

## PAPER

# Pilot-Assisted Adaptive Channel Estimation for Coded MC-CDMA with ICI Cancellation

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**SUMMARY** One of the promising wireless access techniques for the next generation mobile communications systems is multi-carrier code division multiple access (MC-CDMA). MC-CDMA can provide good transmission performance owing to the frequency diversity effect in a severe frequency-selective fading channel. However, the bit error rate (BER) performance of coded MC-CDMA is inferior to that of orthogonal frequency division multiplexing (OFDM) due to the residual inter-code interference (ICI) after frequency-domain equalization (FDE). Recently, we proposed a frequency-domain soft interference cancellation (FDSIC) to reduce the residual ICI and confirmed by computer simulation that the MC-CDMA with FDSIC provides better BER performance than OFDM. However, ideal channel estimation was assumed. In this paper, we propose adaptive decision-feedback channel estimation (ADFCE) and evaluate by computer simulation the average BER and throughput performances of turbo-coded MC-CDMA with FDSIC. We show that even if a practical channel estimation is used, MC-CDMA with FDSIC can still provide better performance than OFDM.

**key words:** MC-CDMA, interference cancellation, channel estimation

## 1. Introduction

In the next generation mobile communications systems, broadband packet services will be in great demand [1]. Hybrid automatic repeat request (HARQ), which is a combination of ARQ and error correcting coding (e.g., turbo code [2]), is an indispensable error control technique for high quality packet data transmissions [3], [4]. However, for such high-speed data transmissions, the wireless channel becomes severely frequency-selective and the throughput performance significantly degrades due to a large inter-symbol interference (ISI) [5], [6].

Recently, multi-carrier code division multiple access (MC-CDMA), which uses a number of low rate subcarriers to reduce the ISI, has been attracting much attention as a broadband wireless access technique because of its high flexibility in variable rate data transmissions by code-multiplexing [7], [8]. Since the property of code orthogonality among different orthogonal spreading codes used in the code-multiplexing is distorted due to the channel frequency-selectivity, the inter-code interference (ICI) is produced. Frequency-domain equalization (FDE) based on minimum mean square error (MMSE) criterion can suppress ICI while minimizing the noise enhancement and signifi-

cantly improve the transmission performance of multi-code MC-CDMA in a severe frequency-selective fading channel [7], [8]. However, the frequency-distorted signal cannot be completely equalized by MMSE-FDE and the residual ICI after MMSE-FDE degrades the performance as the code multiplex order increases. So far, we have shown [9], [10] that the use of frequency-domain soft interference cancellation (FDSIC) can reduce the residual ICI and hence, significantly improve the transmission performance of multi-code MC-CDMA. If the residual ICI is perfectly removed, MC-CDMA can provide a good performance than that of orthogonal frequency division multiplexing (OFDM). Accurate estimation of the channel transfer function and the received signal-to-noise power ratio (SNR) is necessary for FDSIC. However, in [9], [10], ideal channel estimation (ideal CE) was assumed.

In this paper, we propose adaptive decision-feedback channel estimation (ADFCE) using time-multiplexed pilot. The impact of channel estimation error on the average BER and throughput performances of turbo-coded MC-CDMA with FDSIC is evaluated by computer simulation.

The remainder of this paper is organized as follows. Section 2 introduces the turbo-coded MC-CDMA HARQ system. Section 3 describes the proposed ADFCE. In Sect. 4, we show the computer simulation results on the average BER and throughput performances. This paper is concluded in Sect. 5.

## 2. MC-CDMA HARQ System Model

The transmitter/receiver structure for turbo-coded MC-CDMA HARQ with FDSIC is illustrated in Fig. 1 and transmitted packet structure is illustrated in Fig. 2. In this paper, we consider Chase combining [11] as a packet combining.

### 2.1 Transmitted and Received Signals

At the transmitter, after turbo coding and puncturing, the turbo-coded bit sequence is stored in the transmitter buffer. The turbo-coded bit sequence is transformed into a data modulated symbol sequence and the symbol sequence is serial-to-parallel (S/P) converted to  $U$  symbol streams  $\{d_u(n); n = 0 \sim N_c/SF - 1\}$ ,  $u = 0 \sim U - 1$ , where  $N_c$  is the fast Fourier transform (FFT) block size and  $SF$  is the spreading factor. Data symbols in each stream are spread by an orthogonal code  $\{c_u(k); k = 0 \sim SF - 1\}$  and then  $U$  streams are added and multiplied by a scrambling code

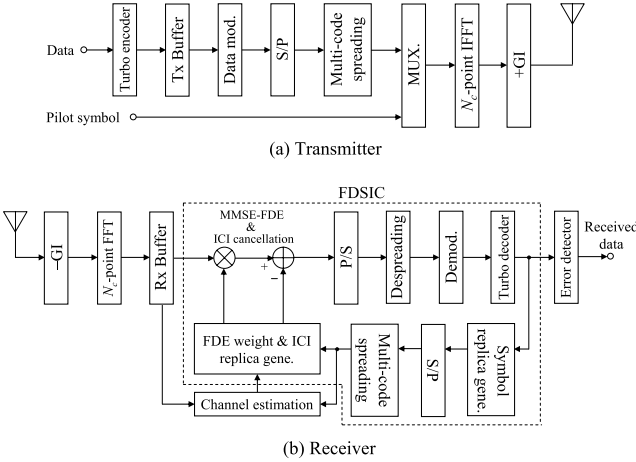
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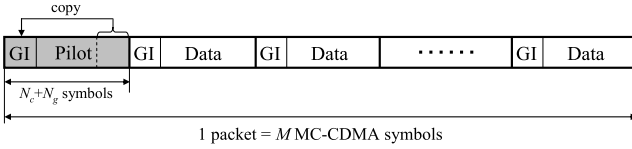
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**Fig. 1** Turbo-coded MC-CDMA HARQ system model.



**Fig. 2** Transmitted packet structure.

$\{c_{scr}(k); k = 0 \sim N_c - 1\}$ . The resultant chip sequence is transformed by  $N_c$ -point inverse fast Fourier transform (IFFT) into an MC-CDMA symbol  $\{s(t); t = 0 \sim N_c - 1\}$ . Then, the last  $N_g$  samples are copied as a cyclic prefix and inserted into the guard interval (GI) at the beginning of the MC-CDMA symbol. A packet to be transmitted is composed of 1 pilot MC-CDMA symbol for channel estimation and  $M - 1$  data MC-CDMA symbols. One turbo-coded bit sequence is transmitted by one MC-CDMA symbol; therefore,  $M - 1$  turbo-coded bit sequences are transmitted in one packet.

The  $m$ th MC-CDMA symbol  $s_m(t)$ ,  $m = 1 \sim M - 1$ , in a packet can be expressed as

$$s_m(t) = \sqrt{\frac{2P}{SF}} \frac{1}{\sqrt{N_c}} \sum_{k=0}^{N_c-1} S_m(k) \exp\left(j2\pi t \frac{k}{N_c}\right), \quad (1)$$

where  $P$  is the transmit power per data symbol and  $S_m(k)$  denotes the  $k$ th subcarrier component.  $S_m(k)$  is expressed as

$$S_m(k) = \sum_{u=0}^{U-1} d_{m,u} (\lfloor k/SF \rfloor) c_u(k \bmod SF) c_{scr}(k), \quad (2)$$

where  $\lfloor x \rfloor$  denotes the largest integer smaller than or equal to  $x$ .

The transmitted signal is received at the receiver via a frequency-selective fading channel. The propagation channel is assumed to be a frequency-selective block fading channel having sample-spaced  $L$  discrete paths, each subjected to independent fading. The channel impulse response  $h_m(t)$  can be expressed as [12]

$$h_m(t) = \sum_{l=0}^{L-1} h_{m,l} \delta(t - \tau_l), \quad (3)$$

where  $h_{m,l}$  and  $\tau_l$  are the complex-valued path gain and time delay of the  $l$ th ( $l = 0 \sim L - 1$ ) path, respectively, with  $E[\sum_{l=0}^{L-1} |h_{m,l}|^2] = 1$  ( $E[\cdot]$  denotes the ensemble average operation). The maximum delay-time  $\tau_{L-1}$  is assumed to be shorter than GI length. We assume a block fading channel so that the path gains remain constant over 1 MC-CDMA symbol.

At the receiver, channel estimation is carried out first by using pilot MC-CDMA symbol. Channel estimation proposed in this paper is described in Sect. 3. Next, the following operation is performed every data MC-CDMA symbol. We assume that the  $m$ th MC-CDMA symbol  $s_m(t)$  has been transmitted  $Q$  times.

The  $q$ th received MC-CDMA symbol  $r_m^{(q)}(t)$ ,  $q = 0 \sim Q - 1$ , can be expressed as

$$r_m^{(q)}(t) = \sum_{l=0}^{L-1} h_{m,l}^{(q)} s_m(t - \tau_l) + \eta_m^{(q)}(t), \quad (4)$$

where  $\eta_m^{(q)}(t)$  is a zero-mean complex Gaussian noise having variance  $2\sigma^2$ . After the removal of the GI,  $N_c$ -point FFT is applied to decompose  $\{r_m^{(q)}(t); t = 0 \sim N_c - 1\}$  into  $N_c$  subcarrier components  $\{R_m^{(q)}(k); k = 0 \sim N_c - 1\}$ .  $R_m^{(q)}(k)$  is represented as

$$\begin{aligned} R_m^{(q)}(k) &= \frac{1}{\sqrt{N_c}} \sum_{t=0}^{N_c-1} r_m^{(q)}(t) \exp\left(-j2\pi k \frac{t}{N_c}\right) \\ &= \sqrt{\frac{2P}{SF}} H_m^{(q)}(k) S_m(k) + \Pi_m^{(q)}(k), \end{aligned} \quad (5)$$

where  $H_m^{(q)}(k)$  and  $\Pi_m^{(q)}(k)$  are the channel gain and the noise at the  $k$ th subcarrier frequency, respectively. They are given by

$$\begin{cases} H_m^{(q)}(k) = \sum_{l=0}^{L-1} h_{m,l}^{(q)} \exp\left(-j2\pi k \frac{\tau_l}{N_c}\right) \\ \Pi_m^{(q)}(k) = \frac{1}{\sqrt{N_c}} \sum_{t=0}^{N_c-1} \eta_m^{(q)}(t) \exp\left(-j2\pi k \frac{t}{N_c}\right) \end{cases}. \quad (6)$$

$R_m^{(q)}(k)$  is stored in the receiver buffer and then, FDSIC is performed.

## 2.2 FDSIC

In FDSIC, a series of joint MMSE-FDE and Chase combining, ICI cancellation, despreading, and turbo decoding is iterated  $I$  times.

Joint MMSE-FDE and Chase combining for the  $i$ th ( $i = 0 \sim I - 1$ ) iteration can be expressed as [9], [10]

$$\begin{aligned} \hat{R}_{m,i}^{(Q-1)}(k) &= \sum_{q=0}^{Q-1} R_m^{(q)}(k) w_{m,i}^{(q)}(k) \\ &= \frac{1}{\sqrt{SF}} \left( \frac{1}{SF} \sum_{k'=\lfloor k/SF \rfloor SF}^{(\lfloor k/SF \rfloor + 1)SF - 1} \hat{H}_{m,i}^{(Q-1)}(k') \right) S_m(k) \end{aligned}$$

$$+ M_{m,i}^{(Q-1)}(k) + \sum_{q=0}^{Q-1} \Pi_m^{(q)}(k) w_{m,i}^{(q)}(k), \quad (7)$$

where

$$\hat{H}_{m,i}^{(Q-1)}(k) = \sum_{q=0}^{Q-1} \sqrt{2\overline{P}H_m^{(q)}}(k) w_{m,i}^{(q)}(k). \quad (8)$$

$M_{m,i}^{(Q-1)}(k)$  is the frequency-domain representation of the residual ICI and is given by [9]

$$\begin{aligned} & M_{m,i}^{(Q-1)}(k) \\ &= \frac{1}{\sqrt{SF}} \left\{ \hat{H}_{m,i}^{(Q-1)}(k) - \frac{1}{SF} \sum_{k'=\lfloor k/SF \rfloor SF}^{\lfloor (k/SF+1)SF-1 \rfloor} \hat{H}_{m,i}^{(Q-1)}(k') \right\} S_m(k). \end{aligned} \quad (9)$$

$w_{m,i}^{(q)}(k)$  is the MMSE weight and can be expressed as

$$\begin{cases} w_{m,i}^{(q)}(k) = \frac{\{\sqrt{2\overline{P}H_m^{(q)}}(k)\}^*}{\frac{1}{SF} \sum_{q=0}^{Q-1} |\sqrt{2\overline{P}H_m^{(q)}}(k)|^2 \sum_{u=0}^{U-1} \rho_{m,u,i-1}^{(Q-1)} (\lfloor k/SF \rfloor) + 2\sigma^2}, \\ \rho_{m,u,i-1}^{(Q-1)}(n) = E[|d_{m,u}(n)|^2] - |\tilde{d}_{m,u,i-1}^{(Q-1)}(n)|^2 \end{cases} \quad (10)$$

where  $\rho_{m,u,i-1}^{(Q-1)}(n)$  is an interference coefficient and  $\rho_{m,u,-1}^{(Q-1)}(n) = 1 \cdot E[|d_{m,u}(n)|^2]$  is the expectation of  $|d_{m,u}(n)|^2$  for the given received packet and  $\tilde{d}_{m,u,i-1}^{(Q-1)}(n)$  is the soft symbol replica generated from the result of the  $(i-1)$ th iteration.  $E[|d_{m,u}(n)|^2]$  and  $\tilde{d}_{m,u,i-1}^{(Q-1)}(n)$  are given by [9]

$$\begin{aligned} E[|d_{m,u}(n)|^2] &= \sum_{d \in D} |d|^2 \prod_{b_{m,u,x}(n) \in d} p(b_{m,u,x}(n)) \\ &= \begin{cases} 1 & \text{for QPSK} \\ \frac{4}{10} \tanh\left(\frac{\lambda_{m,u,i-1,1}^{(Q-1)}(n)}{2}\right) + \frac{4}{10} \tanh\left(\frac{\lambda_{m,u,i-1,3}^{(Q-1)}(n)}{2}\right) + 1 & \text{for 16QAM,} \end{cases} \end{aligned} \quad (11)$$

and

$$\begin{aligned} \tilde{d}_{m,u,i-1}^{(Q-1)} &= \sum_{d \in D} d \prod_{b_{m,u,x}(n) \in d} p(b_{m,u,x}(n)) \\ &= \begin{cases} \frac{1}{\sqrt{2}} \tanh\left(\frac{\lambda_{m,u,i-1,0}^{(Q-1)}(n)}{2}\right) + j \frac{1}{\sqrt{2}} \tanh\left(\frac{\lambda_{m,u,i-1,1}^{(Q-1)}(n)}{2}\right) & \text{for QPSK} \\ \frac{1}{\sqrt{10}} \tanh\left(\frac{\lambda_{m,u,i-1,0}^{(Q-1)}(n)}{2}\right) \left\{ 2 + \tanh\left(\frac{\lambda_{m,u,i-1,1}^{(Q-1)}(n)}{2}\right) \right\} \\ + j \frac{1}{\sqrt{10}} \tanh\left(\frac{\lambda_{m,u,i-1,2}^{(Q-1)}(n)}{2}\right) \left\{ 2 + \tanh\left(\frac{\lambda_{m,u,i-1,3}^{(Q-1)}(n)}{2}\right) \right\} & \text{for 16QAM,} \end{cases} \end{aligned} \quad (12)$$

where  $d$  represents the candidate symbol in the symbol set  $D$  and  $\lambda_{m,u,i-1,x}^{(Q-1)}(n)$  is the turbo decoder output LLR associated with the  $x$ th bit  $b_{m,u,x}(n)$  in the  $n$ th symbol  $d_{m,u}(n)$  ( $x = 0 \sim X-1$ , where  $X$  is equal to 2 and 4 for QPSK and 16QAM data modulation, respectively).  $p(b_{m,u,x}(n))$  is the *a posteriori* probability of  $b_{m,u,x}(n)$  for the given received packet [9].

After joint MMSE-FDE and Chase combining,  $i$ th ICI cancellation is carried out as [9]

$$\tilde{R}_{m,i}^{(Q-1)}(k) = \hat{R}_{m,i}^{(Q-1)}(k) - \tilde{M}_{m,i}^{(Q-1)}(k), \quad (13)$$

where  $\tilde{M}_{m,i}^{(Q-1)}(k)$  is the replica of  $M_{m,i}^{(Q-1)}(k)$  obtained by replacing in Eq. (9) with the MC-CDMA signal replica  $\tilde{S}_{m,i-1}^{(Q-1)}(k)$ , which is generated by substituting  $\tilde{d}_{m,u,i-1}^{(Q-1)}(n)$  into Eq. (2) instead of  $d_{m,u}(n)$ . We assume  $\tilde{S}_{m,-1}^{(Q-1)}(k) = 0$ . Despreading is carried out as

$$\tilde{d}_{m,u,i}^{(Q-1)}(n) = \frac{1}{SF} \sum_{k=nSF}^{(n+1)SF-1} \tilde{R}_{m,i}^{(Q-1)}(k) c_u^*(k \bmod SF) c_{scr}^*(k) \quad (14)$$

to obtain a sequence of log-likelihood ratios (LLRs) [9] and the turbo decoding is carried out.

After carrying out FDSIC on  $M-1$  turbo-coded bit sequences, error information is feedback to the transmitter. In the next packet transmission, the same turbo-coded bit sequence is retransmitted in the same MC-CDMA symbol position in the packet; otherwise a new turbo-coded bit sequence is transmitted. In this paper, ideal error detection and no transmission error of the error information are assumed.

### 3. Adaptive Decision-Feedback Channel Estimation (ADFCE)

$\sqrt{2\overline{P}H_m}(k)$  and  $\sigma^2$  in Eqs. (8), (9) and (10) are unknown to the receiver and need to be estimated. In this paper, we propose ADFCE, in which pilot-assisted channel estimation (PACE) [13] is carried out to obtain the instantaneous channel gain estimate at  $m=0$  and decision-feedback channel estimation (DFCE) [14], [15] is carried out to obtain the instantaneous channel gain estimate at  $m \geq 1$ , then, infinite impulse response (IIR) filtering [15], [16] is performed to improve the channel gain estimate. The IIR filter coefficient is adapted to track the fading variation by recursive least-square (RLS) algorithm [16] in this paper.

#### 3.1 PACE and DFCE

The pilot MC-CDMA symbol is represented by  $m=0$ . Without loss of generality, we assume  $S_0(k) = \pm 1 + j0$ . The estimate  $\tilde{H}_0(k)$  of  $\sqrt{2\overline{P}H_0}(k)$  can be expressed as

$$\begin{aligned} \tilde{H}_0(k) &= R_0(k) \cdot \frac{S_0^*(k)}{|S_0(k)|^2} \\ &= \sqrt{2\overline{P}H_0}(k) + \Pi_0(k) \frac{S_0^*(k)}{|S_0(k)|^2}, \end{aligned} \quad (15)$$

where  $R_0(k)$  is the  $k$ th subcarrier component of received pilot symbol and can be expressed as

$$R_0(k) = \sqrt{2P}H_0(k)S_0(k) + \Pi_0(k). \quad (16)$$

Then, the delay time-domain windowing technique [17] is carried out to get the noise-reduced estimate. The noise power  $\sigma^2$  in Eq. (10) can be estimated as described in [14].

When  $m \geq 1$ ,  $\tilde{H}_m(k)$  is obtained by replacing pilot symbol by MC-CDMA signal replica. We use MMSE channel estimation [14], [15] and the MC-CDMA signal replica  $\tilde{S}_{m,I-1}(k)$  generated at the  $(I-1)$ th iteration in FDSIC.  $\tilde{H}_m(k)$  is given by

$$\tilde{H}_m(k) = R_m(k) \cdot X_m(k), \quad (17)$$

where  $X_m(k)$  is the reference signal to remove data modulation from  $R_m(k)$ . We define the channel estimation error  $e_m(k)$  as [14]

$$e_m(k) = R_m(k) \cdot X_m(k) - \sqrt{2P}H_m(k). \quad (18)$$

Solving  $\partial E[|e_m(k)|^2]/\partial X_m(k) = 0$  gives

$$X_m(k) = \frac{\sqrt{SF}\tilde{S}_{m,I-1}^*(k)}{|\tilde{S}_{m,I-1}(k)|^2 + \left(\frac{1}{SF} \frac{P}{\sigma^2}\right)^{-1}}. \quad (19)$$

The delay time-domain windowing technique is carried out to reduce the noise due to AWGN and decision error. The SNR  $P/\sigma^2$  in Eq. (19) is estimated in the delay time-domain [14]. The expectation of  $\tilde{H}_m(k)$  is not the same as  $\sqrt{2P}H_m(k)$  [14]. The channel gain estimate  $\tilde{H}'_m(k)$ , whose expectation is equal to  $\sqrt{2P}H_m(k)$ , can be obtained as

$$\tilde{H}'_m(k) = \frac{\tilde{H}_m(k)}{A_m}, \quad (20)$$

where

$$A_m = \frac{1}{N_c} \sum_{k=0}^{N_c-1} \frac{|\tilde{S}_{m,I-1}(k)|^2}{|\tilde{S}_{m,I-1}(k)|^2 + \left(\frac{1}{SF} \frac{P}{\sigma^2}\right)^{-1}}. \quad (21)$$

### 3.2 Adaptive IIR Filtering

$\tilde{H}'_m(k)$  is still perturbed due to AWGN and decision error. To improve the channel gain estimate, we use the first-order IIR filter [16].  $\tilde{H}_m(k)$  to be used in FDSIC can be expressed as

$$\tilde{H}_m(k) = \begin{cases} \tilde{H}_0(k) & \text{if } m = 1 \\ \{1 - \beta(k)\}\tilde{H}_{m-1}(k) + \beta(k)\tilde{H}'_{m-1}(k) & \text{if } m \geq 2 \end{cases}, \quad (22)$$

where  $\beta(k)$  ( $0 \leq \beta(k) \leq 1$ ) is an important design parameter to tradeoff between the noise reduction and the tracking ability against fading.  $\beta(k) = 0$  corresponds to PACE using pilot only. In a slow fading channel,  $\beta(k)$  should be smaller to better reduce the noise due to AWGN and decision error.

As the value of  $\beta(k)$  increases, the IIR filter can achieve better tracking ability, but the noise reduction tends to be lost. Therefore, there is an optimum value in  $\beta(k)$ , which depends on the received SNR and the Doppler spread. We adapt  $\beta(k)$  by using RLS algorithm. The cost function  $J_m(k)$  defined as the sum of exponentially weighted channel estimation errors is expressed as [16]

$$J_m(k) = \sum_{m'=1}^{m-1} \lambda^{m-m'-1} |\tilde{H}'_{m'}(k) - \tilde{H}_{m'}(k)|^2, \quad (23)$$

where  $\lambda$  is the forgetting factor ( $0 < \lambda \leq 1$ ) and  $\tilde{H}'_{m'}(k)$  is the reference. Substituting Eq. (22) into Eq. (23) and solving  $\partial J_m(k)/\partial \beta(k) = 0$ , we obtain

$$\beta_m(k) = \frac{\sum_{m'=2}^{m-1} \lambda^{m-m'-1} \text{Re} \left\{ \left\{ \tilde{H}'_{m'}(k) - \tilde{H}_{m'-1}(k) \right\} \left\{ \tilde{H}'_{m'-1}(k) - \tilde{H}_{m'-1}(k) \right\}^* \right\}}{\sum_{m'=2}^{m-1} \lambda^{m-m'-1} |\tilde{H}'_{m'-1}(k) - \tilde{H}_{m'-1}(k)|^2}. \quad (24)$$

We have the following recursive algorithm:

$$\begin{cases} \beta_m(k) = \frac{\Theta_m(k)}{\Omega_m(k)} \\ \Theta_m(k) = \lambda \Theta_{m-1}(k) + \text{Re} \left\{ B_{m-1}(k) C_{m-1}^*(k) \right\} \\ \Omega_m(k) = \lambda \Omega_{m-1}(k) + |C_{m-1}(k)|^2 \\ B_{m-1}(k) = \tilde{H}'_{m-1}(k) - \tilde{H}_{m-2}(k) \\ C_{m-1}(k) = \tilde{H}'_{m-2}(k) - \tilde{H}_{m-2}(k) \end{cases} \quad (25)$$

with  $m \geq 3$ . The initial condition is set as  $\Theta_2(k) = 0$  and  $\Omega_2(k) = \varepsilon$  (small positive value). Since the statistical property of  $H_m(k)$  is identical for all  $k$ ,  $\beta_m(k)$  should converge with the identical value irrespective of  $k$ . This can be exploited to reduce the noise and hence to achieve the faster convergence. We use  $\beta_m$  given as

$$\beta_m = \frac{1}{N_c} \sum_{k=0}^{N_c-1} \beta_m(k). \quad (26)$$

## 4. Simulation Results

Table 1 shows the simulation condition. The single-user transmission with full code-multiplexed ( $U = SF$ ) MC-CDMA is assumed so that the same transmission rate is achieved as OFDM. A turbo encoder with two (13, 15) RSC encoders and a decoder with Log-MAP algorithm are used. The length of the coded bit sequence is 512 and 1024 for QPSK and 16QAM data modulation, respectively. We assume the FFT block size of  $N_c = 256$  and the GI length of  $N_g = 32$ . The channel is assumed to be a frequency-selective block Rayleigh fading channel having a sample-spaced  $L = 16$ -path exponential power delay profile with decay factor  $\alpha$ . The length  $M$  of the transmitted packet is 64.

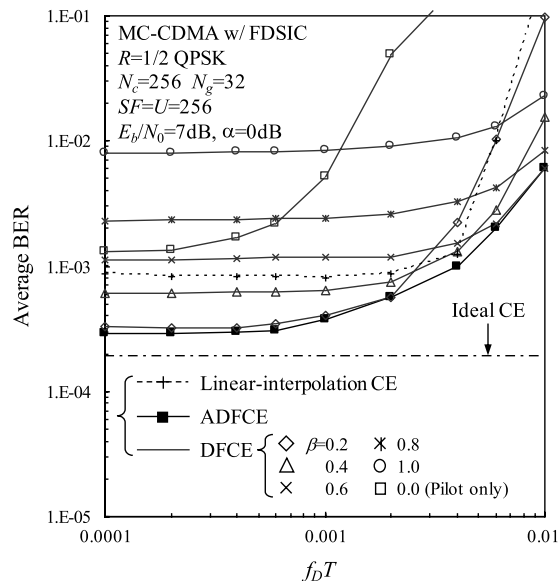
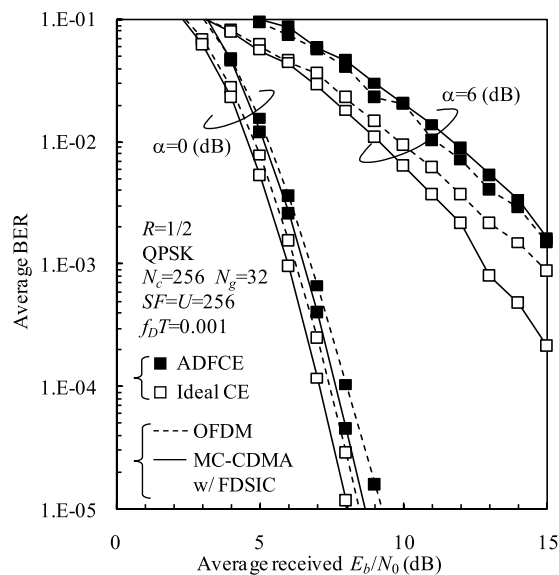
**Table 1** Simulation condition.

Turbo coding	Encoder	(13, 15) RSC encoders
	Decoder	Log-MAP decoding
	Coding rate	$R=1/2, 3/4, 8/9$
	Channel interleaver	Block
Transmitter	Data modulation	QPSK, 16QAM
	No. of FFT points	$N_c=256$
	GI length	$N_g=32$
	Spreading sequence	Product of Walsh sequence and long PN sequence
	Spreading factor	$SF=1, 256$
	Code multiplex order	$U=SF$
Channel	Fading model	Frequency-selective block Rayleigh fading
	Power delay profile	Exponential power delay profile
	Decay factor	$\alpha=0, 6\text{dB}$
	Time delay	$\tau_l=l, l=0\sim 15$
Receiver	Channel estimation	ADFCE
	No. of iterations	$l=8$

The forgetting factor  $\lambda$  of RLS algorithm is set to  $\lambda = 1.0$ , since we are assuming the fading statistical property remains the same over one packet in this paper (note that  $\lambda$  should be less than 1.0 when the fading statistical property changes during one packet). The number of iterations in FDSIC is assumed to be 8. For fair comparison, the number of iterations in turbo decoding is set to 8 for MC-CDMA without FDSIC (i.e., no ICI cancellation) and OFDM. As  $SF$  increases, the residual ICI gets stronger since we assume  $U = SF$ . However, when FDSIC is used, the residual ICI can be suppressed. Therefore, we set  $SF$  to  $SF = N_c = 256$  to obtain the largest frequency diversity gain.

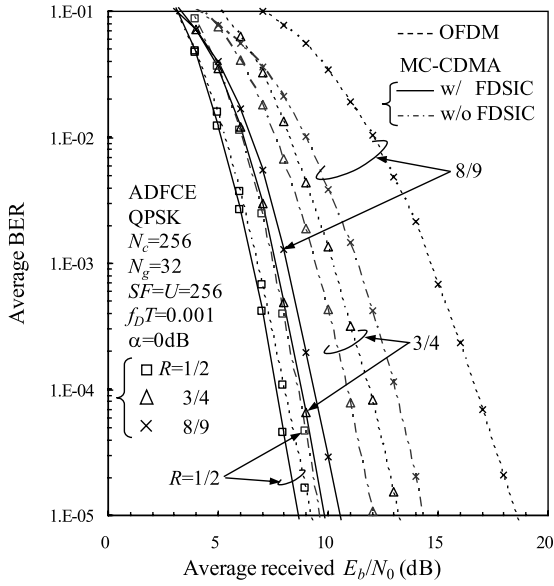
#### 4.1 Average BER Performance

Figure 3 plots the average BER performance of MC-CDMA with FDSIC using ADFCE as a function of  $f_D T$  when the average received bit energy-to-AWGN power spectrum density ratio  $E_b/N_0 = 7$  dB and coding rate  $R = 1/2$ .  $f_D T$  is the normalized maximum Doppler frequency, where  $f_D$  is the maximum Doppler frequency and  $T = (N_c + N_g)T_c$  ( $T_c$  is the FFT sampling interval) is the length of the MC-CDMA symbol. As in [19], assuming 100 MHz signal bandwidth at 5 GHz band as an example of the next generation systems,  $f_D T = 0.001$  corresponds to a moving speed of 75 km/h for a symbol rate of 100 Msps. For comparison, the BER performances of non-adaptive DFCE ( $\beta = 0.0 \sim 1.0$ ) and linear-interpolation CE [18] are also plotted. DFCE with  $\beta = 0$  corresponds to PACE. When  $f_D T \leq 0.001$  (in a slow fading channel), the BERs of DFCE are almost constant except  $\beta = 0$  and smaller BER is achieved as  $\beta$  decreases. This is because the noise is reduced more. However, the BER starts to increase when  $f_D T$  increases beyond 0.002 since the tracking ability against fading starts to deteriorate. Therefore, as  $f_D T$  increases, the optimum value of  $\beta$  becomes bigger. It is seen from Fig. 3 that ADFCE which can adapt the

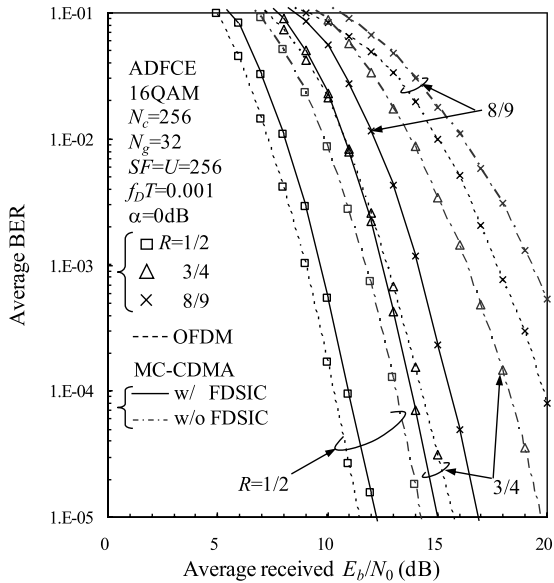
**Fig. 3** Impact of Doppler spread.**Fig. 4** Impact of channel estimation error on BER.

value of  $\beta$  to the optimum value gives the best BER performance for all  $f_D T$  values and provides better BER performance than linear-interpolation CE. This is the reason why we use ADFCE in the following simulation.

Figure 4 shows the average BER performances of MC-CDMA with FDSIC and OFDM as a function of  $E_b/N_0$  with the decay factor  $\alpha$  as a parameter. Note that ADFCE is used in both MC-CDMA with FDSIC and OFDM. For comparison, the BER performances with ideal CE are also plotted. When  $\alpha = 0$  dB, in OFDM, the  $E_b/N_0$  degradation for BER =  $10^{-5}$  from ideal CE is about 0.3 dB (about 0.07 dB is due to the pilot insertion). On the other hand, in MC-CDMA with FDSIC, it is about 0.4 dB (about 0.07 dB is due to the pilot insertion). The difference of the  $E_b/N_0$  degradation be-



(a) QPSK



(b) 16QAM

Fig. 5 BER performance comparison between MC-CDMA and OFDM.

tween MC-CDMA with FDSIC and OFDM is only 0.1 dB. On the other hand, when  $\alpha = 6$  dB, the frequency-selectivity is weak and the frequency diversity gain reduces. As a consequence, the BER performance of MC-CDMA with FDSIC degrades. Also the OFDM with ADFCE shows the same behavior due to the reduced coding gain.

Figure 5 plots the average BER performances of MC-CDMA with/without FDSIC and OFDM as a function of  $E_b/N_0$  with  $R$  as a parameter. ADFCE is used for both MC-CDMA and OFDM. It is seen from Fig. 5 that the introduction of FDSIC to MC-CDMA can significantly improve the BER performance, since the FDSIC can sufficiently suppress the residual ICI while achieving frequency diversity gain. For QPSK, MC-CDMA with FDSIC can reduce the

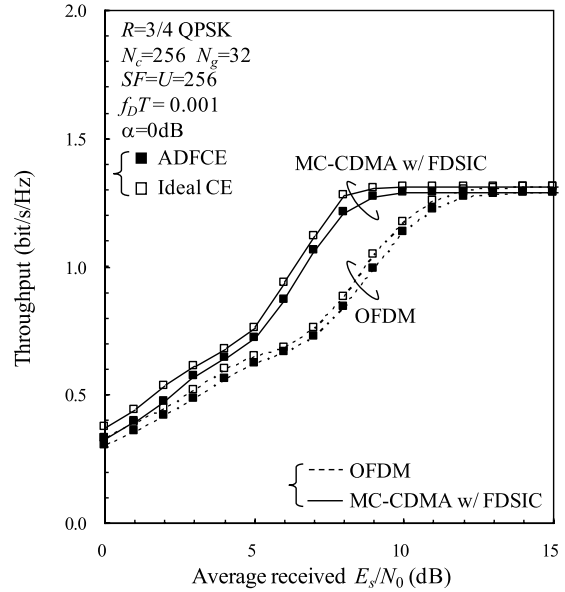


Fig. 6 Impact of channel estimation error on throughput.

required  $E_b/N_0$  for BER =  $10^{-5}$  by about 0.6 dB, 3.4 dB and 8.0 dB, compared to OFDM, when  $R = 1/2, 3/4$  and  $8/9$ , respectively. On the other hand, for the case of 16QAM, a slight BER performance degradation is seen for  $R = 1/2$ . This is because the Euclidian distance between neighboring symbols is shorter for 16QAM and therefore, the residual ICI produces the decision error more likely than for QPSK. However, when  $R = 3/4$  and  $8/9$ , MC-CDMA with FDSIC achieves better BER performance than OFDM (note that the same conclusion can be drawn in the ideal CE case).

#### 4.2 Throughput Performance

Figure 6 plots the throughput performances of MC-CDMA with FDSIC and OFDM as a function of the average received signal energy-to-AWGN power spectrum density ratio  $E_s/N_0$  assuming ADFCE. For comparison, the throughput performances with ideal CE are also plotted. In OFDM, the  $E_s/N_0$  degradation from ideal CE at the throughput of 1.0 bit/s/Hz is about 0.3 dB (about 0.07 dB is due to the pilot insertion), while in MC-CDMA with FDSIC, it is about 0.4 dB (about 0.07 dB is due to the pilot insertion). However, MC-CDMA with FDSIC provides better throughput performance than OFDM because of the larger frequency diversity gain.

Figure 7 plots the throughput comparison between MC-CDMA with/without FDSIC and OFDM as a function of  $E_s/N_0$  with  $R$  as a parameter. It can be seen from Fig. 7 that MC-CDMA with FDSIC gives better throughput performance than MC-CDMA without FDSIC and OFDM. When  $R = 1/2$ , MC-CDMA with FDSIC provides almost the same throughput performance as OFDM for QPSK and 16QAM. When  $R = 3/4$  and  $8/9$ , MC-CDMA with FDSIC can reduce the required  $E_s/N_0$  to achieve the same throughput compared to OFDM as much as 2.4(5.8)dB for QPSK and

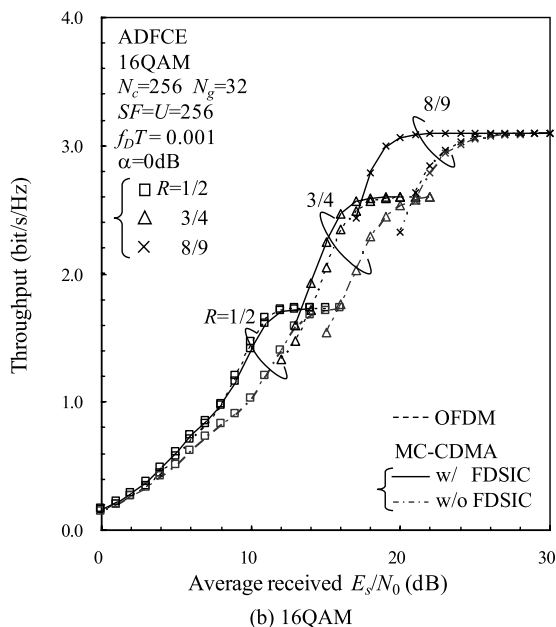
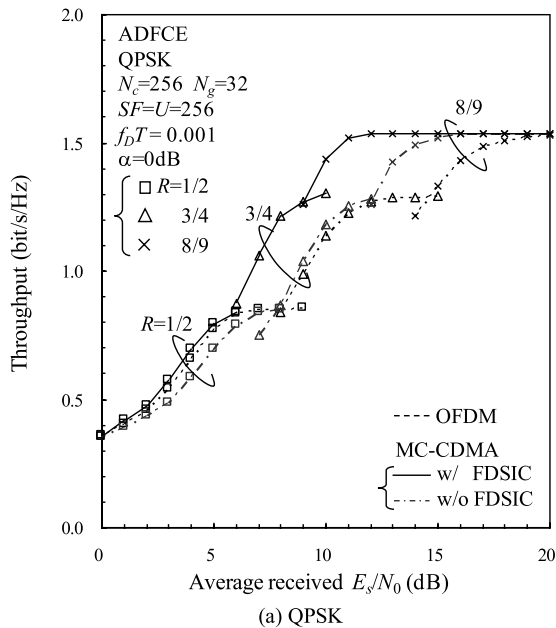


Fig. 7 Throughput comparison between MC-CDMA and OFDM.

0.6(3.6) dB for 16QAM at  $R = 3/4(8/9)$ , respectively.

5. Conclusion

In this paper, we proposed an adaptive channel estimation using time-multiplexed pilot and decision-feedback and evaluated by computer simulation the average BER and throughput performances of MC-CDMA with FDSIC in a frequency-selective Rayleigh fading channel. We have shown that the use of FDSIC significantly improves the performance and MC-CDMA with FDSIC provides almost the same BER/throughput as OFDM when  $R = 1/2$  and provides better BER/throughput than OFDM when  $R = 3/4$ ,

8/9. The performance degradation of MC-CDMA with FD-SIC from ideal CE is less than 0.5 dB.

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