Analysis and Design of CMOS Ultra Wideband Receivers

A dissertation submitted in partial satisfaction of the requirements for the degree Doctor of Philosophy in Electrical and Computer Engineering (Computer Engineering)

by

Mahim Ranjan

Committee in charge:

Professor Lawrence Larson, Chair
Professor Peter Asbeck
Professor Chung-Kuan Cheng
Professor Lawrence Milstein
Professor William Trogler

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The dissertation of Mahim Ranjan is approved, and it is acceptable in quality and form for publication on microfilm:

Chair

University of California, San Diego

2006
To my wife Preeti,
    my mother,
my brother Aseem, my sister Megha
    and the memory of my father,
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VITA

1999
B.Tech., Electrical Engineering,
Indian Institute of Technology, Bombay

1999–2000
Research Assistant, Dept. of Electrical and Computer En-
gineering,
University of California, San Diego

2000
Summer Research Intern, Conexant Systems,
Newbury Park, California

2000
M.S., Electrical and Computer Engineering,
University of California, San Diego

2001–2003
RF Design Engineer,
Magis Networks, Inc, San Diego

2004 - Present
RF Circuit Design Engineer, Qualcomm, Inc, San Diego

2006
Ph.D., Electrical and Computer Engineering,
University of California, San Diego

PUBLICATIONS

M. Ranjan and L. Larson, “Distortion analysis of ultra wideband OFDM receiver front-
ends,” accepted for publication in IEEE Trans. of Microwave Theory and Techniques.

M. Ranjan and L. Larson, “A low-cost and low-power CMOS receiver front-end for

M. Ranjan and L. Larson, “An Analysis of Cross-modulation Distortion in Ultra Wide-

M. Ranjan and L. Larson, “A sub-1 mm$^2$ dynamically tuned CMOS MB-OFDM 3-to-

M. Ranjan, K.H. Koo, G. Hanington, C. Fallesen, and P. Asbeck, “Microwave power
amplifiers with digitall-controlled power supply for high efficiency and high linearity”,
ABSTRACT OF THE DISSERTATION

Analysis and Design of CMOS Ultra Wideband Receivers

by

Mahim Ranjan

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Professor Lawrence Larson, Chair

This research focuses on the design and analysis of Ultra Wideband receivers for Multiband Orthogonal Frequency Division Multiplexing systems. A comprehensive analysis of distortion effects in UWB receivers for different interference scenarios is performed. Analytical expressions for the power spectral density of all relevant distortion products (cross-modulation, intermodulation and harmonic distortion) are derived for the first time. Calculations are presented to show the effect of these distortion effects on overall system performance. Expressions developed will help circuit and system designers to come to an optimum power consumption versus performance trade-off.

An RF receiver front-end for MB-OFDM based UWB systems is designed. The receiver is the first to be designed without using any on-chip inductors or off-chip matching components. The receiver occupies only 0.35mm\(^2\) in a 0.18\(\mu\)M CMOS process, consists of a low-noise amplifier, downconverter and a bandpass filter. The measured receiver gain is 21 dB, Noise Figure is less than 6 dB, input \(IIP_3\) is -5 dBm and the receiver consumes 19.5 mA from a 2.3V supply. The receiver covers all the MB-OFDM bands from 3.1 GHz to 8 GHz.
I

Introduction

I.A  The Need for Short-Range High Speed Wireless Networks

The proliferation of consumer electronics devices, both in the home and the office, is driving a need for robust and high speed wireless networks. There is a very strong case for short-range wireless networks, which operate only over a few feet and provide data rates in the hundreds of Mega Bits Per Second (Mbps).

It is fairly common for an average home to have a few digital cameras, a few portable music players and one or more portable digital assistants, in addition to cellular phones and external hard drives. As the number of devices in a home grows, so does the number of wires connecting these peripherals together. It is not unusual to find an average user frustrated by the sheer mess of cables surrounding his home or office desk, and the need for carrying these cable when travelling. Additionally, an average desk also contains a computer monitor, which requires one more cable to be connected to any of these devices. A typical home or office desk is shown in Fig. I.1

An average entertainment room in a home could consist of a high definition cable/satellite receiver, a digital video recorder and a DVD player. All of these need to be connected to a television and to an audio/video receiver via cables. In an era where companies are spending a significant amount of money to make their flat panel televisions look aesthetically appealing, it is a discouraging to see a user’s entertainment
Figure I.1: A typical home or office desk.

center, such as the one shown in Fig. I.2 cluttered with a mess of cables. It is very easy to see how a short-range wireless network could ease the life of a user by removing some or all of these wires. Consumer electronics have not seen an explosion of short-range wireless devices mainly due to the low speed the present wireless networks have to offer.

Devices such as digital cameras and portable music players frequently require data transfer of the order of a few gigabytes. Transfer of data between these portable devices and personal computers has so far been limited to the wired USB 2.0 connection which offers transfer rates of up to 400 Mbps (Mega bits per second). High resolution computer monitors have resolutions in the 5 mega pixel range and high definition cable TV and DVD players require transfer rates in excess of 20 Mbps.

I.B Existing Wireless Data Networks

Existing wireless data technologies can be divided into the following categories:

1. Wireless Wide Area Networks (WWAN): These are terrestrial and satellite tech-
nologies that cover regions several miles wide. These encompass CDMA based EVDO [2], direct satellite links and the recently proposed WiMax [3]. These can provide data rates up to a few Mbps. The drawback of these is that they are power hungry, since they are designed to cover a few miles in distance.

2. Wireless Local Area Networks (WLAN): WLAN systems have an operating range of a few hundred feet [4]. The primary WLAN solution in existence today is WiFi (802.11 a/b/g). WiFi has the potential for very high data rates. 802.11a and 802.11g offer raw data rates of up to 54 Mbps. However, these communication protocols were designed for relatively long range communication (over 100 feet) and consequently, are very power hungry. The technology serves no purpose if, for example, a user has to connect the power cord to his digital camera in order to use the WiFi network to transfer the photos to his or her computer.

3. Wireless Personal Area Networks (WPAN): WPAN networks are networks that cover only a few feet in range. Examples of such networks are Infra-red commu-
nications and Bluetooth [5]. Infra-red communication offers only a few kilo-bits per second of data rate. This technology is now limited mainly to line-of-sight remote controllers in the realm of consumer electronics devices.

Bluetooth is also meant for short range wireless networking of consumer electronic devices. However, the Bluetooth standard was designed for low speed communications, and the current standard (Bluetooth 2.0) offers only 3.0 Mbps of raw data rate. Adding MAC overhead and non line-of-sight operation, the average speed seen by a user could be far less. An average digital camera uses a 1 Giga-byte (GB) memory card. Assuming a data rate of 3.0 Mbps, this transforms to a data transfer time of about 45 minutes, which is too long for practical purposes.

I.C Ultra Wideband for Wireless Networking

Recently, the Federal Communications Commission (FCC) has opened up a 7.5 GHz wide spectrum for unlicensed ultra wide-band usage. This has created a lot of activity in both the industry and the academia. Various new proposals are being put forth to efficiently utilize this newly available spectrum.

The main requirements of a UWB system are:

1. A UWB system needs to occupy at least 500 MHz bandwidth.

2. A UWB system should not affect the performance of systems already existing in this frequency spectrum. The spectrum allocated is between 3.1 - 10.6 GHz, which overlaps with 802.11a and WiMax [3]. Its power spectral density (PSD) should be less than -41.25 dBm/MHz. The allowed spectral mask for UWB systems is shown in Fig. 1.3.

UWB technology has been around since the 1960’s, when it was known as Time-Domain Electromagnetics [6]. In most UWB applications developed so far, the system operates by sending short pulses in the time domain. This signalling scheme proved to be very effective in position location. Industry and academia are now pondering the most effective method to utilize this spectrum in the area of wireless communications. Since
the transmit power is very low, these systems inherently lend themselves to low power, short range applications.

Figure I.3: FCC designated spectral mask for Ultra Wideband systems. Allowed emission level is in dBm/MHz.

**I.D Single-Carrier and Multi-Carrier Transmission Schemes**

Various transmission schemes exist for creating efficient wireless networks, each suitable for a specific operating environment. For example, Code Division Multiple Axis (CDMA) [7] exists as an efficient network protocol for cellular phones, due to its superiority in efficient bandwidth utilization. Frequency Modulation (FM) is a low-cost low-power and simple solution for radio station broadcast. The choice of correct transmission scheme is probably one of the most important decisions in creating a pervasive and successful wireless technology. Transmission schemes can broadly be divided into two main categories: Single carrier and multi-carrier transmission.
For robustness, the chosen transmission scheme needs to have the following properties

1. Robustness to multipath fading: Multipath fading occurs due to the superposition, at the receiver antenna, of multiple signals generated from the same source, but shifted in time/phase and amplitude, due to reflection from multiple objects. There can be constructive or destructive superposition of these time shifted waveforms, depending upon the phase difference between the arriving signals at the antenna, as shown in Fig. I.4. This results in variations in amplitude of the received signal, over the bandwidth of the signal [8].

There are various models in existence which model the propagation of a wireless signal in the presence of these fades. One of the models which is suitable for short-range indoor environments is the Saleh-Valenzuela (S-V) model [9]. This model splits the received signal rays into groups of signals called clusters. The idea is that in a short-range indoor environment, there may be many signals generated from the transmitter, arriving at the receiver at different times, but there are only a few objects nearby which cause this reflection. Hence the fades can be roughly divided into clusters, where each cluster contains a group of time-shifted signals with similar statistics. The model distinguishes between two different arrival rates:

(a) Cluster arrival rate, which starts at time $t = 0$ and
(b) Ray arrival rate, which is the arrival rate of a ray relative to the start time of the cluster.

The impulse response of the multipath model is described as

$$h_i(t) = X_i \sum_{l=0}^{L} \sum_{k=0}^{K} \alpha_{k,l}^i \delta\left(t - T_{l,1}^i - \tau_{k,l}^i\right)$$  \hspace{1cm} (I.1)

where

- $\alpha_{k,l}^i$ are the multipath gain coefficients, $i$ refers to the impulse response realization,
- $l$ refers to the cluster, and $k$ refers to arrival within the cluster.
- $T_{l,1}^i$ is the delay of the $l$th cluster for the $i^{th}$ channel realisation
\( \tau_{k,l}^i \) is the delay of the \( k^{th} \) multipath component relative to the \( l^{th} \) cluster arrival time \( T_l^i \).

\( X_i \) represents the log-normal shadowing for the \( i^{th} \) channel realization.

\begin{figure}[h]
\centering
\includegraphics[width=\textwidth]{multipath.png}
\caption{(a) Constructive and (b) destructive forms of multipath phenomenon for sinusoidal signals [8].}
\end{figure}

In an indoor environment, there can be reflections from various sources. Since UWB systems will occupy 3.1 - 10.6 GHz, there will inevitably be multiple fades present in that frequency range. The UWB system has to maintain link quality in the presence of multiple fades.

In a single carrier transmission scheme, the digitally modulated baseband spectrum is upconverted using one carrier and transmitted. To counter the effects of multipath fading, a multifinger RAKE receiver is employed [8]. The number of RAKE fingers depends on the number of fades the receiver can experience over
its entire bandwidth. In an indoor UWB system, which operates between 3.1 - 10 GHz, the complexity of this RAKE receiver rises significantly [10].

In a multi-carrier transmission system, the baseband signal consists of multiple carriers, each modulated to reflect the modulation scheme (such as QPSK). These multiple carriers are then upconverted and transmitted. In such a system, a frequency selective fade can only remove a few subcarriers, but the overall link remains robust, as shown in Fig. I.5.

![Schematic of Single Carrier and Multi-Carrier Systems](image)

Figure I.5: Effect of fading on single carrier and multicarrier transmission systems.

2. Robustness to narrowband jammers: Various narrowband jammers are present in the 3.1 - 10.6 GHz spectrum, such as 802.11a, WiMax, marine radar. The UWB system needs to maintain its link quality in the presence of these narrowband jammers. Multi-carrier systems are less susceptible to narrowband jammers. Since UWB systems are open to reception of narrowband jammers, this is a very significant advantage of multi-carrier systems.
I.E  Orthogonal Frequency Division Multiplexing

In a traditional multi-carrier system, the spectrum is divided into $N$ non-overlapping subchannels. The baseband data is modulated onto each of the subcarriers. In traditional multi-carrier systems, some extra spacing is introduced between the subcarriers to reduce Inter Symbol Interference (ISI), as shown in Fig. I.6. However, this results in wastage of valuable spectrum and a drop in spectral efficiency of the system. OFDM [11] was proposed to increase the spectral efficiency of multi-carrier transmission systems. Data is transmitted over a collection of carriers called subcarriers. OFDM has the advantages of robustness against frequency selective fading, narrowband interference and inter-symbol interference.

![Figure I.6: Reduced spectral efficiency of traditional multi carrier systems.](image)

I.E.1  Generation of OFDM Signals

In an OFDM system, the baseband signal is composed of a sum of sinusoids, where each sinusoid represents a baseband symbol. These sinusoids are modulated to reflect the baseband modulation pattern, such as Quadrature Phase Shift Keying (QPSK). This baseband signal, which is a sum of sinusoids, is then upconverted to RF and transmitted in bursts. This means that each baseband symbol is multiplied by a rectangular
waveform in the time domain before transmission. This operation shapes the spectrum of the baseband signal, and the spectrum, which would have been a summation of Dirac Delta functions, is now a summation of \( \text{sinc} \) functions. Let \( F(x(t)) \) denote the operation of taking the Fourier transform of a function \( x(t) \). Then

\[
F(\cos(\omega_o t)) = \pi [\delta_d(\omega - \omega_o) + \delta_d(\omega + \omega_o)]
\]  

(I.2)

and

\[
F\left(\text{rect}\left[\frac{t}{T_s}\right] \cos(\omega_o t)\right) = \frac{T_s}{2} \left[ \text{sinc}\left(\frac{(\omega - \omega_o T_s)}{2}\right) + \text{sinc}\left(\frac{(\omega + \omega_o)T_s}{2}\right) \right]
\]

(I.3)

Where \( \delta_d(t) \) is the Dirac Delta function, defined as

\[
\delta_d(t) = 1 \quad \text{if} \quad t = 0
\]

\[
= 0 \quad \text{otherwise}
\]

(I.4)

and \( \text{rect}\left[\frac{t}{T_s}\right] \) is the rectangular function, defined as (Fig. I.7)

\[
\text{rect}\left[\frac{t}{T_s}\right] = 1 \quad \text{if} \quad -\frac{T_s}{2} < t < \frac{T_s}{2}
\]

\[
= 0 \quad \text{otherwise}
\]

(I.5)

If \( T_s \) contains an integral number of cycles of each of the sinusoids, then, at the peak of one of these \( \text{sinc} \) functions in the frequency domain, all the other \( \text{sinc} \) functions have a null. This is what gives rise to orthogonality in an OFDM systems. All subcarriers are orthogonal to each other.

Since each subcarrier is multiplied by a rectangular function in the time domain, the spectrum of each subcarrier looks like a \( \text{sinc} \) function, as shown in Fig. I.8. It is important to note that the nulls in the spectrum occur at integer multiples of \( \frac{1}{T_s} \). This means that if each subcarrier has an integral number of cycles in the symbol time \( T_s \), then at the peak of one particular subcarrier in the frequency domain, all the other subcarriers would have a null. In other words, if \( T_s \) is chosen to be equal to \( \frac{1}{\Delta f} \), the subcarrier spacing, then at the peak of one particular subcarrier, all the other subcarriers will have a null. This property is what gives rise to the orthogonality between carriers.
in OFDM. An implication of this orthogonality is that Inter-Symbol Interference (ISI) is significantly reduced.

An OFDM symbol can be represented as

\[ S(t) = \text{rect} \left[ \frac{t + \psi_t}{T_s} \right] \sum_{n=0}^{N-1} \text{Re} \{ \exp(j2\pi(f_c + n\Delta f)t) \} \quad (I.6) \]

Where \( T_s \) is the symbol time period
\( f_c \) is the center frequency
\( \Delta f \) is the subcarrier spacing
\( N \) = total number of subcarriers and

A spectrum of an OFDM symbol, comprising several subcarriers is shown in Fig. I.9

The generation of an OFDM signal is shown in Fig. I.10. To generate an OFDM symbol, the baseband data is first converted from serial to parallel. Each data stream then modulates an OFDM subcarrier. This is achieved by performing an Inverse Fast Fourier Transform (IFFT) of the incoming signal. As is clear from (I.6), OFDM modulation is just an inverse Fourier Transform. Once the complex baseband signal has been generated using an IFFT, it is upconverted to RF with an IQ upconverter.
Figure I.8: Spectrum of the $n^{th}$ OFDM subcarrier where $f_c$ is the center frequency, $\Delta f$ is the subcarrier spacing and $T_s$ is the symbol time.

Figure I.9: Spectrum of an OFDM symbol.

I.E.2 Disadvantages of OFDM

While OFDM has all the benefits of a multi-carrier transmission scheme, it does have some disadvantages.

1. Susceptibility to frequency shifts: Since OFDM relies on orthogonality of subcarriers, a frequency offset in the subcarriers can cause the link to degrade. There are two main processes which create a drift in frequency: phase noise from the frequency synthesizer and Doppler effect.

   (a) Doppler effect: A relative movement between and transmitter and receiver
can cause the frequency of the subcarriers to shift. The Doppler frequency shift of one subcarrier can be calculated by using

\[ f_D = \frac{v_r f_c}{c} \cos \alpha \]  

(1.7)

where

- \( f_D \) = frequency deviation due to Doppler effect
- \( v_r \) = relative speed of the transmitter and receiver
- \( f_c \) = center frequency of the subcarrier
- \( c \) = speed of light
- \( \alpha \) = angle of the velocity vector

It is clear from (1.7) that Doppler effect changes the frequency of all the subcarriers by the same percentage, \( \eta \). The fact that all the subcarriers change in frequency by the same percentage destroys the orthogonality of OFDM, since the subcarrier spacing, which was \( \frac{m T_s}{T_a} \) is now \( \frac{(1+\eta)m}{T_s} \), where \( m \) is an integer. While designing an OFDM system, careful attention needs to be paid to the environment the system will operate in. The relative speeds of the transmitter and receiver need to be taken into account while deciding on a subcarrier spacing. The subcarrier spacing needs to be large enough so that the Doppler shift plays an insignificant role in determining system performance.
(b) Phase noise in the local oscillator or frequency synthesizer can cause the center frequency of the subcarriers to drift. OFDM receivers, therefore, have stringent phase noise requirements, which can complicate the design of the Voltage Controlled Oscillator (VCO) and the accompanying Phase Locked Loop (PLL).

2. High peak to average ratio: OFDM signals have a high Peak to Average Power Ratio (PAPR) [12]. It is not uncommon for an OFDM system to exhibit 10 - 12 dB of PAPR. This high PAPR can severely degrade the power efficiency of the power amplifier in the transmitter, which is usually the highest power consuming circuit in the transceiver. The average transmitted power is low, but the power amplifier has to be able to handle the high peaks which occur infrequently. This implies that the transmitter has to operate at a significant back-off from the maximum transmit power, where the power efficiency is usually significantly lower.

I.F Multi-Band OFDM

Ultra Wideband (UWB) Orthogonal Frequency Division Multiplexing (OFDM) systems have been proposed as an emerging solution to wireless communication applications requiring high data rates (up to 400 Mbps) over short distances. In the Multi-Band Orthogonal Frequency Division Multiplexing (MB-OFDM) version [1], a 528 MHz wide OFDM signal is created from 128 subcarriers, with a subcarrier spacing of 4.125 MHz, as shown in Fig. I.11. Each of these subcarriers is modulated with Quadrature Phase Shift Keying (QPSK).

MB-OFDM is a frequency hopping system. To enable operation of multiple UWB systems at the same time, the carrier hops around in frequency. The carrier can hop to one of fourteen channels \((2904 + 528n \ MHz, \ n = 1, 2 \ldots 14)\), divided into four groups of three channels and one group of two channels, as shown in Fig. I.12. An MB-OFDM system can operate using one or more of the band groups. Operation in Band Group 1 is mandatory. An MB-OFDM system operating in Band Group-1 only is
known as a Mode-1 system.

The time between frequency hops (symbol interval) is 312.5 nS and there is a 9.5 nS guard interval for transmit/receive turnaround time. This representative time-frequency interleaving for a Group 1-only system is depicted in Fig. I.13.

Time-frequency interleaving in MB-OFDM has multiple benefits. Firstly, it presents high flexibility in frequency planning. Different countries may have different
Figure I.13: Time frequency interleaving for a Group-1 only MB-OFDM system [10].

regions of the spectrum available for UWB devices. MB-OFDM has the potential to address this issue, since individual frequency bands can be turned on and off. The carrier may avoid hopping to parts of the spectrum that are not allocated for UWB. This can easily be achieved in software, and therefore the same hardware can be used in multiple countries.

Since the carrier hops at a fast rate, the average power is lower than the instantaneously power. MB-OFDM systems can therefore transmit higher instantaneous power than true wideband systems. Consequently, the instantaneous SNR available is higher. Additionally, the design of some parts of the hardware becomes easier, since the instantaneous bandwidth is only 528 MHz.

However, MB-OFDM does present some challenges in hardware design. The main drawback of this approach is that hardware needs to settle within 9.5 nS, which is the guard interval, while hopping. This could present some significant challenges in synthesizer design, since the Phase Locked Loop (PLL) has to switch frequencies within 9.5 nS. The generation and reception of an MB-OFDM signal is shown in Fig. I.14 and Fig. I.15.
**I.G MB-OFDM Modes of Operation**

MB-OFDM systems are anticipated to support high data rates over short ranges. Multiple modes of operation are provided to accommodate different frequency planning and availability of hardware. Group-1 support is mandatory, Groups 2-4 are optional. Various data rates are available and system specifications differ depending on what data rates are supported. This section gives a brief of the system specifications in each mode.

The two lower data rates (110 Mbps and 200 Mbps) are deemed mandatory and the highest data rate of 480 Mbps is optional.
Table I.1: Targeted bit rates and range for MB-OFDM [1].

<table>
<thead>
<tr>
<th>Bit Rate</th>
<th>Distance</th>
</tr>
</thead>
<tbody>
<tr>
<td>110 Mbps</td>
<td>10 m</td>
</tr>
<tr>
<td>200 Mbps</td>
<td>4 m</td>
</tr>
<tr>
<td>480 Mbps</td>
<td>2 m</td>
</tr>
</tbody>
</table>

I.H MB-OFDM Timing

The target data rates and distance of operation are summarized in Table I.4. The main timing related parameters for an MB-OFDM system are shown in Table I.2.

Table I.2: MB-OFDM timing related parameters [1].

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$N_{SD}$: Number of data subcarriers</td>
<td>100</td>
</tr>
<tr>
<td>$N_{SDP}$: Number of defined pilot carriers</td>
<td>10</td>
</tr>
<tr>
<td>$N_{SG}$: Number of guard subcarriers</td>
<td>12</td>
</tr>
<tr>
<td>$N_{ST}$: Number of total subcarriers used</td>
<td>122 ($=N_{SD} + N_{SDP} + N_{SG}$)</td>
</tr>
<tr>
<td>$\Delta_f$: Subcarrier spacing</td>
<td>4.125 MHz ($=528\text{ MHz}/128$)</td>
</tr>
<tr>
<td>$T_{FFT}$: IFFT/FFT Period</td>
<td>242.42 nS ($=1/\Delta_f$)</td>
</tr>
<tr>
<td>$T_{ZP}$: Zero pad duration</td>
<td>70.08 nS ($=37/528\text{ MHz}$)</td>
</tr>
<tr>
<td>$T_{SYM}$: Symbol internal</td>
<td>312.5 nS</td>
</tr>
</tbody>
</table>

MB-OFDM is a frequency hop system. In one particular instance of the system, the channels are allowed to hop in one particular Band Group. For example, systems based on Group-1 can have the carrier hop between Bands 1, 2 and 3. Systems designated for Group-2 can have the carrier hop between Bands 4, 5 and 6. To support multiple piconets, different frequency hop patterns, or Time Frequency Codes (TFCs) are defined for each piconet. There are up to four TFCs available for Groups 1, 2, 3 and 4 and two TFCs for Group 5. The TFCs are defined in Table I.3. Table I.3 shows the TFCs for a Group-1 only system. For systems in Groups 2, 3 and 4, the relevant Band ID should be replace with numbers appropriate for that group. For example, Band IDs 1,2 and 3 for a Group-1 system should be replaced with Band ID 4,5 and 6, for Group 2
and so on.

Table I.3: Time frequency codes for a Group-1 MB-OFDM system [1].

<table>
<thead>
<tr>
<th>TFC</th>
<th>Hopping Sequence (Band ID)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>1 2 3 1 2 3</td>
</tr>
<tr>
<td>2</td>
<td>1 3 2 1 3 2</td>
</tr>
<tr>
<td>3</td>
<td>1 1 2 2 3 3</td>
</tr>
<tr>
<td>4</td>
<td>1 1 3 3 2 2</td>
</tr>
</tbody>
</table>

I.I MB-OFDM Data Rates

Available data rates and the corresponding PHY parameters are listed in Table I.4.

$R = \text{Coding rate}$

Table I.4: MB-OFDM data rates and corresponding PHY parameters, QPSK modulation is used for all data rates [1].

<table>
<thead>
<tr>
<th>Data Rate (Mbps)</th>
<th>R</th>
<th>Conj Symmetry</th>
<th>$T_{SF}$</th>
<th>$G_{SPR}$</th>
<th>CBS</th>
</tr>
</thead>
<tbody>
<tr>
<td>53.3</td>
<td>1/3</td>
<td>Yes</td>
<td>2</td>
<td>4</td>
<td>100</td>
</tr>
<tr>
<td>80</td>
<td>1/2</td>
<td>Yes</td>
<td>2</td>
<td>4</td>
<td>100</td>
</tr>
<tr>
<td>110</td>
<td>11/32</td>
<td>No</td>
<td>2</td>
<td>2</td>
<td>200</td>
</tr>
<tr>
<td>160</td>
<td>1/2</td>
<td>No</td>
<td>2</td>
<td>2</td>
<td>200</td>
</tr>
<tr>
<td>200</td>
<td>5/8</td>
<td>No</td>
<td>2</td>
<td>2</td>
<td>200</td>
</tr>
<tr>
<td>320</td>
<td>1/2</td>
<td>No</td>
<td>1</td>
<td>1</td>
<td>200</td>
</tr>
<tr>
<td>400</td>
<td>5/8</td>
<td>No</td>
<td>1</td>
<td>1</td>
<td>200</td>
</tr>
<tr>
<td>480</td>
<td>3/4</td>
<td>No</td>
<td>1</td>
<td>1</td>
<td>200</td>
</tr>
</tbody>
</table>

$T_{SF} = \text{Time spreading factor}$

$G_{SPR} = \text{Total spreading gain}$

$CBS = \text{Coded bits per OFDM symbol.}$
I.J MB-OFDM Link Budget Analysis

A basic link budget analysis for a Mode-1 MB-OFDM system is presented in Table I.5. The required sensitivity for a Mode-1 (Group 1 only) system is -80.5dBm for a data rate of 110 Mbps. For all MB-OFDM systems, Mode-1 operation is mandatory. Inclusion of higher groups is optional.

Table I.5: MB-OFDM link budget analysis for a Mode-1 system [10].

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Low rate</th>
<th>Medium rate</th>
<th>High rate</th>
</tr>
</thead>
<tbody>
<tr>
<td>Data Rate ($R_b$)</td>
<td>110 Mbps</td>
<td>200 Mbps</td>
<td>480 Mbps</td>
</tr>
<tr>
<td>AverageTx Power ($P_T$)</td>
<td>-9.9 dBm</td>
<td>-9.9 dBm</td>
<td>-9.9 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>$F_C$ (geometric center frequency)</td>
<td>3882 MHz</td>
<td>3882 MHz</td>
<td>3882 MHz</td>
</tr>
<tr>
<td>Path Loss at 1 meter ($L_1$)</td>
<td>44.2 dB</td>
<td>44.2 dB</td>
<td>44.2 dB</td>
</tr>
<tr>
<td>Path loss at d meters ($L_d$)</td>
<td>20 dB (10m)</td>
<td>12 dB (4m)</td>
<td>6 dB (2m)</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>-74.1 dBm</td>
<td>-66.1 dBm</td>
<td>-60.1 dBm</td>
</tr>
<tr>
<td>Average $E_b/N_0$ at RF ($N$)</td>
<td>-93.6 dB</td>
<td>-91.0 dBm</td>
<td>-87.2 dBm</td>
</tr>
<tr>
<td>Rx NF ($N_F$)</td>
<td>6.6 dB</td>
<td>6.6 dB</td>
<td>6.6 dB</td>
</tr>
<tr>
<td>Average $E_b/N_0$ at baseband</td>
<td>-87 dBm</td>
<td>-84.4 dBm</td>
<td>-80.6 dBm</td>
</tr>
<tr>
<td>Required $E_b/N_0$ ($S$)</td>
<td>3.6 dB</td>
<td>4.3 dB</td>
<td>4.6 dB</td>
</tr>
<tr>
<td>Implementation Loss ($I$)</td>
<td>2.9 dB</td>
<td>2.9 dB</td>
<td>3.4 dB</td>
</tr>
<tr>
<td>Link Margin ($P_R - P_N - S - I$)</td>
<td>6.4 dB</td>
<td>11.1 dB</td>
<td>12.5 dB</td>
</tr>
<tr>
<td>Mix. Rx Sensitivity</td>
<td>-80.5 dBm</td>
<td>-77.2 dBm</td>
<td>-72.6 dBm</td>
</tr>
</tbody>
</table>

I.K Dissertation Focus

This dissertation can be divided into two broad sections. The first part of this dissertation focuses on distortion analysis of UWB receivers. While distortion has been studied extensively for narrow-band systems, distortion analysis of UWB systems has not received much attention thus far.

Traditional distortion analysis of RF front-ends uses a single-tone approximation for all RF input signals. A true wideband analysis needs to be undertaken to accurately predict performance of UWB systems in various interference scenarios. In
Chapter II, accurate analytical expressions are developed for various distortion products in a UWB receiver front-end.

The second part of this dissertation focuses on the design of a low-power and low-cost UWB receiver front-end. Traditional receiver architectures are explored for their applicability for a low-power low-cost UWB system.

The specifications of the receiver are developed using distortion analysis presented in Chapter II. This guarantees satisfactory operation of the UWB system in dense wireless environments. A flow-chart of the progression of this dissertation is presented in Fig. I.16.

Chapter II presents a wideband analysis of the impact of receiver nonlinearities on UWB system performance. For the first time, a true wideband analysis is developed, utilizing the properties of a QPSK OFDM system. Analytical expressions relating the power spectral density to unwanted signal powers and receiver distortion coefficients are presented. The accuracy of these expressions is verified by comparing them with the results of a full system simulation.

Chapter III presents the design of the first UWB receiver front-end that does not use any on-chip or off-chip inductors. The receiver is designed in a 0.18 $\mu M$ CMOS technology, and consumes only 0.35 $mm^2$ of die area, while meeting all the specifications of an MB-OFDM UWB system.

I.I Conclusion

In this chapter, it was identified that there is a growing need for a robust high speed network for short range operation, mainly in the area of consumer electronics. Proliferation of storage devices such as portable music players, digital cameras, portable hard drives has created a significant market for such high speed wireless networks.

Existing wireless solutions like 802.11a/b/g and Bluetooth are unable to satisfy the demands of these networks. While 802.11a and 802.11g promise data rates of up to 54 Mbps, these data rates are not high enough to satisfy the needs of high speed
and high capacity storage devices such as digital cameras and portable music players. Bluetooth 2.0 data rate of 3.0 Mbps falls far short of what is required.

UWB MB-OFDM systems have recently been proposed to satisfy these demands. UWB systems have not been widely deployed so far, and their performance in a real world situation is yet unknown. The wide bandwidth required of these systems presents new challenges for circuit and system designers. This research tries to address both of these issues by analyzing the behavior of MB-OFDM systems in a dense interference environment. A new receiver design presented in this research meets the goal of a low-cost, low-power and interference robust solution.
Figure I.16: Progression of dissertation.
II

Distortion Analysis of UWB Receivers

II.A Introduction

This chapter presents a new wideband analysis of all the important distortion products that can occur in an MB-OFDM receiver, due to nonlinearity in the receiver front-end. True wideband analysis, utilizing the statistical properties of an MB-OFDM system, is performed. Analytical expression relating the power spectral densities of these distortion products, and receiver nonlinearities are developed.

The calculated expressions are compared with the results of a full system simulation to determine the accuracy of these expressions. A methodology is outlined to predict the performance of the system in interference scenarios, which can help system designers specify the receiver optimally.

Ultra wideband MB-OFDM receivers are anticipated to operate in a hostile interference environment. In a typical usage environment, there could be multiple wireless sources nearby, and the receiver can face multiple jamming scenarios from in-band and out-of-band jammers. The front-end needs to operate between 3.1 - 10.6 GHz. There are many interference sources present within this frequency band, such as WiMax, 802.11a and marine radar [14]. Some of the interferers present at the input of a typical MB-OFDM receiver are shown in Fig. II.2.

There are two kinds of jammers present at an MB-OFDM receiver input, wide-
band jammers and narrowband jammers. These can be either in-band or out-of-band. In-band jammers are jammers that lie within the MB-OFDM spectrum of 3.1 - 10.6 GHz. Out-of-band jammers are out of the MB-OFDM spectrum, and are either below 3.1 GHz in frequency, or higher than 10.6 GHz. Examples of narrow in-band jammers are 802.11a and marine radar. In-band jammers can also be other UWB transmitters. Examples of out-of-band jammers are WiFi (802.11b/g) and Bluetooth (802.15.1), which operate in the 2.4 - 2.5 GHz frequency range.

Various mechanisms exist in the receiver front-end, which can create spurious outputs. The chief mechanisms are cross-modulation [15], intermodulation and harmonic distortion. Traditional interference analysis approximates jammers with a single tone and uses two-tone analysis to predict distortion components [16]. This approach
works well for narrowband systems, however, it can be highly inaccurate when modeling a UWB system. In narrowband systems, second-order distortion is usually considered only for computation of DC offset [18]. However, in UWB systems, second-order harmonics and second-order intermodulation products can also land in-band and destroy the SNR of a desired channel.

The methodology used to calculate all the distortion products and their impact on system performance is shown in Fig. II.3.

**II.B Receiver Model**

To proceed with the analysis, we first need to develop a model for the receiver. We model the receiver front end as a memoryless nonlinear system with its response to an input $x(t)$ given by

$$y(t) = c_1 x(t) + c_2 x^2(t) + c_3 x^3(t) \tag{II.1}$$

The transfer function will also consist of higher order terms, but in most situations, the front-end amplifier is operated well below its 1-dB compression point, in which case a third-order approximation models the receiver front end quite accurately [16]. Fig. II.4 plots the input-output characteristics of an amplifier with hard saturation, such as a power amplifier, with a third-order curve fitted approximation of its characteristics. The figure shows that, even for a high input signal power of -20 dBm, the relative error is only 0.2 dB. For a receiver front-end, where the amplifier usually shows softer compression, this error would be even less.

A nonlinear system represented by (II.1) produces various spurious outputs. Typically, the nonlinearities of these systems are characterized by presenting a two-tone input and observing the output of the system. The distortion products at the output of the system represented by (II.1) with a two-tone input of frequency $\omega_1$ and $\omega_2$ are shown in Fig. II.5. The linearity of the system is characterized by specifying its Second-Order Input Intercept Point (IIP$_2$) and Third-Order Input Intercept Point (IIP$_3$).
RF Signal Model

Receiver Model

Compute time domain distortion product

Compute autocorrelation function of time domain product

Take Fourier Transform of autocorrelation function to compute PSD

Integrate PSD to obtain total in-band power

Figure II.3: Methodology for calculating spurious outputs.
The coefficient \( c_1 \) in (II.1) is the linear gain of the receiver. The coefficient \( c_2 \) represents the second-order distortion in the receiver. It can be related to the well known second-order intercept point (\( IIP_2 \)) of the receiver using

\[
c_2 = \frac{c_1}{IIP_2}
\]  
(II.2)

Similarly, \( c_3 \) represents third-order distortion in the receiver, and can be related to the third-order intercept point (\( IIP_3 \)) by

\[
IIP_3 = \sqrt[3]{\frac{4c_1}{3c_3}}
\]  
(II.3)

Although we have neglected all memory effects in the power series approximation, some memory effects can be incorporated if we use \( c_3 \) derived from a Volterra
Third-order Power Series

\[ A_1 A_2 A_1 A_2 - A_1^2 - A_2 A_1 A_2 + A_2^2 - A_1^2 - A_2^2 + A_1^2 + A_2^2 \]

Frequency

Figure II.5: Spurious output in a system with second and third-order nonlinearity, represented by (II.1). Input is a sum of two sinusoids at \( \omega_1 \) and \( \omega_2 \).

analysis of a nonlinear system. With a Volterra analysis, \( c_3 \) and receiver \( IIP_3 \) can be related by [15]

\[ c_3^2 = \frac{4G}{9IIP_3^2} \]  

where \( G \) is the power gain of the LNA.

II.C MB-OFDM Signal Model

The next step in the analysis is to develop a model to represent the RF signal present at the UWB receiver input. MB-OFDM signals are QPSK modulated and the symbols are carried over 128 carriers, occupying a total bandwidth of 528 MHz [1]. The transmitted RF MB-OFDM signal can be written as [1]

\[ r_{rf}(t) = Re\left(\sum_{k=0}^{N-1} r_k(t - kT_s) \exp(j2\pi f_{kmod}t)\right) \]  

where \( r_k(t) \) = complex baseband variable representing the \( k^{th} \) OFDM symbol

\( N \) = total number of transmitted OFDM symbols

\( T_s \) = symbol period

\( f_{kmod} \) is the RF center frequency and \( kmod6 \) represents the frequency hop algorithm that dynamically changes the center frequency of the MB-OFDM signal. The allowed hopping sequences are shown in Table I.3.

\( r_k(t) \) is defined as

\[ r_k(t) = \sum_{n=-N/2}^{N/2} C_n \exp(j2\pi n \Delta f t) \]  

II.5

II.6
where $N_s = \text{total number of OFDM subcarriers}$

$C_n = \text{complex baseband data (QPSK modulation for MB-OFDM)}.$

$\Delta f = \text{subcarrier spacing}.$

For a single frame (consisting of $N_s \text{ subcarriers/symbols}$), the OFDM signal can be reduced to

$$r_{rf}(t) = A_{tx} \text{rect} \left[ \frac{t + \psi_t}{T_s} \right] \sum_{n=-N_s/2}^{N_s/2} \left\{ a_n \cos((\omega_{tx} + n\Delta \omega)t + \theta) + b_n \sin((\omega_{tx} + n\Delta \omega)t + \theta) \right\}$$ (II.7)

where $A_{tx} = \text{rms amplitude of the transmitted signal}$

$\omega_{tx} = \text{angular frequency of the transmitted carrier}$

$\Delta \omega = \text{subcarrier spacing}$

$\theta = \text{random phase}$

$\psi_t = \text{random time delay for the symbol. The variable } \psi_t \text{ also serves to make } r_{rf}(t) \text{ a wide sense stationary process, since now the symbol start time is also a random variable.}$

It also has important consequences in the final analysis.

$\text{rect} \left[ \frac{t}{T_s} \right] \text{ is the rectangular function which transforms the OFDM signal from a sum of Dirac Delta functions in the frequency domain, to a sum of sinc functions. The function}$

$\text{rect} \left[ \frac{t}{T_s} \right] \text{ is defined as (Fig. II.6)}$

$$\text{rect} \left[ \frac{t}{T_s} \right] = 1 \quad \text{if} \quad -\frac{T_s}{2} < t < \frac{T_s}{2}$$

$$= 0 \quad \text{otherwise}$$ (II.8)

$a_n$ and $b_n$ are the real and imaginary parts of the baseband QPSK data $C_n$, as shown in Fig. II.7.

Since $a_n$ and $b_n$ are real and imaginary parts of independent baseband QPSK data, we state their following properties:

$$E\{a_kb_ia_n b_m\} = E\{a_ka_n\} E\{b_kb_m\}$$ (II.9)
The RF signal model of (II.7) and the properties (II.9) and (II.12) will be used in the following sections to analyze the performance of an MB-OFDM system under various interference situations.
Figure II.7: QPSK constellation diagram. \( a_n \) and \( b_n \) are the real and imaginary parts of the complex baseband data \( C_n \).

II.D MB-OFDM Received Power

In this section, we will relate the average transmitted MB-OFDM signal power with \( A_{tx} \) in II.7. To calculate the total power, we need to first calculate the autocorrelation function of the MB-OFDM signal. The autocorrelation function of a process \( x(t) \) is given by

\[
R_x(\tau) = E \{ x(t)x(t + \tau) \} \quad \text{(II.14)}
\]

We assume that the process \( r_{rf}(t) \) in (II.7) is a stationary process. This can partly be justified by the presence of the random variable \( \psi_t \) in (II.7). The variable \( \psi_t \) adds a random start time to the OFDM symbol. Under this assumption, the autocorrelation function can be written as

\[
R_x(\tau) = E \{ r_{rf}(0)r_{rf}(\tau) \} \quad \text{(II.15)}
\]
Now
\[ r_{rf}(0) = A_{tx} \text{rect} \left[ \frac{t}{T_s} \right] \sum_{n=-N_s}^{N_s} (a_n \cos(\theta) + b_n \sin(\theta)) \]  \hspace{1cm} (II.16)

\( R_{xx}(\tau) \) can then be written as

\[ R_{xx}(\tau) = E \left\{ A_{tx}^2 \text{rect} \left[ \frac{\tau + \psi_{\tau}}{T_s} \right] \text{rect} \left[ \frac{\psi_{\tau}}{T_s} \right] \sum_{n=-N_s}^{N_s} [a_n \cos((\omega_{tx} + n\Delta\omega)\tau + \theta) + b_n \sin((\omega_{tx} + n\Delta\omega)\tau + \theta)] \right\} \]  \hspace{1cm} (II.17)

It can be shown that
\[ E \left\{ A_{tx}^2 \text{rect} \left[ \frac{\tau + \psi_{\tau}}{T_s} \right] \text{rect} \left[ \frac{\psi_{\tau}}{T_s} \right] \right\} = A_{tx}^2 \Delta_{T_s}(\tau) \]  \hspace{1cm} (II.18)

where \( \Delta_{T_s}(\tau) \), shown in Fig. II.8 is defined as

\[ \Delta_{T_s}(\tau) = 1 - \frac{\mid \tau \mid}{T_s} \quad \text{if} \quad -T_s < \tau < T_s \]
\[ = 0 \quad \text{otherwise} \]  \hspace{1cm} (II.19)

We know, from (II.12), that all the terms involving the product \( a_n b_n \) will go to zero, when the expectation operation is taken inside the parenthesis, therefore we will ignore them. Using (II.12)

\[ R_{xx}(\tau) = A_{tx}^2 \Delta_{T_s}(\tau) E\left\{ \sum_{m=-N_s}^{N_s} \sum_{n=-N_s}^{N_s} [a_m a_n \cos((\omega_{tx} + n\Delta\omega)\tau + \theta) \cos(\theta) + b_m b_n \sin((\omega_{tx} + n\Delta\omega)\tau + \theta) \sin(\theta)] \right\} \]  \hspace{1cm} (II.20)

Using (II.11)
Figure II.8: Definition of $\Delta_{T_s}(\tau)$.

$$R_{xx}(\tau) =$$

$$A_{T_x}^2 \Delta_{T_x}(\tau) \sum_{m=-\frac{N}{2}}^{\frac{N}{2}} \sum_{n=-\frac{N}{2}}^{\frac{N}{2}} \left[ \delta_{mn} E_\theta \{ \cos((\omega_{T_x} + n\Delta\omega)\tau + \theta) \cos(\theta) \} ight. + \left. \delta_{mn} E_\theta \{ \sin((\omega_{T_x} + n\Delta\omega)\tau + \theta)) \sin(\theta) \} \right]$$  \hspace{1cm} (II.21)

Where the function $E_\theta \{ x(\theta) \}$ denotes the expectation of a function $x(\theta)$ over its variable $\theta$. We know that

$$\sum_{m=\frac{-N}{2}}^{\frac{N}{2}} \sum_{n=\frac{-N}{2}}^{\frac{N}{2}} \delta_{mn} f(m, n) = \sum_{m=\frac{-N}{2}}^{\frac{N}{2}} f(m, m)$$  \hspace{1cm} (II.22)

Therefore

$$R_{xx}(\tau) =$$

$$A_{T_x}^2 \Delta_{T_x}(\tau) \sum_{n=\frac{-N}{2}}^{\frac{N}{2}} E_\theta[\cos((\omega_{T_x} + n\Delta\omega)\tau + \theta) \cos(\theta)$$

$$+ \sin((\omega_{T_x} + n\Delta\omega)\tau + \theta)) \sin(\theta)]$$  \hspace{1cm} (II.23)
Also,

\[ E_\theta \{ \cos((\omega t_x + n\Delta \omega)t + \theta) \cos(\theta) \} = \]
\[ \frac{1}{2\pi} \int_0^{2\pi} \{ \cos((\omega t_x + n\Delta \omega)t + \theta) \cos(\theta) \} \, d\theta \]
\[ = \frac{1}{2} \cos(\omega t) \]  \hspace{1cm} (II.24)

and similarly

\[ E_\theta \{ \sin((\omega t_x + n\Delta \omega)t + \theta) \sin(\theta) \} = \]
\[ = \frac{1}{2} \cos(\omega t) \]  \hspace{1cm} (II.25)

Using II.23, II.24 and II.25,

\[ R_{xx}(\tau) = A^2_{tx} \Delta T_s(\tau) N_s \cos(\omega \tau) \]  \hspace{1cm} (II.26)

Equation II.26 gives the autocorrelation function of the input UWB signal. To compute its power spectral density, we use the Wiener-Khinchin theorem [17], which states that the power spectral density of a wide sense stationary process is given by the Fourier transform of its autocorrelation function.

\[ PSD(f) = \int_{-\infty}^{\infty} r_{xx}(\tau) e^{-j2\pi f \tau} \, d\tau \]  \hspace{1cm} (II.27)

Therefore, to compute the PSD of an MB-OFDM signal, we need to take the Fourier transform of II.26. Denoting the Fourier transform operation by \( F(,) \), we know that

\[ F \{ \Delta T_s(\tau) \cos(\omega_0 \tau) \} = T_s \left\{ \text{sinc}^2((\omega - \omega_0)T_s) + \text{sinc}^2((\omega + \omega_0)T_s) \right\} \]  \hspace{1cm} (II.28)

Then, using (II.28) in II.26, the power spectral density of an MB-OFDM signal is given by

\[ PSD_{in}(\omega) = A^2_{tx} N_s T_s \left\{ \text{sinc}^2((\omega - \omega_0)T_s) + \text{sinc}^2((\omega + \omega_0)T_s) \right\} \]  \hspace{1cm} (II.29)
To compute the total input power of a given MB-OFDM signal, represented by (II.7), we need to integrate its PSD, given by (II.29) over the bandwidth of the signal. We assume that the power contained in each sinc function in (II.29) is almost completely contained in the bandwidth of interest. Since it is only a few outer carriers whose power will lie outside the band of interest, this is a very accurate estimate. Integrating (II.29) from $-\infty$ to $\infty$, we get

$$P_{IN} = \frac{A_{tx}^2 N_s}{\pi}$$

Equation (II.30) represents the total input power of any OFDM signal, comprised of $N_s$ subcarriers with an rms amplitude of $A_{tx}$ per carrier.

II.E Received Signal Power at Antenna

For all the calculations in the following sections, we need to assume some input power level at the receiver antenna. We need to estimate the received powers of both an MB-OFDM UWB signal, and a narrowband jammer.

II.E.1 Wideband Jammer Power

The maximum output power of an MB-OFDM transmitter is defined to be $-10.3$ dBm [1]. In a real world application, a receiver may be as close as 0.1m to the transmitter. For example, if a user has a laptop connected to his digital camera with a Wireless USB link (using MB-OFDM), then the camera could be right next to the laptop transmitter. To compute the MB-OFDM received power from a transmitter 0.1m away, we need to subtract the free space path loss of 0.1m, given by [10]

$$P_L = 20 \log_{10} \left[ \frac{4\pi f g d}{c} \right]$$

(II.31)

Where

$P_L = \text{Free space path loss}$

$d = \text{distance in meters}$
\( c \) = the speed of light

\( f_g \) = geometric center frequency, defined as

\[
    f_g = \sqrt{f_u f_l}
\]  

(II.32)

Where \( f_u \) and \( f_l \) are the upper and lower frequencies of the band of interest. Assuming a 0.1m distance, \( f_l = 3.0 \) GHz, \( f_u = 3.5 \) GHz, the maximum received power at the antenna \( (P_{\text{MAX, UWB}}) \) is approximately -35 dBm. In all the distortion analysis scenarios, -35 dBm will be used as a representative maximum power level of an MB-OFDM UWB signal at the receiver antenna input.

The minimum received power level \( (P_{\text{MAX, UWB}}) \) at an MB-OFDM antenna has been defined as -80.5 dBm [1].

\textbf{II.E.2 Narrowband Jammer}

The narrowband jammer with the highest transmit power is 802.11a. Three different output power levels are specified for 802.11a.

\begin{table}[h]
\centering
\begin{tabular}{|c|c|c|}
\hline
Band & Center Frequency & Maximum Transmit Power \\
\hline
Lower & 5.2 GHz & 16 dBm \\
\hline
Middle & 5.3 GHz & 23 dBm \\
\hline
Upper & 5.775 GHz & 29 dBm \\
\hline
\end{tabular}
\end{table}

Although the highest frequency band transmits the maximum output power, it is only the lowest frequency band which is designated for indoor usage. The upper 802.11a transmit bands are designated for outdoor applications only. We will assume that a user will primarily use a UWB system for short-range indoor applications only. We will assume that the closest 802.11a transmitter is 0.3m away from the MB-OFDM receiver. For such a case, the maximum received power from an 802.11a transmitter \( (P_{\text{MAX, NB}}) \) 0.3m away is approximately -30 dBm.

Table II.2 shows all the power levels at the receiver antenna input to consider for interference analysis.
Table II.2: Received signal power levels at receiver antenna for interference analysis.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Maximum UWB Power ($P_{\text{MAX,UWB}}$)</td>
<td>-35 dBm</td>
</tr>
<tr>
<td>Minimum UWB Power ($P_{\text{MIN,UWB}}$)</td>
<td>-80.3 dBm</td>
</tr>
<tr>
<td>Maximum narrowband jammer power ($P_{\text{MAX,NB}}$)</td>
<td>-30 dBm</td>
</tr>
</tbody>
</table>

II.F Receiver Specifications

We also need to assume a gain and Noise Figure for the receiver front-end. Following the analysis in [10], for calculations in the rest of the analysis, the receiver front-end specifications assumed are summarized in Table II.3.

Table II.3: Assumed receiver specifications.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Gain</td>
<td>20 dB</td>
</tr>
<tr>
<td>Noise Figure</td>
<td>4.9 dB</td>
</tr>
</tbody>
</table>

II.G Out-of-band Jammers

Out-of-band jammers are defined as the jammers which lie outside of the MB-OFDM band, which spans 3.1 - 10.6 GHz. All interferers present at the receiver antenna input which are either lower than 3.1 GHz or higher than 10.6 GHz are considered out-of-band jammers. Examples of out-of-band jammers are:

1. 802.11b/g, Bluetooth and microwave ovens, all of which operate in the 2.4 - 2.5 GHz frequency range.
2. Mobile telephony jammers, such as CDMA which operate in the 800 - 900 MHz and 1800 - 1900 frequency range.

Out-of-band jammers can impact the performance of an MB-OFDM system in two ways. If the power level of an out-of-band jammer is significantly high, it can cause
the receiver to compress. Additionally, they can increase the average baseband noise per bit, causing an increase in the Bit Error Rate (BER). These out-of-band jammers can be rejected at the front-end and/or at the baseband output. It is advantageous to provide as much rejection at the front-end as possible, since rejecting the jammers at the front-end reduces the linearity requirement of the entire receiver chain.

Out-of-band rejection can be achieved with a front-end high-pass filter (HPF) or bandpass filter (BPF). While both can provide adequate rejection of lower frequency jammers, bandpass filters provide the additional advantage of rejection of jammers higher in frequency than the UWB spectrum (beyond 10.6 GHz).

For analysis in this chapter, it will be assumed that significant rejection of out-of-band jammers can be achieved via front-end and baseband filtering and that out-of-band jammers will no longer be an issue. An argument to this end can be made that if the receiver is designed to tolerate (without any gain compression) an in-band 802.11a jammer, which transmits equal amount of power as an 802.11b system, then it will automatically be able to tolerate an out-of-band 802.11b jammer, which has been suppressed significantly with a front-end pre-select filter, without any gain compression.

II.H In-Band Jammers

In-band jammers are defined as jammers which lie inside of the 3.1 GHz - 10.6 GHz MB-OFDM band. Examples of some in-band jammers are (Fig. II.2):

1. 802.11a systems which operate in the 5.1-5.3 GHz and 5.7 - 5.8 GHz frequency range.

2. WiMax which may be present in the 3.1 - 4 GHz frequency range,

3. Marine radar

4. Other UWB systems transmitting at a high power

A three-fold distortion analysis is needed to model the system. Some typical interference situations that need to be analyzed are:
1. A narrow-band jammer and the desired signal at the receiver input

2. A narrow-band jammer, a wideband jammer and the desired signal at the receiver input

3. One or more wide-band jammers at the receiver input, along with the desired signal

II.H.1 Narrow-Band Jammer Case

The highest power narrow in-band jammer to consider is 802.11a, which occupies the spectrum from 5.2 - 5.4 GHz and 5.7 - 5.9 GHz. As mentioned before, OFDM systems are fairly robust to narrow-band interferers, since an in-band narrow jammer can at most interfere with a few subcarriers, but may not destroy the whole link. For example, an 802.11a signal is 20 MHz wide and the subcarrier spacing in MB-OFDM is 4.125 MHz. This means that a high power 802.11a jammer can interfere with roughly five subcarriers out of a total of 128. The performance impact of removing interference with these subcarriers is not significant.

A bigger problem with in-band narrow jammers is receiver compression. If the received power is too high, it can cause the gain of several blocks in the receive chain to compress, causing distortion and a severe degradation in the Signal-to-Noise ratio (SNR). Since in-band jammers cannot be rejected by any kind of filtering, the receiver gain compression specification has to take into account the required separation of the receive antenna from an 802.11a transmitter.

From Table II.2, the largest in-band interferer that the receiver has to tolerate is an 802.11a signal at -30 dBm. This means that the receiver input 1-dB compression point has to be higher than -30 dBm.
II.H.2 Narrow-Band and Wideband Jammers at Input: Cross-Modulation

Although OFDM systems are immune, to a certain degree, to narrow-band interferers, cross-modulation of a narrow-band interferer with a wideband interferer can reduce the SNR of an OFDM system.

For a robust system solution, the design of an MB-OFDM receiver needs to accommodate these interferers. However, since these systems are intended for mainly consumer electronics devices, the robustness of the system should come at little or no added cost or power. This calls for a detailed analysis of the performance of an MB-OFDM system in the presence of all these interferers.

Consider the following interference situation, which can limit the performance of an MB-OFDM system: the receiver is tuned to an MB-OFDM transmitter transmitting in Band 1 of Group 1 (3.4 GHz center frequency)- this is the desired signal. There are two other undesired transmitters in the vicinity: (1) a narrowband jammer (for example, a relatively narrowband 3.0 GHz WiMAX) and (2) another MB-OFDM transmitter transmitting in one of the other bands in Group 1 (at 3.9 GHz or 4.4 GHz). This situation is depicted in Fig. II.9. The receiver then accepts, in addition to the desired signal it is tuned to, another interfering input given by

\[ s(t) = A_j \cos(\omega_j t) + \]
\[
A_{tx} \text{rect} \left[ \frac{t + \psi_t}{T_s} \right] \sum_{n=0}^{N_s-1} \{ a_n \cos((\omega_{tx} + \Delta \omega n)t + \theta) \\
+ b_n \sin((\omega_{tx} + \Delta \omega n)t + \theta) \} 
\] (II.33)

where \( \omega_j \) = frequency of the narrowband jammer
\( \omega_{tx} \) = center frequency of the second unwanted MB-OFDM transmitter
\( A_j \) = amplitude of the narrowband jammer
\( A_{tx} \) = \text{rms} amplitude of the second unwanted MB-OFDM transmitter and
\( \theta \) = random phase.

The first term in (II.33) represents the narrowband jammer and the second term represents the unwanted MB-OFDM transmitter. We have assumed that the narrowband jammer can be approximated by a single tone, as in equation (II.33). In reality, the jammer will be modulated. To keep the analysis simple and generic, we employ the single tone approximation for the analysis.

Replacing \( x(t) \) in (II.1) with (II.33) will result in a number of distortion products at the receiver output, which could lie in the UWB spectrum, as shown in Fig. II.10.

Figure II.10: Distortion products at receiver output, when the receiver is modeled with (II.1). Input is the sum of a narrowband jammer at 3.0 GHz and a UWB signal at 4.4 GHz.
Substituting (II.33) into (II.1), we need to expand

\[ s(t) = c_3 \left[A_j \cos(\omega_j t) + A_{tx} \text{rect} \left[\frac{t + \psi_t}{T_s} \right] \sum_{-N_s}^{N_s} \left\{ a_n \cos((\omega_{tx} + \Delta \omega n)t + \theta) \right\} + b_n \sin((\omega_{tx} + \Delta \omega n)t + \theta) \right]\] ^3 \]  

(II.34)

Writing (II.34) as

\[ c_3 [a + b + c]^3 = a^3 + b^3 + c^3 + 3(ab^2 + ac^2 + a^2b + a^2c + bc^2 + b^2c) + 6abc \]  

(II.35)

where

\[ a = c_3 A_j \cos(\omega_j t) \]
\[ b = c_3 A_{tx} \text{rect} \left[\frac{t + \psi_t}{T_s} \right] \sum_{-N_s}^{N_s} \left\{ a_n \cos((\omega_{tx} + \Delta \omega n)t + \theta) \right\} \]
\[ c = c_3 A_{tx} \text{rect} \left[\frac{t + \psi_t}{T_s} \right] \sum_{-N_s}^{N_s} \left\{ b_n \sin((\omega_{tx} + \Delta \omega n)t + \theta) \right\} \]

The crossmodulation terms (around \( \omega_j \)) will arise from

- \( 6abc \)
- \( 3ab^2 \)
- \( 3ac^2 \)

Expanding, the three terms determining cross-modulation distortion are

\[ 6abc = 6c_3 A_j A_{tx}^2 \cos(\omega_j t) \text{rect} \left[\frac{t + \psi_t}{T_s} \right] \sum_{-N_s}^{N_s} \left\{ a_n \cos((\omega_{tx} + \Delta \omega n)t + \theta) \right\} \]
\[ \sum_{-\frac{N_s}{2}}^{\frac{N_s}{2}} \left\{ a_n \cos((\omega_{tx} + \Delta \omega n)t + \theta) \right\} \]
\[ \sum_{-\frac{N_s}{2}}^{\frac{N_s}{2}} \left\{ b_n \sin((\omega_{tx} + \Delta \omega n)t + \theta) \right\} \]  

(II.36)
In (II.37), we neglect the first two terms that represent the component at $2\omega_j$. The second two terms cancel each other out. To show this, the last two terms in (II.37) can be written as

$$
6abc = \frac{3}{2} c_3 A_j A_{tx}^2 \left[ \frac{t + \psi_t}{T_s} \right] \\
\sum_{m=-\frac{N_s}{2}}^{\frac{N_s}{2}} \sum_{n=-\frac{N_s}{2}}^{\frac{N_s}{2}} a_n b_m \left\{ \sin \left( (2\omega_c + \Delta \omega (m + n) + \omega_j) t + 2\theta \right) \\
+ \sin \left( (2\omega_c + \Delta \omega (m + n) - \omega_j) t + 2\theta \right) + \\
\sin \left( (\Delta \omega (m - n) - \omega_j) t + \sin \left( (\Delta \omega (m - n) + \omega_j) t \right) \right\} 
$$

(II.37)

Now, since $m = n$, when (II.38) is expanded, for every $\Delta \omega (m - n)$, there exists a term with the opposite sign at $\Delta \omega (n - m)$. Equation (II.38) therefore goes to zero.

Similarly, computing the rest of the terms $3a^2b$ and $3ab^2$, it can be shown that the time-domain cross-modulation signal is given by

$$
x_{mod}(t) = \\
\frac{c_3}{4} A_j A_{tx}^2 \left[ \frac{t + \psi_t}{T_s} \right] \sum_{m=-\frac{N_s}{2}}^{\frac{N_s}{2}} \sum_{n=-\frac{N_s}{2}}^{\frac{N_s}{2}} (a_m a_n + b_m b_n) \\
\left\{ \cos \left( (\omega_j + \Delta \omega (m - n)) t + \theta \right) \\
+ \cos \left( (\omega_j - \Delta \omega (m - n) t + \theta \right) \right\} 
$$

(II.39)

Since $a_n$ and $b_m$ are random variables, to compute the PSD of the cross-modulation, we need to compute the Fourier transform of the autocorrelation function of $x_{mod}(t)$.

Again, using (II.14)
\[ x_{\text{mod}}(0) = \frac{c_3^2}{2} A_j A_{tx}^2 \text{rect} \left[ \frac{\psi_t}{T_s} \right] \sum_{m=-N_s}^{N_s} \sum_{n=-N_s}^{N_s} (a_m a_n + b_m b_n) \cos(\theta) \] (II.40)

We know that all the terms involving a product such as \( a_m b_n \) will go to zero when the expectation is performed, therefore, to simplify the analysis, we neglect those terms. Then

\[ x_{\text{mod}}(0)x_{\text{mod}}(\tau) = \frac{c_3^2}{8} (A_j A_{tx}^2)^2 \text{rect} \left[ \frac{\tau + \psi_t}{T_s} \right] \text{rect} \left[ \frac{\psi_t}{T_s} \right] \sum_{m=-N_s}^{N_s} \sum_{n=-N_s}^{N_s} \sum_{p=-N_s}^{N_s} \sum_{q=-N_s}^{N_s} (a_m a_n a_p a_q + b_m b_n b_p b_q) \cos(\theta) \cos((\omega_j + \Delta \omega (m - n))\tau + \theta) \]

\[ + \cos((\omega_j - \Delta \omega (m - n))\tau + \theta)) \] (II.41)

Assuming stationarity, and using properties (II.9) - (II.12), the autocorrelation function of \( x_{\text{mod}}(t) \) is given by:

\[ R_{xx}(\tau) = \frac{c_3^2}{2} (A_j A_{tx}^2)^2 T_s \sum_{m=-N_s}^{N_s} \sum_{n=-N_s}^{N_s} \{ \cos((\omega_j + \Delta \omega (m - n))\tau) \} \] (II.42)

The power spectral density of the cross-modulation component can be obtained by taking the Fourier transform of (II.42). Taking the Fourier transform of (II.42), the PSD is given by

\[ PSD_{XMOD}(\omega) = \frac{c_3^2}{2} (A_j A_{tx}^2)^2 T_s \sum_{m=-N_s}^{N_s} \sum_{n=-N_s}^{N_s} \left[ \text{sinc}^2((\omega + (\omega_j + \Delta \omega (m - n))T_s) \right. \]

\[ + \text{sinc}^2((\omega - (\omega_j + \Delta \omega (m - n))T_s) \] (II.43)
It is clear from (II.43) that the spectrum of the jammer is spread to twice the bandwidth of the wideband signal, through cross-modulation distortion. If the jammer is close enough to the desired signal, this cross-modulation spectrum can land on top of the desired signal, reducing the SNR, with the extent of the reduction in SNR depending on the $IIP_3$ and the narrowband and wideband jammer powers.

To verify the validity of (II.43), a full system simulation of an MB-OFDM system was performed in MATLAB. The cross-modulation spectrum obtained was then compared with the prediction of (II.43). The simulation was run for an MB-OFDM receiver, as described in [10], with an $IIP_3$ of -10 dBm, input narrowband jammer power of $-10$ dBm and undesired wideband jammer power of $-10$ dBm. Simulated and calculated spectra are plotted in Fig. II.11 and match very closely, verifying the validity of the prediction.

II.H.3 Effect of Single-Tone Approximation

Since in a real world application, the narrow-band jammer in (II.33) would be modulated, it is of interest to compare the cross-modulation spectrum when the jammer is a single-tone and when it is modulated. Simulations were performed to compare the calculated cross-modulation spectrum with a single-tone narrowband jammer, an FM modulated jammer with 5 MHz bandwidth, and an FM modulated jammer with 10 MHz bandwidth. The results of these simulations are plotted in Fig. II.12. It can be seen that the modulated jammers spread the cross-modulation spectrum by approximately their modulation bandwidth. Therefore, for relatively narrow-band jammers, the cross-modulation spectrum obtained agrees well with the spectrum calculated using a single-tone approximation of (II.33).

II.H.4 Effect of Cross-Modulation on Link Margin

Cross-modulation distortion degrade the performance of the receiver by reducing the SNR at the receiver output. This results in a degraded system link margin. To evaluate the effect of this spread jammer on the total signal-to-noise ratio at the output
Figure II.11: Simulated and calculated cross-modulation products. $IIP_3 = -10$ dBm, Narrowband jammer power = -10 dBm, UWB jammer power= -10 dBm.
Figure II.12: Comparison of cross-modulation PSD with single-tone narrowband jammer and FM modulated narrowband jammer. $IIP_3 = -10$ dBm, Narrowband jammer power = -10 dBm, UWB jammer power = -10 dBm.
of the LNA, we need to integrate (II.43) within the bandwidth of the desired received signal.

For example, consider the case when the receiver is tuned to a MB-OFDM transmitter transmitting in Band 1 of Group 1 (3.15 - 3.65 GHz). There is also present another MB-OFDM transmitter in the vicinity transmitting in Band 3 of Group 1 (4.15-4.65 GHz) and a narrowband transmitter transmitting at 3.0 GHz. Since the cross-modulation product is twice the bandwidth of the MB-OFDM signal, it will occupy 2.5 - 3.5 GHz. This means that the cross-modulation product overlaps with the desired channel and we need to integrate (II.43) from 3.15 - 3.5 GHz to calculate the total distortion power contributed by this cross-modulation product.

To compare the total noise power, (II.43) was numerically integrated with input parameters of $IIP_3 = -10$ dBm, $P_j = -10$ dBm and $P_{TX} = -10$ dBm. MATLAB simulations show the complete noise power to be 2.35 dBm and numerical integration of (II.43) revealed excellent agreement, with a total noise power of 2.31 dBm.

Using the cross-modulation power calculated above, the effect of receiver $IIP_3$ on the total link budget can be calculated by adding this cross-modulation power to the total output noise power of the LNA. Using the link budget analysis in [10] for a 110 Mbps MB-OFDM system, the effect of cross-modulation on total receiver link budget is shown in Fig. II.13 - II.15. The figures show the total receiver link margin for the situation described above, as a function of $IIP_3$ in the presence of a narrowband jammer of -20, -30 and -40 dBm, when the wideband jammers are at -30, -35 and -40 dBm. It can be seen that, in the absence of cross-modulation distortion, the total link budget is 6 dB. But this number degrades rapidly in the presence of a strong narrowband jammer. Specifying the LNA $IIP_3$ without due consideration to cross-modulation can lead to inadequate system performance.

**II.H.5 Example Calculation**

The goal of this research is to present a framework for system designers to easily specify the linearity requirements of an MB-OFDM based UWB receiver. This
Figure II.13: Receiver link margin vs. $IIP_3$ in the presence of cross-modulation, single tone jammer power = -20 dBm. Link margin in the absence of cross-modulation = 6 dB. 

can be achieved either with full system simulations, or through analytical methods. The RF simulation of MB-OFDM systems is very computationally intensive, due to the large fractional bandwidth of the system, and also due to the statistical nature of the signal, which makes it necessary to simulate multiple random symbols to obtain an average value for the distortion products. The expressions derived so far ease the analysis to a simple numerical integration.

To further reduce computation time, estimates of the above distortion products can also be made. We demonstrate a method of arriving at this estimate by computing the noise power for the cross-modulation scenario of Fig. II.9. We assume, without
Figure II.14: Receiver link margin vs. $I_{IP3}$ in the presence of cross-modulation, (From [13]) single tone jammer power = -30 dBm. Link margin in the absence of cross-modulation = 6 dB.

significant loss of accuracy, that the total noise power in each $sinc$ function in the summation in (II.43) lies inside the bandwidth of integration. Also, the integral of each term inside the summation reduces to $1/2\pi T_s$.

Now, for the frequency of interest (3.4 GHz), we need to consider the cross-modulation products between 3.15 - 3.5 GHz. In the summation in (II.43), the term $sinc^2((\omega + (\omega_j + \Delta \omega(m - n))T_s)$ occurs $N_s - (m - n)$ times, and there are two $sinc$ terms in the summation.

Therefore, for the given frequencies and assuming $\Delta f = 4$ MHz (MB-
Figure II.15: Receiver link margin vs. $IIP_3$ in the presence of cross-modulation, single tone jammer power = -40 dBm. Link margin in the absence of cross-modulation = 6 dB.

For $N_s = 128$ (MB-OFDM), (II.6), this results in $n = 3652$. We therefore need to substitute $2 \times 3652 \times \frac{1}{2\pi f_s}$ for the summation in (II.43) to compute the total noise power.
Using the above mentioned values for $IIP_3$, $P_j$ and $P_{TX}$, this gives a total noise power of 0.4 dBm, which is very close to the previous result of 2.31 dBm, based on numerical integration.

This method of cross-modulation estimation reduces the complicated problem of wideband RF simulation to a simple summation, reducing the simulation time from hours to almost instantaneous. Even without the above approximation, the cross-modulation formula (II.43) reduces the problem drastically from a full system simulation and numerical integration of output power. For multiple scenarios (varying TX or jammer powers, receiver $IIP_3$ or $IIP_2$) one would need to run a simulation for each situation to determine its effect on receiver SNR. However, using (II.43), after the initial numerical integration of the terms inside the summation of the PSD equation, cross-modulation power can be immediately calculated for different scenarios.

## II.I Intermodulation Distortion

In this section, we consider the effects of Intermodulation (IM) distortion in MB-OFDM systems. IM distortion can arise from both second and third-order nonlinearities, and we will consider the two separately.

### II.I.1 Second-Order Intermodulation Distortion

Consider the following situation: the receiver is tuned to an MB-OFDM channel at 7128 MHz (Band 8). There are two other nearby transmitters transmitting at 3432 MHz (Band 1) and 3960 MHz (Band 2). Second-order nonlinearity in the receiver will result in these two unwanted received signals intermodulating and creating spurs at their sum and difference frequencies, with the bandwidth of the spur spread to twice that of a single MB-OFDM signal. The sum frequency would be 7392 MHz which will reduce the SNR for a receiver tuned to either Band 8 or Band 9 of an MB-OFDM system. This in-band spur will reduce the SNR and ultimately degrade system performance. Such a situation is depicted in Fig. II.16.
Figure II.16: Second-order intermodulation scenario.

Unlike narrowband systems, where second-order distortion at the front-end is important mainly due to the occurrence of a DC offset, second-order distortion in wide-band systems such as MB-OFDM can directly result in degradation of system performance. An accurate prediction of this distortion product is required to correctly specify second-order distortion performance of the RF front-end.

To derive an expression for this second-order IM PSD, we begin with the power series approximation for the LNA given by (II.1). The signal at the input of the receiver is a sum of two interfering MB-OFDM signals and can be written as

\[
 r_{\text{rf}}(t) = A_{\text{tx}1} \text{rect} \left( \frac{t + \psi_{\text{t}1}}{T_s} \right) \sum_{n=-N_s/2}^{N_s/2} \left\{ a_n \cos((\omega_{\text{tx}1} + n\Delta \omega)t) + b_n \sin((\omega_{\text{tx}1} + n\Delta \omega)t) \right\} + \\
 A_{\text{tx}2} \text{rect} \left( \frac{t + \psi_{\text{t}2}}{T_s} \right) \sum_{n=-N_s/2}^{N_s/2} \left\{ c_n \cos((\omega_{\text{tx}2} + n\Delta \omega)t + \theta) \\
 + d_n \sin((\omega_{\text{tx}2} + n\Delta \omega)t + \theta) \right\} \tag{II.46}
\]

Where \( A_{\text{tx}1} \) and \( A_{\text{tx}2} \) are the rms amplitudes of the two MB-OFDM signals and \( \omega_{\text{tx}1} \) and \( \omega_{\text{tx}2} \) are their angular frequencies.

A signal defined by (II.46) will produce unwanted spurs in the presence of a second-order nonlinearity. Therefore, in analyzing the effect of second-order nonlinear-
ity with an input given by (II.46), we need to compute the PSD of components at

1. $\omega_{tx1} + \omega_{tx2}$
2. $\omega_{tx1} - \omega_{tx2}$
3. $2\omega_{tx1}$
4. $2\omega_{tx2}$

We compute each of these individually.

**Intermodulation product at $\omega_{tx1} + \omega_{tx2}$**

The intermodulation term at $\omega_{tx1} + \omega_{tx2}$ arises from second-order distortion.

We rewrite (II.46) as

$$r_{rf}(t) = a + b + c + d \quad (II.47)$$

where

$$a = A_{tx1} rect\left[\frac{t + \psi_{t1}}{T_s}\right] \sum_{n=-\frac{N_s}{2}}^{\frac{N_s}{2}} \{a_n \cos((\omega_{tx1} + n\Delta\omega)t)\}$$

$$b = A_{tx1} rect\left[\frac{t + \psi_{t1}}{T_s}\right] \sum_{n=-\frac{N_s}{2}}^{\frac{N_s}{2}} \{b_n \sin((\omega_{tx1} + n\Delta\omega)t)\}$$

$$c = A_{tx2} rect\left[\frac{t + \psi_{t2}}{T_s}\right] \sum_{n=-\frac{N_s}{2}}^{\frac{N_s}{2}} \{c_n \cos((\omega_{tx2} + n\Delta\omega)t + \theta)\}$$

$$d = A_{tx2} rect\left[\frac{t + \psi_{t2}}{T_s}\right] \sum_{n=-\frac{N_s}{2}}^{\frac{N_s}{2}} \{d_n \sin((\omega_{tx2} + n\Delta\omega)t + \theta)\}$$

(II.48)

Squaring (II.47), the distortion component at $\omega_{tx1} + \omega_{tx2}$ arises from $2(ac + ad + bc + bd)$.

Then

$$2ac = 2c_2 A_{tx1} A_{tx2} rect\left[\frac{t + \psi_{t1}}{T_s}\right] \sum_{m=-\frac{N_s}{2}}^{\frac{N_s}{2}} \sum_{n=-\frac{N_s}{2}}^{\frac{N_s}{2}} \{a_m c_n \cos((\omega_{tx1} + \Delta m\omega)t + \theta) \cos((\omega_{tx1} + \Delta n\omega)t + \theta)\} \quad (II.49)$$
\[ 2ac = c_2 A_{tx1} A_{tx2} \text{rect} \left[ \frac{t + \psi_{t1}}{T_s} \right] \text{rect} \left[ \frac{t + \psi_{t2}}{T_s} \right] \sum_{m=\frac{N_s}{2}}^{\frac{N_s}{2}} \sum_{n=\frac{N_s}{2}}^{\frac{N_s}{2}} a_m c_n \cos((\omega_{tx1} + \omega_{tx2} + \Delta \omega(m + n)t + \theta)) + \cos((\omega_{tx1} - \omega_{tx2} + \Delta \omega(m - n)t)) \] (II.50)

Similarly

\[ 2ad = c_2 A_{tx1} A_{tx2} \text{rect} \left[ \frac{t + \psi_{t1}}{T_s} \right] \text{rect} \left[ \frac{t + \psi_{t2}}{T_s} \right] \sum_{m=\frac{N_s}{2}}^{\frac{N_s}{2}} \sum_{n=\frac{N_s}{2}}^{\frac{N_s}{2}} a_m d_n \sin((\omega_{tx1} + \omega_{tx2} + \Delta \omega(m + n)t + \theta)) + \sin((\omega_{tx1} - \omega_{tx2} + \Delta \omega(m - n)t)) \] (II.51)

\[ 2bc = c_2 A_{tx1} A_{tx2} \text{rect} \left[ \frac{t + \psi_{t1}}{T_s} \right] \text{rect} \left[ \frac{t + \psi_{t2}}{T_s} \right] \sum_{m=\frac{N_s}{2}}^{\frac{N_s}{2}} \sum_{n=\frac{N_s}{2}}^{\frac{N_s}{2}} b_m c_n \sin((\omega_{tx1} + \omega_{tx2} + \Delta \omega(m + n)t + \theta)) + \sin((\omega_{tx1} - \omega_{tx2} + \Delta \omega(m - n)t)) \] (II.52)

and

\[ 2bd = -c_2 A_{tx1} A_{tx2} \text{rect} \left[ \frac{t + \psi_{t1}}{T_s} \right] \text{rect} \left[ \frac{t + \psi_{t2}}{T_s} \right] \sum_{m=\frac{N_s}{2}}^{\frac{N_s}{2}} \sum_{n=\frac{N_s}{2}}^{\frac{N_s}{2}} b_m d_n \cos((\omega_{tx1} + \omega_{tx2} + \Delta \omega(m + n)t + \theta)) + \cos((\omega_{tx1} - \omega_{tx2} + \Delta \omega(m - n)t)) \] (II.53)

Neglecting all products apart from components at \( \omega_{tx1} + \omega_{tx2} \), the time domain signal at \( \omega_{tx1} + \omega_{tx2} \) is given by

\[ S_{IM2}(t) = c_2 A_{tx1} A_{tx2} \text{rect} \left[ \frac{t + \psi_{t1}}{T_s} \right] \text{rect} \left[ \frac{t + \psi_{t2}}{T_s} \right] \sum_{m=\frac{N_s}{2}}^{\frac{N_s}{2}} \sum_{n=\frac{N_s}{2}}^{\frac{N_s}{2}} \]
\begin{equation}
\{a_mc_n \cos((\omega_{tx1} + \omega_{tx2} + \Delta\omega (m+n)t + \theta)) \\
+ a_md_n \sin((\omega_{tx1} + \omega_{tx2} + \Delta\omega (m+n)t + \theta)) \\
+ b_mc_n \sin((\omega_{tx1} + \omega_{tx2} + \Delta\omega (m+n)t + \theta)) \\
- b_md_n \cos((\omega_{tx1} + \omega_{tx2} + \Delta\omega (m+n)t + \theta))\} \quad (II.54)
\end{equation}

Since $a_n$, $b_n$, $c_m$ and $d_m$ are random variables, to compute the PSD of this intermodulation product, we need to compute the Fourier transform of the autocorrelation function of $S_{IM2}(t)$ [17], which is given by

\begin{equation}
R_{ss}(\tau) = E\{S_{IM2}(t)S_{IM2}(t + \tau)\} \quad (II.55)
\end{equation}

Without loss of generality, we set the variable $t$ to zero, corresponding to the assumption of $a$, $b$, $c$ and $d$ being stationary random processes. We also state that $a$, $b$, $c$ and $d$ are independent processes, from (II.9) - (II.12), we have following properties:

\begin{equation}
E\{a_kb_lc_md_m\} = E\{a_k\}E\{b_l\}E\{c_n\}E\{d_m\} \quad (II.56)
\end{equation}

\begin{equation}
E\{a_ka_n\} = E\{b_kb_n\} = E\{c_kc_n\} = E\{d_kd_n\} = \delta_{kn} = \begin{cases} 1 & \text{if } k = n \\ 0 & \text{otherwise} \end{cases} \quad (II.57)
\end{equation}

Now

\begin{equation}
S_{IM2}(0) = A_{tx1}A_{tx2}\text{rect}\left[\frac{\psi_{t1}}{T_s}\right]\text{rect}\left[\frac{\psi_{t2}}{T_s}\right]\sum_{m=\frac{-N_s}{2}}^{\frac{N_s}{2}} \sum_{n=\frac{-N_s}{2}}^{\frac{N_s}{2}} \{a_mc_n \cos \theta + a_md_n \sin \theta + b_mc_n \sin \theta - b_md_n \cos \theta\} \quad (II.58)
\end{equation}

Then, using (II.54), (II.55), (II.56), (II.58), and neglecting the terms which have a coefficient that is the product of two different random variable (such as $a_nb_m$), since its expectation will be zero, we get the following expression for the autocorrelation function of (II.54),

\begin{equation}
R_{xx,im2}(\tau) = c_2^2 A_{tx1}^2 A_{tx2}^2 \Delta T_s(\tau) \sum_{m=\frac{-N_s}{2}}^{\frac{N_s}{2}} \sum_{n=\frac{-N_s}{2}}^{\frac{N_s}{2}} \cos((\omega_{tx1} + \omega_{tx2} + \Delta\omega (m+n))\tau) \quad (II.59)
\end{equation}
Following analysis similar to [13], the PSD of the intermodulation product can be written as

\[ PSD_{IM2}(\omega) = c^2 A_{tx1}^2 A_{tx2}^2 T_s \sum_{m=-N_s/2}^{N_s/2} \sum_{n=-N_s/2}^{N_s/2} \{ \text{sinc}^2(\omega + (\omega_{tx1} + \omega_{tx2} + \Delta \omega(m+n)) T_s) \\
+ \text{sinc}^2(\omega - (\omega_{tx1} + \omega_{tx2} + \Delta \omega(m+n)) T_s) \} \]  

(II.60)

Figure II.17: Simulated and calculated second-order intermodulation products. $IIP_2 = -10$ dBm, UWB jammer powers = -20 dBm at 3.5 GHz and 8 GHz.

To verify the validity of (II.60), a full system simulation of an MB-OFDM system was performed in MATLAB. The intermodulation spectrum obtained was then compared with results from (II.60). The simulation was run for an MB-OFDM receiver,
as described in [10], with an $IIP_2$ of -10 dBm and undesired Tx power of -20 dBm for each of the unwanted MB-OFDM transmitters. Simulated and calculated spectra are plotted in Fig. II.17 and match very closely, verifying the validity of the prediction.

Using the intermodulation power calculated above, the effect of receiver $IIP_2$ on the total link budget can be calculated by adding this second-order intermodulation power to the total in-band output noise power of the LNA. For the situation depicted in Fig. II.16, this can be computed by integrating (II.60) from 6864 MHz to 7392 MHz.

Following the link budget analysis in [10] for a 110 Mbps MB-OFDM system,
the effect of second-order intermodulation on total receiver link budget for the situation depicted in Fig. II.16 is shown in Fig. II.18. Fig. II.18 shows the total receiver link margin as a function of \( IIP_2 \) in the presence of unwanted UWB transmitters transmitting at -30, -35 and -40 dBm. It can be seen that, in the absence of intermodulation distortion, the total link budget is 6 dB. But this number degrades rapidly in the presence of strong unwanted MB-OFDM transmitters. Specifying the LNA \( IIP_2 \) without due consideration to intermodulation can lead to inadequate system performance.

**Intermodulation product at \( \omega_{tx1} - \omega_{tx2} \)**

The PSD of the intermodulation product at \( \omega_{tx1} - \omega_{tx2} \) will be the same as the PSD at \( \omega_{tx1} + \omega_{tx2} \). Therefore the PSD of the intermodulation product at \( \omega_{tx1} - \omega_{tx2} \) is also given by (II.60). It’s effect on the total link margin will therefore be the same as that of \( \omega_{tx1} + \omega_{tx2} \).

**Intermodulation product at \( 2\omega_{tx1} \)**

Again, substituting (II.46) for \( x(t) \) in to (III.14), the distortion product at \( 2\omega_{tx1} \) is given by:

\[
S_{IM2}(t) = \frac{c_2}{2}A_{tx1}^2 rect\left(\frac{t + \psi_{tx1}}{T_s}\right) \sum_{m=-\frac{N_s}{2}}^{\frac{N_s}{2}} \sum_{n=-\frac{N_s}{2}}^{\frac{N_s}{2}} \left\{ a_n a_m \cos((2\omega_{tx1} + \Delta \omega(m + n)t + \theta)) - b_n b_m \cos((2\omega_{tx1} + \Delta \omega(m + n)t + \theta)) + 2a_m b_n \sin((2\omega_{tx1} + \Delta \omega(m + n)t + \theta)) \right\} \tag{II.61}
\]

Again, assuming stationarity, and using the properties (II.9) - (II.12), computing the Fourier transform of the autocorrelation function of the (II.61), the PSD of the product at \( 2\omega_{tx1} \) is given by:
To compute the link margin degradation due to this distortion, the distortion power can be integrated over one MB-OFDM band and added to the output noise power of the LNA. Comparing (II.60) and (II.62), it is observed that, if the amplitudes of the two incoming UWB signals are the same, the magnitude of the distortion generated at \( \omega_{tx1} + \omega_{tx2} \) is four times the magnitude of the distortion product at \( 2\omega_{tx1} \). Therefore, if the link margin is maintained at the channel lying at \( \omega_{tx1} + \omega_{tx2} \) at \( \omega_{tx1} \). Therefore, there is no need to further investigate this distortion component.

II.I.2 Third-Order Intermodulation Distortion

In this section, we analyze the effect of third-order intermodulation distortion on system performance. The third-order intermodulation components exist at \( 2\omega_{tx1} - \omega_{tx2}, 2\omega_{tx1} + \omega_{tx2}, 2\omega_{tx2} - \omega_{tx1} \) and \( 2\omega_{tx2} + \omega_{tx1} \). By symmetry, all of these terms will have the exact same spectrum and magnitude.

Third-order nonlinearities in the receiver can also result in unwanted signals falling in-band and reducing SNR. For example, transmitters at 3432 MHz (Band 1) and 3960 MHz (Band 2) can, in the presence of third-order nonlinearities, produce an intermodulation product at 4488 MHz, which is in-band for a Group 1 MB-OFDM system. This situation is depicted in Fig. II.19.

To compute the third-order distortion terms, we need to cube (II.46) and multiply the resulting components by \( c_3 \), which is the third-order distortion coefficient for the nonlinearity represented by the power series (II.1). Representing (II.46) by (II.47), after cubing (II.47), the distortion component at \( 2\omega_{tx1} - \omega_{tx2} \) will arise from

\[
PSD_{2\omega_{tx1}}(\omega) = \frac{c_3^2}{4} A_{tx1}^4 T_s \sum_{m=-\frac{N}{2}}^{\frac{N}{2}} \left\{ \text{sinc}^2(\omega + (2\omega_{tx1} + \Delta\omega(2m))T_s) + \text{sinc}^2(\omega - (2\omega_{tx1} + \Delta\omega(2m))T_s) \right\}
\]
Figure II.19: Third-order intermodulation scenario.

\[
3(ac^2 + ad^2bc^2 + bd^2) + 6(acd + bcd) \tag{II.63}
\]

Where \(a, b, c \) and \(d\) are defined by (II.48). Then, using

\[
\left( \text{rect} \left[ \frac{t + \psi_{t1}}{T_s} \right] \right)^2 = \text{rect} \left[ \frac{t + \psi_{t1}}{T_s} \right] \tag{II.64}
\]

the intermodulation product at \(2\omega_{tx1} - \omega_{tx2}\) is given by

\[
S_{IM3}(t) = \frac{c^3}{4} A_{tx1}^2 A_{tx2}
\]

\[
\text{rect} \left[ \frac{t + \psi_{t1}}{T_s} \right] \text{rect} \left[ \frac{t + \psi_{t2}}{T_s} \right] \sum_{k=-N_s/2}^{N_s/2} \sum_{m=-N_s/2}^{N_s/2} \sum_{n=-N_s/2}^{N_s/2} \{3a_k b_m b_n
\]

\[
\left[ \cos(2\omega_{tx1} - \omega_{tx2} + \Delta\omega(m - n + k))t + \theta \right]
\]

\[
+ \cos((2\omega_{tx1} - \omega_{tx2} + \Delta\omega(m - n - k))t + \theta)
\]

\[
- \cos((2\omega_{tx1} - \omega_{tx2} + \Delta\omega(m + n - k))t + \theta)
\]

\[
+ 3a_m a_n b_k
\]
$$\begin{align*}
&[−\sin((2ω_{tx1} − ω_{tx2} + Δω(m + n − k))t + θ) \\
&+ \sin((2ω_{tx1} − ω_{tx2} + Δω(m − n + k))t + θ) \\
&+ \sin((2ω_{tx1} − ω_{tx2} + Δω(−m + n + k))t + θ)] \\
&+ a_k a_m a_n \{\cos((2ω_{tx1} − ω_{tx2} + Δω(m + n − k))t + θ) \\
&+ \cos((2ω_{tx1} − ω_{tx2} + Δω(m − n + k))t + θ) \\
&+ \cos((2ω_{tx1} − ω_{tx2} + Δω(−m + n + k))t + θ)] \\
&+ b_k b_m b_n \{\sin((2ω_{tx1} − ω_{tx2} + Δω(m + n − k))t + θ) \\
&+ \sin((2ω_{tx1} − ω_{tx2} + Δω(m − n + k))t + θ) \\
&+ \sin((2ω_{tx1} − ω_{tx2} + Δω(−m + n + k))t + θ)]\} \quad (II.65)
\end{align*}$$

The autocorrelation function of this distortion product is given by

$$R_{xx}(τ) =$$

$$\frac{9}{4} c_3^2 A^4_{tx1} A^2_{tx2} rect\left[\frac{t + \psi_{11}}{T_s}\right] rect\left[\frac{t + \psi_{12}}{T_s}\right] \sum_{k=-\frac{N_s}{2}}^{\frac{N_s}{2}} \sum_{m=-\frac{N_s}{2}}^{\frac{N_s}{2}} \sum_{n=-\frac{N_s}{2}}^{\frac{N_s}{2}} \{\cos(((2ω_{tx1} − ω_{tx2} + Δω(m + n + k))τ) \} \quad (II.66)$$

Computing the Fourier Transform of (II.66), the power spectral density of this third-order intermodulation product, computed as the autocorrelation function of (II.65) is given by

$$PSD_{IM3}(ω) =$$

$$\frac{9}{4} c_3^2 A^4_{tx1} A^2_{tx2} T_s \sum_{k=-\frac{N_s}{2}}^{\frac{N_s}{2}} \sum_{m=-\frac{N_s}{2}}^{\frac{N_s}{2}} \sum_{n=-\frac{N_s}{2}}^{\frac{N_s}{2}} \{sinc^2((ω + (2ω_{tx1} − ω_{tx2} + Δω(m + n + k))T_s) +
\quad sinc^2((ω − (2ω_{tx1} − ω_{tx2} + Δω(m + n + k))T_s) \} \quad (II.67)$$

To evaluate the accuracy of (III.5), a full system simulation of an MB-OFDM receiver with an $IIP_3$ of -10 dBm was run with equal unwanted TX powers of -30 dBm. A comparison of simulated and calculated PSD of this third-order intermodulation product is presented in Fig. II.20 and the results match closely.
Using the intermodulation power calculated in (III.5), the effect of receiver $IIP_3$, due to third-order intermodulation of two nearby unwanted UWB TX jammers, on the total link budget can be calculated by adding this third-order intermodulation power to the total output noise power of the LNA. For the situation depicted in Fig. II.19, this intermodulation noise power can be computed by integrating (III.5) from 4224 MHz to 4752 MHz.

Following the link budget analysis in [10] for a 110 Mbps MB-OFDM system, the effect of third-order intermodulation on total receiver link budget for the situation depicted in Fig. II.19 is shown in Fig. II.21. Fig. II.21 shows the receiver link margin.
for a receiver tuned to Band 3 of Group 1 as a function of $IIP_3$ in the presence of two unwanted UWB transmitters transmitting at -30, -35 and -40 dBm in Bands 1 and 2 of Group 1 of an MB-OFDM system.

II.J Harmonic Distortion

Unlike narrowband systems, harmonic distortion can also be a significant contributor to a degradation in the output SNR of the system. While with narrowband systems, harmonics of in-band signals are far away from the band of interest, in wide band systems such as MB-OFDM, the harmonics of the signal can also land in band. For
example, if the receiver is tuned to an MB-OFDM transmitter at 6.8 GHz, and there is present a nearby transmitter, transmitting at 3.4 GHz, second-order distortion in the receiver front-end will result in the second harmonic of the unwanted MB-OFDM transmitter falling in-band at 6.8 GHz. As shown in Fig. II.22, this will reduce the SNR at the output of the receiver front-end, and this effect needs to be quantized to accurately predict system performance in such situations.

Representing the input signal by (II.7) and substituting in (II.1), the second harmonic is given by

\[
S_{HD2}(t) = \frac{c_2}{2} A_{tx}^2 \text{rect} \left[ t + \frac{\psi_t}{T_s} \right] \sum_{m=-\frac{N_s}{2}}^{\frac{N_s}{2}} \sum_{n=-\frac{N_s}{2}}^{\frac{N_s}{2}} \left\{ a_m a_n \cos((2\omega_{tx} + \Delta\omega(m + n))t + \theta) + b_m b_n \cos((2\omega_{tx} + \Delta\omega(m + n))t + \theta) + 2a_m b_n \cos((2\omega_{tx} + \Delta\omega(m + n))t + \theta) \right\}
\]

(II.68)

Following analysis similar to [13], the power spectral density of this second harmonic, computed as the Fourier transform of the autocorrelation function of (II.68) is given by
$$PSD_{HD2}(\omega) =$$

$$\frac{c^2}{4} A_{tx}^4 T_s \sum_{\frac{N_s}{2}}^{\frac{N_s}{2}} \left\{ \text{sinc}^2((\omega + (2\omega_{tx} + \Delta\omega(2m))T_s) +$$

$$\text{sinc}^2((\omega - (2\omega_{tx} + \Delta\omega(2m))T_s) \right\} +$$

$$\frac{c^2}{2} A_{tx}^4 T_s \sum_{m=-\frac{N_s}{2}}^{\frac{N_s}{2}} \sum_{n=-\frac{N_s}{2}}^{\frac{N_s}{2}} \left\{ \text{sinc}^2((\omega + (2\omega_{tx} + \Delta\omega(m + n))T_s) +$$

$$\text{sinc}^2((\omega - (2\omega_{tx} + \Delta\omega(m + n))T_s) \right\}$$

(II.69)

Fig. II.23 compares the simulated and calculated PSD of the second harmonic product for a UWB receiver with $IIP_2$ of -10 dBm and an unwanted UWB jammer at -10 dBm.

Again, to compute the degradation in link margin of a UWB receiver tuned to 6.8 GHz, due to the second harmonic of an unwanted UWB jammer at 3.4 GHz, we add the in-band noise power, computed by integrating (II.69) from 6.55 GHz to 7.05 GHz to the output noise of the LNA. Fig. II.24 shows the degradation in link margin due to second harmonic distortion in an MB-OFDM system.

II.K Inaccuracies Resulting from Use of Two-Tone Intermodulation Test in UWB Systems

A very simple and attractive way of estimating harmonic and intermodulation components in narrowband systems is by predicting them through a traditional two-tone analysis. For example, the third-order intermodulation product of a narrow-band system can be approximated by using the well-known formula.

$$IIP_3 = P_{in} + \frac{\Delta P}{2}$$

(II.70)
Figure II.23: Simulated and calculated second-order harmonic. $IIP_2 = -10$ dBm, UWB jammer = -10 dBm at 4 GHz.

Where $P_{in}$ is the power of the single tones at the input and $\Delta_p$ is the observed difference between the single tone output and the IM3 product, as shown in Fig. II.25. For example, for a receiver with an $IIP_3$ of -10 dBm, gain of 15 dB and input single tone powers of -30 dBm, this would yield an output IM3 power of -55 dBm. However, spread-spectrum analysis and simulation of this receiver show the output intermodulation power in a single MB-OFDM receive band to be -50 dBm. This example demonstrates the high levels of inaccuracy that can be introduced in UWB system analysis if single tone approximations of these receivers are utilized to predict system performance. The more accurate analysis presented here presents a realistic estimate of the nonlinear
distortion that can be expected in these systems.

II.L Conclusion

UWB receivers are anticipated to be deployed in dense wireless environments. The successful and robust operation of a UWB receiver in such an environment requires accurate prediction of distortion products produced at the output in the presence of interfering spurs at the input, and nonlinearities in the receiver.

Various real world situations were analyzed where these nonlinearities and interferers could reduce the SNR of the system, degrading performance. A thorough
Figure II.25: Estimation of intermodulation products using single-tone approximation.

wideband analysis, using the statistical properties of QPSK baseband signals was carried out, and analytical expressions were developed relating the receiver distortion coefficients and amplitudes of interferers at the input.

The developed expressions were compared with the results of a full system simulation to determine their accuracy. The analytical results were found to match with simulations very well, confirming their validity.

These expressions were used to predict the degradation of received SNR in various real world scenarios. It was found that in a typical wireless environment, SNR degradation is dominated by intermodulation distortion in the presence of second-order distortion, and by cross-modulation distortion in the presence of third-order distortion.

This chapter has been published in part in the following publications:


III

UWB Receiver Design

III.A Introduction

Most communication protocols in existence today are narrow-band. For example, Code Division Multiple Axis (CDMA) has a bandwidth of only 1.2 MHz. Wireless broadband access protocols, such as 802.11 a/b/g occupy 20 MHz of frequency spectrum. Consequently, the majority of RF circuit design techniques in operation today are unsuitable for UWB transceiver design. For example, the most popular technique for designing a Low Noise Amplifier (LNA) is to inductively degenerate the LNA, use a series inductor for input matching, and provide a tuned LC resonant tank as the output load and matching. While this technique is highly effective for the design of narrow-band LNAs, it is very unsuitable for UWB LNAs. Inductive degeneration would make the gain of the amplifier significantly frequency dependent. Input matching with a series inductor also makes the LNA highly frequency selective. A new design approach is needed to counter the challenges presented by UWB systems.

Apart from individual circuits, most of the architectures of receivers in existence today are geared towards narrow-band operation. A standard heterodyne receiver may use a quadrature Local Oscillator (LO) for image rejection, which could be hard to generate in UWB systems, given the large fractional tuning range of the frequency synthesizer. Standard direct conversion receivers have a time-varying DC offset, which
can be very detrimental to the performance of a UWB system. These standard receiver architectures need to be examined very carefully to verify their viability in a UWB system.

To meet the objective of a widely deployed high-speed wireless network, an MB-OFDM system needs to be cheap, low power and reliable. Meeting all these objectives at the same time is not an easy task. For example, a designer could increase linearity in the receiver to make the design robust. But this increases the power consumption of the system. To reduce the power consumption in the front-end, a designer may add extra degeneration inductance in the low noise amplifier. But this would increase die area and, therefore, the cost of the system. A very careful trade-off needs to be executed for an optimum design.

In the design presented here, we try to meet the following objectives:

1. Low Cost: No on-chip inductors, minimum die area, minimum off-chip components
2. Low Power: Optimum receiver design
3. Interference robust: Estimate interference scenarios before-hand and design the receiver to withstand them

**III.B UWB Receiver Cost Analysis**

In this section, we will discuss the costs associated with designing a UWB system. Some of the cost components of an average receiver chain are listed in Table III.B

While the external bandpass filter is an essential component in the receiver chain for filtering out-of-band jammers, there is a possibility of eliminating most of the other components. This has the potential to save almost $1.25 from the manufacturing cost of a UWB system. Considering these systems are meant as peripherals for consumer electronics devices, this is a very significant saving.
Table III.1: Approximate cost of components in a UWB receiver chain

<table>
<thead>
<tr>
<th>Component</th>
<th>Quantity</th>
<th>Cost ($)</th>
</tr>
</thead>
<tbody>
<tr>
<td>External BPF</td>
<td>1</td>
<td>0.2</td>
</tr>
<tr>
<td>Input matching inductors</td>
<td>2</td>
<td>0.2</td>
</tr>
<tr>
<td>Input matching capacitors</td>
<td>2</td>
<td>0.05</td>
</tr>
<tr>
<td>Off-chip interstage SAW filter</td>
<td>1</td>
<td>0.5</td>
</tr>
<tr>
<td>On-chip inductors</td>
<td>5</td>
<td>0.5</td>
</tr>
</tbody>
</table>

This receiver will be targeted for the lower three Band Groups spanning 3.1 - 8 GHz. Although current UWB systems only operate as Mode-1 devices, operating in Band group 1 only, this design will cover the two higher band groups designated for future use. The goal of the design will be to cover all the three band groups, while adding little or no cost and power, when compared to the receiver design for a Mode-1 system only.

### III.C Direct Conversion Receiver

When it comes to designing a low-power and low-cost receiver, the traditional choice is a direct conversion architecture. A direct conversion receiver, along with some issues typically seen in a direct conversion receiver, is shown in Fig. III.1.

A direct conversion receiver provides the following advantages:

1. Simplicity of LO scheme: A direct conversion receiver requires only one local oscillator, which eases the design of the frequency synthesizer. Having only one frequency synthesizer also reduces the power consumption. It also provides the added advantage of the ability to use the same synthesizer for both the transmitter and receiver. A frequency synthesizer in a heterodyne receiver can usually share the same module only if the transmitter is also a heterodyne transmitter.

2. Reduced power consumption and die area: A direct conversion receiver uses fewer components than a heterodyne receiver. For example, a direct conversion receiver requires only one downconverter, while a heterodyne receiver requires two down-
converters. A heterodyne receiver, thus, is more efficient in both power consumption and die area.

However, a direct conversion receiver does have some disadvantages, which need to be carefully considered before making a design decision. The main issues seen with a direct conversion receiver are:

1. DC Offset: DC offset in direct-conversion receiver can

   (a) Corrupt the spectrum of the output signal. For example, in an MB-OFDM direct conversion receiver, DC offset at the output of the mixer will destroy the SNR of the center carriers.

   (b) Compress the baseband Analog-to-Digital (A/D) converter. In a typical direct-conversion receiver, the mixer may consume approximately 10 mA of current and have a resistive load of approximately 100 Ω. This implies
a DC voltage drop of 1V. Even a 0.1% mismatch in the load resistors or the bias current can produce a 1mV DC offset at the mixer output. A typical MB-OFDM direct-conversion receiver lineup is shown in Fig. III.2. As shown in the figure, a 1mV DC offset at the mixer output can results in a 1V DC offset at the input to the baseband A/D converter. Most A/D converters operate in the 0.5 - 1V range and when this 1V DC offset, combined with the RF signal which may also be 1V, appears at the input to the A/D converters, the converters can easily compress.

![Diagram](image)

Figure III.2: Approximate magnitude of DC offset in a direct-conversion receiver.

There are two primary processes in a receiver that can produce a DC offset at either the LNA output or the mixer output.

(a) Imbalance in differential paths: In a differential circuit, if there is any mismatch in the differential pairs, it will create a DC offset at the output of the circuit. Large device sizes are usually employed to keep this mismatch at a minimum.
(b) LO self mixing: In most downconverting mixers, there is a local oscillator driving a pair of differential switches. This LO can leak back to the RF input port of the mixer, and then mix with itself to produce an undesired DC output.

For example, in a Gilbert Cell based mixer, the LO can couple to the RF input port through the $C_{gs}$ (gate-source capacitance) and $C_{gd}$ (gate-drain capacitance). This LO would then mix with itself and produce an undesired output at DC.

If the LO and RF are far apart in frequencies, this DC offset can be eliminated by simply placing a DC blocking capacitor at the mixer output. However, in a direct conversion receiver, the LO and the incoming RF signal are at the same frequency. Placing a DC blocking capacitor at the output can significantly degrade the desired signal also.

DC offset is one of the biggest issues that haunts direct conversion receivers. Almost all direct conversion receivers employ some kind of a DC calibration loop. DC offset is measured at the baseband output and is fed back to the mixer input or output to cancel it. These DC cancellation loops are usually very slow to respond. The slowness of these calibration loops works fine for narrow-band systems. The DC offset changes over time due to

(a) Change in temperature

(b) Change in RF and LO frequencies

A change in DC offset with a change in RF and LO frequencies is not an issue for narrowband systems, where the frequency tuning range is relatively small. Additionally, most narrow-band systems are static in frequency. The operating frequency does not change dynamically. In contrast, MB-OFDM systems use time-frequency interleaving, and, as shown in Fig. I.13, the operating frequency could change from 3.4 GHz to 10 GHz in less than 400 nS. Not only will the DC offset change significantly every 400nS, offset calibration has to settle in less than
9 nS, which is the specified guard time interval for MB-OFDM. This creates a major problem for DC offset calibration, as shown in Fig. III.3. Designing a DC offset calibration loop that settles this quickly is an enormous challenge.

Figure III.3: DC offset and second-order distortion issues in direct conversion MB-OFDM receivers.

The second method to remove DC offset is to use a DC blocking capacitor at the baseband output. This creates a different problem for MB-OFDM. When the received signal hops in frequency, the received signal amplitude will change, due to the change in free space path loss. A 1MHz pole at the baseband output produces a time constant of 1μs, as shown in Fig. III.3. This means that with every frequency hop, the receiver will take approximately 1μs to settle down. Since the frequency hops themselves are less than 400nS long, a 1μs settling time is not acceptable.

2. Second-order distortion: Since DC offset is higher in a direct-conversion receiver, due to self mixing of the LO and RF, direct-downconversion receivers exhibit
worse $IIP_2$ performance. In a UWB system the second-order distortion of in-band channels can also land in-band. Therefore, second-order distortion is a major concern for MB-OFDM receivers. Direct-conversion receivers for MB-OFDM system are, therefore, not as robust as heterodyne receivers in this situation. Additionally, harmonics of the received signal can mix with the harmonics of the LO and downconvert to baseband, posing an additional distortion problem.

3. **LO Re-Radiation:** Since the RF frequency is equal to the frequency of the received signal, the LO will always be in-band for an MB-OFDM system. Although MB-OFDM is not a full-duplex system (the transmitter and receiver are not on at the same time), the LO is typically not turned off, since the transmitter and receiver share the same LO. Any LO leakage from the receiver can corrupt the spectrum of the accompanying transmitter.

In contrast, in a heterodyne receiver, the LO and RF are at different frequencies. If the frequency plan is carefully chosen, the LO may always be out-of-band.

4. **Baseband Channel Select Filtering:** In a direct-conversion receiver, there can be no intermediate channel select filtering. All the channel select filtering has to happen after the downconversion process. As a consequence, the baseband low pass filter will have to be a high order filter to suppress the adjacent channels. Since the signal at baseband is fairly high in amplitude (of the order of 1V), an active LPF with multiple poles is usually very power hungry [23].

In a heterodyne receiver, an inter-stage channel select filter (after the first downconversion stage) will ease the specification on the baseband LPF. However, this usually comes at a cost of die area, power consumption and/or external component cost.

The issues mentioned above are a very big hurdle to a practical MB-OFDM direct conversion receiver. It is important to take a step back and see if a heterodyne receiver might be a more suitable design.
III.D Heterodyne Receiver

A typical heterodyne receiver is shown in Fig. III.4. A heterodyne receiver provides some advantages for MB-OFDM systems, such as:

1. Improved Distortion Performance: Heterodyne receivers are less susceptible to harmonics of the LO mixing with harmonics of the received signal and corrupting the SNR of the received signal. With careful frequency planning, all the lower order harmonic products can be pushed away from the downconverted signal, and can then be filtered with a channel select filter at the intermediate frequency.

2. Reduced DC-Offset: Since the LO and RF are at different frequencies, the DC offset in the first downconversion stage of a heterodyne receiver is lower, and arises mainly due to imbalances in the circuits. Any DC offset present at the output of the first downconversion stage can easily be removed by a DC blocking capacitor, without effecting the received signal. The LO and signal frequencies will be equal in the second downconversion stage, since the signal has to be downconverted to baseband. This can create additional DC offsets in the second downconversion stage. However, the first stage of downconversion in a heterodyne receiver usually has in excess of 20 dB of gain, and the effect of DC offsets in the second downconversion stage is far lower.
3. Inter-stage channel select filter: If the heterodyne receiver is designed with a fixed Intermediate Frequency (IF), it provides an opportunity for some inter-stage channel select filtering, which eases the adjacent channel suppression requirements of the baseband LPF.

The above mentioned advantages of a heterodyne receiver make it a very suitable candidate for MB-OFDM receivers. However, a heterodyne receiver has its own disadvantages:

1. Need for image rejection: This is the classical issue with all heterodyne receivers. Every received signal, in a heterodyne receiver, has an image that can downconvert to the desired channel and reduce the SNR of the signal. This is shown in Fig. III.5.

![Image Downconversion in a Heterodyne Receiver](image)

Figure III.5: Image downconversion in a heterodyne receiver.

Rejection of images is essential to maintain the SNR of the output signal. Image rejection is usually achieved in one or both of the following ways:

(a) Image reject downconverter: An image reject downconverter utilizes an LO in quadrature to reject the images, as shown in Fig. III.29 [24]. The incoming RF and image are multiplied, using a mixer, with an LO in quadrature. With the correct polarity of the LO, the image can be cancelled at the output of the downconverter, as shown in Fig. III.29.

The key to obtaining good image rejection is to maintain a perfect 90° phase of the LO and RF and to balance the amplitudes of the RF signal in both branches of the downconverter. The Image Rejection Ratio (IRR) is defined
as the ratio of the desired signal amplitude at $\omega_{lo} - \omega_{rf}$ to the amplitude of
the image signal, at $\omega_{im} - \omega_{lo}$. In the presence of amplitude and/or phase
imbalance at the input of the summing block in Fig. III.29, the $IRR$ is given
by,

$$IRR = \frac{1 + \delta^2 + 2\delta \cos(\theta)}{1 + \delta^2 - 2\delta \cos(\theta)} \quad (III.1)$$

Where $\delta = \text{amplitude imbalance of the signal in the two branches and}$
$\theta = \text{phase imbalance of the signal in the two branches.}$

Usually, a $90^0$ hybrid coupler or a polyphase filter are used to generate the
$90^0$ phase shift for the RF signal. Hybrid couplers are difficult to manufac-
ture on-chip, since they occupy a large die are. For example, to generate a
$90^0$ phase shift for a 3 GHz signal, one needs a transmission line of length
2mm in Silicon, which is very large. Off-chip couplers are expensive and
increase the cost of the total system significantly.
Polyphase filters can be active or passive. Active polyphase filters can have good wideband performance, but usually have poor linearity, high current consumption and high Noise Figure [25] [26]. Passive polyphase filters use RC poles to generate a phase shift in the RF signal [27]. Passive filters in the 3 - 10 GHz frequency range occupy smaller die area. For example, an RC pole at 3 GHz, with a resistor of 100 Ω, needs a capacitor of 3.3pF. Most CMOS processes offer on-chip capacitors with a density in the range of $5fF/\mu M$. This implies a capacitor of $660\mu M^2$, which is small. However, designing a polyphase filter which operates all the way from 3 GHz - 10 GHz is extremely difficult. A filter with this high a bandwidth requires multiple RC poles, which introduces significant loss in the RF signal [27]. The result would be a severely degraded noise figure for the downconverter. Additionally, it is difficult to maintain amplitude and phase balance when multiple RC poles are employed, due to mismatch in R and C values and circuit parasitics.

Quadrature LO is usually generated by operating the frequency synthesizer at twice the desired LO frequency, and then using a frequency divider to provide a $90^\circ$ phase shift. MB-OFDM systems operate up to 10 GHz, and the design of a low power synthesizer and Voltage Controlled Oscillation (VCO) which operates at 20 GHz is very challenging.

(b) Front-end image reject filter: The images are rejected before entering the LNA at the front-end. This requires the images to always be out of the MB-OFDM band, and to achieve sufficient rejection, be far away from the receive band. Since MB-OFDM systems utilize a front-end BPF for rejection of out-of-band signals, it would be advantageous to design a heterodyne frequency plan which would take advantage of this and provide significant image rejection.

2. Need for an inter-stage filter: Since the IF output is fairly high in amplitude, there
needs to be some inter-stage filtering to reduce the adjacent channels, to ease the linearity specification of the IF downconverter. A bandpass filter is required at the IF output to mitigate the linearity requirements on the following IF downconverter chain. The filter needs to suppress the adjacent channels present in a multiple-piconet UWB system. On-chip passive bandpass filters require high Q inductors, which can add significant die area and force the use of a thick top metal, which will add to the total cost of the die. Active filters, on the other hand, can be very power hungry and can also cause a significant increase in the Noise Figure and distortion of the system [23]. This challenge needs to be overcome for a viable heterodyne receiver.

3. Need for two LO frequencies: Since there are two downconversion stages, each mixer will need a separate LO frequency. This could result in two separate LO generators, which adds to cost and die area.

The design chosen for this research is heterodyne downconverter. The goal of this research is to come up with a heterodyne receiver design which avoids all the classical problems of a direct downconversion receiver. The design has to overcome all the challenges of a heterodyne receiver, but should cost the same or less than a direct downconversion receiver, both in terms of power consumption and cost.

III.E Receiver Architecture and Frequency Planning

The choice of IF frequency for the heterodyne needs to meet the following objectives:

1. Allow for one frequency synthesizer for the whole chip.
2. Provide image rejection.
3. Reduce susceptibility to interferers.
For a given IF and RF frequencies ($\omega_{if}, \omega_{rf}$), there exist two LO frequencies that can downconvert the RF to IF.

\[
\omega_{lo1} = \omega_{rf} - \omega_{if} \\
\omega_{lo2} = \omega_{rf} + \omega_{if}
\]  

(III.2)

For a high frequency system such as MB-OFDM, $\omega_{lo1}$ is a better choice, since the operating frequency of the VCO and LO synthesizer is lower, resulting in easier design and lower power consumption. If the LO frequency is $\omega_{lo1}$, as given by (III.2), then the image of the received signal is at

\[
\omega_{im} = \omega_{lo1} - \omega_{if} = \omega_{rf} - 2\omega_{if}
\]  

(III.3)

After careful inspection of the channel frequencies, 2.64 GHz turns out to be a very good choice for the IF frequency. If the IF is at 2.64 GHz, all the images fall below 2.64 GHz, as shown in Table III.E. An external BPF with a cut-off frequency of 3.1 GHz will significantly attenuate all the images.

Table III.2: LO and image frequencies for and IF of 2.64 GHz.

<table>
<thead>
<tr>
<th>Band No</th>
<th>Band Freq(MHz)</th>
<th>LO (MHz)</th>
<th>Image (MHz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>3432</td>
<td>792</td>
<td>1848</td>
</tr>
<tr>
<td>2</td>
<td>3960</td>
<td>1320</td>
<td>1320</td>
</tr>
<tr>
<td>3</td>
<td>4488</td>
<td>1848</td>
<td>792</td>
</tr>
<tr>
<td>4</td>
<td>5016</td>
<td>2376</td>
<td>264</td>
</tr>
<tr>
<td>5</td>
<td>5544</td>
<td>2904</td>
<td>264</td>
</tr>
<tr>
<td>6</td>
<td>6072</td>
<td>3232</td>
<td>792</td>
</tr>
<tr>
<td>7</td>
<td>6600</td>
<td>3960</td>
<td>1320</td>
</tr>
<tr>
<td>8</td>
<td>7128</td>
<td>4488</td>
<td>1848</td>
</tr>
<tr>
<td>9</td>
<td>7656</td>
<td>5016</td>
<td>2376</td>
</tr>
</tbody>
</table>

It is also clear from Table III.E that all the required LO frequencies (including the IF LO) can be generated using the same synthesizer as the transmitter, as they are all harmonically related. Therefore, both the objectives of image rejection and one
frequency synthesizer for the chip can be met if the IF is set to 2.64 GHz. The receiver design is therefore targeted towards an IF frequency of 2.64 GHz.

The LNA will be a broadband LNA that accommodates RF signals from 3.1 GHz - 8 GHz. However, the goal will be to avoid using any on-chip or off-chip inductors, to reduce the cost of the chip.

Since it is anticipated that the external BPF will provide significant image rejection, the RF downconverter will not be an image reject mixer, but a standard Gilbert cell based mixer. There will also be an on-chip bandpass filter to provide for channel selection and to ease the linearity specifications of the following IF downconverter. However, this filter also has to be designed without any on-chip or off-chip inductors.

A schematic of the RF front-end along with the frequency plan is show in Fig. III.7.

![Diagram of the RF front-end and frequency plan](image)

Figure III.7: (a) Proposed receiver architecture. (b) Receiver frequency plan. All the images are outside the desired band.
III.F Receiver Specifications

III.F.1 Receiver Linearity

Receiver $IIP_3$

Although UWB systems have not been widely deployed, it is recognized that the interference environment for these devices is a challenge. Potential in-band interferers include WiMax and WiFi devices, as well as airport and marine radars [14]. For the receiver to maintain adequate link margin and sensitivity, careful consideration needs to be given to cross-modulation distortion [13], [15].

In a typical cross-modulation scenario, the UWB system might receive a narrowband jammer at -30 dBm (a WiMax signal for one example), along with a wideband jammer from another transmitting UWB system, which could be located as close as 0.1m from the receiver. The average transmitted power of a UWB transmitter is -10.3 dBm [10]. At 0.1m, this power would drop to roughly -34.5 dBm for a Mode 1 system. The cross-modulation product of a narrowband jammer with an MB-OFDM transmitter is given by equation (II.43).

The effect of receiver $IIP_3$ on the SNR due to this cross-modulation distortion can be computed by integrating (II.43) over the bandwidth of one MB-OFDM receive band. This cross-modulation power should be added to the output in-band noise power of the receiver, to compute the final SNR of the output signal.

For example, assuming a Noise Figure of 6 dB and an overall gain of 15dB for the LNA/downconverter, the output noise power at the desired channel of 3.5 GHz, due to thermal noise at the input and the noise generated inside the LNA and downconverter is -65.84 dBm. With a narrowband jammer of -40 dBm at 3 GHz and a wideband jammer of -34.5 dBm at 4.4 GHz [13], integrating (II.43) from 3150 MHz to 3500 MHz, corresponds to a total cross-modulation power at the mixer output of -66.6 dBm. This cross-modulation effectively increases the Noise Figure of the receiver by 2.6 dB. This implies a reduction of approximately 2.6 dB in receiver sensitivity.

To optimize the required $IIP_3$ of the receiver, we need to compute this degra-
dation in link margin over a variety of UWB jammer powers.

For an operating distance of 0.1m, the maximum received UWB signal power is -34.5 dBm. Fig. III.8 plots the link margin of the receiver as a function of receiver \( IIP_3 \) for a receiver gain of 15 dB and 20 dB using (II.43), when the narrowband jammer power is -40 dBm and wideband jammer power is -34.5 dBm. It can be seen from the plots that a receiver \( IIP_3 \) of -5 dBm results in a link margin degradation of less than 1 dB. The receiver was therefore designed for an \( IIP_3 \) of -5 dBm.

![Figure III.8: Receiver link margin for the case of a narrowband jammer at -40 dBm and a wideband jammer at -34.5 dBm.](image)

**Receiver \( IIP_2 \)**

Receiver \( IIP_2 \) is usually not very important for heterodyne receiver front ends. However, in an MB-OFDM system, since the second harmonic of an undesired trans-
mitter may land in a desired channel, close attention needs to be given to receiver $IIP_2$ specification. For example, the second harmonic of a Band 1 transmitter is 6864 MHz, which will reduce the SNR for a receiver tuned to Band 7 (6600 MHz) or Band 8 (7128 MHz), since the second harmonic is spread to twice the bandwidth of the original signal. This situation is depicted in Fig. II.22.

The power spectral density of this second-harmonic product can be computed using (II.69)

The effect of second harmonic distortion in the receiver on the total link margin, for the situation depicted in Fig. II.22 is shown in Fig. III.9. The receiver gain is assumed to be 20 dB and the Noise Figure is 6 dB. It can be seen that the effect of second harmonic distortion is negligible if the receiver $IIP_2$ is greater than 20 dBm. Therefore, the receiver shall be designed for an $IIP_2$ of 20 dBm.

![Figure III.9: Effect to receiver $IIP_2$ on total link margin. Receiver gain is 20 dB and Noise Figure is 6 dB.](image-url)
III.F.2 Receiver Gain and Noise Figure

To reduce the impact of the IF downconverter Noise Figure on the total receiver Noise Figure, it is advantageous to provide as large a gain in the front-end as possible. However, this can lead to two complications. First, since the LNA is very wideband, providing a large gain will prove to be expensive in terms of power consumption.

Additionally, a high gain in the receiver front-end will increase the linearity requirement of the RF mixer and the IF downconverter chain, which could lead to even higher power consumption.

For example, consider the situation of three nearby MB-OFDM transmitters, transmitting at their maximum power of -10.3 dBm. At 0.1m, the front-end receives three MB-OFDM signals with a total maximum received power of -29.7 dBm. To avoid gain compression in the front-end, the input referred 1-dB compression point has to be greater than -29.7 dBm. Adding another 10 dB of margin, the receiver input referred 1-dB compression point should be -19.7 dBm. Representing the receiver nonlinearity as a third-order power series and using the well known approximation

\[ IIP_3 \approx P1dB + 10 \]  

(III.4)

The required LNA \( IIP_3 \) will be -9.7 dBm.

The trade-off between required IF downconverter \( IIP_3 \) and Noise Figure is shown in Fig. III.10

From Fig. III.10 it is clear that there is a substantial trade-off between the required \( IIP_3 \) of the mixer and the IF downconverter chain, and their Noise Figure. Since the LNA is wideband, its gain is expected to vary over frequency. The goal will be to keep the front end gain between 15 dB and 20 dB. To ease the Noise Figure requirement on the mixer, all the gain will be provided in the LNA.

III.G Existing UWB Receivers

Table III.3 lists the features of some recently published UWB receivers.
Figure III.10: Required IF downconverter $IIP_3$ and Noise Figure to maintain a total receiver $IIP_3$ of -9.7 dBm and total receiver Noise Figure of 6.6 dB. Front-End Noise Figure is 6 dB.

The goal of this research will be to design the receiver such that the performance and power consumption of the receiver is similar to the numbers in Table III.3. The additional constraint is to avoid using any inductors to save die area and reduce the cost of the chip. All the major components of the receiver (LNA, mixer and BPF) will be designed with these constraints.

III.H Low-Noise Amplifier Design

III.H.1 Existing UWB LNA Techniques

The receiver is designed to cover all the UWB bands from 3.1 GHz - 8 GHz. The three standard techniques of designing a wideband LNA are (a) resistive input and output matching (b) multiple section LC input and output match and (c) distributed
Table III.3: Performance of some recently published UWB receivers.

<table>
<thead>
<tr>
<th>Ref</th>
<th>Tech.</th>
<th>Band Groups</th>
<th>Inductors</th>
<th>Power</th>
<th>$IIP_3$</th>
<th>NF</th>
</tr>
</thead>
<tbody>
<tr>
<td>[19]</td>
<td>SiGe</td>
<td>1</td>
<td>4</td>
<td>117.5 mW</td>
<td>-3 dBm</td>
<td>3 dB</td>
</tr>
<tr>
<td>[20]</td>
<td>SiGe</td>
<td>1, 3</td>
<td>5</td>
<td>59.4 mW</td>
<td>NR</td>
<td>3.3 dB</td>
</tr>
<tr>
<td>[21]</td>
<td>CMOS</td>
<td>1</td>
<td>4</td>
<td>105 mW</td>
<td>-37.5 dBm</td>
<td>5.5 dB</td>
</tr>
</tbody>
</table>

amplifier.

Figure III.11: Broadband LNA with shunt resistive matching and resistive load.

1. Broadband LNA with resistive input and output match (Fig. III.11):

   Broadband resistive LNAs provide an input match via RC feedback and output loading with a resistor, as shown in Fig. III.11. Resistive LNAs occupy very little die area, since they only consist of a few transistors and only resistors and capacitors for passives, but they do have the following drawbacks:

   (a) Poor high frequency performance

   Although broadband matching is good in a resistive LNA, it is difficult to provide good high frequency response. For example, with a 100 $\Omega$ load and a 1 pF output device capacitance (which is quite common), there is a pole in the output at approximately 1.9 GHz. The gain will therefore be
severely degraded near 8 GHz, where the LNA for this receiver needs to operate. Techniques such as multiple resistive feedback have been employed to counter this issue [28]. However, such techniques come at the expensive of higher power consumption and poor Noise Figure.

(b) Poor Noise Figure and linearity
A resistive input and output matched LNA is the best solution in terms of die area. However, this will result in a significantly increased current consumption and high Noise Figure. For example, the LNA in [28] consumes 17 mA from a 2.5 V supply, but the $IIP_3$ is only -11.75 dBm. To achieve the linearity requirements of a realistic UWB receiver, this current, which is already quite high, will have to be increased significantly.

2. Broadband LNA with multiple LC sections (Fig. III.12):

Multiple LC section LNA can provide very good broadband matching, high gain and low Noise Figure. However, multiple LC sections result in a very large die area. For example, the LNA in [29] designed in a SiGe BiCMOS process, uses a ladder matching network with three on-chip inductors, the resulting die area is 1.8$mm^2$.

The LNA presented in [30] uses multiple LC sections for input match and a shunt peaking load, resulting in five on-chip inductors. While covering a similar number of bands, the LNA in [30] consumes 1.1$mm^2$ of die area. For comparison, the total die area in the design presented here (including the LNA, mixer and BPF) is only 0.35 $mm^2$. Additionally, a multiple LC match would require high Q inductors, which might require a thick top metal, adding additional cost to the chip.

3. Distributed LNA (Fig. III.13):

Distributed amplifiers can provide very good wideband gain and linearity [32], however, their noise performance is quite poor, power consumption is high and they occupy a very large die area. Their applicability as a low noise amplifier for a low-cost low-power system is therefore very limited.
Figure III.12: Broadband LNA with multiple LC section input match and shunt peaking load [30].

### III.H.2 Improved LNA Design

All of the LNA design techniques presented above are not suitable for a low cost, low power UWB receiver. With all the design techniques presented above, either the power consumption is too high or die area is too large. A new creative design is needed to overcome all of these challenges.

The major issue with all the above presented designs is that they all try to target a UWB LNA, without taking any advantage of the special time-frequency interleaving of MB-OFDM. A new design technique is need which exploits the time-frequency interleaving of MB-OFDM systems and provides a low-cost, low-power design for MB-OFDM.

Since MB-OFDM is a frequency hopping system, it is of interest to note that at any given time, the required bandwidth of the LNA is only 528 MHz. Although the total frequency coverage required is from 3.1 - 8 GHz, at no point in time does the LNA need to tune to the complete MB-OFDM spectrum.
One major goal of this design is to avoid using any on-chip or off-chip inductors. Since resistive LNA and distributed LNA have been ruled out due to high power consumption, the only remaining technique is a broadband match using multiple LC sections. This conflicts with our goal of not using any on-chip or off-chip inductors. Every chip needs to be eventually packaged to be converted into a product. Most packages inadvertently offer bondwires as inductors for free, since bondwires are employed to connect the die to the package itself. In most cases bondwires are considered a nemesis by designers. They are treated as parasitics that need to be dealt with somehow. However, we propose a solution that uses these inductors to a very significant advantage.

We propose an LNA that uses a pair of bondwires as a differential load, as shown in Fig. III.14. The package bondwire inductance for our package (MLF24) was estimated to be approximately 1.2 nH. The device output capacitance (of the LNA and mixer) is chosen to resonate with the bondwires at 7 GHz to create a resonant tank. The frequency response is then broadened by a shunt resistor. This provides sufficient gain for the three highest bands (6.3 - 7.9 GHz). However, even with a shunt resistor, the frequency response of a single resonant tank is fairly narrow band. Fig. III.16 shows the normalized frequency response of a resonant LCR tank, as shown in Fig. III.15, with $L = 1.2nH$, $C = 0.43pF$ and $R_{\text{shunt}} = 200\Omega$.

It can be seen from Fig. III.16 that the frequency response of the tank drops significantly beyond a 2 GHz bandwidth. This frequency response can be broadened by reducing $R_{\text{shunt}}$. However, this will cause a drop in the gain of the LNA and will also
To achieve frequency tuning, we take advantage of the frequency hopping nature of MB-OFDM and switch in extra capacitance to re-tune the LNA when the received signal is in one of the lower band groups, as shown in Fig. III.14. When the received signal is in the lower bands of Group 2 or Group 1, extra shunt capacitance is switched in, by setting bits $b_0$ and $b_1$ “high” successively, to dynamically maximize the gain of the LNA at 5.5 GHz and 4 GHz.

This scheme allows for load tuning of the LNA without using any on-chip inductors. However, it does have a drawback in sensitivity to bondwire length. The bondwire inductance for this package at the LNA input and LNA output was approxi-
approximately 1.2 nH. In a typical bonding process, the variation in bondwire length is usually ± 10%. Additionally, the tank capacitance can vary by ±20% due to variation in device output capacitance and variation in the shunt tuning capacitors. This variation would result in a center frequency variation of ±15%. However, the simulation results of Fig. III.17 show that this will result in at most a 1 dB drop in gain at band edges, which is not significant. The reduced susceptibility to variations in bondwire length and on-chip capacitance can be attributed to the low Q of the LNA load tank.

A differential design is employed to reduce second-order distortion, which is a serious concern for wideband systems since the second-harmonic distortion of a received jamming transmitter can appear in the band of interest. Also, if a single-ended amplifier were employed, the downbond inductance of the package would act as source degeneration for the amplifier. This degeneration would be frequency dependent and the amplifier gain would be a strong function of the input frequency.
Figure III.16: Normalized frequency response of an LCR resonant with $L = 1.2 \text{nH}$, $C = 0.43 \text{pF}$ and $R_{\text{shunt}} = 200\Omega$.

Broadband input matching of the amplifier was achieved by a shunt RC feedback network, as shown in Fig. III.14. Although this provides for a good wideband input match, it degrades the Noise Figure of the LNA. The simulated worst case Noise Figure of this LNA was 4.5 dB, which is higher than that reported in [29]-[31]. However, this Noise Figure is adequate to maintain a 6.0 dB Noise Figure for the entire front-end. Even with a worst case LNA Noise Figure of 4.5 dB and gain of 15 dB, the required RF mixer Noise Figure is 16 dB or less, to achieve a receiver Noise Figure of 6 dB.

The LNA consumes 11 mA from a 2.3V supply, and simulated gain was between 15-18.5 dB. The simulated $IIP_3$ was -1.2 dBm and the worst case input return
loss was -7 dB. The performance and power consumption of this LNA is comparable to that reported in [29]-[31], but this LNA has a significantly smaller die area, since no on-chip inductors are needed. A comparison of the performance of the LNA designed in this research work to that of other published work is shown in Table III.4.

Simulated LNA Noise Figure was less than 4 dB. The Noise Figure in the low frequency band is worse due to the lower gain of the LNA and the noise added by the switches and tuning capacitors.

Additionally, this LNA provides rejection of undesired jammers and images. For example, when the receiver is tuned to Group 3, from the simulated LNA frequency response in Fig. III.17, it can be seen that the LNA rejects the signal from any transmitter transmitting in Group 1 or Group 2 by at least 4 dB. The LNA frequency response also provides for a minimum of 8 dB rejection of 802.11b/g and Bluetooth jammers.
Table III.4: Comparison of the performance of the LNA designed in this research, to that of other published work.

<table>
<thead>
<tr>
<th>Ref.</th>
<th>Tech.</th>
<th>Freq. (GHz)</th>
<th>Inductors</th>
<th>Current</th>
<th>$IIP_{3}$</th>
<th>NF</th>
</tr>
</thead>
<tbody>
<tr>
<td>[29]</td>
<td>SiGe</td>
<td>3-10</td>
<td>4</td>
<td>10 mA</td>
<td>-1 dBm</td>
<td>2.5 dB</td>
</tr>
<tr>
<td>[30]</td>
<td>CMOS</td>
<td>3-10</td>
<td>5</td>
<td>5 mA</td>
<td>-6.7 dBm</td>
<td>4 dB</td>
</tr>
<tr>
<td>[31]</td>
<td>CMOS</td>
<td>2-5.2</td>
<td>4</td>
<td>19 mA</td>
<td>-34 dBm</td>
<td>4.7 dB</td>
</tr>
<tr>
<td>This research</td>
<td>CMOS</td>
<td>3 - 8</td>
<td>0</td>
<td>11 mA</td>
<td>-1.2 dBm</td>
<td>4.5 dB</td>
</tr>
</tbody>
</table>

transmitting in the 2.4 GHz unlicensed band.

For the frequency plan described in Section III.7, all the images are below 2.64 GHz. The received signal and its image are closest when the received signal is in Band 1 at 3432 MHz. The image is then at 1848 MHz. It can be seen from Fig. III.17 that the LNA provides approximately 10 dB of image rejection in this case. Therefore, significant image rejection has been achieved with this LNA design, reducing the need for an image reject mixer or a front end image reject filter.

### III.I RF Downconverting Mixer

The downconverter (Fig. III.18) is a wideband, double balanced Gilbert Cell based mixer. The LNA provides for a minimum of 10 dB of image rejection and further image suppression is obtained by the external bandpass filter before the LNA. Therefore, the mixer does not need to provide any additional image rejection. Resistive loading and degeneration was employed in the mixer for wideband performance.

The simulated gain for the mixer was 0 dB and its simulated Noise Figure was 13 dB. The mixer consumes 8.5 mA from a 2.3 V supply.

### III.J IF Bandpass Filter

Since there might be multiple MB-OFDM transmitters operating in the vicinity, it is necessary to provide for some on-chip IF channel selection to ease linearity requirements on the remainder of the receive chain and eliminate the need for an external IF filter. Unfortunately, active filters have well known dynamic range and power
limitations, and an on-chip LC filter would require high-Q inductors and significant die area. The goal of this design is to keep the area as small as possible, while maintaining adequate performance.

A standard single-ended bandpass Chebyshev filter (Fig. III.20) would require three inductors, one of which would be very large. Usually, the ratio of the inductor in the series resonator to the inductor in the shunt resonator is of the order of the square of the fractional bandwidth of the filter [33]. For our purposes, the fractional bandwidth is approximately $\frac{1}{5}$. Even if the shunt inductors were as small as 1 nH, the series inductor would have to be about 25 nH. Filter loss and adjacent channel rejection are directly related to the Q of these inductors. If such a filter were to be designed on-chip, it would require a very large die area and thick top metal for high Q inductors.

An impedance inversion transformation [33] can be applied to convert a series LC network to a parallel LC network, as shown in Fig. III.19. The negative capacitors in the transformation network can be absorbed by the capacitors in the parallel LC tank.
Figure III.19: Impedance transformation to convert a series LC resonator to a parallel resonator.

in either side of the Chebyshev filter.

Applying this transformation to a standard single-ended third-order passive bandpass Chebyshev filter, results in a “top-C” coupled structure, with fixed equal-valued shunt grounded inductors, as shown in Fig. III.20. It is important to note that this impedance transformation is valid for only one frequency, so the filter rejection is limited compared to the third-order prototype filter. All the inductors in this filter are grounded and can therefore be replaced with downbonds, which have a very high Q. The resulting filter in its differential form is shown in Fig. III.21. The downbond inductance for our package was determined to be 0.8nH and the capacitor values were chosen appropriately.

The filter was designed for an IF center frequency of 2.64 GHz and a bandwidth of 528 MHz. A 3D illustration of this BPF is shown in Fig. III.22.

Since the on-chip capacitance can vary by ±20% and the bondwire inductance can vary by approximately ±10%, it is important to provide some mechanism for tuning the bandpass filter. To this end, the shunt capacitors in the bandpass filters were replaced by MOS varactors. The measured tuning range was ±300 MHz, which is enough to accommodate the capacitance and bondwire inductance variation.

A potential problem with this filter is the possibility of noise coupling onto the bond-wires. But since the receiver has almost 20 dB of gain before the filter, this is not a significant issue.
III.J.1 Impact of IF Bandpass Filter

The filter is especially important in a high interference scenario, since it eases both the gain compression and the intermodulation specification of the IF downconverter. If an off-chip filter were to be employed, it would increase the cost of the system since high frequency off-chip bandpass filters are expensive, extra pins would be required for the signal to be routed off-chip and the off-chip filter can occupy significant board area.

IF Downconverter Gain Compression

In the high interference scenario of three adjacent MB-OFDM transmitters, transmitting at their maximum power of -10.3 dBm, the signal at the receiver antenna is -34.5 dBm from each of the transmitters, for a total input power of -29.7 dBm. With a 20 dB gain in the RF chain, this results in a receiver output of -9.7 dBm.
However, an IF bandpass filter with an adjacent rejection of 8 dB, suppresses the two unwanted sidebands by 8 dB. The power at the IF downconverter input will be -14.5 dBm in the desired channel, and -22.5 dBm in each of the two adjacent channels, resulting in an output power of -13.3 dBm. The IF bandpass filter, therefore, eases the gain compression specification of the IF downconverter by 3.6 dB.

**IF Downconverter Intermodulation Distortion**

The BPF also reduces the IM distortion requirement on the IF downconverter. In a typical intermodulation distortion scenario, adjacent channels present at the IF input can be downconverted to baseband in the presence of third-order nonlinearity in the IF downconverter, resulting in reduced baseband SNR.

As shown in Fig. III.23, a channel select filter at the RF downconverter output can improve the SNR at the baseband output. This implies that, for a given baseband output SNR, a channel select filter reduces the third-order intermodulation specification of the IF downconverter.
Figure III.22: Three-dimensional view of the BPF created entirely with on-chip capacitors, varactors and downbond inductors.

To calculate the intermodulation output at baseband, we can use [13]

\[
PSD_{IM3}(\omega) = \frac{9}{4} e_3^2 A_{tx1}^4 A_{tx2}^2 T_s \sum_{k=-\frac{N_s}{2}}^{\frac{N_s}{2}} \sum_{m=-\frac{N_s}{2}}^{\frac{N_s}{2}} \sum_{n=-\frac{N_s}{2}}^{\frac{N_s}{2}} \left\{ \text{sinc}^2 \left( (\omega + (2\omega_{tx1} - \omega_{tx2} + \Delta\omega(m + n + k)) T_s \right) + \text{sinc}^2 \left( (\omega - (2\omega_{tx1} - \omega_{tx2} + \Delta\omega(m + n + k)) T_s \right) \right\}
\]

(III.5)

Where \(A_{tx1}\) and \(A_{tx2}\) are the \textit{rms} amplitudes of the two undesired MB-OFDM signals and can be related to the corresponding RF input power using (II.30), \(\omega_{tx1}\) and \(\omega_{tx2}\) are the angular frequencies. Equation (III.5) can be integrated over the bandwidth of one MB-OFDM channel to obtain the total output noise power due to intermodulation.

In the situation described in Fig. III.23, the degradation in baseband SNR due to third-order distortion in the IF downconverter can be calculated by adding the intermodulation distortion power from (III.5) to the total output noise at baseband. From [10], the required baseband SNR is 6.5 dB (required \(\frac{E_b}{N_0}\) of 4 dB plus an implementation loss of 2.5 dB). Assuming an IF downconverter gain of 60 dB and Noise Figure of 10
Figure III.23: Third-order IM distortion in the IF downconverter. (a) Without an IF channel select filter. (b) With an IF channel select filter.

dB, using (III.5), it can be shown that the IF downconverter $IIP_3$ has to be 7.5 dBm to maintain a 6.5 dB SNR in the presence of two jamming signals at -14.5 dBm. However, with the channel select filter, which provides an 8 dB rejection of the closest channel and a 12 dB rejection of the second adjacent channel, to maintain the same SNR at baseband, the required IF downconverter $IIP_3$ is only -8 dBm. This channel select filter, therefore, relaxes the $IIP_3$ distortion specification of the IF downconverter by approximately 15 dB.

### III.K Chip Layout and Packaging

Since this design relies on accurate modeling, careful positioning of individual blocks and bondwires is essential. Fig. III.24 shows a block diagram of the chip and the bondwires.

Inductance coupling between bondwires is an issue since it can result in spurious tones and even result in instability in the circuit. Simulations determined that coupling between the down-bonds of the bandpass filter created undesired poles and
zeroes and significantly affected the performance of the filter. In addition, since it is difficult to accurately estimate the coupling coefficient between the bondwires, it is hard to predict the resulting performance of the filter. Electromagnetic simulations demonstrated that the mutual coupling coefficient had to be less than 0.1 for the coupling to be negligible and that if the filter downbonds were separated by at least one bondwire, the coupling coefficient drops to less than 0.1. Therefore, all the downbonds on the chip are separated by at least one bondwire. In addition, the two differential halves of the bandpass filter were placed on perpendicular edges of the chip to reduce coupling between the two sides. Similarly, the LNA input and load tank bondwire pins were placed on perpendicular edges of the chip to reduce the coupling that would lead to potential instability. Fig. III.24 shows the layout of the chip and positioning of bondwires and Fig. III.25 is a micrograph of the chip.

Figure III.24: Layout of UWB receiver.
The chip was fabricated in the Jazz Semiconductor 0.18 $\mu M$ RF CMOS process and was packaged in a 24 pin MLF package.

### III.L Measured Results

#### III.L.1 Receiver Gain

The measured gain of the receiver in each of the MB-OFDM Bands is shown in Fig. III.26. The gain varies from 15 dB in Band 1 (3.5 GHz) to 20 dB in Band 9 (7.7 GHz). This gain variation is expected due to the loss in the switches of the shunt capacitors C1/C2 in Fig. III.14. Although the gain varies by 6 dB between the bands, this is not a significant issue since the received signal strength will vary significantly over frequency. The receiver gain drops by roughly 2 dB at each of the band edges, but this has a small effect, since the ten carriers at the band edge are guard carriers [10].
III.L.2 Receiver Noise Figure and Linearity

The measured receiver $IIP_3$ and Noise Figure are plotted in Fig. III.27. The Noise Figure varies between 5 and 6.5 dB. This is due to both the variation in LNA gain and the variation in image rejection from the LNA (Fig. III.29). The measured $IIP_3$ is between -5.5 dBm and -2.5 dBm, which is sufficient for the receiver to perform adequately in the cross-modulation scenario discussed in Section III.F.1.

Second harmonic distortion for the receiver is important, since the second harmonic of a transmitter, transmitting in Band 1 or Band 2 can reduce the SNR of a receiver tuned to Band 6, Band 7 or Band 8. $IIP_2$ of the receiver was measured by injecting a single tone in Band 1 and measuring the second-harmonic at the output, while tuning the receiver to Band 6. The measured $IIP_2$ of the receiver was 33 dBm, which
meets the specification. The $IIP_3$ performance of this receiver is very good, without the use of any calibration scheme due to the LNA frequency response.

### III.L.3 Bandpass Filter Rejection

The bandpass filter achieves approximately 7 dB rejection of the closest adjacent channel (528 MHz away) and greater than 12 dB rejection of the second adjacent channel (1056 MHz away).

Since the LNA does not have a flat frequency response, it also contributes to the adjacent channel rejection. The variation in the adjacent channel rejection over bands is due to the frequency response of the LNA. The total measured tuning range of the BPF was ±300 MHz.
Figure III.28: Measured BPF Adjacent Channel Rejection and Input Return Loss.

### III.L.4 Receiver Image Rejection

The measured Image Rejection of the receiver is shown in Fig. III.29. The image rejection is determined by the load tuning of the LNA. In the lowest band, since the image is only 1.6 GHz away, the image rejection is lowest. Image Rejection increases to 35 dB as the center frequency of the band increases, since the image moves closer to DC. The overall image rejection will be significantly improved by the off-chip filter.

The receiver consumes only 19.5 mA at a dc bias voltage of 2.3 V. Total active die area was 0.35 \( mm^2 \). The measured results for the receiver are summarized in Table III.L.4.
This chapter presented a new design approach for designing the front-end of an MB-OFDM UWB system. The main challenge for a UWB receiver was the design of a wideband front-end which operates between 3.1 GHz - 8 GHz. At the IF output, the UWB signal is at a fixed IF of 2.64 GHz, with a bandwidth of 528 MHz.

This UWB signal needs to be downconverted to DC for baseband processing. Since there is already about 20 dB of gain in the front-end, along with some channel select filtering, the specifications for the IF downconverter are greatly relaxed. The design of this receiver is, therefore, fairly straightforward. In fact, most Bluetooth receiver
Table III.5: Summary of measured receiver performance.

<table>
<thead>
<tr>
<th>Band no</th>
<th>Freq (MHz)</th>
<th>Gain (dB)</th>
<th>NF (dB)</th>
<th>$IIP_3$ (dBm)</th>
<th>BPF Rej (dB)</th>
<th>Im Rej (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>3432</td>
<td>15</td>
<td>5.8</td>
<td>-2.6</td>
<td>10.4</td>
<td>6.8</td>
</tr>
<tr>
<td>2</td>
<td>3960</td>
<td>18</td>
<td>5.2</td>
<td>-4</td>
<td>10</td>
<td>16.2</td>
</tr>
<tr>
<td>3</td>
<td>4488</td>
<td>15.9</td>
<td>6</td>
<td>-3</td>
<td>13.5</td>
<td>25.4</td>
</tr>
<tr>
<td>4</td>
<td>5016</td>
<td>18.9</td>
<td>6.38</td>
<td>-4.4</td>
<td>13</td>
<td>36</td>
</tr>
<tr>
<td>5</td>
<td>5544</td>
<td>21</td>
<td>6.28</td>
<td>-4.7</td>
<td>9.8</td>
<td>37</td>
</tr>
<tr>
<td>6</td>
<td>6072</td>
<td>20.2</td>
<td>6.57</td>
<td>-4.4</td>
<td>7.9</td>
<td>36</td>
</tr>
<tr>
<td>7</td>
<td>6600</td>
<td>20.3</td>
<td>5.46</td>
<td>-4.7</td>
<td>7.5</td>
<td>27</td>
</tr>
<tr>
<td>8</td>
<td>7128</td>
<td>21</td>
<td>5</td>
<td>-5.2</td>
<td>9.1</td>
<td>22</td>
</tr>
<tr>
<td>9</td>
<td>7656</td>
<td>20</td>
<td>5.3</td>
<td>-5.6</td>
<td>7.5</td>
<td>19</td>
</tr>
</tbody>
</table>

front-ends, which operate at 2.4 GHz could easily be re-designed with some modifications to act as IF downconverters for this UWB system.

In this chapter, we presented the specifications for an accompanying IF downconverter which would operate in unison with the RF front-end designed in Chapter III to create a complete UWB receiver.

**III.M.1 IF Downconverter Architecture**

The IF downconverter needs to provide three main functions

1. IQ Downconversion to baseband: The 2.64 GHz output from the front-end needs to be downconverted to provide an in-phase and quadrature output for complex baseband processing.

2. Variable Gain: Since the received signal at the antenna varies in power, the IF downconverter needs to provide some gain control, to control the baseband output amplitude.

3. Channel Select Filtering: To maintain baseband SNR, the adjacent channels from other UWB transmitters need to be suppressed significantly.

The usual design technique is to have one block in the IF chain to provide each of these functions. The most straightforward architecture to accomplish this is shown in Fig. III.30.
### III.M.2 IF Downconverter Gain

The minimum received signal from an MB-OFDM transmitter at the antenna input is -80.3 dBm. From Chapter III, the lowest gain in the front-end is 15 dB. Therefore, at the IF output, the received signal is at -64.3 dBm.

To calculate the gain for IF downconverter, we need to assume a maximum peak input level for the baseband A/D converter. A maximum input voltage of 0.5V is a fairly common number, and, therefore, we will assume that the A/D converter requires a 1V input signal.

Since the output of the BPF filter in Chapter III requires a 250 Ω impedance match, we will assume that the input impedance of this IF downconverter is 250 Ω.

\[
P_{IF,IN} = -65.3 \text{ dBm} \\
V_{IF,IN} = 0.38 \text{ mV} \\
V_{BB,OUT} = 500 \text{ mV} \\
Gain_{MAX} = 62.3 \text{ dB}
\]

(III.6)
The maximum receiver UWB signal power from a transmitter 0.1m away is -35 dBm. To maintain the same baseband output voltage of 0.5 V, therefore

\[ P_{IF,IN} = -35 \text{ dBm} \]
\[ V_{IF,IN} = 12.6 \text{ mV} \]
\[ V_{BB,OUT} = 500 \text{ mV} \]
\[ Gain_{MIN} = 32 \text{ dB} \] \hspace{1cm} (III.7)

The resulting gain requirement for the IF downconverter are shown in Table III.6.

Table III.6: IF downconverter gain requirements.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Gain, Maximum</td>
<td>62.3 dB</td>
</tr>
<tr>
<td>Gain, Minimum</td>
<td>32 dB</td>
</tr>
</tbody>
</table>

### III.M.3 IF Downconverter Noise Figure

The required system Noise Figure is different for different operating modes. For Mode 1 system (Band Group-1 only), the required Noise Figure is 6.6 dB. For higher Band Groups, the required Noise Figure is 8.0 dB. Then, using the measured receiver gain and Noise Figure from Chapter III, the Noise Figure specifications for the IF downconverter are shown in Table III.M.3.

Table III.7: Noise Figure specifications for IF downconverter

<table>
<thead>
<tr>
<th>Mode</th>
<th>Reqd IF NF (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Group-1</td>
<td>13.7</td>
</tr>
<tr>
<td>Higher Groups</td>
<td>21.8</td>
</tr>
</tbody>
</table>

### III.M.4 IF Downconverter Linearity

The IF downconverter linearity linearity requirements are determined by the following two factors:
Table III.8: IF downconverter linearity requirements.

<table>
<thead>
<tr>
<th>Specification</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Input 1dB Compression Point</td>
<td>-13.3 dBm</td>
</tr>
<tr>
<td>$IIP_3$</td>
<td>-8 dBm</td>
</tr>
</tbody>
</table>

1. Gain compression due to receiver overloading

2. Intermodulation distortion to maintain SNR

It was mentioned in Section III.J.1 that the on-chip bandpass filter significantly mitigates the linearity requirement due to both of these factors. Following the discussion in Section III.J.1, the linearity requirements for the IF downconverter are summarized in Table III.M.4.

III.M.5 IF Downconverter LPF Rejection

Although the on-chip BPF provides some adjacent rejection, additional rejection is required from a baseband LPF to maintain the average $\frac{E_b}{N_0}$.

III.N Conclusion

This chapter presented the design of an Ultra Wideband receiver front-end for MB-OFDM based systems. While the current standard calls for a mandatory operation mode covering the frequency bands from 3.1 GHz - 4.6 GHz, this receiver was designed to cover all the bands up to 8 GHz. This ensures compatibility of the receiver with future UWB systems. This additional band coverage was provided with negligible additional cost to the system. Two capacitors and four MOS switches were the only extra components required to ensure coverage of all these bands.

Traditional receiver design techniques were analyzed to study their viability for UWB systems. Arguments were presented which proved that the traditional approach to low-power low-cost receiver design, which utilizes a direct-conversion architecture is very challenging for UWB systems. Challenges specific to UWB receivers,
such as frequency varying DC offset and second-order intermodulation at RF are difficult to overcome in a direct conversion receiver.

It was also shown that, while heterodyne receiver consume larger die area and power, they are immune to a lot of problems that occur in a direct-conversion receiver. A new design approach was presented which provides all the benefits of a heterodyne receiver, while consuming power and die area comparable to direct conversion receivers.

A heterodyne receiver was designed in a 0.18\(\mu\)m CMOS process, which largely avoids all the issues that plague a heterodyne receiver. All the components of this receiver were integrated on-chip, and the receiver does not need any off-chip matching components. An on-chip bandpass filter was also provided for channel selection.

Distortion analysis, developed in Chapter II was used to specify receiver second-order and third-order distortion in a real world interference environment. This ensures robust operation of the receiver in dense wireless networks.

No inductors were used in this design. An inductor-less design can significantly reduce the cost of the system, as the die area is very small. This receiver front-end occupies only 0.35 mm\(^2\), which is smaller than any UWB receiver presented in the literature to date. Additionally, an inductor-less design avoids the use of any thick top metal, and therefore, this receiver can be integrated along with a baseband processor, for a single-chip UWB solution.

This chapter has been published in part in the following publications:


Conclusion

Wireless Ultra Wideband systems are expected to alleviate some very significant issues facing the consumer electronics arena. Connecting high speed devices with cables is becoming cumbersome and more and more expensive. For example, a standard USB 2.0 cable, used to connect multimedia devices, such as the Ipod, to a computer, regularly cost in excess of about $15. A standard Digital Video Interface (DVI) cable used to connect a high definition DVD player with a high definition TV costs close to $50. In an era where consumer electronics devices are becoming cheaper every day, it is inevitable that consumers will look for a cheaper and easier to use connectivity solution.

Ultra Wideband systems, if implement properly, have a tremendous potential to satisfy the need for cheap, high speed, short range wireless networks. This dissertation has focussed on two main issues:

1. Design of Wireless Ultra Wideband systems that would be robust to interferers

2. Design of inexpensive and low-power Wireless Ultra Wideband systems

The two issues mentioned above are probably the most important issues in designing any successful wireless technology.

Traditional interference analysis, such as a two-tone analysis, is geared towards analyzing narrowband systems. Using an inaccurate analysis framework, such as two-tone analysis, could lead to the highly undesired scenarios of either a heavily
overdesigned receiver, which consumes too much power and die area, or an underdesigned receiver, which fails to perform in a dense interference scenario.

A comprehensive wideband analysis of receiver nonlinearities and various interference scenarios was required to predict system performance in real world scenarios. This thesis has presented a new analytical framework for circuit and system designers to analyze the performance of a nonlinear UWB receiver, in the presence of various interferers. The following interference scenarios, which cover most of the undesired interferes, were analyzed:

1. Narrow out-of-band interferer, such as Bluetooth, 802.11 b/g, Microwave ovens
2. Narrow in-band interferer, such as 802.11a, marine radar
3. Narrow in-band interferer and a UWB in-band interferer
4. Multiple UWB interferers

Where needed, analytical expressions, using rigorous wideband analysis of modulated signal, was performed. Closed-form expressions were developed to calculate the PSD of the distortion products arising from receiver nonlinearities and the above mentioned interferers. The statistical model of a QPSK modulated MB-OFDM signal was used for all of the analyses. The receiver was modeled as a third-order nonlinearity, represented by a power-series. Although the power series expansion may consist of higher order terms, it was shown that, for the expected signal power levels, a third-order power series expansion models the receiver quite accurately.

Although narrow in-band interferers, such as 802.11a, are expected to have very little impact on UWB OFDM systems, it was shown that, in the presence of another UWB interferer and a third-order nonlinearity in the receiver, their impact on system performance can be quite detrimental, due to cross-modulation distortion. Neglecting cross-modulation distortion while specifying the system can result in a severely underperforming system.
While second-order distortion in a traditional receiver is usually considered for calculation of DC offset only, it was shown, in UWB systems, second-order distortion can reduce the SNR of the desired signal, due to either mixing of undesired jammers, or due to the generation of second-harmonic of one undesired jammer. Analytical expressions were developed to calculate the PSD of each of both second-order intermodulation and second harmonic distortion products. While the $IIP_2$ specification arrive at with this analysis might still be lower that required by DC offset specification, in a heterodyne receiver, where DC offset is rarely specified, this may prove to be an important specification.

Design of UWB receivers presents an interesting challenge. Designing a UWB receiver is both expensive due to large die area, and also power hungry. For infiltration into consumer electronics, a wireless UWB system needs to be both inexpensive and power efficient. The two requirements are usually in conflict with each other.

The second goal of this dissertation was to try and overcome the paradox of a low-cost low-power UWB receiver. While the industry usually focuses on traditional, low-risk design techniques to rein in development cost and time-to-market, it is often up to the academia to investigate and suggest alternate approaches, which, in the longer term, might prove to be a lot more power efficient and cost-effective.

That is the approach this dissertation has taken to suggest a second-look at the often neglected technology of a heterodyne receiver. Suggestion of a heterodyne receiver for a low-cost low-power often, and rightly so, raises the eyebrows of circuit and system designers. This dissertation has tried to conquer the usual suspects that raise the cost and power consumption of a heterodyne receiver.

1. Image Rejection: This is one of the biggest issues with heterodyne receivers. It raises power consumption by employing two mixers. This was solved by employing an architectural approach to image rejection. Instead of employing an image reject mixer, the frequency plan was chosen such that a standard UWB BPF provides significant image rejection. The LNA was also designed with a dynamic frequency response, which provided additional image rejection.
2. Need for an interstage-filter: An interstage filter is needed to alleviate the $IIP_3$ and Noise Figure requirements of the second downconversion stage. This interstage filter is usually a high frequency SAW filter, which is expensive. On-chip filters consume a lot of power and increase the die area. The design presented in this dissertation countered this issue by designing an on-chip passive bandpass filter that consumes very little die area. Inductors are usually required in an on-chip BPF which consume a lot of die area. The performance of these on-chip passive filters is usually not very good, due to the low Q of on-chip inductors.

The design presented here uses bondwires, which are available for a very small cost in any packaging process, as inductors in the bandpass filters. The Q of these bondwires is usually greater than 50, which can be used to design on-chip bandpass filters with very good performance. The one issue with using bondwires as inductors is that, while on-chip inductors vary very little with process, bondwire lengths can vary by as much as $\pm 10\%$. This variation bondwire length, coupled with the variation in on-chip capacitance, can change the center frequency of the BPF by as much as $\pm 15\%$. This issue was addressed by providing a tuning mechanism in the filter, with on-chip MOS varactors, that can tune the center frequency by $\pm 15\%$.

One of the biggest challenges in a UWB receiver, is the design of the front-end LNA. Most designs presented in the literature consume either too much die area (using multiple LC sections) or too much power (resistive LNAs) to achieve the desired performance. A novel approach to UWB LNA design was presented in this dissertation. The proposed design approach exploits the time-frequency interleaving of MB-OFDM systems to dynamically tune the center frequency of the LNA. On-chip inductors were avoided by

1. Using package bond-wires to create the load resonant tank
2. Using RC feedback for input matching

The receiver was designed in a $0.18\mu M$ CMOS process. The receiver uses
no on-chip inductors, or any other specialized RF devices, and is therefore a very good candidate for integrating the RF chip with the baseband chip.

The measured receiver performance matches the performance and power consumption numbers of other published UWB receivers, but this receiver is the smallest of all reported UWB receiver front-ends, consuming only 0.35 mm$^2$ of die area in a 0.18 $\mu M$ CMOS process.

This dissertation has presented

1. The first true wideband nonlinearity analysis of OFDM systems, which will assist circuit and system designers in optimizing the power-consumption vs performance trade-off.

2. The first inductor-less UWB receiver front-end. The receiver front-end is the smallest of all reported receivers, with comparable power consumption and performance.
Bibliography


