Performance of Decision Feedback Equalizer over Frequency Selective Fading Channel in W-CDMA Systems

B.Praveen Kumar, D.Vishnu Vardhan

***Abstract*—The CDMA systems provide the greater capacity in wireless communication environment. But a major obstacle for this is multipath fading which causes inter symbol interference (ISI). The conventional sub optimum receiver consisting of a bank of matched filters is often inefficient because interference is treated as noise. The device used to combat with ISI is named as equalizer. Among linear and non linear equalizers, non linear equalizer outperforms over linear equalizers. Here decision feedback equalizer is non linear equalizer. This paper considers a non linear equalizer such as decision feedback equalizer (DFE) to mitigate inter symbol interference (ISI) in code division multiple access (CDMA) systems.**

***Index Terms*—Feed-forward and feedback filters, linear adaptive transversal equalizer (LTE), zero-forcing equalizer, direct sequence code division multiple access (DS-CDMA), and correlation.**

I. INTRODUCTION

The high flexibility and low cost of wireless communication systems increases demand for them. This led to development of multiple access schemes like wideband DS-CDMA systems [1] to achieve higher capacities. In digital wireless communication systems the major obstacle for reliable communication is inter-symbol interference (ISI), encountered in multipath transmission channels. To compensate the ISI channel equalization is used. The fading effects of multipath channel proportional to the complexity of the equalizer design [2]. The maximum likelihood sequence detection (MLSD) is used to combat with ISI. MLSD uses trellis based Viterbi algorithm [3]. The complexity Viterbi algorithm exponentially increases with number of states in trellis. The number of schemes, generally known as reduced-state sequence estimation (RSSE), were proposed in[4,5]. The recent research devoted to the receiver design using zero forcing, minimum MSE (MMSE) linear equalization methods combat with multipath fading channels [6,7]. The decision feedback equalizer minimizes the MSE between the input and decision device. The DFE generally outperforms the traditional linear equalizer, particularly if the channel has deep spectral null in its response.

A decision feedback equalizer based on MMSE criterion provides better performance than linear MMSE equalizers over highly dispersive channel. These findings are motivated to design a chip level DFE for CDMA downlink channel.

Manuscript received December 2013.

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To develop this type of system offers numerous challenges. First one is to develop the method for the chip level feedback from symbol decision. The second one is to suppress pre-cursor ISI, post cursor ISI from all users and current symbol interference from other users. The third one is to develop a method to detect signals buried in inter chip interference (ICI). Here ICI is interference between chips caused by frequency selective fading channels.

In this paper section II describes the system model for CDMA systems at transmitter end and also describes the channel model for CDMA systems. Section III describes the equalizers. In this section also derived the MMSE optimum feed-forward filter (FFF) tap weights and feedback filter tap weights. Section IV describes the DFE filter receiver for CDMA systems.

II. SYSTEM MODEL

A. Transmission end for CDMA systems

Consider a base station with J mobile stations in a WCDMA system. The j^{th} user is assigned is assigned to $f_{j,n}$ (k) (k=0,1,…..,P-1)of length P to spread bit w_j (n) at time i. After multiplexing, each mobile station through a common multipath channel. Assume the channel is FIR and has order q (q<P) with q+1 chip rate coefficients. Also assume, we oversample the channel output to generate M sub-channels. If we use h_m (n) to represent the mth composite sub-channel impulse response, including the transmitter, the physical channel and the receiver, then the output of the mth sub-channel due to user *j* is

$$u_{j,m}(n)=\sum_{l=0}^{q}h_m(l)s_j(n-l) \quad (1)$$

where

$$s_j(n)=\sum_{k=-\infty}^{\infty}w_j(k)f_{j,m}(n-kP)$$

The following assumptions are made in respect of codes and symbols mentioned above;

1. The spreading code is an independent identically distributed (i.i.d) white sequence with unit variance.
2. These spread codes are orthogonal to each other.

The BPSK modulated spread signal propagates through a dispersive multipath channel with the impulse response h(t). The impulse response of a multipath channel can be expressed as

$$h(t)=\sum_{l=0}^{N-1}h_m\delta(t-mT_c) \quad (2)$$

Where, N is total number of delayed paths and δ (t) denotes the Dirac-delta function. Finally, the received signal at receiver is

$$r(t)=\sum_{l=0}^{N-1} h_m u(t-mT)+\eta(t) \tag{3}$$

Where, η(t) additive white Gaussian noise (AWGN) with noise variance of σ^2. Assume that the received signal r(t) is discrete time signal, then received signal can be expressed as

$$r_k=\sum_{m=0}^{N-1} h_m u_{k-m}+\eta_k \tag{4}$$

Where r_k is the received signal sequence, h_m are tapped delay line (TDL) filter coefficients (m=0, 1,…..N-1),and η_k is white noise sequence. A single user is assumed to be considered to be approximated as a Gaussian noise in this paper without loss of generality.

B. *Channel estimation for CDMA systems*

The transmission signal and the impulse response of j^{th} user are expressed in (1) & (2) . From (4), the received multipath CDMA signals are depicted in fugure.1.

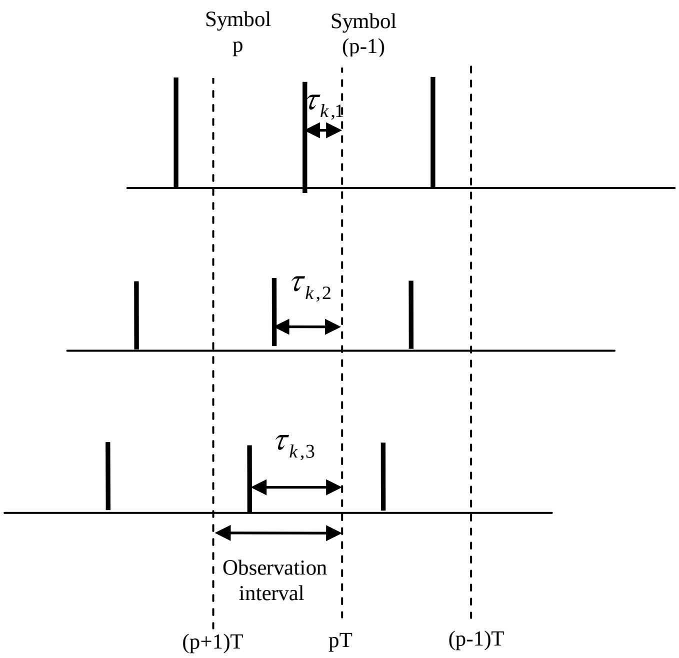

Figure.1: CDMA multipath channel

Here we assume that auto-correlation of multipath channel is

$$E\left[\sum_{l=0}^{N-1}\left|h_m\right|^2\right]=1 \tag{5}$$

Where E [.] denotes ensemble average. Now the impulse response multipath channel is $[h_0\ \ h_1 \ldots\ldots h_{N-1}]^T$. Where $[.]^T$ denotes transposition of a matrix.

III. EQUALIZERS

Equalization compensates for Inter Symbol Interference (ISI) created by within time dispersive channels. Coherence bandwidth is statistical measure of the range of frequencies over the channel can be considered “flat” (i.e., a channel which passes all spectral components with approximately equal gain and linear phase). If the transmitted signal bandwidth exceeds the coherence band width of radio channel ISI occurs and transmitted signal pulses are spread in time into adjacent symbols. ISI is major obstacle to high speed data transmission over wireless channels. An equalizer within a receiver compensates for the average range of expected channel amplitude and delay characteristics. Equalizers must be adaptive since the channel is generally unknown time varying. The classification of equalizers is depicted in figure.2.

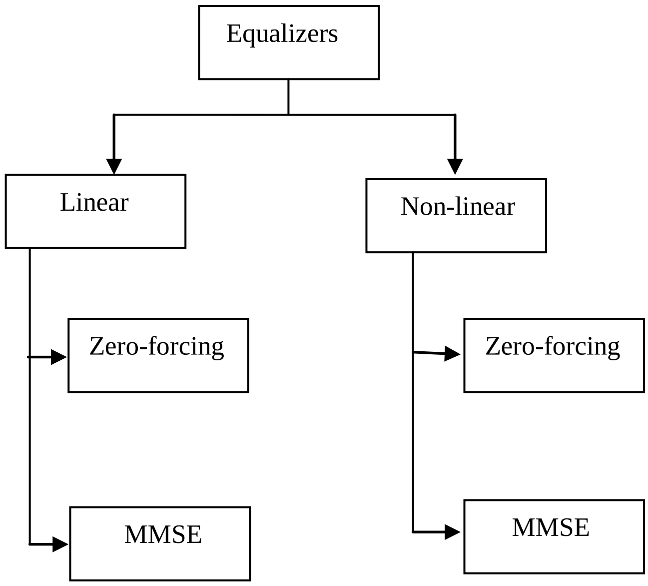

Figure.2: classification of equalizers.

The optimum tap weights adaptive linear equalizer is obtained by forcing the main lobe of received signal and all side lobes (ISI terms) to zero, named as zero forcing linear equalizer. The tap weights adaptive linear equalizer is obtained by Minimum Mean Square Error (MMSE) at the output of the linear adaptive equalizer. Linear equalizers do not perform well on channels which have deep spectral nulls in passband. In an attempt to compensate for the linear equalizer places too much gain in the vicinity of the spectral null, thereby enhancing the noise present in those frequencies. The basic idea behind the decision feedback equalization is that once an information symbol has been detected and decided upon, the ISI that it induces on future symbol can be estimated and subtracted out before detection of subsequent symbols. The baseband DFE filter depicted in figure.3.

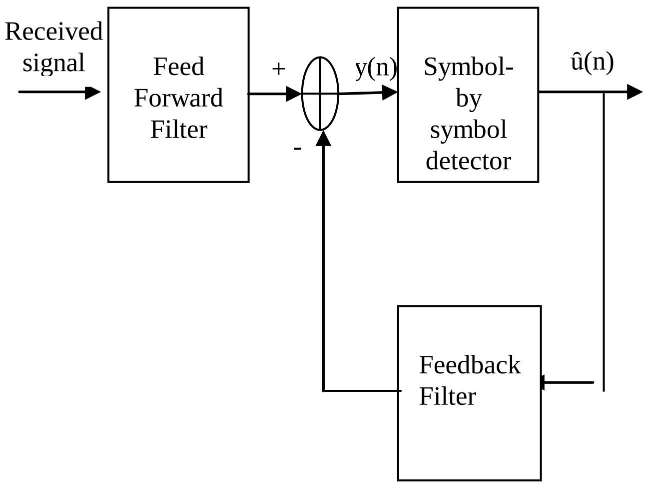

Figure.3: Baseband representation DFE Filter

Figure 1 depicts the digital baseband equivalent of the system. The received signal is equalized using a DFE which consists of a feed-forward filter (FFF) and a feedback filter (FBF). The feed-forward filter has L_f taps and its impulse response is denoted by the vector $\bar{f}=[f(0)\ f(1)\ldots\ldots f(L_f\text{-}1)]^T$ and the feed-back filter has L_b taps and its impulse response is denoted by the vector $\bar{b}=[b(0)\ b(1)\ldots\ldots b(L_b)]^T$. Then the input to the de-spreading box can be derived y_{n+1} as

$$y(n+1)=u(n+1)+b(1)[u(n)-d(n)]+z(n) \qquad (6)$$

Where $d(n)$ decision box output and $z(n)$ comprises of modeled post-cursor ISI residual precursor ISI, residual post-cursor ISI, filtered noise. $Z(n)$can be expressed as

$$z(n)=\sum_{i=2}^{L_b} b(i)(u(n-i)-d(n-i))+\sum_{i=2}^{\Delta-1} c(i)u(n+\Delta-i)$$

$$+\sum_{i=\Delta+L_b+1}^{L_h-1} c(i)u(n+\Delta-i)+\sum_{i=0}^{L_f-1} f(i)\eta(n-i). \qquad (7)$$

Where $c=h*f$, symbol * denotes convolution operation, is combined impulse response of both channel and feed-forward filter. And Δ denotes overall delay caused by system. Assume all previous decisions are correct i.e. d(n-1)=u(n-1).The output of decision device can be expressed as

$$d_n=\sum_{l=0}^{L_f-1} f(l)r(n-l)-\sum_{l=1}^{L_b} b(l)u(n-l-\delta) \qquad (8)$$

The error and error variance of the signal can be expressed as

$$e(n)=d(n)-u(n-\delta) \qquad (9)$$

$$J=E\left[\left|e(n)\right|^2\right] \qquad (10)$$

Here, invoke the orthogonality principle then the error variance is

$$J=\bar{f}^H(\phi_{hh}+\sigma_n^2 I)\bar{f}+\bar{b}^H\bar{b}+1-\bar{f}^H H\bar{b}$$

$$-\bar{b}^H H^H\bar{f}-\bar{f}^H h-h^H\bar{f} \qquad (11)$$

To obtain optimum filter weights differentiate J with respect to $\boldsymbol{f}^*$ *and* $\boldsymbol{b}^*$, respectively, we obtain as $\boldsymbol{f}_{opt}$ and $\boldsymbol{b}_{opt}$ as

$$\bar{f}_{opt}=((\phi_{hh}-HH^H)+\sigma_n^2 I)^{-1}h \qquad (12)$$

$$\bar{b}_{opt}=H^H\bar{b} \qquad (13)$$

The MMSE of DFE is given by

$$J=1-\bar{f}_{opt}^H h \qquad (14)$$

The FFF and FBF are cancels the perfectly cancels the precursor and post-cursor ISI respectively.

IV. DFE FILTER FOR CDMA SYSTEMS

To mitigate the chip level ISI in CDMA systems, the DFE filter is depicted in figure.4. Here assume that previous detected symbols are correct, z(n+1) is conditionally uncorrelated to [$e(n)$|y(n)]. The filtered noise is uncorrelated with the decision error $e(n)$. The residual ISI components has some correlation with the decision error $e(n)$.but assume that the correlation is small enough to be ignored.

In chip level DFE system de-spreading of chips is done at the input of decision box. The decisions are again spreaded to achieve the chip level subtraction at output of FFF of DFE

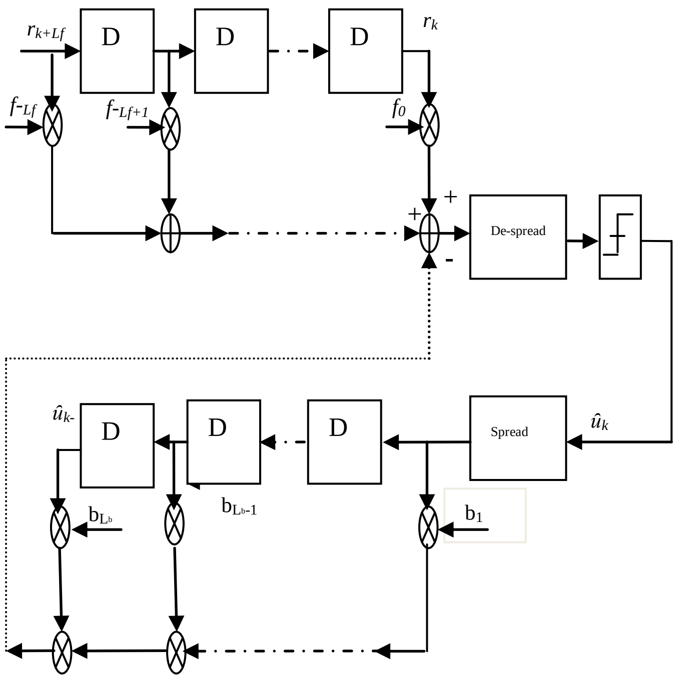

Figure.4: The DFE filter for CDMA systems

To minimize the The optimization of filter taps of DFE can be achieved by the forcing the ISI terms zero equating the main lobe to unity. the optimized filter taps of DFE can also achieved by MMSE-DFE from the identity (13) and (14).

V. SIMMULATION RESULTS

For simulations at transmitter end generate N random bits with mean zero and unit variance. Convert these bits into BPSK modulated symbols. Generate orthogonal spreading code whose length is multiple of 2, for example 2,4,8,16….. . Spread the BPSK modulated symbols into chips by using the generated orthogonal spreading code. Orthogonal spreading codes are generated by Walsh-Hadamard orthogonal codes or pseudo random noise orthogonal codes. Orthogonal codes are reduces inter-channel interference in CDMA systems. Generate the multipath channel impulse response whose length is greater than the orthogonal spreading code. This make the channel delay spread is greater than the chip rate of symbol. Now the channel is frequency selective faded channel. Transmit the chips through this multipath channel. In this channel additive white Gaussian noise is added to each chip. The additive white Gaussian noise channel has power spectral density N_o /2.

At receiver end, to attenuate the precursor ISI FFF used. Here the minimum number of FFF taps used to mitigate the ISI in chips is equal to product of spreading code and the channel memory. The channel auto-correlation function is calculated using channel impulse response. From this calculate the optimum feed forward filter tap weights from, using identity(12) feedback filter tap weights are calculated. The maximum number of feedback filter taps is not greater than feed forward filter taps. Assume that all past decisions of decision device are correct. The feedback filter tap weights are calculated from the identity (13).

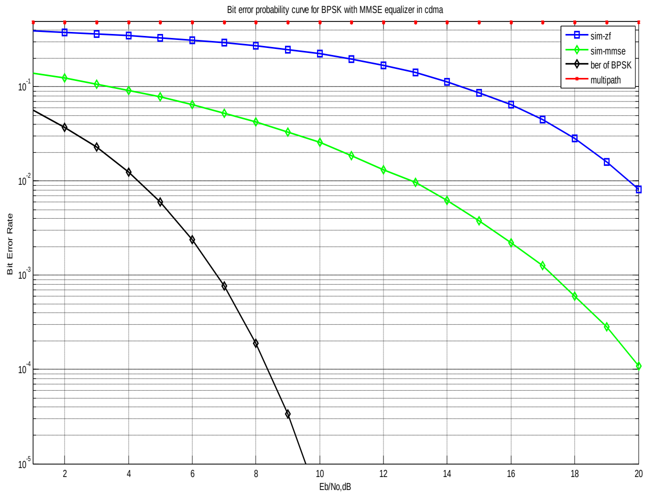

Figure.5:BER performance comparison between zero-forcing Linear equalizer and MMSE linear equalizer.

The bit error rate (BER) response of linear equalizer in CDMA is shown in figure(5). Here the BER response of zero-forcing DFE filter, MMSE DFE filter over AWGN channel compared with the BPSK signal BER performance over AWGN channel.

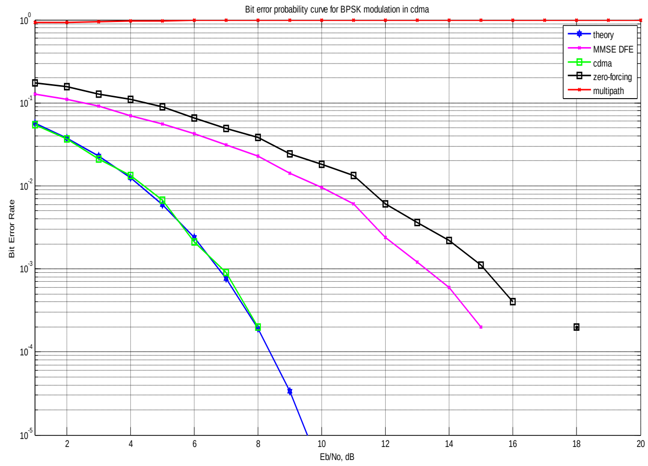

Figure.5:BER performance comparison between zero-forcing Linear equalizer and MMSE DFE equalizer

The bit error rate (BER) response of DFE in CDMA is shown in figure. Here the BER response of zero-forcing DFE filter, MMSE DFE filter over AWGN channel compared.

By comparing the simulation results of linear equalizer and DFE it is clear that at BER of 10^{-4} the difference of Eb/No ratio is about 4dB. This difference is increasing at higher Eb/No. by this simulation results the performance of MMSE DFE equalizer better than the zero-forcing DFE and also the DFE performance much better than the linear equalizer.

VI. CONCLUSION

In this paper, decision feedback equalizer using zero-forcing and MMSE criterion developed for CDMA systems, based on correlation technique for channel estimation. It has been shown that the performance MMSE-DFE system is better than zero-forcing DFE and linear equalizer. We would like to investigate this to turbo DFE environment. This issue can taken up as future scope the work.

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