PAPER

DS-CDMA HARQ with Overlap FDE

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SUMMARY Turbo coded hybrid ARQ (HARQ) is known as one of the promising error control techniques for high speed wireless packet access. However, in a severe frequency-selective fading channel, the HARQ throughput performance significantly degrades for direct sequence code division multiple access (DS-CDMA) system using rake combining. This problem can be overcome by replacing the rake combining by the frequency-domain equalization (FDE) based on minimum mean square error (MMSE) criterion. In a system with the conventional FDE, the guard interval (GI) is inserted to avoid the inter-block interference (IBI). The insertion of GI reduces the throughput. Recently, overlap FDE that requires no GI insertion was proposed. In this paper, we apply overlap FDE to HARQ and derive the MMSE-FDE weight for packet combining. Then, we evaluate the throughput performance of DS-CDMA HARQ with overlap FDE. We show that overlap FDE provides better throughput performance than both the rake combining and conventional FDE regardless of the degree of the channel frequency-selectivity.

key words: *DS-CDMA, HARQ, overlap FDE*

1. Introduction

For the next generation mobile communication systems, high-speed and high-quality packet data services are demanded. Turbo coded hybrid ARQ (HARQ) is known as one of the promising error control techniques to realize high speed wireless packet access [1]. In the 3rd generation mobile communication systems, direct sequence code division multiple access (DS-CDMA) with rake combining is adopted to provide packet data services of around a few Mbps [2]. Since the wireless channel is composed of many propagation paths having different time delays, a frequency-selective fading channel is produced [3]. For data rate transmission higher than a few Mbps, the channel frequency-selectivity gets severe and the throughput performance of DS-CDMA HARQ using rake combining severely degrades.

It was shown in [4], [5] that frequency-domain equalization (FDE) based on minimum mean square error (MMSE) criterion can replace rake combining while offering much improved DS-CDMA transmission performance. Since MMSE-FDE is a block processing, MMSE-FDE needs the insertion of guard interval (GI) to avoid the inter-block interference (IBI). However, the GI insertion reduces the throughput. Furthermore, if the time delays of the propagation paths exceed the GI length, the throughput of DS-CDMA HARQ with MMSE-FDE significantly degrades due to the IBI. Recently, overlap FDE that needs no GI insertion was proposed [6], [7]. By using overlap FDE, the impact of IBI can be sufficiently suppressed even without GI. In this paper, we apply overlap FDE to HARQ and derive the MMSE-FDE weight for packet combining. Then, we show that overlap FDE provides better throughput performance than the rake combining and conventional FDE irrespective of the degree of the channel frequency-selectivity.

The remainder of this paper is organized as follows. Section 2 presents the overall transmission system model of DS-CDMA HARQ using overlap FDE. The MMSE-FDE weight for packet combining is derived. In Sect. 3, the throughput of DS-CDMA HARQ using overlap FDE in a frequency-selective Rayleigh fading channel is evaluated by computer simulation. The simulated throughput performance is compared with those using the rake combining and conventional FDE. Section 4 concludes the paper.

2. Packet Combining and Overlap FDE for DS-CDMA HARQ

Figure 1 shows the overall transmission system model of DS-CDMA HARQ using overlap FDE. Turbo encoded HARQ with Chase combining (CC) [8], [9] is considered. The original coding rate is $R = 1/3$. In this paper, we assume a single-user packet transmission using full multicode transmission (i.e., all the orthogonal spreading codes are assigned to a single-user for high-speed packet access).

2.1 Signal Representations

At the transmitter, an information bit sequence is turbo encoded and bit-interleaved before transforming into the data-modulated symbol sequence. The data-modulated symbol sequence is serial-to-parallel (S/P) converted into U parallel streams $\{d_u(i);\ i = \ldots, -1, 0, 1, \ldots\}$, $u = 0 \sim U-1$. Then, each stream is spread by using an orthogonal spreading code with spreading factor SF $\{c_u(t);\ t = 0 \sim SF-1\}$, $u = 0 \sim U-1$. After multiplexing U chip sequences, the multicode chip sequence is multiplied by a scramble code $\{c_{scr}(t);\ t = \ldots, -1, 0, 1, \ldots\}$ to obtain

$$s(t) = \sqrt{\frac{2E_c}{T_c}} \sum_{u=0}^{U-1} d_u\left(\lfloor t/SF \rfloor\right) c_u(t \bmod SF) c_{scr}(t), \quad (1)$$

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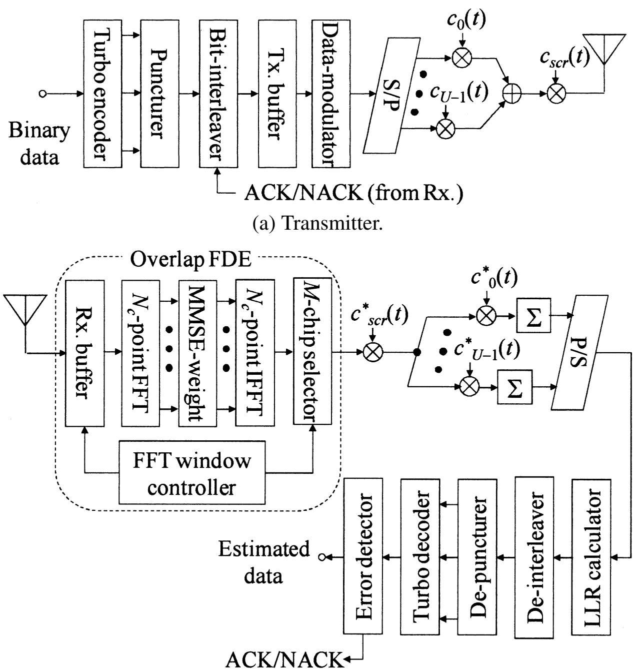


(a) Transmitter.
(b) Receiver.

Fig. 1 DS-CDMA HARQ with overlap FDE.

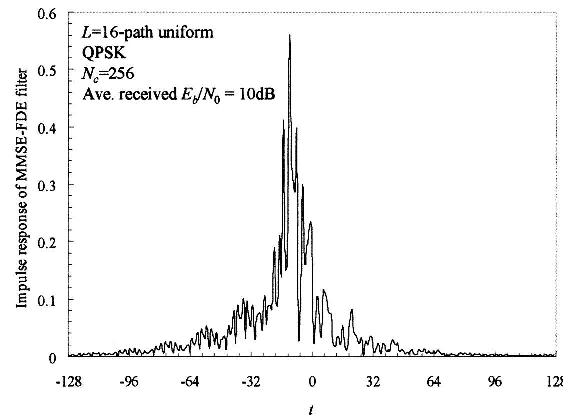


Fig. 2 Impulse response of MMSE-FDE filter.

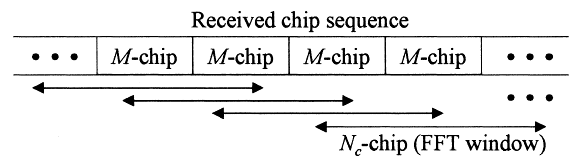


Fig. 3 Overlap FDE.

where E_c and T_c respectively denote the chip energy per parallel stream and the chip duration and $\lfloor x \rfloor$ represents the largest integer smaller than or equal to x.

The transmitted packet is received via a frequency-selective fading channel. We assume a chip-spaced L-path frequency-selective block Rayleigh fading channel. Its impulse response at the reception of the trth retransmitted packet ($tr \leq Q-1$) is expressed as

$$h^{(tr)}(\tau) = \sum_{l=0}^{L-1} h_l^{(tr)} \delta(\tau - \tau_l), \tag{2}$$

where $h_l^{(tr)}$ and τ_l denote the complex valued path gain and delay time of the lth path, respectively, and $\sum_{l=0}^{L-1} E[|h_l^{(tr)}|^2] = 1$.

The received packet for the trth retransmission is expressed as

$$r^{(tr)}(t) = \sum_{l=0}^{L-1} h_l^{(tr)} s(t - \tau_l) + \eta^{(tr)}(t), \tag{3}$$

where $s(t)$ is the transmitted packet and $\eta^{(tr)}(t)$ is the noise due to the additive white Gaussian noise (AWGN) having the one-sided power spectrum density N_0.

2.2 Overlap FDE

For conventional FDE, the multicode chip sequence is divided into a sequence of N_c-chip blocks. The last N_g-chip portion of each N_c-chip block is copied and inserted as a cyclic prefix into the guard interval (GI) placed at the beginning of the block. Due to the GI insertion, the received chip block becomes a circular convolution of the channel impulse response and the transmitted N_c-chip block and therefore, the inter-block interference (IBI) can be avoided [4], [5]. However, when the GI is not used, the IBI is present at the beginning of the received N_c-chip block. MMSE-FDE is a linear circular convolution filter; the residual IBI after MMSE-FDE is a circular convolution of the IBI and the impulse response of MMSE-FDE filter. As seen from Fig. 2, the MMSE-FDE filter impulse response concentrates at a vicinity of $t = 0$. Therefore, the residual IBI after MMSE-FDE is localized only near the both ends of N_c-chip FFT block. The overlap FDE that requires no GI insertion is based on this observation. The received packet is divided into a sequence of M-chip blocks ($M < N_c$). Then, N_c-point FFT is applied to an N_c-chip interval centering an M-chip block of interest. After MMSE-FDE, M-chip block is picked up from the equalized N_c-chip block to reduce the residual IBI. The FFT intervals for consecutive M-chip blocks are overlapped as shown Fig. 3. As M reduces, the residual IBI can be better suppressed, however, the number of FFT/inverse FFT (IFFT) operations increases N_c/M times. Therefore, M should be as large as possible in order not to increase the computational complexity excessively while sufficiently suppressing the IBI.

2.3 Frequency-Domain Packet Combining

We consider the reception of chip block in a time interval of $t = 0 \sim N_c - 1$. The received signal is given by Eq. (3). The desired signal component must be expressed as a circular convolution of the channel impulse response and the transmitted N_c-chip block. Since the GI is not used, the IBI component appears unlike the conventional FDE with GI insertion. We rewrite Eq. (3) as

$$r^{(tr)}(t) = \sum_{l=0}^{L-1} h_l^{(tr)} s((t-\tau_l) \bmod N_c) + \nu^{(tr)}(t) + \eta^{(tr)}(t), \quad \text{for } t = 0 \sim N_c - 1, \tag{4}$$

where the first term represents the desired signal component and $\nu^{(tr)}(t)$ is the IBI component which can be expressed as [7]

$$\nu^{(tr)}(t) = \sum_{l=0}^{L-1} h_l^{(tr)} \{s(t-\tau_l) - s((t-\tau_l) \bmod N_c)\} \times \{u(t) - u(t-\tau_l)\}, \tag{5}$$

with $u(t) = 0$ (1) for $t < 0$ $(0 \le t)$. First, the received N_c-chip block $\{r^{(tr)}(t);\ t = 0 \sim N_c - 1\}$ is transformed into the frequency-domain signal $\{R^{(tr)}(k);\ k = 0 \sim N_c - 1\}$. $R^{(tr)}(k)$ is given by

$$\begin{aligned} R^{(tr)}(k) &= \sum_{t=0}^{N_c-1} r^{(tr)}(t) \exp\left(-j2\pi k \frac{t}{N_c}\right) \\ &= H^{(tr)}(k)S(k) + N^{(tr)}(k) + \Pi^{(tr)}(k), \end{aligned} \tag{6}$$

where $S(k)$ is the kth frequency-component of $s(t)$, and $H^{(tr)}(k)$, $N^{(tr)}(k)$, and $\Pi^{(tr)}(k)$ are respectively the channel gain, the IBI component, and the noise component at the kth frequency and are given as

$$\begin{cases} S(k) = \sum_{t=0}^{N_c-1} s(t) \exp\left(-j2\pi k \frac{t}{N_c}\right) \\ H^{(tr)}(k) = \sum_{l=0}^{L-1} h_l^{(tr)} \exp\left(-j2\pi k \frac{\tau_l}{N_c}\right) \\ N^{(tr)}(k) = \sum_{t=0}^{N_c-1} \nu^{(tr)}(t) \exp\left(-j2\pi k \frac{t}{N_c}\right) \\ \Pi^{(tr)}(k) = \sum_{t=0}^{N_c-1} \eta^{(tr)}(t) \exp\left(-j2\pi k \frac{t}{N_c}\right) \end{cases} \tag{7}$$

with

$$E[|\Pi^{(tr)}(k)|^2] = N_c \frac{2N_0}{T_c}. \tag{8}$$

Below we assume that the same data packet has been retransmitted Q times (including the original packet). In CC, these Q received packets are combined based on the MMSE criterion as

$$\begin{aligned} \hat{R}(k) &= \sum_{tr=0}^{Q-1} w^{(tr)}(k) R^{(tr)}(k) \\ &= \hat{H}(k)S(k) + \hat{N}(k) + \hat{\Pi}(k), \end{aligned} \tag{9}$$

where

$$\begin{cases} \hat{H}(k) = \sum_{tr=0}^{Q-1} w^{(tr)}(k) H^{(tr)}(k) \\ \hat{N}(k) = \sum_{tr=0}^{Q-1} w^{(tr)}(k) N^{(tr)}(k) \\ \hat{\Pi}(k) = \sum_{tr=0}^{Q-1} w^{(tr)}(k) \Pi^{(tr)}(k) \end{cases}. \tag{10}$$

$\hat{H}(k)$ is called the equivalent channel gain after packet combining. $w^{(tr)}(k)$ is the MMSE weight, which will be derived in Sect. 2.4, taking into account the IBI as well as the noise.

After packet combining, N_c-point IFFT is applied to obtain the time-domain chip sequence $\{\hat{r}(t);\ t = 0 \sim N_c - 1\}$. $\hat{r}(t)$ is given as

$$\begin{aligned} \hat{r}(t) &= \frac{1}{N_c} \sum_{k=0}^{N_c-1} \hat{R}(k) \exp\left(j2\pi k \frac{t}{N_c}\right) \\ &= \left(\frac{1}{N_c} \sum_{k=0}^{N_c-1} \hat{H}(k)\right) s(t) + \mu(t) + \hat{\nu}(t) + \hat{\eta}(t), \end{aligned} \tag{11}$$

where the first, second, third and fourth terms represent the desired signal, the residual inter-chip interference (ICI), the IBI and the noise, respectively. They are given as

$$\begin{cases} \mu(t) = \frac{1}{N_c} \sum_{k=0}^{N_c-1} \hat{H}(k) \left[\sum_{\substack{\tau=0 \\ \neq t}}^{N_c-1} s(\tau) \exp\left(j2\pi k \frac{t-\tau}{N_c}\right) \right] \\ \hat{\nu}(t) = \frac{1}{N_c} \sum_{k=0}^{N_c-1} \hat{N}(k) \exp\left(j2\pi k \frac{t}{N_c}\right) \\ \hat{\eta}(t) = \frac{1}{N_c} \sum_{k=0}^{N_c-1} \hat{\Pi}(k) \exp\left(j2\pi k \frac{t}{N_c}\right) \end{cases}. \tag{12}$$

The IBI after FDE exists only near both the ends of N_c-chip block $\{\hat{r}(t);\ t = 0 \sim N_c - 1\}$ [7]. To suppress the IBI, we pick up only the M-chip block $\{\tilde{r}(t);\ t = N_c/2 - M/2 \sim N_c/2 + M/2 - 1\}$ from the equalized N_c-chip block.

The above packet combining and equalization operation are repeated for obtaining a sequence of equalized M-chip blocks. Finally, despreading is applied to obtain a soft decision symbol for $d_u(i)$ as

$$\hat{d}_u(i) = \frac{1}{SF} \sum_{t=iSF}^{(i+1)SF-1} \tilde{r}(t) c_{scr}^*(t) c_u^*(t \bmod SF). \tag{13}$$

2.4 MMSE Weight for Packet Combining

We represent the equalization error $e(k)$ of the kth frequency component using a vector notation as

$$e(k) = \boldsymbol{w}^H(k)\boldsymbol{R}(k) - S(k). \tag{14}$$

where $(.)^H$ denotes the Hermitian transpose, $\boldsymbol{w}(k) = [\{w^{(0)}(k)\}^*, \cdots, \{w^{(Q-1)}(k)\}^*]^T$ and $\boldsymbol{R}(k) = [R^{(0)}(k), \cdots, R^{(Q-1)}(k)]^T$. The MMSE weight vector $\boldsymbol{w}(k)$ can be obtained by solving

$$\frac{\partial}{\partial w^{(tr)}(k)} E\left[|e(k)|^2\right] = 0 \ \text{ for } \ tr = 0 \sim Q-1. \tag{15}$$

According to the Wiener theory [10], we have

$$\boldsymbol{w}(k) = \boldsymbol{\Psi}^{-1}(k)\boldsymbol{p}(k). \tag{16}$$

where

$$\begin{cases} \mathbf{\Psi}(k) = E[\mathbf{R}(k)\mathbf{R}^H(k)] \\ \quad = N_c \dfrac{U}{SF}\dfrac{E_s}{N_0}\dfrac{2N_0}{T_c}\mathbf{H}(k)\mathbf{H}^H(k) + \mathbf{A} \\ \mathbf{p}(k) = E[\mathbf{R}(k)S^*(k)] \\ \quad = N_c \dfrac{U}{SF}\dfrac{E_s}{N_0}\dfrac{2N_0}{T_c}\mathbf{H}(k) \end{cases}, \tag{17}$$

where E_s denotes the average received symbol energy. In the above equation, $\mathbf{H}(k) = [H^{(0)}(k), \cdots, H^{(Q-1)}(k)]^T$ is the channel gain vector and $\mathbf{A}$ is given by

$$\mathbf{A} = \begin{bmatrix} 2\sigma_0^2 & & \mathbf{0} \\ & \ddots & \\ \mathbf{0} & & 2\sigma_{Q-1}^2 \end{bmatrix}, \tag{18}$$

where

$$2\sigma_{tr}^2 = E\left[\left|N^{(tr)}(k)\right|^2\right] + N_c\frac{2N_0}{T_c} \tag{19}$$

is the variance of IBI plus noise.

We find $2\sigma_{tr}^2$ of Eq. (19). From Eq. (7), we have

$$\begin{aligned} &E\left[\left|N^{(tr)}(k)\right|^2\right] \\ &= \sum_{t=0}^{N_c-1}\sum_{t'=0}^{N_c-1} E[\nu^{(tr)}(t)\{\nu^{(tr)}(t')\}^*]\exp\left(-j2\pi k\frac{t-t'}{N_c}\right) \\ &= 2\frac{U}{SF}\frac{E_s}{N_0}\frac{2N_0}{T_c}\sum_{t=0}^{N_c-1}\sum_{t'=0}^{N_c-1}\sum_{l=0}^{L-1}\sum_{l'=0}^{L-1} h_l^{(tr)}\{h_{l'}^{(tr)}\}^* \\ &\qquad \delta(t-t'+l-l') \\ &\times\{u(t)-u(t-\tau_l)\}\{u(t')-u(t'-\tau_{l'})\} \\ &\qquad \exp\left(-j2\pi k\frac{t-t'}{N_c}\right) \end{aligned} \tag{20}$$

for the given $\{h_l^{(tr)};\ l = 0 \sim L-1\}$, where we have used the following equation:

$$\begin{aligned} &E[\{s(t-\tau_l) - s((t-\tau_l) \bmod N_c)\} \\ &\quad \times\{s^*(t'-\tau_{l'}) - s^*((t'-\tau_{l'}) \bmod N_c)\}] \\ &\qquad = 2\frac{U}{SF}\frac{E_s}{N_0}\frac{2N_0}{T_c}\delta(t-t'+l-l'). \end{aligned} \tag{21}$$

It is quite difficult, if not impossible, to find a closed-form expression for $E[|N^{(tr)}(k)|^2]$. Assuming that the IBI variance is equally likely for all k, we first obtain the block averaged total IBI variance. Using Perseval's equality [11], the total IBI variance is given by

$$\sum_{k=0}^{N_c-1} E\left[\left|N^{(tr)}(k)\right|^2\right] = N_c\sum_{t=0}^{N_c-1} E\left[\left|\nu^{(tr)}(t)\right|^2\right]. \tag{22}$$

Since

$$\begin{aligned} &\sum_{t=0}^{N_c-1} E\left[\left|\nu^{(tr)}(t)\right|^2\right] \\ &= 2\frac{U}{SF}\frac{E_s}{N_0}\frac{2N_0}{T_c}\sum_{l=0}^{L-1}\left|h_l^{(tr)}\right|^2\sum_{t=0}^{N_c-1}\{u(t)-u(t-\tau_l)\} \\ &= 2\frac{U}{SF}\frac{E_s}{N_0}\frac{2N_0}{T_c}\sum_{l=0}^{L-1}\left(\left|h_l^{(tr)}\right|^2\tau_l\right), \end{aligned} \tag{23}$$

we obtain

$$\begin{aligned} \sigma_{tr}^2 &= N_c\frac{N_0}{T_c}\left(2\frac{1}{N_c}\frac{U}{SF}\frac{E_s}{N_0}\sum_{l=0}^{L-1}\left(\left|h_l^{(tr)}\right|^2\tau_l\right)+1\right) \\ &= N_c\frac{N_0}{T_c}\tilde{\sigma}_{tr}^2, \end{aligned} \tag{24}$$

where

$$\tilde{\sigma}_{tr}^2 = 2\frac{1}{N_c}\frac{U}{SF}\frac{E_s}{N_0}\sum_{l=0}^{L-1}\left(\left|h_l^{(tr)}\right|^2\tau_l\right)+1. \tag{25}$$

From Eq. (25) with replacing tr by q and from Eqs. (17)–(19), $\mathbf{\Psi}^{-1}(k)$ can be derived, using the matrix inversion lemma [10] as

$$\begin{aligned} &\mathbf{\Psi}^{-1}(k) = \mathbf{A}^{-1} \\ &\quad - \frac{N_c\dfrac{2N_0}{T_c}}{\displaystyle\sum_{q=0}^{Q-1}\frac{\left|H^{(q)}(k)\right|^2}{\tilde{\sigma}_q^2} + \left(\frac{U}{SF}\frac{E_s}{N_0}\right)^{-1}}\mathbf{A}^{-1}\mathbf{H}(k)\mathbf{H}^H(k)\mathbf{A}^{-1}. \end{aligned} \tag{26}$$

Substitution of Eqs. (17) and (26) into Eq. (16) gives

$$\mathbf{w}(k) = \frac{N_c\dfrac{2N_0}{T_c}}{\displaystyle\sum_{q=0}^{Q-1}\frac{\left|H^{(q)}(k)\right|^2}{\tilde{\sigma}_q^2} + \left(\frac{U}{SF}\frac{E_s}{N_0}\right)^{-1}}\mathbf{A}^{-1}\mathbf{H}(k), \tag{27}$$

from which, the MMSE weight for the trth received packet can be obtained as

$$w^{(tr)}(k) = \frac{\{H^{(tr)}(k)\}^*}{\displaystyle\sum_{q=0}^{Q-1}\left(\frac{\tilde{\sigma}_{tr}^2}{\tilde{\sigma}_q^2}\left|H^{(q)}(k)\right|^2\right) + \left(\frac{U}{SF}\frac{E_s}{N_0}\right)^{-1}\tilde{\sigma}_{tr}^2}. \tag{28}$$

2.5 Log-Likelihood Ratio (LLR)

A sequence of log-likelihood ratio (LLR) $L(b_m)$ is generated using the soft decision symbol sequence for turbo decoding. In this paper, we assume M is set small enough to avoid the IBI, so the IBI is not considered in LLR calculation. ICI is approximated as a complex Gaussian noise and the ICI plus noise is treated as a new zero-mean complex Gaussian variable with variance $2\sigma^2$. The LLR is given by [12]

$$L(b_m) = \frac{1}{2\sigma^2}\left|\hat{d}_u(i) - \sqrt{\frac{2E_c}{T_c}}\left(\frac{1}{N_c}\sum_{k=0}^{N_c-1}\hat{H}(k)\right)d^{\min}_{b_m=0}\right|^2 - \frac{1}{2\sigma^2}\left|\hat{d}_u(i) - \sqrt{\frac{2E_c}{T_c}}\left(\frac{1}{N_c}\sum_{k=0}^{N_c-1}\hat{H}(k)\right)d^{\min}_{b_m=1}\right|^2, \quad (29)$$

where $d^{\min}_{b_m=0,1}$ represents the data symbol having the largest LLR within a set of $\{d_{b_m}\}$, and $2\sigma^2$ is given by

$$2\sigma^2 = \frac{2}{SF}\frac{N_0}{T_c}\frac{1}{N_c}\sum_{k=0}^{N_c-1}\sum_{tr=0}^{Q-1}|w^{(tr)}(k)|^2 + \frac{2}{SF}\frac{N_0}{T_c}\left(\frac{U}{SF}\frac{E_s}{N_0}\right)\left(\frac{1}{N_c}\sum_{k=0}^{N_c-1}|\hat{H}(k)|^2 - \left|\frac{1}{N_c}\sum_{k=0}^{N_c-1}\hat{H}(k)\right|^2\right). \quad (30)$$

2.6 Turbo Decoding and Error Detection [13]

After de-interleaving and depuncturing, turbo decoding is carried out using LLR. In this paper, ideal error detection is assumed. If no error is detected in the received packet after turbo decoding, the ACK signal is transmitted to the transmitter. If error is detected, the NACK signal is transmitted to the transmitter to request the retransmission of the same packet. When the same packet is received, the packet combining is carried out using the updated MMSE-FDE weight as described in Sect. 2.4.

3. Computer Simulation

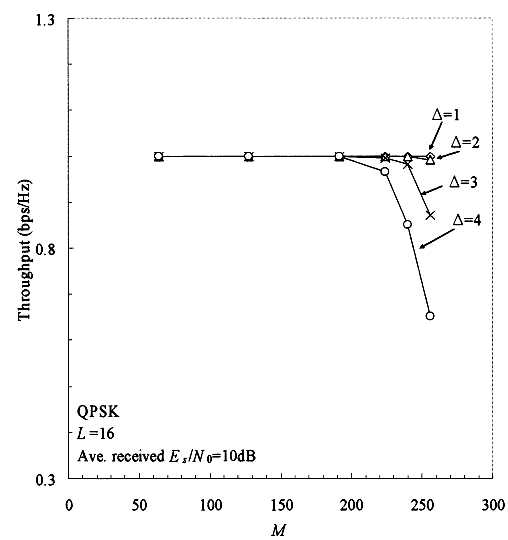
The simulation conditions are summarized in Table 1. We assume QPSK and 16 QAM data-modulations. The single-user transmission with full code-multiplexing is assumed (i.e., U=SF). The maximum number Q_{max} of retransmissions (including the first packet) is set to Q_{max}=100. For turbo coding, we use an original rate R=1/3 turbo encoder having two (13, 15) recursive systematic component encoders, which are concatenated with S-random interleaver followed by puncturer is used [11]. The following puncturing matrix is used to obtain the rate R=1/2 turbo code.

$$\begin{bmatrix} 1 & 1 \\ 1 & 0 \\ 0 & 1 \end{bmatrix}$$

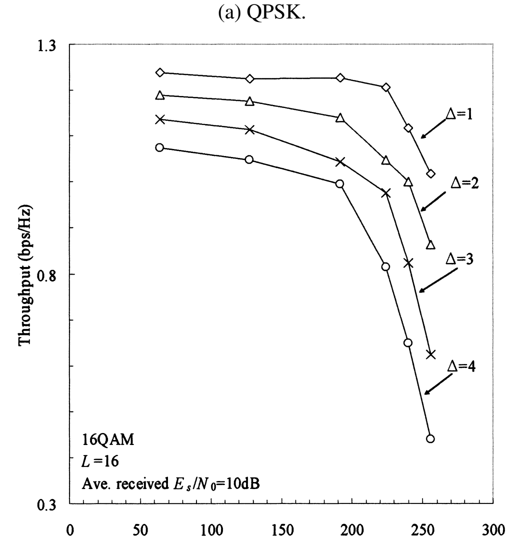
The size M for overlap FDE is an important design parameter. Figure 4 plots the simulated throughput of overlap FDE as a function of M for various values of Δ at the average received symbol energy-to-noise power spectrum density ratio $E_s/N_0(= E_cSF/N_0)$ = 10 dB, where the throughput in bits per second per Hertz (bps/Hz) is defined as the number of bits per data-symbol times the ratio of the number of information bits transmitted successfully to the total number of transmitted bits. With QPSK data modulation, for a

Table 1 Simulation conditions.

Turbo coding	R=1/2 (13, 15) RSC encoder Log-MAP decoding with 8 iterations	
Channel interleaver	S-random interleaver	
Data modulation	QPSK, 16QAM	
DS-CDMA	Spreading codes	Walsh codes
	Spreading factor	SF=16
	Code-multiplexing order	U=SF
ARQ	Chase combining	
	Max no. of retransmissions	Q_{max}=100
Channel model	Frequency-selective block Rayleigh fading	
	No. of paths	L=16
	Power delay profile	Uniform
	Delay time	τ_l=$l\Delta$ with Δ=1, 2, 3, 4
Channel estimation		Ideal
Overlap FDE	FFT window size	N_c=256
	FDE weight	MMSE
	No. of chips to pick up	M=64~256

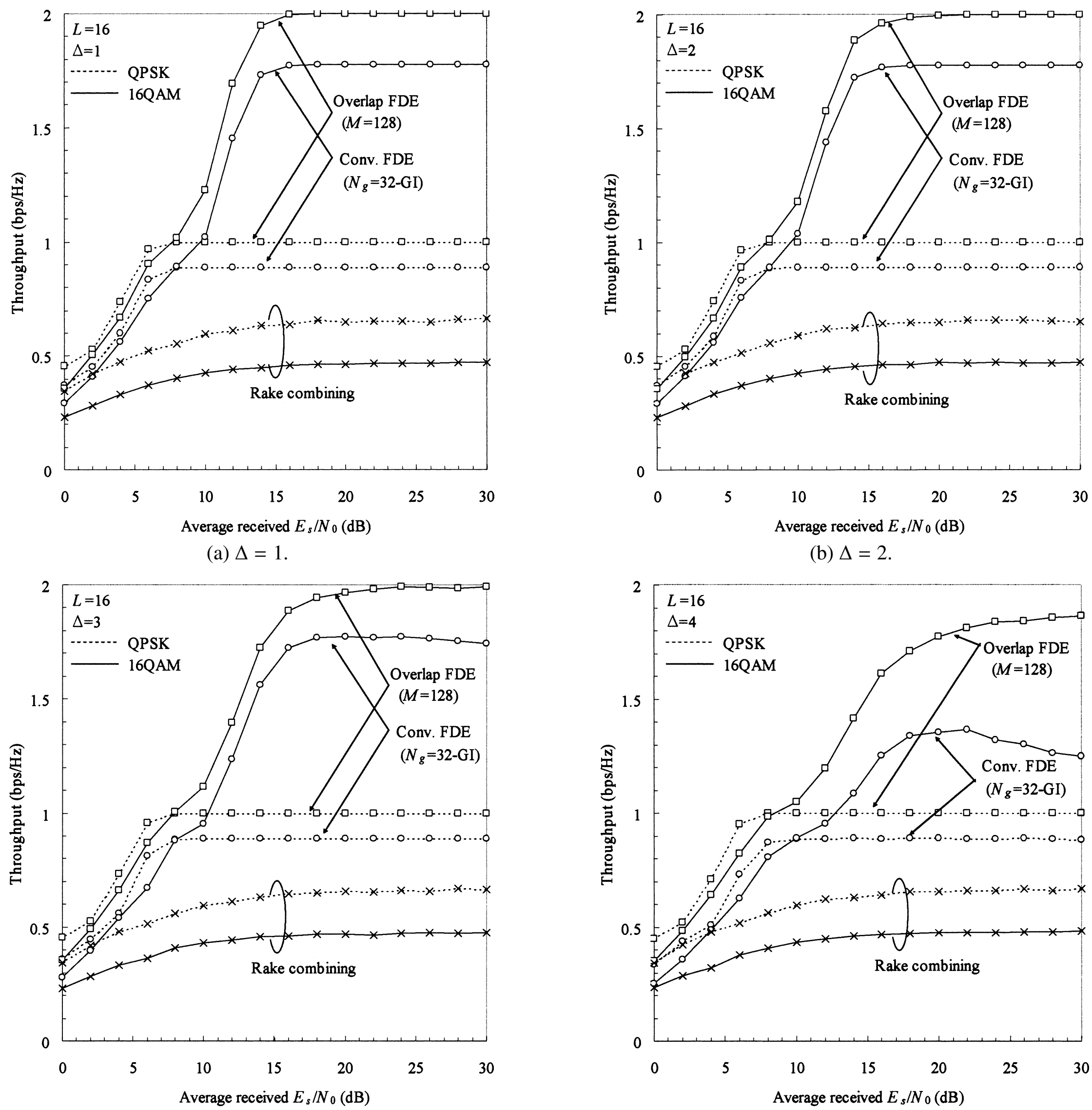


(a) QPSK.



(b) 16 QAM.

Fig. 4 Throughput of overlap FDE as a function of M.



(a) Δ = 1. (b) Δ = 2. (c) Δ = 3. (d) Δ = 4.

Fig. 5 Throughput as a function of E_s/N_0.

weak channel frequency-selectivity case (e.g., $\Delta = 1$), almost the same throughput is obtained regardless of M since the IBI can be sufficiently suppressed by overlap FDE. On the other hand, for a strong channel frequency-selectivity case (e.g., $\Delta = 4$), the throughput is the same if $M \leq 192$ but rapidly degrades due to the increasing residual IBI as M increases beyond 192. When 16 QAM data modulation is used, the throughput is very sensitive to the choice of M since the Euclidean distance between the signal points in the signal space is much shorter for 16QAM than for QPSK and hence, packet error is more likely produced due to the residual IBI. However, by choosing $M \leq 128$, the influence of the residual IBI can be sufficiently suppressed for both QPSK and 16 QAM.

Figure 5 compares the throughput performances of overlap FDE with $M = 128$, the conventional FDE with $N_g = 32$-chip GI, and rake combining. Overlap FDE does not require the GI insertion and therefore, increases the throughput by a factor of $(1+N_g/N_c)$ compared to the conventional FDE with N_g-chip GI insertion. It can be seen from Fig. 5 that overlap FDE can always provide better throughput performance than the conventional FDE irrespective of the degree of the channel frequency-selectivity for both QPSK and 16 QAM. In the case of conventional FDE, when $\Delta > 2$, the time delays of 16 paths exceed the GI length and the residual IBI degrades the throughput. Furthermore, since the IBI variance is not considered in obtaining the MMSE-FDE weight, the throughput using the conventional FDE degrades as E_s/N_0 increases. On the other hand, almost no throughput degradation is seen for the overlap FDE. Also seen from Fig. 5 is that the throughput performance with rake combining is much lower than those with overlap FDE and conventional FDE due to strong inter-path interference (IPI).

In Fig. 5, $M = 128$ is considered for $\Delta = 1 \sim 4$. However, as understood from Fig. 4, larger M can be used as Δ decreases (i.e., as the channel frequency-selectivity becomes weaker). From the view point of computational complexity, the use of larger M is desirable since the number of N_c-point FFT/IFFT operations increases by a factor of N_c/M. When using overlap FDE, almost the same throughput can be achieved irrespective of the degree of the channel frequency-selectivity, simply by changing the value of M. This is a promising advantage of using overlap FDE.

4. Conclusions

In this paper, we applied overlap FDE to DS-CDMA HARQ and derived the MMSE-FDE weight for packet combining. The throughput performance of DS-CDMA HARQ with overlap FDE was evaluated by computer simulation. We have shown that overlap FDE can always achieve higher throughput than the conventional FDE using the GI and also rake combining regardless of the degree of channel frequency-selectivity. Since the GI insertion is not required, overlap FDE can be applied to HARQ used in the present DS-CDMA packet access using rake combining to significantly improve the throughput performance by modifying the receiver structure only.

In this paper, we have assumed ideal channel estimation for computing the MMSE-FDE weight, i.e., the receiver has a perfect knowledge of the channel gain $H(k)$ and E_s/N_0. However, in a practical system, $H(k)$ and E_s/N_0 need to be estimated. The estimation error degrades transmission performance. This is a practically important problem and is left as an important future research.

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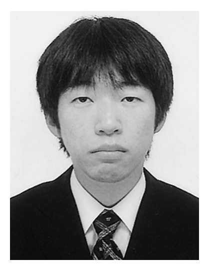

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