LDPC-coded HARQ Throughput Performance of MC-CDMA using ICI Cancellation

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***Abstract*—Broadband packet data services are demanded in the next generation mobile communications systems. Multi-carrier CDMA (MC-CDMA) is considered to be a promising wireless technique. Hybrid ARQ (HARQ) is a powerful error control technique. It is known that the HARQ throughput performance of MC-CDMA degrades due to the residual inter-code interference (ICI) after frequency-domain equalization (FDE). The use of frequency-domain soft interference cancellation (FDSIC) technique can reduce the residual ICI and improve the throughput performance. An important technical problem is the generation of accurate residual ICI replica for FDSIC. In this paper, we consider low-density parity-check coded (LDPC-coded) MC-CDMA HARQ and generate the residual ICI replica from *a-posteriori* log-likelihood ratio (LLR) obtained by the LDPC decoder. We 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 better throughput performance than orthogonal frequency division multiplexing (OFDM).**

Keywords-component; MC-CDMA; ICI cancellation; LDPC coding

I. INTRODUCTION

For the next generation mobile communications systems, broadband packet transmission techniques are required [1]. Hybrid ARQ (HARQ), which is a combination of ARQ and error-correction coding, is a well-known error control technique [2]-[3]. Packet transmission using HARQ type II can achieve high throughput performance. As an error-correction coding, low-density parity-check (LDPC) coding has been gaining much attention [4]-[6].

Wireless channels for broadband data transmission become severely frequency-selective [7]. Multi-carrier CDMA (MC-CDMA) is known as a promising wireless access technique [8]-[9]. Although MC-CDMA has a high flexibility in variable rate data transmission and in user multiplexing, the presence of residual inter-code interference (ICI) after frequency-domain equalization (FDE) degrades the transmission performance in a severe frequency-selective fading channel. In [10], a frequency-domain soft interference cancellation (FDSIC) technique, which repeats minimum mean square error-FDE (MMSE-FDE) and ICI cancellation a sufficient number of times, was proposed to reduce the residual ICI and improve the bit error rate (BER) performance. In this paper, we apply FDSIC to LDPC-coded MC-CDMA HARQ in order to improve the throughput performance. An important technical problem is the generation method of accurate residual ICI replica for ICI cancellation. We generate the residual ICI replica from *a-posteriori* log-likelihood ratio (LLR) [11] obtained by the LDPC decoder. We 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 better throughput performance than orthogonal frequency division multiplexing (OFDM).

The remainder of this paper is organized as follows. First, we present the LDPC-coded MC-CDMA HARQ system model in Sect. II. Then, we present FDSIC and the residual ICI replica generation in Sect. III. In Sect. IV, we present the computer simulation results on the throughput performance and compare the HARQ throughput performances of LDPC-coded MC-CDMA using FDSIC and orthogonal frequency division multiplexing (OFDM) [12]. This paper is concluded in Sect. V.

II. LDPC-CODED MC-CDMA HARQ SYSTEM MODEL

In general, HARQ can be classified into three types [13]: HARQ type I, type II and type III. In this paper, we consider the HARQ type II S-P2 [3], as illustrated in Fig. 1. An information bit sequence of length K is encoded with coding rate R=1/3 and systematic bit sequence of length K and parity bit sequence of length $2K$ are generated. The parity bit sequence is divided into two different blocks of length K. The first packet to be transmitted is the systematic bit sequence. If any error is detected in the received packet, the receiver requests the transmission of the first parity bit block. Upon the reception of the first party bit block, LDPC decoding is carried out using the previously received 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 transmitter/receiver structure for LDPC-coded MC-CDMA HARQ using FDSIC is illustrated in Fig. 2. Below, we assume that the same packet (information and parity) has been transmitted Q times. The data-modulated symbol sequence is serial-to-parallel (S/P) converted to C symbol streams $\{d_c(n)$; c=0~C-1, n=0~$N_c/SF\}$ and each symbol in C streams is spread by an orthogonal code $\{c_c^{\mathrm{oc}}(k)$; k=0~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 $\{c^{\mathrm{scr}}(k)$; k=0~N_c-1}. Using N_c-point inverse fast Fourier transform (IFFT) with the sampling period T_c, the resultant sequence is transformed into an MC-CDMA signal $\{s(t)$; t=0~N_c-1}, given as

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

where $S(k)$ denotes the kth subcarrier component and is expressed as

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

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_g-sample guard interval (GI), the MC-CDMA signal is transmitted over a fading channel.

The received signal block $\{r^{tr}(t);\ t=-N_g\sim N_c-1\}$ at the trth ($tr=0\sim Q-1$) reception of the same packet can be expressed as [3]

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

where h_l^{tr} and τ_l are the complex-valued path gain and time delay of the lth ($l=0\sim L-1$) path, respectively, with $\sum_{l=0}^{L-1} E\left[\left|h_l^{tr}\right|^2\right]=1$ and $\eta^{tr}(t)$ is the additive white Gaussian noise (AWGN) with the single-sided power spectrum density N_0. 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 $\{R^{tr}(k);\ k=0\sim N_c-1\}$, where $R^{tr}(k)$ is represented as

$$\begin{aligned} R^{tr}(k)&=\frac{1}{N_c}\sum_{t=0}^{N_c-1} r^{tr}(t)\exp\left(-j2\pi k\frac{t}{N_c}\right) \\ &=H^{tr}(k)S(k)+\Pi^{tr}(k) \end{aligned} \quad .(4)$$

$H^{tr}(k)$ is the channel gain at the kth subcarrier frequency and is given by

$$H^{tr}(k)=\sum_{l=0}^{L-1} h_l^{tr}\exp\left(-j2\pi k\frac{\tau_l}{N_c}\right) \quad .(5)$$

$\Pi^{tr}(k)$ is the noise component. $R^{tr}(k)$ and $H^{tr}(k)$ are stored in the receiver buffer for performing packet combining [3] and FDSIC, followed by LDPC decoding. FDSIC is explained in Sect. III. In our HARQ type II, only systematic bit sequence is received at 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, LDPC decoding is carried out.

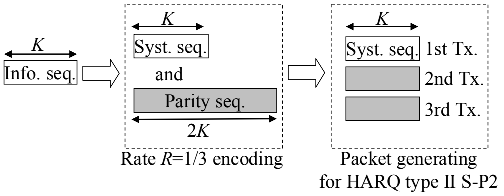
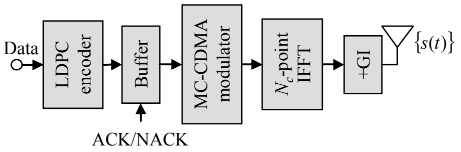
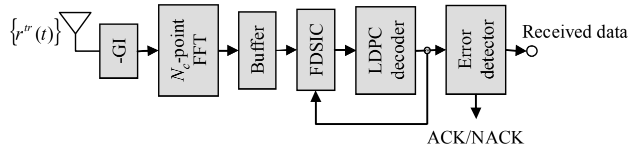


Figure 1. HARQ type II S-P2.



(a) Transmitter



(b) Receiver

Figure 2. Structure of transmitter and receiver.

III. FDSIC FOR LDPC-CODED MC-CDMA HARQ

A combination of FDSIC and LDPC decoding, which consists of steps of A~D, is illustrated in Fig. 3. Step A performs MMSE-FDE and ICI cancellation. De-spreading and LLR computation are performed in step B, followed by step C (LDPC decoding based on the sum-product algorithm [6], in which *a-posteriori* LLR is computed once). Step D generates the residual ICI replica from *a-posteriori* LLR. A series of steps A~D is iterated $N_{\max}$ times.

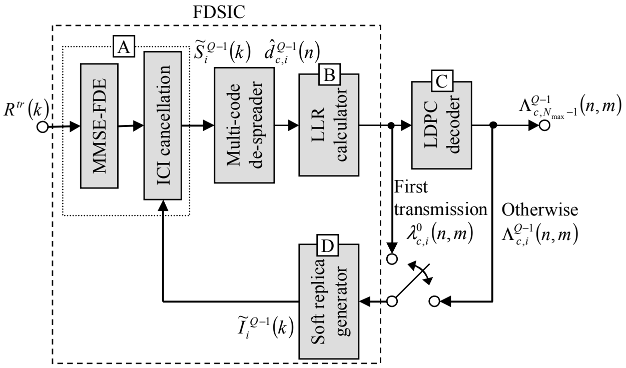


Figure 3. A combination of FDSIC and LDPC decoding.

A. MMSE-FDE and ICI cancellation

Joint MMSE-FDE and packet combining for the ith iteration ($i=0\sim N_{\max}-1$) can be expressed as

$$\tilde{R}_i^{Q-1}(k)=\sum_{tr=0}^{Q-1} w_i^{tr}(k)R^{tr}(k) \quad ,(6)$$

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

$$w_i^{tr}(k)=\frac{H^{tr^*}(k)}{\sum_{tr'=0}^{Q-1}\left|H^{tr'}(k)\right|^2+\left\{\frac{1}{SF}\frac{E_s}{N_0}\sum_{c=0}^{C-1}\rho_{c,i-1}^{Q-1}\left(\left\lfloor\frac{k}{SF}\right\rfloor\right)\right\}^{-1}} \quad ,(7)$$

where $E_s=PT_cN_c$ is the symbol energy and $\rho_{c,i-1}^{Q-1}(n)$ represents the contribution from the residual ICI obtained as [10]

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

where $\tilde{d}_{c,i-1}^{Q-1}(n)$ is the soft symbol replica generated at step D and $\bar{d}_{c,i-1}^{Q-1}(n)$ is the hard decision result obtained from $\tilde{d}_{c,i-1}^{Q-1}(n)$.

Eq. (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

$$\tilde{H}_i^{Q-1}(k)=\sum_{tr=0}^{Q-1}w_i^{tr}(k)H^{tr}(k) \quad .(10)$$

In Eq. (9), the first term is the desired signal component and the second term is the residual ICI, 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]

$$\tilde{S}_i^{Q-1}(k)=\tilde{R}_i^{Q-1}(k)-\tilde{I}_i^{Q-1}(k) \quad ,(11)$$

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

B. LLR computation

After performing ICI cancellation, parallel-to-serial (P/S) conversion and de-spreading are done to obtain

$$\begin{aligned}\hat{d}_{c,i}^{Q-1}(n)&=\frac{1}{SF}\sum_{k=nSF}^{(n+1)SF-1}\tilde{S}_i^{Q-1}(k)\left\{c_c^{\mathrm{oc}}(k\bmod SF)c^{\mathrm{scr}}(k)\right\}^*\\&=\sqrt{\frac{2P}{SF}}\hat{H}_i^{Q-1}(n)d_c(n)+\mu_{i,\mathrm{residual\,ICI}}^{Q\text{-}1}(n)+\mu_{i,\mathrm{noise}}^{Q\text{-}1}(n)\end{aligned} \quad ,(12)$$

where

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

In Eq. (12), the first term represents the desired signal component. The second term and third term are the residual ICI and noise, respectively. After de-spreading, the LLR $\lambda_{c,i}^{Q-1}(n,m)$, m=0~(log$_2M$)-1, of the mth bit belonging to the symbol $d_c(n)$, where M is the modulation level, is computed using [3]

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

where $d_{b(m)=0}^{\min}$ (or $d_{b(m)=1}^{\min}$) is the most probable symbol whose mth 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

$$\begin{aligned}&2\left\{\sigma_{c,i}^{Q-1}(n)\right\}^2=\frac{1}{SF^2}\frac{2E_s}{T_c}\\&\times\left[\begin{aligned}&\sum_{\substack{c'=0\\c'\neq c}}^{C-1}\rho_{c',i-1}^{Q-1}(n)\left\{\frac{1}{SF}\sum_{k=nSF}^{(n+1)SF-1}\left|\tilde{H}_i^{Q-1}(k)\right|^2-\left|\hat{H}_i^{Q-1}(n)\right|^2\right\}\\&\qquad+\left(\frac{1}{SF}\frac{E_s}{N_0}\right)^{-1}\sum_{tr=0}^{Q-1}\sum_{k=nSF}^{(n+1)SF-1}\left|w_i^{tr}(k)\right|^2\end{aligned}\right]\end{aligned} \quad .(15)$$

C. LDPC decoding

Let **A**=$\{A^{xy}$; x=0~(1-R)K/R-1, y=0~K/R-1$\}$ be a ((1-R)K/R-by-K/R) parity-check matrix [6], where (1-R)K/R and K/R are the bit lengths of the parity bit sequence and 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]

$$\begin{aligned}\alpha_i^{xy,Q-1}=&\left(\prod_{\substack{y'\in\{A^{xy'}=1\}\\y'\neq y}}\mathrm{sign}\left(\lambda_i^{y',Q-1}+\beta_i^{xy',Q-1}\right)\right)\\&\times f\left(\sum_{\substack{y'\in\{A^{xy'}=1\}\\y'\neq y}}f\left(\left|\lambda_i^{y',Q-1}+\beta_i^{xy',Q-1}\right|\right)\right)\end{aligned} \quad ,(16)$$

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

$$\begin{cases}\mathrm{sign}(x)=\begin{cases}1 & ,x\geq 0\\ \text{-}1 & ,\text{otherwise}\end{cases}\\ f(x)=\ln\dfrac{\exp(x)+1}{\exp(x)-1}\end{cases} \quad .(17)$$

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

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

for the (i+1)th iteration. The decoder outputs a sequence of *a-posteriori* LLR's $\{\Lambda_i^{y,Q-1}\}$, which is computed as [6]

$$\Lambda_i^{y,Q-1} = \lambda_i^{y,Q-1} + \sum_{x' \in \{A^{x'y}=1\}} \alpha_i^{x'y,Q-1} \quad , (19)$$

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 as

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

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

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

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

$$\tilde{I}_i^{Q-1}(k) = M_i^{Q-1}(k)\hat{S}_i^{Q-1}(k) \quad , (22)$$

where $\hat{S}_i^{Q-1}(k)$ is the MC-CDMA signal replica, obtained by spreading $\tilde{d}_{c,i}^{Q-1}(n)$, and $M_i^{Q-1}(k)$ is the cancellation weight given as [10]

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

IV. SIMULATION RESULTS

We evaluate by the computer simulation the HARQ throughput performances of LDPC-coded MC-CDMA using FDSIC. The throughput η is defined as

$$\eta = \frac{\text{Total number of transmitted infomation bits}}{\text{Total number of transmitted bits}} \cdot \frac{N_g}{N_c + N_g} \quad . \quad (24)$$

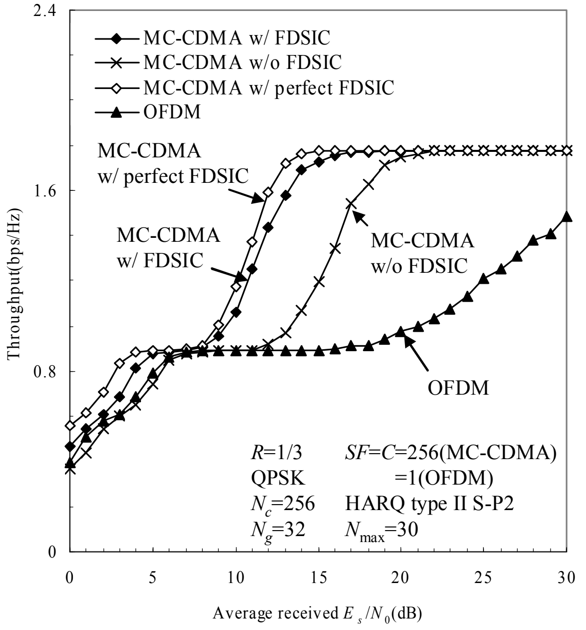
The second term in the RHS of Eq. (24) is due to the GI insertion loss.

The simulation condition is summarized in Table I. We assume a frequency-selective block Rayleigh-fading channel having a T_c-spaced L-path uniform power delay profile. Furthermore, we assume ideal channel estimation, ideal error detection, and no transmission error in ACK/NACK. No frequency-domain interleaving is considered for both MC-CDMA and OFDM. For fair comparison, *a-posteriori* LLR computation is repeated $N_{\max}$ times as in conventional LDPC decoder also for MC-CDMA without FDSIC and OFDM.

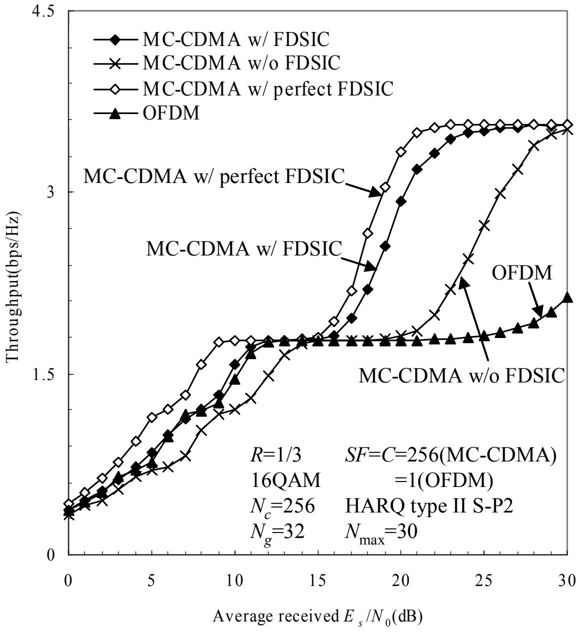
TABLE I. SIMULATION CONDITION

Channel code	R=1/3 LDPC code	
Data modulation	QPSK, 16QAM	
MC-CDMA	No. of FFT points	N_c=256
	No. of 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 τ_l=l (l=0~L-1)	

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 $N_{\max}$ (the number of iterations)=30. It is seen from Fig. 4 that the introduction of FDSIC significantly improves the throughput performance. When FDSIC is used, the E_s/N_0 reduction from the no FDSIC case for the throughput=0.8bps/Hz (1.6bps/Hz) is as much as 1.5dB (4dB) 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, the E_s/N_0 reduction for 16QAM is bigger than for QPSK and is about 2dB (5dB) for the throughput=1.5bps/Hz (3bps/Hz) (see Fig. 4(b)). The throughput performance gets closer to the perfect FDSIC case (i.e., $\tilde{I}_i^{Q-1}(k) = I_i^{Q-1}(k)$) by about 1dB for QPSK and about 2dB for 16QAM. It is also seen from Fig. 4 that the throughput performance of MC-CDMA using FDSIC is higher than using OFDM. This is because the frequency diversity gain can be obtained while sufficiently suppressing the residual ICI by de-spreading process in case of MC-CDMA. An about 1dB (12dB) E_s/N_0 reduction from the OFDM case is achieved at the throughput=0.8bps/Hz (1.6bps/Hz) for QPSK (see Fig. 4 (a)).



(a) QPSK



(b) 16QAM

Figure 4. Throughput performance.

V. CONCLUSION

The residual inter-code interference (ICI) after frequency-domain equalization (FDE) degrades the throughput performance of MC-CDMA HARQ. In this paper, for obtaining higher throughput performance, we applied the frequency-domain soft interference cancellation (FDSIC) technique to suppress the residual ICI. We presented the generation method of residual ICI replica for performing FDSIC from *a-posteriori* LLR obtained by the LDPC decoder. 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 better throughput performance than 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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