Decision Feedback Equalization Receiver for DS-CDMA with Turbo Coded Systems

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Abstract — **In this paper, adaptive equalizer receiver for a turbo code direct sequence code division multiple access (DS-CDMA) by using least mean square (LMS) adaptive algorithm is presented. The proposed adaptive equalizer is using soft output of decision feedback adaptive equalizer (DFE) to examines the output of the equalizer and the Log- maximum a posteriori (Log-MAP) algorithm for the turbo decoding process of the system. The objective of the proposed equalizer is to minimize the bit error rate (BER) of the data due to the disturbances of noise and intersymbol interference (ISI) phenomenon on the channel of the DS-CDMA digital communication system. The computer program simulation results shown that the proposed soft output decision feedback adaptive equalizer provides a good BER than the others one such as conventional adaptive equalizer, infinite impulse response adaptive equalizer.**

Keywords — **adaptive equalizer, turbo code, DS-CDMA.**

1. Introduction

Direct sequence code division multiple access (DS-CDMA) is a candidate access technique for a third generation mobile system[12]. In DS-CDMA systems, signal transmission over time-varying multipath mobile radio channel produces intersymbol interference (ISI), multiple access interference (MAI), and other time-varying effect from the channel. In this case, a DS-CDMA receiver with channel equalization is generally required to mitigate the effect of ISI and MAI.

A turbo code, which were discovered by C. Berrou *et al.* in 1993 provide significant improvements in the quality of data transmission over a noisy channel [1]. Iterative Log-MAP algorithm is an important kind of turbo decoding algorithms [6, 11].

In order to improve the performance of the DS-CDMA system. In this paper proposes a new adaptive equalizer based on decision feedback equalizer using soft output to calculates the output of the equalizer and the Log-MAP algorithm for the turbo decoding process of the system. First, the system model for the DS-CDMA system is presented. In section 3 presents the turbo codes herein including turbo encoder and the turbo decoder. Section 4, conventional adaptive equalizer, infinite impulse response adaptive equalizer and soft output decision feedback equalizer are presented. In section 5 presents the adaptive least mean square algorithm for the equalizers. Computer simulations for several situations to demonstrate the performance of the system and conclusions are given in section 6 and 7, respectively.

2. System Model

The block diagram of the CDMA system model is shown in Fig. 1. We consider the uplink of a DS-CDMA system with K users. We assume that the signals of the users arrive at the receiver synchronously and that the spreading codes are known at the receiver. The modulation scheme is linear. d_i^k is the kth symbol of duration T_s , transmitted by user i. The spreading code $c_i(n)$ of length L_c assigned to user i can be written as [3]:

$$c_i(n) = \sum_{k=0}^{Lc-1} c_i^k x(n-kT_c) \tag{1}$$

with

$$x(n) = \begin{cases} 1\, if\ n \in [0, T_c] \\ 0\ otherwise \end{cases}$$

where T_c is the chip duration ($T_c = T_s / L_c$) and $c_i^k \in \{-1,1\}$.
The spread signal is transmitted by user i given by:

$$s_i(n) = \sum_{k=0}^{\infty} d_i^k c_i(n-kT_s) \tag{2}$$

The spread signal is transmitted over the frequency and time selective channel $h_i(n)$:

$$h_i(n) = \sum_{k=0}^{W-1} a_i^k \delta(n-kT_c) \tag{3}$$

The path gain a_i^k is a complex gaussian random process with zero mean. Let us now introduce a guard interval before the transmission of information symbol in order to cancel ISI. T_s must therefore be replaced by τ in Eq. (2). The minimum value of τ is: τ

$= T_s + (W\text{-}1)\ T_c$. The received signal can be expressed as:

$$r(n) = \sum_{i=0}^{\infty} \sum_{k=0}^{K-1} d_i^k c_k(n-i\tau) * h_k(n) + n(n) \tag{4}$$

where $n(n)$ is a complex additive white gaussian noise with zeros mean and a variance of σ^2.

The waveform $p_k(n)$ assigned to user k is: $p_k(n) = c_k(n) * h_k(n)$. We suppose a perfect channel estimation, hence $p_k(n)$ is perfectly know at the receiver. This channel estimation can be performed by transmitting bursts of data and inserting a training sequence in each burst. In the following the channel is assumed to be constant over the duration of a burst. Eq. (4) can be rewritten as:

$$r(n) = \sum_{i=0}^{\infty} \sum_{k=0}^{K-1} d_i^k \, p_k(n - i\tau) + n(n) \qquad (5)$$

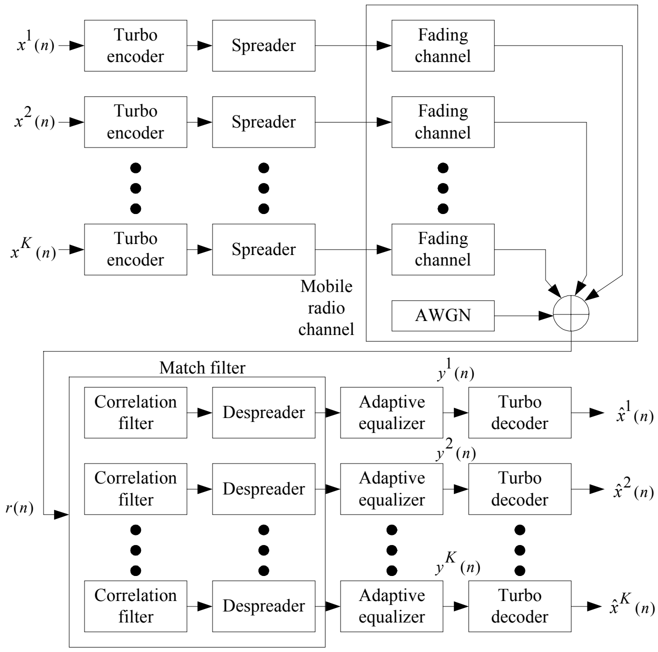

Figure 1. The system model

3. Turbo Codes [11]

3.1 Turbo Encoder

Turbo coding, illustrated in Fig. 2, uses a parallel forward error control (FEC) encoding scheme. With this scheme, the information is systematically encoded by two separate encoders. Often, the two encoders are identical, and to ensure that they do not produce the same output, the information bits are reordered through the use of a pseudorandom interleaver before the second encoding stage. This pseudorandom interleaver is often referred to as the turbo interleaver to distinguish it from channel interleavers. The information bits and the parity bits generated by the two encoders are then transmitted over the channel. In a typical turbo encoding scheme, the parity bits are punctured in a repeating pattern to increase the code rate, but this is not a requirement and puncturing refers to the omission of certain parity check bits in the code so as to increase the data rate.

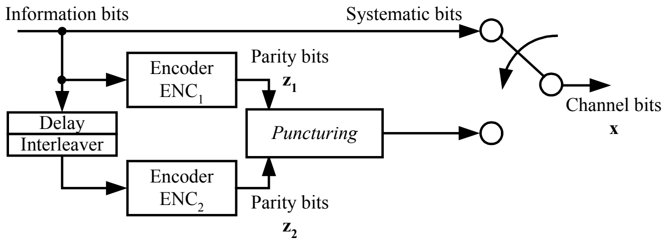

Figure 2: Rate 1/2 turbo encoder.

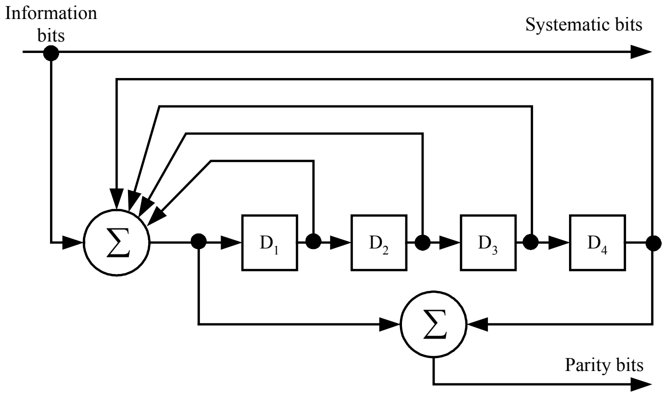

Figure 3: A systematic recursive convolution code structure.

Although any valid code is permissible, the component codes that are recommended for turbo encoding are short constraint length, systematic, recursive convolutional codes. With a systematic code, the original information bits are part of the transmitted sequence, shown in Fig. 2. A recursive convolutional code means that the code has a feedback structure, as opposed to the feedforward structure that characterizes the convolutional codes. The original turbo component code described by C. Berrou *et al.* [1] is illustrated in Fig. 3. This is a simple 16 state code. The feedback nature of these component codes means that a single bit error correspond to an infinite sequence of channel errors. To see this, consider an information sequence consisting of the binary digit 1, followed by infinite number of zeros, as the input to the encoder shown in Fig. 3. If the encoder starts in the all zero state, then it is straightforward to show that it never returns to that state with that kind of input [11].

3.2 Turbo Decoder

In addition to turbo codes having a novel parallel encoding structure, the decoding strategy applied to turbo codes is highly innovative. The scheme draws its name from the analogy of the structure of the decoding algorithm to the turbo engine principle. A block diagram of the decoder structure is shown in Fig. 4. To illustrate the processing of the turbo decoder, let ($\mathbf{x}$, $\mathbf{z}_1$, $\mathbf{z}_2$) be the vector of outputs from the turbo encoder of Fig. 2. Corresponding to these inputs applied to the channel, let ($\mathbf{u}, \zeta_1, \zeta_2$) be the vector of noisy outputs from the demodulator after matched filtering.

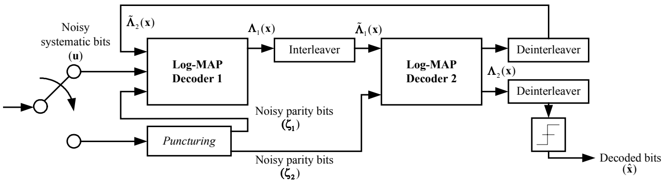

Figure 4: Turbo decoder.

In Fig. 4, the inputs to the first decoding stage are the channel samples ($\mathbf{u},\zeta_1$) corresponding, respectively, to the systematic (information) bits (**x**), the channel samples corresponding to the parity bits of the first encoder ($\mathbf{z}_1$), and the extrinsic information about the systematic bits that were determined from previous decoder stage ($\tilde{\Lambda}_2(\mathbf{x})$). We will define extrinsic information later, but suffice it to say that, on the first decoding iteration, the extrinsic information is zero.

Crucial to the performance of turbo decoding is the use of a soft input soft output (SISO) decoding algorithm. That is, not only does the algorithm accept a soft input, a real value whose magnitude indicates the reliability of the input, but it also produces a soft output, generally the probability that a particular bit is 0 and 1. We will describe a maximum a posteriori (MAP) probability decoding algorithm that meets this requirement.

The first decoding stage uses the MAP algorithm to produce a refined (soft) estimate, $\text{Prob}(x(n)\,|\,\mathbf{u},\zeta_1,\tilde{\Lambda}_2(\mathbf{x}))$, of the systematic bits. However, for the implementation, it is more convenient to express this soft estimate as the equivalent log-likelihood ratio:

$$\Lambda_1(x(n)) = \log\left(\frac{\text{Prob}(x(n)=1\,|\,\mathbf{u},\zeta_1,\tilde{\Lambda}_2(\mathbf{x}))}{\text{Prob}(x(n)=0\,|\,\mathbf{u},\zeta_1,\tilde{\Lambda}_2(\mathbf{x}))}\right) \tag{6}$$

In practice, calculation of this log-likelihood ratio is mush simple than equation (6) would suggest. Before entering the second stage of processing on the first iteration, the vector of refined estimates, Λ_1, is reordered to compensate for the turbo interleaving at the transmitter, where after it is combined with the second set of parity samples from the channel as the input to the second decoding stage.

The second decoding stage uses the MAP algorithm and the second set of parity bits to produce a further refined estimate, $\text{Prob}(x(n)\,|\,\tilde{\Lambda}_1((\mathbf{x}),\zeta_2))$, of the systematic bits. This second refined estimate is also expressed as a log-likelihood ratio, namely, $\tilde{\Lambda}_2(x(n))$. The estimate so produced can be hard detection, if so desired, to provide bit estimates at this point. Alternatively, the output of the second stage can be used to provide extrinsic information to the first stage. In either case, the estimates must be reordered to compensate for the turbo interleaving. As can be seen from Fig. 4, the first decoding stage, the interleaver, the second decoding stage, and the deinterleaver constitute a single loop feedback system, thereby making it possible to iterate the decoding process in the receiver as many times as is deemed necessary for a satisfactory performance. Indeed, this iterative process constitutes the turbo coding principle at the receiver. However, that during each set of iterations, the noisy input vector **u** is unaltered.

The decoding scheme of Fig. 4 relies on the implicit assumption that the bit probabilities remain independent from one processing stage to the next, the turbo algorithm makes use of the concepts of intrinsic and extrinsic information. Intrinsic information refers to the information inherent in a sample prior to a decoding operation. On the other hand, extrinsic information refers to the incremental information obtained through decoding. To maintain as much independence as is practically possible from one iteration to the next, only extrinsic information is fed from one stage to the next.

An important property of the MAP algorithm is that it includes both intrinsic and extrinsic information in a product relationship. However from a computational perspective, it is more convenient to express this relationship in terms of log-likelihood ratios, because then the product becomes a sum and the scaling term assumes a nonsignificant role. In particular, the extrinsic information at the output of the second stage is:

$$\tilde{\Lambda}_2(\mathbf{x}) = \Lambda_2(\mathbf{x}) - \tilde{\Lambda}_1(\mathbf{x}) \tag{7}$$

That is, the extrinsic information (expressed in logarithmic form) is the difference between the input and output log-likelihood ratios for the systematic bits.(On the first pass, $\tilde{\Lambda}_1(\mathbf{x}) = \Lambda_1(\mathbf{x})$.) Similarly, at the output of the first stage, the extrinsic information supplied to the second stage by the first stage is given by:

$$\tilde{\Lambda}_1(\mathbf{x}) = \Lambda_1(\mathbf{x}) - \tilde{\Lambda}_2(\mathbf{x}) \tag{8}$$

The basic MAP decoding algorithm is not changed in the turbo decoder. Details of the actual computation of Λ_1 and Λ_2 are omitted in Fig. 4 to simplify the exposition. The differences occur only in the processing, which is performed on the inputs to, and outputs from, the MAP algorithm.

On the last iteration of the decoding of the decoding process, a hard decision is applied to the output of the second decoder (after deinterleaving) to produce an estimate of the *n*th information bits[11]:

$$\hat{x}(n) = sign(\Lambda_2(x(n))) \tag{9}$$

4. Equalization

4.1 Conventional Adaptive Equalizer with Finite Impulse Response (FIR) Filter Structure [8]

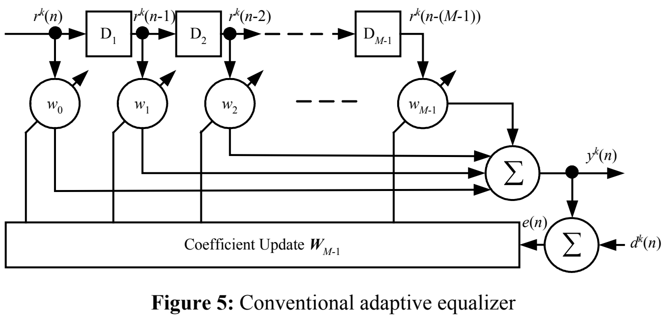

Figure 5: Conventional adaptive equalizer

A symbol-space adaptive equalizer is employed shown in Fig. 5, which has an FIR filter structure defined by

$$y^k(n) = \sum_{i=0}^{M-1} w_i r^k(n - i\tau) = \mathbf{w}^T \mathbf{r}^\mathbf{k}(n) \tag{10}$$

where M is the equalizer order, $\mathbf{w}(n)=[w_0\ w_1 \ldots w_{M-1}]^T$ is the equalizer weight vector, and $\mathbf{r}^\mathbf{k}(n)=[r^k(n)\ r^k(n-1) \ldots r^k(n-(M-1))]^T$ the equalizer input vector for user index number k.

4.2 Infinite Impulse Response (IIR) Adaptive Equalizer [8]

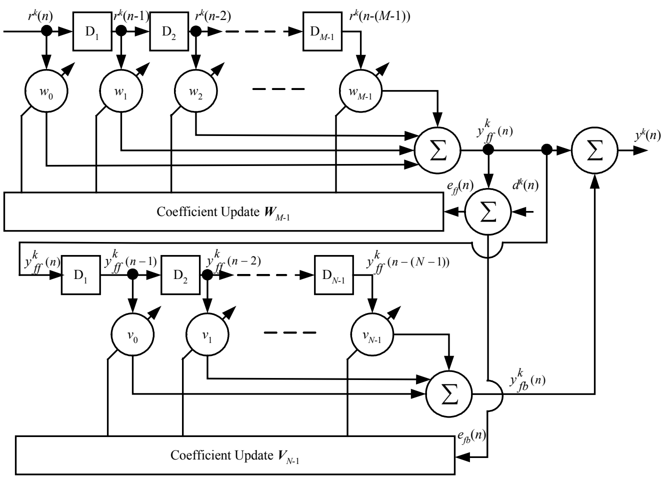

Figure 6: Infinite impulse response adaptive equalizer

The IIR adaptive equalizer, shown in Fig. 6, consist of two sections, a feed-forward section and feedback section. The feed-forward section is identical to conventional adaptive equalizer with M-1 Taps. The feedback section, on the other hand, is also a transversal equalizer with N-1 Taps whose function is to remove the ISI portion due to the previously detected symbols. The output is given as:

$$y_{ff}^k(n) = \sum_{i=0}^{M-1} w_i r^k(n - i\tau) \tag{11}$$

$$y_{fb}^k(n) = \sum_{i=0}^{N-1} v_i y_{ff}^k(n - i\tau) \tag{12}$$

$$y^k(n) = y_{ff}^k(n) + y_{fb}^k(n) \tag{13}$$

where $y_{ff}^k(n)$ is the output of feed-forward equalizer, and $y_{fb}^k(n)$ is the output of feedback equalizer.

4.3 Soft Output Decision Feedback Adaptive Equalizer

This section, we proposed the modified version of the decision feedback adaptive equlizer called soft output decision feedback adaptive equalizer (SO-DFE) as shown in Fig. 7. It is suitable for use in the CDMA communication turbo decoding system. The input of feedback path is given as [5]:

$$y_{df}^k(n) = sign(y_{ff}^k(n)) \tag{14}$$

where $y_{ff}^k(n)$ is the output of feed-forward part, as same as Eq. (11). The output of soft output decision feedback adaptive equalizer is given by :

$$y^k(n) = y_{ff}^k(n) + y_{fb}^k(n) \tag{15}$$

where $y_{fb}^k(n)$ is the output of feedback equalizer, as same as Eq. (12).

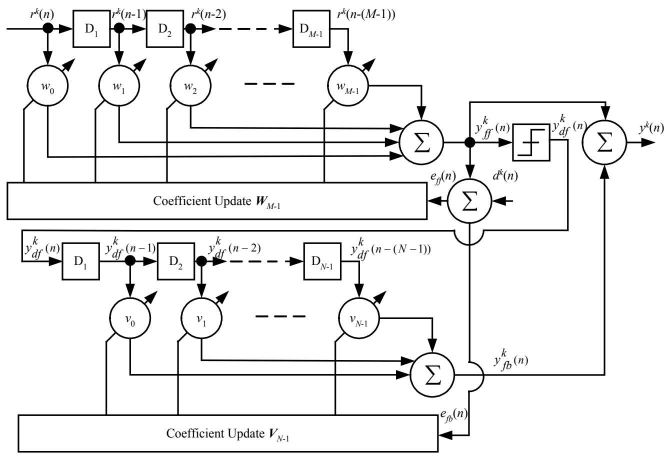

Figure 7: Soft output decision feedback adaptive equalizer

5. Adaptive Least Mean Square (LMS) Algorithm

The receiver structure consists of an adaptive LMS and a turbo decoder as shown in Fig. 1. Demodulation of DS-CDMA signals is conventionally achieved with a matched filter receiver. Herein, we assumed M Taps for the feed-forward section and N Taps for feedback section of the adaptive equalizer, respectively. In this paper, we used a well-known LMS adaptive algorithm to examines an appropriate of weights for the adaptive equalizer.

The filter weights are updated for conventional equalizer and feed-forward section of the equalizer can be expressed as:

$$w(n+1)=w(n)+\mu e_{ff}(n) r^k(n) \tag{16}$$

where $r^k(n)$ is the input signal of the receiver. $e_{ff}(n)$ is the error signal between the desired signals and the received signal. μ is the positive step-size parameter.

The filter weights are updated for feedback section of IIR adaptive equalizer given as:

$$v(n+1)=v(n)+\mu e_{fb}(n)\ y_{ff}^k(n) \tag{17}$$

where $y_{ff}^k(n)$ is the output signal of feed-forward section or the input signal of feedback section of the adaptive equalizer. $e_{ff}(n) = e_{fb}(n)$ is the error signal. And the adaptive algorithm for updated the filter weights of the feedback section of the soft output decision feedback adaptive equalizer given as:

$$v(n+1)=v(n)+\mu e_{fb}(n)\ y_{df}^k(n) \tag{18}$$

where $y_{df}^{k}(n)$ is the output signal of the decision feedback adaptive equalizer.

6. Numerical Results

The BER performance of the proposed adaptive equalizer with turbo code for the CDMA system receiver is presented in this section. The simulation results are obtained from the communication system over both ISI and AWGN channels. The main system parameters are defined as follows: length of transmission information 15 kbits (15 x 1024), support 4 users, filter length of feed-forward part $M = 11$, feedback part $N = 10$, power of each user = 8 dB, with 10 ensembles and the discrete channel response $a(n)$ is given by [7, 8]:

$$a(n) = \begin{cases} 0.5\left\{1+\cos\left(\frac{2\pi}{A}(n-2)\right)\right\}, & n = 1,2,3 \\ 0 & ,otherwise \end{cases} \quad (19)$$

where the factor A is introduced to allow scaling to customize the simulated ISI, herein $A = 2.9$.

And for turbo code: code generator metric = [1 1 1; 1 0 1] with constraint length = 3, channel code rate R = 1/2, and the number of iteration for turbo decoder = 1.

Fig. 8 shows the average BER performance of the perfect power control versus Eb/No for various values of the receiver. From Fig. 8, it is seen that the proposed adaptive equalizer provide a lowest BER performance for every Eb/No values. Fig. 9 shows the convergence process of the adaptive algorithm for the equalizers with the Eb/No = 8 dB. Fig. 10 shows mean square error (MSE) of each equalizer structure with the Eb/No = 8 dB.

7. Conclusion

This paper proposed the soft output decision feedback adaptive equalizer for DS-CDMA communication and using the attractive Log-MAP algorithm for the turbo decoding process of the system. From the simulation results, it is observed that the proposed adaptive equalizer show an improvement in the BER performance of the system as compared to the other adaptive equalizers.

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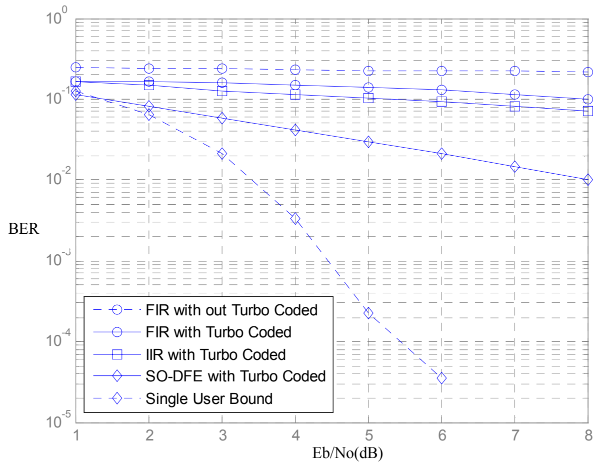

Figure 8: Performance of BER for Adaptive Equalizers

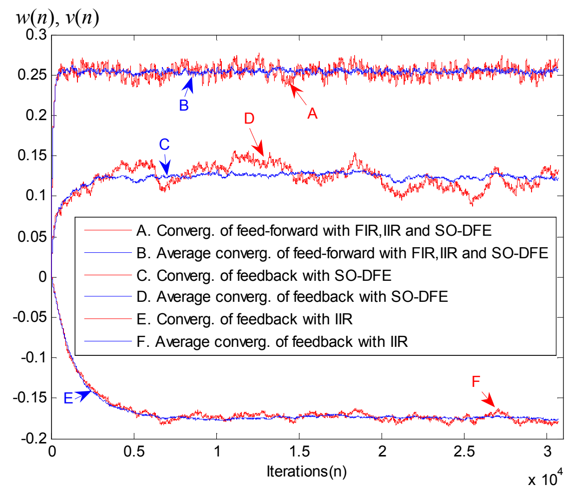

Figure 9: The convergence process with Eb/No = 8dB

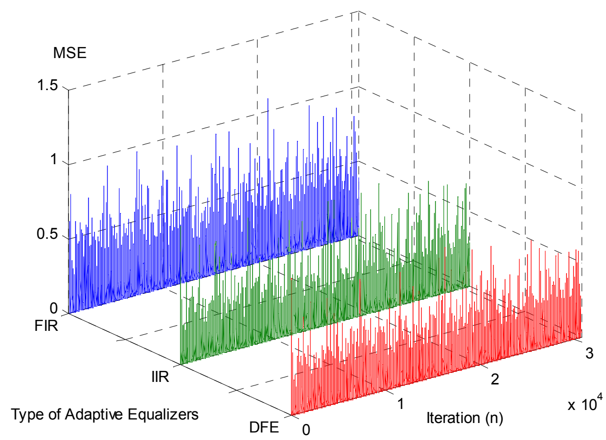

Figure 10: MSE with Eb/No = 8dB