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Overlap-Cut Frequency Domain Equalizer with Decision Feedback in the HSDPA Downlink

Chi-Min Li, Shao-Min Wen, Po-Jen Wang, Jia-Chyi Wu and I-Tseng Tang

Abstract— **In this paper, we propose a Frequency Domain Equalizer (FDE) without inserting the Guard Interval (GI) at the transmitter in the Wide-band Code Division Multiple Access (WCDMA) system. The proposed FDE adopts the Overlap-Cut (OC) method to avoid the Inter-Block Interference (IBI) and exploits the decision feedback structure to improve the Bit Error Rate (BER) performance. Without inserting the GI, the proposed FDE will be more compatible to the frame format of the current High Speed Downlink Packet Access (HSDPA).**

Index Terms— **FDE, IBI, Overlap-Cut, WCDMA**

I. Introduction

Recently, a Wideband Code Division Multiple Access (W-CDMA) system Release 5 [1] uses the High Speed Downlink Packet Access (HSDPA) technique to increase the system downlink data rate up to 14Mbps. Due to the multipath propagation, the orthogonality of the spreading codes is destroyed and the Bit Error Rate (BER) performance of the receiver will be degraded. Therefore, an equalizer has to be adopted at the receiving site to restore the code orthogonality and suppress the Multiple Access Interference (MAI). Generally, equalizers can be classified into two types: one is the Time Domain Equalizer (TDE) and the other is Frequency Domain Equalizer (FDE). The optimum BER solution for the TDE is via using the Minimum Mean Square Error (MMSE) criteria [2]. However, the MMSE-TDE involves the Direct Matrix Inverse (DMI) operation that requires very intensive computation to obtain its weighting. Zhang, Bhatt and Mandyam [3] proposed a FFT-Circulant approximation algorithm that provides a good tradeoff between computational complexity and BER performance. The proposed algorithm simplifies the MMSE-TDE approach by exploiting the Toeplitz autocorrelation matrix of the received signal [4]. Unfortunately, the FFT-Circulant algorithm has the problem of large condition number at a high signal-to-noise ratio (SNR) multipath channel [5]. Besides, other suboptimum solutions such as Conjugate Gradient (CG) [6] and Least Mean Square (LMS) [2] can reduce computation complexity. However, their BER performances are not satisfactory compared with the MMSE method. Another kind of equalizer: the FDE [7] reduces the computation complexity by using an efficient Fast Fourier Transform (FFT) operation. To avoid the influence of the Inter-Block Interference (IBI), the transmitter of the FDE has to insert the Guard Interval (GI) to the signal before transmission. This GI insertion reduces the system capacity and is also not compatible with the frame format of the current WCDMA systems [1].

In this paper, a FDE without GI insertion for the WCDMA systems is proposed. To avoid the IBI, the proposed FDE adopts the Overlap-Cut (OC) technique [7]. Besides, the proposed FDE exploits the decision feedback structure to improve the Bit Error Rate (BER) performance. This paper is organized as follows. In Section II, we will describe the proposed FDE in detail. Section III illustrates the simulation results and the computational comparisons for different FDE methods. Some conclusions of this paper are given in Section IV.

II. Method Description

Consider a W-CDMA downlink channel with J active users, the transmitted signal can be described as

$$s(i) = c_{bs}(i)\sum_{k=1}^{K}\sum_{m=0}^{N_s-1} a_k b_k(m) c_k(i - SFm) \qquad (1)$$

where $K=\sum_{j=1}^{J} K_j$ is total number of the spreading code. K_j is the number of j-th user's spreading code. m and k are the symbol and spreading code indices. $c_{bs}(i)$ is the base-station dependent long scrambling code, b_k is the information bit sequences and c_k is the spreading code with length *SF*. N_s is the number of the symbols transmitted during a given time window.

Let $S(n)=[s((n-1)M+1), s((n-1)M+2),....., s(nM)]^T$ be the nth $M\times 1$ transmitted signal block and the channel impulse response be denoted as $h=\{h_0,...,h_L\}$, the nth received signal block $r(n)$ is given by

$$r(n) = H_0 S(n) + H_1 S(n-1) + n(n) \qquad (2)$$

where $n(n)$ is the Additive White Gaussian Noise (AWGN) vector, H_0 and H_1 are the $M\times M$ channel matrix defined as

$$H_0=\begin{bmatrix} h_0 & 0 & \cdots & \cdots & 0 \\ \vdots & \ddots & \ddots & & \vdots \\ h_L & \ddots & \ddots & \ddots & 0 \\ 0 & \cdots & h_L & \cdots & h_0 \end{bmatrix} \quad (3)$$

$$H_1=\begin{bmatrix} 0 & \cdots & h_L & \cdots & h_1 \\ \vdots & & & \ddots & \vdots \\ \vdots & & & & h_L \\ & & & & \vdots \\ 0 & \cdots & \cdots & \cdots & 0 \end{bmatrix} \quad (4)$$

Define the matrices C, and C_{ISI}, C_{IBI} as

$$C=\begin{bmatrix} h_0 & 0 & \cdots & 0 & h_L & \cdots & h_1 \\ \vdots & \ddots & \ddots & & \ddots & \ddots & \vdots \\ \vdots & & \ddots & \ddots & & \ddots & h_L \\ h_L & & & \ddots & \ddots & & 0 \\ 0 & \ddots & & & \ddots & \ddots & \vdots \\ \vdots & \ddots & \ddots & & & \ddots & 0 \\ 0 & \cdots & 0 & h_L & \cdots & \cdots & h_0 \end{bmatrix} \quad (5)$$

$$C_{ISI}=H_1 \quad (6)$$

$$C_{IBI}=H_1 \quad (7)$$

We can rewrite the received signal block $r(n)$ as

$$r(n)=CS(n)-C_{ISI}S(n)+C_{IBI}S(n-1)+n(n) \quad (8)$$

The propose FDE with decision feedback is depicted in Fig.1. The FDE uses the OC technique to reduce the IBI interference and adopts the feedback signals to cancel the $C_{ISI}S(n)$ and $C_{IBI}S(n-1)$. It is expected that after these interference cancellations, the received signal is distorted only by a circulant channel matrix C and the AWGN noise.

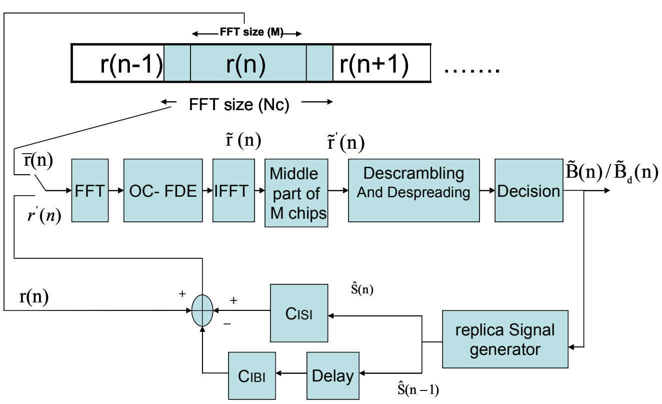

Fig.1 Overlap-Cut FDE with Decision Feedback

The OC method was firstly proposed in [7] and can be briefly illustrated in Fig.2. The length of the FFT operation of the FDE is selected as N_c chips where $N_c=M+2P$, M is the size of the signal block, P denotes the overlapped length with the post- and pre- signal blocks. After the compensation of the FDE, the middle M chips are chosen to avoid the IBI for the signal processing afterward.

Besides, an interesting observation can be deduced from an example shown in Fig.3 and Fig.4. In Fig.3, if we perform the overlapped expansion (in this case, P=2, M=6), the original channel matrix A will expand to the $(M+2P)\times(M+2P)=N_c\times N_c$ matrix **B**. After adding a corner matrix **D** to **B**, i.e., **C=B+D**, the middle M-chips signal $r(n)=[r_1(n),r_2(n)....r_6(n)]^T$ in Fig.3 is the same as the $r(n)'$ in Fig.4. According to OC method and this observation, we can assume that if the overlapped length P is sufficient long, the received signal can be approximated as Eq.(9) after the overlapped expansion.

$$r(n)=CS(n)+C_{IBI}S(n-1)+n(n) \quad (9)$$

To determine a proper set of the OC-FDE weighting W, let the output signal of the OC-FDE can be given by

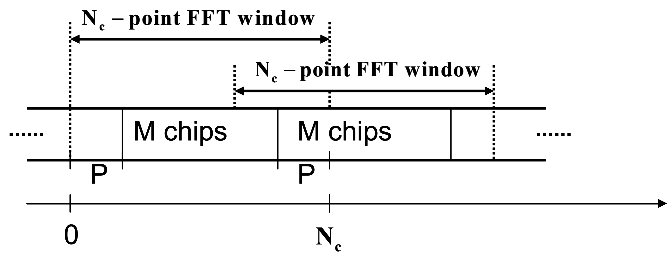

Fig.2 Illustration of the OC Method

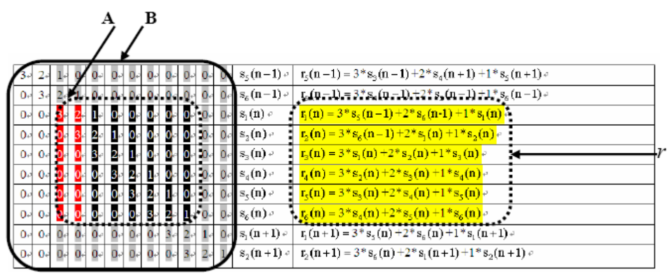

Fig.3 Overlapped Expansion of a Signal Block

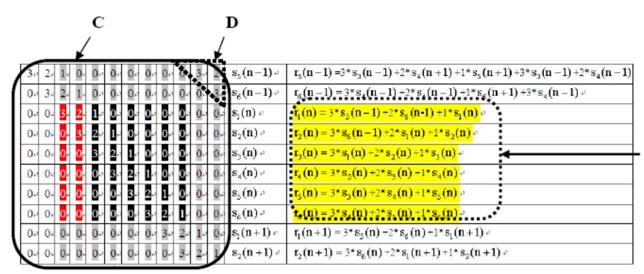

Fig.4 Circulant Approximation of the Overlapped Expansion

$$\hat{S}(n)=D^H W\Lambda DS(n)+D^H WDC_{IBI}S(n-1)+D^H WDn(n) \quad (10)$$

where C= $D^H\Lambda D$, Λ is a diagonal matrix with diagonal $\Lambda=diag(\lambda_0,\lambda_1,....\lambda_{M-1})$, D is DFT matrix. Eq.(10) is derived form the well-known fact that a circulant matrix can be diagonalized by the DFT matrix. Define a cost function Jx to be minimized.

$$\begin{aligned} Jx&=E[tr(\hat{S}(n)-S(n)(\hat{S}^H(n)-S(n)^H))] \\ &=\sigma_s^2 tr(W\Lambda\Lambda^H W^H+WDC_{IBI}C_{IBI}^H D^H W^H-W\Lambda-\Lambda^H W^H+I)+\sigma_n^2 tr(WW^H) \quad (11) \\ &=\sigma_s^2\sum_{m=0}^{N_c-1}(|\lambda_m|^2|w_m|^2+|w_m|^2\sum_{i=0}^{Nc-1}|g_{m,i}|^2-\lambda_m w_m-\lambda_m^* w_m^*)+\sigma_s^2|w_m|^2 \end{aligned}$$

where E[.] and tr[.] denote the ensemble average and trace of the matrix, I is the identity matrix. By solving $\frac{\partial Jx}{\partial w_m^*}=0$, we can obtain the OC-FDE weighting

$$\frac{\partial Jx}{\partial w_m^*} = \sigma_s^2\left(|\lambda_m|^2 w_m + w_m \sum_{i=0}^{Nc-1} |g_{m,i}|^2 - \lambda_m^*\right) + \sigma_n^2 w_m = 0 \quad (12)$$

$$w_m = \frac{\lambda_m^*}{|\lambda_m|^2 + \sum_{i=0}^{Nc-1} |g_{m,i}|^2 + \frac{\sigma_n^2}{\sigma_s^2}} \quad (13)$$

where σ_n^2 is the noise power, σ_s^2 is the signal power, $g_{m,i}(m,i=0,.......,M-1)$ can be calculated from

$$g_{m,m} = \frac{1}{N}\sum_{l=0}^{L-1}\sum_{i=0}^{l} h_{L-i} e^{j\frac{2\pi}{M}n(M-L+l-i)} \quad (14)$$

After the OC-FDE, we choose the middle M chips for the de-scrambling and de-spreading to obtain the M/SF symbols $\tilde{B}(n)=[\tilde{b}_0,\tilde{b}_1 \ldots \tilde{b}_{(M/SF)-1}]^T$. Then, the replica signal $\hat{S}(n)$ of the transmitted chip block and the delay signal $\hat{S}(n-1)$ can be generated to cancel the interferences of the received signal via the feedback. Therefore, the received signal vector $r'(n)$ can be obtained as

$$r'(n) = r(n) + C_{ISI}\hat{S}(n) + C_{IBI}\hat{S}(n-1) \approx CS(n) + n(n) \quad (15)$$

Ideally, the compensated signal $r'(n)$ depends only on the transmitted signal $S(n)$ with a circulant matrix C . Hence, we can exploit a conventional MMSE-FDE or the proposed OC-FDE to obtain the final decided signal $\tilde{B}_d(n)$.

III. BER PERFORMANCE AND COMPLEXITY COMPARISONS

In this section, we compare the BER performance and computational complexity of different FDE equalizers including the SCBT-FDE[8] 、OC-FDE and OC-FDE with Decision Feedback(OC-FDE-FB). We simulated the BER performance under the VA50 channel model (Table-I) defined in the 3GPP specification [9]. All the simulations are with 14 MAI interferences. The chip rate for the transmitted signal is 3.84 Mcps. Orthogonal-Variable-Spreading-Factor (OVSF) codes with spreading factor 16 and Quadrature Phase Shift Keying (QPSK) modulation are used in accordance with the 3GPP HSDPA standard [1].

TABLE-I 3GPP VEHICULAR-A CHANNEL MODEL

ITU vehicular A Speed 50 km/h (VA50)	
Relative Delay [ns]	**Relative Mean Power [dB]**
0	0
310	-1.0
710	-9.0
1090	-10.0
1730	-15.0
2510	-20.0

A. BER Performance

In Fig.5, the BER Performance of the Single Carrier Block Transmission-FDE (SCBT-FDE) [8], OC-FDE and OC-FDE-FB are compared under the VA50 channel. The FDE employs the OC method with overlap length P=16,32 and M=128. Simulation results show that the BER performance of the SCBT-FDE will be saturated at high SNR channel because the IBI can not be perfectly cancelled. The proposed OC-FDE-FB scheme can achieve the best performance among the three methods because it adopts the feedback signals to cancel the IBI caused by the multipath channel.

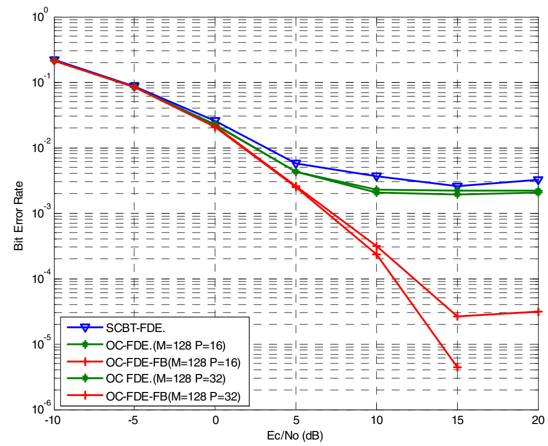

Fig.5 BER Performance under the VA50 channel

B. Complexity Comparison

TABLE-II lists the computational complexity required for different equalizers. The length of channel is $L+1$. The proposed OC-FDE-FB method requires a little increment in the computational loading. However, its BER performance is better than the SCBT-FDE method especially at the high SNR scenarios.

IV. CONCLUSION

In this paper, we proposed an OC-FDE with decision feedback for the HSDPA downlink system. We also derived the calculation of weighting for the OC-FDE. The proposed FDE adopts the OC method to avoid the IBI and exploits the decision feedback structure to improve the BER performance. According to the simulation results, the proposed method has a satisfactory performance. Besides, without inserting the GI, the proposed FDE is more compatible to the frame format of the current High Speed Downlink Packet Access (HSDPA).

TABLE-II 3GPP Vehicular-A Channel Model

Operation	SCBT FDE N_c =M	OC-FDE N_c =M+2P	OC-FDE-FB N_c =M+2P
Equalizer weights	$N_c \log_2(N_c) +$ $3N_c \times \frac{L\times(L+1)}{2}$ $+L\times N_c + 5\times N_c$	$N_c \log_2(N_c) +$ $2Lf \times \frac{L\times(L+1)}{2}$ $+L\times N_c + 3\times N_c$	$N_c \log_2(N_c)$ $+M \log_2(M)$ $+2N_c \times \frac{L\times(L+1)}{2}$ $+L\times N_c + 3\times N_c$
FFT	$N_c \log_2(N_c)$	$N_c \log_2(N_c)$	$N_c \log_2(N_c)$ $+M \log_2(M)$
IFFT	$N_c \log_2(N_c)$	$N_c \log_2(N_c)$	$N_c \log_2(N_c)$ $+M \log_2(M)$
Equalization	N_c	N_c	$N_c + M$
Replica Signal Generator	$\frac{L\times(L+1)}{2}$ $+2\times(N_c(1+K))$	0	$L\times(L+1)$ $+2\times M(1+K)$
Total (N_c =160, M=128,P=32, L=15, K=15)	5.5×10^4	5.4×10^4	6.1×10^4

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