This thesis is dedicated to my parents

My father, who wished me to see at the top of everything…
It is the memories of his personality which have been a guide for me always…
It is the great love that he filled me with, that I could stand the most difficult of circumstances…

My mother, for her immeasurable sacrifices and profound love…
Her hard-work and dedication to do the best that she can for me…

For them my love and gratitude is immense

Abstract

Orthogonal Frequency Division Multiplexing (OFDM) has gained enormous popularity in the recent years due to its inherent ability to counter multipath effects and support high data rates. Its implementation has been made simpler through the efficient calculation of Discrete Fourier Transform (DFT) via the Fast Fourier Transform (FFT) algorithm. However, timing and frequency synchronization issues in OFDM affect its performance severely. Carrier Frequency Offset (CFO) in particular presents a huge problem as a minor offset disturbs the orthogonality of the carriers. In addition, it suffers from high Peak to Average Power ratio (PAPR) at the transmitter and requires the transmitter to operate with a large linear range.

To alleviate these problems, a Single Carrier with Frequency Domain Equalization (SC-FDE) is considered which gives performance comparable to OFDM. CFO does not pose that big a problem in SC-FDE. Various frequency domain equalizers have been implemented which include Linear Equalizer, Decision Feedback Equalizer (DFE) and Iterative Block DFE. Inorder to avoid the error propagation phenomenon of the DFEs and to increase the robustness of the Equalizer, we propose the use of a training sequence to optimize the feedforward and feedback filter coefficients. The new technique Data Aided Decision Feedback Equalizer (DAB-DFE), assumes that the transmission takes place within the duration of coherence time and hence, the channel remains constant for the subsequent blocks in the frame. A number of different channels are considered for performance evaluation. The results confirm that DAB-DFE outperforms the contemporary equalizers.

Acknowledgements

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Chapter 1

Introduction

This thesis analyses various frequency domain equalization techniques for a single carrier communication system which has been proposed as an alternative to OFDM to combat multipath fading. As the need for higher data rates is growing, there is an increase in the requirement of greater signal bandwidth. If the signal bandwidth becomes greater than the coherence bandwidth of the channel, the signal will experience frequency selective fading. In order to mitigate the effects of multipath, various techniques are in use. These include Single Carrier with Time Domain Equalization and OFDM.

Single Carrier modulation with time domain equalization has been in use in voice band telephone modems where the channel dispersion is not a dominant factor. But as the channel dispersion and hence the multipath delay spread increases, the time domain equalizer grows more and more complex rendering it unfavorable to be used.

Orthogonal Frequency Division Multiplexing (OFDM) is another technique which provides us with low complexity and offers a good solution for multipath fading. The reason being, that OFDM transmits multiple slowly modulated carriers in parallel. This is done by taking IFFT of the signal at the transmitter side which modulates the data symbols on orthogonal sub-carriers. For each sub-carrier, the channel appears to be frequency flat causing only one attenuation and phase rotation. At the receiver side the data can be recovered by doing a simple FFT operation followed by a single-tap equalizer. OFDM with its simplicity is a very efficient anti-multipath approach. However, due to the orthogonal sub-carriers, its Peak to Average Power ratio (PAPR) is very high, requiring predistortion techniques or large back off for the power amplifier.

Lately, Frequency Domain Equalization has been suggested to be used with Single Carrier Modulation which gives the performance comparable to OFDM with almost the same overall complexity with no PAPR problem [1]. This comparison involves the Frequency Domain Linear Equalizer. Further studies involve Frequency Domain Decision Feedback Equalizer (FD-DFE) [2], [3] which in fact is a Hybrid DFE as the Forward Filter is in the Frequency Domain whereas the Feedback Filter is in the Time Domain. This Equalizer shows a little improvement in the performance at high SNR values at the expense of increase in complexity. Another FDE technique has been presented which has both the filters in the frequency domain. This is the Iterative Block DFE (IBDFE) [4] which gives us a filter with lower complexity than the FD-DFE.

This thesis explores some of these equalization techniques. A complete Single Carrier Communication system has been implemented which includes the blocks of Timing and Frequency Synchronization, hence, they are not assumed perfect. Channel Estimation algorithm [5] has also been added to the algorithm. Algorithms [1], [2], [3] and [4] have been simulated and their performances are compared in various scenarios i.e. AWGN, Multipath Channel and Rayleigh Faded Channel. In the end we propose an improvement in the IBDFE algorithm by using training sequence for the estimation of the filter coefficients which makes the algorithm robust.

1.1 Organization

The outline of the thesis is as follows:

- **Chapter 2** explains the existing anti-multipath techniques i.e. Single Carrier with Time Domain Equalization and OFDM. It gives a slight comparison between the existing techniques and elaborates the motivation behind preferring Single Carrier system with Frequency Domain Equalization and its basic principles.
- **Chapter 3** introduces the SC-FDE System Model that we have simulated. It explains each block of the model thoroughly.

- **Chapter 4** elaborates the characteristics of a mobile channel including different types of fading. It further explains an algorithm which is used for the estimation of slow fading channels.
- **Chapter 5** investigates various Frequency Domain Equalization Techniques. It includes Frequency Domain Linear Equalizer (FD-LE), Frequency Domain Decision Feedback Equalizer (FD-DFE), Iterative Block DFE (IBDFE). We will give details of these all these algorithms as well as their comparisons.
- **Chapter 6** proposes improvement in the IBDFE by making use of a training sequence. The detail of this algorithm; Data-Aided Block DFE (DAB-DFE) is given.
- **Chapter 7** gives the simulation results for all these Equalization Techniques. These algorithms are tested in AWGN, Multipath Channel and Rayleigh Faded Channel with different Doppler Frequencies.
- **Chapter 8** gives a summary and suggests the future work which can be carried out in this field.

Chapter 2

Anti-Multipath Approaches

2.1 Single Carrier with Time Domain Equalization

A conventional anti-multipath approach which was pioneered in voiceband telephone modems and has been applied in many other digital communications systems is to transmit a single carrier, modulated by data using Quadrature Phase Shift Keying (QPSK) or Quadrature Amplitude Modulation (QAM) and to use an adaptive equalizer at the receiver to compensate for the Inter-Symbol Interference (ISI) [2]. Its main components are one or more transversal filters for which the number of adaptive tap coefficients is on the order of the number of data symbols spanned by the multipath. For example, for a 20μs delay spread we would require a transversal filter with at least 100 taps and at least several hundred multiplication operations per data symbol. For tens of megasymbols per second and more than about 30-50 symbol ISI, the complexity and required digital processing speed become exorbitant, and this time-domain equalization approach becomes unattractive.

2.2 Orthogonal Frequency Division Multiplexing (OFDM)

One way to mitigate the frequency-selective fading seen in a wide band channel is to use a multicarrier technique which subdivides the entire channel into smaller subbands, or subcarriers. Orthogonal frequency division multiplexing (OFDM) is a multicarrier modulation technique which uses orthogonal subcarriers to convey information. It transmits multiple modulated sub-carriers in parallel [2]; each occupies only a very narrow bandwidth. Since the channel affects only the amplitude and phase of each sub-carrier, equalization of each sub-carrier's gain and phase does compensation for frequency selective fading. In the frequency domain, since the bandwidth of a subcarrier is designed to be smaller than the coherence bandwidth, each subchannel is seen as a flat fading channel which simplifies the channel equalization process. In the time domain, by splitting a high-rate data stream into a number of lower-rate data stream that are transmitted in parallel, OFDM resolves the problem of ISI in wide band communications. More technical details on OFDM are at [8], [13], [14], [15], [16], and [17].

In summary, OFDM has the following advantages:

- For a given channel delay spread, the implementation complexity is much lower than that of a conventional single carrier system with time domain equalizer.
- Spectral efficiency is high since it uses overlapping orthogonal subcarriers in the frequency domain.
- Modulation and demodulation are implemented using inverse discrete Fourier transform (IDFT) and discrete Fourier transform (DFT), respectively, and fast Fourier transform (FFT) algorithms can be applied to make the overall system efficient.
- Capacity can be significantly increased by adapting the data rate per subcarrier according to the signal-to-noise ratio (SNR) of the individual subcarrier.

Because of these advantages, OFDM has been adopted as a modulation of choice by many wireless communication systems such as wireless LAN (IEEE 802.11a and 11g) and DVB-T (Digital Video Broadcasting-Terrestrial).

However, it suffers from the following drawbacks [18], [19]:

- High peak-to-average power ratio (PAPR): The transmitted signal is a superposition of all the subcarriers with different carrier frequencies and high amplitude peaks occur because of the superposition.
- High sensitivity to frequency offset: When there are frequency offsets in the subcarriers, the orthogonality among the subcarriers breaks and it causes intercarrier interference (ICI).
- A need for an adaptive or coded scheme to overcome spectral nulls in the channel: In the presence of a null in the channel, there is no way to recover the data of the subcarriers that are affected by the null unless we use rate adaptation or a coding scheme.

Typically, the FFT block length M is at least 4-10 times longer than the maximum impulse response span. One reason for this is to minimize the fraction of overhead due to the insertion of a cyclic prefix at the beginning of each block. The cyclic prefix is a repetition of the last data symbols in a block. Its length in data symbols exceeds the maximum expected delay spread. The cyclic prefix is discarded at the receiver. Its purpose is to:

- Prevent contamination of a block by ISI from the previous block
- Make the received block appear to be periodic with period M

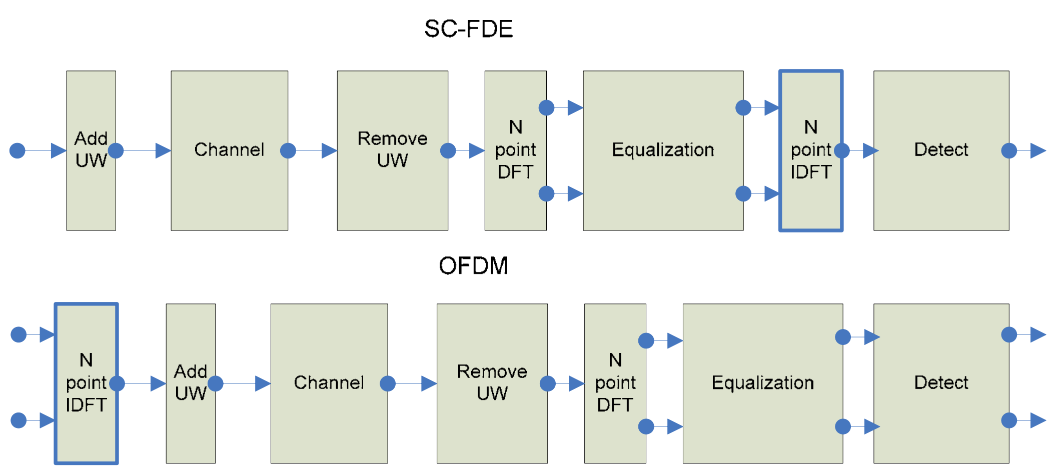

Figure 2-1 Block Diagram of SC-FDE and OFDM

2.3 Single Carrier Modulation with Frequency Domain Equalizer Processing (SC-FDE)

An SC system transmits a single carrier, modulated, for example, with QAM at a high symbol rate [2]. For broadband multipath channels, conventional time domain equalizers are impractical because of the complexity (very long channel impulse response in the time domain). Frequency domain equalization (FDE) is more practical for such channels. Single carrier with frequency domain equalization (SC-FDE) technique is another way to fight the frequency-selective fading channel. Frequency domain linear equalization in an SC system is simply the frequency domain analog of what is done by a conventional linear time domain equalizer. It delivers performance similar to OFDM with essentially the same overall complexity, even for long channel delay [18], [19]. For channels with severe delay spread, frequency domain equalization is computationally simpler than corresponding time domain equalization for the same reason OFDM is simpler: because equalization is performed on a block of data at a time, and the operations on this block involve an efficient FFT operation and a simple channel inversion operation. Figure 2-1 shows the block diagram of SC-FDE and compares it with that of OFDM.

In the transmitter of SC-FDE, we add a cyclic prefix (CP), which is a copy of the last part of the block, to the input data at the beginning of each block in order to prevent inter-block interference (IBI) and also to make linear convolution of the channel impulse response look like a circular convolution. Apart from the CP, a Unique Word (UW) can also be added. It can be a PN sequence and since it is repeated for every block of data, it performs the same work as that of the CP. It should be noted that circular convolution problem exists for any FDE since multiplication in the DFT-domain is equivalent to circular convolution in the time domain [20]. When the data signal propagates through the channel, it linearly convolves with the channel impulse response. An equalizer basically attempts to invert the channel impulse response and thus channel filtering and equalization should have the same type of convolution either linear or circular convolution. One way to resolve this problem is to add a CP or UW in the transmitter that will make the channel filtering look like a circular convolution and match the DFT-based FDE.

Another way is not to use CP or UW but perform an “overlap and save” method in the frequency domain equalizer to emulate the linear convolution [20]. However, the use of cyclic prefix gives protection form inter-block interference which should not be neglected. For either OFDM or SC-FDE broadband wireless systems operating in severe outdoor multipath environments, typical values of FFT size could be 256 - 1024. SC-FDE receiver transforms the received signal to the frequency domain by applying DFT and does the equalization process in the frequency domain. Most of the well-known time domain equalization techniques, such as minimum mean-square error (MMSE) equalization, decision feedback equalization, and turbo equalization, can be applied to the FDE After the equalization, the signal is brought back to the time domain via IDFT and detection is performed.

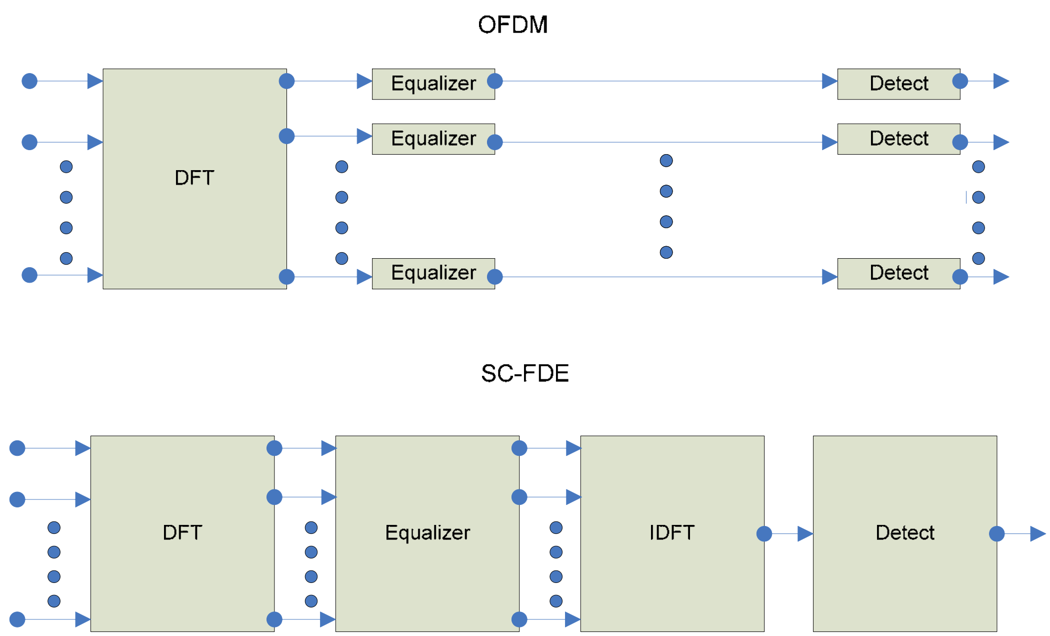

Figure 2-2 Differences between SC-FDE and OFDM

Comparing the two systems in Figure 2-1, it is interesting to find the similarity between the two. Overall, they both use the same communication component blocks and the only difference between the two diagrams is the location of the IDFT block.

Thus, one can expect the two systems to have similar link level performance and spectral efficiency.

However, there are distinct differences that make the two systems perform differently as illustrated in Figure 2-2. In the receiver, OFDM performs data detection on a per-subcarrier basis in the frequency domain whereas SC-FDE does it in the time domain after the additional IDFT operation. Because of this difference, OFDM is more sensitive to a null in the channel spectrum and it requires channel coding or power/rate control to overcome this deficiency. Also, the duration of the modulated time symbols are expanded in the case of OFDM with parallel transmission of the data block during the elongated time period.

Also notable is that a frequency domain receiver processing SC modulated data shares a number of common signal processing functions with an OFDM receiver as can be seen in Figure 2.1.

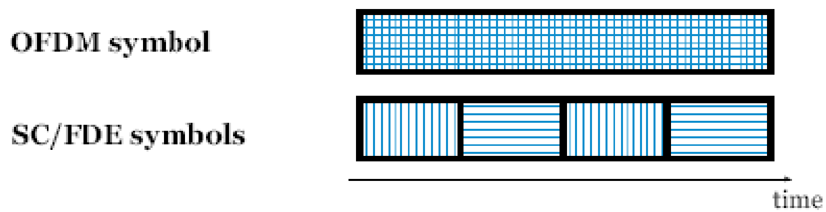

Figure 2-3 Differences between SC-FDE and OFDM symbols

Overlap-add or Overlap-save techniques could also be used to avoid the extra overhead of the cyclic prefix.

2.3.1 Why Use SC-FDE?

The use of SC modulation and FDE by processing the FFT of the received signal has several attractive features:

- SC modulation has reduced peak-to-average ratio requirements from OFDM, thereby allowing the use of less costly power amplifiers.

- Its performance with FDE is similar to that of OFDM, even for very long channel delay spread.
- Frequency domain receiver processing has a similar complexity reduction advantage to that of OFDM: complexity is proportional to log of multipath spread.
- Coding while desirable, is not necessary for combating frequency selectivity as it is in adaptive OFDM
- A further sensititivity of OFDM, not shared to the same degree by single carrier, is phase noise and frequency offsets, due to the close spacing in frequency of its sub-carriers. This sensitivity leads to tighter local oscillator requirements for OFDM systems.
- SC modulation is a well-proven technology in many existing wireless and wireline applications, and its RF system linearity requirements are well known.

A comparison of the various anti-multipath schemes is given in Table 2.1

Table 2-1 A comparison of Anti-multipath schemes

	OFDM	SC-FDE	SC-TDE
Signal PAPR	High	Low	Low
Computational Complexity	Low	Low	High
Coding Requirement	Strict	Flexible	Flexible

Chapter 3

System Model

Figure 3.1 gives the block diagram of the Single Carrier Communication system that we have simulated in MATLAB. This chapter will explain the whole model in detail.

3.1 Transmitter

3.1.1 Constellation Mapping

Random binary data is generated. This data is then modulated using BPSK, QPSK, 8PSK, and 16 QAM.

3.1.2 Addition of Unique Word

Frequency Domain Multiplication is equivalent to time domain circular convolution. Hence we require that the convolution of (1) is forced to be circular i.e the transmitted data must satisfy the following requirement

$$s(kP-n) = s[(k+1)P-n] \tag{3.1}$$

We can arrange the data sequence in various types in order to make the above condition hold for the transmitted signal. These include pseudo-noise (PN) extended or zero-padded (ZP) transmission [3], [13] and the cyclic prefix (CP) extended transmission [11].

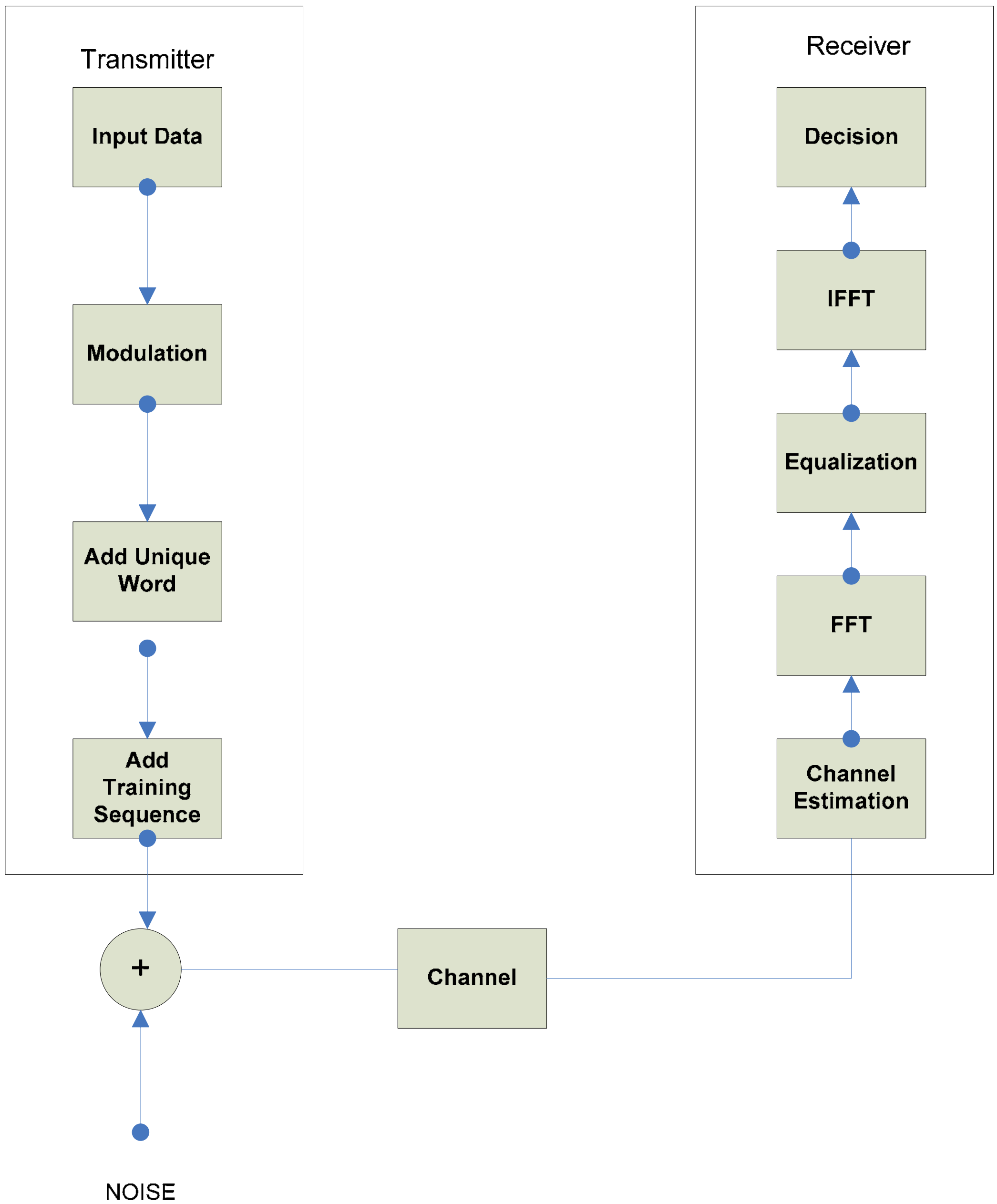

Figure 3-1 The system block diagram

In the PN-extended transmission, each information data block of length M is extended with a fixed sequence of L symbols, for example, a PN sequence $\{p(n), n = 0,1,\ldots,L-1\}$ with $L \geq N_h - 1$. The transmitted data block of $P = M + L$ symbols is

$$s(k) = [s(k(M+L)), s(k(M+L)+1), \dots, s(k(M+L)+(M+L)-1)] \tag{3.2}$$

$$s(k) = [d(kM), d(kM+1), \dots, d(kM+M-1), p(0), p(1), \dots, p(L-1)] \tag{3.3}$$

where the last L symbols are the PN sequence. An additional PN extension is required before the first data block. This format can reap a few benefits. For example, the channel estimation technique of [14] can be implemented. ZP is a particular case of this format.

In the CP extension, the information data signal at the end of a block is appended in the beginning to make it cyclic. The transmitted block as a result becomes

$$s(k) = [d(kM+M-L), d(kM+M-L+1, \dots, d(kM+M-1), d(kM), d(kM+1), \dots, d(kM+M-1)] \tag{3.4}$$

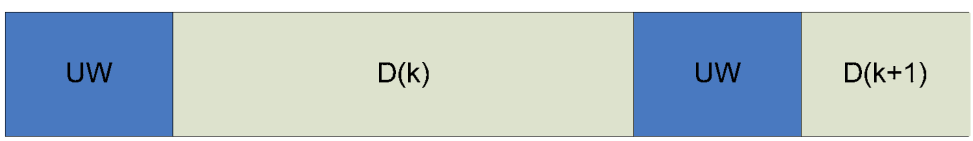

Figure 3-2 Frame format for PN-extension

where the first L symbols coincide with the last symbols of the block. The convolution thus gets circular on blocks of size $P=M$. The CP-extension format is also used in multi-carrier communications [13] and has been proposed also for SC transmission with linear FD equalization [11]. We work with PN-extension format. It is found that a reduced BER is obtained for PN-extension format.

3.1.3 Appending the Training Sequence

We add the training sequence prior to the transmission of data. It serves the following purposes

- Timing Synchronization
- Frequency Synchronization
- Channel Estimation

Various frequency domain equalizers are tested and compared two cases:

1. Assuming ideal timing and frequency synchronization
2. Using the timing and frequency synchronization algorithms [6], in order get the performance evaluation for realistic scenarios.

One training symbol is used which serves all the three above mentioned purposes, the format of which is given in the following chapter. Training Sequence is also used for training of the equalizer coefficients. However, this training sequence should be different from the one used for synchronization and channel estimation.

3.2 Channel Model

We have simulated the frequency domain equalizers in the following channel conditions:

- AWGN
- Flat Fading Channel
- Multipath Fading
- Multipath Rayleigh Faded Channel

We created our own channel model with differing path gains to make a multipath channel. For the Rayleigh Fading effect we used the MATLAB command ‘rayleighchan’. We tested the simulations on carrier frequency of 426 MHz. For this carrier frequency, the coherence time for the channel was calculated. We assume that the transmission time of the signal is less than or equal to the coherence time of the channel. This assumption is very important for synchronization algorithms and the channel estimation algorithm as well.

3.3 Receiver

3.3.1 Channel Estimation

Channel estimation is done on the receiver side. It makes use of the training symbol sent by the transmitter. Details of the algorithm are in the preceding chapter. This is a very important block of the system as all the equalizers are heavily dependent on the performance of the channel estimation.

3.3.2 Frequency Domain Equalization

Equalizers have the information of the channel impulse response. For frequency domain equalization, FFT of the received signal as well as that of the channel impulse response is taken. We must know the exact starting point of our data blocks as we have to take the FFT of the blocks of data in the same format as transmitted. The importance of the synchronization algorithms becomes evident here. If the transmitter and the receiver are not synchronized, not only the performance of the channel estimator will be highly affected but also that of the equalizer.

After the equalization, IFFT of the block of data is taken and then sent to the slicer. It takes the decision and recovers the transmitted data. This completes our system.

Chapter 4

Channel Characteristics and Channel Estimation

The mobile channels in use today frequently experience multipath and other Doppler spread degradations. These degradations can be characterized in certain categories [23].

4.1 Mobile Channel Characteristics

Wireless mobile communications are characterized by two types of fading effects: Large scale fading and small scale fading. The average signal power attenuation due to motion over large areas is called large scale fading. Its statistics express the path loss as a function of distance. Small scale fading is the phenomenon of changes in the signal amplitude and phase due to small changes in the spatial separation between the transmitter and the receiver. This type of fading is constituted of two mechanisms: time spreading of the channel and time variant nature of the channel. In this context, few terms are worth defining; Delay spread, Coherence Bandwidth, Doppler spread, and Coherence time.

Delay Spread is a type of distortion that is caused when an identical signal arrives at different times at its destination. The signal usually arrives in a number of different paths and with different angles of arrival. The time difference between the arrival moment of the first multipath component, usually the line of sight component, and the last one, is called delay spread.

Coherence bandwidth is a statistical measure of the range of frequencies over which the channel can be considered "flat" i.e. a channel which passes all spectral components with approximately equal gain and linear phase. In other words, coherence bandwidth is the range of frequencies over which two frequency components have a strong potential for amplitude correlation.

Doppler spread B_D is a measure of the spectral broadening caused by the time rate of change of the mobile radio channel and is defined as the range of frequencies over which the received Doppler spectrum is essentially non-zero. When a pure sinusoidal tone of frequency f_C is transmitted, the received signal spectrum, called the Doppler spectrum, will have components in the range $f_c - f_d$ to $f_c + f_d$, where f_d is the Doppler shift. The amount of spectral broadening depends on f_d which is a function of the relative velocity of the mobile, and the angle θ between the direction of motion of the mobile and direction of arrival of the scattered waves.

Coherence Time is the time interval during which two data points spaced by a certain distance in time experience sufficient correlation after passing through the channel. In other words, it is the measure of the staticness of the channel. Typically for mobile communications, it is defined as

$$T_C = \frac{0.423}{f_d} \tag{4.1}$$

4.2 Fading due to Multipath Effects

4.2.1 Flat Fading

In this type of fading, the multipath structure of the channel is such that the spectral characteristics of the transmitted signal are preserved at the receiver. However, the strength of the received signal changes with time, due to fluctuations in the gain of the channel caused by multipath.

If the channel gain changes over time, a change of amplitude occurs in the received signal. Over time, the received signal $r(t)$ varies in gain, but the spectrum of the transmission is preserved. In a flat fading channel, the reciprocal bandwidth of the transmitted signal is much larger than the multipath time delay spread of the channel, and $h_b(t,\tau)$ can be approximated as having no excess delay (i.e. a single delta function with $\tau = 0$). Flat fading channels are also known as *amplitude varying channels* and are sometimes referred to as *narrow band channels*, since the bandwidth of the applied signal is narrow as compared to the channel flat fading bandwidth. Hence, a signal undergoes flat fading if

$$B_S \ll B_C \tag{4.2}$$

$$\text{and } T_S \gg \sigma_\tau$$

Where T_S is the reciprocal bandwidth (e.g. symbol period) and B_S is the bandwidth, respectively, of the transmitted modulation, and σ_τ and B_C are the rms delay spread and coherence bandwidth, respectively, of the channel.

4.2.2 Frequency Selective Fading

Frequency selective fading is due to the time dispersion of the transmitted symbols within the channel. Thus the channel includes *intersymbol interference* (ISI). Viewed in the frequency domain, certain frequency components in the received signal spectrum have greater gains than others.

For frequency-selective fading, the spectrum $S(f)$ of the transmitted signal has a bandwidth which is greater than the coherence bandwidth B_C of the channel. Viewed in the frequency domain, the channel becomes frequency selective, where the gain is different for different frequency components. Frequency selective fading is caused by multipath delays which approach or exceed the symbol period of the transmitted symbol. Frequency selective fading channels are also called wideband channels because the bandwidth of the signal $s(t)$ is wider than the bandwidth of the channel impulse response. As time varies, the channel varies in gain and phase across the spectrum of $s(t)$, resulting in time varying distortion in the received signal $r(t)$. Hence, a signal undergoes frequency selective fading if

$$B_S > B_C \tag{4.3}$$

and $T_S < \sigma_\tau$

4.3 Fading effects due to Doppler Spread

4.3.1 Fast Fading

In a fast fading channel, the channel impulse response changes rapidly within the symbol duration. That is the coherence time of the channel is smaller than the symbol period of the transmitted signal. This causes frequency dispersion (also called time-selective fading) due to Doppler spreading, which leads to signal distortion. Viewed in the frequency domain, signal distortion due to fast fading increases with increasing Doppler spread relative to the bandwidth of the transmitted signal. Therefore, a signal undergoes fast fading if

$$T_S > T_C \tag{4.4}$$

and $B_S < B_D$

A flat fading, fast fading channel is a channel in which the amplitude of the delta function varies faster than the rate of change of the transmitted baseband signal. In the case of a frequency selective fast fading channel, the amplitude, phase, and time delays of any one of the multipath components vary faster than the rate of change of the transmitted signal. In practice, fast fading only occurs for very low data rates.

4.3.2 Slow Fading

In a slow fading channel, the channel impulse response changes at a rate much slower than the transmitted baseband signal s(t). In this case, the channel maybe assumed to be static over one or several reciprocal bandwidth intervals. In the frequency domain this implies that the Doppler spread of the channel is much less than the bandwidth of the baseband signal. Therefore, a signal undergoes slow fading if

$$T_S \ll T_C \tag{4.5}$$

$$\text{and } B_S \gg B_D$$

The velocity of the mobile (or the velocity of the objects in the channel) and the baseband signaling determines whether a signal undergoes fast fading or slow fading.

Channel Estimation is required for all the equalizers. Various techniques are used to estimate the channel response, some of which include [5] and [6]. We found the technique of [6] to be very efficient and we will describe it in detail.

4.4 Design of the Training Symbol

The training symbol has been designed both for an OFDM-type Frequency Domain training or single carrier-type Time Domain training. This training symbol is composed of L identical parts to handle the frequency offsets up to $\pm L/2$ carrier spacing. The L identical parts are supplied with specific sign patterns to give a steep roll off timing metric trajectory. The training symbol can be designed with the help of Golay complementary sequences which are known to result in low PAPR values. The basic structure of the training symbol is

$$s = \left[\pm A, \pm A, \pm A ...., \pm A\right] \tag{4.6}$$

where A is the repeated part which is given a predetermined sign pattern whose knowledge is common to both transmitter and receiver. The sign patterns for different values of L were determined by a computer search and are given in Table 4.1.

Table 4.1 Training Symbol Sign Pattern

L	Sign Pattern
4	$(-+--)$
	$(+++-)$
8	$(++--+---)$
	$(-++---+-)$
16	$(+--+++--+-++-+--)$
	$(--++-++----++++-)$

4.5 Channel Impulse Response Estimation

This algorithm assumes channel to be quasi-static which is perfectly valid in our case as the frequency domain equalization techniques are tested on the assumption that the signal duration is less than or equal to the coherence time of the channel. Hence, we are dealing with the slow fading case.

Let us define the following:

$$\mathrm{r}(0) \triangleq [\, r(0) \quad r(1) \quad \cdots \quad r(N-1)\,]^{\mathrm{T}} \tag{4.7}$$

$$\mathrm{h} \triangleq [\, h_0 \quad h_1 \quad \cdots \quad h_{K-1}]^{\mathrm{T}} \tag{4.8}$$

$$\mathrm{W}(v) \triangleq diag\{1, e^{j2\pi v/N}, e^{j2\pi 2v/N}, ..., e^{j2\pi(N-1)v/N}\} \tag{4.9}$$

$$\mathrm{n} \triangleq [\, n(0) \quad n(1) \quad \cdots \quad n(N-1)\,]^{\mathrm{T}} \tag{4.10}$$

$$S \triangleq \begin{pmatrix} s(0) & s(-1) & \cdots & s(-K+1) \\ s(1) & s(0) & \cdots & s(-K+2) \\ \ddots & & & \\ s(N-1) & s(N-2) & \cdots & s(N-K) \end{pmatrix} \tag{4.11}$$

where $h_0, h_1, ..., h_{K-1}$ give the instantaneous path gains, $\{s(k): k = -N_g, -N_g + 1, ..., N-1\}$ are the samples of the transmitted training symbol(including the cyclic prefix part), $\{r(k): k = 0,1,..., N-1\}$ the corresponding received samples (excluding the cyclic prefix part), $\{n(k): k = 0,1,..., N-1\}$ the noise samples, and v the normalized frequency offset. The received sample vector can then be expressed as

$$r(0) = e^{j\phi} W(v) \cdot S \cdot h + n \tag{4.12}$$

If we use the frequency offset estimate $\hat{v}$, the ML channel impulse response estimate can be realized by

$$\hat{h} = [S^H \cdot S]^{-1} \, S^H \cdot W^H(\hat{v}) \cdot r(0) \tag{4.13}$$

In this channel estimation, the knowledge of the maximum delay spread of the channel is required.

Chapter 5

Frequency Domain Equalizers

This chapter describes various Frequency Domain Equalizers which include the Frequency Domain Linear Equalizer [1], Frequency Domain Decision Feedback Equalizer [3], and the Iterative Block Decision Feedback Equalizer [4]. These equalizers have been simulated in MATLAB.

5.1 Frequency Domain Linear Equalizer (FD-LE)

his equalizer has been in use for OFDM systems, however, its use for Single Carrier modulation was proposed by Hikmet Sari in [1].

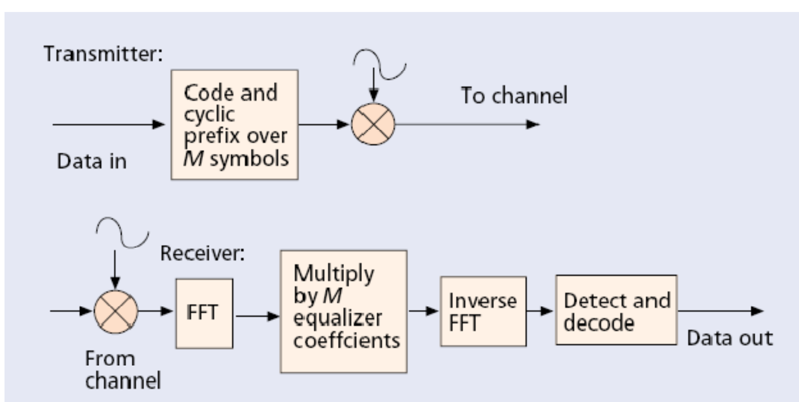

Figure 4.1 SC-FDE with Linear Equalization.

Let the estimate of the channel impulse response be given by $h(t)$, then the FD-LE requires that the N point FFT of the received signal block be multiplied by N equalizer coefficients which are defined as

$$C_k = \frac{1}{H_k} \tag{5.1}$$

The FFT of the channel impulse response is given by H_k where $k = 0,1,...,N$. This is the optimization based on Zero-Forcing Criterion canceling all the ISI irrespective of the presence of the noise. The equalizer coefficients can also be optimized on the basis of finding a compromise between canceling the noise and ISI by minimizing the mean square error and are given by

$$C_k = \frac{H_k^*}{|H_k|^2 + \frac{\sigma_n^2}{\sigma_a^2}} \tag{5.2}$$

Where σ_n^2 and σ_a^2 denote the noise and signal variance respectively.

Note that the ZF criterion does not have a solution if there are nulls in the channel transfer function. Moreover, it leads to infinite noise enhancement at the spectral nulls. Hence, the MMSE optimization yields better and more efficient filter coefficients and is particularly suitable in the scenarios where there are deep spectral nulls in the channel bandwidth.

5.2 Frequency Domain Decision Feedback Equalizer (FD-DFE)

The Decision Feedback Equalizer in the frequency domain is has been proposed by [2], [3] and [5]. These algorithms suggest the forward filter in the frequency domain and the feedback filter in the time domain.

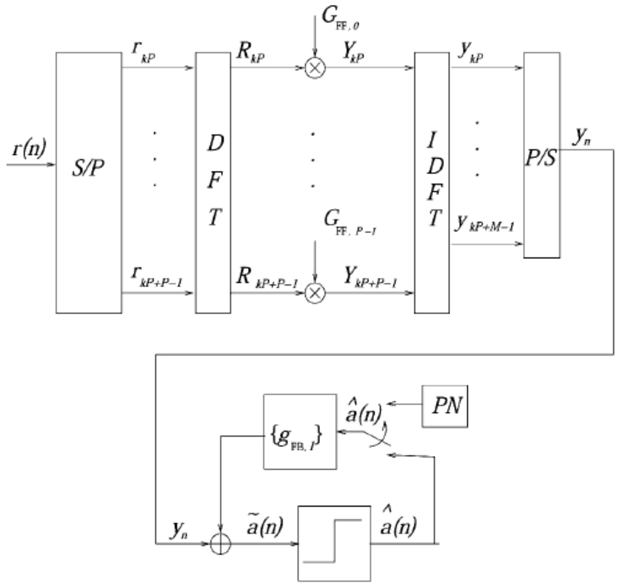

Figure 4.2 FD-DFE Equalizer for single carrier systems

We will describe the DFE in [3] in detail. Let the received signal be given by

$$r_n = \sum_{l=0}^{L} h_l s_{n-l} + w_n \qquad (5.3)$$

where $\{s_n\}$ is generated as in (3.1), and w_n is a zero mean complex Gaussian white noise with variance σ_w^2. The FD-DFE structure is shown in Fig. 4.2. The frequency-domain filtering must be performed on per-block basis, while the feedback section must be fed with the previous detected symbols whose decision is performed in the time domain.

As shown in Figure 4.2, after the DFT of the received samples, the FD feedforward filter, with coefficients $\{G_{FF,n}\}$ $n = 0,1,...,P-1$ is applied to yield the block signal $Y(k)$ with elements

$$Y_{kP+n} = R_{kP+n} G_{FF,n}, \qquad n = 0,1,...,P-1$$

Through the inverse DFT, block $Y(k)$ is then transformed in time domain to give

$$y(k) = [y_{kP}, y_{kP+1,...,} y_{kP+P-1}]$$

Let us define the impulse response of the cascade of the equivalent discrete-time channel impulse response and the feedforward equalizer as the IDFT of the product of $\{H_n\}$ and $\{G_{FF,n}\}$, i.e.

$$u_l = \sum_{n=0}^{P-1} G_{FF,n} H_n e^{j2\pi(nl/P)}, \qquad l = 0,1,...,L_u$$

assuming it has a support of $L_u + 1$ coefficients. Then, each block $y(k)$ can be written as the circular convolution of $s(k)$ and $\{u_l\}$ plus a noise term, namely

$$y_{kP+n} = (s \otimes u)_{kP+n} + \tilde{w}_{kP+n,} \\ n = 0,1,..., P-1$$

where from (7) and (8)

$$\tilde{w}_{kP+n} = \sum_{l=0}^{P-1} G_{FF,l} \left(\sum_{m=0}^{P-1} w_{kP+m} e^{-j2\pi(ml/P)} \right) e^{j2\pi(ml/P)} \\ n = 0,1,..., P-1$$

In (11) we note that if $L_u \leq L$ then the condition (2) holds for the useful part of $\{y_n\}$, which can be written as a linear convolution, and

$$y_{kP+n} = \sum_{l=0}^{L_u} s_{kP+n-l} u_l + \tilde{w}_{kP+n,} \qquad n = 0,1,..., M-1$$

Note that in (13) the last L samples of each block $y(k)$ have not been considered, since they are a noisy and interfered version of the PN sequence, which is already known. On the other hand, if $L_u > L$ then (13) does not hold and $\{y_{kP+n}\}$ depends also on $\{s_{kP+(n-l) \bmod P}\}, l = L+1, L+2,..., L_u$. In this case the feedback filter will not be able to cancel all the interference.

As mentioned previously, the feedforward filter operates in the frequency domain, while the feedback section operates in the time domain. Let’s indicate the estimated data sequence with $\{\hat{d}_n\}$ and the extended estimated sequence with $\{\hat{s}_n\}$. From (4) it holds

$$\hat{s}_{kP+n} = \begin{cases} \hat{d}_{kM+n}, & n = 0,1,\ldots,M-1, \\ p_{n-M}, & n = M, M+1,\ldots,P-1 \end{cases}$$

Then if $\{g_{FB,l}\}, l = 1,2,\ldots,N_{FB}$, are the coefficients of the FB filter, the signal at the input of the decision element is

$$\hat{d}_{kM+n} = y_{kP+n} + \sum_{l=1}^{N_{FB}} g_{FB,l}\hat{s}_{kP+n-l}$$
$$n = 0,1,\ldots,M-1$$

As in the case of Linear Equalizer the coefficients can be optimized on the basis of ZF or MMSE criterion and are given as follows

5.2.1 Zero Forcing FD-DFE

According to ZF criterion, all interferers must be cancelled by the feedback part. If the support of $\{u_l\}$ is $L_u \le L$, (13) holds true and interference can be cancelled by the feedback filter. Hence let's set $L_U = L$. The zero forcing condition can be expressed as

$$u_0 = 1, \qquad u_l = 0, \qquad \text{for} \quad l < 0, l > L$$

and only coefficients u_l, $l = 1,2,\ldots,L$, can be chosen freely. Once the $\{u_l\}$ coefficients are known, by selecting $N_{FB} = L$ and

$$g_{FB,l} = -u_{l,} \qquad l = 1,2,\ldots,L$$

the feedback filter cancels all residual interferers. In turn, from (10) coefficients of the feedforward filter $\{G_{FF,n}\}$ can be computed as

$$
\begin{aligned}
G_{FF,n} &= \frac{1}{H_n}\sum_{l=0}^{L} u_l e^{-j2\pi(nl/P)} \\
&= \frac{1}{H_n}\left(1-\sum_{l=1}^{L} g_{FB,l} e^{-j2\pi(nl/P)}\right)
\end{aligned}
\tag{5.4}
$$

where $n = 0,1,...,P-1$ and assuming $H_n \neq 0$ and using (17)

Here, the L coefficients $g_{FB,l}, l = 1,2,...,L$ are chosen to minimize the power of the filtered noise, which from (10) and under the condition (16) can be written as

$$
\begin{aligned}
J_{ZF} &= \frac{\sigma_w^2}{P}\sum_{n=0}^{P-1} | G_{FF,n} |^2 \\
&= \frac{\sigma_w^2}{P}\sum_{n=0}^{P-1}\frac{1}{| H_n |^2}\left|1-\sum_{l=1}^{L} g_{FB,l} e^{-j2\pi(\ln/P)}\right|^2
\end{aligned}
\tag{5.5}
$$

Minimizing the gradient of the above equation with respect to the feedback filter coefficients, we obtain a linear system of L equations

$$
A_{ZF}\, g_{FB} = b_{ZF} \tag{5.6}
$$

where

$$
[A_{ZF}] = \sum_{n=0}^{P-1}\frac{e^{-j2\pi((n(l-m))/P)}}{| H_n |^2} \qquad 1 \le m,l \le L \tag{5.7}
$$

$$
[b_{ZF}] = \sum_{n=0}^{P-1}\frac{e^{j2\pi(nm/P)}}{| H_n |^2} \qquad 1 \le m \le L \tag{5.8}
$$

5.2.2 Minimum Mean Square Error FD-DFE

According to the MMSE criterion, the coefficients of the FF and FB filters are chosen to minimize the sum of the power of the filtered noise, and the power of the residual interference. In particular, the mean square error at the detector is given by

$$J_{MMSE} = E[|\tilde{d}_n - d_n|^2] \quad (5.9)$$

By assuming that the past decisions are correct and that $N_{FB} \le L$ and $L_u \le L$, from (13) and (15) we obtain

$$J_{MMSE} = E\left[\left|\sum_{l=0}^{L_u} s_{kP+n-l} + \sum_{l=1}^{N_{FB}} s_{kP+n-l}\, g_{FB,l} + \tilde{w}_{kP+n} - d_{kM+n}\right|^2\right] \quad (5.10)$$

Where $n = 0,1,\ldots,M-1$. Now, we rewrite (24) in the frequency domain. Firstly, we introduce the P-size DFT of the FB filter

$$G_{FB} = \sum_{l=1}^{N_{FB}} g_{FB,l}\, e^{-j2\pi(lp/P)}, \qquad p = 0,1,\ldots,P-1 \quad (5.11)$$

Moreover, from (10) the gain of the useful data at the decision point can be written as

$$u_0 = \sum_{p=0}^{P-1} H_p G_{FF,p} \quad (5.12)$$

Hence, from (24), (25) and (26), according to the minimum mean square error criterion, the functional to be minimized is

$$J_{MMSE} = \frac{1}{P}\sum_{p=0}^{P-1}\left[\sigma_w^2 |G_{FF,p}|^2 + \sigma_d^2 |1-(G_{FF,p}H_p + G_{FB,p})|^2\right] \quad (5.13)$$

where σ_d^2 is the power of the signal .

Due to the PN-extension structure, the FB filter is not able to cancel more than L interferers; hence, we must impose that $N_{FB} \le L$. Here we consider the case $N_{FB} = L$. In order to compute the design of the FF and FB filters, we write the functional J_{MMSE} only as a function of $\{G_{FB,p}\}$. In particular, we observe that, given the feedback filter, by applying the gradient method to the above equation, the feedforward filter is given by

$$G_{FF,p} = \frac{H_p^*(1-G_{FB,p})}{|H_p|^2 + \frac{\sigma_w^2}{\sigma_d^2}} \tag{5.14}$$

Now minimizing J_{MMSE} as a function of the feedback filter coefficients, we obtain the following system of linear equations.

$$A_{MMSE}\, g_{FB} = b_{MMSE} \tag{5.15}$$

where

$$\begin{aligned} [A_{MMSE}] &= \sum_{n=0}^{P-1} \frac{e^{-j2\pi((n(l-m))/P)}}{|H_n|^2 + \sigma_w^2/\sigma_d^2} \\ [b_{MMSE}] &= \sum_{n=0}^{P-1} \frac{e^{j2\pi(nm/P)}}{|H_n|^2 + \sigma_w^2/\sigma_d^2} \end{aligned} \tag{5.16}$$

5.2.3 Iterative Block DFE (IBDFE)

Similar to the previously mentioned frequency domain equalizers, DFT is applied to successive blocks of P received samples. This equalizer consists of two parts [4]:

1. FF filter with coefficients $\{C_p\}$, $p = 0,1,...,P-1$, in the FD. It partially equalizes for the interference; and
2. FB filter with coefficients $\{B_p\}$, $p = 0,1,...,P-1$, and output $\{Y_p\}$ in the FD, which removes part of the residual interference.

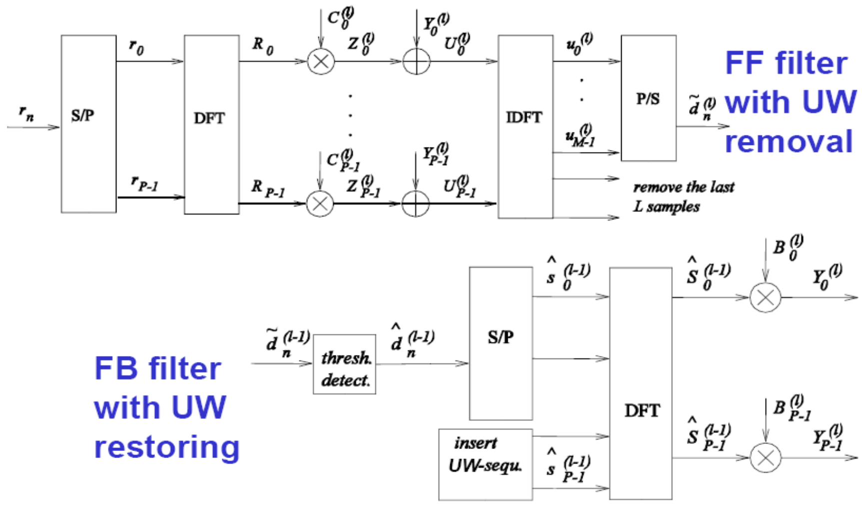

Figure 4.3 General architecture of the IBDFE

The filter coefficients and the detected data is iterated N_I times. The filter coefficients C_p obtained at the *lth* iteration are element-wise multiplied with R yielding the vector signal Z^l, with elements

$$Z_p^{(l)} = C_p^{(l)} R_p, \quad p = 0,1,\ldots,P-1 \tag{5.17}$$

At the feedback side, FFT of the detected data at iteration $(l-1)$, $\hat{s}^{(l-1)}$ and the PN sequence is taken to get $\hat{S}^{(l-1)}$ which is element-wise multiplied by the FB filter coefficients to get the FB output vector signal $Y^{(l)}$ given by

$$Y_p^{(l)} = B_p^{(l)} \hat{S}_p^{(l-1)} \tag{5.18}$$

The signal at the detection point is given by

$$U^{(l)} = Z^{(l)} + Y^{(l)} \tag{5.19}$$

where $Y^{(l)}$ removes both the precursors and the postcursors of the global impulse response. As $Y^{(l)}$ depends on the detected at iteration $(l-1)$, for $l=1$, when no detected data is available,

$$\hat{d}_p^{(0)} = 0, \quad p = 0,1,...,M-1 \tag{5.20}$$

In the end, IFFT is taken to obtain the time domain vector signal at the detection point

$$u^{(l)} = \frac{1}{P} W^H U^{(l)} \tag{5.21}$$

The first M samples of $\hat{d}$ are obtained by the first M decisions after the detector. The last points are from the PN sequence. The paper [4] proposes *Hard Detection IBDFE (HD-IBDFE)* and *Soft Detection IBDFE (SD-IBDFE).* We have, however, simulated only the HD-IBDFE and its description is given below.

5.2.4 Hard Detection IBDFE

For the design of FF and FB filters, MSE criterion is used where expectations are taken with respect to the transmitted data, noise and the detected data assuming the apriori statistic of the involved signals. In particular both the sequences are assumed independent and identically distributed (i.i.d), with zero mean and statistically independent of noise. Hence, the FB filter doesnot depend on the particular transmitted data format, but only on the correlation between the transmitted and the detected data.
The power of the involved signals in the frequency domain is given by

$$M_{S_p} = E[|S_p|^2]$$

and

$$M_{\hat{S}_P^{(l)}} = E[|\hat{S}_P^{(l)}|^2]$$

And the correlation between the transmitted and the detected data sequence:

$$\hat{r}_1^{(l)} = \frac{1}{P}\sum_{p=0}^{P-1}\frac{R_P}{H_P}\hat{S}_P^{(l-1)*} \tag{5.22}$$

The mean square error at the detection point is given by

$$J_{HD}^{(l)} = E\left[|\tilde{d}_n^{(l)} - d_n|^2\right] = \frac{1}{P}\sum_{i=0}^{P-1}E\left[|u_i^{(l)} - s_i|^2\right] \tag{5.23}$$

By applying the Parseval's theorem, and using the above two equations $J_{HD}^{(l)}$ can be written as

$$J_{HD}^{(l)} = \frac{1}{P^2}\sum_{p=0}^{P-1}E\left[|C_P^{(l)} + B_p^{(l)}\hat{S}_P^{(l-1)} - S_P|^2\right] \tag{5.24}$$

Minimizing this equation by applying the gradient method with respect to the FB filter coefficients, we get

$$B_p^{(l)} = -\frac{r_{S_P,\hat{S}_P^{(l-1)}} \times (H_P C_p^{(l)} - \gamma^{(l)})}{M_{\hat{S}_P^{(l-1)}}} \tag{5.25}$$

with $p = 0,1,\ldots,P-1$ and

$$\gamma^{(l)} = \sum_{p=0}^{P-1}H_P C_p^{(l)} \tag{5.26}$$

Now minimizing the gradient with respect to the forward filter coefficients, we obtain,

$$C_p^{(l)} = \frac{H_P^*}{M_W + M_{S_P}\left(1 - \frac{|r_{S_P,\hat{S}_P^{(l-1)}}|^2}{M_{\hat{S}_P^{(l-1)}}M_{S_P}}\right)|H_P|^2} \tag{5.27}$$

5.2.5 Parameter Estimation

For the design of both IBDFEs, the estimate fo the FD response of the channel is required. A technique for channel estimation has been presented in [5]. Additionally for the HD-IBDFE, receiver requires the estimate of powers M_{S_P} and $M_{\hat{S}_P^{(l)}}$, and the correlation(5.22). Let us give the various methods for the estimation of these parameters

5.2.5.1 Estimation of the Correlation Factor

A first estimate can be obtained by correlating the equalized received signal and the detected data signal, namely

$$\hat{r}_1^{(l)} = \frac{1}{P}\sum_{p=0}^{P-1}\frac{R_P}{H_P}\hat{S}_P^{(l-1)*} \tag{5.28}$$

By introducing the variance of the reciprocal of the channel DFT coefficients

$$M_H = E\left[\left|\frac{1}{H_p}\right|^2\right] \tag{5.29}$$

we can compute the variance of the estimate (5.28) as follows

$$\begin{aligned} E\left[\left|\hat{r}_1^{(l)} - r_{S_p,\hat{S}_p^{(l-1)}}\right|^2\right] &= E\left[\frac{1}{P}\sum_{p=0}^{P-1}\left|\frac{W_p\hat{S}_p^{(l-1)*}}{H_p}\right|^2\right] \\ &= M_W M_H \frac{1}{P}\sum_{p=0}^{P-1} M_{\hat{S}_p^{(l-1)}} \end{aligned} \tag{5.30}$$

Hence for very dispersive channels, M_H yields a significant increase of the estimate variance. A more reliable estimate is obtained by taking into account only the frequencies whose channel gain is larger than a given specific threshold H_{TH}.

The following set of frequencies are defined

$$S = \{ p : |H_p| \geq H_{TH} \} \tag{5.31}$$

having cardinality P_S . A new estimate of the correlation is

$$\hat{r}_2^{(l)} = \frac{1}{P_S} \sum_{p \in S} \frac{R_p}{H_p} \hat{S}_p^{(l-1)*} \tag{5.32}$$

where

$$|H_{TH}|^2 = \frac{M_W}{M_{S_p}} \tag{5.33}$$

A good correlation estimate must be determined such that the probability of overestimating $|r_{S_P, \hat{S}_P^{(l-1)}}|$ is sufficiently small.

5.2.5.2 Estimation of the Data Signal Power

The power of the detected data, $M_{\hat{S}_P^{(l)}}$ can be approximated with the average power PM_S , where M_S is the statistical power of s_n . If the values of s_n are equally likely

$$M_s = E[|s_n|^2] = \frac{1}{N} \sum_{\alpha \in A} |\alpha|^2 \tag{5.34}$$

with N the alphabet size.

However, in accordance with (5.32) and , a better estimate is given by

$$\hat{M}_{S_P^{(l-1)}} = \frac{1}{P_S} \sum_{p \in S} \left| \hat{S}_p^{(l-1)} \right|^2 \tag{5.35}$$

5.3 Observations on the Frequency Domain Equalizers

All the above mentioned frequency domain equalizers have been simulated in MATLAB. The FD-LE works very well giving performance comparable to OFDM. The FD-DFE performs a little better than FD-LE in high SNR conditions at the expense of increase in complexity. This is because it can cancel the postcursors of the ISI. This DFE design has an added time domain feedback filter which requires matrix inversion for Equations (5.6) and (5.15) which has high computational value. The IBDFE, on the other hand, has the ability to cancel both the precursors and the postcursors. In both the DFEs the error propagation is limited to one block only. IBDFE also has the feedback filter in the frequency domain, hence decreasing the complexity as compared to H-DFE.

As explained for IBDFE, the filter coefficients are totally dependant on the correlation factor of the transmitted and the detected data given by(5.22). If the estimation of this correlation is correct, the design of the filter coefficients will show improved performance. But as this design scheme does not use a training symbol, we cannot get a good correlation estimation, as the channel conditions become more severe.

Chapter 6

Data-Aided Block Decision Feedback Equalizer (DAB-DFE)

In this This chapter explains the Data-Aided Block Decision Feedback Equalizer which shows improvement over the IBDFE.

6.1 Drawbacks of IBDFE

As we had seen in the previous chapter, the design of filter coefficients of IBDFE depends on the estimation of some parameters. For HD-IBDFE, these parameters are Correlation Factor which gives the correlation between the transmitted and detected data and the Data Signal Power. As the transmitted data is unknown, the correlation factor cannot be estimated correctly. This incorrect estimation has the following effects on the equalizer performance [4]:

- If the correlation factor is underestimated, it will not cancel the ISI very efficiently and will be a cause of slower convergence.
- If the correlation factor is over-estimated, the feedback filter may cancel the ISI completely. As the correlation factor is calculated by the detected data and not the actual transmitted data, it will lead to worse performance.

It has been verified that the overestimation of the correlation factor leads to worse performance than the previous iteration, with catastrophic results on the equalizer performance [4].

6.2 Training Sequence Format

In order to overcome this problem, we append a training sequence prior to the data symbols. The training sequence should be of the same modulation scheme as that of the transmitted signal. For example, if the modulation of the transmitted data is QPSK then the training sequence should be randomly generated sequence with the QPSK constellation mapping.

Since we assume that the channel remains the same for the duration of our burst transmission, the training symbol hence designed will deliver us an accurate result that can be used for the filter optimization and the same filter calculations will suffice for the rest of the symbols in the frame as they have experienced the same channel due to transmission that proceeds within the duration of the coherence time.

6.3 Estimation of the Correlation Factor

The correlation factor is calculated by the following relation

$$r_{S_P,\hat{S}_P^{(l-1)}} = E[T_P \hat{T}_P^*] \tag{6.1}$$

Where T_P is the transmitted training sequence and $\hat{T}_P$ is the output of the slicer for the training symbol. Now the correlation factor is calculated only once and not for multiple iterations causing a reduction in the computational complexity.

Also the calculation of the correlation factor by the above procedure would allow us to determine this correlation perfectly since we feedback the correct decisions. Hence there will not be a problem of underestimating or overestimating this correlation factor. Using this process, we alleviate the basic discrepancy in IBDFE as the correlation factor is involved in determining the necessary feedforward and feedback filter coefficients.

The reduction in computational complexity is another big advantage in using DAB-DFE. Since we do not need to calculate this correlation factor in successive iterations, its value can be calculated once and can be stored to reference it for use in subsequent blocks in the frame.

6.4 Filter Coefficients

The coefficients of the DAB-DFE are calculated by minimizing the Mean Square Error at the detection point. Because of the perfect knowledge of the correlation factor, the forward filter coefficient is given by

$$C_P = \frac{H_P^*}{M_W} \tag{6.2}$$

Thus the forward filter is perfectly matched to the channel. The feedback filter reduces to

$$B_P = -(H_P C_P - \gamma) \tag{6.3}$$

where

$$\gamma = \sum_{p=0}^{P-1} H_P C_P \tag{6.4}$$

Now the feedback filter removes all the ISI. DAB-DFE shows improvement in performance over the contemporary DFEs. The use of training sequence for the optimization of the filter coefficients makes it a robust algorithm, although we have to compromise on the spectral efficiency. It eliminates the need for iterations as the training of the equalizer is done on the basis of reliable data.

For the assumption that the channel does not much change much over the duration of transmission, this transmission must occur within the coherence time. This is a common case with the design of mobile communication channels where the fast fading phenomenon is avoided by using a frame size whose duration does not exceed the designated coherence time.

Chapter 7

Performance Evaluation

7.1 Comparison with Imperfect Channel Estimate

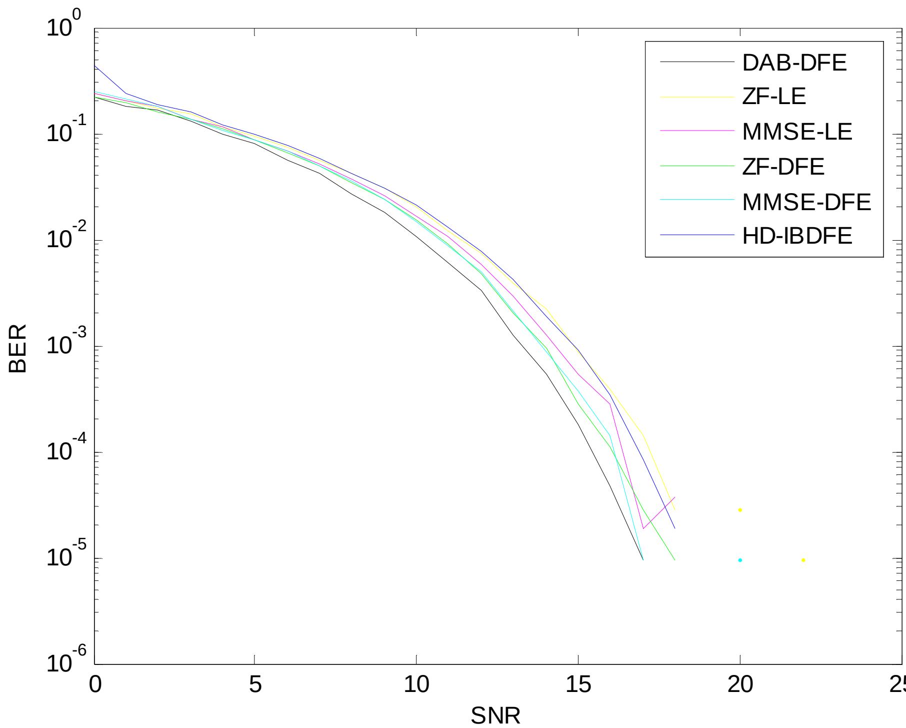

Figure 7-1 Comparison in case of Imperfect Channel Estimate

First of all we compare all the equalization techniques by feeding them all with imperfect channel estimation. Simulations are run for all the equalizers in frequency selective channel. Figure 7.1 gives the BER plots for the QPSK modulation. For all the comparisons we have used Minn's algorithm [6] for timing synchronization; frequency synchronization has been assumed to be perfect. As evident from the figure, DFEs show better performance than the Linear Equalizer particularly for high

SNR values. This is because as the SNR values increase more correct decisions are fed back which increases the affectivity of the DFEs to cancel ISI. We can see that the DAB-DFE performs better than or equal to the other equalization techniques.

7.2 Comparison of Equalization Schemes in Flat Fading Channel

In all the following comparisons perfect channel estimation is assumed. The assumptions on timing and frequency synchronization remain the same. First the comparison is carried out in a flat fading Rayleigh Channel. In this case all the equalizers show almost the same performance except for the DAB-DFE which gives an improvement in performance.

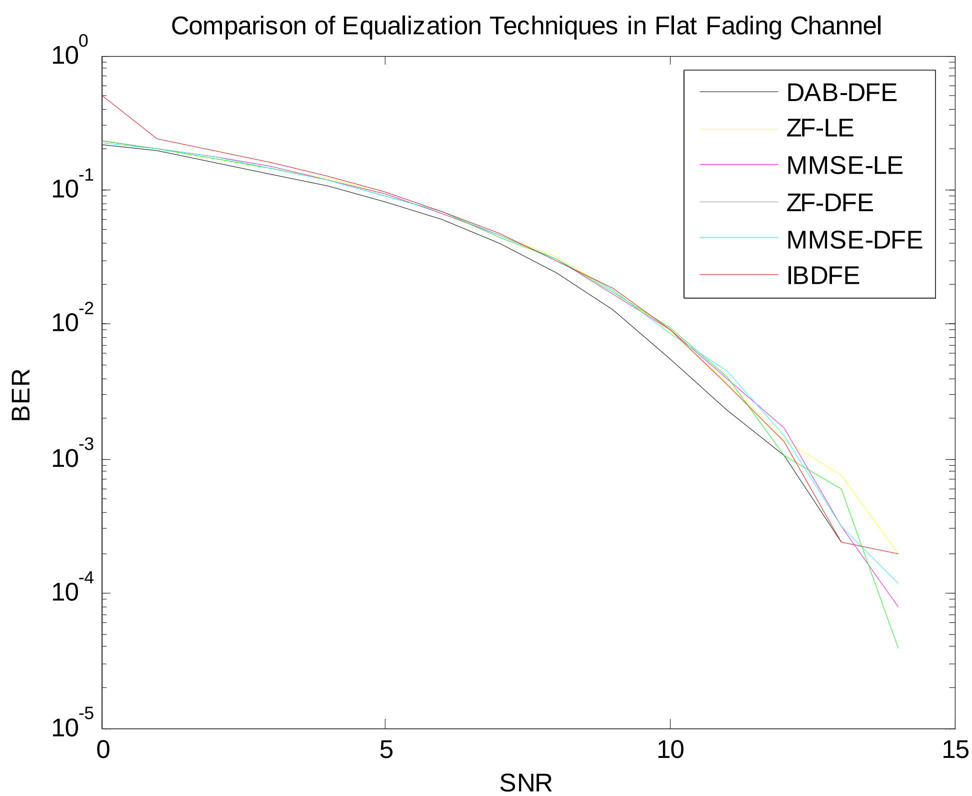

Figure 7-2 Comparison of Equalization Techniques in Flat Fading Channel

7.3 Comparison of Equalization Schemes in Frequency Selective Channel

Comparison in frequency selective channel gives the results of Figure 7.3. For another comparison we give all the equalizers perfect channel estimation and for DFEs we feedback the perfect decisions. DAB-DFE estimates the channel with the algorithm. We can see that even in these conditions DAB-DFE's performance is not worse than the rest of the equalizers.

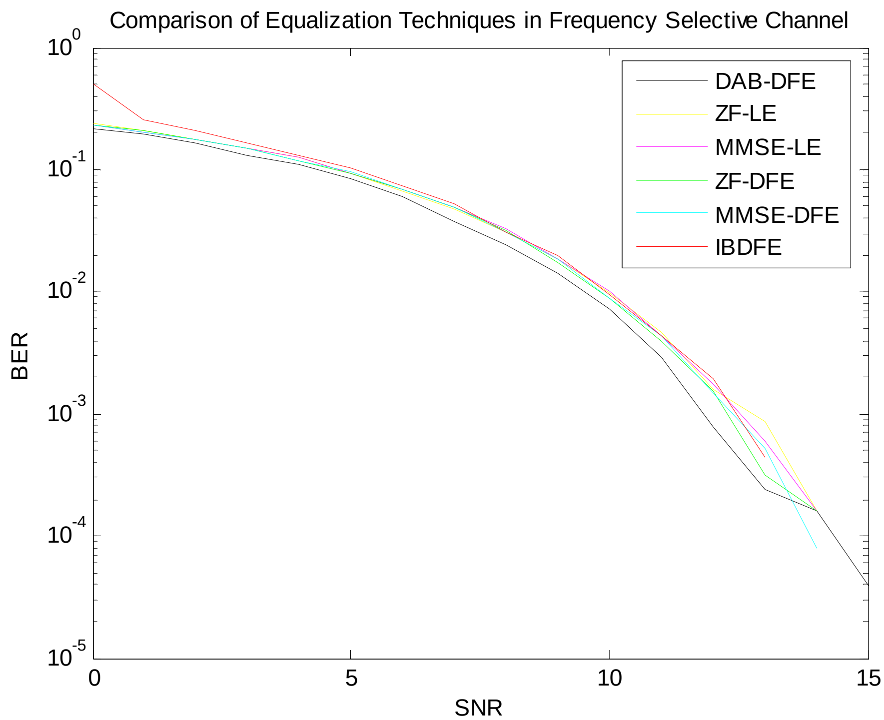

Figure 7-3 Comparison in Frequency Selective Fading Channel

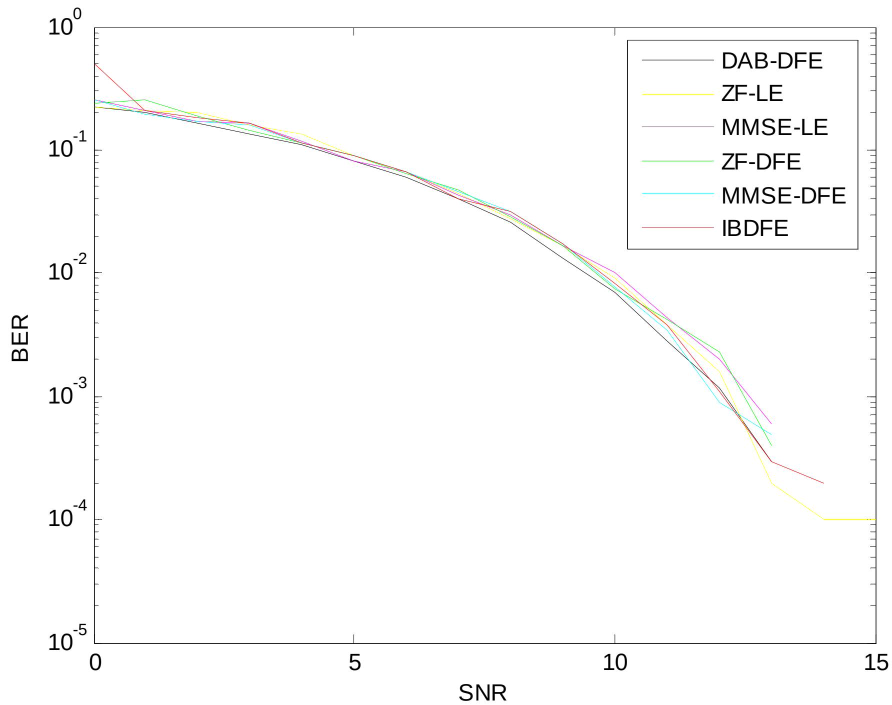

Figure 7-4 Perfect channel estimation for all the equalizers while imperfect channel estimation for DAB-DFE

7.4 Channel Estimation of a Frequency Selective Channel

Figure 7.5 gives the frequency response of a multipath channel that we have used for simulations. Figure 7.6 gives the impulse response of the same channel. The estimate of the channel as given by the channel estimation algorithm is given in Figure 7.7. As we can see, this algorithm gives a very good performance.

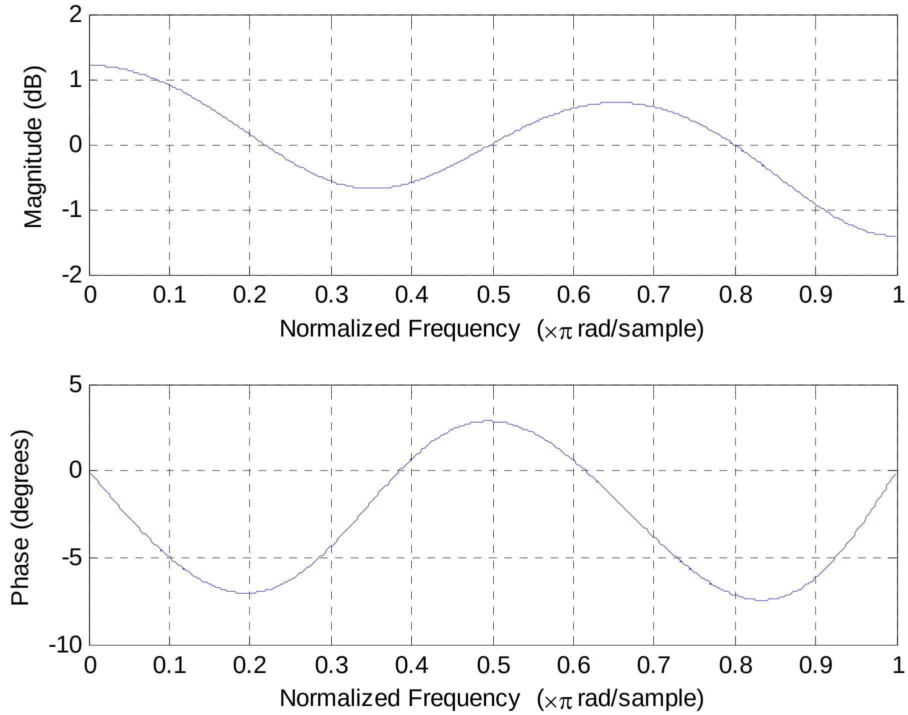

Figure 7-5 Frequency Selective Channel Model

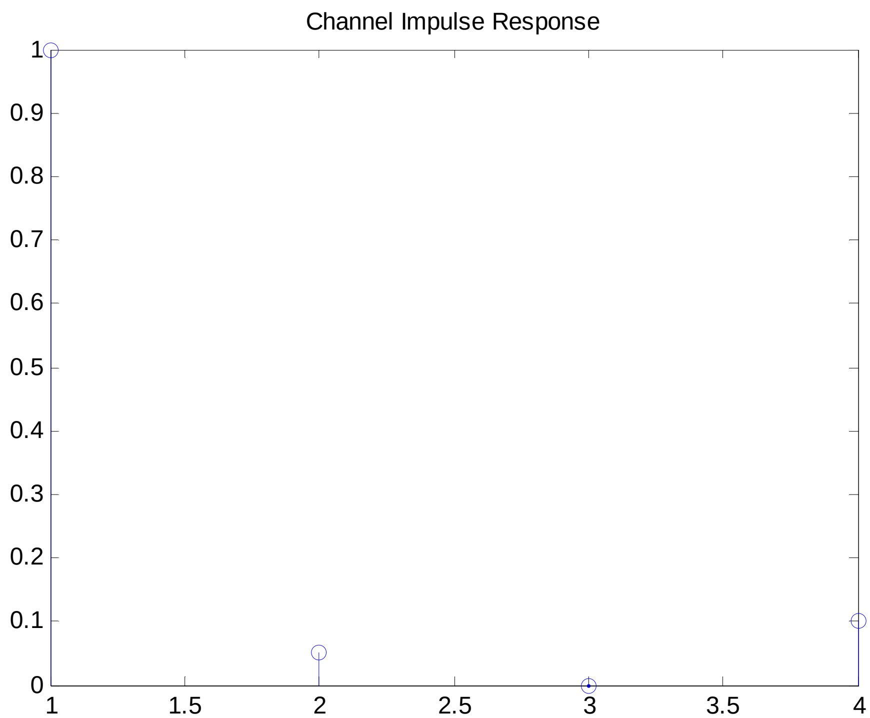

Figure 7-6 Impulse Response of the Frequency Selective Channel

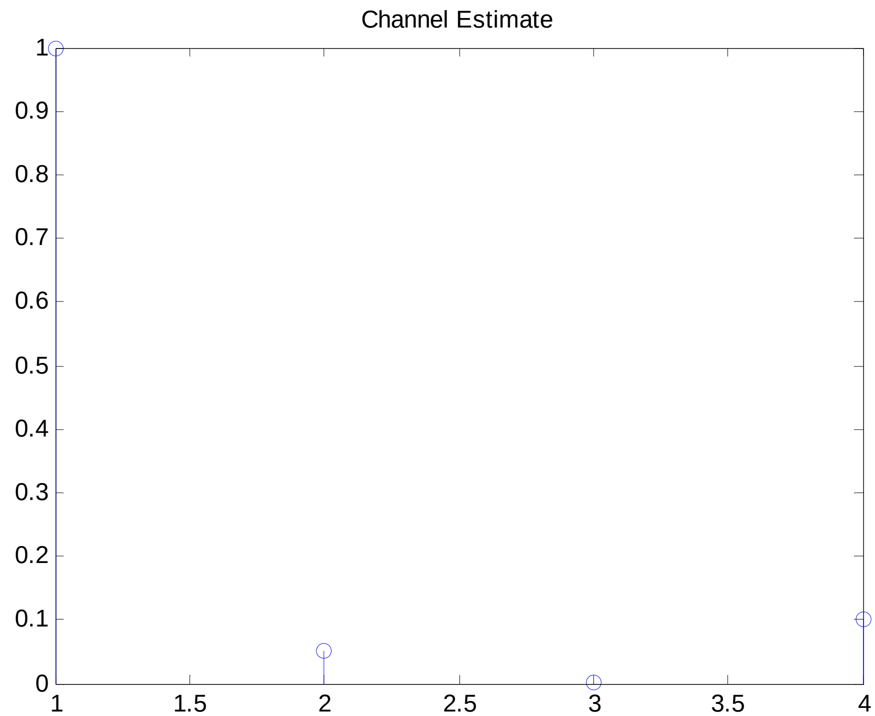

Figure 7-7 Estimate of the Channel

7.5 Comparison of the DAB-DFE with OFDM

We have compared Single Carrier system with Frequency Domain Equalization with OFDM system. Figure 7.8 gives the comparison of DAB-DFE with OFDM. Both the systems are evaluated in the same conditions. This gives us a very encouraging result and supports the claim that Single Carrier with frequency domain equalization gives performance comparable to OFDM. Comparisons with other equalizers will also give the same results. We have however, compared them with non-coded OFDM. OFDM systems are not used without coding. Hence, if the coding is used for OFDM, its performance will also improve.

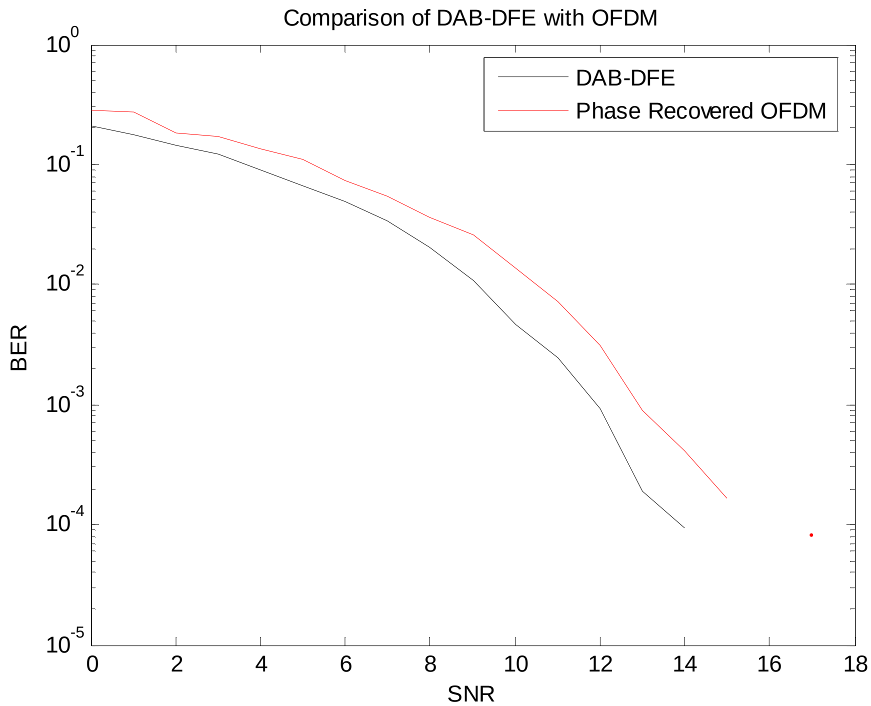

Figure 7-8 Comparison of the DAB-DFE with OFDM

Chapter 8

Conclusions

The need for high data rates calls for the need of greater bandwidth. If the signal bandwidth is increased it becomes larger than the coherence bandwidth of the channel. The signal, thus, experiences frequency selective fading or the multipath fading. Several techniques have been investigated to provide reliable data communication in frequency selective channels. Single Carrier with Time domain Equalizer processing at the receiver is one. This approach combats the multipath phenomenon for small channel impulse responses. However, as the impulse response gets larger the complexity of time domain equalizer grows exorbitantly, rendering it improper for use in communication systems. Orthogonal Frequency Division Multiplexing is another technique used for this purpose. This has been popular recently and has been employed in various communication systems due to its good performance and simplicity of structure. It transmits the data by sending orthogonal sub-carriers in parallel by using FFT operator. At the receiver the data is retrieved by demodulating the received signal with an IFFT operator and equalizing it with one-tap equalizer. There are however, some drawbacks associated with this mode of communication. OFDM exhibits high Peak to Average Power ratio and therefore requires a large linear range of the amplifier. In addition it is very sensitive to carrier frequency offsets. Single Carrier modulation with Frequency Domain Equalization has been proposed to give performance comparable to OFDM with almost the same overall complexity. It is a topic of recent interests and various comparisons have been done on the performance of the two schemes. In addition various equalization techniques i.e. Linear Equalization, Decision Feedback Equalization etc have been implemented in the frequency domain as well. We have simulated the Single Carrier communication system and different frequency domain equalizers available in literature in MATLAB.

We have found that Single Carrier with Frequency Domain Equalization gives performance comparable to OFDM. If uncoded transmission is used for both the modulation schemes, SC-FDE outperforms OFDM as the later has strict coding requirements. We have compared different frequency domain equalization techniques and have found that FD-DFE gives a slightly better performance than FD-LE for high SNR values with an increased complexity. Iterative Block DFE has a reduced complexity and gives better performance than the two but is highly dependant on the correlation factor, which if not estimated correctly can worsen the performance. We propose the use of Data-Aided Block DFE which sends a training sequence prior to the data symbols to get optimized filter coefficients and have found that it outperforms all the contemporary equalizers. Hence, we can conclude that Single Carrier with Frequency Domain Equalization gives very efficient performance in multipath fading channels and can be used for reliable high data rate communication.

8.1 Future Work

Further work can be done on DAB-DFE to design a new training sequence. This training sequence should be such that it should be used for timing and frequency synchronization, channel estimation and Equalizer Coefficients' optimization. This will increase the spectral efficiency with the use of DAB-DFE yet reaping its benefits as well. We had assumed that the transmission time is lesser than the coherence time. SC-FDE in fast fading environments is also a point of lot of interest [10]. Future research should be carried out to extend the design of this system for fast fading channels.

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