Pilot Assisted Channel Estimation for MC-CDMA Signal Transmission Using Overlap FDE

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***Abstract*—Recently, multi-carrier code division multiple access (MC-CDMA) has been attracting much attention as a broadband wireless access technique for the next generation mobile communications systems. Frequency-domain equalization (FDE) based on the minimum mean square error (MMSE) criterion can take advantage of the channel frequency-selectivity and improve the average bit error rate (BER) performance due to frequency-diversity gain. However, conventional MC-CDMA requires the insertion of guard interval (GI) and this reduces the transmission efficiency. Overlap FDE technique has been proposed that requires no GI insertion. Recently, we showed that MC-CDMA using overlap FDE can provide almost the same BER performance as the conventional MC-CDMA downlink using the GI insertion. However, our previous work assumed the ideal channel estimation. In this paper, we propose a pilot assisted channel estimation technique suitable for MC-CDMA downlink using overlap FDE and evaluate the BER performance by computer simulation.**

Keywords-component; Frequency-selective fading channel, overlap FDE, Channel estimation, MC-CDMA.

I. INTRODUCTION

Broadband data services are demanded in the next generation mobile communication systems. However, the broadband channel is composed of many propagation paths having different time delays, thereby resulting in severe frequency-selective channel; the transmission performance degrades due to severe inter-symbol interference (ISI) [1, 2]. Recently, multicarrier-code division multiple access (MC-CDMA), which uses a number of low rate orthogonal subcarriers to reduce the ISI resulting from frequency-selective channel, has been attracting much attention [3-5]. A good bit error rate (BER) performance can be achieved by using frequency-domain equalization based on minimum mean square error criterion (MMSE-FDE) [5].

The conventional MC-CDMA requires the insertion of guard interval (GI) to make the received signal to be a circular convolution of the transmit signal and the channel impulse response. However, the GI insertion reduces the transmission efficiency. Recently, an FDE technique that requires no GI insertion, called overlap FDE, was proposed for the single-carrier transmission [6, 7]. The overlap FDE can also be applied to MC-CDMA. We have shown [8, 9] that overlap FDE can provide almost the same BER performance as that of MC-CDMA using the conventional FDE with GI insertion. However, our previous work assumed the ideal channel estimation. In this paper, we propose a pilot assisted channel estimation technique suitable for MC-CDMA downlink using overlap FDE, and evaluate its BER performance by computer simulation.

The remainder of this paper is organized as follows. Sect. II describes the transmission system model of MC-CDMA downlink using the overlap FDE. The proposed channel estimation technique is present in Sect. III. In Sect. IV, the average BER performance is evaluated by computer simulation. Sect. V offers some conclusions.

II. TRANSMIT SYSTEM MODEL

Figure 1 illustrates the transmitter and receiver structure for MC-CDMA downlink with overlap FDE. Throughout the paper, sample-spaced discrete-time signal representation is used.

A. Received signal representation

At the transmitter, U data symbol sequences $\{d_u(i); i=0\sim(N_c/SF-1)\}$, $u=0\sim(U-1)$, to be transmitted are respectively spread by orthogonal spreading codes $\{c_u(t); t=0\sim(SF-1)\}$, $u=0\sim(U-1)$, to obtain the multi-code chip sequence, where SF denotes the spreading factor. A sum of U chip streams is multiplied by a scrambling sequence $c_{scr}(t)$. To generate the MC-CDMA signal block with N_c subcarriers, N_c-point inverse fast Fourier transform (IFFT) having T_c as the sampling period is applied. In the conventional MC-CDMA transmitter, the guard interval (GI) is inserted to the transmit signal. However, overlap FDE requires no GI insertion, thereby, improving the transmission efficiency.

The MC-CDMA chip stream is transmitted over a frequency-selective fading channel and is received at a receiver. We assume a sample-spaced L-path frequency-selective block fading channel. The complex-valued path gain and time delay of the l-th propagation path are denoted by h_l and τ_l, respectively. The channel impulse response $h(\tau)$ is given by

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

The received MC-CDMA chip stream is divided into a sequence of M ($<N_c$)-chip blocks. For performing FDE on the m-th chip block, we pick up the MC-CDMA chip stream over a time interval of $t=(m-1/2)N_c+M/2\sim(m+1/2)N_c+M/2-1$, which can be expressed as

$$r(t)=\sqrt{\frac{2P}{SF}}\sum_{l=0}^{L-1}h_l s((t-\tau_l)\bmod N_c)+\mu(t)+\eta(t)$$
$$,\ t=(m-1/2)N_c+M/2\sim(m+1/2)N_c+M/2-1, \tag{2}$$

where P is the average received signal power and $s(t)$ is the MC-CDMA chip stream. $\eta(t)$ is the additive white Gaussian noise (AWGN) with zero mean and variance $2N_0/T_c$ with N_0 being the single-sided power spectrum density. $s(t)$ can be expressed, using the equivalent low-pass representation, as

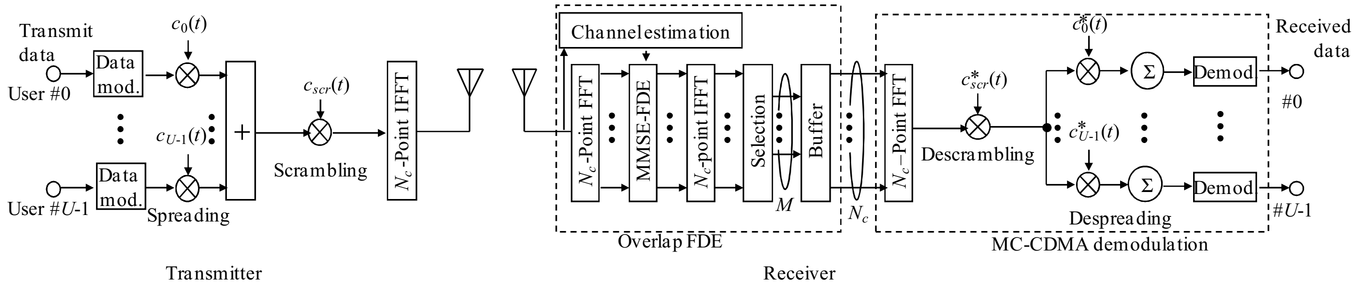


Fig. 1. Transmitter/receiver structure of MC-CDMA downlink using overlap FDE

$$s(t) = \sum_{m=-\infty}^{\infty} s_m(t - mN_c) \tag{3}$$

with

$$s_m(t) = \begin{cases} \sum_{k=0}^{N_c-1} S_m(k)\exp\left(j2\pi k\frac{t}{N_c}\right), t = 0 \sim (N_c - 1) \\ 0 \qquad\qquad\qquad\qquad\qquad , \text{otherwise} \end{cases}, \tag{4}$$

where

$$S_m(k) = \sum_{u=0}^{U-1} c_{scr}(k)c_u(k \bmod SF)d_u\left(\lfloor k/SF \rfloor + m\frac{N_c}{SF}\right), \tag{5}$$

and $\lfloor x \rfloor$ is the largest integer smaller than or equal to x. $\mu(t)$ is the inter-block interference (IBI), which is given by

$$\mu(t) = \sum_{l=0}^{L-1}\begin{Bmatrix} h_l\{s_{m-1}((t-\tau_l) \bmod N_c) \\ - s_m((t-\tau_l) \bmod N_c)\}(u_0(t) - u_0(t-\tau_l)) \end{Bmatrix}, \tag{6}$$

where $u_0(t)$ is the unit step function.

B. *Overlap FDE*

$r(t)$ is decomposed by N_c-point FFT into N_c frequency components $\{R(k); k=0\sim(N_c-1)\}$. $R(k)$ is given by

$$\begin{aligned} R(k) &= \frac{1}{N_c}\sum_{t=(m-1/2)N_c+M/2}^{(m+1/2)N_c+M/2-1} r(t)\exp\left(-j2\pi k\frac{t}{N_c}\right) \\ &= H(k)\widetilde{S}(k) + \mathrm{N}(k) + \Pi(k) \end{aligned}, \tag{7}$$

where

$$\begin{cases} H(k) = \sqrt{\frac{2P}{SF}}\sum_{l=0}^{L-1} h_l \exp\left(-j2\pi k\frac{\tau_l}{N_c}\right) \\ \widetilde{S}(k) = \frac{1}{N_c}\sum_{t=(m-1/2)N_c+M/2}^{(m+1/2)N_c+M/2-1} s(t)\exp\left(-j2\pi k\frac{t}{N_c}\right) \\ \mathrm{N}(k) = \frac{1}{N_c}\sum_{t=(m-1/2)N_c+M/2}^{(m+1/2)N_c+M/2-1} \mu(t)\exp\left(-j2\pi k\frac{t}{N_c}\right) \\ \Pi(k) = \frac{1}{N_c}\sum_{t=(m-1/2)N_c+M/2}^{(m+1/2)N_c+M/2-1} \eta(t)\exp\left(-j2\pi k\frac{t}{N_c}\right) \end{cases}. \tag{8}$$

One-tap FDE is performed on

$$\widetilde{R}(k) = R(k)w(k) \qquad . \tag{9}$$

We use the MMSE-FDE weight that minimizes the mean square error (MSE) ε between $\widetilde{S}(k)$ and $\widetilde{R}(k)$. ε is defined as

$$\varepsilon = E[|\widetilde{R}(k) - \widetilde{S}(k)|^2] \ , \tag{10}$$

where $E[\cdot]$ represents the expectation operation. MMSE-FDE weight $w(k)$ can be obtained by solving $\partial\varepsilon/\partial w(k) = 0$. After some manipulations, we obtain the following MMSE-FDE weight $w(k)$:

$$w(k) = \frac{H^*(k)}{U|H(k)|^2 + 2\sigma^2} \qquad , \tag{11}$$

where σ^2 is the variance of the IBI plus noise power.

The N_c-point IFFT is applied to $\{\widetilde{R}(k); k=0\sim(N_c-1)\}$ to obtain the equalized N_c-chip block $\{\widetilde{r}(t);\ t=(m+1/2)M-N_c/2\sim(m+1/2)M+N_c/2-1\}$. In order to suppress the IBI, we only pick up its center part of M chips, $\{\widetilde{r}(t);\ t=mM\sim((m+1)M-1)\}$, and store it into the buffer.

The resulting sequence of equalized M-chip blocks is the equalized MC-CDMA chip stream. For MC-CDMA demodulation, N_c-point FFT is applied to decompose the m-th MC-CDMA chip block $\{\widetilde{r}(t);\ t=mN_c\sim((m+1)N_c-1)\}$ into N_c subcarrier components $\{\hat{R}_m(k);\ k=0\sim(N_c-1)\}$. After descrambling and despreading, the sequence of decision variable, $\{\hat{d}_u(i);\ i=0\sim(N_c/SF-1)\}$, associated with $\{d_u(i);\ i=0\sim(N_c/SF-1)\}$ is obtained.

If M is too small, the residual IBI can be sufficiently suppressed, but the number of FFT/IFFT operations per MC-CDMA chip block increases [8]. In this paper, we assume $M=N_c/2$.

III. CHANNEL ESTIMATION

FDE requires the accurate channel estimation. We consider the pilot-assisted channel estimation (CE) [10-15]. In Refs. [12-15], a pilot assisted channel estimation, called PACE in this paper, was presented for OFDM using GI insertion. If no GI insertion is used for the pilot block similar to the overlap FDE, the accuracy of channel estimation significantly degrades due to the IBI from the previous data block. In Refs. [16, 17], another channel estimation technique was presented, in which a short pilot sequence is inserted at both ends of each data block. The pilot sequence inserted at the end of the present data block plays as a cyclic prefix for the pilot sequence inserted at the beginning of next data block. Therefore, channel estimation can be carried out without suffering from the IBI. However, the insertion of pilot sequence results in the loss of data rate and power.

In this paper, we propose a new channel estimation technique that requires no GI insertion, in which a short pilot sequence is repeated in the pilot chip block. We call this channel estimation technique as cyclic PACE (CPACE). In the following, we consider a short pilot sequence with a period of $N_c/2$ is repeated twice in the pilot chip block.

A. Principle of CPACE

CPACE uses the time-multiplexed pilot chip block to be periodically transmitted every N data chip blocks as shown in Fig. 2. The pilot chip block $\{s_p(t);\ t=0\sim(N_c-1)\}$ is given by

$$s_p(t) = a(t \bmod (N_c/2)) \quad . \qquad (12)$$

The received pilot chip block $\{r_p(t);\ t=0\sim(N_c-1)\}$ is given by

$$r_p(t) = \sqrt{\frac{2PU}{SF}} \sum_{l=0}^{L-1} h_l a((t-\tau_l) \bmod (N_c/2)) + \mu(t) + \eta(t) \ . \quad (13)$$

1) CE using the latter half of pilot chip block

The first half of the pilot chip block, $\{s_p(t);\ t=0\sim(N_c/2-1)\}$, plays a role of the cyclic prefix for the latter half of the pilot chip block, $\{s_p(t);\ t=N_c/2\sim(N_c-1)\}$. By applying $N_c/2$-point FFT to $\{r_p(t);\ t=N_c/2\sim(N_c-1)\}$, as shown in Fig. 3(a), the received pilot chip block can be transformed into $N_c/2$ frequency components $\{R_1(q); q=0\sim(N_c/2-1)\}$ without causing the IBI if the maximum delay time τ_{max} of the channel is shorter than $N_c/2$.

$R_1(q)$ is given by

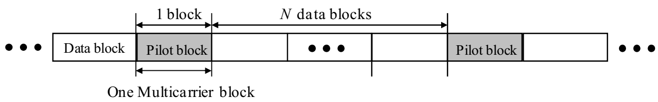
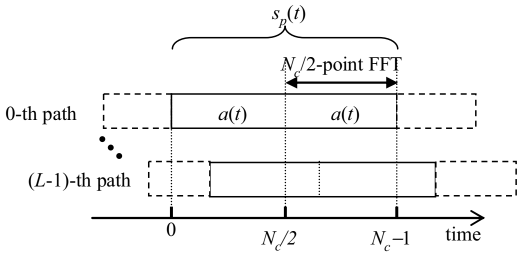
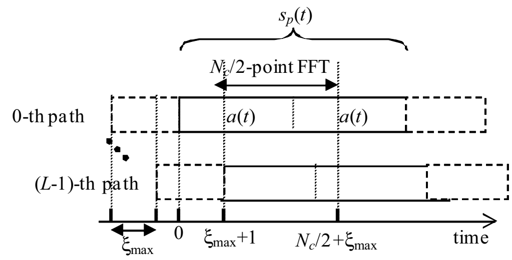


Fig. 2 Frame format.



(a) FFT window for the latter half of the pilot block



(b) FFT window for the first half of the pilot block

Fig. 3 Received pilot block and FFT window timing for CE.

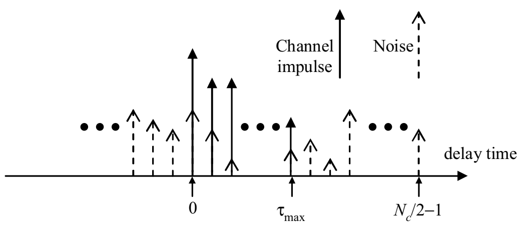


Fig. 4 Instantaneous channel impulse response.

$$\begin{aligned} R_1(q) &= \frac{1}{N_c/2} \sum_{t=0}^{N_c/2-1} r_p(t + N_c/2) \exp\left(-j2\pi q \frac{t}{N_c/2} \right) \\ &= \sqrt{U} H(2q) A(q) + \Pi(2q) \end{aligned} , \quad (14)$$

where

$$A(q) = \frac{1}{N_c/2} \sum_{t=0}^{N_c/2-1} a(t) \exp\left(-j2\pi q \frac{t}{N_c/2} \right) \qquad (15)$$

with $|A(q)|^2=2$. The initial channel gain estimate $\widetilde{H}_1(2q)$, $q=0\sim(N_c/2-1)$, is obtained as

$$\widetilde{H}_1(2q) = R_1(q)/A(q) = \sqrt{U} H(2q) + \Pi(2q)/A(q) \ . \quad (16)$$

However, an interpolation technique is necessary since FDE requires N_c channel gain estimates $\{\overline{H}_1(k);\ k=0\sim(N_c-1)\}$. Furthermore, $\widetilde{H}_1(2q)$ is noisy. Therefore, we apply the delay time-domain windowing technique [14] to reduce the noise while interpolating the initial channel gain estimates $\{\widetilde{H}_1(2q)$ $q=0\sim(N_c/2-1)\}$ to obtain $\{\overline{H}_1(k);\ k=0\sim(N_c-1)\}$.

The instantaneous channel impulse response estimate $\{\widetilde{h}_1(\tau);\ \tau=0\sim(N_c/2-1)\}$, can be obtained by performing $N_c/2$-point IFFT on $\{\widetilde{H}_1(2q);\ q=0\sim(N_c/2-1)\}$ as

$$\widetilde{h}_1(\tau) = \sum_{q=0}^{N_c-1} \widetilde{H}_1(2q) \exp\left(-j2\pi\tau \frac{q}{N_c/2} \right) \quad . \qquad (17)$$

The channel impulse response is present only over $\tau=0\sim\tau_{max}$, while the noise due to the AWGN spreads over the entire delay time-domain as shown in Fig. 4. To reduce the noise, $\widetilde{h}_1(\tau)$ is replaced by zeros for $\tau=\tau_{max}+1\sim(N_c-1)$ and an N_c-point FFT is applied. Improved channel gain estimates $\{\overline{H}_1(k);\ k=0\sim(N_c-1)\}$ can be obtained as [14, 15]

$$\begin{aligned} \overline{H}_1(k) &= \sum_{\tau=0}^{\tau_{max}} \widetilde{h}_1(\tau) \exp\left(-j2\pi k \frac{\tau}{N_c} \right) \\ &= \sum_{q=0}^{N_c/2-1} \widetilde{H}_1(2q) \left\{ \frac{\sin\left(\pi \frac{k-2q}{N_c} \right) \tau_{max}}{\sin\left(\pi \frac{k-2q}{N_c} \right)} \exp\left(-j\pi \frac{k-2q}{N_c} (\tau_{max}-1) \right) \right\} \end{aligned} \quad . \qquad (18)$$

In CPACE, the maximum delay time τ_{max} must be known. τ_{max} can be estimated using the average power delay profile $\{E[|\widetilde{h}_1(\tau)|^2];\ \tau=0\sim(N_c/2-1)\}$. Although the actual channel impulse response is present only over $\tau=0\sim\tau_{max}$, we introduce the threshold L_{th} to determine the maximum delay time as $\tau_{max}=\arg\max\ \{E[|\widetilde{h}_1(\tau)|^2] \le L_{th}\}$. The optimum L_{th} that minimizes the BER is found by computer simulation.

2) CE using the first half of pilot block

So far, we have described the channel estimation scheme which uses the latter half of the pilot chip block $\{s_p(t);\ t=N_c/2\sim(N_c-1)\}$. If the maximum delay time τ_{max} is known, we can apply an $N_c/2$-point FFT to the first half of the received pilot chip block $\{r_p(t);\ t=(\tau_{max}+1)\sim(N_c/2+\tau_{max})\}$, as shown in Fig. 3(b), to obtain $N_c/2$ frequency components $\{R_2(q); q=0\sim(N_c/2-1)\}$ without causing IBI. $R_2(q)$ is given by

$$R_2(q) = \frac{1}{N_c/2} \sum_{t=\tau_{max}+1}^{N_c/2+\tau_{max}} r_p(t) \exp\left(-j2\pi q \frac{t}{N_c/2}\right) \quad . \quad (19)$$
$$= \sqrt{U} H(2q) A(q) + \Pi(q)$$

The initial channel gain estimate $\widetilde{H}_2(2q)$ can be obtained as

$$\widetilde{H}_2(q) = R_2(q)/A(q) = \sqrt{U} H(2q) + \Pi(q)/A(q) \quad . (20)$$

An $N_c/2$-point IFFT is applied to decompose $\{\widetilde{H}_2(2q)$ q=0~($N_c/2-1$)} into the instantaneous channel impulse response estimate $\{\widetilde{h}_2(\tau)$; τ=0~($N_c/2-1$)} as

$$\widetilde{h}_2(\tau) = \sum_{q=0}^{N_c/2-1} \widetilde{H}_2(2q) \exp\left(-j2\pi\tau \frac{q}{N_c/2}\right). \qquad (21)$$

Replacing $\widetilde{h}_2(\tau)$ by zeros for $\tau=\tau_{max}+1$~(N_c-1) and applying an N_c-point FFT, N_c channel gain estimates $\{\overline{H}_2(k)$; k=0~(N_c-1)} can be obtained. The final channel estimate is obtained as

$$\overline{H}(k) = (\overline{H}_1(k) + \overline{H}_2(k))/2 \quad . \qquad (22)$$

B. Estimation of IBI plus noise power σ^2

For performing MMSE-FDE, the IBI plus noise power σ^2 must be known (see Eq. (11)). σ^2 can be estimated as follows.

The received pilot chip block $\{r_p(t)$: t=0~(N_c-1)} is decomposed by using an N_c-point FFT into N_c subcarrier components $\{R_p(k)$; k=0~(N_c-1)} as

$$R_p(k) = \frac{1}{N_c} \sum_{t=0}^{N_c-1} r_p(t) \left(-j2\pi k \frac{t}{N_c}\right) \quad , \qquad (23)$$
$$= \sqrt{U} H(k) S_p(k) + \mathrm{N}(k) + \Pi(k)$$

where

$$S_p(k) = \frac{1}{N_c} \sum_{t=0}^{N_c-1} s_p(t) \exp\left(-j2\pi k \frac{t}{N_c}\right) = \begin{cases} A(k) & \text{for } k = 2q \\ 0 & \text{for } k = 2q+1 \end{cases}, \quad q=0\sim N_c/2-1. \quad (24)$$

$R_p(2q+1)$, q=0~$N_c/2-1$, contains the IBI plus noise component only. The estimate of IBI plus noise power $\hat{\sigma}^2$ can be obtained using

$$2\hat{\sigma}^2 = \frac{2}{N_c} \sum_{q=0}^{N_c/2-1} |R_p(2q+1)|^2 \quad . \qquad (25)$$

IV. COMPUTER SIMULATION

The simulation condition is summarized in Table 1. Quadrature phase shift keying (QPSK) data modulation, N_c=256-subcarrier MC-CDMA, and a frequency-selective Rayleigh fading channel having an L=16-path uniform power delay profile are assumed. The normalized maximum Doppler frequency $f_D T_c N_c$ is assumed to be 0.001 (this corresponds to a mobile terminal speed of about 80 km/h for 100Msps and 5 GHz carrier-frequency).

One pilot chip block is transmitted every N=15 data blocks (binary phase shift (BPSK) modulation is used). A partial sequence taken from an M-sequence of 4095 bits is used as the pilot $\{A(q)$;q=0~($N_c/2-1$)}. To estimate the channel at data chip blocks between the two pilot blocks, linear interpolation [18] is used. For estimation of τ_{max}, the optimum threshold value L_{th} to minimize the BER was found by computer simulation.

Table 1 Simulation condition

Data modulation	Data	QPSK
	Pilot	BPSK
MC-CDMA	No. of subcarriers	N_c=256
	Scrambling code	4095-chip PN
	Spreading codes	Walsh codes
	Spreading factor	SF=1, 16
	No. of users	U=1, 16
Channel model	No. of paths	L=16
	Power delay profile	Uniform
	Time delay	$\tau_l=l\Delta T_c$, l=0~$L-1$, Δ=1, 2
Overlap FDE	FFT window size	256 (=N_c)
	FDE weight	MMSE

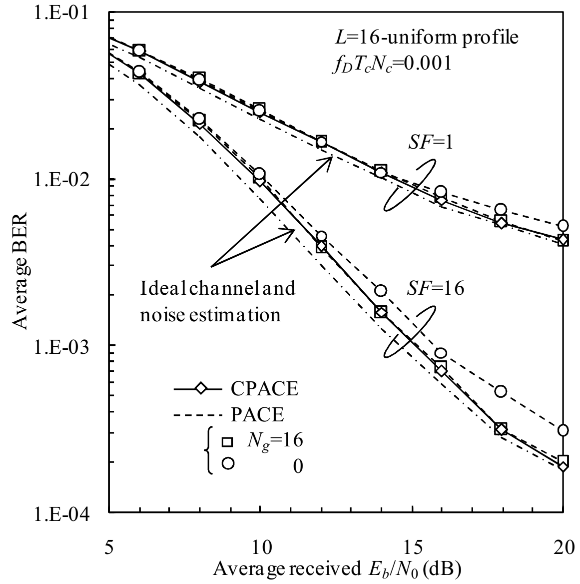


Fig. 5 BER performance with perfect knowledge of τ_{max}.

Figure 5 shows the BER performance with CPACE as a function of the average received E_b/N_0 when the maximum delay time τ_{max} is known. For comparison, the BER performance is also plotted for PACE using pilot with and without GI insertion (N_g=16) [13]. It can be seen from Fig. 5 that CPACE provides almost the same BER performance as PACE with GI-inserted pilot. The degradation in the required E_b/N_0 for BER=10^{-2} from the ideal channel estimation case is as small as 0.5 (0.6) dB when SF=1 (16) (out of which about 0.28 dB is due to the pilot insertion). On the other hand, PACE with no GI-inserted pilot degrades the BER performance due to the residual IBI.

Figure 6 shows that the BER performance of CPACE when the maximum delay time τ_{max} is estimated. It can be seen from Fig. 6 that almost the same BER performance can be obtained as the known τ_{max} case.

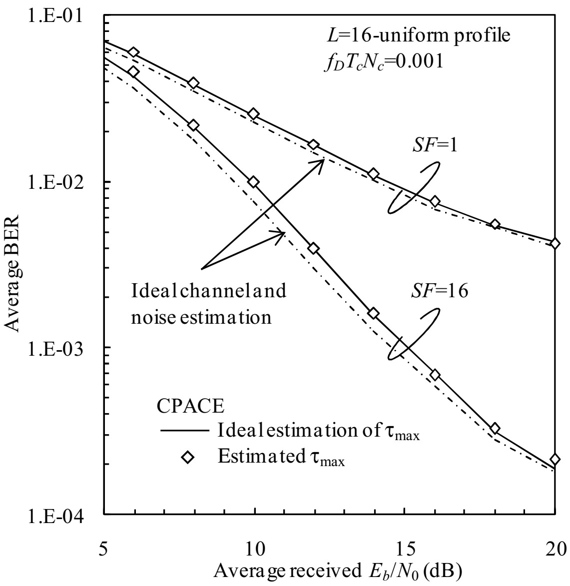


Fig. 6 BER performance with estimation of τ_{max}.

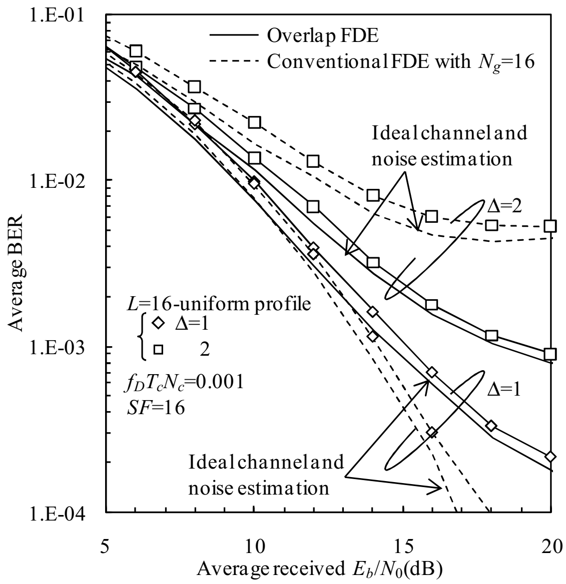


Fig. 7 Comparison between overlap FDE and conventional FDE.

Figure 7 compares the BER performances of the overlap FDE using CPACE and the conventional FDE using PACE with N_g=16-chip GI when SF=U=16. The BER performances are plotted with the time delay difference Δ between the paths as a parameter. As Δ increases from 1 to 2, the BER performance of the conventional FDE significantly degrades due to the IBI caused by delayed paths whose time delays exceed the GI length. The BER performance of overlap FDE also degrades since the residual IBI gets stronger; however, the performance degradation is much smaller than the conventional FDE even using the channel estimation.

In the real fading channel environment, the channel selectivity changes. Even in a weak frequency-selective channel (e.g., Δ=1), the conventional FDE must use a fixed length GI which is longer than the expected maximum path time delay. On the other hand, the overlap FDE can simply change the value of M to adapt to the changing in channel frequency-selectivity at the cost of increased computational complexity.

V. CONCLUSION

In this paper, we proposed a cyclic pilot assisted channel estimation technique (CPACE) suitable for MC-CDMA using overlap FDE. It was shown by the computer simulation that CPACE provides a good BER performance and the degradation of required E_b/N_0 for BER=10^{-2} from the ideal channel estimation case is as small as 0.6 dB.

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