IEEE 802.15.4a – Understanding the Protocol and Reducing Multi-User Interference

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IEEE 802.15.4a – Understanding the protocol and Reducing Multi-User Interference.

by

TAHA HARNESSWALLA

M.S., University of Colorado, 2013

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Prof. Dr. Douglas Sicker
Prof. Jose Ramon Santos

A thesis submitted to the
Faculty of the Graduate School of the
University of Colorado in partial fulfillment
of the requirement for the degree of
Masters of Science
Interdisciplinary Telecommunication Program
2013
This thesis entitled:
IEEE 802.15.4a – Understanding the protocol and Reducing Multi-User Interference
written by Taha Harnesswalla
has been approved for the Interdisciplinary Telecommunication Program

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4/09/2013
Date ________________

The final copy of this thesis has been examined by the signatories, and we
Find that both the content and the form meet acceptable presentation standards
Of scholarly work in the above mentioned discipline.
Abstract

Experimental results from various studies have proved that the 2.4 GHz ISM band has become increasingly congested. Widespread use of Wi-Fi (802.11) along with incumbent technologies already utilizing this band have made it difficult to provide other wireless services within the same band without an increase in bit error rate and a drop in throughput. The 802.15.4 protocol which has been defined for low data rate networks is currently also being deployed in the 2.4 GHz band.

This paper suggests the use of Ultra Wide Band (UWB) as a feasible alternative to deploying a Zigbee based Home Area Network (HAN) on the 2.4 GHz spectrum using Direct Spread Spectrum (DSSS). The FCC has approved the 802.15.4a standard for an UWB PHY layer in a sensor based networks. This study gives an analysis of the standard. The performance evaluation for this standard leads our focus to multi-user interference effects on preamble detection which affects synchronization and timing acquisition in a non-coherent energy detector receiver. The paper also considers two approaches that can be taken up to overcome multi-user interference so that, the inherent robustness of UWB is not overcome by inaccuracy in data reception introduced by interference.
Acknowledgements

This research project would not have been possible without the support of many people. I wish to express my gratitude to my supervisor, Prof. Dr. Frank Barnes who was abundantly helpful and offered invaluable assistance, support and guidance. Deepest gratitude also due to the members of the supervisory committee, Prof. Dr. Douglas Sicker and Prof. Jose Santos without whose knowledge and assistance this study would not have been successful.

Special thanks to Prof. Dr. Jim Lansford, for taking interest in my work and providing it much needed structure and direction. His feedback and expert insight gave me the confidence to keep going. Not forgetting Karan Shrivastava, Aamir Shahpurwalla, Noopur Bodke and Ritika Rahate for encouraging and motivating me through every step. I also wish to express my love and gratitude to my beloved parents; Iqbal and Durriya Harnesswalla alongwith my siblings Huzefa and Fatema Harnesswalla for their understanding & endless love, through the duration of my studies.
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CHAPTER 1

Introduction

1.1 Motivation

The motivation for writing this paper is to understand how an Ultra Wideband (UWB) physical layer can be used to implement a Home Area Network (HAN) and the challenges involved. The development of the 802.15.4a UWB physical layer as an option for the traditional Zigbee 802.15.4 PHY layer seems to have opened various avenues that will help overcome barriers such as congestion and cost. UWB will give the significant advantage of robust multipath performance, as a result immunity to the effects of interference, especially in a congested urban environment. Study by Flury, M. et. al. (2007) has shown, that even though theoretically UWB does have superior multiuser characteristics, a practical implementation using non-coherent energy detectors does not give an ideal performance. The study thus, puts impetus on finding a means to overcome these interference based performance issues. As a result, this thesis will give an insight on the performance of the IEEE 802.15.4a standard, followed by a look at methods that could help decrease the packet error introduced in an 802.15.4a compliant receiver due to interference.

1.2 Background

The concept of the HAN has been evolving over the years. Currently, Zigbee communications based on the 802.15.4 protocol form the basis of the implementation of the Home Area Network.
The Zigbee protocol itself has evolved out of convenience more than anything else. A collection of major corporations—the most significant eight being Ember, Freescale, Honeywell, Invensys, Mitsubishi, Motorola, Philips, and Samsung—are all committed to standardizing cost-effective, low-power and wirelessly networked, monitoring and control products based on an open global standard (Howlander, Kiger, Ewing, 2007) resulting in the formation of the Zigbee Alliance. As a result, these companies were looking for a protocol that does not use a large amount of bandwidth and is not very complex, because higher bandwidth and complexity leads to increased cost and power consumption. The Zigbee protocol based on the 2.4 GHz PHY layer has become popular over the years and is now, possibly the most preferred option for the HAN environment.

1.3 Goal of the Paper

The main goal of this thesis is to identify and analyze the features of a low power and low data rate UWB physical layer. The significant contribution of the thesis will be in understanding the advantages the implementation of the alternate PHY layer holds in comparison to the traditional Zigbee PHY layer. This study will consider 2 approaches to timing acquisition for a non-coherent UWB energy detector receiver which will be used in while implementing that alternate PHY Layer and will fill the gap in determining various means to overcome multi-user interference for the IEEE 802.15.4a UWB standard. It will highlight the drawbacks with the current standard that are proving to be a hindrance to a commercial implementation of this alternate UWB PHY layer for sensor networks.
1.4 Related Research

Time Hopping Impulse Radio (TH-IR) is the UWB PHY layer used for the IEEE 802.15.4a standard. There is plenty of literature on TH–IR. The 802.15.4a standard was developed in 2007 (IEEE P802.15 Working Group, 2007) as an amendment to the 802.15.4 protocol developed in 2006. López-Salcedo (2007) gives a brief overview on the methods that differentiate coherent and non-coherent UWB approaches. Vitavasiri (2007) is an excellent end to end implementation of a non-coherent UWB scheme even though is does not consider a TH-IR PHY layer. The paper by Fischer, Adalan, Scholtz, and Mecklenbrauker (2009) provides a schematic for the entire 802.15.4a transmission channel. This research provides the ideal building block for further investigation on this protocol.
CHAPTER 2

Drawbacks with current protocol

The 802.15.4 Wireless Personal Area Network (WPAN) protocol was approved by the IEEE in early 2006. It is considered as the basis of the physical layer on which the HANs will operate. The IEEE 802.15.4 physical layer uses three frequency bands in the 868, 915 MHz and 2.4GHz bands. The 2.4GHz band is the unlicensed ISM band available worldwide. This band offers 16 channels and a raw transmission data rate of 250kbps using offset quadrature phase shift keying (O-QPSK) as the modulation scheme.

The salient features of the current Zigbee protocol are listed as follows (Chen, 2007):

- Star and peer-to-peer topologies
- Optional Frame Structure
- Association between Nodes
- CSMA-CA channel access mechanism
- Packet validation and message rejection
- Optional guaranteed time slots
- Guaranteed packet delivery
- Facilitates low-power operation

The popular architecture for a home area network consists majorly of 2 kinds of devices. The full-functional device (FFD) is a transceiver and the hub of the network. The network will consist of a single or maybe a few FFDs. These devices have the ability to aggregate data from several reduced-functional devices (RFD) or end nodes and forward this data to a central
aggregator. An FFD is more immune to pricing, spatial and power consumption constraints, since these will be fewer in number and are expected to be connected to the central power source. The FFD can behave as the PAN coordinator and as a result performs initiation and termination of route communication and allocation of 16 bit addresses to devices (Coates and Rabbat, 2005).

An RFD on the other hand, is expected to be extremely low cost. Being battery powered an RFD will be far more sensitive to power consumption as well. It will communicate only with an FFD. A rule of thumb for the cost of an RFD is that it should cost about 1% of the device it is placed on. It should also not be too large so it can be placed strategically to gather the required data.

Recent research has shown that there might be several constraints that we might come across while using the 2.4 GHz band, as our band of choice. This band is popular among several existing technologies like Wi-Fi and Bluetooth. Being a spread spectrum technology, Zigbee will not be influenced much by the presence of Bluetooth. Wi-Fi, on the other hand, has increasingly become the preferred choice when it comes to Wireless Local Area Networks (WLANs), and will be a continuous interference to the smooth functioning of a parallel Zigbee network and vice versa. Also, both these networks will not only, be overlapping each other in the same frequency band, but also spatially, since they ought to be the choice for WLAN and HAN PHY layer in the same house, serving crucial roles as part of modern day urban wireless technology infrastructure. Currently, Zigbee has an application specific range of about 10m – 70m (Howlander et. al. 2007). Wi-Fi on the other being a higher power protocol has a range that is larger and offers superior data rates.
The following results were obtained when the interference characteristics of Zigbee with Wi-Fi were observed:

1. The different data rates of IEEE 802.11b/g have different impact on the link performance of IEEE 802.15.4 (Chen, 2007).

2. Distance between a ZigBee node and a Wi-Fi interferer should be a minimum of 3m (Yang & Yu, 2008).

3. For Zigbee to operate without any loss due to interference from Wi-Fi, the frequency offset between channel center frequencies should be at least 12 MHz (Howlander et. al. 2007).

As a result we realize that reliable Zigbee communication in the presence of Wi-Fi will always be a hindrance due to the overlapping bands. Figure 1, illustrates the Zigbee channels that fulfill the above requirements.

![Figure 1: ISM Band Channel Layout](https://example.com/f1.png)

Source: Digi-Key Wireless Solutions

When dealing with a fully operational Wi-Fi spectrum, successful Zigbee transmissions with no interference are available in primarily 4 channels. This significantly reduces the overall capacity offered by Zigbee in this spectrum. Thus, if we are to consider moving to higher physical layer frequency ranges we stand a chance of utilizing the inherent nature of these frequencies to achieve a low rate, low power and a low cost wireless sensor network. In a frequency range that is sparsely populated and by using a protocol more robust to interference, the HAN will be a more reliable.
The report and white paper by Thonet, Allard-Jacquin and Colle (2008) for Schneider Electric in France is an in-depth study on the difficulties that might be encountered while dealing with coexistence issues between Wi-Fi and Zigbee. They have conducted tests in a variety of test environments, but have not considered the interference effects produced by multiple Wi-Fi networks on a single Zigbee network as will be the case in a dense urban environment. The results obtained by Mao et. al. though, indicate that in the 2.4 GHz spectrum interference between the two standards will adversely affect the throughput on both transmissions. As a result the IEEE has considered various PHY layer options. These options are available for the 2.4 GHz ISM band as well as in higher, less congested frequency ranges.
CHAPTER 3

Ultra Wide Band (UWB) – A feasible option

Ultra Wideband is a wireless technology that works in higher range frequencies. It requires a large bandwidth in the range of about 500 MHz. In the US, the frequency range from 3.1 GHz to 10.6 GHz has been allocated for the possible deployment of UWB systems. The following will be the main features that a UWB system offers over normal radio communication technologies (Radio Communication Study Groups, 2003).

1) Low Susceptibility to Multipath Fading: Multipath within and around buildings causes’ significant deterioration in the performance of conventional communication systems. The short duration of UWB pulses makes them less sensitive to multipath effect compared to narrowband signals. This is because with transmission of pulses shorter than a nanosecond in duration, the reflected pulse has an extremely short window of opportunity to collide with the line of sight (LOS) pulse to cause signal degradation.

2) Ability to work with low signal-to-noise ratios: UWB wireless communication systems exhibit excellent immunity to a high noise environment. It can easily co-exist with other narrowband and wideband services and noise is generally overcome by the inherent robustness offered by and UWB PHY layer.

3) Secure Communications: Because UWB signals can be made noise-like, communication using transmitter/receiver pairs with a unique timing code at millions of bits per second, have such low energy and spectral density below the noise floor of conventional receivers, and occupy such a wide bandwidth, they are more covert and potentially harder to detect than conventional radio. These advantageous characteristics result in UWB
transmissions with a very low extant RF signature, providing intrinsically secure transmissions with low probability of detection (LPD) and low probability of interception (LPI).

4) System Simplicity: The relative simplicity of the UWB impulse radio architecture compared to the super heterodyne architecture transceiver is due to the fact that the UWB transceiver has no phase-lock loop synthesizer, voltage-controlled oscillator (VCO), mixer or power amplifier, which translates to lower material and assembly costs.

The reduction of required fade margins and power spectral density of the UWB spectrum makes UWB technology a favorable technology for low power and short-range communications, as required for operation in indoor environments. A very important advantage of UWB radios is their huge processing gain, a measure of a radio's resistance to jamming (Radio Communication Study Groups, 2003) but, this technology suffers from drawbacks that might need to be overcome during implementation. Adding to this, UWB is implemented at extremely high frequencies and as a result will not offer the same range and penetration offered by other PHY Layer technologies in the 2.4 GHz ISM band.

3.1 Ultra Wide Band Signals and Information Carriers

Traditionally, harmonic oscillations have been the primary carrier of information signals. UWB does not necessarily use harmonic oscillations as the sole information carrier. Methods described below stress on the varied means used for modulation when dealing with UWB signals.

Firstly, a signal with center frequency $F_c$ and bandwidth $B$ can be considered an UWB signal if its fractional bandwidth, $\Delta = \frac{\text{Bandwidth}}{\text{Center Frequency}} > 0.2 - 0.25$. Based on the regulations of the US Federal Communications Commission (FCC) which permitted unlicensed usage of UWB
signals in wireless-communications devices, signals with a bandwidth greater than 500 MHz in the frequency band 3.1–10.6 GHz can be treated as UWB signals.

Initially, ultra-short pulses that resulted in the broad spreading of power in the frequency domain were considered UWB, but nowadays there are multiple variations to this option, that accommodate for modulation, coding and multiple access (MCM) in unique ways that help fulfill the unique and disparate requirements for both and FFD and an RFD within the network (Dmitriev et. al., 2008)

Today the list of Ultra Wide Band technologies is as follows:

1. **Ultra-short pulses** are the oldest and the most traditional means of producing an Ultra Wide Band signal. In general, the duration of the pulse depends on the frequency band used. A pulse of 2 ns results in the widest bandwidth of 500 MHz, which categorizes the signal as UWB. The relation of the pulse duration and the width of the power spectral density is rigid, and the power spectrum ranges from \( f = 0 \) to \( f = 1/T \) along the frequency axis.

2. **Short Radio pulse-oscillation trains.** According to this approach, a signal is formed in a given frequency band. Here too, the relation between the duration of the pulse and the power spectrum of the signal is rigid. Thus, to obtain a more uniform spectral density in the frequency band, the pulse envelope is bell shaped.

3. **Chaotic Radio pulses.** The overall shape of the power spectral density spectrum in this method is determined by the duration of the initial spectrum of the continuous chaotic radio pulses generated. In, this method there is no direct relation between the radio pulses and the eventual power spectral density envelope.
4. **Burst of short pulses.** Like in the case of a single short pulse the duration of the pulse is matched to envelope produced in the frequency domain. In this case though, we have multiple short pulses, as a result, the frequency envelope has a direct relation to the total number of pulses sent in a single burst.

5. **Direct sequence - UWB.** In this case to transmit 1 bit, that bit is divided into a number of chips and used to modulate a high frequency sine wave. This is the traditional approach on which technologies like Code Division Multiple Access (CDMA) are based. The power spectral density is directly related to the number of chips used to transmit a single bit.

There are other methods of spreading signals like Orthogonal Frequency Division Multiplexing (OFDM) and Voltage controlled frequency modulation that are also popular. OFDM is being used currently in various wireless protocols (Dmitriev et. al., 2008).

At this juncture though, it is useful to note that the idea of a mass application of UWB technology is very closely related to impulse UWB technology. Creation of a cheap and simple wireless communication standard is the main aim. Figure 2 helps illustrate the UWB carriers we have talked about earlier.
One of the main problems with UWB is the synchronization required between the transmitter and the receiver. For example, for efficient coherent reception, the receivers need to be synchronized to no worse than 10 ps for a pulse duration of 150 ps (Dmitriev et. al., 2008). In order to overcome this, complex circuitry with high power consumption would be needed. As a result non–coherent designs for UWB signals are an extremely attractive option as they comply with the low complexity and low cost criteria of a HAN node.
CHAPTER 4

Design Consideration for an UWB System

4.1 THE UWB Channel

The environment of propagation for the UWB signals has a huge impact on the design of the system. If an Additive White Gaussian Noise (AWGN) Channel is assumed, a simple energy detector serves the purpose of receiving UWB signals. Most studies conducted on UWB consider indoor Non Line of Sight (NLOS) environment and as a result an AWGN channel. We too will assume an AWGN channel in the rest of the paper.

An UWB signal has a large bandwidth, resulting in a fine delay resolution, which means large number of independently fading multipath components are available at the receiver. The advantage is that an UWB signal has a high degree of delay diversity, thus, small-scale fading fluctuations are potentially eliminated (Kumar et. al.). Conversely, a delay of 5-50 ns among UWB signal multipath components results in the Rake receiver component of the coherent receiver design requiring a large number of fingers to collect the available energy. This goes against the basic requirement of low cost and low complexity. A simple energy detector receiver is the most attractive option for the UWB receiver. Since we have considered the non-coherent reception option for UWB, another important consideration is identification of the first component or the leading edge of the signal. An inability to do this efficiently even in the presence of Multi User Interference, will not only hinder the ranging feature of the new PHY layer, but will also drastically reduce the robustness of this protocol in a noisy environment.
There are a couple of other reasons that help make a case for the UWB technology. We consider Shannon’s Capacity theorem for information theory. It is defined in Equation 1.

\[ C = W \log_2 \left( 1 + \frac{S}{N} \right) \]  

(1)

In Shannon’s theorem, the capacity is directly proportional to the bandwidth \( W \). Thus, as long as transmission rate \( R < C \) capacity, theoretically, error free transmissions can be obtained. Since UWB signals have such a large bandwidth to the order of 500 MHz, the overall capacity achievable on the channel is extremely high. For an UWB channel we will be transmitting in the GHz range reducing chip fabrication costs significantly. The antenna length will now be to the order of the dimensions of the chip and can thus, theoretically could be fabricated onto the chip itself. This theoretical concept has still not been implemented yet. Since the duty cycle of the transmission is very low, heat dissipation from the chip becomes less of a concern and so the heat dissipation requirements are lower. Thus, components on the chip now can be closer to each other, reducing size of the chip and as a result, reducing overall fabrication costs, and bringing its size within the spatial requirements of a HAN node. These are some of the less important advantages that add to the more significant advantages of an UWB based system listed earlier.

### 4.2 Global Regulation on UWB

The US Federal Communications Commission (FCC) defines UWB signals as signals having an absolute bandwidth greater than 500 MHz or a relative bandwidth larger than 20\%. The absolute bandwidth of a given UWB signal is calculated as the difference between the upper and lower frequencies at the -10 dB emission level. Spreading the signal over an ultra-wide
bandwidth results in the construction of a wireless system that interferes minimally with other wireless systems. As a result, several national frequency regulating bodies around the world have issued rulings that allow the unlicensed use of UWB spectrum, even if the UWB spectrum overlaps with spectrum assigned to existing systems.

In the US, the FCC allows emissions in between the 3.1 GHz and 10.6 GHz range, while setting a set of stringent rules in order to prevent harmful interference from UWB signals. Emission limits are given in terms of Equivalent Isotropically Radiated Power (EIRP) or the peak power density observed in the direction of maximum antenna gain. According to the FCC, the EIRP in any direction should not exceed -41.3 dBm/MHz which is approximately equivalent to the transmission by an unintentional emitter. This results in an average EIRP of -14.31 dBm for a 500 MHz channel.

Also considering that we are referring to TH-IR form of UWB, we are looking at a gated version of UWB transmission. A set of pulses of a certain very short time length are transmitted to achieve the desired UWB spreading of the signal. The FCC in 2006 also placed further limits on the peak transmission power. The paper by Corral, Emami and Rasor (2006) illustrates the peak transmission power limits imposed by the FCC. According to the paper for a gated channel like a TH-IR UWB, the $P_{\text{peak}} = P_{\text{ave}} + W$, where W is the peak-to-average power ratio in the peak resolution bandwidth. Thus, in a signal like a TH-IR UWB which is temporal in nature, not a continuous transmission, the peak power of the pulse will also have to be maintained within a certain limit. This will ensure the peak power transmission is also within FCC regulations.

FCC indoor and outdoor limits differ over some frequencies (Zhang, Orlik, Sahinoglu, Molisch, Kinny 2009). Figure 3 illustrates the limits placed by the FCC on indoor and outdoor emission levels.
4.3 Multiple Access Considerations

In a wireless network, the method in which the transmission medium is shared amongst devices is called a multiple access technique. Since wireless communications are inherently broadcast in nature, all the devices on a wireless communication channel must have a protocol in place to share the medium. Thus, a major goal of the MAC layer is to minimize interference within the network. There are several well-known methods by which wireless devices can share a channel.

Network Topologies: Constructing a network topology that supports multiple users transmitting on the channel simultaneously, helps with the multiple user interference characteristics of the network. For a given HAN implemented with the current Zigbee standards at our disposal we talk mainly about 2 kinds of nodes; an FFD and an RFD. The FFD contains the software that enables initiation, network formation, and control of the wireless channel for multiple access.
among the RFDs. An FFD is commonly referred to as a coordinator due to its ability to perform the above mentioned activities (Zhang et. al. 2009).

A simple star topology for a sensor network has the FFD as the center of the configuration while the RFDs surround the coordinator. The RFD devices are logically associated with the coordinator. A tree network can be viewed as an amalgamation of star networks, where the star networks are connected to each other by connecting the respective FFDs (Zhang et. al. 2009). A third topology ideally suited for sensor networks is a mesh topology. The mesh topology provides redundancy to the network, so in the event of device or link failure data can still be routed. While considering techniques for multiple access, we have to understand that the topology will play a significant role in determining the necessary requirements.

A simple topology leads to simple multiple access designs. Fewer devices accessing the channel simultaneously results in a lower probability of interference. Also, simple technologies offer the option to control access at a central point. Complex technologies require more careful planning of the channel access, but they also allow coverage of larger areas by a single network even with severe constraints on transmit power, similar to those in UWB networks (Zhang et. al. 2009).

Typically centralized schemes offer better efficiency and reliability since collisions can be more easily avoided. The coordinator device though is increased in complexity and requires network-wide information and is a single point of failure on the network. Decentralized systems are less complex, but have a higher probability of causing harmful interference. Use of handshaking is also another method to enable multiple users to access the channel without a high level of interference (Zhang et. al. 2009). For a low cost sensor network, a decentralized system fits the bill.
**Medium Sharing:** Time Division and Frequency Division are other means of allowing multiple access when several devices share the same wireless infrastructure. Sharing the total transmission time amongst the transmitters is called Time Division. Sharing the available frequency is called frequency division. None of these methods apply to the 802.15.4a standard since the inherent method of transmission is Aloha i.e. random back off and transmit. Thus, the using the right network topology will lead to optimum results.

The aim of a HAN is to achieve coverage where needed. Every device desired to be on the network should be part of the HAN. Generally UWB offers a range of 10’s of meters and is adversely affected by any physical barrier. Thus, designing custom networks which are an amalgamation of standard RFD, repeaters and limited and necessary FFD components to obtain the required coverage would be the main aim. These networks can be designed to compensate for changes in physical layout and could have variable transmission power options on the RFD to increase or decrease range as needed within the UWB limits.

Keeping the above characteristics in mind for an UWB based PHY layer the IEEE approved an alternate PHY layer for Zigbee in 2007 naming it 802.15.4a, as an alternate standard to the already approved 802.15.4. The following chapters will go through the basics of this protocol and evaluate constraints in communication produced Multi User Interference (MUI) at the receiver.
CHAPTER 5

UWB PHY Layer – IEEE 802.15.4a Standard

5.1 Modulation and Spreading

Various modulating and spreading options can be considered for the UWB channel. The main technologies that could be applied to an UWB channel are Frequency Hopping (FH), Orthogonal frequency-division multiplexing (OFDM), Direct-Sequence Spread Spectrum (DS-SS) and Time-Hopping Impulse Radio (TH-IR). Time Hopping impulse radio is based on the principle that each data symbol is represented by a sequence of pseudorandom delays assuming pulse position modulation (PPM). A strong duality exists between FH and TH-IR. FH hops sequentially in frequency domain, while TH-IR hops in the time domain. For a low rate data transmission neither OFDM nor DS-SS is suitable. They require high rate sampling, analog to digital conversion, processing at a high rate resulting in both high complexity and large energy consumption. This would go against the core requirements for a HAN node. FH and TH-IR offer a better performance to complexity tradeoff. FH would create large interference to legacy systems and is prohibited in several domains by regulatory bodies worldwide. TH-IR too has it’s set of limitations in terms of delay spread that result in a limit to data rate. There is also a cap on peak power by the FCC. As a result TH-IR is the preferred option for UWB PHY layer (Zhang et. al., 2009).
5.2 PHY LAYER

The TH-IR PHY layer option is designed to provide robust performance and communication over extended distances. The wide bandwidth characteristics are meant to overcome harsh multipath and inference conditions. Concatenated forward error correction is provided using a Reed Solomon coding for a more flexible performance. Variance in data rates available to cope with a variety of channel conditions. These features make this PHY layer an extremely attractive alternative. The HAN would operate without the concerns we might face in the near future with respect to interference and information security.

The 802.15.4a PHY layer provides for a mandatory data rate of 850 kb/s. Data rates in the variety of 110 kb/s, 1.70 Mb/s, 6.81 Mb/s and 27.24 Mb/s can also be achieved. The UWB PHY design enables heterogeneous networking. The network would have at least one FFD. For an FFD, which is less cost sensitive, high processing complexity and power consumption is generally not a hurdle. The HAN nodes on the other end of the communication, which are attached to various devices within the home have extremely stringent limits on complexity, size and energy consumption. The 802.15.4a PHY layer has been designed with modulation, coding and multiple access schemes that allow both FFDs and RFDs to achieve optimum performance. Having mentioned earlier, that a non-coherent reception scheme is preferred to maintain power consumption and cost within its desirable range, the protocol allows FFDs to employ coherent or non-coherent reception while RFDs use simple energy detectors. The remainder of this section describes in greater detail some specific features and designs for the IEEE 802.15.4a standard (IEEE Std. 802.15.4a, 2007).
5.2.1 Band Plan

The first step to developing the UWB standard is selecting the frequency and bandwidth of the signal. The total transmit power is a function of the signal bandwidth. Thus, an increase in the signal bandwidth allows a higher transmit power. For non-coherent receivers though, the bandwidth preferably should be less than the inverse of the channel delay spread. With a larger bandwidth the receiver cannot optimally combine the resolved multipath components. For coherent receivers the trade off is generally between the delay spread and the amount of signal energy that can be collected with a given number of Rake fingers (Zhang et. al. 2009). Cost requirements tend to restrict bandwidth to as low as possible. Keeping these considerations in mind the IEEE 802.15.4a decided a mandatory signal bandwidth of 500 MHz with optional bandwidths greater than 1 GHz.
Table 1 denotes the center frequency and bandwidths of the defined bands.

<table>
<thead>
<tr>
<th>Channel Number</th>
<th>Center Frequency</th>
<th>Bandwidth</th>
<th>Channel Type</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>499.2</td>
<td>499.2</td>
<td>Sub - gigahertz</td>
</tr>
<tr>
<td>1</td>
<td>3494.4</td>
<td>499.2</td>
<td>Low Band</td>
</tr>
<tr>
<td>2</td>
<td>3993.6</td>
<td>499.2</td>
<td>Low Band</td>
</tr>
<tr>
<td>3</td>
<td>4492.8</td>
<td>499.2</td>
<td>Low Band / Mandatory</td>
</tr>
<tr>
<td>4</td>
<td>3993.6</td>
<td>1331.1</td>
<td>Low Band</td>
</tr>
<tr>
<td>5</td>
<td>6489.6</td>
<td>499.2</td>
<td>High Band</td>
</tr>
<tr>
<td>6</td>
<td>6988.8</td>
<td>499.2</td>
<td>High Band</td>
</tr>
<tr>
<td>7</td>
<td>6489.6</td>
<td>1081.6</td>
<td>High Band</td>
</tr>
<tr>
<td>8</td>
<td>7488.0</td>
<td>499.2</td>
<td>High Band</td>
</tr>
<tr>
<td>9</td>
<td>7987.2</td>
<td>499.2</td>
<td>High Band</td>
</tr>
<tr>
<td>10</td>
<td>8486.4</td>
<td>499.2</td>
<td>High Band</td>
</tr>
<tr>
<td>11</td>
<td>7987.2</td>
<td>1331.2</td>
<td>High Band / Mandatory</td>
</tr>
<tr>
<td>12</td>
<td>8985.6</td>
<td>499.2</td>
<td>High Band</td>
</tr>
<tr>
<td>13</td>
<td>9484.8</td>
<td>499.2</td>
<td>High Band</td>
</tr>
<tr>
<td>14</td>
<td>9984.0</td>
<td>499.2</td>
<td>High Band</td>
</tr>
<tr>
<td>15</td>
<td>9484.8</td>
<td>1354.9</td>
<td>High Band</td>
</tr>
</tbody>
</table>

*Table 1: IEEE 802.15.4a Frequency Bands*
Figure 4 depicts the frequency band plan for 802.15.4a.

The center frequencies are chosen such that they can be derived from readily available crystal oscillators. Also to avoid congestion, the 5 GHz ISM band is overlooked. Thus, this IEEE standard will operate without contributing to congestion and interference to legacy systems on the ISM band or other bands. This gives an ideal, reliable and readily available spectrum option for the HAN networks without having to worry about the presence of other narrowband technologies in the free spectrum. The challenge that lies ahead is successfully overcoming Multi User Interference produced by other users on the same UWB channel. This will be discussed in detail in following chapters.

5.2.2 Hybrid Modulation and Multiple Access

To achieve optimum performance for both the FFD and the RFD based on their complexity and power consumption limitations, the modulation coding and multiple access schemes (MCM)
must work with both, coherent and non-coherent receivers. The research by Zhang, Orlik, Molisch, Liu, and Zhang (2007) helped define the hybrid modulation scheme that has been accepted for the IEEE 802.15.4a standard. The transmit waveform defined by Zhang et. al. (2007) is represented by Equation 2.

$$\omega^{(k)}(t) = \sum_{i=0}^{\infty} \sum_{n=0}^{N-1} b_i^{(k)} p\left(t - nT_c - c_i^{(k)}T_b - iT_s - b_i^{(k)}T_{PPM}\right)d_{i,n}^{(k)}$$  \hspace{1cm} (2)

The superscript \((k)\) denotes the \(k^{th}\) user, \(b_i^{(k)}\) is the \(i^{th}\) data bit to be transmitted that modulates the position of the pulse in each symbol duration. \(c_i^{(k)}\) is the parity check bit associated with the \(i^{th}\) data bit. The parity check bit is also to be transmitted and modulated onto the phase of the pulse. \(T_c\) is the chip or pulse duration which is approximately 2 ns for the low band mandatory 499.2 MHz channel. \(T_b = NT_c = 32\) ns, \(n\) indexes the \(N = 16\) pulses that are transmitted during each data burst. \(c_i^{(k)}\) is the time-hopping sequence for multiuser access while \(T_{PPM}\) is the modulation interval for pulse position modulation. \(T_{PPM} = 16T_b\) while \(T_s\) is the symbol duration. \(d_{i,n}^{(k)}\) represents a pseudorandom scrambling sequence drawn from \([-1, 1]\). The pulse \(p(t)\) is the raised-cosine pulse that is used as the “basis pulse” for the 802.15.4a PHY layer (Zhang et. al., 2009).

Depending on the value of the data bit, it will be determined whether the burst of pulses will be in the first or the second half of the symbol. The resultant waveform achieved can be denoted as \(S\). It is also referred to as the “basis waveform”. Figure 5 depicts the modulation scheme while equation 3 represents the resultant waveform.
This basis waveform is modulated by both PPM and binary phase-shift keying. The modulation interval $T_{\text{ppm}}$ is 512 ns. This ensures that non-coherent receivers can detect the PPM even in channels with heavy delay dispersion since, the interval is much larger than the typical channel delay spreads. The duration of the burst waveform is in the order of tens of ns, or shorter than typical delay spreads. Thus, for successful detection at the receiver end the duration over which the receiver integrates is determined by the propagation channel. Reducing length of the burst will not reduce the optimum integration as the propagation channel is still the same. The coherent receiver can perform a correlation with $s_i^{(k)}(t)$ to enhance the signal-to-noise ratio by a factor $N$. A coherent receiver, also has additional information available by detecting the additional bit $\tilde{b}_i$ which is different from $b_i$. Thus, the coherent reception technique has significant advantages over the non-coherent method.
The position of the burst waveform is shifted by multiples of $T_b = 32\ ns$ in a pseudorandom way according to the time-hopping (TH) sequence. Multi-user access is achieved since each user has a unique TH sequence and so the position shifts for different users are different. TH is provided for both coherent and non-coherent receivers. The maximum possible time shift is $8T_b$ while the PPM time shift is $16T_b$. An interval $8T_b = 256\ ns$ serves a guard interval for channels with heavy dispersion (Zhang et. al., 2009).

Each user has a unique TH sequence which produces a unique order in which the positions for the bursts to appear for that user. Match filtering at the receiver input provides multi-user interference suppression. The extent of suppression depends on the cross-correlation between the TH sequences. The value of the TH sequence changes from one symbol to the next.

### 5.2.3 Coding

A coherent receiver can maintain double the bit rate as a non-coherent receiver. As was made apparent in the earlier section, a coherent scheme can receive 2 bits per transmit symbol while the non-coherent scheme receives only one bit per transmit symbol. Doubling the payload rate if the receiver can perform coherent reception is not a practical idea in sensor networks like a HAN, as we are looking for a low data rate scheme, but one that fulfills other primary criteria.

When considering broadcast situations, we need both coherent and non-coherent devices to be able to receive the message. Most relay nodes are non-coherent receivers as well, even if the final destination might be an FFD device capable of coherent reception. As a result we use the extra bits for the coherent devices to provide higher coding gain, thus, improving the robustness of the system. The code has to be designed such that, non-coherent receivers can still decode the signal in spite of not receiving the extra coding bits. A systematic code is a coding scheme in which the original bits are transmitted unchanged along with the extra parity check bits. This
coding scheme would be an appropriate option in this case. The original symbol bits or systematic bits modulate the position of the burst, while the parity check bits are phase modulated onto the burst itself. A coherent receiver can thus, see both systematic and parity check bits, while a non-coherent receiver can demodulate only the systematic bits. The information is protected using a systematic (51,43,8) Reed-Solomon code (IEEE Std. 802.15.4a, 2007).

5.2.4 Preamble and Synchronization

The preamble and synchronization is the most crucial and error prone part of UWB reception. Successful preamble detection and synchronization is the start of payload reception. It is also the process most susceptible to noise and interference as is proved later on in the paper. The literature on this process is limited and this paper helps fill the gap on aspects concerning the preamble and synchronization. Before any data detection is performed by the receiver, it is necessary to acquire, synchronize and perform channel estimation. A scheme called “perfectly balanced ternary sequences” is used to achieve and support channel acquisition. A specific preamble detectable by both coherent and non-coherent receivers is designed. When using ternary sequences as the preamble, the periodic autocorrelation function for both coherent and non-coherent receivers is perfectly matched. IEEE 802.15.4a transmits this preamble multiple times to improve Signal-to-Noise ratio (SNR) via processing gain. As a result, a processing gain suitable for channel estimation is achieved (Zhang et. al., 2009). The IEEE 802.15.4a standard will use either a length-31 or length-127 preamble code. In order to facilitate ranging, the preamble should have a single autocorrelation peak value and zeros elsewhere for the different receiver architectures. To avoid interference from simultaneously operating networks, the codes should also have minimal cross-correlation (Kwok et. al., 2006). Table 2 illustrates the 31 bit
ternary codes employed by the standard for various channels. The mandatory channel uses codes 5 and 6.

<table>
<thead>
<tr>
<th>Code Index</th>
<th>Code sequence</th>
<th>Channel number</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>-1,0,0,0,0,+1,0,-1,0,+1,+1,0,+1,-1,0,0,0,+1,-1,1,+1,1,0,0,-1,1,0,-1,0,0</td>
<td>0, 1, 8, 12</td>
</tr>
<tr>
<td>2</td>
<td>0,+1,0,+1,-1,0,+1,0,0,0,-1,+1,+1,0,-1,-1,-1,0,0,0,+1,+1,0,0,0,+1,0,0,0,0,0,0</td>
<td>0, 1, 8, 12</td>
</tr>
<tr>
<td>3</td>
<td>-1,+1,0,+1,0,0,0,-1,+1,-1,+1,0,0,+1,+1,0,0,-1,0,0,0,-1,0,+1,0,-1</td>
<td>2, 5, 9, 13</td>
</tr>
<tr>
<td>4</td>
<td>0,0,0,0,+1,-1,0,0,-1,0,+1,+1,+1,0,0,0,+1,0,-1,0,0,0,+1,0,0,0,-1,0,+1,0,-1</td>
<td>2, 5, 9, 13</td>
</tr>
<tr>
<td>5</td>
<td>-1,0,+1,-1,0,0,+1,+1,-1,0,0,0,-1,+1,0,+1,+1,0,-1,+1,0,0,0,-1,0,0,0,-1,0,0</td>
<td>3, 6, 10, 14</td>
</tr>
<tr>
<td>6</td>
<td>+1,+1,0,0,+1,0,0,-1,-1,-1,0,0,0,+1,-1,0,0,0,+1,0,-1,+1,0,0,0,0,0,0</td>
<td>3, 6, 10, 14</td>
</tr>
<tr>
<td>7</td>
<td>+1,0,0,0,0,+1,-1,0,0,+1,0,0,0,+1,0,+1,-1,-1,0,-1,0,0,-1,1</td>
<td>4, 7, 11, 15</td>
</tr>
<tr>
<td>8</td>
<td>0,+1,0,0,-1,-1,0,0,+1,0,0,0,-1,-1,0,0,0,+1,0,+1,+1,-1,+1,0,+1,0,0,0,0,0,0</td>
<td>4, 7, 11, 15</td>
</tr>
</tbody>
</table>

*Table 2: Perfectly Balanced Ternary Code Sequences for IEEE 802.15.4a Preamble*

This preamble code is used to construct the symbols that constitute the SYNC portion of the SHR preamble. As seen in Table 2, each preamble code is a sequence of code symbols drawn from a ternary alphabet belonging to \{-1, 0, 1\}, selected for use in the UWB PHY for their perfect periodic autocorrelation properties. Which codes are used in a particular UWB PHY channel is restricted. For a HAN using the ternary code which is indexed by \(i\), the SYNC part of the preamble shall consist of \(N_{\text{sync}}\) repetitions of the symbol \(S_i\), where \(S_i\) is the ternary code \(C_i\) spread by the delta function \(\delta_L\). The spreading operation where the code \(C_i\) is extended to the preamble symbol duration and is described mathematically in Equation 4.

\[
S_i = C_i \otimes \delta_L(n)
\]  

\[
\delta_L(n) = \begin{cases} 
1 & n = 0 \\
0 & n = 1, 2, ..., L - 1
\end{cases}
\]
After the Kronecker operation a preamble symbol is formed as depicted in Figure 10, where L-1 zeroes are inserted between each ternary element of \( C_i \) (IEEE Std. 802.15.4a, 2007). All devices in the same network are required to use the same preamble, enabling differentiation between transmissions from other networks. The IEEE 802.15.4a standard makes provisions for heavy multipath environments, where the delay spread is longer resulting in intersymbol interference. Hence, we can have adaptive settings for the pulse repetition frequency variable between 15.6 and 3.90 MHz depending on the level of noise or interference on the propagation channel.

5.3 MAC LAYER

The MAC layer design for this IEEE standard is very simple. It banks on the robust Multi User performance available to a UWB based system. Multiple-access is already ensured by the use of different ternary sequences between devices on various networks and time hopping codes between various devices on the same network. The mandatory medium access control mode for the 802.15.4a standard is ALOHA.
Each user transmits without checking if other users are on the air or not. Throughput improvement can be achieved by using traditional MAC layer techniques like TDMA. Thus, the option to enable TDMA does exist for this IEEE standard. By enabling the TDMA option in this protocol, a provision for allotting guaranteed slots for devices is made. Devices make slot requests to the PAN coordinator, which is responsible for maintaining synchronization among all the devices that it serves as PAN coordinator and signaling the allocation of slots back to the transmitters. There is a limit of only seven time slots per superframe. This allocation of time slots is done only between PAN coordinator and its associated device. Work on extending the timing of these time slots across several PAN coordinators has been taken up for the original Zigbee protocol in the past, but is not applicable to the 802.15.4a standard.

Another manner in which throughput can be improved is by carrier sensing, backoff scheduling, and handshaking. Several methods permitting the clear channel assessment are described in the IEEE 802.15.4a standard. For example, the correlation peaks of the received preamble are used to detect the preamble and are thus, indicative of a signal present. This approach if utilized significantly improves the throughput of the protocol, but increases the cost of the devices as there is an increase in complexity (IEEE Std. 802.15.4a, 2007).

5.4 Ranging

UWB networks generally use the time-of-arrival for determining the range between different nodes. These ranges form the basis of the actual location estimation. The 2.4GHz PHY is not suitable for ranging because its bandwidth is too small but, UWB is very wide by definition, so high resolution timing information is available making ranging possible. The IEEE 802.15.4a standard does not define the algorithm and implementation of signal detection and ranging, but
only has some provisions in place to realize accurate ranging. The two-way ranging protocol can be implemented to achieve accurate ranging.

5.4.1 Two-Way Ranging Protocol

Figure 7 depicts the ranging protocol. It is uncomplicated and works well for non-coherent and low cost devices. As a result it was the preferred protocol kept in mind when developing ranging features in the IEEE 802.15.4a standard.

![Figure 7: Exchange of Message in two-way ranging](image)

As we are talking about non-coherent devices, there is no presence of clock synchronization between the two ranging devices. The device A which initiates the ranging process uses its own clock as a time reference. The session begins with device A sending a range request message to device B. Device B can measure the absolute time of arrival of the message, but without synchronization with device A, it does not know the time of departure of the message and thus, cannot determine $t_p$. Rather device B waits a time $t_{replyB}$, known to both devices and sends a request to device A. Now device A can measure the round trip time $t_{round} = 2t_p + t_{replyB}$ with respect to its own reference time. The concern with this protocol though is the finite tolerance of
the device crystal reference frequency, since frequency error introduces error in the measurement of $t_p$. Using a high quality crystal that offers tolerances to the order of 2 parts per million is not good enough. As a result, some enhancements are suggested to accommodate these frequency errors better (IEEE Std. 802.15.4a, 2007).

### 5.4.2 Start Frame Delimiter (SFD)

The start frame delimiter is added after the preamble and before the PHY header of an 802.15.4a packet. It indicates the end of the preamble. Detection of the SFD helps frame synchronization and accurate ranging. The standard specifies a long and short SFD. The SFD consists of either 0, 1, or -1. The short SFD is 8 symbols while the long SFD which is used only for the lowest data rate is 64 symbols long. On detecting the SFD of a received range request, the ranging timing counter begins. The instant the SFD range reply packet leaves, the counter is stopped. The difference in these two counter values corresponds to the turnaround time. Processing gain for detection of SFD is 6 dB higher than that for the individual preamble signal. As a result SFD detection instants offer better accuracy to manage timing counters.

### 5.5 IEEE 802.15.4a Transmitter and Receiver

Figure 8 depicts the symbol generation block for an IEEE 802.15.4a compliant transmitter. The 802.15.4a standard offers variable data rates ranging from 0.11 Mb/s to 27.24 Mb/s. Each bit from the MAC layer is first Reed Solomon encoded. Then, each bit is systematically mapped onto data symbols with two bits each by a convolutional encoder. The resultant bits from the convolutional encoder are the actual symbols transmitted by the protocol. One bit determines the position of a pulse burst, the other selects the sign of the spreading sequence used in the burst. In
a non-coherent receiver scheme, the extra parity bit which is used to determine the sign of the burst is ignored, only the pulse position modulated bit, which represents the actual data bit itself, can be detected (Fischer M., et. al., 2009). Thus, this modulation scheme is called hybrid modulation as it consists of both position modulations for the data bit, which can be detected by a simple non-coherent energy detector receiver, and phase modulations for the parity bit which can be detected by the more complicated coherent receiver designs adding processing gain to the overall transmission.

The overall modulation scheme is called burst position modulation - binary phase shift keying (BPM-BPSK). According to the BPM, which is the modulation scheme of concern for non-coherent energy detector receivers, the burst is active in the first or the third period depending on the position bit of the data symbol. The remaining periods are used as guard intervals to compensate for multi-path delay spread. The actual position of the burst inside the selected period is defined by a time hopping sequence to mitigate multi-user interference.

The block diagram in Figure 8 shows the proposed signal source when implemented on a Xilinx Virtex-4-FX20 field programmable gate array (FPGA). We are going through the symbol generation schematic to understand the difference between data symbol and preamble code generation, as later on in the paper we will be considering the effects of multi-user interference (MUI) on the synchronization and timing acquisition, which contributes significantly to the packet error rate (PER) of the protocol. The lower part in figure 12 generates the actual data symbols by performing the RS encoding, followed by the convolutional encoding resulting in two outputs per data bit i.e. one parity bit, one data bit. A linear feedback shift register (LFSR) generates the spreading sequence used for each burst along with the time hopping sequence. The time hoping sequence is used to provide robust multiple access features for devices on the same
network, while the spreading sequence determines the phase of the burst which is used to provide higher processing gain in coherent UWB receivers. The BPM modulator uses the position bit to insert zeros before and after the burst where necessary. The result is a symbol stream at the peak PRF of 499.2 MHz with a very low duty cycle (Fischer M., et al., 2009).

![Block Diagram - Symbol Generation](image)

**Figure 8: Block Diagram - Symbol Generation**

Source: Fischer M., et al., 2009

The PPDU encoding process will help differentiate between the coding process followed for the preamble and the rest of the symbols. Each IEEE 802.15.4a PHY service data unit (PSDU) block is preceded by a preamble (synchronization header, SHR), and a PHY header (PHR). The SHR is used for synchronization and for determining the Time of Arrival (TOA) in ranging mode, while the latter carries information about the actual transmission parameters used in the PSDU. The PHR is encoded in a similar way as the PSDU – although with a more restricted set of symbol/bit rates (Fischer M., et al., 2009).

The SHR is not modulated using BPM-BPSK like the PHR and PSDU bits. A ternary code sequence of length 31 or 127 is spread by inserting a setup dependent number of zeros. The SYNCO part of the SHR plainly repeats the sequence which equals to multiplying it with +1. The number of repetitions for the SYNCO part is indicated in the PHR. After the insertion of zeros the
overall structure of the SHR is similar to the PHR and PSDU modulated symbols. The resultant chip rate is 499.2 MHz and each preamble symbol is 993.59 ns long (Fischer M., et. al., 2009).

Given in Figure 9 is a diagram depicting the schematic for transmission after symbol generation has been accomplished. In this schematic it is important to understand that each block can be realized in multiple ways to achieve the same overall behavior.

![Diagram](image)

**Figure 9: IEEE 802.15.4a Transmitter Schematic**

For the receiver schematic, the key is the manner in which the data detection is achieved. It can be done using a transmitted reference signal, and comparing the density function of the reference signal to the received signal in order to determine the value of the symbol. The low cost and less complexity option though is to use an energy detector.

### 5.6 Average Battery Life

The average battery life for an average AA battery powering an IEEE 802.15.4a standard node can be calculated as using equation 5.

\[
T_{oper} = \frac{I_b \cdot V_b}{P_{av}}
\]  

(5)
\( P_{av} \) has to be calculated keeping in mind that power is utilized in transmission, reception and within the control unit itself. Thus, \( P_{av} \) can be computed using equation 6.

\[
P_{av} = P_{TX} + P_{RX} + P_{CU}
\]  

(6)

In equation 6, \( P_{TX}, P_{RX}, \) and \( P_{CU} \) represent transmitted power, received power and control unit power respectively. The control unit power is relatively low in the range of a few microwatts. The transmitter and receiver power can be calculated using equation 7 (Chong and Yong, 2008).

\[
P_{TX} = P_{RX} = \frac{P_e}{\eta} = \frac{\frac{1}{2} \cdot D \cdot P_{in} \cdot T_{sym} \cdot R}{\eta}
\]  

(7)

In equation 5, \( P_e, D, P_{in}, R, T_{sym}, \) and \( \eta \) are the total average emitted power, duty cycle, instantaneous power, transmission rate, symbol duration and efficiency of transmission or reception. For the IEEE 802.15.4a standard has the following values for these variables. Duty cycle for the standard \( D = 1/32 \), \( R = 0.98 \) MHz, \( T_{sym} = 1025.64 \) ns while the efficiency \( \eta = 0.05 \).

Instantaneous power \( P_{in} \), can be calculated assuming transmission at the allowable level according to FCC standards. The FCC limit on power for UWB signals is \(-41.3\) dBm/MHz. As a result the instantaneous transmission power for an UWB signal of bandwidth 499.2 MHz is given by \( P_{in} = 0.037 \) mW \( = 37 \) µW.

We assume the use of a standard AA battery which has a rating of \( I_b = 1000 \) mA.h, while the value for voltage \( V_b = 1.5 \) V. The control unit power \( P_{CU} = 750 \) µW. Using these values, we
calculate the average time a battery will last while transmitting signals in accordance to this IEEE standard.

The complete equation for $P_{TX} = P_{RX}$ is given below.

$$P_{TX} = P_{RX} = \frac{1}{2} \cdot \frac{1}{32} \times 0.037 \times 1025.64 \times 10^{-9} \times 0.98 \times 10^6}{0.05}$$

Thus, $P_{TX} = P_{RX} = 0.1162 \text{ mW}$.

$$P_{av} = P_{TX} + P_{RX} + P_{CU}$$

$$P_{av} = 0.01162 + 0.01162 + 7.5 \times 10^{-3} (Assumed)$$

Thus, $P_{av} = 0.03074 \text{ mW}$. To calculate the lifetime for the battery we use equation 3.

$$T_{oper} = \frac{1.5 \times 1000}{0.03074} = 48796.36 \text{ hrs} \approx 5 \text{ years 6 months}$$

As a result the we can have calculated the operational time for a standard AA battery to be 5 years and 6 months when operating a node transmitting at the IEEE 802.15.4a standard. This time frame matches up to the battery life obtained from any other household device. Also, this calculation does not incorporate variable like idle state for the nodes and other battery related inefficiencies like self-discharge. Overall, the range for battery life will remain at between 5 to 7 years.
CHAPTER 6

Performance Evaluation

The performance evaluation considered is for a complete packet based system. In the past there have been several studies that consider the overall performance of an energy detector when detecting a Time Hopping signal undergoing interference. For the IEEE 802.15.4a, the study put together by Flury, Merz, Le Boudec, and Zory (2007) takes into account packet detection, timing acquisition, the estimation of the power delay profile of the channel, and the recovery of the encoded payload. The results obtained indicate that in spite of all the emphasis given to a low complexity energy detector producing good results, without the use of high complexity channel estimation, very few studies have concentrated on the effects of MUI on the energy detector itself. UWB is ideally meant to work extremely well under MUI and multi path fading effects. On the other hand the energy detector also, ideally works without any problem under the effects of other wideband and narrowband noise. The effects of other user interference though on an energy detector functioning according to the IEEE 802.15.4a standard as proven by this study seem to degrade the performance of the standard significantly. Thus, this study on the standard negates some of the most attractive features of UWB, namely robustness to MUI and as a result the possibility to allow parallel transmissions.

One of the main reasons we consider a non-coherent receiver using an energy detector over and above a coherent receiver using the concept of a rake receiver is its lower sampling requirements and its robustness to timing impairments. A rake receiver needs to properly estimate the channel characteristics, while an energy detector only needs to estimate the channel power delay profile. A non-coherent receiver is robust to timing impairments, but unfortunately
is less robust to multi-user interference as it cannot take advantage of the full diversity offered by the UWB channel.

To define MUI in the IEEE 802.15.4a standard, we go back to the MAC layer defined for this standard. The defined MAC layer is the uncontrolled ALOHA format. Relying on the robustness to MUI offered by UWB, the MAC layer does not incorporate carrier sensing, collision detection or collision avoidance before transmission. If an acknowledgement of successful reception for a packet is not obtained, there is a random backoff time before retransmission. In this MAC format, concurrent transmissions inevitably occur. Choosing the ALOHA format for the MAC layer is justified also due to the infrequent nature of transmission, as HAN network nodes are expected to transmit once every few minutes. The study and results obtained by Flury et. al. (2007) prove that, 802.15.4a compliant energy detectors show very little capture effect. Packets are lost if several transmissions occur concurrently. In this paper since, we consider a dense urban environment, the probability of concurrent transmissions from devices on the same network is accompanied by the probability of transmissions by devices on various overlapping networks as well. In a near-far scenario, with a single strong interferer, there is no capture effect at all. This increases the importance of this performance evaluation.

As has been described earlier an IEEE 802.15.4a packet consists of a preamble and the payload. The preamble is used to serve the following functionalities:

a. Packet Detection
b. Timing Acquisition
c. Estimation of Channel mask used by the Energy Receiver

Without going into the details of the synchronization process, we understand that on detecting the start-frame-delimiter, the symbol detection procedure begins. We keep in mind that the
channel mask estimation has already been performed in between preamble and SFD detection. The channel mask is a sampled and quantized version of the power delay profile of the channel. It is represented as a binary vector \([m_0, m_1, \ldots, m_{N_{ch}−1}]\), \(m_i\) belongs to \(\{0, 1\}\). The channel estimation is done by taking \(N_{ch}\) samples of the received signal averaging it over \(G\) blocks of \(N_{ch}\) samples. A threshold is then applied to quantize the \(N_{ch}\) values to 0 or 1. The channel mask is used to reduce the amount of noise accumulated by the energy receiver. As part of this averaging process, received signal level contained in the channel mask is also estimated. Mathematically, symbol detection can be represented as the scalar product in equation 8 (Flury et. al. 2007). For each frame, two scalar products are computed, one for an expected 1, the other for an expected 0.

\[
S_j = \sum_{i=0}^{N_{ch}−1} r_j[i] \cdot m_j 
\]

\(j = 0, 1\)

The array \(r_0\) indicates the power delay profile expected when a 0 bit is transmitted, while the array \(r_1\) will be the power delay profile expected when a 1 bit is transmitted. A comparator compares the outputs of the two scalar products; the larger value is the binary output being forwarded to the Reed Solomon decoder.

The simulation performed by Flury et. al. (2007), accurately depicts the 802.15.4a standard, including course and fine synchronization, and estimation of the channel mask. The entire MAC protocol though is not accurately represented, to reduce complexity. If the simulation were to represent the entire MAC protocol accurately, reception and decoding at for every packet from any device would have to be simulated at its destination; instead, the simulation is concerned
only with receiving and decoding packets from the user of interest. We denote $N_u$ as the number of users.

For MUI generation, each user is considered to have a queue with a packet arrival rate given as $\lambda_i$, $i = 0, 1, \ldots, N_u - 1$. According to the IEEE 802.15.4a standard, the backoff for every packet is drawn with the backoff exponent set to maximum. Thus, when the backoff exponent expires and there is a packet at the head of the queue, this packet is transmitted without carrier sensing or any provision made to avoid collision. The simulation considers the mandatory band 3 with center frequency 4.49 GHz with preamble codes 5 and 6, with each packet consisting of 1014 bits corresponding to 1208 RS encoded symbols. The channel model (CM1) is utilized, producing an RMS delay of 18 ns (Balakrishnan et. al., 2004) which is a slight variation to the traditional AWGN channel. Before we get into the results of this simulation, a short discussion on how the SNR for the simulation is defined is necessary.

$$SNR = \frac{E_p}{N_0}$$

(9)

where $E_p$ is the energy received per impulse, or convolution of the received signal with the impulse response of the channel. A Gaussian noise process is band limited to the bandwidth of the signal and $N_0/2$ is the variance of the zero mean. The method defined by Beaulieu, N., and Tan, C. (1997), is used to correlate the Gaussian noise samples, since the sampling frequency is more than 2 times the bandwidth of the signal. For simulations performed on a multipath channel with no MUI, it is proved that the receiver is well balanced between the synchronization procedure and decoding of data.
For simulations with MUI, Flury et. al. (2007) have considered 2 separate cases for the frequency band 3. The first case has all the devices using the same preamble code, while in the second case, the user of interest uses code 5 while the rest of the users use code 6. In both cases the packet arrival rate for all users will be identical. The high traffic case has \( \lambda = 200 \) packet/s which corresponds to an effective data rate of 241 kbit/s, while the low traffic case has \( \lambda = 10 \) packets/s resulting in an effective data rate of 12.1 kbit/s (Flury et. al. 2007).

The results obtained from the simulation are compared to two specific capture models to help understand the results better. In the “Destructive Capture Model”, the packet is lost whenever there is more than one active transmission at a time. If there is a single active transmission, single user performance is assumed. In a “Perfect Capture Model”, when there are multiple active users might compete for the packet detection and timing acquisition. One of these users who is randomly chosen, succeeds and now experiences single user performance. If the transmission was perfectly orthogonal the “Perfect Capture Model” would be the ideal performance received.

The first scenario considered has 2 users, with the low packet rate of \( \lambda = 10 \) packets/s. In this scenario there seems to be an error-floor at which packet error rate becomes stable and does not drop anymore even with increase in SNR. In their simulation, Flury et. al. (2007) have also considered two near far cases. The first case with 2 users with the received power of the second user 10 dB higher than the user of interest, while the other case considers 4 users with their received power 3 dB below the user of interest. When comparing the PER results obtained from each of these simulations to the results for a Destructive or Perfect capture models, it is noticed that the PER for the cases considered is much closer to that of the destructive capture model than
the perfect capture model (Flury et. al. 2007). Thus, we seem to be losing out on the most appealing features of UWB which is its robustness to MUI.

The point of interest from this simulation is the analysis of what is the cause of this packet loss. Ideally an UWB PHY layer ought to give an extremely robust performance when operating under the influence of MUI.

The analysis reveals 2 primary reasons for packet error:

1. Packet is missed during the synchronization phase.
2. Errors in the packet are too many to be corrected by RS code.

When we consider packets receiver errors due to misses in synchronization phase, these can be broadly categorized into two possibilities:

a) Missed Detection (MD): The receiver is trying to acquire a packet, but is unable to do so, or it acquires the preamble, but misses the SFD.

b) False Alarm (FA): The receiver wrongly assumes to be already successfully synchronized thus, it is not trying to acquire a packet at all.

When Flury et. al. (2007) considered the cases with users using same and different preamble code in the high traffic case, they expected a result in which the scenario A with same preamble code would lead to much higher packet reception errors in comparison to the scenario B with different preamble codes. The result they obtained though indicated that even though packet errors in scenario A were larger than the packet errors in scenario B, the difference was extremely small with no significant impact on performance. Table 3 gives a summary on the packet error analysis.
What is clear from the results in Table 3 is that irrespective of the nature of preamble codes used, whether identical or different, a large chunk of the packet error is produced due to FA. This indicates that packets are lost in one of two scenarios.

i. The receiver acquires timing correctly, but noise or interfere signal lead to the receiver exceeding the SFD threshold as a result, wrongly declaring detection of the SFD.

ii. Receiver acquires wrong timing due to noise and interference, as a result wrongly declares the SFD.

From Table 3, it is also evident that, of all the packets that are not received more than 50% are lost due to an FA, which means that whether the interferer and the user of interest use the same or different preamble codes, the receiver is still synchronizing often with the interferer. The reason for this becomes apparent when we consider the preamble codes, their autocorrelation and cross-correlation properties when used as a preamble in an energy receiver detector.
MUI Mitigation in Noncoherent UWB ranging via Nonlinear Filters

The concept of using nonlinear filters for the mitigation of interference was first introduced while considering interference mitigation in image processing. We have already looked at the packet structure for the 802.15.4a standard and have evaluated the performance of the protocol.

Figure 10: UWB Energy Detector Receiver

The preamble of each 802.15.4a packet consists of multiple repetitions of a base symbol waveform. The symbols transmitted are determined depending on a pre-determined ternary code. The receiver can detect and process the preamble either by template matching (coherent) or by Energy Detection (ED). We consider an ED for mitigation of MUI, since the effects of MUI are far more prominent on ED. Also, Energy detectors are more resilient to pulse-distortion, simple and operate at sub-Nyquist sampling rates.
The ED receiver in consideration is depicted in Figure 10. It feeds the band-passed received signal into a square-law device, integrates its output, and then samples this signal periodically. Generated energy samples are denoted as \( z[n] \), while the sampling interval and the number of samples per symbol as \( t_s \) and \( n_b = T_{sym}/t_s \) respectively. \( z[n] \) is then regrouped into a 2D matrix. On the formation of the matrix, it is passed through a nonlinear filter to enhance desired signal energy parts and remove MUI. The matrix is then converted back into a 1D time series to locate the leading edge, by means of adaptive search back and thresholding techniques. (Sahinoglu and Guvenc, 2006)

### 7.1 Energy Matrix Formation

Ranging and detection for an UWB signal depends heavily upon SNR. Now the SNR can be improved significantly on increasing processing gain, which is achieved by coherently combining received signal energy samples as explained by Gezici et. al. (2005). In the presence of MUI though, we do not get the desired result with the coherent energy combining approach. Sahinoglu and Guvenc (2006) have illustrated this through their paper that signal design itself and coherent energy combining is not good enough to overcome the drawbacks of a Multiuser interference. While using nonlinear filters, the solution to this problem lies is considering the collected energy samples with a different view.

In the example we consider, the symbol is meant to consist of 4 frames. Signal energy is integrated and sampled such that it produces 4 samples per frame. This results in a total of 16 samples per symbol. Assume the TH code for desired signal is \{0, 4, 4, 3\} and that for the interfering signal is \{0, 4, 5, 4\}. This method leads to the formation of an energy matrix \( Z \), which consists of \( M \) rows and \( N \) columns. \( M \) is the number of frames processed while \( N \) is the number
of samples per frame. As per the example considered, $Z$ for the user of interest will be a 4 X 4 matrix.

z[0 + 0] & z[1 + 0] & z[2 + 0] & z[3 + 0] \end{pmatrix}$$

Each column of $Z$ is grouped with samples produced according to the received signal’s TH pattern. The left-most vertical line indicates the time of arrival of the signal. With the interference following a different TH pattern, the energy matrix does not form a vertical line as it should for the column that indicates the leading edge in the matrix. As a result, applying conventional leading edge detection techniques causes erroneous detection of the leading edge (Sahinoglu and Guvenc, 2006).

### 7.2 Energy Matrix for TH-IR

The energy samples collected are represented as the following equation 10 given below.

$$z^{(th)}[n] = \int_{(n-1)\tau_s}^{n\tau_s} \left| r^{(th)}(t) \right|^2 dt$$ (10)

These samples are grouped together in accordance to the TH code. The samples for the same group are used to populate the columns of an energy matrix.

Interference caused due to MUI and self-interference that is induced due to improper auto-correlation properties of the TH code resulting in the presence of short discrete lines. Energy samples are grouped according to the desired users’ TH codes. It is thus possible that only some
of the energy groups contain energy due to a partial overlap with the signal’s TH pattern. Also, if the uncertainty period for the energy collection process is greater than the time period for a frame $T_f$, there is an increase in MUI and self-interference (Sahinoglu and Guvenc, 2006).

### 7.3 NonLinear Matrix Filtering

Two popular nonlinear filters used in image processing are the minimum filter and the median filter. We consider these two to reduce MUI in the 802.15.4a standard system. $W$ is considered the length of the filter which is applied to each column of the matrix formed. The minimum filter replaces the center sample with the minimum of the samples within the filter window. Thus, the new energy matrix can be represented as in equation 11.

$$z^{(\text{min})}[\lambda, k] = \min\{z[\lambda,k], z[\lambda+1,k], ..., z[\lambda+W-1,k]\}$$  \hspace{1cm} (11)

Where $\lambda \in \{1, 2, ..., N_{s}N_{\text{sym}} - W + 1\}$. $N_{s}$ is the number of frames per symbol while $N_{\text{sym}}$ is the number of symbols per packet. Once the interference is overcome, $z^{(\text{min})}$ is converted into a vector by column sum operation.

Median filters apply a similar approach to a minimum filter, but a median filter replaces the center value of a given data set with the median of the data set. A longer median filter is less effective to mitigate multiuser interference. Any unsuppressed interference may be passed on to neighboring samples. The best means to overcome this drawback is using non-overlapping windows for filtering (Sahinoglu and Guvenc, 2006). The new energy matrix obtained can be represented as in equation 12.
Both these forms of nonlinear filtering though, increasingly add to the complexity of the energy detection receiver. Memory for storing energy matrices is not only expensive, but would significantly add to the cost of the receiver. The question at this stage is whether the advantages of using a nonlinear filter justify the increase in cost of the system and at what point the additional complexity and power consumption is approximately the same as a coherent receiver, which would still need MUI mitigation.

7.5 Trade off

While implementing mitigation using nonlinear filters we assume the use of a standard 16 bit Analog-to-Digital converter. Thus, to store a given matrix $z$, we will require $2^*M^*N$ bytes of storage. For an IEEE 802.15.4a standard transmission, in the mandatory low band, we have 2 frames per symbol and 1209 symbols per packet. Thus per packet for a given user, $M$ is a constant at 1209 symbols and 2 frames per symbol. On the other hand depending on sampling rate varying from 2ns i.e. chip rate to higher, $N$ has a maximum value of 128 samples.

Thus, considering a best case sampling rate of 4ns per sample, we are produce $D$ bytes of data which is determined by equation 13.

\[
D = 2^*M^*N
\]

\[
D = 2^*(1209^*2)^*128
\]

\[
D = 619008 \text{Bytes}
\]
For every matrix we need to consider a maximum of 619 Kbytes of data. Every node would need an ability to store at least 619 Kbytes on board. This would significantly increase the cost of the each node and the complete HAN. Accommodating extra storage capacity while fulfilling other requirements for a HAN node like low complexity, low cost, low power consumption and spatial constraints would be a difficult task. A trade off needs to be made between the requirement for MUI mitigation using nonlinear filters and increase in the end to end cost of the system.
CHAPTER 8

Autocorrelation and Cross-correlation properties of ternary codes

Up-sampled versions of the ternary codes in Table 2 are used to generate the preamble for the IEEE 802.15.4a standard. These codes are meant to have ideal autocorrelation and crosscorrelation properties, such that, it produces a distinct peak during autocorrelation and no peaks when crosscorelated. We notice a squaring circuit while considering the schematic for an energy detector receiver in Figure 10. The squaring circuit reduces the effectiveness of the autocorrelation and crosscorelation properties of the ternary code. Thus, the characteristics vary significantly between a receiver that can consider pulses of both polarities and an energy detector receiver as in the case of an IEEE 802.15.4a non-coherent receiver. In the charts shown below, ternary codes 5 and 6 have been considered. Ternary codes 5 & 6 are used for preamble creation for devices on networks operating in the most common low mandatory band of the 802.15.4a standard.

Autocorrelation – Figure 11 and 12 depict the difference between the ideal autocorrelation and observed autocorrelation for an energy detector receiver while using ternary code 5. The amplitude of the central peak remains the same. Even when the autocorrelation is not ideal, there is a significant difference observed between the central peak and secondary maximums within the autocorrelation.
Figure 11: Ideal Autocorrelation properties for Ternary Code 5

Figure 12: Autocorrelation properties for Ternary Code 5 for an Energy Detector Receiver
Crosscorrelation – The maintenance of ideal crosscorrelation properties of the ternary codes ensures mitigation of interference caused by other users on the network. As seen in Figure 14, the crosscorrelation properties suffer a significant deviation from ideal for an energy detector receiver. Ideal crosscorrelation properties for ternary codes 5 and 6 are plotted in Figure 13. Ternary codes 5 and 6 are the codes chosen for the Low Mandatory channel i.e. channel 3. The peaks produced by crosscorrelation of ternary code 5 and 6 are significantly high in amplitude and can in fact, be mistaken for a peak in the presence of noise, leading to a false detection. Thus, in the presence of noise or with a high power interference, the crosscorrelation properties used to detect and mitigate interference, will in fact detect the presence of a legitimate packet. The plots in Figure 13 and 14 help differentiate between the ideal crosscorrelation properties of ternary codes 5 and 6 against the properties observed for an energy detector receiver.

Crosscorrelation for ternary code 5 and 6

Figure 13: Ideal Crosscorrelation between Ternary codes 5 and 6
Crosscorrelation for ternary code 5 and 6 at an IEEE 802.15.4a Receiver

Figure 14: Crosscorrelation properties at an Energy Detector receiver for ternary codes 5 & 6
CHAPTER 9

**Power-Independent Detection and Preamble Code**

**Interference Cancellation (PICNIC)**

The PICNIC algorithm was suggested by Le Boudec et. al. (2009) as a solution to overcome the MUI problem encountered in energy detector non-coherent receivers for the IEEE 802.15.4a UWB standard. We did look at nonlinear filters earlier as a possible solution, but increase in complexity and cost due to the formation of the energy matrix might just make an implementation of the nonlinear filter method less of a possibility. Also, the use of a nonlinear filter is not adaptive and thus, cannot guarantee accurate performance under changing MUI conditions.

As discussed earlier, the IEEE 802.15.4a standard packet consists of a preamble and a payload. We are mainly interested in overcoming missed detections and false alarms when dealing with the preamble. Nonlinear filters are more suitable when dealing with the payload. The preamble consists of the SYNC and the SFD. The SFD marks the beginning of the payload. The time unit for the 802.15.4a standard is a chip of length $T_c$. In the SYNC part of the preamble pulses are sent at every $L^{th}$ chip. The preamble pulses have no time hopping associated with them and are based solely on the preamble ternary codes of length $C$. The receiver model consists of an antenna, a band pass filter, a squaring device and an integrator sampled at rate $1/T$. Prior to integration or sampling the received SYNC preamble can be given as $r(t) = x(t - r_0) + \omega(t)$ where $x(t)$ is the contribution of the user of interest (UOI). $r_0$ is the propagation delay and $\omega(t)$ accounts for MUI and thermal noise.
Thermal noise is generated by the receiver circuitry. It is a zero-mean AWGN process with power spectral density (PSD) $N_0/2$. We assume that the extent on thermal noise is known and can be calibrated or estimated in a robust fashion. The signal for the UOI is given by

$$x(t) = \sum_{i=0}^{N_{sync}-1} \sum_{j=0}^{C-1} s_j \cdot h(t - (j + iC) LT_c)$$  \hspace{1cm} (13)$$

where $s_j$ belongs to $\{+1, 0, -1\}$ is the preamble code used by the UOI and $h(t)$ denotes the unknown channel response. The discrete time signal obtained after sampling is $y_n = \int_{nT}^{(n+1)T} [r(t)]^2 dt$. We assume $T = L/M \cdot T_c$ and $M$ is a divisor of $L$. Thus, $LT_c$ is the minimum inter-pulse spacing and, we obtain $M$ discrete samples $y_n$ per pulse. The payload on the other hand consists of short and continuous bursts of a constant number of pulses, say $L_b$. Pulses are characterized by pseudo-random polarity and time-hopping (Le Boudec et. al., 2009). Thus, a received payload signal from an interfering transmitter contains $L_b$ times more energy than an interfering preamble pulse. There are two possible preamble codes, with perfect auto-correlation but, not the best cross-correlation properties. Also, if there is no user signal and only the received signal is purely AWGN, the distribution $y_n$ can be approximated with a chi-square distribution with $2BT$ degrees of freedom, whose cumulative distribution function will be represented as $F_{\chi^2_{2BT}}$.

Three approaches to countering the effects of MUI have been suggested by Le Boudec et. al. (2009). Their approaches include the salient features of the nonlinear filter suggested earlier and are in increasing order of robustness. The approaches considered here also mainly are meant to
overcome MUI introduced due to cross-correlation between ternary codes, which are selected for preamble symbol creation and do not possess the desired ideal crosscorrelation properties.

The classical base algorithm that can be employed involves the correlation of the receiver output with a template derived from the known preamble code sequence of the UOI. The most important factor is that even though the preamble is formed with multiple symbols. Each symbol in the preamble is formed by up sampling pre-decided ternary codes. Each letter in ternary code belongs to \{-1, 0, +1\}. Since the non-coherent receiver uses energy detection, we use a squaring device before sampling as a result, a squared amplitude of every symbol is received (Le Boudec et al., 2009). Thus, the cross-correlation properties of the ternary code results in multiple undesirable peaks as seen in Figure 17. Squaring a signal also introduces possible aliasing contamination of sampled signals; increasing sampling requirements. The template is formed by repeating the preamble code \(N_G\) times to obtain a processing gain. Length of the template is given as \(M_T = N_G \cdot C \cdot M\). \(C\) is the length of the preamble code while \(M\) is the number of samples created per pulse. The template \(t_i\) is represented as Equation 14.

\[
t_i = \sum_{k=0}^{N_G-1} \sum_{j=0}^{C-1} s_j^2 \cdot \delta(i - (j + kC)M) \tag{14}
\]

After sampling, a correlation output is generated that is defined in Equation 15.

\[
z_n = \sum_{i=0}^{M_T-1} t_i \cdot y_n (M_T - 1) + i \tag{15}
\]
Now, the preamble of the signal for the UOI can be represented as $LT_c \cdot C - periodic$ samples. Thus, the samples produced for the UOI, $z_n = M \cdot C - periodic$, if the signal is present. The baseline algorithm processes the samples in blocks of $MC$ consecutive samples. Thus the $i^{th}$ block is $z_i = \{z_{iMC}, z_{iMC+1}, \ldots, z_{(i+1)MC-1}\}$. The baseline algorithm has 2 steps, namely, detection and verification. The detection step declares the presence of a signal by comparing the samples of the current block to $U_{base}$, which is a known base value, if the received signal is AWGN only.

$$U_{base} = \frac{N_0}{2} F_{\chi_{2BT_c C_1 N_G}}^{-1} \left(1 - p_{AWGN}^{base}\right) \quad (16)$$

$C_1 = \sum_{j=0}^{C-1} s_j^2$ denotes the number of non-zero code symbols of the UOI preamble code. Thus, $C_1 \cdot N_G$ samples are combined due to the template. The design parameter $p_{AWGN}^{base}$ dictates the threshold, which is based on the probability of AWGN signal only exceeding the threshold.

For a given block of samples, if a sample exceeds the threshold, a signal detection is declared on that $i^{th}$ block. Initial timing acquisition is determined on a particular sample within the block based on the sample, $j_{l}^{max}$ having the highest correlation output value. For verification, it is required that for $N$ sample blocks following this sample block, the maximum sample value which is represented as $z_{(i+k)MC} + j_{i+k}^{max} > U_{base}$. Also, it is necessary that the index of this sample $j_{i+k}^{max}$ differs only by the minimum inter-pulse distance $M$ from the synchronization point $j_{i+k-1}^{max}$ of the previous block. This ensures that both maxima stem from the same preamble. If these clauses for verification succeed for $N$ consecutive blocks, the verification succeeds and
synchronization is declared. On the other hand, even if a single verification were to fail, detection starts afresh.

This baseline method works efficiently for a single user. When we take MUI into account, especially a strong interference signal, there is a very high chance on exceeding the threshold $U_{\text{base}}$ based purely on the interference noise, even when not aligned with the template. Interference from other users can produce missed detections (MD) by producing maxima in following sample blocks leading to a failed verification. It also can lead to false alarms (FA) when interference produces a maximum for $N+I$ sample blocks.

Failures in the baseline algorithm to counter MUI, brings us to the 2\textsuperscript{nd} degree of robustness that can be achieved in an ED UWB receiver by using Power-Independent Detection (PID) using thresholding. The PID method applies a threshold check at the input of the correlation rather than the output of the correlation as done in the baseline algorithm. This prevents any form of interference from dominating the correlation output leading to MD and FA. The algorithm is very simple, in which samples above the threshold are set to 1, while those below the threshold are set to 0. Like a limiter this algorithm produces a capture effect that can effectively suppress an interferer if the C/I is somewhat positive (in dB). Thus the correlation output can be indicated as in Equation 17.

$$Z(n) = \sum_{i=0}^{M_T-1} t_i \cdot 1_{[\gamma_n (M_T-1)+i > U_{\text{pid}}]}$$

(17)

In Equation 17, $1_{[\cdot]}$ denotes the indicator function, which decides whether the sample is a 1 or a zero, while the threshold $U_{\text{pid}}$ is represented in Equation 18.
is parameterized by \( P_{AWGN}^{pid} \) which is the probability that a pure noise signal can exceed the threshold. Apart from the threshold check before correlation, all the steps in the PID method are identical to the baseline method.

Preamble Code Interference Cancellation is used to overcome the significant performance loss indicated in the earlier performance evaluation of the IEEE 802.15.4a standard. This performance loss is caused mainly due to imperfect cross-correlation properties of the various ternary codes involved with symbol formation for the preambles. This method uses nonlinear filters to further increase the robustness of the PID method suggested earlier. The PICNIC algorithm pre-processes each block \( z_l \), but still takes its output from the PID block attempting to detect and cancel out interference by looking for the specific cross-correlation pattern between a pair of ternary codes within each block (Le Boudec et. al., 2009). This is done before it hands it over for coarse synchronization. The algorithm calculates the cross-correlation pattern between the ternary codes. On doing so a fixed number of sub-blocks \( C_{peak} \) of length \( M \) have high energy, while \( C_{trough} \) sub-blocks have a low energy, while the rest of the blocks \( C_{mid} = C − C_{peak} − C_{trough} \) have a mid-energy within this cross-correlation template. The algorithm for comparing the characteristics of the received signal to the cross-correlation template already formed proceeds in the following three steps.

1. **Detecting the presence of an interfering preamble code**: The first step in the PICNIC algorithm tries to determine if an interfering preamble is present or not. If a preamble is not detected, then the process moves on to the timing acquisition method. This first step

\[
U_{pid} = \frac{N_0}{2} F^{-1}_{X_{2BT}} (1 - P_{AWGN}^{pid})
\]  

(18)
tries to determine sub-blocks with energy levels corresponding to the cross-correlation properties of the ternary signal. This is done by the creation and correlation of two ternary vectors of length \( C \), namely \( z_{tern} \) and \( x_{tern} \). \( z_{tern} \) is constructed from all the sub-blocks of samples \( z_i \) that have been collected so far. The maximum value for every sample position over all the sub-blocks is considered and denoted as \( z_j^{max} = \max \left( z_{jM}, z_{jM+1}, \ldots, z_{(j+1)M-1} \right) , j \in \{0,\ldots,C-1\} \). Then \( z_{\max}^{\text{max}} \) is converted to the ternary vector \( z_{\text{tern}} \) by replacing its \( C_{\text{peak}} \) highest values with “+1”, its \( C_{\text{trough}} \) lowest values with “-1” and the rest with “0”. Again, \( x_{\text{tern}} \) is obtained by taking the cross-correlation of the ternary codes and replacing its highest and lowest values with “+1” and “-1” respectively, while zero takes the place of the rest of the values. From here on we correlate \( z_{\text{tern}} \) and \( x_{\text{tern}} \), and compare it to the cross-correlation vector (Le Boudec et. al., 2009). If the number of coinciding peaks and troughs between the cross-correlation and the sample vector exceed the threshold, \( \frac{C_{\text{peak}} + C_{\text{trough}}}{2} + 1 \), interference is assumed to be present. If not, the algorithm moves on to the PID method.

2. **Determination of First Interference Multipath Component:** If it is determined that interference is present, we try and obtain a rough estimate of the energy delay-profile of the interference signal. To do this, the first multipath component of the interfering signal must be determined, only then can we subtract this interference from the set of sampled sub-blocks \( z_i \). The set of samples \( z_j^{\max} \) is considered. Attention is paid to the original indices for each of the samples \( C_{\text{peak}} \) within \( z_j^{\max} \) whose corresponding values in \( z_{\text{tern}} \) is a “+1”. The majority on the original indices of the \( C_{\text{peak}} \) sub-blocks from their original blocks \( z_i \) helps determine the index of the strongest path in a sub-block of length \( M \). For a given set of samples, we start from that index of the high-energy sub-block \( C_{\text{peak}} \) for a
window of length $W$ from the strongest path and search for the first path above the noise threshold level given by $U^b_{pid}$.

$$U^b_{pid} = \frac{N_0}{2} F^{-1}_{Bin \left(C_1, N_G, P^{pid}_{AWGN} \right)} \left( 1 - P^{pid, sb}_{AWGN} \right)$$ (19)

In Equation 19, $F_{Bin \left(C_1, N_G, P^{pid}_{AWGN} \right)}$ is the cumulative distributive function of a binomial distribution with parameters $C_1$, $N_G$, and $P^{pid}_{AWGN}$. This binomial distribution corresponds to the distribution output in (13) only if AWGN were present (Le Boudec et. al., 2009).

The threshold is set by fixing $P^{pid, sb}_{AWGN}$, which is the probability that the AWGN can exceed this threshold. Eventually, the first path index is the lowest path index for more than half of the $C_{peak}$ individual searches.

3. **Interference Cancellation:** Cancellation of interference is eventually achieved by subtracting the estimated channel energy delay profile of the interference from the received signal. A given set of samples $z_i$ is split into $C_{peak}$ high energy sub-blocks, $C_{trough}$ low energy sub-blocks, and $C_{mid}$ medium energy sub-blocks. A separate energy delay profile exists for high energy, low energy and medium energy samples.

$$e^j_{type} = \{e^0_{type}, e^1_{type}, ..., e^M_{type} \}$$

$Type = \{high, medium, low\}$. Thus, $e^j_{type}$ represents the energy delay profile for high, low and medium energy sub-blocks, respectively. If there are $i$ sub-blocks $z_i$, with $j$ samples each, then $z^i_{type}^j$ is the $j^{th}$ energy sample of the $i^{th}$ energy sub-block. Thus the channel energy delay profile is calculated as in equation 20.
The delay profile is subject to high, low and medium energy sub-blocks and is calculated for each. It is here that the nonlinear filtering aspect for MUI mitigation is applied in considering the median of the values to generate an energy delay profile. The median is used instead of the mean to be robust to outliers. $\omega_{AWGN}$ is the expected noise level at the output of the correlation. $e_{\text{type}}$ is now subtracted from high, low and medium energy samples, to cancel out interference in each one of the energy sub-blocks.

In the research conducted by Le Boudec et. al. (2009) on the PICNIC algorithm, a network was simulated in compliance with the IEEE 802.15.4a. The scenario considered is for a single receiver and $N_u$ transmitters. The packets are placed in a queue by every transmitter and are transmitted in accordance to the Aloha back-off procedure. A 100% utilization of the capacity available would yield an average packet rate of about 200 packets/s, per transmitter. Le Boudec et. al. have thus considered a packet rate of 100 packets/s. What is important to note is that on a real IEEE 802.15.4a implementation a transmitter would transmit in bursts of packets and remain dormant for the rest of the time, thus resulting in an average transmission rate lower than 100 packets/s.

The simulation considers the mandatory mode on the IEEE 802.15.4a: Frequency Band 3. Preamble codes 5 and 6 are in Low Pulse Repetition Frequency (LPRF) mode. The codes have 31 bits, the spreading factor $L = 64$ and the preamble length $N_{\text{sync}} = 64$. The protocol samples at the chip rate i.e. every 2 ns. The template used consists of $G = 10$ repetitions while the window length for verification is $N = 16$. This combination of parameters ensures that timing acquisition is fast enough such that channel estimation etc. can be performed on the preamble. The

\[
e_{\text{type}} = \text{median}(z_{0,j}^{\text{type}}, z_{1,j}^{\text{type}}, \ldots, z_{\text{type}-1,j}^{\text{type}}) - \omega_{AWGN}
\]
simulation sets $p_{AWGN}^{base} = 0.999$, $p_{AWGN}^{pid} = 0.8$, $p_{AWGN}^{pid, sb} = 0.9999$. The results for a NLOS channel model have been shared. The main metric of performance considered is the packet synchronization error rate which consists of mainly of FAs and MDs. Figure 15 shows the results for a simulation with two interferers with different ternary codes, but power levels equal to the user of interest.

![Figure 15: Results for Simulation of MUI Mitigation Algorithm](image)

Source: Le Boudec et. al., 2009

As is observed from the results obtained, the PICNIC method performs at least two orders better than the other methods of MUI reduction in this scenario. Simulations performed by Le Boudec et. al. (2009) have considered various other scenarios, but the one most applicable to MUI mitigation for the preamble and the payload is the one depicted in figure 18. Thus, PICNIC is a possible solution to MUI mitigation to ensure timing acquisition during the preamble.
CHAPTER 10

Conclusion

The IEEE 802.15.4a standard based on UWB reception is a great example of a standard that theoretically is extremely feasible and qualifies all the criteria for a HAN. It is a standard that ensures operation in high frequency band thus, reducing the possibility of interference with other protocols already operating in the 2.4 GHz spectrum. The standard uses UWB as the means for the PHY layer which is inherently immune to narrowband and wideband interference, even if it were to arise in these high frequency ranges. It offers a range of about 10m. With the presence of relay devices in a given house or apartment, the network can easily be designed to obtain the coverage needed. Non-coherent UWB transmission is cheap and not complex and is the preferred option for transmission.

The major problem is overcoming MUI at the receiver, both to ensure detection of the preamble and overcome interference from other users in the payload. The methods considered in this paper give a guideline to what the best means to resolve this issue might be. Yet, the strict constraints on every HAN node, might make it difficult to make an appropriately functioning 802.15.4a standard system when under the influence of heavy MUI. Thus, a receiver that employs a combination of statistical interference modeling and thresholding is a slightly more complex solution, but provides a better performance in comparison to purely thresholding algorithms. Considering that a given node does not transmit very, it is important to consider how severe these effects will be when looking at a practical system that does not have continuous transmission for the user of interest. The reserved time slot option for a user might also be a reliable option when dealing within urban HANs.
REFERENCES

Amendment to IEEE Std. 802.15.4, IEEE Std. 802.15.4a – 2007 (Aug. 31, 2007), IEEE Computer Society.


APPENDICES
## APPENDIX 1 - UWB PHY rate-dependent and timing-related parameters

<table>
<thead>
<tr>
<th>Channel Number</th>
<th>Peak PRF MHz</th>
<th>Bandwidth MHz</th>
<th>Preamble Code Length</th>
<th>Modulation &amp; Coding</th>
<th>Data Symbol Structure</th>
<th>Data</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td></td>
<td></td>
<td></td>
<td>Viterbi Rate  RS Rate Overall FER Rate</td>
<td># Burst Positions per Symbol</td>
<td># Hop Bursts</td>
</tr>
<tr>
<td>(0.3, 5.6, 8.10, 12.14)</td>
<td>499.2</td>
<td>499.2</td>
<td>31</td>
<td>0.5</td>
<td>0.87</td>
<td>0.44</td>
</tr>
<tr>
<td>(0.3, 5.6, 8.10, 12.14)</td>
<td>499.2</td>
<td>499.2</td>
<td>31</td>
<td>0.5</td>
<td>0.87</td>
<td>0.44</td>
</tr>
<tr>
<td>(0.3, 5.6, 8.10, 12.14)</td>
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<td>31</td>
<td>0.5</td>
<td>0.87</td>
<td>0.44</td>
</tr>
<tr>
<td>(0.3, 5.6, 8.10, 12.14)</td>
<td>499.2</td>
<td>499.2</td>
<td>31</td>
<td>0.5</td>
<td>0.87</td>
<td>0.44</td>
</tr>
<tr>
<td>(4, 11)</td>
<td>1331.2</td>
<td>127</td>
<td>31</td>
<td>0.5</td>
<td>0.87</td>
<td>0.44</td>
</tr>
<tr>
<td>(4, 11)</td>
<td>1331.2</td>
<td>127</td>
<td>31</td>
<td>0.5</td>
<td>0.87</td>
<td>0.44</td>
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<tr>
<td>(4, 11)</td>
<td>1331.2</td>
<td>127</td>
<td>31</td>
<td>0.5</td>
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<tr>
<td>(4, 11)</td>
<td>1331.2</td>
<td>127</td>
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<td>0.5</td>
<td>0.87</td>
<td>0.44</td>
</tr>
<tr>
<td>7</td>
<td>1081.6</td>
<td>31</td>
<td>0.5</td>
<td>0.87</td>
<td>0.44</td>
<td>32</td>
</tr>
<tr>
<td>7</td>
<td>1081.6</td>
<td>31</td>
<td>0.5</td>
<td>0.87</td>
<td>0.44</td>
<td>32</td>
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<tr>
<td>7</td>
<td>1081.6</td>
<td>31</td>
<td>0.5</td>
<td>0.87</td>
<td>0.44</td>
<td>32</td>
</tr>
<tr>
<td>7</td>
<td>1081.6</td>
<td>31</td>
<td>0.5</td>
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</tr>
<tr>
<td>15</td>
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<td>0.5</td>
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<td>0.5</td>
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<td>0.5</td>
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<td>0.44</td>
<td>32</td>
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<tr>
<td>15</td>
<td>1354.97</td>
<td>31</td>
<td>0.5</td>
<td>0.87</td>
<td>0.44</td>
<td>32</td>
</tr>
</tbody>
</table>

Source: IEEE Std. 802.15.4a, 2007
APPENDIX 2 – PPDU Encoding process

<table>
<thead>
<tr>
<th>Step</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>From MAC via PHY SAP</td>
<td>PSDU</td>
</tr>
<tr>
<td></td>
<td>Variable length: 0-127 octets</td>
</tr>
<tr>
<td>Reed-Solomon encoding</td>
<td>Data field (non-spread &amp; before convolutional encoding)</td>
</tr>
<tr>
<td></td>
<td>Variable length: 0-1208 bits</td>
</tr>
<tr>
<td>Add PHY Header</td>
<td>PHY Header (PHR) 13 bits</td>
</tr>
<tr>
<td></td>
<td>Data field (non-spread &amp; before convolutional encoding)</td>
</tr>
<tr>
<td></td>
<td>Variable length: 0-1208 bits</td>
</tr>
<tr>
<td>Add SECDED bits</td>
<td>PHY Header (PHR) 19 bits</td>
</tr>
<tr>
<td></td>
<td>Data field (non-spread &amp; before convolutional encoding)</td>
</tr>
<tr>
<td></td>
<td>Variable length: 0-1208 bits</td>
</tr>
<tr>
<td>Convolutional encoding</td>
<td>PHY Header (PHR) 38 bits</td>
</tr>
<tr>
<td></td>
<td>Data field (after coding, before spreading)</td>
</tr>
<tr>
<td></td>
<td>Variable length: 0-2418 bits</td>
</tr>
<tr>
<td>Spreading</td>
<td>PHY Header (PHR) 19 symbols @ 850 or 110 kb/s</td>
</tr>
<tr>
<td></td>
<td>Data field</td>
</tr>
<tr>
<td></td>
<td>0-1209 symbols @ variable rate</td>
</tr>
<tr>
<td>Insert SHR preamble</td>
<td>SHR Preamble 16, 64, 1024 or 4096 symbols</td>
</tr>
<tr>
<td></td>
<td>PHY Header (PHR) 19 symbols @ 850 or 110 kb/s</td>
</tr>
<tr>
<td></td>
<td>Data field</td>
</tr>
<tr>
<td></td>
<td>0-1209 symbols @ variable rate</td>
</tr>
<tr>
<td>Modulate</td>
<td>SHR Preamble 16, 64, 1024 or 4096 symbols</td>
</tr>
<tr>
<td></td>
<td>PHY Header (PHR) 19 symbols @ 850 or 110 kb/s</td>
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<td>Data field</td>
</tr>
<tr>
<td></td>
<td>0-1209 symbols @ variable rate</td>
</tr>
<tr>
<td></td>
<td>coded @ base rate</td>
</tr>
<tr>
<td></td>
<td>BPM-BPSK coded @ 850 kb/s or 110 kb/s</td>
</tr>
<tr>
<td></td>
<td>BPM-BPSK coded @ Rate indicated in PHR</td>
</tr>
</tbody>
</table>

Source: IEEE Std. 802.15.4a, 2007