PAPER Throughput Performance of MC-CDMA HARQ Using ICI Cancellation

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SUMMARY Multi-carrier code division multiple access (MC-CDMA) is a promising wireless access technique for the next generation mobile communications systems, in which broadband packet data services will dominate. Hybrid automatic repeat request (HARQ) is an indispensable error control technique for high quality packet data transmission. The HARQ throughput performance of multi-code MC-CDMA degrades due to the presence of residual inter-code interference (ICI) after frequency-domain equalization (FDE). To reduce the residual ICI and improve the throughput performance, a frequency-domain soft interference cancellation (FDSIC) technique can be applied. An important issue is the generation of accurate residual ICI replica for FDSIC. In this paper, low-density parity-check coded (LDPC-coded) MC-CDMA HARQ is considered. We generate the residual ICI replica from a-posteriori log-likelihood ratio (LLR) of LDPC decoder output and evaluate, by computer simulation, the throughput performance in a frequency-selective Rayleigh fading channel. We show that if the residual ICI is removed, MC-CDMA can provide a throughput performance superior to orthogonal frequency division multiplexing (OFDM). key words: MC-CDMA, HARQ, ICI cancellation, LDPC decoding

1. Introduction

For the next generation mobile communications systems, in which broadband packet data services will dominate, high speed packet transmission techniques are required [1]. Hybrid automatic repeat request (HARQ), which is a combination of ARQ and error-correcting coding, is an indispensable error control technique for high quality packet data transmission [2], [3]. As an error-correcting code, low-density parity-check (LDPC) coding [4]–[6] can be used.

Wireless channels for a high speed data transmission become severely frequency-selective [7]. Multi-carrier code division multiple access (MC-CDMA) is a promising wireless access technique [8], [9] because of its high flexibility in variable rate data transmission (this can be achieved by code-multiplexing) and user multiplexing. 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 intercode interference (ICI) is produced. The minimum mean square error frequency-domain equalization (MMSE-FDE) can suppress ICI while minimizing the noise enhancement [8], [9]. However, the presence of residual ICI after MMSE-FDE degrades achievable bit error rate (BER) performance in a severe frequency-selective fading channel. To reduce the residual ICI and improve the BER performance, a frequency-domain soft interference cancellation (FDSIC) technique was proposed in [10]. The residual ICI replica is generated and is subtracted from the MMSE-FDE output. In FDSIC, MMSE-FDE and ICI cancellation is repeated a sufficient number of times.

In this paper, we apply FDSIC to LDPC-coded MC-CDMA HARQ in order to reduce the residual ICI and improve the throughput performance. The single-user case is considered. An important issue is the generation of accurate residual ICI replica for FDSIC. We generate the residual ICI replica from *a-posteriori* log-likelihood ratio (LLR) [11] of LDPC decoder output and evaluate, by computer simulation, the throughput performance of LDPC-coded MC-CDMA HARQ using FDSIC in a frequency-selective Rayleigh fading channel. We will show that if the residual ICI is removed, MC-CDMA can provide a throughput performance superior to orthogonal frequency division multiplexing (OFDM) [12].

The remainder of this paper is organized as follows. First, we present the LDPC-coded MC-CDMA HARQ system model in Sect. 2. Then, we introduce FDSIC and present the residual ICI replica generation in Sect. 3. In Sect. 4, we show the computer simulation results on the throughput performance and compare the HARQ throughput performances using MC-CDMA and OFDM. This paper is concluded in Sect. 5.

2. MC-CDMA HARQ System Model

In general, HARQ can be classified into two types: type I and type II. In this paper, we consider HARQ type II and assume an S-PX strategy [3], as illustrated in Fig. 1, which is a family of the well-known incremental redundancy (IR) strategy. After an information bit sequence of length K is encoded, a systematic bit sequence of length K bits and a parity bit sequence of length $(R^{-1} - 1)K$ bits are generated, where R is the coding rate. The parity bit sequence is divided into X different parity bit blocks of length $(R^{-1} - 1)K/(X)$. The systematic bit sequence is transmitted first. If any error is detected in the received systematic bit sequence, the receiver requests the transmission of the first parity bit block. LDPC decoding is carried out using the systematic bit sequence and the first parity bit block. If any error is detected, the receiver requests the transmission of second parity bit block. This is repeated until no error is detected. The above

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signal is transmitted over a frequency-selective fading channel.

At the receiver, the MC-CDMA signal is received by N_r antennas. The received signal block $\{r_{n_r}^{tr}(t); t = -N_g \sim N_c - 1\}$ at the *tr*th ($tr = 0 \sim Q - 1$) reception of the same packet data on the n_r th ($n_r = 0 \sim N_r - 1$) antenna can be expressed as [3]

$$r_{n_r}^{tr}(t) = \sum_{l=0}^{L-1} h_{n_r,l}^{tr} s\left(t - \tau_l\right) + \eta_{n_r}^{tr}(t) , \qquad (3)$$

where $h_{n_r,l}^{tr}$ and τ_l are the complex-valued path gain and time delay of the *l*th $(l = 0 \sim L - 1)$ path, respectively, with $\sum_{l=0}^{L-1} E\left[\left|h_{n_r,l}^{tr}\right|^2\right] = 1$, and $\eta_{n_r}^{tr}(t)$ is a zero-mean complex Gaussian process having variance $2N_0/T_c$ with N_0 being the single sided power spectrum density of additive white Gaussian noise (AWGN) and T_c being the IFFT/FFT sampling interval. After removing the N_g -sample GI, the N_c -point FFT is applied to decompose the received MC-CDMA signal into N_c subcarrier components { $R_{n_r}^{tr}(k)$; $k = 0 \sim N_c - 1$ }, where $R_{n_r}^{tr}(k)$ is represented as

$$R_{n_r}^{tr}(k) = \frac{1}{N_c} \sum_{t=0}^{N_c-1} r_{n_r}^{tr}(t) \exp\left(-j2\pi k \frac{t}{N_c}\right)$$

= $H_{n_r}^{tr}(k) S(k) + \Pi_{n_r}^{tr}(k).$ (4)

 $H_{n_r}^{tr}(k)$ and $\Pi_{n_r}^{tr}(k)$ are respectively the channel gain and the noise at the *k*th subcarrier frequency and are given by

$$\begin{cases} H_{n_r}^{tr}(k) = \sum_{l=0}^{L-1} h_{n_r,l}^{tr} \exp\left(-j2\pi k \frac{\tau_l}{N_c}\right) \\ \Pi_{n_r}^{tr}(k) = \frac{1}{N_c} \sum_{t=0}^{N_c-1} \eta_{n_r}^{tr}(t) \exp\left(-j2\pi k \frac{t}{N_c}\right). \end{cases}$$
(5)

 $R_{n_r}^{tr}(k)$ and $H_{n_r}^{tr}(k)$ are stored in the receiver buffer for performing packet combining [3] and FDSIC, followed by LDPC decoding. FDSIC is explained in Sect. 3. For HARQ type II, since only systematic bit sequence is received at the first transmission, LDPC decoding can't be carried out. After FDSIC, error detection is performed. If any error is detected, retransmission is requested; otherwise, a new information sequence is transmitted. After the second retransmission onwards, the joint use of FDSIC and LDPC decoding are carried out.

3. Introduction of FDSIC

A combination of FDSIC and LDPC decoding is illustrated in Fig. 3. A series of steps $A \sim D$ is iterated N_{max} times. Step A performs MMSE-FDE and ICI cancellation. Despreading and LLR computation are performed in step B, followed by step C (LDPC decoding based on the sumproduct algorithm [6], in which *a-posteriori* LLR is computed once). Step D generates the residual ICI replica from *a-posteriori* LLR of the LDPC decoder output. Note that the updating of *a-posteriori* LLR is done only once in each



Fig. 1 HARQ type II S-PX.



Fig. 2 Transmitter/receiver structure.

described HARQ type II S-PX is only possible if systematic error correction coding like LDPC coding or turbo coding is used.

The transmitter/receiver structure for LDPC-coded MC-CDMA HARQ using FDSIC is illustrated in Fig. 2. Below, we assume that the same data packet (consisting of the systematic and parity bit sequences of total length KR^{-1} bits) has been transmitted Q times. The data-modulated symbol sequence is serial-to-parallel (S/P) converted to C symbol streams { $d_c(n)$; $n = 0 \sim N_c/SF - 1$ }, $c = 0 \sim C - 1$, and each symbol in C streams is spread by an orthogonal code { $c_c^{oc}(k)$; $k = 0 \sim SF - 1$ }, where SF is the spreading factor and N_c is the number of subcarriers. C streams are added and multiplied by a scrambling code { $c_c^{scr}(k)$; $k = 0 \sim N_c - 1$ }. Using N_c -point inverse fast Fourier transform (IFFT), the resultant sequence is transformed into an MC-CDMA signal {s(t); $t = 0 \sim N_c - 1$ }, given as

$$s(t) = \sum_{k=0}^{N_c - 1} S(k) \exp\left(j2\pi k \frac{t}{N_c}\right),$$
(1)

where S(k) denotes the *k*th subcarrier component and is expressed as

$$S(k) = \sqrt{\frac{2P}{SF}} \sum_{c=0}^{C-1} d_c \left(\left\lfloor \frac{k}{SF} \right\rfloor \right) c_c^{oc} (k \mod SF) c^{scr}(k)$$
(2)

with *P* being the transmit power per symbol and $\lfloor k \rfloor$ denoting the largest integer smaller than or equal to *k*. After inserting an N_q -sample guard interval (GI), the MC-CDMA



Fig. 3 FDSIC and LDPC decoding.

iteration (step C). The reason for this is explained below. In LDPC decoder, *a-posteriori* LLR is updated by using the soft decision result (obtained after FDSIC/de-spreading) and *a-priori* LLR [11]. The *a-priori* LLR, obtained after many times of updating of *a-posteriori* LLR, tends to have large absolute value. If the residual ICI replica is generated from the latest *a-posteriori* LLR and then, FDSIC is carried out, the quality of soft decision result may improve. This improved soft decision result is input to the LDPC decoder for the updating of *a-posteriori* LLR. However, since the *a-priori* LLR stored in LDPC decoder has much larger absolute value than the soft decision result, FDSIC cannot make a sufficient contribution to the updating of *a-posteriori* LLR. Therefore, the updating of *a-posteriori* LLR is done only once in each iteration.

After N_{max} iterations, the error detection is made.

A. MMSE-FDE and ICI Cancellation

Joint MMSE-FDE and packet combining (the same data packet has been transmitted Q times) for the *i*th iteration $(i = 0 \sim N_{\text{max}} - 1)$ can be expressed as

$$\tilde{R}_{i}^{Q-1}(k) = \sum_{tr=0}^{Q-1} \sum_{n_{r}=0}^{N_{r}-1} w_{n_{r},i}^{tr}(k) R_{n_{r}}^{tr}(k),$$
(6)

where $w_{n_{r,i}}^{tr}(k)$ is the MMSE-FDE weight and can be derived, from [10], as

where $E_s = PT_c N_c$ is the average received symbol energy (note that we assume $\sum_{l=0}^{L-1} E\left[\left|h_{n_r,l}^{tr}\right|^2\right] = 1$) and $\rho_{c,i-1}^{Q-1}(n)$ represents the contribution from the residual ICI, given in [10], as

$$\rho_{c,i-1}^{Q-1}(n) = \begin{cases} 1, \ i = 0\\ \left| \bar{d}_{c,i-1}^{Q-1}(n) \right|^2 - \left| \tilde{d}_{c,i-1}^{Q-1}(n) \right|^2, \ i \ge 1 \end{cases}, \tag{8}$$

where $\tilde{d}_{c,i-1}^{Q-1}(n)$ is the soft symbol replica derived in step D and $\bar{d}_{c,i-1}^{Q-1}(n)$ is the hard decision result obtained from $\tilde{d}_{c,i-1}^{Q-1}(n)$. $\rho_{c,i-1}^{Q-1}(n)$ is the expectation of squared error $E\left[\left|d_{c}(n)-\tilde{d}_{c,i-1}^{Q-1}(n)\right|^{2}\right]$ between the transmit symbol $d_{c}(n)$ and $\tilde{d}_{c,i-1}^{Q-1}(n)$. Remembering that $\tilde{d}_{c,i-1}^{Q-1}(n)$ is the expectation of $d_{c}(n)$, we obtain $\rho_{c,i-1}^{Q-1}(n) = E\left[\left|d_{c}(n)\right|^{2}\right] - \left|\tilde{d}_{c,i-1}^{Q-1}(n)\right|^{2}$. Since $d_{c}(n)$ is unknown, it is replaced by its hard decision $\bar{d}_{c,i-1}^{Q-1}(n)$, giving Eq. (8).

Equation (6) can be written as

$$\tilde{R}_{i}^{Q-1}(k) = \tilde{H}_{i}^{Q-1}(k) S(k) + I_{i}^{Q-1}(k) + N_{i}^{Q-1}(k), \quad (9)$$

where

$$\tilde{H}_{i}^{Q-1}(k) = \sum_{tr=0}^{Q-1} \sum_{n_{r}=0}^{N_{r}-1} w_{n_{r},i}^{tr}(k) H_{n_{r}}^{tr}(k).$$
(10)

In Eq. (9), the first term is the desired signal component and the second term is the residual ICI component, which should be removed for improving the throughput performance. The third term is the noise component due to AWGN. After joint MMSE-FDE and packet combining, ICI cancellation is carried out as [10]

$$\tilde{S}_{i}^{Q-1}(k) = \tilde{R}_{i}^{Q-1}(k) - \tilde{I}_{i}^{Q-1}(k), \qquad (11)$$

where $\tilde{I}_i^{Q-1}(k)$ is the residual ICI replica generated from *a*-posteriori LLR of LDPC decoder output. $\tilde{I}_i^{Q-1}(k)$ is derived in step D.

B. LLR Computation

After performing ICI cancellation, P/S conversion and despreading are done to obtain the following soft decision result

$$\hat{d}_{c,i}^{Q-1}(n) = \frac{1}{SF} \sum_{k=nSF}^{(n+1)SF-1} \tilde{S}_{i}^{Q-1}(k) \{c_{c}^{oc}(k \mod SF) c^{scr}(k)\}^{*}$$
$$= \sqrt{\frac{2P}{SF}} \hat{H}_{i}^{Q-1}(n) d_{c}(n) + \mu_{i,\text{residual ICI}}^{Q-1}(n)$$
$$+ \mu_{i,\text{noise}}^{Q-1}(n), \qquad (12)$$

where

$$\hat{H}_{i}^{Q-1}(n) = \frac{1}{SF} \sum_{k=nSF}^{(n+1)SF-1} \tilde{H}_{i}^{Q-1}(k).$$
(13)

In Eq. (12), the first term represents the desired signal component and the second term and third term are the residual ICI and noise, respectively. Approximating the residual ICI plus noise as a new Gaussian variable, the LLR $\lambda_{c,i}^{Q-1}(n,m)$, $m = 0 \sim (\log_2 M) - 1$, of the *m*th bit of the symbol $d_c(n)$, where *M* is the modulation level, is computed using [3]

$$\lambda_{c,i}^{Q-1}(n,m) = \frac{1}{2\left\{\sigma_{c,i}^{Q-1}(n)\right\}^{2}} \left(\begin{vmatrix} \hat{d}_{c,i}^{Q-1}(n) - \sqrt{\frac{2P}{SF}} \hat{H}_{i}^{Q-1}(n) d_{b(m)=0}^{\min} \end{vmatrix}^{2} \\ - \left| \hat{d}_{c,i}^{Q-1}(n) - \sqrt{\frac{2P}{SF}} \hat{H}_{i}^{Q-1}(n) d_{b(m)=1}^{\min} \right|^{2} \end{vmatrix},$$
(14)

where $d_{b(m)=0}^{\min}$ (or $d_{b(m)=1}^{\min}$) is the most probable symbol whose *m*th bit b(m) is 0 (or 1), for which the Euclidean distance from $\hat{d}_{c,i}^{Q-1}(n)$ is minimum. $2\left\{\sigma_{c,i}^{Q-1}(n)\right\}^2$ is the variance of the residual ICI plus noise and can be found, from [10], as

$$2\left\{\sigma_{c,i}^{Q-1}(n)\right\}^{2} = \frac{1}{SF^{2}} \frac{2E_{s}}{T_{c}} \\ \times \left[\sum_{\substack{c'=0\\c'\neq c}}^{C-1} \rho_{c',i-1}^{Q-1}(n) \left\{\frac{1}{SF} \sum_{k=nSF}^{N-1} \left|\tilde{H}_{i}^{Q-1}(k)\right|^{2} - \left|\hat{H}_{i}^{Q-1}(n)\right|^{2}\right\} \right] \\ + \left(\frac{1}{SF} \frac{E_{s}}{N_{0}}\right)^{-1} \sum_{tr=0}^{Q-1} \sum_{n_{r}=0}^{N_{r}-1} \sum_{k=nSF}^{(n+1)SF-1} \left|w_{n_{r},i}^{tr}(k)\right|^{2} \right].$$
(15)

C. LDPC Decoding

Let $A = \{A^{xy}; x = 0 \sim (1 - R)K/R - 1, y = 0 \sim K/R - 1\}$ be a $((R^{-1} - 1)K \times R^{-1}K)$ parity-check matrix [6], where $(R^{-1} - 1)K$ and $R^{-1}K$ are the length of the parity bit sequence and that of LDPC-coded bit sequence, respectively. At first, for each (x, y) satisfying $A^{xy}=1$, the extrinsic information $\alpha_i^{xy,Q-1}$ is computed using [6]

$$\alpha_{i}^{xy,Q-1} = \left(\prod_{\substack{y' \in [A^{xy'}=1] \\ y' \neq y}} \operatorname{sign}\left(\lambda_{i}^{y',Q-1} + \beta_{i}^{xy',Q-1}\right)\right) \\ \times f\left(\sum_{\substack{y' \in [A^{xy'}=1] \\ y' \neq y}} f\left(\left|\lambda_{i}^{y',Q-1} + \beta_{i}^{xy',Q-1}\right|\right)\right), \quad (16)$$

where $\lambda_i^{y,Q-1}$ is the LLR of the *y*th bit of LDPC-coded bit sequence obtained from $\{\lambda_{c,i}^{Q-1}(n,m); m = 0 \sim \log_2 M - 1\}, \beta_i^{xy,Q-1}$ is *a-priori* LLR [11] with $\beta_0^{xy,Q-1} = 0$, and the functions, sign(*x*) and *f*(*x*), are defined as

$$\begin{cases} \operatorname{sign}(x) \equiv \begin{cases} 1, \ x \ge 0\\ -1, \ \text{otherwise} \end{cases} \\ f(x) \equiv \ln \frac{\exp(x) + 1}{\exp(x) - 1} \end{cases}$$
(17)

After obtaining $\alpha_i^{xy,Q-1}, \beta_i^{xy,Q-1}$ is updated as [6]

$$\beta_{i+1}^{xy,Q-1} = \sum_{\substack{x' \in \{A^{x'y}=1\}\\x' \neq x}} \alpha_i^{x'y,Q-1}$$
(18)

for the (i+1)th iteration. The decoder outputs a sequence of *a-posteriori* LLR's { $\Lambda_i^{y,Q-1}$ }, where $\Lambda_i^{y,Q-1}$ is computed as

[6]

$$\Lambda_{i}^{y,Q-1} = \lambda_{i}^{y,Q-1} + \sum_{x' \in \{A^{x'y}=1\}} \alpha_{i}^{x'y,Q-1},$$
(19)

which is used to generate the residual ICI replica in step D.

D. Residual ICI Replica Generation

First, the soft symbol replica $\tilde{d}_{c,i}^{Q-1}(n)$ is computed using the LLR as

$$\tilde{d}_{c,i}^{Q-1}(n) = \begin{cases} g\left(\left\{\lambda_{c,i}^{Q-1}(n,m); m = 0 \sim \log_2 M - 1\right\}\right) \\ \text{for the first transmission} \\ g\left(\left\{\Lambda_{c,i}^{Q-1}(n,m); m = 0 \sim \log_2 M - 1\right\}\right), \\ \text{for otherwise} \end{cases}$$
(20)

where $\lambda_{c,i}^{Q-1}(n,m)$ is given by Eq. (14) and $\Lambda_{c,i}^{Q-1}(n,m)$ is *a*posteriori LLR of the *m*th bit belonging to the symbol $d_c(n)$. $g(\lambda(n,0), \lambda(n,1), \dots, \lambda(n, \log_2 M - 1))$ is defined as [10]

$$g\left(\lambda(n,0), \lambda(n,1), \dots, \lambda(n,\log_2 M - 1)\right) \\ = \begin{cases} \frac{1}{\sqrt{2}} \tanh\left(\frac{\lambda(n,0)}{2}\right) + j\frac{1}{\sqrt{2}} \tanh\left(\frac{\lambda(n,1)}{2}\right) \\ \text{for QPSK } (M = 4) \\ \frac{1}{\sqrt{10}} \tanh\left(\frac{\lambda(n,0)}{2}\right) \left\{2 - \tanh\left(\frac{\lambda(n,1)}{2}\right)\right\} \\ + j\frac{1}{\sqrt{10}} \tanh\left(\frac{\lambda(n,2)}{2}\right) \left\{2 - \tanh\left(\frac{\lambda(n,3)}{2}\right)\right\} \end{cases}.$$
(21)
for 16QAM $(M = 16)$

The residual ICI replica $\tilde{I}_i^{Q-1}(k)$ in Eq. (11) is generated as

$$\tilde{I}_{i}^{Q-1}(k) = M_{i}^{Q-1}(k) \,\hat{S}_{i}^{Q-1}(k) \,, \tag{22}$$

where $M_i^{Q-1}(k)$ is the cancellation weight given as

$$M_i^{Q-1}(k) = \begin{cases} 0, \ i = 0\\ \tilde{H}_i^{Q-1}(k) - \hat{H}_i^{Q-1}\left(\left\lfloor \frac{k}{SF} \right\rfloor\right), \ i \ge 1 \end{cases}$$
(23)

and $\hat{S}_i^{Q-1}(k)$ is the MC-CDMA signal replica given as

$$\hat{S}_{i}^{Q-1}(k) = \sqrt{\frac{2P}{SF}} \sum_{c=0}^{C-1} \tilde{d}_{c,i-1}^{Q-1}\left(\left\lfloor \frac{k}{SF} \right\rfloor\right) c_{c}^{oc} (k \mod SF) c^{scr}(k) .$$
(24)

4. Simulation Results

The simulation conditions are summarized in Table 1. The single-user transmission with full code-multiplexed (*C*=*SF*) MC-CDMA using N_c =256 is assumed. HARQ type II S-P2 is considered (*X*=2). For S-P2, as explained in Sect. 2, the parity bit sequence of length ($R^{-1} - 1$)*K* is divided into two different blocks of length ($R^{-1} - 1$)*K*/2. Each parity bit block is assumed to be transmitted on request over one MC-CDMA signal period (i.e., N_c subcarriers). The information

Table 1 Simulation condition.		
Channel code	LDPC code (<i>R</i> =1/3, 1/2, 2/3)	
Data modulation	QPSK, 16QAM	
MC-CDMA	No. FFT points	N _c =256
	GI	$N_g=32$
	Spreading factor	SF=1, 256
	Code multiplex order	C=SF
HARQ scheme	Type II S-P2	
Channel model	L=16-path block Rayleigh fading	
	$\tau_l = l \ (l = 0 \sim L - 1), f_D T = 0.001$	

bit length K depends on coding rate R and data modulation level M. Therefore, in the case of QPSK (M=4), we have K = 512 for R=1/3, K=1024 for R=1/2, and K=2048 for R=2/3(this means that the information bit sequence is transmitted over one, two, and four MC-CDMA signal periods when R=1/3, 1/2, and 2/3, respectively). The symbol energy E_s is kept constant for every (re)transmission. We assumed a frequency-selective block Rayleigh-fading channel having a T_c -spaced L-path uniform power delay profile and a normalized maximum Doppler frequency $f_D T = 0.001$, where f_D is the maximum Doppler frequency and $T = (N_c + N_a)T_c$ is the block length of MC-CDMA signal including GI. In this paper, ideal channel estimation, ideal error detection, and no transmission error of ACK/NACK command are assumed. For fair comparison, a-posteriori LLR is updated $N_{\rm max}$ times, as in conventional LDPC decoder, also for MC-CDMA without FDSIC and OFDM. Note that since we assume the full code-multiplexed MC-CDMA (C=SF), the same transmission rate is achieved as OFDM.

As SF increases, the larger frequency diversity gain can be achieved, but the residual ICI gets stronger since we are assuming SF=C. Without FDSIC, the throughput improvement obtained by the frequency diversity gain is offset by the residual ICI. However, when FDSIC is used, the residual ICI can be mitigated and the throughput can be improved due to frequency diversity gain. Since the largest frequency diversity gain can be obtained when $SF=N_c$, we set SF to SF=256 for the throughput performance evaluation.

The throughput performance of LDPC-coded MC-CDMA HARQ using FDSIC is plotted as a function of the average received signal energy per symbol-to-the AWGN power spectrum density ratio E_s/N_0 in Fig. 4, when coding rate R=1/3, N_r (the number of receive antenna) = 1, N_{max} (the number of iterations)=30 and SF=C=256. It is seen in Fig.4 that the introduction of FDSIC significantly improves the throughput performance. When FD-SIC is used, the E_s/N_0 reduction from the without FD-SIC case at the throughput of 0.8 bps/Hz (1.6 bps/Hz) is as much as 1.5 dB (4 dB) for QPSK. 16QAM has shorter Euclidean distance and is more sensitive to the residual ICI. This suggests that the use of FDSIC can be more effective than for QPSK. Therefore, for 16QAM, the E_s/N_0 reduction is bigger than for QPSK and an about 2 dB (5 dB) reduction is seen from Fig. 4(b) at the throughput of 1.5 bps/Hz (3 bps/Hz). The throughput performance gets closer to perfect FDSIC case (the residual ICI is suppressed completely,



i.e., $\tilde{I}_i^{Q-1}(k) = I_i^{Q-1}(k)$) by about 1 dB for QPSK and 2 dB for 16QAM. It is also seen from Fig. 4 that the throughput performance of MC-CDMA using FDSIC is higher than using OFDM. About 1 dB (12 dB) E_s/N_0 reduction from the OFDM case is achieved at the throughput of 0.8 bps/Hz (1.6 bps/Hz) for QPSK (see Fig. 4(a)). This is because the frequency diversity gain can be achieved through frequencydomain de-spreading in MC-CDMA. On the other hand, in OFDM which corresponds to MC-CDMA with SF=1, since each data symbol is transmitted on one subcarrier, the frequency diversity gain cannot be achieved. Furthermore, in HARQ type II S-P2, since only the uncoded information bit sequence is transmitted at the first transmission, both MC-



Impact of coding rate.

CDMA and OFDM can not achieve the coding gain; however, MC-CDMA can obtain the frequency diversity gain. After retransmission, OFDM can achieve the coding gain while MC-CDMA can achieve not only the coding gain but also the frequency diversity gain. As a consequence, MC-CDMA with FDSIC can provide better throughput performance compared to OFDM.

Figure 5 shows the impact of the coding rate R on the throughput performance. It is seen that MC-CDMA using FDSIC can achieve higher throughput performance than OFDM irrespective of R. The E_s/N_0 reduction from the OFDM case is almost the same irrespective of R and is about 1 dB at the throughput of 0.8 bps/Hz.

Figure 6 shows the impact of maximum number N_{max} of iterations on the throughput performance. Compared to OFDM, the use of FDSIC can reduce the required E_s/N_0 at the throughput of 0.8 bps/Hz by about 14 dB, 12.5 dB and 1 dB when N_{max} = 3, 10 and 30, respectively. Even if N_{max} is small, the throughput performance using FDSIC improves in the lower E_s/N_0 region. As was discussed earlier, this is because MC-CDMA can obtain the frequency diversity gain while OFDM cannot.

In Fig. 7, the throughput performance is plotted to show the effect of N_r =2-antenna diversity reception. It is seen that FDSIC can achieve almost the same throughput performance as the perfect FDSIC case. There are two possible



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reasons for this result. One is the improved accuracy of the residual ICI replica due to antenna diversity and the other is an assumption of perfect channel estimation. In practical systems, however, it is necessary to estimate the channel. Therefore, the throughput performance may degrade due to the channel estimation error (the throughput performance evaluation in the presence of the channel estimation error is left as an important future study).

5. Conclusion

The residual inter-code interference (ICI) degrades the throughput performance of MC-CDMA HARQ. In this paper, we applied the frequency-domain soft interference can-

cellation (FDSIC) technique to MC-CDMA HARQ to improve the throughput performance. We presented the generation of residual ICI replica, for FDSIC, from *a-posteriori* LLR of the LDPC decoder output. By computer simulation, we have shown that the use of FDSIC significantly improves the throughput performance in a frequency-selective Rayleigh fading channel. We have also shown that if the residual ICI is removed, MC-CDMA can provide a throughput performance superior to OFDM.

In this paper, we assumed ideal channel estimation. The impact of the channel estimation error on FDSIC is left for the future work.



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