CDMA/TDD 다중코드 전송에서 주파수 도메인 등화기와 결합된 Pre-Rake 와 Cyclic Prefix 최소화 방법

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Introductions of Pre-Rake with Frequency Domain Equalizer and Cyclic Prefix Reduction Method in CDMA/TDD Multi-code Transmission

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요 약

본 논문에서는 SC-CDMA/TDD (Single Carrier CDMA/Time Division Multiplexing) 다중코드 전송에 있어서 송신단에서의 Pre-rake와 수신기에서 주파수 도메인 등화기를 결합한 Pre/Post-FDE (Frequency Domain Equalizer)를 제안하고, 무선 채널 환경 및 사용되는 멀티 코드들의 수에 따라 Post-FDE, Pre-rake, Pre/Post-FDE와 같은 모드들이 선택적으로 구동해야 함을 제안한다. 또한, Pre-rake가 적용되는 시스템에서 Cyclic Prefix의 길이를 최소화 하는 방법이 소개된다. Pre-rake로 전송되는 신호는 수신기에 부가적인 Signal to Noise ratio(SNR)를 제공할 수 있다. 이를 위하여 수신기에서는 시간 도메인에서의 rake combiner를 고려할 수 있다. 그러나, 기존의 Pre-rake만을 고려하는 단말에 비하여 rake combiner와 같은 더 많은 하드웨어를 요구하게 된다. 한편 SC-FDE(CP-CDMA)와 같은 광대역 싱글 케리어 전송 시스템에서 채용되고 있는 주파수 도메인 등화기는 시간 도메인에서의 하드웨어 복잡성을 피하고, 다양한 등화 방법을 고려할 수 있는 특징이 있다. 본 논문에서는 주파수 선택적 특성을 갖는 무선 채널과 사용되는 멀티코드 수에 따라 제안하고자 하는 Pre/Post-FDE와 기존의 Post-FDE와의 성능을 보여 주고자 한다. 제안하는 시스템의 검증을 위하여 멀티 코드 전송에서 가우지안 모델을 이용한 이론적인 분석 및 Pre/Post-FDE와 Post-FDE의 성능을 비교 분석한다.

Key Words : Pre-rake; Post-rake; Diversity; Frequency Domain Equalizer; Cyclic Prefix

ABSTRACT

In this paper we propose a Pre-rake system applied with a frequency domain equalizer in TDD/CDMA multi-code transmission. The Pre-rake system has been well known technique in TDD/CDMA to make a receiver simple. However, it still has residual losses of path diversity and signal to noise ratio (*SNR*). However, gathering all the residual paths demands an additional hardware such as a rake combiner at the receiver. For the reason Pre/Post-rake system has already been proposed at up/downlink correlated channel condition under the assumption of noisier channel. There is a trade-off between the first purpose of Pre-rake that makes hardware simple at the receiver and the performance improvement. From the point the frequency domain equalizer (FDE) can be considered in Pre/Post-rake to supply the receiver with the flexible equalizing methods with rather reduced complexity compared with time domain rake combiner or equalizers. Pre-rake itself increases the number of multipath, which results from the convolution of Pre-rake filter and wireless channel, and FDE must be well matched to Pre/Post-rake, while it considers the relationship of hardware complexity and the performance. In this paper, the Pre-rake/Post-FDE system is introduced at TDD/CDMA multi-code transmission. In addition, the cyclic prefix reduction method in the proposed system is introduced, and the theoretical analysis to the proposed system is given by assuming Gaussian approximation, and finally the numerical simulation results are provided.

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I. Introduction

The capacity increasing techniques for the high data rate downlink transmission are very demanding. The SC (Single Carrier)-CDMA has been widely researched with the strength against the frequency selectivity and the effective resource utilization. The SC-CDMA/TDD (Time Division Duplex) was adopted as a wireless access method in IMT-2000, known for WCDMA/TDD. The WCDMA/TDD offers an extension to the WCDMA/FDD when the available pair frequency bands are not symmetrical or when higher data rates are required at the cost of a reduced mobility. Its advantages are to achieve high capacity, high performance, and low cost system design. Among advantages especially, the share of same center frequency and the ease of slot allocation of up/downlink largely have the attraction to the CDMA/TDD Wireless Internet Services [1]. In the meantime wideband SC-CDMA applied with FDE (frequency domain equalizer) has been already adopted in the fixed wireless broadband IEEE 802.16b as a standard [2]. Unlikely of the FDD where the limited channel information such as DOA (Direction of Arrival) and channel delay from the uplink can be utilized to downlink transmission (Frequency Division Duplex), the TDD can use the wireless multipath channel characteristics obtainable by sharing the center frequency in up/downlink. When the channel reciprocity of up/downlink is satisfied, it can result in low complexity of hardware and signal processing at MS. The idea was firstly proposed by M. Nakagawa, named as the Pre-rake [3], and one of its applications can also be found in [4]. Even though it can design MS simple, it still has diversity loss, SNR loss, and the performance degradation due to interferences in a frequency selective channel. Additionally, the channel reciprocity is not satisfied, it malfunctions as other pre-coding techniques do [5][6]. Under the correlated channels of up/downlink the pre-coded transmit signals with the Pre-rake filter are still imperfectly combined at the receiver side, and increases the number of multipath at the receiver. The increased multipath results from the linear convolution of Pre-rake filter and the wireless channel. Meantime to increase SNR gain the Pre/Post-rake system was proposed to combine the residual multipath by adding the time domain post-rake to MS [7][8]. And also, the principal eigenvector was also applied as an optimum combining weight where the noisier channel was assumed in [9]. However, it still can not reduce the interferences due to frequency selective channel and also requires many fingers to combine the residual multipath signals. In other words, the countermeasure for those is required when the Pre-rake is applied in the transmitter. Balancing the hardware complexity and performance improvement at MS leads to the combination with frequency domain equalizer (FDE). The FDE is firstly proposed by [10], and it has been actively researched on SC-CDMA with FDE, called SC-FDE or CP-CDMA for the hardware complexity and performance in [11][12][13][14]. To perform FDE the transmitter inserts CP every data block where the length of CP and data block depends on the maximum channel delay and fading time selectivity, respectively. By inserting CP the inter-block interference (IBI) is avoided and the wireless channel matrix looks like a circular, so that the channel compensation is simply performed by the complex multiplication in a frequency domain [14]. On the other hand, the CP length should be carefully considered in the proposed system since the Pre-rake filter looks like a wireless channel. From (1) and (2), the CP length directly related to the power and frequency efficiency losses. This will be visited in the next section. This paper is organized as follows. In section 2 the proposed system model and CP reduction method are introduced, and the probability error of the proposed system is evaluated using the Monte Carlo integration with the assumption of Gaussian approximation. The comparison of the simulation results will be presented in section 3. Finally, the concluding remarks are drawn in section 4.

$$data\ loss = (1 - \frac{data\ length}{data\ length + CP\ length}) \times 100[\%] \quad (1)$$

$$power\ loss = 10\log 10(1 + \frac{CP\ length}{data\ length})[dB] \quad (2)$$

II. Pre/Post-FDE System

Figure 1 shows the characteristics of the channel impulse response at the receiver when transmitted with the Pre-rake under correlated and uncorrelated channels of up/downlink.

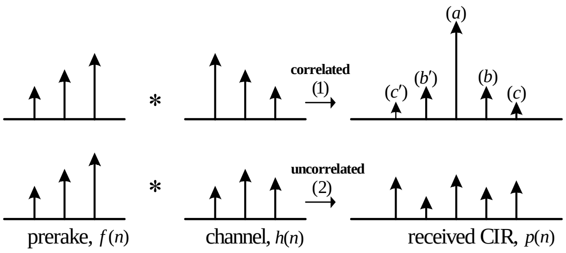

Figure 1. Channel impulse response (CIR) at MS

In case that the channels of up/downlink are correlated, the demodulation at the conventional Pre-rake starts at (a) with simple matched filter for the despreading. However, the residual multipath of (c'), (b'), (b), and (c) still can be used to maximize the SNR gain and obtain imperfect diversity gain at the receiver so that the Pre-rake can be supported by the Post-rake at MS in time domain [7][8]. In a severe noisier channel under multi-user with high spreading factor, the Pre/Post-rake has shown the best performance because the maximal ratio combining (MRC) at both BS and MS must be an optimal solution. However, the real channel could not always be like that. In Fig. 2 it shows the proposed system block diagram.

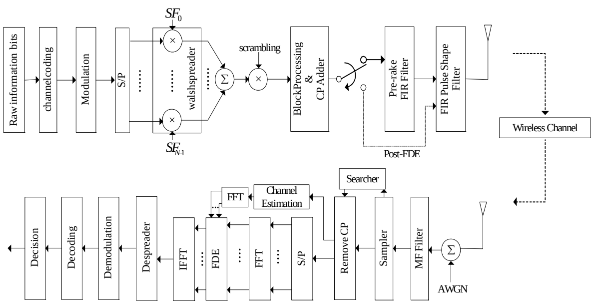

Figure 2. Proposed system block diagram

The system works in a dual switching mode. One is for the Post-FDE where the Pre-rake at the transmitter does not operate and the other is for the Post-FDE with the Pre-rake. The switching parameters are the status of the wireless channel and the number of multi-code used. The spreaded and scrambled data sequences with modulated symbols go through a block processor that divides data sequences into blocks of which the length is determined by the duration of fading channel variation, and CP is added to the head of each data block to eliminate IBI, and its length is the same as the maximum channel delay.

The wireless channel impulse response is expressed by (3).

$$h_k(n) = \sum_{l=0}^{L-1}\beta_{k,l}\delta(n-lT_c) \tag{3}$$

where $\beta_{k,l}$ denotes the Rayleigh distributed channel complex coefficient of the lth uniform distributed multipath for k^{th} user. The baseband model of the transmitted signal can be written as (4).

$$s_k(n) = \sqrt{P}\sum_{m=0}^{M-1}\frac{1}{\sqrt{U_k}}\sum_{l=0}^{L-1}\beta^{*}{}_{k,L-l-1}b_{k,m}(n-lT_c)a_m(n-lT_c) \tag{4}$$

where P is the transmit power, here β is the Pre-rake filter coefficients, b is BPSK modulated symbol, and $a$is spreading chip (assumed with BSPK scrambling). The sub-fix of k, m, and l denotes user, multicode, and multipath, respectively. And U_k is transmit power normalization factor in (5).

$$U_1 = \sum_{l=0}^{L-1}\left|\beta_{1,l}\right|^2 \tag{5}$$

The transmit symbols spreaded with each walsh code are multiplexed and scrambled with cell specific code. The CP inserted data blocks are transmitted through Rayleigh fading channel and added by AWGN at the receiver. After removing CP, the received signal is given by (6).

$$\begin{aligned} r_1(n) &= \mathrm{Re}\left\{\sum_{j=0}^{L-1}\beta_{1,j}s_1(n-jT_c)+n(n)\right\} \\ &= \sqrt{P}\sum_{m=0}^{M-1}\sqrt{\frac{1}{U_1}}\sum_{j=0}^{L-1}\sum_{l=0}^{L-1}[\beta_{1,j}\beta^{*}{}_{1,L-1-l}b_{1,m}(n-jT_c-lT_c) \\ &\quad \cdot a_m(n-jT_c-lT_c)]+n(n) \end{aligned} \tag{6}$$

where n(n) is the zero mean AWGN with two sided power spectral density $N_0/2$.

In Fig. 1 the linear convolution of the Pre-rake filter and wireless channel, $p(n)$ can be given by (7).

$$p(n) = \sum_{j=0}^{L-1}\sum_{l=0}^{L-1}\beta_{1,j}\beta^{*}_{1,L-1-l}\delta(n-(j+l)T_c) \tag{7}$$

In the other hand, the received signal is the output of circular convolution of the transmit signal without Pre-rake filter and (7). The circular convolution is simply converted to the frequency domain as a multiplication form in (8).

$$R_1(k) = S(k)\cdot P(k) + N(k) \tag{8}$$

where NC denotes FFT number.

$$R_1(k) = \sum_{n=0}^{N_C-1} r_1(n)\exp\left(-j\frac{2\pi}{N}nk\right)$$

Likewise, $s(k)$, $p(k)$, and $N(k)$ are the FFT of $s(n)$, $p(n)$, and $n(n)$.

To compensate channel distortion the weight vector is obtained in MRC and MMSEC as follows.

MRC : $$W(k) = \hat{P}^*(k) \tag{9}$$

MMSEC : $$W = \arg\min_w E[|WR - D|^2]$$

$$\begin{aligned}\nabla \mathbf{J}(W) &= \nabla(W^H E[RR^H]W) - \nabla(W^H E[RD^H]) - \nabla(E[DR^H]W) \\ &\quad + \nabla(E[DD^H]) \\ &= \nabla(W^H E[RR^H]W) - \nabla(W^H E[RD^H]) \\ &= 2E[RR^H]W - 2E[RD^H] = 0\end{aligned}$$

where ∇ denotes a gradient, R is the received signal vector and D is the reference pilot vector. Using definition below, the weight per sub-carrier is obtained in (10).

$$\nabla(\mathbf{w}^H \mathbf{A}\mathbf{w}) = 2\mathbf{A}\mathbf{w}$$
$$\nabla(\mathbf{w}^H \mathbf{c}) = 2\mathbf{c}$$
$$\nabla(\mathbf{c}^H \mathbf{w}) = 0$$

$$W(k) = \frac{\sigma_d^2 \hat{P}(k)}{\sigma_n^2 + \sigma_d^2 |\hat{P}(k)|^2} = \frac{\left(\frac{\sigma_d}{\sigma_n}\right)^2 \hat{P}^*(k)}{1 + \left(\frac{\sigma_d}{\sigma_n}\right)^2 |\hat{P}(k)|^2} \tag{10}$$

where σ_d^2 and σ_n^2 are the variance of signal and noise, respectively. In MMSEC the required information is SINR (Signal to Interference Noise Ratio), and its optimal and sub-optimal methods can be found in [16].

After FDE, the time domain signal with IFFT can be given by (11).

$$\hat{r}(n) = IFFT(P(k)S(k)W(k)) + IFFT(N(k)W(k)) \tag{11}$$

Now to derive SINR, (11) is despreaded and the decision statistics per symbol is obtained by (12).

$$\begin{aligned}\hat{z}_{i,m} &= \frac{1}{SF}\sum_{j=0}^{SF-1} c_m(j) \cdot \hat{r}(j + i \cdot SF) \\ &= D + SCI + MCI + \eta\end{aligned} \tag{12}$$

where SF is the spreading factor. D is the desired signal, SCI is the self-code interference, MCI is multi-code interference, and η is the AWGN thermal noise term.

For the analysis of the proposed system, MRC is considered here. Firstly, the desired signal, D is given by (13).

$$D(k) = b_1^0 \cdot SF \sqrt{\frac{P}{U_1}\left[\sum_{n=0}^{2L-2} |\xi_1(n)|^2\right]} \tag{13}$$

where $\xi_i(k) = \sum_{\substack{l,j \\ l-j=k-L+1}} \beta_{i,j}\beta_{1,l}$.

In addition, $\sum_{n=0}^{2L-2}\sum_{\substack{j,l \\ l-j=k-L+1}}$ denotes the summation of all the $(l,\ j)$ combinations, which satisfies $k-L+1$, but $0 \le l,j \le (L-1)$ where L denotes the number of multipath.

Secondly, the self-code interference, SCI is given by (14).

$$\begin{aligned}SCI &= \sqrt{\frac{P}{U_1}}\sum_{n=0}^{2L-3}\sum_{j=n+1}^{2L-2}\xi(n)\xi^*(j) \\ &\quad \{b_1^0 C_1(SF - j + n) + b_1^0 C_1(SF - j + n) + 2b_1^0 C_1(j - n)\} \\ &= \sqrt{\frac{P}{U_1}}\sum_{n=0}^{2L-3}\sum_{j=n+1}^{2L-2}\xi(n)\xi^*(j)\{2SF - (j - n)\}\end{aligned} \tag{14}$$

Thirdly, the multi-code interference, MCI is given by

$$MCI = \sqrt{\frac{P}{U_1}}\sum_{m=0}^{M-1}\sum_{n=0}^{2L-2}\sum_{j=0}^{2L-2}\xi_1(n)\xi^*(j)\left\{\sum_{t=0}^{SF-1} b_m(t-(j-n))a_m(t-(j-n))a_1(t)\right\} \tag{15}$$

where M denotes multi-code. In the other hand, (15) can be divided into two terms. One is when $n = j$, and the other is when $n \ne j$.

$$MCI|_{n=j} = W\sqrt{\frac{P}{U_1}}\sum_{m=0}^{M-1}\sum_{j=0}^{2L-2}\xi_1^2(j) b_m^0 C_1(0) \tag{16}$$

$$MCI|_{n \ne j} = \sqrt{\frac{P}{U_1}}\sum_{m=1}^{M-1}\sum_{n=0}^{(2L-3)}\sum_{j=n+1}^{(2L-2)}\xi_1(n){\xi^*}_1(j)\begin{Bmatrix} b_m^0 C_{m,1}(n-j) + \\ b_m^0 C_{m,1}(N+n-j) + \\ b_m^0 (j-n-N) + \\ b_m^0 C_{m,1}(j-n) \end{Bmatrix} \tag{17}$$

where W='0' for orthogonal code, W='1' for random code.

The discrete aperiodic cross correlation function $C_{i,j}(\bullet)$ is considered as random code interference as in [17] because the transmitted signal is scrambled by cell specific PN code. In (18) their expected values are given.

$$\begin{aligned} &E\{C_i^2(n)\} = N - |n| \qquad m \ne 0 \\ &E\{C_{i,j}^2(n)\} = N - |n| \\ &E\{C_{i,j}(n)C_{i,j}(m)\} = 0 \qquad m \ne n \ and \ i \ne j \end{aligned} \tag{18}$$

Finally, the signal to interference ratio (SIR) can be found by (13), (14) and (15). The variance of (13), (14) and (15) can be given by (19), (20), and (21), respectively.

$$E\{D^2\} = \frac{P \cdot SF^2}{U_1}\chi^2 \tag{19}$$

$$E\{SCI^2\} = \frac{P}{U_1}\{2N\mu - \upsilon\} \tag{20}$$

$$\begin{aligned} E\{MCI^2\} &= E\left\{\left(MCI|_{n=j}\right)^2 + \left(MCI|_{n\neq j}\right)^2\right\} \\ &= \frac{2P}{U_1}SF(M-1)\mu \end{aligned} \tag{21}$$

where orthogonal codes are assumed so that the first term of $E\{MCI^2\}$ in (21) is equal to '0', and χ, μ, and υ are expressed as below.

$$\chi = \left[\sum_{n=0}^{2L-2}|\xi_1(n)|^2\right]$$

$$\mu = \sum_{n=0}^{2L-3}\sum_{j=n+1}^{2L-1}\left|\xi(n)\xi^*(j)\right|^2$$

$$\upsilon = \sum_{n=0}^{2L-3}\sum_{j=n+1}^{2L-1}\left|\xi(n)\xi^*(j)\right|^2(j-n)$$

The noise variance of (12) is given in (22)

$$\frac{N_0 \cdot SF}{2}\chi \tag{22}$$

After all, the $SINR$ can be expressed by (23) and the error probability is obtained with (24).

$$SINR = \frac{D^2}{SCI^2 + MCI^2 + \eta} \tag{23}$$

$$P_e = Q(\sqrt{SINR}) = \frac{1}{2}erfc\left(\sqrt{\frac{SINR}{2}}\right) \tag{24}$$

When (23) is substituted by (19), (20), (21) and (22)

$$\begin{aligned} Y &= \left[\frac{2}{SINR}\right]^{-1} \\ &= \frac{\dfrac{4SF \cdot P}{U_1}\mu - \dfrac{2P\upsilon}{U_1} + \dfrac{4SF \cdot P(M-1)\cdot\mu}{U_1} + N_0 \cdot SF \cdot \chi}{\dfrac{P \cdot SF^2 \cdot \chi^2}{U_1}} \\ &= \left[\frac{4\mu}{SF\chi^2} + \frac{2\upsilon}{SF^2\chi^2} + \frac{4(M-1)\mu}{SF\chi^2} + \frac{LU_1}{\gamma_b\chi}\right] \end{aligned} \tag{25}$$

where $\overline{\gamma_b} = \dfrac{SF \bullet PL}{N_0}$, $\overline{\gamma_b}$ is the average signal to noise ratio.

The bit error probability P_e is obtained by substituting (25) for (24). To evaluate the theoretical analysis, the Monte-Carlo integration method was used in [18]. And also the analysis was compared with the simulation result in section 3.

In designing Pre/Post-FDE the maximum channel delay for the required CP length should include the Pre-rake filter tap delay. However, the CP insertion is directly related to the loss in (1), and (2). In the following the CP reduction method is introduced which results in the required CP length equal to that in Post-FDE.

Ⅲ. Cyclic Prefix Control

In Pre/Post-FDE system the cyclic prefix (CP) insertion method can be considered in two ways. Briefly speaking, CP reduction method is to make a linear convolution to a circular convolution before transmission. Without an additional hardware of FFT and IFFT, the CP reduction can be implemented.

1. CP length with $N_{tap} + N_L - 1$ (Without Control)

As shown in Fig. 4, there are CP control block before and after Pre-rake filter. The Pre-CP adder inserts N_{tap} chips, equal to N-tap Pre-rake filter length, and after Pre-rake filtering it is removed.

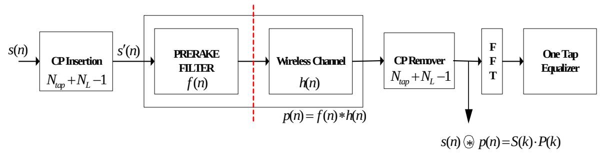

Figure 3. CP insertion without control

We assume that the number of Pre-rake filter taps is assumed to be equal number of the channel multipath delay (N_L). This is the total channel delay to be the linear convolution output delay of Pre-rake filter and wireless channel. Accordingly, the maximum delay of the transmitted signal is equal to $2N_L - 1$. It does not require any CP control, but it sacrifices almost two times of loss comparing with Post-FDE. The $*$, and $\circledast$ denotes linear and circular convolution, respectively in Fig. 3.

2. CP length with N_L (With Control)

The Post-CP adder inserts N_L chips at the end of every block to avoid IBI due to the wireless channel delay. To construct a circular channel matrix at the receiver the CP part to be removed is the N_L chips from every block start even if N_L chips are inserted at the end of every block. This is because the additional operation is not required for the dispreading in MS. To help more understanding, the received circular channel matrix is composed as in (26) and (27) under the Tab. 1.

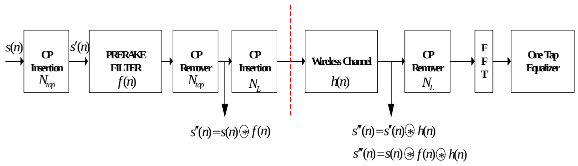

Figure 4. CP insertion with control

Table 1. Assumed system parameter

Spreading factor	8
Symbols per block	Slow fading duration
Max. multipath delay (N_L)	2 chips
Prerake filter tap number (N_{tap})	2
Prerake filter tap distance	T_C

From Tab. 1 the number of received multipath is 3 (Ntap+NL−1=3).Without considering CP reduction method, the minimum required CP length is 3 chips, and its circular channel matrix is shown in (26). Since the first column vector is used for the frequency domain equalization, it is stressed with a dotted block. In (27) the circular channel matrix through CP reduction method is shown.

$$\begin{bmatrix} h_0h_1^* & & & & h_1h_0^* & h_1h_1^*+h_0h_0^* \\ h_1h_1^*+h_0h_0^* & h_0h_1^* & & & & h_1h_0^* \\ h_1h_0^* & h_1h_1^*+h_0h_0^* & h_0h_1^* & & & \\ & h_1h_0^* & h_1h_1^*+h_0h_0^* & h_0h_1^* & & \\ & & \ddots & \ddots & \ddots & \\ & & & h_1h_0^* & h_1h_1^*+h_0h_0^* & h_0h_1^* \end{bmatrix} \tag{26}$$

$$\begin{bmatrix} h_1h_1^*+h_0h_0^* & h_0h_1^* & & & & h_1h_0^* \\ h_1h_0^* & h_1h_1^*+h_0h_0^* & & & & \\ & h_1h_0^* & & & & \\ & & \ddots & & & \\ & & & h_1h_0^* & h_1h_1^*+h_0h_0^* & h_0h_1^* \\ h_0h_1^* & & & & h_1h_0^* & h_1h_1^*+h_0h_0^* \end{bmatrix} \tag{27}$$

where h_0, and h_1 are channel impulse response per multipath and h_0^*, and h_1^* are Pre-rake filter tap coefficients. All other elements at the matrix are zeros. The manipulation of the CP reduction can reduce the power and frequency efficiency losses up to those of Post-FDE. The CP reduction method, introduced in this paper can be replaced by the Pre-rake in frequency domain with FFT and IFFT [19] because the multiplication in frequency domain equalization means the circular convolution of equalizer coefficients and the input transmit signal in time domain. However, the time domain operation without needing FFT and IFFT can be implemented. The additional benefit from it is that the decision on the sampling position at the receiver becomes easier because the Pre-raked pilot signal with the maximum peak is firstly received.

Ⅳ. Simulation Results

In this section the numerical results will show the performance comparison of Pre-rake, Post-FDE, and the proposed Pre/Post-FDE under the correlated channel condition of up/downlink. Before proceeding, the characteristics of each system need to be classified. First of all, the performances of Pre-rake, Post-rake, and Post-FDE(MRC) are the same in multi-code transmission because SINR is the same in all cases if CP loss is ignored in Post-FDE(MRC). The performance comparison of the Post-rake in time domain and Post-FDE(MRC) are well described in [20]. Likewise, the performances of Pre/Post- rake in time domain and the Pre/Post-FDE (MRC) in frequency domain are the same if CP loss is ignored. After all, showing the performance comparison of Pre/Post-FDE and Post-FDE is to see the performances of all mentioned systems. As frequency domain equalization methods, MRC and MMSEC are applied here. The simulation parameters are summarized in Tab. 2.

Table 2. Simulation parameters

Symbols per block	8 symbols
Spreading factor	64 (Orthogonal Code)
FFT Size	512
multipath distribution	- Exponential delay profile with 3 [dB] decaying model. - Uniform delay profile.
Cyclic prefix	32 chips
Modulation	BPSK
Channel estimation	Perfect assumed.
Channel Coding	Uncoded

First of all, the commencement of the Pre/Post-FDE research comes from the result in Fig. 5.

The Pre-rake detects the maximum peak for the demodulation. On the other hand Pre/Post-FDE (MRC) combines the entire residual multipath so that the SNR and diversity gains can be achieved more. As Fig. 5 shows, the obtainable gain is about 2 [dB] at 4 multipath in single code transmission. Although there is no IBI by inserting CP in Pre/Post-FDE (MRC), it does not show any gain over the Pre/Post-rake in time domain. This is because MRC does not compensate the destruction of the orthogonality due to channel delay.

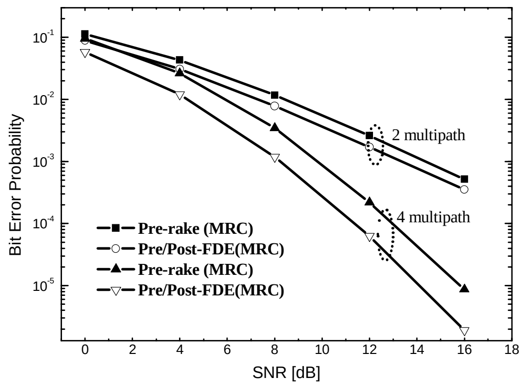

Figure 5. Pre-rake, Pre/Post-rake at single code

The Fig. 6 shows the performance comparison of Pre/Post-rake and Pre/Post-FDE(MRC). As it was mentioned, there is a difference due to CP insertion loss in (2) where the simulation was performed with CP length of 32 chips. In this the power loss is about 0.28 [dB]. For the verification of the analysis in (24) the simulation result in Fig. 7 is shown.

For the performance comparison the number of multi-code is 8, 30 and 50 in the uniformly distributed Rayleigh multipath channels. The dotted line is for the analysis and the solid line is for the simulation in Pre/Post-FDE(MRC). In the next the BER performances of Pre/Post-FDE(MRC) with CP reduction and without CP reduction are shown in Fig. 8. As was explained in section 2, both should show the same performance. Without CP reduction in section 2, however, the power and frequency efficiency loss are doubled. Since the simulation was performed without including a power loss by CP insertion, the same results are shown. In Referring to [7][8], the performance of Pre/Post-rake is better than other two systems of Pre-rake, and Post-rake. This is because MRC and the principal eigenvector is the optimum solution for a known signal and for a Rayleigh faded random signal in AWGN, respectively [21].

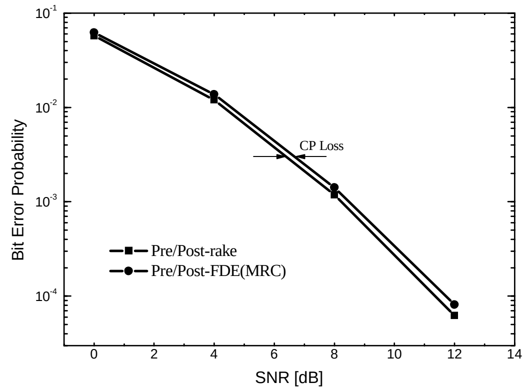

Figure 6. Pre/Post and Pre/Post-FDE (MRC)

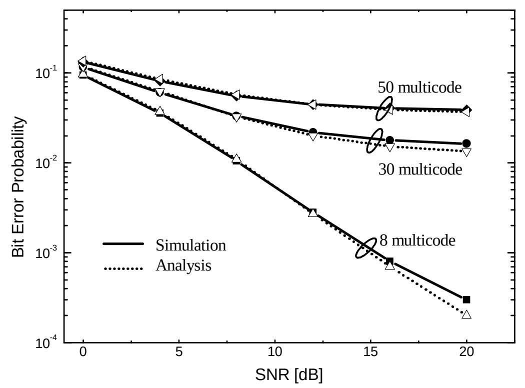

Figure 7. Comparison of Simulation and Analysis

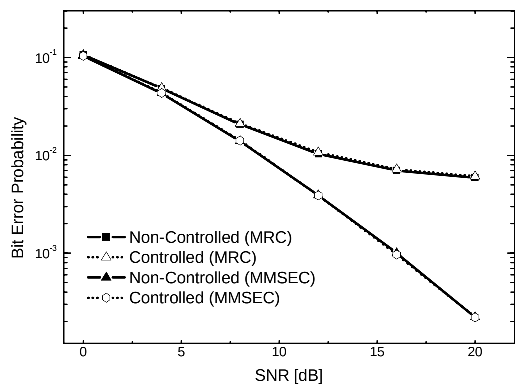

Figure 8. Performance of CP with/without reduction

Namely, the assumed channel was a severely noisier channel with the high spreading gain. However, there is a big difference between Pre/Post-FDE (MRC) and Pre/Post-FDE (MMSEC) as is shown in Fig. 9. The difference of the performance can be explained by the effect of MCI (Multiple

Code Interference).

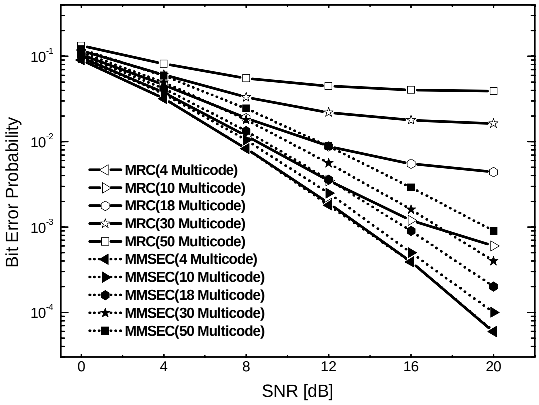

Figure 9. BER Comparison of Pre/Post-FDE in MRC and MMSEC

And also the result can be explained by the characteristics of MMSEC that works as MRC in low SNR(noisier channel), and works as ZF(Zero Forcing) in high SNR. As a result, Pre/Post-FDE(MMSEC) shows the better performance than Pre/Post-FDE(MRC). As seen in (10), the MMSEC requires SNR or SINR estimation per sub-carrier after FFT. In [16] its optimal and sub-optimal methods for MMSEC can be found.

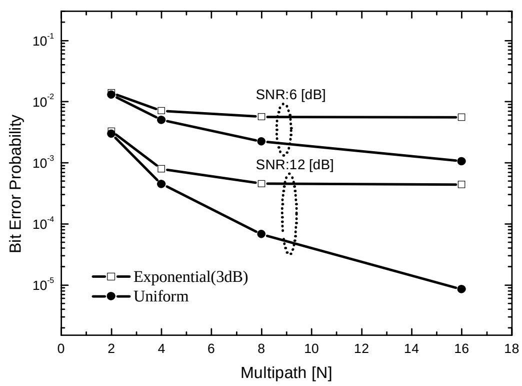

Figure 10. Effects of two different delay profiles at 8 multi-code

In Fig. 10 the performances of Post-FDE in the exponential and uniform channel delay profile are shown. The number of multi-code used is 8 and the result shows the degree of time diversity gain in a different channel delay profile. The exponential delay profile with 3 [dB] decaying factor has been considered. As is shown, the exponential delay profile has less diversity gain compared with the uniform delay profile. In addition, the obtainable gain in 6 [dB] of SNR is smaller than that in 12 [dB] of SNR. It can be explained that the multipath signal is below the noise level. This result will be referred to the following simulation results, which will be shown. In Fig. 11 the BER performances of Pre/Post-FDE and Post-FDE are compared in uniform delay profile with 2 multipath. The simulation is evaluated in 12 [dB] of SNR. The performance can be evaluated from the relationship of SNR gain and MCI. First of all, the results can be summarized in the following.

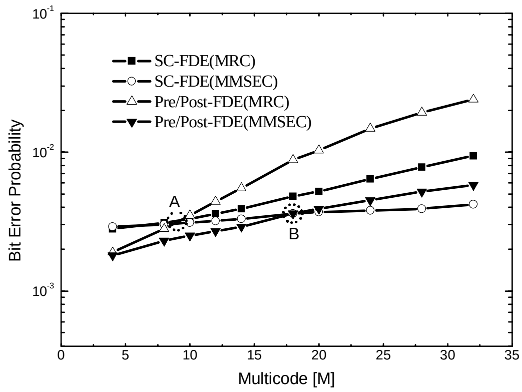

Figure 11. Comparison of Post-FDE and Pre/Post-FDE

The performance of Pre/Post-FDE(MRC) is getting worse than others over 8 multi-code. On the other hand the performance of Pre/Post-FDE(MMSEC) shows the best performance up to 18 multi-code. (about 28 [%], multi-code use efficiency). To be more in detail, from the cross point of 'A' where the 8 multicodes are used, the performance of Post-FDE(MRC) is getting worse than that of Post-FDE(MMSEC). From the fact that Post-FDE(MRC) and Post-FDE (MMSEC) show the same performance in low load system, it can be seen that MCI is getting stronger from the cross point of 'A' even for Post-FDE(MRC). In the meantime, the performances of Pre/Post-FDE(MRC) and Pre/Post-FDE(MMSEC) have been already separated before reaching to the cross point of 'A'. This results from the fact that the Pre-rake in the transmitter increases the multipath at the receiver. However, the transmitter with the Pre-rake still shows the better performance than Post-FDE. It can be seen that the obtainable diversity and SNR gain in Pre/Post-FDE system achieves the better system performance than that of Post-FDE even if the effect of MCI is relatively large in the Pre/Post-FDE. However, the performances of Pre/Post-FDE(MMSEC) and Post-

FDE(MMSEC) branches off from the cross point of 'B' because the obtainable SNR gain could not be greater than MCI(Multi-codeInterference) at Pre/Post-rake (MMSEC). Both systems with MMSEC have a capability of reducing MCI, so the overall performances are better than other two systems with MRC. In the next the BER performances in the exponential and uniform delay profile with 8 multipath are shown in Fig. 12 and Fig. 13, respectively.

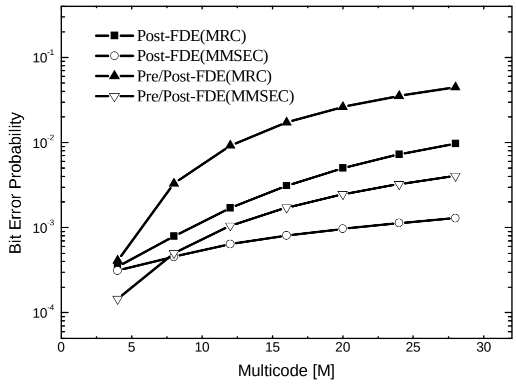

Figure 12. BER performance in exponential delay profile with
8 multipath

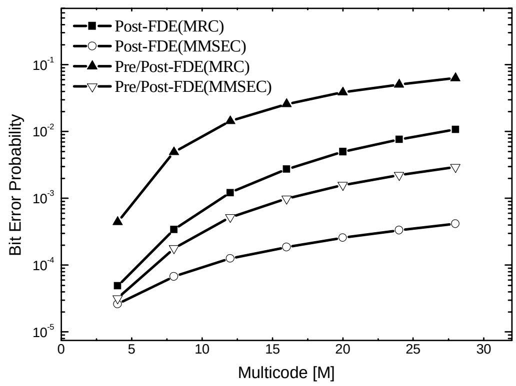

Figure 13. BER performance in uniform delay profile with 8 multipath

The overall performances are better than the case of 2 multipath, but the active range of the Pre/Post-FDE are reduced up to 8 multi-code (about 12.5 [%]) in the exponential channel delay profile, while up to 4 multi-code in the uniform delay profile. From Fig. 10 the diversity gain can be achieved more in the uniform delay profile. It also means that the increase of MCI due to the Pre-rake filter is greater than that in the exponential delay profile. Referring the channel models in [22][23], the real channel model is close to the exponentially distributed channel model. Therefore, the Pre/Post-FDE mode up to 8 multi-code in this scenario still can be effectively switched with Post-FDE mode.

The Fig. 14 shows the BER performances in uniform delay profile with 16 multipath. The tendency of the performance is almost same as in Fig. 13, but the comprehensive performances are improved in Pre/Post-FDE and Post-FDE with MMSEC. On the other hand the two systems with MRC do show very little gain due to the MCI. After all, the Post-FDE in a severe multipath channel outperforms the Pre/Post-FDE due to the increased interferences.

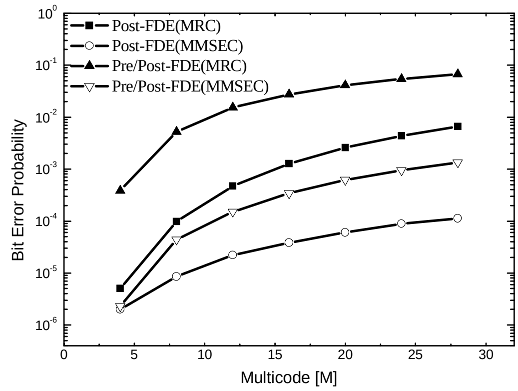

Figure 14. BER performance in uniform delay profile with 16 multipath

V. Conclusions

In this paper we have introduced the Pre-rake applied with Post-FDE (Pre/Post-FDE) in CDMA/TDD multi-code transmission. For the verification of the proposed system its BER performance was expressed for the signal to interference and noise ratio (SINR) with the Gaussian approximation. For the fair comparison, the numerical performances were also compared with the Post-FDE. Moreover, through the CP reduction method without using FFT and IFFT the power and frequency efficiency losses could be reduced as in Post-FDE and makes it easy to design a dual mode system with the Pre-rake and the Post-FDE. The concept of the rake combiner at both BS and MS is to maximize the SNR, which is optimal in a noisier channel environment, but the interferences such as ISI and MCI can not be prevented. Since the transmitter with the Pre-rake increases the number of multipath at the receiver, the time domain operation is not desirable. Accordingly, the improvement of SNR/imperfect diversity gains and the countermeasure for the interference suppression in the frequency selective fading are required for the receiver to combine a frequency domain equalizer. From the results the Pre/Post-FDE can achieve SNR and imperfect diversity gains at low loaded transmission while the Post-FDE outperforms the Pre/Post-FDE in high loaded transmission. To sum it all up, the proposed system with dual switching mode can be an alternative solution in CDMA/TDD multi-code transmission over the single mode such as the Pre-rake only or Post-FDE only.

References

[1] R. Esmailzadeh, M. Nakagawa, and A.Jones, "TDD-CDMA for the 4th generation of wireless communication," *IEEE Wireless. Commun.*, vol. 10, no. 4, pp. 8-15, Aug. 2003.

[2] Air Interface for Fixed Broadband Wireless Access Systems Part A: Systems Between 2 and 11 GHz, *IEEE802.16ab-01/01*, June. 2001.

[3] R. Esmailzadeh, and M. Nakagawa, "Pre-rake Diversity Combination for Direct Sequence Spread Spectrum Mobile Communications Systems," *IEICETrans.OnCommun.*, pp. 1008-1015, Aug. 1993.

[4] Incheol Jeong, Jaesang Cha, and Mungeon Kyeong, "A Pre-rake CDMA System with Modified ZCD Spreading Code". International Workshop on Telecommunications 2009, Sao Paulo, Brazil, pp. 156-160, Feb. 2009.

[5] B. R. Vojcic and W. M. Jang, "Transmitter precoding in synchronous multi-user communications," *IEEETrans.onCommun.*, pages 1347-1355, Oct. 1998.

[6] M. Meuerer, P. W. Baier, T. Weber, Y. Lu and A. Papathanasiou, "Joint transmission: advantageous downlink concept for CDMA mobile radio systems using time division duplexing," *IEE Electronics Letters*, 36:900-901, May. 2000.

[7] J. B. Wang, M. Zhao, S. D. Zhou, and Y. Yao, "A Novel Multipath Transmission Diversity Scheme in TDD-CDMA Systems," *IEICE Trans. Commun.*, vol. E82-B,Oct.1999.

[8] A. N. Barreto and G. Fettweis, "Performance Improvement in DS-SPREAD Spectrum CDMA Systems Using A PRE-AND A POST-RAKE," *Broad Commun.*, pp. 39-46, Feb. 2000.

[9] R. Irmer, A. N.Barreto, and G. Fettweis, "Transmitter Precoding For Spread-Spectrum Signals In Frequency-Selective Fading Channels," *Proc.3G Wireless 2001*, pp. 939-944 ,May. 2001.

[10] T. Walzman, and M. Schwartz, "Automatic equalization using the discrete Fourier domain," *IEEE Trans. Info. Theory*, vol. IT-19, pp. 59-68, Jan. 1973.

[11] K. L. Baum, and T. A. Tomas F. W.Vook, V. Nangia, "Cyclic Prefix CDMA : An Improved Transmission Method for Broadband DS-CDMA Cellular Systems," *IEEE WCNC 2002*, vol. 1, pp 183-188, Mar. 2002.

[12] I. Martoyo, T. Weiss, F. Capar, and F. K. Jondral, "Low Complexity CDMA Downlink Receiver Based on Frequency Domain Equalization," *IEEE VTC 2003*, vol. 2, pp. 987-991, Oct. 2003.

[13] M. V. Clark, "Adaptive Frequency-Domain Equalization and Diversity Combining for Broadband Wireless Communications," *IEEE Journal on Selected Areas in Commun.*, vol. 16, pp. 1385-1395, Oct. 1998.

[14] F. Adachi, D. Garg, S. Takaoka, K, Takeda, "Broadband CDMA techniques," *IEEE Wireless Commun.*, vol. 12, pp. 8-18, April. 2005.

[15] R. D.Strum, and D. E. Kirk, Discrete Systems And *Digital Signal Processing, Addison Wesley*, 1998.

[16] S. Kaiser, "On the Performance of Different Detection Techniques for OFDM-CDMA in fading Channels," *Global Telecommun.*, vol. 3, pp. 2059-2063, Nov. 1995.

[17] M. Pursley, "Performance Evaluation for Phase-Coded Spread-Spectrum Multiple Access Communications-part.I : system analysis," *IEEE Trans. Commun.*, vol. Com-25, No.8, pp. 795-799, Aug. 1977.

[18] W. H. Press, B. P. Flannery, S. A. Teukolsky, and W. T. Veterling, Numerical Recipes the Art of Scientific Computing. New York : Cambridge Univ. Press, 1986.

[19] R. L. Choi, and R. D. Murch, "Frequency Domain Pre-equalization with Transmit Diversity for MISO Broadband Wireless Communications,"

IEEE VTC 2002, vol. 3, pp. 1787-1791, Sept. 2002.

[20] F. Adachi, and T. Itagaki, "Frequency-Domain Rake Combining for Antenna Diversity Reception of DS-CDMA Signals," *IEICE Trans. Commun.*, vol. E86-B, pp. 2781-2784, Sep. 2003.

[21] S. Haykin, Adaptive filter Theory, Prentice Hall, Englewood Cliffs, NJ, 1986.

[22] L. M. Correia, *Wireless Flexible Personalised Commun.*, John Wiley and Sons Ltd. 2001.

[23] "Guidelines for evaluation of radio transmission technologies for IMT-2000," Recommendation ITU-R M.1225, 1997.

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