hence nonideal series peaking, the D/S conversion does not double the signal swing, but still helps to reduce the size of the output transistor, $M_4$.

To avoid a transmit/receive switch, this design directly shares the antenna between the TX output stage and the LNA. The choice of $W_4$ and hence the capacitance that it introduces at the output entails a trade-off between the transmitter output level and the degradation in the receiver noise figure. Designed for an output power of $-10$ dBm, $M_4$ raises the noise figure by 0.15 dB.

4.6 Toward Full-Band Transceivers

The design issues outlined above for Mode 1 devices prove much more severe as a larger number of MBOA bands are accommodated. In order to cover as many as 14 bands, both a broadband, low-noise signal path, and a large set of PLLs and SSB mixers become necessary.

We introduce here a frequency plan that employs only three phase-locked loops to cover the first 12 bands in Fig. 4.1. Illustrated in Fig. 4.9(a), the idea is to incorporate two RF PLLs at 5,280 and 7,392 MHz as the “base” frequencies and add or subtract increments of $(1, 3, 5, 7) \times 264$ MHz to obtain the center frequencies of all of the bands. Figure 4.9(b) depicts a simplified embodiment of the plan, where each SSB mixer senses one “base” input and one multiplexed increment input. Thus,
Fig. 4.9 (a) Choice of “base” frequencies, (b) simplified illustration of frequency synthesis for 12 bands, (c) generation of the base and increment frequencies

\( f_1 \) covers the center frequencies from 3,432 to 7,128 MHz, depending on whether the increments are added to or subtracted from 5,280 MHz. Similarly, \( f_2 \) covers the range 5,544–9,240 MHz. In practice, two sets of this arrangement are necessary so as to produce quadrature LO phases.

Figure 4.9(c) shows the details of the three-PLL synthesis method. The frequencies \((1, 5, 7) \times 264 \text{ MHz}\) are generated in quadrature form by proper division of the RF PLL outputs.

Of various spurious components produced in the architecture of Fig. 4.9, three require special attention: (1) \( f_{S1} = 7,392 \text{ MHz} - 3 \times (1,848 \text{ MHz}) = 1,848 \text{ MHz} \). This frequency lies within the transmit band of DCS1800 base stations; (2) \( f_{S2} = 7,392 \text{ MHz} - 3 \times (792 \text{ MHz}) = 5,016 \text{ MHz} \). This frequency is in the vicinity of the first IEEE 802.11a band (5,180–5,320 MHz), but still safely far; (3) \( f_{S3} = 5,280 \text{ MHz} - 3 \times (792 \text{ MHz}) = 2,904 \text{ MHz} \). This frequency is 104 MHz above the IEEE 802.11g band. Among these components, only the first appears serious, necessitating filtering of the 1,848-MHz signal in Fig. 4.9(b) to ensure a low third-order harmonic. Such a filter can be realized as a notch in subsequent buffer stages. Note that, by design, the unwanted components arising from phase and gain mismatch fall at the center of the MBOA bands but not on WLAN interferers. For
example, the mismatch-induced sideband at 7,392 – 1,848 MHz = 5,544 MHz lies outside both IEEE 802.11a bands.

An important issue in the above architecture stems from the multiplexing and routing of quadrature signals at frequencies as high as \( f_{2,\text{max}} = 9,240 \text{ MHz} \). Transistor and interconnect mismatches can introduce substantial phase and gain imbalance here, possibly requiring the use of calibration in the LO and/or RF signal path(s).

### 4.7 Experimental Results

The transceiver has been fabricated in 0.13-\( \mu \text{m} \) digital CMOS technology. Figure 4.10 shows a photograph of the die, whose active area measures approximately 1 mm \( \times \) 1 mm. The circuit is tested with a 1.5-V supply.

Figure 4.11 plots the measured frequency response of the LNA across the three bands with the corresponding noise figures shown on each plot. Due to inductor modeling inaccuracies, the center frequencies are shifted down and the gain is reduced in the upper bands. (The dashed plots indicate the desired characteristics.) These deviations can be corrected by adjusting the design of the inductors in subsequent silicon iterations.

Figure 4.12 plots the input return loss for high and low LNA gains. The magnitude of \( S_{11} \) remains above 11 dB across the three bands in both cases.

Depicted in Fig. 4.13 is the output spectrum of one PLL while the other two are off. It is observed that the reference sidebands fall well below 60 dBc as they approach the WLAN bands. The same holds for the other two PLLs.
Figure 4.14 shows the output spectrum of one PLL while the other two are on. The coupling of the oscillators through the supplies results in a relatively high level of spurs at the center frequencies of the other two bands, thereby lowering the tolerance of the receiver to interference produced by other UWB devices (but not WLAN transmitters). On-chip filtering of the supplies is expected to suppress this coupling.
Shown in Fig. 4.15 is the close-in output spectrum of the transmitter, in response to sinusoidal baseband signals, indicating an output level of $-11.7 \text{ dBm}$ (including a cable loss of 1 dB) and a plateau phase noise of approximately $-104 \text{ dBc/Hz}$. The sidebands arise from carrier leakage and I/Q mismatch, but their level appears to be adequately low for QPSK modulation.

Figure 4.16 plots as a function of the input level the data rate that the second channel of the receiver can detect with BER $= 10^{-3}$. In this test, a Tektronix arbitrary waveform generator (AWG520) provides quadrature baseband OFDM signals to a discrete upconverter, whose output is used to drive the UWB receiver. The RX quadrature outputs are sampled by high-speed Gage digitizers and fed to a Matlab
Fig. 4.16 Measured data rate as a function of input level

![Graph showing measured data rate as a function of input level.]

Table 4.4 Measured performance

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<table>
<thead>
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<tbody>
<tr>
<td>Voltage gain</td>
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<tr>
<td>Noise figure</td>
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<td>In-band 1-dB compression point</td>
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<td>High LNA gain</td>
<td>−27.5–29.5 dBm</td>
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<td>Low LNA gain</td>
<td>−9.5–12.5 dBm</td>
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<tr>
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<td>−12 dB</td>
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<tr>
<td>Low LNA gain</td>
<td>−11 dB</td>
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<tr>
<td>TX output 1-dB comp.</td>
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<tr>
<td>Phase noise @ 1-MHz offset</td>
<td>−104–106 dBc/Hz</td>
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<td>Technology</td>
<td>0.13–um CMOS</td>
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program, and the result is compared with the data produced by AWG520 to determine the BER.

Table 4.4 summarizes the transceiver performance. The range of values for the voltage gain, noise figure, $P_{1\text{dB}}$, and phase noise correspond to different bands. The TX output compression point is 4 dB lower than required and can be raised by increasing the bias current of the output stage.

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References


Chapter 5
Pulse-Based, 100 Mbps UWB Transceiver

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Abstract A pulse-based FCC-compliant ultra-wideband (UWB) transceiver is designed and integrated as a four chip and planar antenna solution. The signaling is based on 500 MHz-wide subbanded binary-phase-shift-keyed (BPSK) Gaussian pulses centered in one of 14 bands across the 3.1–10.6 GHz bandwidth. The system includes a UWB planar antenna, a Gaussian BPSK transmitter, a direct-conversion front-end, dual 500 MSps analog-to-digital converters, and a parallelized digital baseband for timing control and data demodulation. The RF local oscillators and baseband gain stages are implemented externally. A 100 Mbps wireless link is established with this chipset. A bit-error rate of $10^{-3}$ is observed at a $-84$ dBm sensitivity. This energy-aware receiver is implemented with strategic hardware hooks such that the quality of service is exchangeable with power consumption.

5.1 Introduction and System Overview

Ever since the Federal Communications Commission (FCC) released 7.5 GHz of unlicensed spectrum in the 3.1–10.6 GHz band for the development of ultra-wideband (UWB) radios [1], there has been a great effort to harness this new resource.

System-level advantages of UWB signaling over traditional narrowband radio signals include exploiting the increased timing resolution requirements of processing UWB signals for locationing and tagging applications [2, 3], biomedical imaging [4], and energy-efficient data transfer [5–7]. However, there are also challenges in implementing UWB wireless systems [8]. The following overview of the UWB signaling environment discusses these issues from a system perspective.
5.1.1 UWB Signaling Environment

UWB is meant to be an overlay technology to existing narrowband radios and appears “noise-like” to those systems. However, for extremely sensitive technologies such as GPS, a notch in the UWB spectral mask suppresses interference upon those systems [9].

Although the FCC limits for UWB transmission have been carefully engineered to avoid interference upon existing narrowband receivers, there are no provisions guaranteeing that in-band narrowband transmitters do not saturate UWB receivers. Furthermore with such large bandwidths, the channel model diverges from a simple gain correction to a complex multi-path problem. Therefore, the system must be designed with a sensible combination of signaling scheme, architecture, linearity and noise figure (NF) specifications, and solution to multi-path effects so that it can achieve a desired quality of service (QoS) for a given sensitivity, power constraint, and data rate.

5.1.2 Pulse-Based, 100 Mbps UWB Transceiver

In this approach, we have designed a pulsed-based FCC-compliant UWB binary-phase-shift-keyed (BPSK) direct-conversion transceiver such that QoS can be scaled as a function of power consumption. Specifically, the analog-to-digital converter (ADC) is designed such that it can scale with 1–5 bits of resolution [10] and the digital baseband can vary the number of RAKE taps it uses to perform channel estimation [11]. A 3.1–10.6 GHz RF Gaussian pulse generator up-converts the 500 MHz-wide BPSK UWB pulses [12] to one of 14 subbands with center frequency $f_{\text{center}}$ (5.1):

$$f_{\text{center}} = 2904 + 528 \cdot n_{ch} \ [MHz] \quad (5.1)$$

where $n_{ch} = 1, 2, \ldots, 14$. These pulses are propagated by a custom UWB planar antenna [13]. The BPSK signaling choice relaxes the linearity specification and the front-end low-noise amplifier (LNA) is unmatched for improved noise performance up to 10 GHz. The bandwidth of the UWB pulses are limited to 500 MHz so that out-of-band interferers can be suppressed after down-conversion by a baseband channel-select filter. An RF notch filter in the receiver is implemented to suppress known interferers at the 5 GHz ISM bands [14]. Figure 5.1 shows a block diagram of the transceiver system. The RF local oscillators (LO) and baseband gain stages are implemented externally.

This chapter is organized as follows. Section 5.2 surveys existing solutions to high data-rate UWB systems. Section 5.3 describes a discrete prototype system that is used to verify design choices and back-end algorithms for the integrated system. Finally, Sections 5.4–5.8 describe the hardware design involved for each block.
5.2 Related Work

Two predominantly implemented solutions for UWB systems are OFDM-based transceivers [15–20] and pulse-based transceivers [21–23]. Both solutions employ a direct-conversion architecture, a subbanded approach to UWB signaling, and achieve comparable data rates. The OFDM-based architectures use fast frequency-hopping of the UWB signals among adjacent bands to capture additional margin [24]. The <9.5 ns hop-time between bands can be met by providing all of the frequencies concurrently with multiple phase-locked loops (PLL) [15] or synthesizing the frequencies serially using a combination of PLLs, dividers, and single-sideband mixers [25–27]. For pulse-based systems, the UWB signals remain in one subband, thereby simplifying the synthesizer requirements. OFDM-based transmitters require a high-speed digital-to-analog converter (DAC) [28, 29], while pulse-based transmitters generate UWB signals through more efficient methods. Common solutions include generating baseband pulses with analog circuits and up-converting them to RF [22, 30], exciting a filter or antenna that shapes the pulse [3, 31–34], or synthesizing the RF pulse directly with no additional filtering [2, 35–37].

For OFDM-based systems, the digital baseband performs signal demodulation using an FFT and a one-tap channel equalizer per tone [38, 39], while pulse-based systems perform data recovery using a matched filter and a RAKE [39–42].
Other architectural methods for UWB pulse reception that reduce ADC sampling speed requirements include frequency channelized receivers [43] which improve immunity to narrowband interferers and analog correlation to trade off optimized bit-recovery for reduced power consumption [44, 45].

To mitigate unwanted interference to the receiver, subbanded LNAs [15], notch filters in the RF [46], and high $P_{1dB}$ front-ends [47] are employed.

### 5.3 Discrete Prototype

In conjunction to silicon development, a discrete prototype UWB system adapted from commercially available components was built to evaluate various design specifications upon system performance and the digital baseband algorithms [48]. These include ADC bit resolution, linearity, DC offsets, channel effects, antenna radiation, I/Q mismatch, and frequency drifts/offsets.

Figure 5.2 shows a block diagram of the UWB transceiver prototype. It can be partitioned into three distinct sections: the transmitter, the receiver, and the ADC with baseband processing. The link between the transmitter and the receiver can be made through wireless transmission using various antennae [13] and spatial configurations to emulate a wide range of channels. It can also be directly connected through a cable with attenuators to emulate an ideal channel. The following sections describe the prototype in more detail.

![Discrete prototype architecture](image-url)
5.3.1 Transmitter

The transmitter uses a direct-conversion architecture to up-convert a baseband signal to a center frequency of 5.355 GHz. This frequency is chosen based on the availability of wideband RF components that operate in this range. The baseband UWB signal is generated using either a programmable arbitrary waveform generator (AWG) or a dedicated pulse generator built with off-the-shelf components. The amount of memory in the AWG limits the transmitted signal to 4 ms of data sampled at 4 GSPs; however, this is long enough to store hundreds of data packets. Using an AWG enables a large amount of flexibility in the shape of the pulses transmitted, modulation scheme, and duration of transmission. For example, although this work focuses on pulse-based systems, OFDM can also be synthesized as long as a quadrature transmitter is not required. Various nonidealities may also be added to the AWG signal, such as nonlinearities or complex multi-path channels. In-band interferers such as 802.11a and random tones may also be added. The AWG is useful for implementing uni-directional, static communication with great flexibility, but cannot be used for transmitting dynamically changing data. In order to implement real-time communication, a dedicated pulse generator is used. This is built using discrete components and generates BPSK pulses with an up-converted bandwidth of 500 MHz. The pulse shape is fixed in the dedicated pulse generator; however, a variable pulse repetition frequency (PRF) up to 100 MHz is supported.

5.3.2 Receiver

The RF front-end is built entirely using discrete components. The received signal is amplified by two cascaded LNAs, then split and applied to two identical passive mixers performing I/Q direct conversion. The 90° phase shift in the LO is implemented by fixed, unequal delays in the LO transmission lines to each mixer. This method of phase shifting not only provides quadrature tones at 5.355 GHz, but also allows for tuning of the I/Q phase error simply by adjusting the RF center frequency. Tunable phase error is desirable in the prototype for testing the robustness of the digital baseband. After I/Q down-conversion, the baseband signals are filtered and amplified with an adjustable gain before digitization.

5.3.3 ADC and Baseband Processing

The baseband I and Q signals from the front-end are sampled by a dual-channel 8 b 500 MSps ADC board that interfaces to a PC directly through the PCI bus. The full-scale voltage, sample rate, and capture record length are all adjustable. The received samples are buffered and captured to a file for post-processing and demodulation. Once the samples are captured to a file, virtually any baseband algorithm not requiring real-time control of the system may be tested. This includes acquisition and fine
tracking, channel estimation, interferer rejection, and demodulation. Feedback loops such as automatic gain control require real-time sampling, and therefore cannot be tested using this acquisition board. An implementation of the digital baseband on an FPGA that provides basic real-time demodulation has also been demonstrated.

5.4 Transmitter

A transmitter has been fabricated in a 0.18 μm SiGe BiCMOS process as part of the custom chipset for a 100 Mbps pulse-based UWB transceiver [12, 49]. The goal of this work is to design a low-power UWB transmitter that emits Gaussian-shaped pulses. By exploiting the exponential behavior of a BJT, the Gaussian pulse can be accurately approximated with an elegant analog circuit that simultaneously performs up-conversion mixing to the 3.1–10.6 GHz band. Pulses are up-converted to one of 14 channel center frequencies described by (5.1).

5.4.1 Tanh Pulse Shaping

The transmitter uses a differential pair of BJTs with a triangle input signal to generate and shape a pulse of one polarity, shown conceptually in Fig. 5.3. For the proper choice of $A$, $PW$, and $V_{off}$, the current $i_{C2}$ will have a shape that approximates that of a Gaussian. For a fixed bias current $I_B$, the collector currents in the differential pair are approximately related by

$$I_B = i_{C1} + i_{C2}. \quad (5.2)$$

By substituting terms into the exponential equations for collector current of a BJT, it can be shown using hyperbolic identities that the collector current $i_{C2}$ is described by

$$i_{C2} = \frac{1}{2} I_B [1 - \tanh \left( \frac{V_{in}}{2 V_{th}} \right)] \quad (5.3)$$

where $V_{th}$ is the thermal voltage, equal to $kT/q$.

![Fig. 5.3 BJT differential pair and input voltage waveform for generating a tanh pulse](image-url)
The output current $i_{C2}$ of the differential pair will be a pulse with tanh-shaped rising and falling edges for the triangle input signal shown in Fig. 5.3. Note that the y-axis is normalized to $V_{in}/V_{th}$, the argument of the tanh function in (5.3). Current $i_{C1}$ is not used and can be terminated to the power supply. The pulse is simultaneously up-converted to the UWB band by additionally modulating the tail current $I_B$ with an LO. Furthermore, BPSK pulses are generated by inverting the LO signal in the tail current.

This architecture has several benefits. (1) The input signal begins and ends at the same level; thus, there is no “reset” phase required as in differentiating pulse generators, eliminating transients. (2) Positive and negative pulses can be generated with the same triangle input signal and inverted LO to improve matching, which is difficult to achieve with complementary circuits. (3) Up-conversion is performed by adding an LO signal to the tail current $I_B$; thus, no additional mixer is required. (4) The triangle signal can be generated with well-known techniques, and the accuracy of the Gaussian approximation is not sensitive to small deviations in the values of $A$ and $PW$.

The minimum mean-squared-error (MSE) between the tanh and Gaussian pulses is broad. This relaxes the requirements on the circuitry used to generate the triangle signal. It also relaxes the dependency of the pulse shape on temperature through $V_{th}$. Varying absolute temperature over a range of 28% results in an 11% variation in bandwidth and 0.2 dB in peak power for the pulse shown in Fig. 5.4.

### 5.4.2 Transmitter Architecture

A block diagram of the transmitter is shown in Fig. 5.5. The triangle signal is implemented off-chip, but is suitable for integration. To generate BPSK pulses, the triangle signal is switched to either the positive or negative input of the mixer. The inactive mixer input is simultaneously switched to a constant voltage. The same triangle signal is used to generate both polarity pulses to improve matching between

![Fig. 5.4 Time and frequency response of the optimized tanh pulse for $\sigma = 1.0, V_{off} = 1.0$](image)
pulses. The triangle signal switch also has an off state to implement variable PRF or a standby mode. The up-converted pulse is filtered and amplified on-chip before being DC-coupled to the off-chip UWB antenna.

5.4.3 Circuit Description

A schematic of the tanh pulse-shaping mixer is shown in Fig. 5.6. At the core are two tanh shaping pulse generators made by transistors \(Q_{3+/-}\) and \(Q_{4+/-}\). The tail currents of the pulse generators are modulated by LO signals. This enables simultaneous pulse shaping and up-conversion mixing. The LO signals to the two pulse generators are 180° out of phase, giving the inversion for BPSK pulse generation.

The LO signal is generated on-chip or can be switched to an external source and can be tuned from 3.1 to 10.6 GHz. The LO signal path is balanced to ensure equal amplitudes and 180° phase difference between the differential signals. The differential LO signals are converted to a current and mirrored, along with a bias current, into the tails of the two differential pairs.

Up-converted positive or negative pulses are generated by applying the triangle input signal to \(V_{in+}\) or \(V_{in-}\), respectively. The triangle signal voltage is relative to \(V_{cm}\), which is at a fixed potential. Applying the triangle signal to \(Q_{3+/-}\) with the bases of \(Q_{4+/-}\) fixed reduces unwanted signals from coupling to the output. The output currents of the differential pairs are summed at node \(V_{out}\). This provides first-order cancellation of LO feedthrough, similar to a double-balanced Gilbert cell mixer.

A schematic of the UWB band select filter and power amplifier (PA) is shown in Fig. 5.7. The mixer output is fed into the filter made by \(L_1, L_2,\) and \(C_1\), providing a 2nd-order roll-off below 3 GHz to reduce out-of-band emissions. The signal is then

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Fig. 5.5 BPSK UWB transmitter block diagram

Fig. 5.6 Schematic of the tanh pulse-shaping UWB mixer
buffered and AC coupled to the PA. The PA is class A, with an RF choke at the output, and can be DC coupled to the antenna.

### 5.4.4 Experimental Results

The transmitter was fabricated in a 0.18 μm SiGe BiCMOS process and wirebonded in a 48 lead MLF/QFN package. Due to a resonance with the bond wires at the input of the mixer, the pulses generated in channels 6–10 were distorted and did not have a Gaussian shape. An external LO was used as the center frequency of the pulses for all measurement results.

The matching between positive and negative pulses was evaluated at each of the center frequencies by measuring the peak power and bandwidth of a train of all positive or all negative pulses. An external LO was used to set the center frequency of the pulses, with an on-chip, single-to-differential converter to generate the differential LO used for inverting pulses. This converter has inherent mismatch in its differential outputs that varies with frequency, which can directly lead to mismatch between positive and negative pulse amplitudes. The spectral measurements fall into the high-PRF region for a PRF of 100 MHz and resolution bandwidth (RBW) of 1 MHz; therefore, a line spectrum is observed for a repetitive pulse train and peak power levels are independent of RBW [12]. The measurements of peak pulse power in each channel are shown in Fig. 5.8. The undistorted pulse responses demonstrated matching better than 2.5 dB.

![Fig. 5.8 Un-modulated peak power matching between positive and negative pulses](image_url)
The peak voltage $V_p$ of the pulse was measured at each center frequency with a high-speed sampling oscilloscope. The effective pulse width was approximated from the spectrum measurement by $\tau_{\text{eff}} \approx 1/(f_{\text{1st null}} - f_{\text{center}})$. This was used to predict the measured peak power level [12]. The calculated and measured results can be compared in Fig. 5.9. Each peak power data point is the RMS average of the negative and positive pulse powers at that center frequency. There is very good agreement between the calculated and measured peak powers. This peak power is of an un-modulated pulse train.

The average power of a pulse train modulated with random data is calculated from the measured peak power data and plotted in Fig. 5.9 [12]. These data are based on an RMS average of the positive and negative pulse peak power levels. The $-41.3$ dBm/MHz FCC limit is exceeded in some channels using nominal biasing; however, this can be corrected by reducing the gain adjustment in the power amplifier. The modulated peak power is expected to be 11 dB above the average power and follows a $10 \log(\text{RBW})$ trend in the high-PRF region [50]. For a 2 MHz RBW, the calculated peak power does not exceed the FCC limit of $-28$ dBm when the $-41.3$ dBm/MHz average power limit is not exceeded. A die photo of the transmitter is shown in Fig. 5.10.
5.5 RF Front-End

A block diagram of the direct-conversion RF front-end is shown in Fig. 5.11. The receiver consists of a single-ended LNA followed by an integrated 2nd-order high-pass filter for improved out-of-band interference rejection at minimal cost to NF. The broadband LNA is chosen as opposed to a 500 MHz subbanded LNA [15], due to the large number of required channels across the 7.5 GHz bandwidth. However, a more feasible trade-off to improve linearity and filtering while maintaining reasonable complexity would be to group multiple adjacent 500 MHz channels together into banks and to design an LNA that switches between these banks with a reduced number of switches and loads. The filter is followed by an RF single-to-differential converter with a switch-able notch filter to suppress unwanted 5 GHz ISM signals. The single-to-differential conversion done after the LNA helps to maintain low system NF and allows the RF notch filters to be easily implemented. The mixers are degenerated double-balanced Gilbert cell mixers. The LO for the mixers is generated externally. The baseband output is filtered, buffered, and interfaced to the remaining receiver chain via AC coupling to suppress 1/f noise and DC offsets.

5.5.1 Low Noise Amplifier

To leverage the high $f_t$ of the SiGe transistors for reduced NF, the UWB LNA here does not attempt to achieve impedance match [14]. Instead, the inductively peaked cascode LNA shown in Fig. 5.12 is viewed as a voltage amplifier and designed such that the NF to a 50 $\Omega$ source is minimized at 10 GHz, where propagation losses are greatest. Because an input match is not required, $Q_1$ is biased at a high $f_t$ for
low-noise performance with minimum current. The drawback to this method is that wideband preselect filters that improve out-of-band interference rejection can no longer be used, since they require a 50 Ω match at both ports. However, this problem can be partially mitigated with a post-LNA on-chip filter and/or an LNA with sufficient linearity. To improve immunity to bond wire, package, and PCB parasitics at the LNA load, a de-Qed capacitor shunts the RF current that would normally be sourced from the power supply to the low-impedance ground established at the LNA emitter by parallel down bonds.

The UWB LNA in Fig. 5.12 is biased with a high-impedance PMOS drain $M_1$ embedded in a $μW$ feedback loop that monitors collector current through $R_L$ and sets the $g_m$ of $Q_1$. This circuit contributes minimal noise due to the large output impedance and small drain noise current of $M_1$. The buffer is optimized to interface the LNA to the high-pass filter and single-to-differential converter. The simulated NF of the LNA is 2–2.3 dB over the entire band and provides a gain of 15 dB.

5.5.2 Channel-Select Filter and Single-to-Differential Converter

The channel-select filter shown in Fig. 5.13 is formed by a three-element high-pass ladder network with two capacitors and shunt inductor. The low-pass portion of the channel-select filter is formed by the cascaded amplifier roll-off.

The cross-coupled low-frequency PMOS loop in the single-to-differential converter equalizes current differences between the differential pairs to ensure balanced operation. The embedded switch-able notch filter uses parallel $LC_1$ and $LC_2$ tanks that degenerate the emitters of the differential pair transistors with a high-impedance at their respective resonances. This reduces the gain at those frequencies. The notch frequencies are fixed at $1/\sqrt{LC}$. The notch frequency of each $LC$ structure is designed at 5.25 and 5.75 GHz, respectively, where 5 GHz ISM signals may appear. The switch to bypass the notch filters is AC-coupled to the emitters of the differen-
tial amplifier transistor pair. This allows the switch to be reasonably sized while still having a low on-resistance because of the large overdrive.

5.5.3 Mixers and LO Amplifier Chain

The mixers are conventional double-balanced Gilbert mixers. The RF input is resistively degenerated to increase dynamic range (DR) and linearity while decreasing the load on the previous stage. Though resistive degeneration increases the NF of the mixer, the effective increase in system NF is marginal due to the two gain stages prior to the mixer. A 250 MHz 1st-order low-pass filter is implemented at the output for attenuating high-frequency mixing products and is assisted with a 3rd-order off-chip filter for further channel selectivity in the system. The AC-coupling capacitors are also off-chip.

The single-ended quadrature LOs are generated off-chip through a frequency generator and a wideband I/Q phase shifter. The on-chip LO gain stages convert single-ended LO signals to differential and refine the signal through a cascade of two differential amplifiers with low common-mode gain. On-chip amplification of the LO reduces the external bond wire coupling of LO–RF and LO–LO signals.

5.5.4 Measured Results

Figure 5.14 shows the measured receiver conversion gain and NF. When the notch filter is activated, the attenuation is 10 dB. The frequency of the notch filter is higher than expected due to over-accounted parasitics, but the functionality of an RF notch in a UWB receiver system can still be examined. This filter allows the receiver to
tolerate an additional 10 dB of interferer power within the notch frequencies. The measured input $P_{1dB}$ increases from $-46$ dBm at 3.1 GHz to $-36$ dBm at 10.6 GHz.

By embedding this front-end in the receiver prototype system, BER measurements are made at $f_c = 3.432$ GHz and $f_c = 7.128$ GHz in a clear channel environment. Figure 5.15 plots the two measured BER curves. Also included in the plot is a simulated ideal BER curve for Gaussian-shaped UWB pulse signals that are limited only by additive white Gaussian noise and quantization noise of a 4 b ADC.

The difference in lateral translation between the BER measurements corresponds with the general increase in NF at 7.128 GHz from the NF at 3.432 GHz. For RX
signals less than $-85$ dBm, the measured curves depart in shape from the ideal curve as they approach a BER of 0.5. This means that the signal to noise (and distortion) ratio of the system is decreasing by more than just the reduction in RX signal power.

To quantify the effectiveness of the notch filter for increasing the signal to interferer-and-noise ratio (SINR), BER measurements are made with two different 500 MHz-wide UWB pulses both with $-63$ dBm of signal power at the receiver, one centered at 3.432 GHz and another centered at 4.488 GHz. An interferer centered at the notch filter frequency is also added to the front-end input. As the interferer power is swept, BER measurements are taken. The interferer reduces SINR by producing intermodulation (IM) products that appear in the baseband frequency with the harmonics of the LO. Figures 5.16 and 5.17 show the oscilloscope plots of the IM products that lie in-band. For the 3.432 GHz pulses, the 114 MHz baseband beat frequency is the $IM_3$ of $2 \cdot f_{LO}$ and $f_{interferer}$. For the 4.488 GHz pulses, the 36 MHz baseband beat frequency is the $IM_5$ of $3 \cdot f_{LO}$ and $2 \cdot f_{interferer}$. These unwanted IM products can be suppressed with the notch filter. Figures. 5.18 and 5.19 plot BER versus increasing interferer power, with and without the notch filter. Though the 3.432 GHz pulses are further away from the 6.75 GHz interferer than the 4.488 GHz pulses, the effect of the notch filter is greater for pulses that suffer from higher order IM products. This can be understood by observing how the $IM_3$ amplitude coefficient is proportional to $P_{interferer}$, while the $IM_5$ amplitude coefficient for the 4.488 GHz pulses is proportional to $(P_{interferer})^2$. By comparing the SINR improvement of Figures. 5.18 and 5.19, the one suffering from $IM_3$ produces a 3 dB performance shift while the one that suffers from $IM_5$ produces a 6 dB shift. Correspondingly, the notch filter demonstrates one order-of-magnitude improvement in

![Fig. 5.16 Oscilloscope plots of pulses amidst $IM_3$ distortion](image1)

![Fig. 5.17 Oscilloscope plot of pulses amidst $IM_5$ distortion](image2)
Fig. 5.18 Measured BER gains for 3.432 GHz pulses with the notch filter

BER for the 3.432 GHz pulses and roughly four orders-of-magnitude improvement in BER for the 4.488 GHz pulses.

To obtain wireless BER measurements, the system is fixed 1 m apart in a line-of-sight configuration, and the transmitter power is varied to achieve the desired RX antenna power. A BER of $10^{-3}$ is achieved at a receiver sensitivity of $-84$ dBm, which matches with the wired BER measurements.

Table 5.1 shows the power breakdown of the front-end, and a summary of chip performance is reported in Table 5.2. The receiver in this work affords up to 10 GHz of subbanded UWB pulse reception in a QFN package with low NF through the band, but linearity could be improved with a coarse gain control setting in the RF gain path and improved linearity in the mixers. The die photo of the chip is shown in Fig. 5.20. The active area is $1 \text{ mm} \times 2.3 \text{ mm}$. 

Fig. 5.19 Measured BER gains for 4.488 GHz pulses with the notch filter
Table 5.1 Power consumption at 1.8 V

<table>
<thead>
<tr>
<th>Block</th>
<th>Power (mW)</th>
</tr>
</thead>
<tbody>
<tr>
<td>LNA</td>
<td>4.23</td>
</tr>
<tr>
<td>Buffer, S2D, and notch</td>
<td>13.5</td>
</tr>
<tr>
<td>2 mixers and buffers</td>
<td>21.6</td>
</tr>
<tr>
<td>2 LO amplifiers</td>
<td>14.4</td>
</tr>
<tr>
<td><strong>Total</strong></td>
<td><strong>53.73</strong></td>
</tr>
</tbody>
</table>

Table 5.2 Chip performance summary

<table>
<thead>
<tr>
<th>Specification</th>
<th>Value from 3.1 to 10.6 GHz</th>
</tr>
</thead>
<tbody>
<tr>
<td>Conversion gain</td>
<td>39 to 29 dB</td>
</tr>
<tr>
<td>Noise figure</td>
<td>3.3 to 5 dB</td>
</tr>
<tr>
<td>Input $P_{1dB}$</td>
<td>$-46$ to $-36$ dBm</td>
</tr>
<tr>
<td>Notch filter attenuation</td>
<td>10 dB (6.25–6.75 GHz)</td>
</tr>
<tr>
<td>Sensitivity (BER = $10^{-3}$)</td>
<td>$-84$ dBm (at 3.432 GHz)</td>
</tr>
<tr>
<td>Process technology</td>
<td>0.18 μm SiGe BiCMOS</td>
</tr>
</tbody>
</table>

Fig. 5.20 Die photo

5.6 ADC

To digitize the down-converted pulses, the 250 MHz-wide analog baseband pulses are sampled at the 500 MHz Nyquist rate. Previous work on pulse-based communication has shown that 4 bits of resolution are sufficient for proper reception in both noise- and interference-limited regimes [51]; in this ADC, 5 bits of resolution have been chosen to be conservative. This section begins with choosing an energy-efficient architecture for I/Q 500 MSps, 5 b ADCs, followed by a description of the circuit details. A bit-scaling mechanism to reduce power under favorable channel conditions is then introduced, and the section concludes with measured results of the prototype ADC implemented in a 0.18 μm CMOS process [10].
5.6.1 Architecture Selection

The flash topology, along with its interpolating and folding variants, has long been the preferred choice for high-speed, low-resolution ADCs [52–56]. However, the time-interleaved successive approximation register (SAR) topology has recently emerged as a low-power alternative [57], primarily due to its decreased complexity; the flash ADC requires $2^b$ comparisons versus $b$ for the SAR ADC. An excellent introduction to the operation of both a SAR and a flash converter can be found in [58].

In the past, time-interleaved SAR has been chosen over flash because of its reduced area [59] and reduced comparator power [60]. If comparator complexity is the sole metric, then SAR will outperform flash even down to the 2 b level. At UWB resolutions, however, overhead from digital logic and the capacitive DAC must be included to provide a meaningful comparison between SAR and flash. The digital power consumption, in particular, is nontrivial because digital feedback is present on the critical path between successive bit decisions. A simple model of the energy per conversion for a SAR ADC has three components:

$$E_{SAR} = E_{\text{comparator}} + E_{\text{caparray}} + E_{\text{digital}}$$

For a comparator with a two-stage preamplifier, the comparator energy scales as $b^2 2^{b/2}$, assuming that the energy is limited by the offset matching between channels and not by thermal noise. The capacitor array sizing is limited by matching and scales as $2^{2b}$. Finally, the digital energy scales linearly with resolution. Using values seeded from HSPICE simulations, (5.4) has been plotted in Fig. 5.21 versus resolution. Also plotted is the theoretical energy of a flash ADC, including only the power from the $2^b$ comparators; the resistive reference ladder and thermometer decoder are ignored. As seen, the digital logic dominates the SAR energy in the region of interest, and the capacitor array is negligible until medium-to-high resolutions. At 4 bits and below, this digital overhead makes the SAR converter less efficient than flash, but at the 5 b node, time-interleaved SAR is lower power and has been chosen for this design. Furthermore, the advantage of SAR is expected to improve in more advanced technologies because of the reduced digital power.

5.6.2 Circuit Design

Each of the I/Q ADCs has six time-interleaved SAR channels (Fig. 5.22) and each conversion requires one clock period to sample the input and five periods to resolve the bits. Therefore, with six channels, the channel’s clock is identical to the overall sampling clock. The channels communicate by passing a start token, which signals when to begin sampling. A start generation block produces the start token for the first channel of both the I and Q ADC. All critical sampling edges are aligned to the same 500 MHz clock; timing skew is minimized through careful layout of this
Fig. 5.21 Theoretical energy scaling with resolution of time-interleaved SAR and flash ADCs. Individual energy contributions of the SAR sub-blocks are also shown.

clock and the input signal. The channel is shown in Fig. 5.23 and consists of a DAC capacitor array, a comparator with two preamplifiers and a regenerative latch, and a digital control block. The comparator must have an input referred offset of less than half an LSB. This is necessary in a time-interleaved converter because any variation in the offset between channels produces spurs in the output spectrum that fall at multiples of the 83 MHz channel sampling rate. While a regenerative latch is an energy-efficient comparator, it tends to have a large offset. Two low-gain preamplifiers are used to amplify the input beyond this high offset level; output offset storage is used after the first preamplifier to mitigate its contribution to the overall offset.

The digital logic is the most power-hungry part of the ADC and limits maximum conversion rate. The logic has been designed using a combination of static CMOS combinational gates and dynamic registers, built upon the C²MOS logic family [61].

Fig. 5.22 Top-level block diagram of the I (or Q) ADC
By embedding most of the logic functions required in a single register, the propagation delay from the rising edge of the clock to valid outputs of the SAR controller is kept to $2t_{c\rightarrow q}$, where $t_{c\rightarrow q}$ is the delay from the rising edge to valid data for a register [10]. This full custom logic design reduces the digital power by 80% and is almost 60% faster than an equivalent SAR built using standard cells and static feedback flip-flops.

The timing of the ADC is shown in Fig. 5.24. The first period is the sampling operation; the assertion of $SAMP$ is always delayed from the rising edge of $CLK$ to avoid a high-frequency current impulse on the analog inputs at the same time that the previous channel finishes sampling. During bit-cycling, the ADC operates in two phases and uses self-timed bit-cycling to reduce the impact of digital propagation delay [62]. When $CLK_{BC}$, the bit-cycling clock, is high, the DAC and preamplifiers are settling and the latch is reset. Upon the falling edge of $CLK_{BC}$, the latch is strobed and begins regeneration. At the end of the first bit-cycle, the latch typically resolves very quickly. The rising edge of $CLK_{BC}$ is then triggered by the $DONE$ signal, and the next bit-cycle starts immediately. In effect, after the latch output has settled, the remainder of the clock period is borrowed by the DAC and preamplifiers for the next bit. In case the latch does not resolve in time (third period of Fig. 5.24), the rising edge of $CLK_{BC}$ is triggered by $CLK$, and bit-cycling continues.

### 5.6.3 Bit Scaling for Power Savings

For flexibility in response to varying system requirements, the ADC is capable of scaling the output resolution $b$ with 1–5 bits at a constant sampling rate. Since each conversion requires $b + 1$ periods, fewer channels are needed to maintain the same throughput; the unused $5 - b$ channels are disabled by halting propagation of the
START signal (Fig. 5.22), gating the clocks, and shutting off all bias currents. The start generation block triggers the first channel at a higher rate. When a channel receives the START signal, it samples during the next clock period, and the sampling operation automatically resets bit-cycling at the correct, lower resolution.

To reduce circuit board requirements between the ADC and the digital baseband, the outputs of the channels are not serialized, yielding an output frequency of 83 MHz. When not operating in the full 5 b mode, some of the channels are deactivated. Rather than burdening the baseband chip to sort the ADC outputs, a mux tree is implemented to maintain a constant 83 MHz output rate independent of the resolution, with the unused LSBs of each word forced to 0. Level conversion circuits are placed between the core and the I/O circuits to allow an internal digital supply lower than 1.8 V while still maintaining the proper voltage for inter-chip communication.

5.6.4 Measured Results

The dual ADC chip has been fabricated in a 0.18μm CMOS technology and occupies an active area of 1.1mm². All test results are for the ADC operating at 500 MSps. The ADC has a measured integral and differential nonlinearity of $-0.26/0.42$ and $-0.39/0.33$ LSBs, respectively. The measured signal-to-noise-plus-distortion ratio (SNDR) is 20 dB at DC and has an effective resolution bandwidth of 30 MHz, which is a result of insufficient settling time in the input sampling network. Still, the SNDR remains above 16 dB through the Nyquist band and is used in the full system prototype. Tones in the output spectrum due to interleaving products (offset and skew) are lower than $-35$ dB and do not limit overall performance. The power consumption of the dual ADC chip, not including digital output buffers,
is 15 mW at the full 5 b resolution; a 40% reduction in power is possible when the resolution is reduced to the 1 b mode.

5.7 Digital Baseband

The digital baseband detects and demodulates packets of UWB pulses. In order to compensate for multi-path and inter-symbol interference, it implements a data-aided channel estimation and uses the results in both a RAKE receiver [39] and a maximum likelihood sequence estimator (MLSE) [63]. Its block diagram is presented in Fig. 5.25. The baseband can adapt the required signal processing to the channel conditions and QoS received. The digital baseband receives six complex samples from the ADC aligned to a clock edge. Each sample is composed of 10 bits, split equally between real and imaginary parts. The signals are processed by a correlator block that performs several functions. It computes the correlations required to detect the presence of a data packet, estimates the channel impulse response, and implements the RAKE receiver. Timing synchronization is achieved with both a delay-locked-loop and a PLL [64]. The MLSE is implemented as an autonomous block.

5.7.1 Synchronization

The time required for packet synchronization is a critical specification of any high data-rate wireless system. The initial synchronization of the packet is achieved by correlating the incoming samples with a local replica of the pseudorandom

---

**Fig. 5.25** Digital baseband block diagram
sequence with the duty cycle that is expected in the received signal. The values of the correlation are compared with a programmable threshold. If the threshold condition is met, packet detection is declared. This baseband is capable of obtaining 150 correlations in parallel and achieves a 20μs average time to achieve packet detection.

5.7.2 Channel Estimation and RAKE Receiver

Channel estimation in UWB communications has been previously addressed in [65–68] to assess the signal energy capture in RAKE receivers as a function of the number of fingers. In these papers, an isolated monocycle is transmitted through the channel and the corresponding received waveform is recorded. The channel is approximated with a reconstructed channel containing $L_c$ branches. The degree of matching improves for larger $L_c$, and the value of $L_c$ is chosen to be the minimum that produces a sufficiently good match. This establishes the number of fingers that a RAKE receiver must possess to efficiently exploit the channel diversity. We are using a data-aided least-square error approach that matches that of [69]. In this article, the authors lump together the effect of the multiuser situation as additional additive white Gaussian noise. For the RAKE receiver used in this baseband, it is not necessary to separate the information of the pulse shape from that of the multi-path. Only the aggregate result of their convolution is required to implement the RAKE receiver as an approximation to the matched filter. This RAKE would implement a matched filter with the following channel impulse response:

$$h_{lp}[n] = \sum_{i=0}^{24} a_i \delta[n - i]$$  (5.5)

where the weights $a_i$ are complex numbers. To obtain this estimation, the channel impulse response obtained from 27 consecutive impulses is averaged and rounded to a maximum of 4 bits. It was ascertained from simulation that a maximum precision of 4 bits was required for the channel estimation, but fewer bits can also work with higher SNRs. It is assumed that the channel coherence time is much longer than the data packet so that the channel impulse response does not change during its duration. It is also assumed that the maximum length of the channel impulse response is 25 samples or, equivalently, 50 ns, implying that the taps associated to longer delays are negligible. For this reason and since the impulses in the preamble are separated by 60 ns, the channel impulse response can be estimated without inter-symbol interference.

The energy of the different taps is estimated and compared to a programmable threshold $T_h$. Those taps that do not meet the threshold are set to zero and not used for the matched filter, reducing the complexity required in the signal processing in the baseband. Figure 5.26, shows the difference in SNR requirements as the
threshold changes for channel models CM1 to CM4. This plot shows that there is a trade-off between QoS or SNR and the complexity of the signal processing employed.

5.8 Antenna

Optimal UWB pulse reception entails minimization of ringing, spreading, and distortion of the pulse at the transmit and receive antennas. This requires sufficient impedance matching and near constant group delay throughout the entire 3.1–10.6 GHz bandwidth. In addition, for compatibility with portable electronic devices, the antenna is required to be low profile, omni-directional, and efficient.

Figure 5.27 shows two ultra thin, low-profile, single-ended elliptical antennas (SEA) for use with UWB IC transceivers. The figure on the left incorporates a horizontal elliptical cutout to explore its effect on bandwidth and frequency.

Figure 5.28 shows a differential version of the elliptical antenna design (DEA), which can be used with UWB transceivers to create a fully differential front-end receiver. This eliminates the need for a single-to-differential converter at the LNA, which eases the RF front-end design. This design is embodied by a common ground plane, in which the transceiver can be housed, with bond wires extending from the positive to negative terminals of the antenna.

These antenna topologies are derived from recent work in circular and elliptical disc monopoles (CDM and EDM) [70–73]. While these antennae are extremely broadband and efficient radiators, they are not low profile. CDM and EDM antennae protrude perpendicularly from their ground planes, and therefore are not optimally compatible with an integrated circuit receiver. The differential and single-ended elliptical antennae presented here are planar and enable easy integration with an IC.

The key intuition behind SEA and DEA design is the understanding of the bandwidth effects at various higher modes within a circular resonator such as a CDM.
The roots of a Bessel function derivative characterize these closely spaced modes [73]. Since the antenna distance from the ground plane consistently increases symmetrically from the antenna feed, the impedance change from one resonant mode to another resonant mode is very small, and therefore enables a very large bandwidth from the fundamental resonant frequency on through much higher frequencies. The designs presented here differ from that of a CDM in that they are coplanar with their ground planes rather than perpendicular, and they are printed on dielectric substrate. As such, they have a tapered clearance area from the ground plane, which increases fringing capacitance and therefore may cause a slight decrease in the fundamental frequency. This has been indicated in simulation. The theoretical lower end frequency for a CDM is given as follows:

\[
\frac{c}{\lambda} = \frac{30 \times 0.24}{L + r} \text{GHz}
\]  

(5.6)

where \( L \) = disc height (cm), and \( r \) = equivalent radius given by

\[
2\pi rL = \pi r^2
\]  

(5.7)
Adjustment for ellipticity is achieved by defining \( L = 2 \times (y \text{ radius}) \) (cm) and \( r = (x \text{ radius})/4 \) (cm). The equivalent radius is derived by equating the planar disc area with that of a cylindrical wire (monopole) of height \( L \). It has been found that this equation can be applied to the DEA and SEA for quite accurate results in design, simulation, and measurements.

Each antenna was designed and simulated using CST Microwave Studio. Measured results were extremely close to simulated results, and impedance matching was easily achieved with MMCX to SMA adapters at the feed. It was found that slightly increasing the ellipticity ratio enabled a better impedance match with an increase in directivity. Also, better impedance matching for the bandwidth was generally achieved with closer placement of the radiating ellipse to the feed point; however, the optimal match was achieved at approximately 0.010”. In the designs presented, the radiating ellipse was placed 0.005” from the ground plane at the unloaded SEA feed, and 0.010” from the loaded SEA and DEA feeds. The slot was placed 0.010” from the feed point in the SEA and 0.005” from the feed point in the DEA. The frequency effects for each of these locations are not negligible, as can be seen in Fig. 5.29, which illustrates the return loss for each of the antennae presented.

The lower-end theoretical frequency for a CDM of this size is 3.15 GHz. Simulations of a CDM of these dimensions were in agreement with theory (achieving a lower-end frequency of 3.13 GHz). The measured lower-end frequencies of the antennae presented here are 3.09 GHz for the loaded SEA and 3.2 GHz for the unloaded SEA and DEA. This further solidifies the argument that the CDM equation can be used in designing the planar elliptical antennas. The loaded SEA seems to have an advantage in achieving better impedance matching throughout the UWB frequency band, as well as achieving a slightly lower fundamental frequency. This suggests that size reduction can be employed with further investigation of antenna-loading techniques. The DEA would be expected to achieve similar characteristics as the loaded SEA; however, the slot load distance from the feed in the DEA is

Fig. 5.29 VSWR for three versions of the elliptical antenna
twice that in the SEA, and the surrounding metal area also alters its frequency characteristic. Slight differences in the antenna feed could have also caused inconsistency in comparison to the loaded SEA and DEA. One notable characteristic of the DEA is that it had a slightly resonant point at 2.46 GHz, although not optimally tuned, which suggests that dual-mode 802.11b and UWB antennae are certainly achievable. This result was also observed in simulation.

Time domain pulse reception from these antennae has been qualitatively compared against a standard wideband double-ridged waveguide horn antenna with 1–18 GHz bandwidth and nominally 10 dBi gain throughout the UWB band. We propose that pulse distortion and dispersion should not occur during transmission and reception of the pulses within the UWB band. The pulse that is transmitted should ideally be the same pulse that is received, such that correct detection can be employed at the digital back-end of the UWB receiver. The transmitter system used to test the UWB antenna designs is based on a design from Intel labs [74], with block diagram shown in Fig. 5.30.

This system uses a clock and data generator, which provides a 100 MHz clock and data synchronized with the clock. This corresponds to a time between pulses of 10 ns. The clock is fed to an impulse generator, which generates sub-nanosecond pulses. The impulse generator is split into positive and negative pulses via a power splitter and pulse inverter. The positive and negative pulses are then fed to a double-pole, single-throw RF switch. The switch output is then filtered through a high-pass filter with a 3 GHz cutoff to provide transmission in the UWB frequency range. The signal is then amplified via a power amplifier, and then transmitted through a 1–18 GHz horn antenna.

Figure 5.31 illustrates the transmitted pulse from the horn antenna superimposed on the received pulse from the loaded SEA. Pulse reception measurement was similar for the unloaded SEA and the DEA. This test setup was conducted in a typical multi-path lab environment, and the reception distance was approximately 1.5 m. The transmitted pulse was measured directly at the amplifier terminals with a 30 dB attenuator. Each measurement was taken on a timescale of 500 ps/div. The measurements of the received pulse was taken directly at the antenna terminals at 10 mV/div. By the theory of reciprocity, it can be inferred that each antenna transmits the same way it receives.

This plot shows clearly that very little pulse distortion can be observed from the transmitted pulse to the received pulse. The same result was indicated in the measurement made with ideal wideband horn antenna at transmit and receive ends.

![Fig. 5.30 Transmit block diagram](image-url)
Figure 5.32 indicates the measured radiation patterns at 3.5 GHz in the azimuth and elevation planes of the antenna.

The maximum gain achieved in the azimuth and elevation planes are 2.1 and 2.7 dB, respectively. These values are consistent to within 2 dB for an octave of frequency. Radiation patterns were measured in MIT’s Lincoln Laboratory Millimeter-wave Anechoic chamber. Measured radiation efficiency of these antennae are approximately 93%.

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References


Chapter 6
Pulse-Based UWB Integrated Transceiver Circuits and Systems*

Yuanjin Zheng, Rajinder Singh, and Yong-Ping Xu

Abstract The chapter briefly describes UWB standards proposals followed by the transceiver system architecture, signaling, and modulation techniques for a pulse-based UWB system. Based on this, the circuit building blocks and transceiver chipset design are illustrated.

6.1 Overview of UWB Standards Proposals

6.1.1 MB-OFDM

Orthogonal frequency division multiplexing (OFDM) is a frequency modulation scheme [1]. OFDM divides the available frequency spectrum into several sub-bands, and then a low-rate data stream is transmitted over each sub-band using a standard modulation scheme, for instance phase shift keying (PSK). This makes equalization simpler at the receiver end as the effects of the channel are roughly constant over a given sub-band. For UWB system, multiband OFDM scheme was proposed to achieve high-data-rate transmission [2] providing a wireless personal area network (WPAN) with data communication capabilities ranging from 55 Mbps, to 480 Mbps. Support for transmitting and receiving at data rates of 55, 110, and 200 Mb/s is mandatory. Time–frequency interleaved OFDM (TFI-OFDM) is proposed in [2] for UWB high-data-rate application. Under this scheme 122 sub-carriers are adopted that are modulated using quadrature phase shift keying (QPSK). Convolutional coding is used with a coding rate of 11/32, 1/2, 5/8, and 3/4 for different data rate and bit-error-rate (BER) performance.

* The chapter briefly describes UWB standards proposals followed by the transceiver system architecture, signaling, and modulation techniques for a pulse-based UWB system. Based on this, the circuit building blocks and transceiver chipset design are illustrated.

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6.1.2 DS-UWB

Direct-sequence spread spectrum (DSSS) [3] is a modulation technique whereby the transmitted signal takes up more bandwidth than the data signal being modulated. Under this scheme carrier signals are needed to up-convert the full signal spectrum to an appropriate frequency band. A continuous string of PN (pseudonoise) code symbols, known as “chips”, are used for modulation. The PN code having a pseudorandom sequence of 1 and -1 values is multiplied with the data signal. Since PN code sequence has a much faster rate than the data signal, the energy of the resultant signal is spread over a much wider band. On the receiver side, de-spreading is done to reconstruct the transmitted data. The PN sequence on the receiver is the same as the transmitted one but needs synchronization for de-spreading to work correctly. Correlation of the synchronized PN sequence with the received signal enables reconstruction of the transmitted data. The synchronization, being challenging, adds to the receiver complexity. By utilizing different PN sequences, multiple access is made possible. This is the basis for the code division multiple access (CDMA) property for DSSS. It allows multiple transmitters to share the same channel so long as it is within the limits of cross-correlation of the PN sequences used for different users.

The UWB system proposed in [4] provides a WPAN with data payload communication capabilities ranging from 28 Mbps to 1320 Mbps. Direct sequence spreading of binary phase shift keying (BPSK) and quaternary bi-orthogonal keying (4BOK) UWB pulses is employed. Forward error correction coding (convolutional coding) is used with a coding rate of 1/2 and 3/4. The UWB system also supports operation in two different bands: the ‘low-band’ nominally occupying the spectrum from 3.1 GHz to 4.85 GHz and the ‘high-band’ nominally occupying the spectrum from 6.2 GHz to 9.7 GHz. In each band, a maximum of six piconet channels with their unique operating frequency and acquisition codes can be supported. A preamble is used for clock and carrier acquisition and receiver training. A short preamble is suitable for short-range communication which supports high bit rate. Thus the preamble length is 5 μS for a channel with high SNR and low channel dispersion. The preamble for a channel with nominal SNR is around 15 μS. A long preamble can be used for applications with an extended range; for example, a 30 μS preamble time is used for channels that have a relatively poor SNR or channels that are highly dispersive [4].

6.1.3 Impulse Radio

Impulse radio (IR) is also a form of spectrum spreading. It works by transmitting baseband pulses of very short duration, typically on the order of nanosecond or sub-nanosecond. By this way, the energy of the radio signals is spread from near DC to a few gigahertz [5]. The shape of the pulse is very important as it specifies
the frequency spectrum of the transmitted signal to ensure that the maximum emitted power is within the FCC-allocated frequency mask [6].

The use of transmitted signals with gigahertz bandwidths ensures that the multipath signals can be resolved with different delays of an order less than nanoseconds. This greatly reduces the multipath fading effects. The transmission power can also be reduced due to the reduction of fading margin and a relatively simpler design can be adopted for the receiver. Signal processing techniques can be used to counter the degrading effects of multipath. However, challenges arise while dealing with small time-resolution pulses. Several acquisition tests might be required in the base band before synchronization can be achieved.

IR scheme does not require sinusoidal carriers or any IF processing and hence greatly reduces the transceiver complexity and overall power consumption. Typically, pulse position modulation (PPM) with time hopping (TH) can be employed for frequency spreading and multi user multiple access requirements [7].

6.2 Pulse-Based UWB Signal, Modulation, and Transceiver System

6.2.1 UWB Signaling

UWB occupies a very large bandwidth for signal transmission. UWB transmission systems are defined as those having an instantaneous spectral occupation in excess of 500 MHz or a fractional bandwidth of more than 20%. The bandwidth and spectral mask for indoor communication systems assigned by FCC is illustrated in Fig. 6.1 [6]. It can be seen that the FCC-regulated power levels are very low (below $-41.3 \text{ dBm}$), which allows UWB technology to coexist with other services and systems due to the spread-spectrum techniques utilized.

Based on the FCC regulation, UWB is mostly suitable for the short-range, high-data-rate applications. From Shannon’s theory [8],

$$C = W \times \log_2 (1 + SNR),$$  \hspace{1cm} (6.1)

where $C$ represents the channel capacity in bits per second, $W$ is the channel bandwidth in Hertz, and $SNR$ is the signal-to-noise ratio. The channel transmission rate grows linearly with channel bandwidth but only logarithmically with $SNR$. In other words, the channel capacity increases at a much faster rate with bandwidth than with power. Thus UWB has the potential to offer very high data rates at low power consumptions.

As mentioned above, the shape of the transmitted pulse determines the frequency spectrum of the transmitted signal. A well-designed pulse shape allows maximum emitted power to be within the FCC frequency mask. This from Shannon’s theory affects the channel capacity as well. A variety of pulse shapes have been proposed and discussed in [9–14]. Among them, the most frequently employed pulse shapes
Fig. 6.1 FCC-regulated spectral mask for UWB indoor communication systems (cited from [6])

are the derivatives of Gaussian function. The time domain and frequency domain representations of the \( n \)th order Gaussian derivative \( p_n(t) \) and \( P_n(f) \) are given in equations (6.2) and (6.3) [14], respectively:

\[
p_n(t) = (-1)^{(3n+1)/2} n! \pi^{-1/4} e^{-\left(\frac{1}{2\sigma^2}\right)^2} \sum_{k=0}^{\lfloor n/2 \rfloor} \frac{(-1)^k 2^{n+1/4-k} \left(\frac{1}{2\sigma^2}\right)^{n/2+1/4-k} t^{n-2k}}{(n-2k)!k!\sqrt{(2n-1)!!}},
\]

(6.2)

where \( \sigma \) is the standard deviation of the Gaussian function which is associated with the pulse duration:

\[
P_n(f) = \frac{(-1)^n i^{n^2} (2\pi)^{n+1/4} (2\sigma^2)^{n/2+1/4}}{\sqrt{(2n-1)!!}} f^n e^{-2\pi^2 f^2 \sigma^2}.
\]

(6.3)

6.2.2 Transceiver System Architecture

The transceiver system architecture for a pulse-based (IR or DS) UWB is relatively simple compared to other wireless transceivers due to the absence of the intermediate frequency (IF) stage. A comparison between a basic UWB transceiver and a conventional narrowband transceiver is shown in Fig. 6.2a. The transmission of
UWB signals can be free of sine-wave carriers (“carrier-less short pulse” technique [15]) and does not necessarily require any IF processing. The required signal processing can be done in the baseband.

The detailed implementation of non-coherent and coherent UWB transceivers can be found in [16] and [17], respectively. Implementations of a few UWB transceiver architectures are discussed in Section 6.4.

6.2.3 Link Budget

An example of the link budget calculation for a pulse-based high-rate UWB system of 110 Mbps, 200 Mbps, and 480 Mbps data rates are shown in Table 6.1.

6.2.4 Modulation Scheme

In the design of a UWB system, the selection of the modulation scheme is a crucial task as it affects the transceiver complexity, data rate, bit error rate (BER), and robustness against interference. Thus, given the application requirements, a UWB system designer will have to consider a trade-off between different factors and select an appropriate modulation scheme. The commonly used modulation schemes include on–off keying (OOK), binary phase shift keying (BPSK), and pulse position modulation (PPM), etc. [18, 19].

6.2.5 Baseband Front-End Signal Processing

In a UWB system, baseband front-end signal processing can be applied for signal demodulation and to improve the bit error rate. Examples of such signal processing
Table 6.1  Link budget for high-rate UWB communication system

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
<th>Value</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Throughput (Rb)</td>
<td>110 Mbps</td>
<td>200 Mbps</td>
<td>480 Mbps</td>
</tr>
<tr>
<td>Average Tx power (P_T)</td>
<td>−11 dBm</td>
<td>−11 dBm</td>
<td>−11 dBm</td>
</tr>
<tr>
<td>Tx antenna gain (G_T)</td>
<td>0 dBi</td>
<td>0 dBi</td>
<td>0 dBi</td>
</tr>
<tr>
<td>f'<em>c = \sqrt{f</em>{\text{min}}f_{\text{max}}} : geometric center frequency of waveform ( f_{\text{min}} and f_{\text{max}} are the −10 dB edges of the waveform spectrum) (GHz)</td>
<td>3.5 + 1*I I = 0, ..., 6</td>
<td>3.5 + 1*I I = 0, ..., 6</td>
<td>3.5 + 1*I I = 0, ..., 6</td>
</tr>
<tr>
<td>Path loss at 1 m (L_1 = 20 \log_{10}(4\pi f'_c/c))</td>
<td>Min: 43.32 dB</td>
<td>Min: 43.32 dB</td>
<td>Min: 43.32 dB</td>
</tr>
<tr>
<td></td>
<td>Max: 52 dB</td>
<td>Max: 52 dB</td>
<td>Max: 52 dB</td>
</tr>
<tr>
<td>Path loss at d m (L_2 = 20 \log_{10}(d))</td>
<td>20 dB (d = 10 m)</td>
<td>12 dB (d = 4 m)</td>
<td>6 dB (d = 2 m)</td>
</tr>
<tr>
<td>Rx antenna gain (G_R)</td>
<td>0 dBi</td>
<td>0 dBi</td>
<td>0 dBi</td>
</tr>
<tr>
<td>Rx power (P_R = P_T + G_T + G_R − L_1 − L_2(dB))</td>
<td>Min: −74.32 dB</td>
<td>Min: −66.32 dB</td>
<td>Min: −60.32 dB</td>
</tr>
<tr>
<td></td>
<td>Max: −82 dB</td>
<td>Max: −75 dB</td>
<td>Max: −69 dB</td>
</tr>
<tr>
<td>Average noise power per bit (N = −174 + 10*I \log_{10}(R_b))</td>
<td>−93.59 dBm</td>
<td>−91 dBm</td>
<td>−87.19 dBm</td>
</tr>
<tr>
<td>(R_b = B)</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Rx noise figure referred to the antenna terminal (N_F)</td>
<td>7.0 dB</td>
<td>7.0 dB</td>
<td>7.0 dB</td>
</tr>
<tr>
<td>Average noise power per bit (P_N = N + N_F)</td>
<td>−86.59 dBm</td>
<td>−84 dBm</td>
<td>−80.19 dBm</td>
</tr>
<tr>
<td>Minimum Eb/N0 (S)</td>
<td>3.5 dB</td>
<td>7.5 dB</td>
<td>10 dB</td>
</tr>
<tr>
<td>Implementation loss (I)</td>
<td>1.0 dB</td>
<td>1.0 dB</td>
<td>1.0 dB</td>
</tr>
<tr>
<td>Link margin (M = P_R − P_N − S − I)</td>
<td>Min: 7.77 dB</td>
<td>Min: 8.18 dB</td>
<td>Min: 8.87 dB</td>
</tr>
<tr>
<td></td>
<td>Max: 0.1 dB</td>
<td>Max: 0.5 dB</td>
<td>Max: 0.18 dB</td>
</tr>
<tr>
<td>Proposed min. Rx sensitivity level</td>
<td>−82.3 dBm</td>
<td>−78.2 dBm</td>
<td>−75.1 dBm</td>
</tr>
</tbody>
</table>

Techniques include synchronization for coherent demodulation, rake receivers to combat multipath fading, and equalization to reduce inter-symbol interference (ISI), as shown in Fig. 6.2b. These techniques are introduced below.

6.2.5.1 Synchronization

A typical UWB pulse width is in the sub-nanosecond range (generally 0.1 ns~1 ns). In comparison with the pulse repetition time (typically 10 ns, for 100 Mbps data rate), the duty cycle is much smaller than 1%. The pulse, therefore, exists only for a relatively short period during each cycle. In order to accurately detect the presence of a pulse in each cycle, timing uncertainty must be small. At the receiver end, a low noise amplifier (LNA) should be followed by a sliding correlation receiver. The received signals are then correlated with locally generated template pulses. The synchronization algorithm utilizes the correlator output and adjusts the delay of the local template pulse. Once the template pulses are aligned with the received pulses, synchronization is achieved. Thus, the synchronization process includes two functions: acquisition and tracking.
Acquisition determines the rough delay of the received pulse with respect to the local template. For acquisition a serial searching through all potential delays is carried out, and the most likely delay is determined. This position is regarded as a coarse timing delay from which the tracking is initiated. A tracking loop is used to maintain the alignment of the two pulses by locking the template pulse delay to the received pulse. The tracking loop is able to minimize the phase difference between the received pulses and the template automatically within the adjustable (but constrained) phase error. A thorough analysis of UWB pulse tracking performance has been presented in [14] and [20].

6.2.5.2 Rake Receiver

In a typical wireless environment, a signal propagates through several paths from the transmitter to the receiver. A rake receiver collects these multipath components, decodes them independently, and sums them finally to reconstruct the original signal [21].

6.2.5.3 Equalization

In a bandwidth-limited channel, strong inter-symbol interference (ISI) can result [19]. An equalizer having a transfer function that is the inverse of the impulse response of the channel can be used to reduce ISI by compensating for the amplitude and phase distortion caused by the channel. To do so, a training sequence can be sent by the transmitter to help the receiver estimate the channel impulse response and then the receiver can adapt to the channel variations. A detailed analysis of equalization can be found in [21].
6.3 UWB Transceiver Building Block Circuits

6.3.1 RF front-end

6.3.1.1 LNA

Most LNAs in UWB transceivers can generally be categorized into two different types of circuit structures, namely a cascode amplifier with LC ladder input network and inductive peaking, such as in [22], and a single stage amplifier with inductive peaking and shunt feedback such as in [23]. Other LNA techniques such as distributed amplifiers and noise canceling types will not be addressed in this chapter. The two above-mentioned amplifiers utilize different input matching networks. As illustrated in Fig. 6.3, the LNA in [22] enables wideband input impedance matching within the UWB spectrum using a three-section passband Chebyshev filter input network. Good noise figure and high S21 are obtained through the optimization of the input transistor M1. Inductive peaking is employed for bandwidth enhancement.

For LNA design, there exists a trade-off between the noise figure and impedance matching. Thus, to obtain good input matching, the noise figure is generally limited at the lower end by the matching requirements. The LNA in [23] does not have this limitation as it allows the noise figure and input matching to be optimized separately through two different transistors. Figure 6.4 shows an application of the LNA in [17]. It uses a two-stage cascade of [23] for gain boosting purposes. In the first stage, the noise figure and input impedance matching are optimized through transistors M1 and M2, respectively, thus making it possible to obtain both good input matching and noise figure simultaneously. R1 and M2 are used to provide biasing for M1 while L1 and C1 at the output are used to boost the gain. In the second stage, the feedback network provides further bandwidth extension in addition to the inductive peaking.

6.3.1.2 Multiplier

For UWB correlation operation, the multiplier is required to have a large bandwidth, high conversion gain, and small systematic mismatch. The transconductance

Fig. 6.3 LNA with Chebyshev filter input network
multipliers to be reviewed here can be grouped into two categories (Types I and II) based on the cancellation method [24]. In both types, the two input signals are injected into the gate and source of the input transistors, and transistors operating in the saturation region are used to perform the multiplication using the $V_{gs}^2$ term in the following equation:

$$I_d = \frac{K}{2} [V_{gs} - V_T]^2 = \frac{K}{2} [V_{gs}^2 - 2V_T V_{gs} - V_T^2]$$ (6.4)

where $K$ and $V_T$ are the transconductance parameter and threshold voltage of the MOS transistor, respectively.

Type I uses the square device cancellation scheme (Fig. 6.5b) while Type II uses the single-quadrant multiplier cancellation scheme (Fig. 6.5a). Examples of the Type I multiplier are shown in Fig. 6.5 [24]. In Fig. 6.5a, the multiplier core consists of M1–M8, R1, and R2. In Fig. 6.5 b, the core consists of M1-M6, R1, R2, L1, and L2 (for bandwidth enhancement). The two schematics are similar as the local oscillator (LO) signals are injected through a source follower. However, Fig. 6.5a uses a separate source follower (M2, M3, M6, M7) for each transistor in the cross-coupled pair (M1, M4, M5, M8). In addition, M9–M16 are introduced in Fig. 6.5a to improve the port-to-port isolation.

It can be proven that both multipliers in Fig. 6.5 yield the output current of the form [24]

$$I_o = I_{o1} - I_{o2} = 4Kxy$$ (6.5)

where $x$ represents the small signal input for ‘RF’ port and $y$ represents the small signal input for the ‘LO’ port. As an illustration, the output voltage of the multiplier in Fig. 6.5b can be derived as follows.

The current in each branch is

$$I_{o1} = K(V_{gs1} + x - V_T)^2 + K(V_{gs3} - x - V_T)^2$$ (6.6)

$$I_{o2} = K(V_{gs2} - x - V_T)^2 + K(V_{gs4} + x - V_T)^2$$ (6.7)
Fig. 6.5 Type I multiplier schematics (a) with separate source follower for each input transistor, (b) with only two source followers

(a)

(b)

The differential output voltage can be expressed as

\[ V_{out} = Z_{out}(I_{o1} - I_{o2}) \] (6.8)

Since \( V_{gs1} = V_{gs2} \), \( V_{gs3} = V_{gs4} \), and M2 and M3 share the same gate terminal, from (6.6), (6.7) and (6.8), the output voltage can be written as

\[ V_{out} = 4Kx (V_{gs1} - V_T) - 4Kx (V_{gs3} - V_T) \]
\[ = 4Kx (V_{gs1} - V_{gs3}) \]
\[ = 4Kx (V_{s3} - V_{s1}) \] (6.9)

Since M5 and M6 act as source followers, if ideal level shifts are performed, the signal variations at the source of M5 and M6 should be the same as that at the gates. Since \( V_{s1} = V_{s5} \) and \( V_{s3} = V_{s6} \), the input voltage of \( y \) terminal is given by
An example of Type II multiplier is shown in Fig. 6.6. It is essentially a Gilbert cell with resistive load. Single-ended RF signal (LNA output) is applied to positive RF input terminal “RF+”, while the “RF−” is connected to a DC voltage reference together with a bypass capacitor to create an AC ground. The output current of this structure can be proven to be, if $x$ and $y$ are sufficiently small,

$$I_o \propto \mu xy$$

(6.12)

where $\mu$ is a gain factor related to transistor sizing $K$ and tail current $I_s$ of $M_7$.

### 6.3.2 Wideband Zero-IF Signal Processing Circuits

Wideband zero-IF circuits include LPF, VGA, and integrator, etc. A linear VGA is shown in Fig. 6.7. A linear transconductance with source degeneration is used as a VGA gain stage [25]. The VGA gain can be adjusted by changing the control voltage $V_{ctl}$. A simple DC offset cancellation technique integrated with the VGA is realized. Transistors $M_1$, $M_2$, biased in the linear region, and capacitors $C_1$ and $C_2$ form a low-pass filter. The low-pass filtered voltage is converted into differential current through transistors $M_3$ and $M_4$ and fed back to the input. Since it is configured as
Fig. 6.7 DC offset rejected VGA

Fig. 6.8 gm-C active low-pass filter

A negative feedback loop, the VGA can suppress gain at low frequencies and hence the DC offset. A differential $g_m - C$-active LPF is shown in Fig. 6.8. Transistors M1–M12 work as a linear V–I converter [26]. Together with on-chip capacitors, a simple $g_m - C$ LPF is realized.

6.3.3 UWB Pulse Generator

Most UWB systems use a Gaussian pulse for data transmission and reception. Assuming that the following Gaussian pulse, $G_0(t)$, is transmitted at time $t=0$

$$G_0(t) = e^{-2\pi \left( \frac{t}{\tau_m} \right)^2}$$

(6.13)

where $\tau_m$ represents the temporal width.

When the pulse travels through the communication channel and reaches the receiver, an attenuated and delayed version of the transmitted Gaussian pulse is received and can be expressed as

$$G_1(t) = -A \frac{4\pi}{\tau_m^2} \left[ 1 - \frac{4\pi \left( t - t_d \right)^2}{\tau_m^2} \right] e^{-2\pi \left( \frac{t - t_d}{\tau_m} \right)^2}$$

(6.14)
where \( A \) and \( t_d \) represent the attenuation factor of the communication channel and the time delay, respectively.

In a coherent communications system, an identical pulse has to be created in the receiver to recover the transmitted data. Thus, to generate a pulse represented by (6.14), a pulse generation circuit that performs the squaring, exponential, and second-order derivation functions has to be constructed. It can be implemented as a cascade of three stages as shown in the block diagram of Fig. 6.9 [27].

The detailed circuit topology is shown in Fig. 6.10 [27].

The squaring circuit is made up of resistor \( R \) and transistor \( M_1 \) operating in the saturation region, where the input source provides the bias current. The output of the squaring circuit can thus be expressed as

\[
V_{\text{sqr}} = -\frac{1}{V_{th}} (V_{\text{in}} - \frac{3V_{th}}{2})^2 + \frac{5V_{th}}{4} \quad (6.15)
\]

where \( V_{th} < V_{\text{in}} < 2V_{th} \) has to be satisfied and \( M_1 \) must be in the saturation region. For \( M_1 \) to operate in the saturation region, \( R \) should be chosen as in (6.16):

\[
R = \frac{2}{KV_{th}} \quad (6.16)
\]

The exponential function is performed by transistor \( M_2 \) in the weak inversion region and the signal can be expressed as follows:
\[ I_{DS2} = \kappa e^{V_{GS2}/\lambda} = \kappa e^{V_{sq}/\lambda} \quad (6.17) \]

Finally, the derivation operation is completed by the RLC network at the output and its trans-impedance can be expressed as

\[ T(s) = \frac{V_{out}(s)}{I_{DS2}(s)} = \frac{sR_L L}{R_L + sL + \frac{1}{sC}} \quad (6.18) \]

The final output voltage of the pulse generated can be simplified into equation (6.19) using (6.16)–(6.18) when \( R_L + sL \ll 1/sC \) is assumed. This approximation is quite accurate (assuming load resistance \( R_L \) is 50 \( \Omega \)) in the desired frequency range of 1–5 GHz as the values of the on-chip inductors and MIM capacitors are typically in the ranges of 1–10 nH and 0.3–6 pF, respectively, in RF circuits:

\[ V_{out}(s) \approx R_L L C s^2 I_{DS2}(s) \quad (6.19) \]

From equation (6.19), it is clear that \( V_{out} \) is a second-order derivative of the current \( I_{DS2} \). Therefore the pulse in (6.14) can be replicated in the receiver for coherent detection.

### 6.3.4 Precision timing generation circuits

Delay locked loop (DLL) [28] can be used to generate delayed clocks whose edge-triggered pulses can be synchronized with the received pulses. Suppose the transmission data rate is 100 Mbits/s, the time interval of two successive pulses is 10 ns. Assuming the width of the pulse to be synchronized is 1 ns, the delayed step of the DLL clock output may be as small as one tenth of the pulse width (i.e., 0.1 ns) to ensure a correct synchronization. Using the conventional delay line based DLL, 100 taps will be needed. This makes it difficult to implement. A two-stage cascaded DLL which can achieve 0.1 ns time delay resolution within 10 ns can be used instead. The block schematic is shown in Fig. 6.11.

The timing diagram of the two-stage DLL is shown in Fig. 6.12. The first DLL has ten delayed outputs (coarse delay) and each delay step is 1 ns. The second DLL further divides the coarse delay step of 1 ns into ten fine steps (fine delay) with 0.1 ns each. Thus, the two-stage cascaded DLL could output an arbitrary 0.1 ns delayed clock within a time span of 10 ns. Furthermore, only 20 delayed taps are required.

### 6.3.5 Synchronization Circuits

Due to the extremely short duty cycle of the UWB pulses, synchronization is of great importance in impulse radio UWB communication systems. The desired performance can only be achieved with precise synchronization. For this purpose, a complete solution to the synchronization problem is proposed in [29], which includes
two parts, namely acquisition and tracking. The aim of acquisition is to roughly determine the delay of the received pulse with respect to the local template. During the acquisition process, serial searching of all the potential delays is performed to locate the most likely delay. This position is regarded as a coarse timing delay from which tracking is initiated. The objective of the tracking loop is to lock the template pulse delay to the received pulse and maintain the alignment. The tracking loop is able to minimize the phase difference between the received pulses and the template automatically so long as it is within the adjustable phase error.

The acquisition loop is first shown in Fig. 6.13. It consists of a correlation receiver, a sampler, four detector branches, a decision maker, a precision time delay, and a pulse generator.

The sampler (at rate $T_f$) samples at the pulse repetition rate of $1 / T_f$. The sampled correlation value is matched to four shifted template Baker codes [29] in parallel. Here PN MF represents pseudo-noise-matched filter. The maximum absolute value of the four detector outputs is selected, which is then compared with a threshold in the decision maker. If this maximum value is larger than the threshold, the corresponding Baker code template is regarded to have synchronized with the received pulse in phase. In this case, decision maker will stop the operation of time delay (by $\tau$ each time) and the loop is locked. Otherwise, it is assumed that no
synchronization is set up. The time delayer then outputs a delayed clock signal, which controls the pulse generator to produce a further delayed pulse. A typical working curve is depicted in Fig. 6.14. The acquisition is accomplished so long as one of the four branches gives a larger value than the threshold.

The tracking loop is shown in Fig. 6.15. The tracking loop is essentially a digital delay lock loop (DLL), which can track the slow fading of the transmission channel. The tracking loop has the ability to minimize the delay difference between the received pulses and the local templates automatically. The sampler samples at the pulse repetition rate. The difference between the leading and the lagging branches is then fed into the loop filter, which is a typical low-pass filter (LPF), to estimate the DC component of the differentiated input. The loop will stabilize and reach its equilibrium point when the outputs of the leading and lagging branches are equal.

The system transfer function of the closed loop is presented. Assuming the impulse response of the loop filter as \( f(t) \) with \( F(s) \) as its Laplace transform, integrator gain as \( g_c \), multiplier gain as \( K_1 \), the transfer function of the loop is therefore [29]

\[
H(s) = \frac{K_1 K_{loop} g_c F(s)}{s + K_1 K_{loop} g_c F(s)}
\]  

(6.20)
As shown in [29], the response of the closed loop system to a delta, step, ramp, and parabola inputs can be pre-determined for various loop filter configurations. Such a tracking loop is a stable system.

6.4 UWB Transceiver Chipset Design

6.4.1 A Low-Power Non-coherent CMOS UWB Transceiver Chipset

6.4.1.1 Overview

As discussed in Sections 6.1 and 6.2, since the UWB technique employs a very wide signaling bandwidth and low emission power density, it can be potentially used in low-cost, low-power, and short-range communication applications. Adoption of a short pulse in UWB systems also offers the ability of precision location and tracking, as for example for systems that use the IEEE 802.15.4a standard.

There are two methods for the data demodulation in UWB communication systems. One is coherent demodulation, which needs precision timing synchronization between transmitter and receiver which greatly increases the system complexity. The other is non-coherent demodulation, which needs special devices such as a step recovery diode (SRD) to generate to detect the pulses [30]. In this section, a novel pulse amplitude modulation and non-coherent demodulation technique is proposed. A fully integrated CMOS transmitter and receiver (TRX) based on this technique were fabricated in a CMOS 0.18 μm technology. The implemented TRX can achieve transmission rate of 1–50 Mbps in non-coherent mode with low power consumption. This system can be potentially used in wireless personal area networks, smart sensor networks, miniature imaging systems, and vehicular radar systems applications.
6.4.1.2 System Architecture

In this design, a novel pulse amplitude modulation and a non-coherent demodulation scheme as shown in Fig. 6.16 is proposed. No synchronization system is required but a relatively high-data-rate transmission can still be achieved. Figure. 6.17 shows the block diagram of the proposed pulse-based UWB transmitter and receiver. The transmitter is accomplished by integrating a pulse generator, a pulse modulator, and a driver amplifier (DA). The receiver consists of a low noise amplifier (LNA), a multiplier, a demodulation driving amplifier (DDA), a variable gain amplifier (VGA), and a low-pass filter (LPF).

The transmitter generates amplitude-modulated signal (Fig. 6.16c) by multiplying a UWB monocycle pulse train (Fig. 6.16b) with a binary information data (Fig. 6.16a), which is emitted by the antenna after amplification by the DA. The UWB pulse occupies the FCC low band (3.1-5.1 GHz) and pulse repetition is at a frequency greater than 200 MHz. The binary modulating data rate can be up to 50 Mbps. The emission power density is less than \(-41.3\) dBm/MHz. To ensure the reliable demodulation in the receiver, the pulse repetition rate is much greater than the binary data rate (\(\geq 4:1\)). This scheme uses UWB pulses to over-sample the modulation data. The over-sampled pulse sequence is an ultra-wideband signal and it can be emitted by the antenna directly with a very low power density.

This UWB system directly modulates the information bits with extremely short pulses. Since the pulse occupies a large bandwidth, it can be directly emitted without

![Figure 6.16](image-url)
using a carrier. The system can achieve high data transmission rates and is robust in a multipath environment and immune to the interferers. However, it is normally necessary to synchronize the local template pulses with the received pulses in sub-nanosecond time scale, which is very challenging and increases the system complexity [30, 31].

The received weak pulse sequence is amplified by the LNA and subsequently fed to the two input ports of a multiplier concurrently, resulting in a squaring operation (Fig. 6.16d). This square operation de-spreads the pulse spectrum and thus has a high processing gain together with integration. The squared output is then amplified by the DDA to further improve the demodulation signal-to-noise ratio (SNR). The VGA is used to maintain a large dynamic range. The LPF is used to extract the envelope of the squared signal. The extracted envelope may be sharpened by a simple comparator, whose output can be regarded as the recovery of the transmitted information data (Fig. 6.16e). This demodulation is non-coherent since it recovers the data only through the received pulse sequence. To cover a distance of 6 m with transmission rates up to 50 Mbps, the receiver is designed with $-72$ dBm sensitivity and 7.5 dB noise figure (NF).

It should be emphasized that the recovered data at RX is already a delayed copy of the baseband binary data transmitted from TX. This means that the proposed transceiver can directly wirelessly convey the digital information by analog modulation/demodulation. An ADC and baseband processor can be further included for signal processing and multiuser capability. However, the requirement for ADC complexity has been greatly reduced, i.e., the sampling rate could be the transmitted data rate (pulse repetition rate), and resolution need only be 1–2 bits whereas in other UWB transceivers, 500 M–1 GHz sampling rate with 4–6 bits resolution.
is required. Therefore, the proposed architecture is simpler and potentially able to achieve very low power consumption.

The TRX can be easily modified as a coherent transceiver if pulse position modulation (PPM) is adopted in the TX and external synchronization is adopted in the RX [5]. In coherent mode, a much higher transmission rate can be achieved but the system complexity is also greatly increased [17].

6.4.1.3 Building Block Design

Pulse Generator and Modulator

Gaussian monocycle pulses are generated by the proposed CMOS pulse generator (PG) circuit shown in Fig. 6.18 [27], which consists of three cascaded stages. The three stages transform an input clock sequentially into square, exponential, and second-order derivative time functions, respectively.

The pulse generator circuit described above is designed based on a Chartered’s 0.18 μm CMOS technology and in-house RF device models with on-chip inductors and MIM capacitors. The simulation result in Agilent ADS is shown in Fig. 6.19. Assuming that the input signal $V_{in}$ of the pulse generator is a clocked sequence (top waveform), it can be seen that a short Gaussian pulse is generated at the output of the pulse generator corresponding to each ramp-up edge of the input voltage (bottom waveform). In our design, non-minimum channel length is chosen for M1 to minimize the channel length modulation effect and enhance the gain. Since the gate-to-drain voltage of M1 cannot change abruptly due to the capacitance of $C_{gd}$, the formation of another Gaussian pulse when the input voltage ramps down is suppressed. This is a desired feature for a UWB modulator, where only one Gaussian pulse needs to be produced during each cycle of the clocked input signal. The resulting pulse is approximately symmetrical around its peak center and has a bandwidth (−10 dB) of approximately 2 GHz. The largest swing (peak to peak) is of the order

![Fig. 6.18 Pulse generator and driver amplifier (reprinted, with permission, from Yuanjin Zheng, Yan Tong, Jiangnan Yan, Yong-Ping Xu, Wooi Gan Yeoh, Fujiang Lin, “A Low Power Noncoherent CMOS UWB Transceiver ICs”, In Digest of Radio Frequency integrated Circuits (RFIC) Symposium, 2005, pp. 347–350, June 2005. © [2005] IEEE)
of 30 mV and the RMS error (w.r.t. ideal) is within 3%, which meets the pulse amplitude requirement for UWB applications. The pulse width can be adjusted by changing the rising time of the input signal while its amplitude can be changed by adjusting the aspect ratio of M3.

The measured pulse, shown in Fig. 6.24, is approximately symmetrical and has a short duration interval of around 0.8 ns, similar to the simulated one in Fig. 6.19.

The monocycle pulse can be modulated by binary information data using a modified Gilbert cell in Fig. 6.20, acting as a pulse over-sampling modulation. The two inputs to the Gilbert cell are the binary information data (input to the top transistors) and the Gaussian pulses from the pulse generator (input to the bottom transistor), respectively. Since the output from the pulse generator is single-ended, the gate of the other bottom transistor is an AC ground. The tail current source is removed from the modulator to improve the output swing and linearity. A cascoded transconductor stage is used to improve isolation between the two input ports. The differential output is finally converted to a single-ended one. The modulator achieves a bandwidth of 2.0 GHz, conversion gain of 2.3 dB, and IIP3 of $-5.2$ dBm.
The DA is used to amplify and shape the modulated pulse signal into FCC spectrum mask and also acts as a matching network to drive the antenna. As shown in Fig. 6.18, it consists of three cascaded stages, where each stage is a common source amplifier with resistor and inductor shunt feedback. The bandwidth of the DA is improved by using the dominant pole bandwidth enhancement technique [32]. The PMOS-NMOS current reuse technique is employed to improve the $g_m$ and hence the power gain. In order to reject the low-frequency interferences, AC coupling capacitors are used. The measured DA has a power gain of 12.5 dB and NF of 5.5 dB with $-3$ dB bandwidth around 2.0 GHz from 3.1 GHz to 5.1 GHz.

The CMOS LNA described in Section 6.3.1.1 is used in this chipset.

Demodulator

The demodulator is realized using a standard Gilbert cell multiplier with resistive loads [24], as shown in Fig. 6.6. For coherent demodulation, the LO signal is applied from the local pulse generator and the LNA provides a single-ended RF signal. Since the RF signal is single-ended and connected to terminal “RF+”, the “RF−” is connected to a DC voltage reference and AC grounded. For non-coherent demodulation, the RF port is connected to the LO port, and thus the multiplier can act as a squarer. A large bandwidth is needed for this pulse multiplier and a good linearity can improve the receiver dynamic range. A measured multiplier has a conversion power gain of 3.2 dB, IIP3 of $-6.3$ dBm, and $-3$ dB bandwidth of 1.75 GHz. To further improve the bandwidth and linearity, an inductor can be serially connected with resistive load for bandwidth peaking [33].

The UWB receiver is essentially a direct conversion receiver (DCR). DC offset is a serious problem in DCR and is produced here by self-mixing between the leaked LO (RF) and the LO (RF) signal in the multiplier. DC offset can corrupt the signal, but more importantly, it may saturate the following gain stages [34]. To reduce the DC offset, some special considerations have been given to the design of the multiplier and VGA.

The squared output signal is still very weak. To improve the demodulation SNR and further drive the VGA, a DDA is inserted. Essentially it is a differential low noise amplifier, as shown in Fig. 6.21. M1–M4 acts as a differential cascode amplifier to provide better isolation and stability. L1 and L2 function as inductive degeneration at the source of M1 and M2, improving the linearity of the DDA. The output LC tanks (C1, L3 and C2, L4) are used to boast the gain in a wide-frequency range of 0.1–2 GHz. This is feasible since the Q factor of the on-chip inductors is only around 3.2. A power gain of $\sim10$ dB is obtained in the above-mentioned bandwidth with a
good IIP3 of $-4.2$ dBm and low NF of 4.5 dB. Another parallel resonant network, C3 and L5, resonates around 1.0 GHz, which improves the output impedance of the current source and the balance of the differential current outputs.

**VGA and LPF**

The reference designs for VGA and LPF are shown in section 6.3.2. The two-stage VGA is used to achieve a large gain range from $-10$ dB to 45 dB. The simulated frequency response of the cascading of VGA with LPF stages is shown in Fig. 6.22. The lower cutoff frequency is around 1.5 kHz and the higher cutoff frequency is around 100 MHz with offset rejection around 15 dB.

**6.4.1.4 Measurement Results**

The transceiver IC and circuits described above are implemented in a 0.18 μm CMOS technology. The chip is mounted directly on a Rogers PCB using
chip-on-board (COB) packaging. Figure 6.23 shows a measured UWB monocycle pulse at the output of the generator. The pulse is the second-order derivative of a Gaussian function as per the design. The maximum swing is around 35 mVp-p, and the pulse width is around 0.8 ns, translating to a bandwidth (−10 dB) of around 2.0 GHz. The pulse repetition rate can be changed by varying the input trigger clock rate since each pulse is generated by the clock rising edge. However, the pulse width and bandwidth are almost unchanged. A typical measured pulse sequence is shown in Fig. 6.24, which matches with the simulation result of Fig. 6.19 very well. Here, the pulse repetition rate is 200 MHz and the pulse duration is less than 1 ns.

The measured LNA gain is 8.5–9 dB with −3 dB bandwidth of 5.5 GHz (3–8.5 GHz). The measured S11 is less than −7 dB within the passband. Both S12 and S22
are good enough within the whole band. The estimated two-stage LNA thus has a gain of 18.0 dB with a bandwidth of 3.5 GHz.

The measured performance of the multiplier is shown in Fig. 6.25. The top two signals are the pulses generated from the above-mentioned pulse generator and fed to the RF and LO input of the multiplier, the bottom waveform is the output of the multiplier. The multiplier functions according to requirements and generates a symmetric output with a conversion voltage gain of approximately 6 dB. It can be seen that DC components are regenerated from the multiplied output, which could be used to determine the polarity of demodulation signals.

The chip microphotographs of the transmitter and receiver ICs are shown in Fig. 6.26 and Fig. 6.27. The chips are mounted on a PCB using COB and tested.

A typical measured transmitted information data pattern (in TX) and demodulated recovered data pattern (in RX) are shown in Fig. 6.28. Here the modulated signal is transmitted wirelessly between the transmitter and the receiver at a distance...
of 1 m through a pair of ultra-wideband antennas. UWB pulse repetition rate is 200 MHz. The modulation information data rate is 20 Mbps. In the receiver, the transmitted binary data pattern are demodulated and recovered without symbol error. The peak–peak amplitudes of demodulated signals are around 100 mV. A comparator can be employed after LPF to reconstruct the binary digital data. The measured transmission rate is up to 50 Mbps. The complete measured performance of the transmitter and receiver is reported in [16].

### 6.4.2 A Low-Power Coherent CMOS UWB Transceiver Chip

#### 6.4.2.1 Overview

With the continued growth in the integration density and clock frequencies of ultra-large-scale CMOS integrated circuits, interconnect technology is emerging...
as a major bottleneck to the improvement of IC technology. The semiconductor industry has sought to address this primary problem by increasing the thickness of the wires, using more exotic substrate materials with lower dielectric loss tangents and employing more sophisticated input/output drivers at the transmitter and receiver. However, all these potential solutions are costly, thereby making wireless interconnect technology an increasingly attractive alternative. A wireless interconnect technology uses a radio technology to provide communications between functions on a large integrated circuit chip (intra-chip) as well as communications between functions on separate chips (inter-chip) located on a multi-chip module or on a motherboard, where distances are measured in a few to tens of centimeters and data rates are in the range of Gbps. The wireless interconnect technology has become possible due to the confluence of wireless communications algorithms with radio-frequency silicon processes. For example, Floyd et al. have demonstrated a wireless interconnect technology with integrated antennas, transmitters, and receivers in a 0.18 μm CMOS process for intra-chip clock distribution at 15 GHz [35]. Chang et al. have implemented a wireless interconnect technology for inter-chip communications using capacitive coupling technique and Mizoguchi et al. using inductive coupling mechanism [36]. Wireless interconnect technology is a very new approach, and thus much work remains to be done before it can be a viable replacement for global wires. UWB employs ultra-wide bandwidth and very low emission power density, is robust in a multipath environment and immune to interference [5]. All of these features make it a great candidate for use in inter and intra-chip communication applications. In this section, we describe a novel wireless interconnect technology that features the use of a UWB impulse radio for inter-chip communications. The performance of impulse radio over an inter-chip wireless interconnect channel is analyzed in section 6.4.2.2. We present the hardware design and measured performance of a UWB impulse radio in 0.18 μm CMOS technology in section 6.4.2.3.

6.4.2.2 Impulse Radio Transceiver Chip Design

We describe here the design of an impulse radio transceiver in a 0.18 μm CMOS process to implement an inter-chip wireless channel that can support data rates as high as PCIe. The block diagram of the proposed UWB impulse radio transceiver architecture is shown in Fig. 6.29. The building blocks of the UWB transmitter comprises a UWB Gaussian pulse generator, modulator, and UWB driver amplifier (DA). A Gaussian pulse generator generates a UWB Gaussian pulse that is subsequently modulated, amplified and transmitted wirelessly. The receiver consists of a UWB low noise amplifier (LNA), a correlator (including a multiplier and integrator), an analog-to-digital converter (ADC), and a clock generation and synchronization circuits. The UWB LNA is matched to the UWB antenna by means of a matching network. The purpose of the UWB LNA is to amplify the received pulses to a suitable level for signal processing as well as to provide enough gain so as to overcome noise in subsequent stages. The data is subsequently recovered by the correlator. The ADC is used to convert the analog demodulated signal
into digital form. The digital baseband provides control for the clock generation, synchronization, and data processing.

LTCC Antenna

An impulse radio requires a UWB antenna. The design of the UWB antenna in CMOS should be avoided because of the lossy nature of the CMOS substrate. A LTCC antenna has been designed for this transceiver [37].

Pulse Generator and Modulator

The monocycle pulse generation circuit of Fig. 6.10 is adopted in this design as well. The pulse modulation circuit used here is shown in Fig. 6.30. A pseudo-differential pulse generator is composed of R1, M1, M2, M3A-M3B, M4A, L1, C1, L2, and C2. M3A works in the weak inversion region to provide the exponential current function. M3B is biased by a DC voltage equal to VGS9 and realizes pseudo differential operation. Thus, two pulses with opposite polarity are produced at the output of the pseudo-differential pulse generator or the drains of M4A and M4B. Cascaded transistors M4A and M4B are employed to improve reverse isolation. M7 and M8 form a current source with high output impedance to provide the tail current. The modulation is performed by two switches, M5A and M5B. M6 and R4 are used to generate a complementary control signal. Thus M5A and M5B turn on alternately, passing the current to the 50Ω load at the output to produce either a positive or a negative monocycle pulse. The polarity of output pulse is determined by Vctrl. The detailed simulation results of pulse generator and modulator have been reported in [27].
Driver Amplifier

The design of the driver amplifier (DA) involves complex trade-offs among noise gain, matching, power consumption, and linearity. The schematic of the UWB DA is the same as that in Fig. 6.18. It consists of three cascaded stages. Each stage is a common source shunt–shunt feedback amplifier with resistor and inductor feedback, where PMOS-NMOS current reuse technique is employed to improve the $g_m$ or the gain without sacrificing the bandwidth.

Low Noise Amplifier and Multiplier

The CMOS LNA design shown in section 6.3.1.1 is adopted in the receiver. A symmetrical multiplier, as shown in Fig. 6.5a, is employed here to implement the squaring function. The multiplier core consists of transistors M1–M8 and R1 and R2. M9–M16 are cascaded to improve the port-to-port isolation for reducing the output DC offset. For non-coherent demodulation, the RF port is connected to the LO port, and thus the multiplier can act as a squarer. When the external synchronization mechanism is available, the multiplier can be used as a correlator for coherent demodulation, where the transmission rate is expected to be much higher. This multiplier achieves 5 dB conversion gain and 1.75 GHz bandwidth. The simulated performance of pulse multiplication is shown in Fig. 6.31. The top two waveforms are RF and LO inputs, respectively. The third one is the output of the multiplier.

Integrator

In order to maintain a steady integration level in response to a pulse input and hold it for at least 5 ns (for 100 Mbits/s transmission rate) to facilitate the A-to-D conversion, the integrator loss should be minimized. This implies a very low $-3$ dB frequency cut-off. On the other hand, the integration response time should be short...
so that the ADC would have enough time to perform conversion. This means that the integrator should have high slew rate and fast settling behavior.

The block diagram of the integrator is shown in Fig. 6.32. Because of Miller effect, the impact of the parasitic capacitance at the output of the Gm cell is reduced and the requirements for output impedance and output swing of transconductor are relaxed. The 3 dB bandwidth of the integrator is given by \( (R_{\text{out, gm}} C A_{\text{OTA}})^{-1} \) where \( R_{\text{out, gm}} \) is the output resistance of the Gm cell and \( A_{\text{OTA}} \) is the DC gain of the OTA. Clearly, it is lower than that of the Gm-C integrator by a factor of OTA dc gain. To achieve high integration speed and short rising time of less than 1 ns, a large output current is used to obtain high slewing rate. Another parallel transconductor is used to create a feedforward path to boost the high-frequency response and quickly build the rising edge of the output signal.

Clock Generator and ADC

An ADC employing full flash architecture and sampling rates of up to 500 Msamples/s can be employed after the integrator and 4-bit resolution is enough for baseband signal processing including synchronization and equalization.
Clocks for the pulse generator, integrator, and ADC are provided externally. Two cascaded delayed lock loops (DLL) are used as multiphase clock generator and synchronizer which provide delayed shifts of the clocks. The DLL can work at clock rates from 100 MHz to 500 MHz. The clock output has an output swing of 1 V, jitter of less than 30 ps and duty cycle of ~50%. The first DLL delays the clock by the minimal step of 1 ns (coarse delay), and the second DLL further delays the clock with a resolution of 0.1 ns (fine delay). By cascading two DLLs with a proper logic control, the output clock can be shifted with a step of 0.1 ns within 10 ns. Coherent demodulation can take place once the DLL is locked to synchronize the local pulses (triggered by a shifted clock) with the received pulses.

Taking 100 Mbps transmission rate as an example, the clock timing for pulse generator, integrator, and ADC is illustrated in Fig. 6.33. During each period, assuming that the clock to the multiplier is synchronized with the received pulse, the same clock is used for the integrator. The integration starts when the clock is high and lasts for half of a clock cycle. The other half cycle is used for discharging the integration capacitor. The tracking period of ADC is slightly ahead of the discharging time so that a stable sample can be obtained.

6.4.2.3 Experimental Results

The measured typical BPSK-modulated pulses after the antenna are shown in Fig. 6.34. Here the pulse repetition rate is 100 MHz, the pulse swing (at the output of transmitter) is adjustable in 40–800 mV with pulse width less than 1 ns. With the further spectrum shaping by the antenna, the measured spectrum of the pulse sequences with a $-10$ dB bandwidth from 3.1 GHz to 4.4 GHz, as shown in Fig. 6.35.

The measured results of the DA are given in Fig. 6.36. The power gain is 19–21 dB and the 3 dB bandwidth is from 1.1 GHz to 9.2 GHz, which covers the full UWB, although the transceiver described in this section is only required to
work in the FCC-released low band of (3.1–5 GHz). The DA has a low noise figure of 3.2–3.6 dB and an IIP3 of $-10 \, \text{dBm}$ measured at the center frequency of the passband, i.e., 5.05 GHz. As shown in Fig. 6.37, the measured LNA has a power gain ($S_{21}$) of 18 dB with 0.5 dB passband ripple, input reflection coefficient ($S_{11}$) $<-10 \, \text{dB}$, NF of 4.0–4.6 dB, and $-3 \, \text{dB}$ bandwidth of around 2.5 GHz.
The measured integrator performance is shown in Fig. 6.38. It is similar to the simulation result for the integration rising and holding time. However, the discharge rate is slower (200 Sample/s). This may be due to the higher leakage current of the discharge path than the simulated values.

The tested maximum pulse transmission rate is 200 Mbits/s. The microphotograph of the impulse radio transceiver chip is shown in Fig. 6.39, and the transceiver performance is summarized in Table 6.2.

### 6.4.3 A CMOS Carrier-Less High-Rate UWB Transceiver for WPAN Applications

#### 6.4.3.1 Overview

Both MB-OFDM and DS-UWB transceivers need carriers for up-and down-conversion of the baseband signal. Therefore a frequency synthesizer is necessary, which increases the system complexity. UWB impulse radio directly modulates...
the information bit with extremely short-duration pulses. Since the pulse occupies a large bandwidth, it can be directly emitted without using a carrier. In this work, a carrier-less impulse radio-based high-rate UWB transceiver utilizing a new modulation scheme is developed. It requires fewer components and has low power consumption. The transceiver, implemented in CMOS technology, works in the FCC-released low band of 3.1 – 5 GHz.

### 6.4.3.2 Transceiver Architecture

Figure 6.40 shows the block diagram of the impulse-based UWB transceiver. The TX generates pulse position modulated (PPM) UWB high-order derivative pulses that are emitted by the UWB antenna. At the RX, the received pulses are weak and first amplified by the LNA. The amplified pulses are then correlated with the local pulses, further amplified and integrated to a constant level for A/D conversion. As such, the signal modulation and demodulation are both completed in the analog domain.
As shown in Fig. 6.41, a PPM scheme is proposed to directly modulate the digital baseband signals to UWB pulses. The baseband NRZ (non-return-to-zero) data (a) is used to drive the TX pulse generator (PG). Since each pair of input data edges generates a pair of UWB fifth-order derivative of Gaussian pulses through the TX circuits, the location of the generated pulses (b) is therefore modulated by digital data (a). At the RX, the first derivative of Gaussian pulses (c) is generated and synchronized to the received pulses (b), and the pulse multiplication of (b) and (c) generates (d). The integration of (d) yields outputs (e) that recovers the transmitted data (a).

The shape of the pulse is of vital importance to the UWB systems. In general, it dictates the frequency spectrum of the transmitted signal. A well-designed pulse shape ensures that the maximum power emitted from the transmitter is within the FCC frequency mask. The most frequently employed pulse shapes are the derivatives of Gaussian function. The time-domain waveform and power spectrum density (PSD) of the \( n \)th derivative of Gaussian pulse are shown in Fig. 6.42 [27]. It can be seen that the second derivative of Gaussian pulse cannot fit into the FCC mask while the fifth derivative of Gaussian pulse can be fit in easily. The pulse generator circuits can generate positive and negative monocycle pulses from the corresponding rising and falling edges of the input clock in [27]. Based on this, in the presented TX circuits, the on-chip low Q LC networks (formed with pulse generator and pulse-shaping amplifier) have a wideband bandpass filter characteristic which acts as a higher-order derivative circuit in the RF band 3.1–5.1 GHz. The measured results show that the approximate fifth derivative of Gaussian pulses is generated [38].
6.4.3.3 Building Blocks

The transmitter circuits are shown in Fig. 6.43. A CMOS pulse generator (PG) circuit is employed to generate monocycle pulses, i.e., second derivative of Gaussian pulses [27]. After being amplified by a wideband pulse amplifier (WPA), the pulses are then shaped by a pulse-shaping amplifier (PSA). The output of the PSA is matched to the TX antenna. A three-stage wideband LNA implemented in 0.18μm CMOS technology is shown in Fig. 6.44. The first stage employs a cascode structure (M1 and M2) with wideband LC ladder matching (L1/ L2 and C1/ C2), which
Fig. 6.43 Pulse TX circuits (reprinted, with permission, from Yuanjin Zheng et al. “A CMOS Carrier-less UWB Transceiver for WPAN Applications,” *ISSCC Dig. Tech. Papers*, pp. 116–117, 2006. © [2006] IEEE)


boosts the gain, minimizes the noise figure (NF), and enables matching to the RX antenna [22]. Each of the last two stages is essentially an inductively peaked shunt feedback amplifier. Figure 6.45 shows the on-wafer testing results of the second stage of the LNA alone; the third stage is identical to the second stage. The measured gain is 8.5–9 dB, −3 dB bandwidth is 5.5 GHz (3–8.5 GHz), and S11 is less than −7 dB to −15 dB in the band of interest. The measured three-stage LNA has a gain of 20.2 dB and bandwidth of 4.5 GHz (3–7.5 GHz).

The subsequent LPF rejects the strong out-of-band interference and suppresses the leaked high-frequency components of pulse signals. The two-stage cascaded variable gain amplifier can achieve a dynamic gain range from −10 dB to 45 dB with 300 MHz bandwidth. A low-pass feedback loop is employed to reject the DC offset with a cutoff frequency of 600 kHz. The measured VGA frequency response is shown in Fig. 6.46. The Gm-C-OTA integrator achieves a low −3 dB bandwidth of 1 MHz and a high unity-gain bandwidth of 1 GHz. This provides a high integration
Fig. 6.45 Measured LNA performance (a) for the first stage and (b) second stage alone

Fig. 6.46 Measured VGA performance

gain and a long holding time. The steady integration value can hold for 10 ns with only <1% error due to charge leakage.

A PLL with ring oscillator is used for clock generation while two cascaded delay locked loops (DLL) are used for synchronization. The two DLLs are capable of delaying the clock with minimum steps of 1 ns and 0.1 ns, respectively, which is illustrated in Section 6.3.4.

6.4.3.4 Transceiver Performance

The transceiver IC is integrated in a 0.18 μm CMOS technology and occupies a die area of 2.6 mm × 1.7 mm. The RX achieves a NF of 7.7–8.1 dB, IIP3 of −12.3 dBm,
sensitivity of $-80$ to $-72$ dBm. The measured power consumption with a 1.8 V supply for the TX and RX is 86 and 81 mW, respectively. The transceiver performance is summarized in [38]. Figure 6.47 shows the chip microphotograph.

6.5 Summary

In this chapter, three UWB standard proposals named MB-OFDM, DS-UWB, and impulse radio have been reviewed. Pulse-based UWB technique is investigated in depth. UWB pulse modulation techniques, signal processing, and building block circuit design techniques are discussed. Three pulse-based UWB transceiver chipsets for different applications are presented that include a low-power, low-rate, non-coherent UWB transceiver for wireless sensor network applications; low-power, high-rate, coherent UWB transceiver for intra–inter chip communications; and carrier-less, high-rate UWB transceivers for WPAN applications. Furthermore, a DS-UWB implementation for a dual-band UWB transceiver can be found in a recent work presented in [39].

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