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Kurumbailmadham, Raju Rakesh

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Mitigating the Effects of Destructive ISI in IR-UWB Wireless Systems

Kurumbailmadham Raju Rakesh

School of Electrical & Electronic Engineering

A thesis submitted to the Nanyang Technological University
in partial fulfillment of the requirement for the degree of
Master of Engineering

2019

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The contributions of the co-authors are as follows:

- I drafted out the manuscript and reviewed this with Prof Lee
- I reformatted the manuscript as advised by Prof Lee

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- I consolidated the simulations results for a multipath channel exhibiting destructive ISI
- Prof Lee suggested I run simulations for other channel conditions to verify expected behavior
- I revised the manuscript with the guidance of Prof Lee.

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Date

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[Kurumbailmadham Raju Rakesh]

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Summary

Impulse Radio Ultra Wideband (IR-UWB) systems used in multipath rich channels are very susceptible to signal distortion due to Inter-Symbol Interference (ISI) as well as signal attenuation due to destructive ISI. Equalizers can be used to undo signal distortion due to Inter-Symbol Interference but any signal attenuation due to destructive ISI would still be present and this would lead to lower effective data transmission rates.

In this report, we propose a solution, which we refer to as the Real-Time Channel Estimator (RTCE), which helps to undo signal distortion due to Inter-Symbol Interference while reducing the effects of destructive ISI as well and thus permitting higher effective data transmission rates. This is particularly useful for Impulse Radio Ultra Wideband systems operating between 3.1 GHz and 10.6 GHz which are subject to the low power emission restrictions imposed by the Federal Communications Commission (FCC).

In essence, the Real-Time Channel Estimator would estimate the Channel Impulse Response (CIR) function for the entire bandwidth of the IR-UWB communication channel and represents this in a well conditioned matrix form. We measure the singularity of this matrix, using the Singular Value Decomposition technique, to minimize the effects of perturbations due to channel noise. It would then attempt to determine the transmitted signal from the received signal using the Gaussian elimination technique.

In the following sections, we review most of the commonly used equalization techniques and compare their performance against our proposed solution in a multipath channel with destructive ISI. We then compare the Bit Error Rate (BER) of these different types of equalizers over various Signal-to-Noise-Ratio (SNR) values and their computational complexity and show that our proposed solution performs better than most commonly used equalizers.

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1.Introduction

The performance of all wireless communication systems is primarily limited by the effect the channel has on the transmission. The transmission via a wireless channel can lead to signal distortion due to channel fading and/or due to channel noise.

Channel fading can be classified into two broad categories: large scale fading and small scale fading. Large scale fading occurs due to free space path loss and defines the radio coverage area of a transmitter for a given transmission power. Small scale fading on the other hand, occurs when a transmitted radio signal reaches the receiving antenna by more than one path due to reflections from large objects (such as walls, buildings), diffraction by objects with sharp edges (such as mountain ridges) and scattering by a large number of small objects (such as foliage) resulting in multipath propagation. When multiple versions of a signal arrive at the receiver antenna at slightly different times, the combined effects of these signals may lead to constructive or destructive fading for narrowband signals and signal distortion due to Inter-symbol Interference (ISI) for wideband signals.

1.1. Small Scale Fading

Small scale fading is caused by the interference between different versions of a transmitted signal arriving at the receiver antenna via different paths. This can result in constructive or destructive fading of the signal depending on how the different multipath signals combine at the receiver. This often results in different frequencies experiencing different levels of attenuation and/or phase changes. This is described as the channel exhibiting frequency selective fading behaviour.

Motion of the receiver, transmitter or objects within the channel could cause the wireless channel to change over time and this could result in the channel exhibiting time selective fading behaviour.

1.1.1. Time Dispersion/ Frequency Selective Fading

Ideally, a wireless communication channel should have a single, line of sight, wireless transmission path and its transfer function (i.e. $h[n]$), in the discrete time domain, should be a

delta function as defined by (1.1). If this is not the case, the channel would exhibit some form of small scale fading.

$$\delta[n] = \begin{cases} 1, & n = 0 \\ 0, & n \neq 0 \end{cases} \quad (1.1)$$

When multiple transmission paths exist between a transmitter and receiver, the signal at the receiver antenna would be the sum of the different multipath signals and hence a multipath channel can be represented by (1.2) where K is the number of multipath components, τ_k is the excess delay and $a_k e^{j\phi_k}$ is the complex gain of the k th multipath component [3].

$$h(t) = \sum_{k=1}^K a_k e^{j\phi_k} \delta(t - \tau_k) \quad (1.2)$$

A multipath channel may exhibit flat fading or frequency selective fading characteristics depending on length of the channel delay spread and the symbol period of the transmitted signal. A multipath channel would exhibit flat fading characteristics if the symbol period is much greater the channel delay spread as shown in the (1.3) where T_s is the symbol period and T_d is the channel delay spread.

$$T_s \gg T_d \quad (1.3)$$

Similarly, if the symbol period is much less than the channel delay spread as shown in (1.4) where T_s is the symbol period and T_d is the channel delay spread, the channel would exhibit frequency selective fading characteristics which would in turn lead to signal distortion due to Inter-Symbol Interference (ISI).

$$T_s \ll T_d \quad (1.4)$$

1.1.2. Frequency Dispersion/ Time Selective Fading

Ideally, a wireless communication channel should behave as a Linear Time-Invariant (LTI) system. However, the motion of the receiver, transmitter or objects within the channel could cause the wireless channel to change over time.

Motion of objects within the channel can result in the channel exhibiting slow fading or fast fading characteristics depending on the length of the channel coherence time and the symbol period of the transmitted signal. A channel would exhibit fast fading characteristics if the symbol period is much greater than the channel coherence time as shown in the (1.5) where T_s is the symbol period and T_c is the channel coherence time.

$$T_s \gg T_c \quad (1.5)$$

Conversely, a channel would exhibit slow fading characteristics if the symbol period is much less than the channel coherence time as shown in the (1.6) where T_s is the symbol period and T_c is the channel coherence time.

$$T_s \ll T_c \quad (1.6)$$

Typically fast fading occurs only at low data rates.

1.2. Large Scale Fading

Large scale fading occurs due to free space path loss and defines the radio coverage area of a transmitter for a given transmission power. The Free-Space Path Loss (FSPL), describes the loss in signal strength that occurs when an electromagnetic signal propagates over a line of sight path through free space, without any nearby obstacles. The FSPL is defined by the (1.7), where d is the distance of the receiver from the transmitter in meters, λ is the signal wavelength in meters, f is the signal frequency in hertz and c is the speed of light in vacuum. From this equation, we can observe that the FSPL for a signal is determined by the frequency of the signal and the distance between the transmitter and receiver.

$$FSPL = \left(\frac{4\pi d}{\lambda}\right)^2 = \left(\frac{4\pi d f}{c}\right)^2 \quad (1.7)$$

1.3. Channel Noise

In wireless communication channels, noise refers to any unwanted electromagnetic signal that distorts a transmitted electromagnetic data signal. This limits the receiver's ability to detect the transmitted signal and hence reduces the overall effective data transmission rate.

Channel noise can be from either natural sources or from manmade sources or both. Natural sources of noise, commonly referred to as Johnson-Nyquist noise, are caused by the random motion of electrons and come from various natural sources such as the atmosphere, the sun and other galactic sources. Thermal noise can be described as a zero mean Gaussian random process in the time domain and hence termed Additive White Gaussian Noise (AWGN). A Gaussian random process is a random variable statistically characterized by the Gaussian probability density function shown in (1.8) where σ is the standard deviation and σ^2 is the variance.

$$p(x) = \frac{1}{\sigma\sqrt{2\pi}} e^{-\frac{1}{2}\left(\frac{x^2}{\sigma^2}\right)} \quad (1.8)$$

Manmade sources of noise are generated by devices which emit electromagnetic waves within the frequency spectrum of a transmitted electromagnetic data signal. Common manmade devices which generate noise include electrical devices such as Microwave ovens, television sets, etc. The underlying probability density function of such noise sources are difficult to model, however the central limit theorem in probability theory suggests that when several independent random variables are added, their summation and mean will tend to have a Normal distribution even if the underlying random variables themselves are not normally distributed. (1.9) below shows the probability density function of a normal distribution and Figure 1.1 [9] shows a plot of a normal distribution where each band has a width of 1 standard deviation where μ is the mean, σ is the standard deviation and σ^2 is the variance.

$$p(x|\mu, \sigma^2) = \frac{1}{\sigma\sqrt{2\pi}} e^{-\frac{1}{2}\left(\frac{(x-\mu)^2}{\sigma^2}\right)} \quad (1.9)$$

Figure 1.1: A plot of normal distribution [9]

1.4. Wireless Communication Systems

Wireless communication systems can be broadly classified into three main categories: (1) Personal Area networks (PAN), (2) Wireless Local Area Networks (WLAN) and (3) Wireless Wide Area Networks (WWAN). All of these categories of wireless communication systems are affected by small scale fading.

1.4.1. Wireless Personal Area Networks (WPAN)

Wireless personal area networks (WPAN) are used to interconnect devices such as mobile phones, tablets, television sets, etc within about 1 meters range and are based on the IEEE 802.15 standards [5]. The two most commonly used techniques to generate UWB signals for PAN are: Multicarrier Orthogonal frequency-division multiplexing (OFDM) transmission and Impulse Radio (IR) UWB transmission.

1.4.2. Wireless Local Area Networks (WLAN)

Wireless Local Area Networks (WLAN) are used to interconnect devices such as laptops, smart phones, tablets, printers, etc within confined areas such as homes and offices and typically have a range of about 10 meters.

Most WLAN are based on the IEEE 802.11 standards [18]. Within the IEEE 802.11 family the IEEE 802.11n and IEEE 802.11ac are more recent and are very widely used. The IEEE 802.11n works in both the 2.4GHz and 5GHz frequency bands and operates with a bandwidth of 20MHz or 40MHz. The IEEE 802.11ac operates only on the 5GHz frequency band and has a bandwidth of up to 160MHz. Both the IEEE 802.11n and IEEE 802.11ac fundamentally rely on OFDM technology to mitigate the effects of channel fading [19].

1.4.3. Wireless Wide Area Networks (WWAN)

Wireless Wide Area Networks (WWAN) are wireless networks designed to function in much larger areas than WLAN and thus use different technologies to cope with the more complex channel conditions. Long-Term Evolution (LTE) and Long-Term Evolution Advanced (LTE-A) are one of the most commonly used mobile telecommunication cellular network technologies to deal with these channel conditions and both of these technologies rely on OFDM for downlink transmission [18]. For uplink transmission, Single-Carrier Frequency

Division Multiple Access (SC-FDMA) is used due to its lower Peak-to-Average Power Ratio (PAPR).

LTE supports a number of different channel bandwidths from 1.4MHz to 20MHz as shown in Table 1.1, all having a subcarrier spacing of 15kHz.

Channel Bandwidth	1.4MHz	3MHz	5MHz	10MHz	15MHz	20MHz
Subcarrier Spacing	15kHz					

Table 1.1: Channel Bandwidths supported by LTE

1.5.Motivation

In 2002, the Federal Communication Commission (FCC) allowed the unlicensed operation between 3.1 GHz and 10.6 GHz for Ultra wideband (UWB) communication, using an appropriate wideband signal format with a low EIRP level (-41.3dBm/MHz) [1].

The low power emission restrictions imposed by FCC makes this very suitable for WPAN and the large bandwidth offers the possibility of high speed data transmission. This opens up the possibility for new high speed short range wireless applications such as wireless video cables, wireless USB, etc. Other benefits of UWB transmission include, robustness against interception and detection, providing resistance to jamming and being better able to pass through walls [4][5].

The two most commonly used techniques to generate UWB signals for PAN are Multicarrier OFDM transmission and Impulse Radio (IR) UWB transmission. [1]

Multicarrier Orthogonal Frequency-Division Multiplexing (OFDM) uses several subcarriers, each of which has a symbol period much larger than the channel delay spread, to transmit wideband signals. This allows OFDM to be able to combat inter-symbol-interference (ISI) very effectively. However, as explained in [1], this method further exacerbates the significant power limitations imposed by the FCC, as each subcarrier would receive only a small fraction of the total power.

Using IR-UWB systems allows us to transmit signals with a much higher power due to the large bandwidth of the signal while still complying with the FCC power restrictions. This has made IR-UWB to be the preferred choice for high data rate Wireless Personal Area Network (WPAN) [12].

IR UWB signals would typically have a symbol period of several nanoseconds. If the channel delay spread is longer than this symbol period, ISI would occur at the receiver which would in turn result in signal distortion. Equalizers could be used to address the problem of signal distortion, however the effects of any destructive ISI introduced by the channel would still be present, exacerbating the power restrictions imposed by FCC.

1.6.Objectives

The primary objective of this thesis is to minimize the effects of destructive ISI when attempting to recover a signal corrupted by ISI for IR-UWB systems operating between 3.1 GHz and 10.6 GHz. As wireless systems operating in this spectrum are subject to the low power emission restrictions imposed by the FCC, minimizing the effects of destructive ISI would allow us to achieve higher effective data transmission rates.

1.7.Major Contribution of the Thesis

There are several solutions to address the problems associated with small scale fading and reduce their effects. In this thesis, we review these solutions and examine their effectiveness and limitations when used in wireless personal area networks (WPAN). We then propose a solution that addresses most of these weaknesses and limitations and thus allows us to achieve higher effective data transmission rates.

1.8.Organization of the Thesis

In chapter 2, we review the existing technologies to reduce the effects of small scale fading. In chapter 3, we examine the weaknesses of these existing technologies and propose a solution that addresses these weaknesses. We refer to this proposed solution as the Real-Time Channel Estimator (RTCE). In Chapter 4, we review the performance of the RTCE in mitigating the effects of AWGN and as well as reducing the effects of destructive ISI for IR-UWB Systems. In Chapter 5, we review the results consolidated in Chapter 4, and arrive at a conclusion and mention the future work that needs to be done. In Chapter 6, we mention the publications that were referenced as part of this thesis.

2. Existing Technologies to Reduce the Effects of Small Scale Fading

2.1. Mitigating Time Dispersion/ Frequency Selective Fading

As explained in Chapter 1, if a signal is transmitted via a multipath channel with a symbol period less than the channel delay spread, the signal would experience frequency selective fading. This would result in signal distortion due to ISI and hence poses as a major obstacle to high speed data transmission.

In this section, we review the commonly used techniques to prevent the occurrence of ISI while transmitting data at high speeds via multipath channels. ISI can be dealt with by avoiding it, mitigating it or by reducing its effects.

ISI could be avoided by either lengthening the symbol period or by transmitting data using the Orthogonal Frequency Division Multiplexing (OFDM) technique, explained in 2.1.4, or by using a combination of both techniques such as the WiMedia UWB as explained in [16] which uses zero postfix in addition to OFDM.

Mitigating the effects of ISI can be done using equalizers which are explained in 2.1.3. However, using equalizers would increase the complexity of the communication system. An alternative solution, to avoid this complexity, would be to reduce the effects of ISI using the shift window approach [10] explained in 2.1.1.

2.1.1. Generalized Shift-Window Approach for UWB ISI Reduction

Shift Window was introduced in [10] as an effective means to reduce ISI in a Wireless Chip Area Network (WCAN). The Generalized Shift Window approach described in [11], attempts to utilize this technique for Wireless Personal Area Networks (WPAN).

The Shift Window approach is based on the assumption that the energy from the interfering previous symbol would gradually get weaker and as a result would not interfere with the current symbol's detection.

The fundamental idea behind the shift window approach is to perform the correlation operation, in a correlation-based receiver, during the window for which the impact of the ISI from previous symbols' transmission is the lowest.

The Shift Window approach primarily involves 2 steps,

Step 1, would involve transmitting $2^{NI/2}$ combination of pilot training symbols to train the system to obtain the preferred correlation window where NI represents the number of interferences and can be defined by equation (2.1).

$$NI = \frac{\text{channel excess delay}}{\text{symbol period}} \quad (2.1)$$

Step 2, would involve using a RAKE receiver to perform the correlation operation, represented by $\otimes$ to produce the test statistic (2.2) which would be used for detection.

$$Z_m = \sum_{ls}^{le} \alpha_l v(t - \tau_l) \otimes R_m(t) \quad (2.2)$$

Where, $R_m(t)$ is the received signal with ISI and $v(t)$ is the template used for the RAKE receiver.

The Shift window approach cannot be applied directly to WPAN though, due to the following reasons. Firstly, as an indoor WPAN would have a much longer channel excess delay than the WCAN, transmitting $2^{NI/2}$ combination of pilot training symbols would be difficult. Secondly, the Shift Window approach limits its time shift search to be within the current symbol period. However, in the case of a non line of sight (NLOS) channel, the peak energy of a symbol may fall in the next symbol period and thus the time shift generated would not be optimal. Lastly, the Shift window approach is based on the assumption that the energy from the interfering previous symbol would gradually get weaker, however, in an indoor WPAN channel is characterized by cluster-based arrival and hence, energy of the replicas may not grow weaker.

However, this approach could be generalized and tailored to allow it to perform better in more severe multipath channels. Figure 2.1 below shows the Generalized view of the Shift Window Approach.

Figure 2.1: Generalized view of the Shift Window Approach [11]

In its tailored form the Generalized Shift Window approach would function as follows,

The transmitter would transmit N_p pilot symbols. In order to reduce the time required by the training process in the Shift Window approach, a simple energy detector is used instead of the correlation window selection algorithm. We then measure and store the energy of different window regions using the energy detector.

The window region could be divided using either using the shift window approach as described in [10]. Alternatively, the window could be divided by fixing the window region period (e.g. 2 ns) and dividing the received pilot signal by this period. Hence, the number of window regions would be defined by (2.3).

$$\text{number of window regions} = \frac{\text{channel excess delay}}{\text{window region period}} \quad (2.3)$$

The correlation window selector selects one or more regions which have the highest energy among all the regions which would then be used as the correlator templates for the next N_d data symbols.

In order to handle scenarios where there are multiple peaks, more than one correlator could be used. The number of correlators should be close to the number of clusters and the correlation window period should be close to the average cluster length. The receiver will correlate and sum all the correlation results to produce the test statistic to be used for detection.

For the correlator, a transmitted reference (TR) receiver is used instead of a selective RAKE in order to reduce the complexity of the system [13] and to simplify it even further, a single reference TR [14] (where $1 \leq N_p < N_d$) is used.

Overall, the generalized Shift-Window approach, provides a good performance improvement without the complexity of the ISI avoidance or mitigation.

2.1.2. Data Directed Estimation

Data Directed Estimation (DDE) is a technique described in [15] used to mitigate the effects of ISI when transmitting a single-carrier signal via multipath channels. DDE technique estimates channel parameters and data blocks by appending training data to the message block when transmitting. This technique estimates the channel using the steepest-descent algorithm and subsequently estimates the source data block using the Levinson recursive algorithm.

Figure 2.2. Signal Structure of DDE [15]

The signal structure of the DDE is shown Figure 2.2 above. It is made up of one block Known Symbols (training data of Length N) followed by one block of Unknown Symbols (message of Length M) and together they form 1 Frame.

The information throughput and the fraction of training are given by (2.4) and (2.5) respectively.

$$\text{Information Throughput} = \frac{M}{M + N} \quad (2.4)$$

$$\text{Fraction of Training} = \frac{N}{M + N} \quad (2.5)$$

In this technique, the channel is modelled as a tapped delay line with time-varying coefficients and the taps are spaced at the symbol interval (T). Hence, a channel with a tapped delay line of length L can be shown in vector as (2.6) where t denotes the transpose.

$$\underline{f}^t = [f_L \ f_{L-1} \ . \ . \ . \ f_0] \quad (2.6)$$

The channel shown by (2.6) would have a multipath delay spread of LT and N would need be larger the L for the DDE technique to function.

When transmitting data via a single carrier signal at a high data rate, the channel would exhibit slow fading characteristics and hence, we could assume that the channel is fixed within each decision block. Hence, the received signal without the additive noise could be represented by (2.7)

$$\underline{r}' = \begin{bmatrix} \underline{f}^t & & & & \phi \\ & \underline{f}^t & & & \\ & & \underline{f}^t & & \\ & & & \cdot & \\ & & & & \cdot \\ \phi & & & & \underline{f}^t \end{bmatrix} \begin{bmatrix} A_1 \\ B \\ A_2 \end{bmatrix} \quad (2.7)$$

Where,

$$A_1^t = [a_0 \ a_1 \ . \ . \ . \ a_{N-1}] \quad (2.8)$$

$$B^t = [b_0 \ b_1 \ . \ . \ . \ b_{M-1}] \quad (2.9)$$

$$A_2^t = [a_N \ a_{N+1} \ . \ . \ . \ a_{2N-1}] \quad (2.10)$$

(2.4) could be rewritten as 2.8,

$$\underline{r}_1' = \underline{M}_1 B + \underline{M}_2 A_3 + \underline{M}_3 A_4 \quad (2.11)$$

Where,

$$\mathbf{A}_3^t = [\mathbf{a}_{N-L} \mathbf{a}_{N-L+1} \cdot \cdot \cdot \mathbf{a}_{N-1}] \quad (2.12)$$

$$\mathbf{A}_4^t = [\mathbf{a}_N \mathbf{a}_{N+L} \cdot \cdot \cdot \mathbf{a}_{N+L-1}] \quad (2.13)$$

In (2.11), $\underline{M}_2 A_3$ and $\underline{M}_3 A_4$ form the post-cursors and precursors, respectively, while A_3 and A_4 are the training symbols. Using the training symbols, the channel weights can be estimated by a steepest-descent adaptive method. The channel equations can then be solved using the Levinson recursive algorithm to determine the data symbols.

In a multitone modulated signal, the transmitter would need to be capable of supporting a high peak-to-average power ratio in the range of 8-9 dB. This would mean that the average transmitted power would be about 8-9 dB below the maximum deliverable power of the transmitter. This would in turn result in significant power loss. Also if there are spectral null in the channel on or near one of the tones used, there would be a significant degradation in performance.

Using the DDE technique on a single tone modulated signal would help to address the above mentioned problems, as it would have a lower peak-to-average power ratio and spectral nulls would have a less detrimental effect.

In the DDE technique though, if a single tone modulated signal is transmitted via a multipath channel of length L , it requires training data of length N to be prefixed to each of the message blocks and the value of N needs to be larger than L . This would greatly reduce the effective data transmission rate as shown in (2.4).

In comparison, the RTCE explained in Chapter 3, does not require training data to be prefixed to each of the message blocks and hence does not suffer the loss in data transmission rate as a result of (2.4) unlike the DDE.

2.1.3. Equalizers

Equalization refers to any technique used to undo the effects of ISI on a received signal. Equalizers can be split into two main categories: Equalizers implemented using the

Maximum Likelihood Sequence Estimation (MLSE) algorithm and Equalizers which are essentially filters designed to undo the effects of ISI.

The MLSE Equalizers compare the received data sequence with all possible data sequences and chooses the sequence with the highest probability of occurrence. They are most commonly implemented using the Viterbi algorithm as it is the most efficient implementation. ISI Filtering Equalizers try to reverse the effects of a frequency selective fading channel and thus attempt to undo the effects of ISI on the received signal.

In the following sections, we will be reviewing the MLSE Equalizer as well as the Linear Equalizers and Non-Linear Equalizers both of which are implementations of ISI Filtering Equalizers.

2.1.3.1. MLSE Equalizer

The Maximum Likelihood Sequence Estimation (MLSE) Equalizer compares the conditional probabilities of all the possible transmitted message sequences with the received message sequence to recover data from a signal corrupted by ISI [20]. (2.14) and (2.15) define the MLSE algorithm, where $\mathbf{Z}$ is the received message sequence, $\mathbf{U}^{(m)}$ is one of the possible transmitted message sequences and z_{ji} and $u_{ji}^{(m)}$ are the j th code symbol of the i th branch of $\mathbf{Z}$ and $\mathbf{U}^{(m)}$ respectively.

$$P(\mathbf{Z}|\mathbf{U}^{(m')}) = \max(P(\mathbf{Z}|\mathbf{U}^{(m)})) \text{ over all } \mathbf{U}^{(m)} \quad (2.14)$$

$$\text{where, } P(\mathbf{Z}|\mathbf{U}^{(m)}) = \prod_{i=1}^{\infty} \prod_{j=1}^{\infty} P(z_{ji}|u_{ji}^{(m)}) \quad (2.15)$$

The Viterbi algorithm is the most efficient implementation of the MLSE equalizer. Compared to other equalizers, theoretically a Maximum Likelihood Sequence Estimation (MLSE) Equalizer yields the best possible performance, but is computationally very intensive.

2.1.3.2. ISI Filtering Equalizers

All Inter-symbol Interference (ISI) filtering equalizers essentially use some form of digital filtering technique, such as a Finite Impulse Response (FIR) Filter or Infinite Impulse Response (IIR) Filter, to remove the signal distortion introduced by ISI.

2.1.3.2.1. Linear Equalizers

The effect a channel has on a wireless signal transmitted through it is as defined by (2.16), where $X(f)$ is the transmitted signal, $H_c(f)$ is the wireless channel and $Y(f)$ is the received signal. Linear Equalizers attempt to remove ISI by applying the inverse of the frequency response of the channel to undo the effects of the channel on the signal [20].

$$Y(f) = X(f)H_c(f) \quad (2.16)$$

Linear Equalizers can be implemented using an FIR filter and can be defined by (2.17). Figure 2.3 shows its block diagram. Linear Equalizers apply a weight to each of the current and previous received signals which are then summed such that the additive interference contributed by the ISI is removed.

$$s'[n] = \sum_{i=0}^N b_i y[n - i] \quad (2.17)$$

As linear Equalizers do not have any feedback component, they are also referred to as feed forward equalizers. The filter weights need to be recomputed over time in order to adapt to changing channel conditions. A commonly used algorithm for this is the LMS algorithm.

Figure 2.3: Linear Equalizers Block Diagram

A linear equalizer has its limitations though. When a signal is transmitted via a frequency selective fading channel, a Linear Equalizer would attempt to remove any variations in gain in the channel frequency by increasing the gain of the frequency components such that the channel appears to have flat fading characteristics. This would unfortunately also amplify the noise in those frequency components. Thus, minimising ISI comes at the cost of increasing

noise which leads to lower Signal to Noise Ratio (SNR) especially when certain frequency components on the channel exhibit very low gain. Non-Linear Equalizer can be used to address this problem to some extent.

2.1.3.2.2. Non-Linear Equalizers

All wireless communication channels essentially behave as low pass filters, attenuating signals at higher frequencies and hence can be modelled as Finite Impulse Response (FIR) filter. By using a Linear Equalizer, to undo the effects of the channel, we would be attempting to create an inverse of the channel which would have an Infinite Impulse Response and cannot be completely modelled by a Linear Equalizer which has a Finite Impulse Response. Also, linear equalizer would amplify noise in frequencies with low channel gain which lowers signal quality. A commonly used non-linear equalizer used to avoid this problem to some degree is the Decision Feedback Equalizer.

Figure 2.4 below shows an example of a power delay profile of a multipath channel. Equalizers attempt to recover the original signal using the signal transmitted via the strongest multipath and removing the interference contributed by the other paths. Multipath components after the strongest multipath are termed post-cursor ISI and those before are termed pre-cursor ISI.

Decision Feedback Equalizers (DFE) attempt remove signal distortion introduced by the post-cursor multipath components by using the symbols that have been detected and subtracting out their ISI contributions from the received signal. Using this approach helps to remove any noise amplification by the equalizer for the post-cursor multipath signals. DFE use a linear equalizer to remove the pre-cursor ISI.

Figure 2.4: Power Delay Profile of a multipath channel

Hence, a DFE consists of two components: a) Feed Forward Equalizer and b) Feedback Equalizer. Figure 2.5 illustrates this in a block diagram.

Figure 2.5: Simplified block diagram of DFE

The implementation of a DFE is similar to that of an IIR Filter and can be defined by (2.18).

$$s'[n] = \sum_{i=0}^N b_i x[n-i] + \sum_{j=1}^P a_j s[n-j] \quad (2.18)$$

A DFE has some limitations though. As mentioned earlier, DFE can only address the problems associated with noise amplification for the post-cursor multipath components. Hence, noise amplification due to the use of a linear equalizer to remove pre-cursor ISI is still a problem.

Also, as the output of the symbol detector is fed back to correct the received signal, if the symbol detector makes an error when detecting a symbol, this error may cause subsequent symbols to be detected in error resulting in error propagation and mean lead to a catastrophic failure. (i.e. The equalizer reaches an unstable state and hence detects all future symbols incorrectly).

Orthogonal Frequency Division Multiplexing (OFDM), discussed in section 2.1.4, can be used to significantly reduce the occurrence of ISI and thus help to address these problems.

2.1.4. OFDM

Orthogonal Frequency Division Multiplexing (OFDM) is a multi-carrier modulation technique used to cope with the frequency selective nature of wireless channels due to multipath. It is widely used in both Wi-Fi Standards such as IEEE 802.11n and IEEE 802.11ac as well as in Cellular Standards such as LTE, LTE-A and WiMax.

2.1.4.1. Overview

The fundamental limitation of a single carrier modulation scheme is its inability to deal with the frequency selective fading nature of wireless channels. This problem further exasperates when transmitting high data rate signals with large bandwidths. For instance if a 20MHz signal is transmitted over a cellular network using a single carrier modulation scheme, the symbol period of 50ns would be significantly smaller than the channel delay spread which is typically about 2 μ s or 3 μ s in outdoor environments. This would lead to severe signal distortion due to the interference between the transmitted signals known as ISI.

OFDM is a key wireless broadband technology used to allow the transmission of wideband signals, thus permitting high data rates, while preventing signal distortion due to the frequency selective nature of the channel. To prevent the occurrence of ISI, OFDM uses a Multi-Carrier Modulation (MCM) scheme with each sub-carrier transmitting data at a low rate thus ensuring that its symbol period is much greater the channel delay spread as shown in the (2.19) where T_s is the symbol period and T_d is the channel delay spread.

$$T_s \gg T_d \quad (2.19)$$

An OFDM signal with a bandwidth of B and with N sub-carriers, would have a subcarrier spacing F_0 as shown in the below (2.20),

$$F_0 = \frac{B}{N} \quad (2.20)$$

And the transmitted signal at the k th subcarrier can be represented by the below (2.21),

$$y_k[t] = X_k e^{j2\pi k F_0 t} \quad (2.21)$$

Thus, the complete OFDM signal would be a sum of the above signals for all the subcarriers and can be represented by the (2.22).

$$y[t] = \sum_{-N/2}^{N/2} X_k e^{j2\pi k F_0 t} \quad (2.22)$$

In order to avoid the need to have guard bands in the frequency domain between signals transmitted in the different subcarriers, the subcarrier frequencies are selected such that they are orthogonal to each other and this allows us to achieve high spectral efficiency. Also, this orthogonality allows us to use the IFFT and FFT algorithms at the transmitter and receiver respectively to modulate and demodulate the signals. This eliminates the need for N local oscillators correctly spaced at the different subcarriers at the transmitter and receiver.

Though the data signals transmitted in each of the subcarriers have a symbol period much larger than the channel delay spread, these signals would still suffer from minor distortion due to the interference between signals transmitted in different OFDM symbol blocks known as Inter-Block Interference (IBI). For instance, if the following signal $x[t]$, shown in (2.23), consists of the signals from two OFDM blocks $x_1[t]$ and $x_2[t]$ and if this signal is transmitted via the channel $h[t]$ of length L , shown in (2.24), the received signal $y[t]$ would be as defined by (2.25). Thus, the received signal $y[0]$, as shown in (2.26), includes signal components from OFDM blocks $x_1[t]$ and $x_2[t]$ resulting in IBI.

$$x[t] = (x_1[0], x_1[1], x_1[N-1], x_2[0], x_2[1], x_2[N-1]) \quad (2.23)$$

$$h[t] = (h[0], h[1], h[L-1]) \quad (2.24)$$

$$y[t] = h[t] * x[t] \quad (2.25)$$

$$y[0] = h[0]x_1[0] + h[1]x_2[N-1] + \dots + h[L-1]x_2[N-L+1] \quad (2.26)$$

To prevent the occurrence of IBI in an OFDM signal cyclic prefixes are used. This involves prefixing each OFDM symbol block with the last L symbols within this symbol block, where L is the length of the channel. If we introduce a cyclic prefix to the signal $x[t]$, we would get a signal as defined by (2.27). If this signal is transmitted via the channel $h[t]$ of length L , the received signal $y[t]$ would be as defined by (2.28). Thus, the received signal $y[0]$, as shown in (2.29), no longer includes signal components from OFDM block $x_2[t]$ thus eliminating in IBI. Examining the convolution of $h[t]$ and $x_{cyclic}[t]$ shows that it is essentially a circular convolution of $h[t]$ and $x[t]$ as shown in (2.30) where $\otimes$ is the circular convolution operator. The Discrete Fourier Transform (DFT) of the circular convolution in the time domain is equivalent to the product of the DFT coefficients in the frequency domain and hence (2.30) can be represented by (2.31).

$$x_{cyclic}[t] = (x_1[N-L], \dots, x_1[N-1], x_1[0], x_1[1], x_1[N-1]) \quad (2.27)$$

$$y[t] = h[t] * x_{cyclic}[t] \quad (2.28)$$

$$y[0] = h[0]x_1[0] + h[1]x_1[N-1] + \cdots + h[L-1]x_1[N-L+1] \quad (2.29)$$

$$y[t] = h[t] \otimes x[t] \quad (2.30)$$

$$Y[k] = H[k] \times X[k] \quad (2.31)$$

Examining (2.31), we can see that each subcarrier experience flat fading thus completely removing any ISI in the signal. The effect of $H[k]$ can be easily removed from the signal by using a single tap equalizer.

3. Real-Time Channel Estimator (RTCE)

In 2002, the Federal Communication Commission (FCC) allowed the unlicensed operation between 3.1 GHz and 10.6 GHz for Ultra wideband (UWB) communication, using an appropriate wideband signal format with a low EIRP level (-41.3dBm/MHz) for WPAN. The two most commonly used techniques to generate UWB signals for PAN are Multicarrier OFDM transmission and Impulse Radio (IR) UWB transmission. [1]

OFDM allows us to combat inter-symbol-interference (ISI) very effectively as it uses several subcarriers, each of which has a symbol period much larger than the channel delay spread, to transmit wideband signals. However, as explained in [1], this method further exacerbates the significant power limitations imposed by the FCC, as each subcarrier would receive only a small fraction of the total power.

As IR-UWB systems do not use subcarriers for transmission, they do not suffer from the above mentioned problem, while still complying with the FCC power restrictions. However, as explained in section 1.5, IR-UWB signals typically suffer from signal distortion due to ISI. Equalizers could be used to address the problem of signal distortion, however, the effects of any destructive ISI introduced by the channel would still be present, exacerbating the power restrictions imposed by FCC.

In this chapter, we propose a different approach that helps to address this issue of destructive ISI by determining the channel transfer function in real-time and using this information to improve the effective throughput of the wireless communication system.

3.1. Destructive ISI in IR-UWB Systems when using Equalizers

In an Impulse Radio UWB communication system, that transmits data at symbol rate of F_s , we would recover the transmitted data from a received IR signal by removing the carrier frequency, F_c , component of the signal from the received signal using a mixer, filtering this signal using a Low Pass Filter (with a cut-off frequency between F_s and F_c) and equalizing this signal as shown in Figure 3.1 below.

As the received signal is a convolution of the transmitted signal and the multipath channel, Low-Pass Filtering(LPF) the received signal would result in filtering the transmission channel as well.

Figure 3.1: Recover Data Signal from IR Signal using an Equalizer

If the wireless multipath channel, exhibits destructive ISI, filtering the transmission channel would result in effectively attenuating the overall channel gain.

We illustrate this concept with an example. A pilot impulse signal, with a pulse duration $T_s/2$, is transmitted via a wireless multipath channel and the received power delay profile of the channel is as shown in Table 3.1. If we LPF the received power delay profile with a cut-off frequency above $1/T_s$ but below F_c , the effective power delay profile of the channel would be as shown in Table 3.2. As we can observe by comparing the power delay profiles shown in Table 3.1 and Table 3.2, filtering the channel may result in effectively attenuating the overall channel gain.

Gain	0.927	-0.465	0.460	-0.230	0.688	-0.344	0.460	-0.230	0.227	-0.114
Time	$0 T_s$	$0.5 T_s$	$1 T_s$	$1.5 T_s$	$2 T_s$	$2.5 T_s$	$3 T_s$	$3.5 T_s$	$4 T_s$	$4.5 T_s$

Table 3.1: Channel Power Delay Profile

Gain	0.462	0.230	0.344	0.230	0.113
Time	$0 T_s$	$1 T_s$	$2 T_s$	$3 T_s$	$4 T_s$

Table 3.2: Filtered Channel Power Delay Profile

3.2. Overview of RTCE

As explained in Chapter 1, if we transmit a wireless signal via a multipath channel, with a symbol period less than the channel delay spread, the frequency selective fading nature of the

channel would cause severe signal distortion due to ISI. As ISI is introduced by the channel, if we have the channel impulse response function, we would be able to undo its effects completely.

A multipath channel can be modelled as a linear time invariant system, for brief periods of time, by the (3.1) where K is the number of multipath components, τ_k is the excess delay and $a_k e^{j\phi_k}$ is the complex gain of the k th multipath component.

$$h(t) = \sum_{k=1}^K a_k e^{j\phi_k} \delta(t - \tau_k) \quad (3.1)$$

$$y(t) = x(t) * h(t) \quad (3.2)$$

If a signal is transmitted via this channel, the received signal would be the convolution of the input signal, $x(t)$, with the multipath channel, $h(t)$, as shown in (3.2). As $h(t)$ consists of only delta functions, which only introduce phase shifts to the input signal, the highest frequency component of the received signal would be the same as that of the transmitted signal. Thus, if we transmitted a data signal through this channel, the highest frequency component of the received signal would be the same as that of the data signal.

To determine the Channel Impulse Response (CIR) function, we would typically transmit a delta function as defined by (3.3) through the channel. However, this would provide us with the CIR for the entire bandwidth of the channel, but in order to undo the distortion of a data signal due to ISI we would only need the CIR for the bandwidth of the data signal.

$$\delta(x) = \begin{cases} +\infty, & x = 0 \\ 0, & x \neq 0 \end{cases} \quad (3.3)$$

Hence, if we transmit a pilot signal with a bandwidth matching that of the data pulse that would be sent later, as shown by (3.4), using the received signal and the pilot we should be able to compute the CIR function. Now, by using (3.2), we could use this CIR function to undo the effects of ISI in the data signal transmitted later.

$$\delta_{RTCE}(x) = \begin{cases} 1, & 0 < x \leq \text{pulse period} \\ 0, & \text{otherwise} \end{cases} \quad (3.4)$$

3.2.1. Undoing the Effects of ISI

Traditionally, given the impulse response of a system and the system input, we would derive the system output by convolving the input with the impulse response as described by (3.2). Likewise, given the system output, we would derive the corresponding system input by deconvolving output with the impulse response. This is normally done by apply the Fourier Transform algorithm to (3.2), which would give us (3.5) and dividing $Y(f)$ by $H(f)$ would give us $X(f)$.

$$Y(f) = X(f)H(f) \quad (3.5)$$

In the discrete time domain, the convolution operation between x and h would be as defined by (3.6). Thus, if x had a length of L and h had a length of M the resulting convolution would have a length of $L+M-1$.

$$(x * h)[n] = \sum_{m=-M}^M x[n-m]h[m] \quad (3.6)$$

Examining (3.6) carefully, it can be seen that the convolution operation can be represented as a set of simultaneous equations. Hence, to recover the message sequence of length L based on the received signal and the CIR we would need only L set of equations as there are only L unknowns. We illustrate this with a simple example below.

If the channel impulse response $h[n]$ is: [0.9, 0.7, 0.5, 0.5] and the transmitted signal $x[n]$ is [0.8, -0.5, 0.7, -0.4] the received signal $y[n]$ would be [0.7200, 0.1100, 0.6800, 0.2800, -0.1800, 0.1500, -0.2000]. If the receiver knows the channel impulse response function using the channel estimation technique described above, we would be able to compute the values of $x[0]$ to $x[3]$ using the values of $y[0]$ to $y[3]$ by solving the equations below.

$$y[0] = x[0]h[0] \quad (3.7)$$

$$y[1] = x[1]h[0] + x[0]h[1] \quad (3.8)$$

$$y[2] = x[2]h[0] + x[1]h[1] + x[0]h[2] \quad (3.9)$$

$$y[3] = x[3]h[0] + x[2]h[1] + x[1]h[2] + x[0]h[3] \quad (3.10)$$

In the above example, as the 1st multipath is the strongest channel multipath, we recover the transmitted signal using the 1st multipath and remove the ISI contributions from the other multipath components. This ensures the “corrected signal” (i.e. the output of the RTCE) would have the highest SNR. If the channel impulse response $h[n]$ was [0.5, 0.9, 0.7, 0.5] instead, we would need to recover the transmitted signal using the 2nd multipath as it is the strongest multipath. However, in-order to use 2nd multipath to recover the signal, we would need an equation relating x , y and h without the 1st multipath component. This can be achieved by ensuring we do not transmit a symbol immediately after the last data symbol (i.e. at $x=5$ we transmit 0). We would now be able to compute the values of $x[0]$ to $x[3]$ using channel impulse response by solving the equations below. As IR-UWB signals inherently need to transmit zeros, this zero value could be positioned over the time unit $x[5]$.

$$y[2] = x[2]h[0] + x[1]h[1] + x[0]h[2] \quad (3.11)$$

$$y[3] = x[3]h[0] + x[2]h[1] + x[1]h[2] + x[0]h[3] \quad (3.12)$$

$$y[4] = x[4]h[0] + x[3]h[1] + x[2]h[2] + x[1]h[3] + x[0]h[4] \quad (3.13)$$

$$y[5] = x[5]h[0] + x[4]h[1] + x[4]h[2] + x[2]h[3] + x[1]h[4] + x[0]h[5] \quad (3.14)$$

(3.14) can be simplified to (3.15) as $x[5]=0$ and $h[5]=0$.

$$y[5] = x[4]h[1] + x[4]h[2] + x[2]h[3] + x[1]h[4] \quad (3.15)$$

3.2.2. Adapting to Channel Changes

Although a multipath channel can be modelled a linear time-invariant system for brief periods for time, due to the movement of the objects within the multipath channel, its impulse response would change over time and it is important we capture these changes to ensure the optimal performance of the RTCE.

As explained in section 3.2.1, the convolution of the transmitted signal with the wireless channel can be represented by a set of simultaneous equations relating x , y and h . By solving these simultaneous equations, we recover the message block, x . This computed value of x would not exactly match the value of the transmitted signal though due to the presence of noise in the channel. However, by passing the recovered message block through a symbol detector (QPSK, 8-PSK, etc) we would be able to recover the exact message block sent and

thus remove the effects of the channel noise. We could then try to re-compute the value of h based on the value of x obtained from the symbol detector. This would ensure we are able to track any changes in the channel. We illustrate this concept using an example.

Given a message sequence, x , is transmitted via a channel, h , of length L and we received a signal, y , we can write the following set of simultaneous equations relating x , y and h as shown in (3.16) to (3.19). We can then solve for h using the values of the received signal y and the symbol detector output x . This would now give us the current channel impulse response.

$$y[L] = x[L]h[0] + x[L-1]h[1] + \cdots + x[1]h[L-1] + x[0]h[L] \quad (3.16)$$

$$y[L+1] = x[L+1]h[0] + x[L]h[1] + \cdots + x[2]h[L-1] + x[1]h[L] \quad (3.17)$$

.

.

.

$$y[2L-1] = x[2L-1]h[0] + x[2L-2]h[1] + \cdots + x[L-1]h[L-1] + x[L-2]h[L] \quad (3.18)$$

$$y[2L] = x[2L]h[0] + x[2L-1]h[1] + \cdots + x[L]h[L-1] + x[L-1]h[L] \quad (3.19)$$

3.2.3. Mitigating the effects of Channel Noise

Like all the other technologies used to reduce the effects of small scale fading, as described in Chapter 2, the channel noise can affect the reliable operation of the RTCE. As described above, the RTCE can be used to recover a message sequence from a signal corrupted by ISI and it is also capable of adapting to changing channel conditions. Both these processes can be affected by channel noise though and in the section we describe techniques that can be used to reduce the effects of noise.

3.2.3.1. Recovering the Message Sequence

As explained in section 3.2.1 to recover a message sequence of length L , we would need L simultaneous equations relating x , y and h . This can be represented by the matrix equation

3.20 and 3.21 where matrixes $\mathbf{H}$, $\mathbf{X}$ and $\mathbf{Y}$ correspond to the coefficients of h , x and y respectively.

To determine the value of $\mathbf{X}$ given matrixes $\mathbf{H}$ and $\mathbf{Y}$ we would need apply the inverse of matrix $\mathbf{H}$ to matrix $\mathbf{Y}$ as shown in (3.22).

$$\mathbf{H}\mathbf{X} = \mathbf{Y} \quad (3.20)$$

$$\begin{pmatrix} h[0] & 0 & \cdots & \cdots & 0 & 0 \\ h[1] & h[0] & \cdots & \cdots & 0 & 0 \\ \vdots & \cdot & \ddots & & \vdots & \vdots \\ \vdots & \cdot & \ddots & \ddots & \vdots & \vdots \\ h[L-2] & h[L-3] & \cdots & \cdots & h[0] & 0 \\ h[L-1] & h[L-2] & \cdots & \cdots & h[1] & h[0] \end{pmatrix} \begin{pmatrix} x[0] \\ x[1] \\ \vdots \\ \vdots \\ x[L-2] \\ x[L-1] \end{pmatrix} = \begin{pmatrix} y[0] \\ y[1] \\ \vdots \\ \vdots \\ y[L-2] \\ y[L-1] \end{pmatrix} \quad (3.21)$$

$$\mathbf{X} = \mathbf{H}^{-1}\mathbf{Y} \quad (3.22)$$

Hence, to compute matrix $\mathbf{X}$, matrix $\mathbf{H}$ needs to be non-singular. However, matrix $\mathbf{H}$ being merely non-singular using does not guarantee that (3.22) would be able to recover the values of $\mathbf{X}$ within a reasonable degree of accuracy. If the determinant of matrix $\mathbf{H}$ is small, any perturbation in matrix $\mathbf{Y}$ due to the presence of channel noise would significantly alter the computed values for $\mathbf{X}$ due to the amplification of $\mathbf{Y}$ by the inverse of the determinant. Ideally, we would like the matrix $\mathbf{H}$ to have a large value for the determinant. This can be done by modelling matrix $\mathbf{H}$ such that the diagonal is the strongest channel multipath. In this case, to ensure there is no ISI between the message blocks, if the strongest multipath is in position P , we would need to transmit zeros for the $P-1$ units of time after each message block, as described in section 3.2.1. This would ensure the matrix $\mathbf{H}$ has the larger determinant value for a given multipath channel, and this would provide better tolerance to channel perturbations. This can be represented by the below matrix (3.23). As IR signals inherently need to transmit zeros, these zeros can be positioned over time units 0 to $P-1$.

$$\begin{pmatrix}
 h[P] & h[P-1] & \dots & \dots & h[P-L+2] & h[P-L+1] \\
 h[1] & h[P] & \dots & \dots & h[P-L+3] & h[P-L+2] \\
 \vdots & \vdots & \ddots & & \vdots & \vdots \\
 \vdots & \vdots & \ddots & \ddots & \vdots & \vdots \\
 h[P+L-2] & h[P+L-3] & \dots & \dots & h[P] & h[P-1] \\
 h[P+L-1] & h[P+L-2] & \dots & \dots & h[P+1] & h[P]
 \end{pmatrix}
 \begin{pmatrix}
 x[P] \\
 x[P+1] \\
 \vdots \\
 \vdots \\
 x[P+L-1] \\
 x[P+L]
 \end{pmatrix}
 =
 \begin{pmatrix}
 y[0] \\
 y[1] \\
 \vdots \\
 \vdots \\
 y[L-2] \\
 y[L-1]
 \end{pmatrix}
 \quad (3.23)$$

3.2.3.2. Determining Channel Impulse Response

As described in section 3.2.2, in order to detect changes in the channel, we would need to recompute the values of h at fixed intervals in time. This can be done using the information obtained from the symbol detector (which should be the same as the transmitted signal, assuming correct detection) and the received signal. We could represent relationship between these terms in the below matrix form, shown by (3.24) and (3.25), where the transmitted signal is a square matrix $\mathbf{X}$ and the channel and the received signal are the column matrixes $\mathbf{H}$ and $\mathbf{Y}$ respectively.

$$\mathbf{X}\mathbf{H} = \mathbf{Y} \quad (3.24)$$

$$\begin{pmatrix}
 x[M-1] & x[M-2] & \dots & \dots & x[1] & x[0] \\
 x[M] & x[M-1] & \dots & \dots & x[2] & x[1] \\
 \vdots & \vdots & \ddots & & \vdots & \vdots \\
 \vdots & \vdots & \ddots & \ddots & \vdots & \vdots \\
 x[2M-3] & x[2M-4] & \dots & \dots & x[M-1] & x[M-2] \\
 x[2M-2] & x[2M-3] & \dots & \dots & x[M] & x[M-1]
 \end{pmatrix}
 \begin{pmatrix}
 h[0] \\
 h[1] \\
 \vdots \\
 \vdots \\
 h[M-2] \\
 h[M-1]
 \end{pmatrix}
 =
 \begin{pmatrix}
 y[0] \\
 y[1] \\
 \vdots \\
 \vdots \\
 y[M-1] \\
 y[M-2]
 \end{pmatrix}
 \quad (3.25)$$

Thus, to determine the value of $\mathbf{H}$ given matrixes $\mathbf{X}$ and $\mathbf{Y}$, we would need apply the inverse of matrix $\mathbf{X}$ to matrix $\mathbf{Y}$ as shown in (3.26).

$$\mathbf{H} = \mathbf{X}^{-1}\mathbf{Y} \quad (3.26)$$

As explained in section 3.2.3.1, to ensure the computed value of $\mathbf{H}$ is not affected by any perturbation in matrix $\mathbf{Y}$ due to the presence of channel noise, we would need to able to measure singularity of matrix $\mathbf{X}$ and choose a set of values for matrix $\mathbf{X}$ such that it is well

conditioned. Using the determinant is not an accurate measure of singularity though as shown by the below example. The determinant of the matrix shown in (3.27), is 3.1250×10^{-22} which makes this matrix appear to be singular. However, as the below matrix is a multiple of the identity matrix, it is non-singular.

We can obtain a better estimate of the singularity of a matrix using the Singular Value Decomposition (SVD) technique. SVD, as shown by (3.28), involves factoring a matrix $A_{[m \times n]}$ into the below form, where $U_{[m \times r]}$ is the left singular matrix, $V_{[n \times r]}$ is the right singular matrix and $\Sigma_{[r \times r]}$ is a matrix of the singular values.

$$\begin{pmatrix} 0.00005 & 0 & 0 & 0 & 0 \\ 0 & 0.00005 & 0 & 0 & 0 \\ 0 & 0 & 0.00005 & 0 & 0 \\ 0 & 0 & 0 & 0.00005 & 0 \\ 0 & 0 & 0 & 0 & 0.00005 \end{pmatrix} \quad (3.27)$$

The diagonal of matrix $\Sigma_{[r \times r]}$ indicates the percentage contribution of each of the columns terms to the matrix $A_{[m \times n]}$. Thus, it can be used to determine if matrix A is well conditioned by taking the ratio of the largest singular value of X and the smallest.

$$A_{[m \times n]} = U_{[m \times r]} \Sigma_{[r \times r]} (V_{[n \times r]})^T \quad (3.28)$$

Using this technique we would be able to choose a set of values for matrix X such that it is well conditioned and ensure the computed value of H is not affected by any perturbation in matrix Y due to the presence of channel noise.

As this computed value of H is used to determine the transmitted signal from a received signal, reducing the effects of the channel noise further on the matrix H would be desirable as it would allow us to more accurately predict the corresponding transmitted signal of a given received signal. (i.e. increase the SNR of the channel estimate). This can be achieved by adopting the following de-noising technique.

If the channel noise is purely due to thermal noise, which can be described as a zero mean Gaussian random process, the effects of the noise on a signal can be reduced by taking the

mean of a sufficiently large number of samples of the signal. We could apply this technique when computing the value of matrix $\mathbf{H}$ by repeating the channel estimation process multiple times and taking a mean of the computed values.

If the channel noise is due to the noise generated by manmade devices such as microwave ovens, television sets, etc removing the effects of this noise would be a more complex process. The underlying probability density function of such noise sources are difficult to model, however the central limit theorem in probability theory suggests that when several independent random variables are added, their summation and mean will tend to have a Normal distribution even if the underlying random variables themselves are not normally distributed.

(3.29) and (3.30) show the Bienaymé formula which suggests that as the sample size, n , of a random variable X increases the standard deviation, σ , of the probability density function of $\bar{X}$ is decreased by a factor of $\sqrt{n}$. Thus, we could estimate the channel noise by taking sufficiently large number of measurement of the channel noise, computing its mean and then subtract this mean value from the value of matrix $\mathbf{H}$ computed by repeating the channel estimation process multiple times and taking a mean of the computed values. As both the noise estimation process and the channel estimation process are repeated a large number of times the standard deviation of the mean value of the noise would be small in both the channel estimate and the noise estimate. Having a smaller noise variance would allow is to remove the effects of the noise on the signal with greater accuracy and hence recover a signal with higher SNR.

$$Var(\bar{X}) = Var\left(\frac{1}{n} \sum_{i=1}^n X_i\right) = \frac{1}{n^2} \sum_{i=1}^n Var(X_i) = \frac{\sigma^2}{n} \quad (3.29)$$

$$Standard\ Deviation(\bar{X}) = \frac{\sigma}{\sqrt{n}} \quad (3.30)$$

3.2.4. Novelty of the RTCE in Comparison with Existing Technologies

As described in Chapter 2, ISI can be dealt with by avoiding it, mitigating it or by reducing its effects.

ISI could be avoided using the OFDM transmission, as explained in 2.1.4, however, in a multitone modulated signal, the transmitter would need to be capable of supporting a high peak-to-average power ratio in the range of 8-9 dB which would mean that the average transmitted power would be about 8-9 dB below the maximum deliverable power of the transmitter. This would in turn result in significant power loss.

Using the DDE technique on a single tone modulated signal would help to address the above mentioned problems, as it would have a lower peak-to-average power ratio and spectral nulls would have a less detrimental effect.

In the DDE technique though, if a single tone modulated signal is transmitted via a multipath channel of length L , it requires training data of length N to be prefixed to each of the message block and the value of N needs to be larger than L . This would greatly reduce the effective data transmission rate as shown in (2.4).

As explained in 2.1.3, equalizers could be used to mitigate the effects of ISI. However, as described in 3.1, an IR-UWB system, that utilizes an equalizer to mitigate the effects of ISI, and operates in a channel that exhibits destructive ISI would still suffer the effects of signal attenuation as a result of this destructive ISI.

The effects of ISI could be reduced using the Generalized Shift-Window Approach for UWB ISI Reduction as explained in 2.1.1. The generalized Shift-Window approach, provides a good performance improvement without the complexity of the ISI avoidance or mitigation.

IR-UWB communication systems typically transmit very short pulses with a low duty cycle. [17]. The RTCE fundamentally works by transmitting a pilot signal with a bandwidth matching that of the data pulse that would be sent later. This pilot signal would then provide the CIR for the entire bandwidth of the channel as shown in (3.31) rather than just the bandwidth of the underlying data signal as shown in (3.32). With this additional information about the entire bandwidth of the channel the RTCE is able to mitigate the destructive ISI introduced by the channel.

$$\text{Bandwidth of IR - UWB Channel} = \frac{1}{\text{Pulse Period}} \quad (3.31)$$

$$\text{Bandwidth of Data Signal} = \frac{1}{\text{Symbol Period}} \quad (3.32)$$

The RTCE initially determines the channel coefficients by transmitting a pilot signal with a bandwidth matching that of the data pulse that would be sent later, as shown by (3.4), and uses the received signal and the pilot to compute the CIR function. (Explained in section 3.2)

In order to detect changes in the channel, we would need to re-compute the channel coefficients at fixed intervals in time. This is done using the information obtained from the symbol detector (which should be the same as the transmitted signal, assuming correct detection) and the received signal. (Explained in section 3.2.2)

As explained in section 3.2.3.1, for the RTCE to be able to minimize the effects of channel noise when recovering the message sequence, the matrix $\mathbf{H}$ used to recover the message sequence would need to have the strongest channel multipath as its diagonal. This can be achieved by removing the effects of the pre-cursor ISI components by transmitting zeros during this period. Hence, if the strongest multipath is in position P, we would need to transmit zeros for P-1 units of time after each message block. As IR signals inherently need to transmit zeros, as long as (3.33) is satisfied, the effects of the pre-cursor ISI components can be removed.

$$\text{Duration of Precursor ISI} < \text{Duration of Postfix Zeros} \quad (3.33)$$

Figure 3.2: Sample Power Delay Profile of an Indoor Wireless Channel [21]

Figure 3.2 above shows a sample power delay profile of an indoor wireless channel. In this example, the strongest multipath component arrives at around 5ns. Hence, in this case, the multipath components that arrive before 5ns would contribute to the precursor ISI.

IR-UWB signals transmit information in very short pulses with a very low duty cycle.[17] Figure 3.3 below shows a sample IR-UWB signal, where P is the Pulse Period of the impulse radio signal, Z is the duration of the postfix zeros transmitted and $P + Z$ is the Symbol Period. (3.34) shows the information throughput of this IR-UWB signal.

Figure 3.3: Sample IR-UWB Signal

$$\text{Information Throughput} = \frac{P}{P + Z} \quad (3.34)$$

Hence, an IR-UWB signal with a Pulse period of 2ns and a duty cycle of 10% would have a symbol period of 20ns and would need to transmit zeros for 18ns. If this IR-UWB signal were to be transmitted via the indoor wireless channel shown in Figure 3.2, the effects of the precursor ISI components would be removed completely by the transmitted zeros as it satisfies (3.33)

As explained in 2.1.2, the signal structure of the DDE is made up of one block Known Symbols (training data of Length N) followed by one block of Unknown Symbols (message of Length M) and together they form 1 Frame (Shown in Figure 2.2).

In the DDE technique, if a single tone modulated signal is transmitted via a multipath channel of length L , it requires training data of length N to be prefixed to each of the message blocks and the value of N needs to be larger than L .

The channel length of the indoor wireless channel shown in Figure 3.2 is about 250ns. Now, if we transmitted an IR-UWB signal with a Pulse period of 2ns and a duty cycle of 10% using

the DDE technique via the indoor wireless channel shown in Figure 3.2, the effective data rate would be 10% of (2.4). In comparison, the RTCE does not require training data to be prefixed to each of the message blocks and hence does not suffer the loss in data transmission as a result of transmitting the training data as shown in (2.4).

In comparison to the Equalizers, the RTCE is able to better combat the destructive ISI of the channel and in comparison to the Generalized Shift-Window Approach for UWB the RTCE would be able better mitigate the effects of ISI.

4. Simulation Results

In this section, we review the performance of the RTCE in mitigating the effects of AWGN and as well as reducing the effects of destructive ISI for IR-UWB Systems.

4.1. Mitigating the Effects of an AWGN Channel

As explained in section 3.2.3.2, in an AWGN channel, the effects of noise on a signal can be reduced by taking the mean of a sufficiently large number of samples of the signal. Using this technique, in most of our simulations, we have assumed that the signal used by the RTCE to estimate the channel has a E_b/N_0 value of 25dB.

In this section we have computed, through simulation, the required sample size to ensure that the mean value of a signal transmitted via an AWGN channel with a given E_b/N_0 value would have an effective E_b/N_0 value of 25dB. Figure 4.1 below shows the histogram of the probability distribution for an AWGN channel with a E_b/N_0 value of 25dB. It has a standard deviation of about 0.04. By examining this distribution, we could say that there is 80% probability that the noise voltage would have a value less than -0.01 or greater than 0.01. Hence, an AWGN channel with any E_b/N_0 value, but with a sample mean value between -0.01 and 0.01, can be said to have an effective E_b/N_0 of about 25dB.

Figure 4.1: Probability distribution for an AWGN channel with a E_b/N_0 value of 25dB

By running simulations, we have determined the standard deviation values of AWGN channels with various Eb/No values as shown in Table 4.1.

Eb/No (dB)	Standard Deviation
15	0.1257
10	0.2236
5	0.3974
0	0.7074

Table 4.1: The Standard Deviation Values of AWGN channels with various Eb/No values

Using the Student's t-distribution and the T-Test shown in (4.1), where $\bar{X}$ is the sample mean and μ is the population mean, σ is the sample standard deviation and n is the sample size, we computed the Confidence Level that the mean value of a given sample size is between -0.01 and 0.01 as shown in Table 4.2. For our calculations we have taken $\bar{X} = 0.01$, $\mu = 0$ and σ to be the values shown in Table 4.1.

$$t = \frac{\bar{X} - \mu}{\sigma/\sqrt{n}} \quad (4.1)$$

Eb/No (dB)	Sample Size					
	100,000	50,000	10,000	5,000	2,000	1,000
15	Not Required	Not Required	Not Required	100%	99.96%	98.80%
10	Not Required	Not Required	100%	99.84	95.44%	84.24%
5	Not Required	100%	98.81%	92.48	73.94%	57.36%
0	100%	99.84%	84.25%	68.24%	47.27%	34.50%

Table 4.2: Confidence Level that the mean value of a given sample size is between -0.01 and 0.01

Based on the above results, we can confirm that with a sufficient sample size, the mean of a signal corrupted by AWGN channel with a Eb/No value between 0dB to 15dB can be de-noised to have an effective Eb/No value of 25dB.

4.2. Mitigating the Effects of Destructive ISI

In this section we compare the performance of the RTCE with the other commonly used equalizers (Linear Equalizer, Zero Forcing Equalizer, Decision Feedback Equalizer and Ideal MLSE) and verify its effectiveness in combating destructive ISI introduced by the channel.

Table 4.3 below shows the configuration parameters of the Linear Equalizer, DFE and Ideal MLSE.

Equalizer Type	Parameters
Linear Equalizer	Number of Weights: 31, RLS algorithm for first data block, LMS algorithm for remaining data blocks
DFE	Number of Feedforward Weights: 15, Number of Feedback Weights: 15, RLS algorithm for first data block, LMS algorithm for remaining data blocks
Ideal MLSE	K= 5, Number of trellis states: 16, Traceback length = 30

Table 4.3: Parameters of the Equalizers

4.2.1. Simulation Model

The Simulation model was written in MATLAB and it essentially compares the performance of the different type of equalizers with the RTCE. It takes the channels described in Table 4.4, Table 4.5, Table 4.6 and Table 4.7 as input and consists of the following modules.

We have written a module that generates a noisy BPSK signal of 10,000 Symbols. As we are running simulating for Eb/No values between 0dB and 14dB this module generates 15 such signals. Each of these signals are then channel filtered by the channels described by Table 4.4, Table 4.5, Table 4.6 and Table 4.7.

We have also written a module that equalizes each of these noisy channel filtered signals and applies the different types of equalization techniques as well as the RTCE to the signal. We

used the MATLAB built in functions to simulate the of equalization techniques and wrote a separate module to simulate the RTCE. This RTCE module was written as described in Chapter 3. The output of each of these ISI mitigating techniques is then compared with the transmitted signal to compute the BER of this technique. This process is then repeated for each of the E_b/N_0 values between 0dB and 14dB.

4.2.2. Verify Simulation Model

To validate the logic used to implement the RTCE, we simulated the transmission of an IR signal via a flat fading channel, not exhibiting any destructive ISI and attempted to recover the data signal using the RTCE as well as the Linear Equalizer, Zero Forcing Equalizer, Decision Feedback Equalizer and Ideal MLSE. Figure 4.2 below shows the performance of the different equalizers over various SNR values when attempting to recover a signal transmitted via a Flat Fading Channel. As expected, the performance of the RTCE is very similar to that of the other equalizers and the Ideal BPSK BER which authenticates the simulation model of the RTCE.

Figure 4.2: Performance Comparison of different Equalizers in a Flat Fading Channel

4.2.3. Multipath Channel Exhibiting Destructive ISI

We simulated the transmission of IR signals via the multipath channel, which exhibits destructive ISI, as described in Table 4.4, with a symbol rate of $1/2T_s$ and pulse period of T_s and attempted to recover the data signal using the RTCE. We describe the filtered channel in Table 4.5 as exhibiting destructive ISI as when the multipath channels combine, the overall channel gain is attenuated by one or more of the multipath channels.

To process the same IR signals using Equalizers, we needed to filter this signal above the symbol rate, prior to equalization and hence, we simulated the transmission via the channel described in Table 4.5 and attempted to recover the data signal using a Linear Equalizer, Zero Forcing Equalizer, Decision Feedback Equalizer and Ideal MLSE.

Gain	0.927	-0.465	0.460	-0.230	0.688	-0.344	0.460	-0.230	0.227	-0.114
Time	0 T_s	0.5 T_s	1 T_s	1.5 T_s	2 T_s	2.5 T_s	3 T_s	3.5 T_s	4 T_s	4.5 T_s

Table 4.4: Channel Power Delay Profile of a Multipath Channel Exhibiting Destructive ISI

Gain	0.462	0.230	0.344	0.230	0.113
Time	0 T_s	1 T_s	2 T_s	3 T_s	4 T_s

Table 4.5: Filtered Channel Power Delay Profile of a Multipath Channel Exhibiting Destructive ISI

Figure 4.3 shows the Channel Frequency Response of the multipath channel described by Table 4.4 and Table 4.5.

Figure 4.4 shows the performance of the different equalizers over various E_b/N_0 values when attempting to recover a signal transmitted via a Multipath Channel Exhibiting Destructive ISI. From this figure, we can observe that the RTCE performs better than the ZFE, Linear Equalizer and DFE. This is because, the RTCE is better able to mitigate the destructive ISI introduced by the channel. This is due to the RTCE having the CIR for the entire bandwidth of the channel as shown in (3.31) rather than just the bandwidth of the underlying data signal as shown in (3.32). With this additional information about the entire bandwidth of the channel the RTCE is able to mitigate the destructive ISI introduced by the channel.

Figure 4.3: Channel Frequency Response of a Multipath Channel Exhibiting Destructive ISI

Figure 4.4: Performance Comparison of different Equalizers in a Multipath Channel Exhibiting Destructive ISI

The Ideal MLSE equalizer performs the best but it come at the price of higher computational complexity. The Ideal MLSE using the Viterbi algorithm requires $L2^{K-1}$ operations to decode

the message sequence of length L and needs to have a path memory of $h2^{K-1}$, where h is the length of the path history.[2]

In the simulations we have run, the Ideal MLSE has a K value of 5 and hence we would need $16 \times L$ operations to recover a message sequence of length L while the RTCE would require $10 \times L$ operations, as the Channel Length for RTCE was 10, to recover a message sequence of length L . Also, the MLSE requires a large path memory of $h2^{K-1}$ which is not required by the RTCE. Hence, we can conclude that the computational complexity of the RTCE would be less than that of the Ideal MLSE.

4.2.4. Multipath Channel Not Exhibiting Destructive ISI

We simulated the transmission of IR signals via a multipath channel, not exhibiting destructive ISI, as described in Table 4.6 with a symbol rate of $1/2T_s$ and pulse period of T_s and attempted to recover the data signal using the RTCE. We describe this channel as not exhibiting destructive ISI as when the multipath channels combine, none of the multipath channels attenuate the overall channel gain.

To process the same IR signals using Equalizers, we needed to filter this signal above the symbol rate, prior to equalization and hence, we simulated the transmission via the channel described in Table 4.7 and attempted to recover the data signal using a Linear Equalizer, Zero Forcing Equalizer, Decision Feedback Equalizer and Ideal MLSE.

The channel described in Table 4.6 is the same as that in Table 4.4 without the destructive ISI components. The filtered channel in Table 4.7 is the same as that in Table 4.5 to ease the comparison of the performance of the RTCE between a channel that exhibits Destructive and Non-Destructive ISI.

Gain	0.462	0	0.230	0	0.344	0	0.230	0	0.113	0
Time	$0 T_s$	$0.5 T_s$	$1 T_s$	$1.5 T_s$	$2 T_s$	$2.5 T_s$	$3 T_s$	$3.5 T_s$	$4 T_s$	$4.5 T_s$

Table 4.6: Channel Power Delay Profile of a Multipath Channel Not Exhibiting Destructive ISI

Gain	0.462	0.230	0.344	0.230	0.113
Time	0 T_s	1 T_s	2 T_s	3 T_s	4 T_s

Table 4.7: Filtered Channel Power Delay Profile of a Multipath Channel Not Exhibiting Destructive ISI

Figure 4.5 below shows the Channel Frequency Response of the multipath channel described by Table 4.6 and 4.7.

Figure 4.5: Channel Frequency Response of a Multipath Channel Not Exhibiting Destructive ISI

Figure 4.6 shows the performance of the different equalizers over various E_b/N_0 values when attempting to recover a signal transmitted via a Multipath Channel Not Exhibiting Destructive ISI. From this figure, we can observe that the RTCE performs better than the ZFE, Linear Equalizer but is similar in performance to the DFE. This is due to the channel not exhibiting any destructive ISI and hence the RTCE performs similar to the DFE.

Figure 4.6: Performance Comparison of different Equalizers in a Multipath Channel Not Exhibiting Destructive ISI

5 Conclusion and Recommendations

5.1 Conclusion

The RTCE fundamentally works by transmitting a pilot signal with a bandwidth matching that of the data pulse that would be sent later. This pilot signal would then provide the CIR for the entire bandwidth of the channel as shown in (3.31) rather than just the bandwidth of the underlying data signal as shown in (3.32). With this additional information about the entire bandwidth of the channel the RTCE is able to mitigate the destructive ISI introduced by the channel. Also, the RTCE, does not require training data to be prefixed to each of the message blocks and hence does not suffer the loss in data transmission rate as a result of (2.4) unlike the DDE.

In Chapter 4, we simulated the transmission of IR-UWB signals, with a symbol rate of $1/2T_s$ and pulse period of T_s , via different a multipath channel exhibiting destructive ISI and a multipath channel not exhibiting any destructive ISI and compared the performance of the RTCE with that of the Linear Equalizer, Zero Forcing Equalizer, Decision Feedback Equalizer and Ideal MLSE.

Looking at the simulation results, when transmitting an IR-UWB signal via a Multipath Channel Exhibiting Destructive ISI, it is clear that the RTCE is able to mitigate the destructive ISI introduced by the channel and performs better than most commonly used equalizers. This is particularly useful for WPAN which need to conform to the low power emission restriction imposed by FCC.

5.2 Recommendations

In Chapter 4, we have shown that the RTCE is able to mitigate the destructive ISI introduced by the channel and performs better than most commonly used equalizers, which is particularly useful for WPAN. The results in Chapter 4 were consolidated mainly by running computer simulations though. To validate that the RTCE indeed performs better in reality, we would need to implement this logic in hardware, and verify its performance.

The RTCE would require $C \times L$ operations, to recover a message sequence of length L transmitted via a Channel of Length C . Hence, to recover a message sequence in a multipath rich channel with long delay spreads using the RTCE would result in increased computation complexity. In such scenarios, we would need to verify if our hardware implementation of the RTCE has sufficient computational power and speed to process the incoming data sufficiently fast enough.

Devices operating in WPANs are typically powered via a battery and hence have a limited power supply. We would need to compare the power consumption of the RTCE with that of the other commonly used equalizers and verify if the performance uptick offered by the RTCE outweighs any additional power utilization it may have.

The RTCE performs better than the decision feedback equalizer only if the channel exhibits destructive ISI. Hence, we would also need to run channel sounding experiments on numerous WPANs and consolidate this data to determine the extent to which destructive ISI is a problem in WPANs.

Lastly, all wireless communication systems suffer from some degree of destructive ISI. Hence, we would need to run channel sounding experiments on numerous WLANs and WWAN to determine the extent to which destructive ISI is a problem for them. If destructive ISI does pose a genuine concern in any of these cases, we would attempt to utilize the RTCE in these scenarios to compare their performance against the most commonly used equalizers.

6. Author's Publications

- [22] R. R. Kurumbailmadham, "Mitigating the Effects of Destructive Fading in IR-UWB Wireless Systems", IRC Conference on Science, Engineering and Technology (IRC-SET), 2018

7. Bibliography

- [1] M. Mirshafiei and L. A. Rusch, "Equalizer Complexity/ Performance Trade-Offs for High Data-Rate IR-UWB Linear Receivers in Multipath Channels", IEEE International Conference on Ultra-Wideband, pp 215-219, 2011.
- [2] B. Sklar, "Digital Communications Fundamentals and Applications", Prentice Hall, Second Edition, pp 408, 2017.
- [3] T. S. Rappaport, "Wireless Communications: Principles and Practice", Prentice Hall, Second Edition, pp 167, 2002.
- [4] J. Yang and P. Dong, "On the Power Spectrum Density of UWB Signals in IEEE 802.15.4a", Intelligent Information Hiding and Multimedia Signal Processing, pp 461-464, 2012.
- [5] A. F. Molisch, J. R. Foerster and M. Pendergrass, "Channel Models For UltraWideband Personal Area Networks", IEEE Wireless Communications, pp 14-21, 2003.
- [6] S.S. Virk and P.M. Nathan, "Comparative Performance Analysis of Zero Forcing, Minimum Mean Square Error and Decision Feedback Equalizer", IEEE International Conference on Green Computing, pp 136 -139, 2013.
- [7] Farebrother, R.W, "Linear Least Squares Computations (Statistics: A Series of Textbooks and Monographs)", CRC Press, First Edition, 1988.
- [8] K. Siwiak and D. McKeown, "Ultra-Wideband Radio Technology", Wiley, Second Edition, pp 59-65, 2004.
- [9] D. Kernler, "Normal Distribution plot", Retrieved from:
<https://commons.wikimedia.org/w/index.php?curid=36506025>

-
- [10] Z. Chen, Y. P. Zhang, A. Q. Hu, and T.-S. Ng, "Bit-error-rate analysis of UWB radio using BPSK modulation over inter-chip radio channels for wireless chip area networks", *IEEE Trans. Wireless Commun.*, vol. 8, no. 5, pp. 2379–2387, 2009.
 - [11] W.C. Tjhi, C.T. Lau, and A. B. Premkumar, "A Generalization of Shift-Window Approach for UWB ISI Reduction", *IEEE Communications Letters*, vol. 14, no. 7, pp 602-604, 2010.
 - [12] L. Yang and G. Giannakis, "Ultra-wideband communications: an idea whose time has come", *IEEE Signal Process. Mag.*, vol. 21, no. 6, pp. 26–54, 2004.
 - [13] R. T. Hoctor and H. W. Tomlinson, "An overview of delay-hopped transmitted-reference RF communications", *Technique Information Series: GE Research and Development Center*, 2002.
 - [14] A. Rabbachin and I. Oppermann, "Comparison of UWB transmitted reference schemes", *IEE Proceedings - Communications*, vol. 153, no. 1, pp. 136– 42, 2006.
 - [15] Frank M. Hsu, "Data Directed Estimation Techniques for Single-Tone HF Modems", *MILCOM 1985 - IEEE Military Communications Conference*, pp 271-280, 1985.
 - [16] G. Heidari, "WiMedia UWB Technology of Choice for Wireless USB and Bluetooth", *Wiley*, 2008.
 - [17] M. Z. Win and R. A. Scholtz, "Impulse radio: How it works", *IEEE Communications Letters*, vol. 2, no. 2, pp. 36-38, 1998.
 - [18] J.F. Kurose and K.W. Ross, "Computer Networking: A Top-Down Approach", *Pearson*, Sixth Edition, pp. 526, 2012.
 - [19] "IEEE Standard for Information Technology—Telecommunications and Information Exchange Between Systems—Local and Metropolitan Area Networks—Specific Requirements—Part 11: Wireless LAN Medium Access Control (MAC) and Physical

- Layer (PHY) Specifications”, IEEE Standard 802.11-2007, Jun. 2007.
- [20] Ali Grami, "Introduction to Digital Communications", Academic Press, First Edition, pp286-292, 2015
- [21] T. S. Rappaport, “Wireless Communications: Principles and Practice”, Prentice Hall, Second Edition, pp 163, 2002