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PAPER

Iterative Frequency-Domain Soft Interference Cancellation for Multicode DS- and MC-CDMA Transmissions and Performance Comparison

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SUMMARY Frequency-domain equalization (FDE) based on the minimum mean square error (MMSE) criterion can significantly improve the BER performance of DS- and MC-CDMA systems in a severe frequency-selective fading channel. However, since the frequency-distorted signal cannot be completely equalized, the residual inter-code interference (ICI) limits the BER performance improvement. 4G systems must support much higher variable rate data services. Orthogonal multicode transmission technique has flexibility in offering variable rate services. However, the BER performance degrades as the number of parallel codes increases. In this paper, we propose an iterative frequency-domain soft interference cancellation (IFDSIC) scheme for multicode DS- and MC-CDMA systems and their achievable BER performances are evaluated by computer simulation.

key words: *DS-CDMA, MC-CDMA, frequency-domain equalization (FDE), multicode, interference cancelation*

1. Introduction

In the next generation mobile communication systems, much higher variable data rate services (e.g., higher than several 10 Mbps) than in the present third generation (3G) systems are required. Wideband direct sequence code division multiple access (DS-CDMA) with coherent rake combining has been adopted in the 3G systems for data transmissions of up to a few Mbps [1]. The transmission channels of 4G systems become severely frequency-selective and the transmission performance significantly degrades due to a large inter-path interference (IPI), even if coherent rake combining is used [2]. Recently, it has been shown [2]–[4] that the application of frequency-domain equalization (FDE) based on the minimum mean square error (MMSE) criterion, similar to multi-carrier (MC)-CDMA [5], [6], can significantly improve the bit error rate (BER) performance of DS-CDMA. Multicode DS- and MC-CDMA have flexibility in offering variable rate data services by simply changing the number of parallel orthogonal spreading codes [7], [8]. However, the frequency-distorted signal cannot be completely equalized by the use of FDE and the residual inter-code interference (ICI) degrades the BER performance as the number of parallel codes increases. Various ICI cancellation techniques have been proposed to solve this problem [9]–[11].

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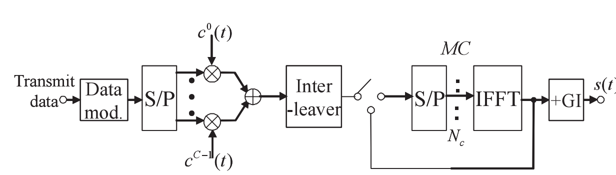
In this paper, we propose an iterative frequency-domain soft interference cancellation (IFDSIC) scheme and evaluate by computer simulation the achievable BER performances of multicode DS- and MC-CDMA in a frequency-selective Rayleigh fading channel. In IFDSIC, soft interference replica is generated using the log-likelihood ratio (LLR) and then, joint MMSE-FDE and ICI cancellation is carried out in the frequency-domain. MMSE and cancellation weights, taking into account the residual ICI, are derived. The remainder of this paper is organized as follows. The multicode CDMA signal transmission system model is presented in Sect. 2. IFDSIC is proposed in Sect. 3 and the MMSE and cancellation weights are derived in Sect. 4. In Sect. 5, the computer simulation results for the BER performances are presented. The paper is concluded in Sect. 6.

2. Multi-Code CDMA Signal Transmission System Model

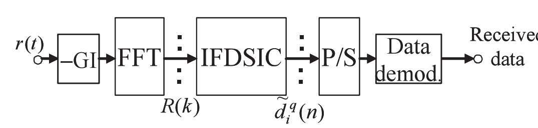
2.1 Transmit/Receive Signal

The transmitter/receiver structure for the multicode CDMA with IFDSIC is illustrated in Fig. 1. Throughout the paper, the chip-spaced discrete time representation is used.

We consider the transmission of one block of N_c chips, where N_c denotes the block length of fast Fourier transform (FFT). At a transmitter, a binary data sequence is transformed into data-modulated symbol sequence and then converted to C parallel streams by serial-to-parallel (S/P) conversion. Then the qth symbol stream $d^q(n)$, $n = 0 \sim (N-1)$



(a) Transmitter



(b) Receiver

Fig. 1 Transmitter/receiver structure for DS- and MC-CDMA.

and $N = N_c/SF$, is spread using an orthogonal spreading code $c^q_{ort}(t)$, $t = 0 \sim (SF-1)$, where SF is the spreading factor. The C chip streams are added and multiplied by a scramble sequence $c_{scr}(t)$. Chip interleaver is used in order to reduce the effect of error propagation due to decision feedback for the interference replica generation. N_c-point inverse FFT (IFFT) is applied to obtain the MC-CDMA signal. The multi-code CDMA signal $s(t)$, $t = 0 \sim (N_c-1)$, can be expressed using the equivalent baseband representation as

$$s(t) = \begin{cases} \sqrt{\dfrac{2E_s}{T_c SF}} \displaystyle\sum_{q=0}^{C-1} d^q\left(\left\lfloor \frac{t}{SF} \right\rfloor\right) \cdot c^q(t), & \text{DS-CDMA} \\ \sqrt{\dfrac{2E_s}{N_c T_c SF}} \displaystyle\sum_{k=0}^{N_c-1} \left\{ \sum_{q=0}^{C-1} d^q\left(\left\lfloor \frac{k}{SF} \right\rfloor\right) \cdot c^q(k) \right\} \exp\left(j2\pi k \frac{t}{N_c}\right), & \text{MC-CDMA} \end{cases} \quad (1)$$

where

$$c^q(t) = c^q_{ort}(t \bmod SF) \cdot c_{scr}(t) \quad (2)$$

and E_s represents the symbol energy, T_c the chip length, and $\lfloor x \rfloor$ the largest integer smaller than or equal to x. Before transmission, the last N_g chips of the N_c-chip block are copied and inserted, as a cyclic-prefix, into the guard interval (GI) placed at the beginning of the block.

Assuming that the channel has L independent propagation paths with chip-spaced distinct time delays, the impulse response $h(t)$ of the channel can be expressed as [12]

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

where h_l and τ_l are the lth path gain and time delay, respectively, with $\sum_{l=0}^{L-1} E[|h_l|^2] = 1$ (here, $E[.]$ is the ensemble average operation). The received signal $r(t)$ can be expressed as

$$r(t) = \sum_{l=0}^{L-1} h_l s(t - \tau_l) + \eta(t), \quad (4)$$

where $\eta(t)$ represents the zero-mean noise process having variance $2N_0/T_c$ with N_0 representing the single-sided power spectrum density of the additive white Gaussian noise (AWGN). Here, we have assumed a block fading, where the path gains remain constant over the time interval of $t = -N_g \sim (N_c - 1)$. After the removal of the GI, the received signal is decomposed into N_c frequency components $\{R(k); k = 0 \sim (N_c - 1)\}$ by applying N_c-point FFT. $R(k)$ is expressed as

$$R(k) = \frac{1}{\sqrt{N_c}} \sum_{t=0}^{N_c-1} r(t) \exp\left(-j2\pi k \frac{t}{N_c}\right) = H(k)S(k) + \Pi(k), \quad (5)$$

where $S(k)$ is the kth frequency component of $s(t)$ and $H(k)$ and $\Pi(k)$ are the channel gain and the noise component at the kth frequency due to the AWGN, respectively. $S(k)$, $H(k)$ and $\Pi(k)$ are given by

$$\begin{cases} S(k) = \dfrac{1}{\sqrt{N_c}} \displaystyle\sum_{t=0}^{N_c-1} s(t) \exp\left(-j2\pi k \frac{t}{N_c}\right) \\ H(k) = \displaystyle\sum_{l=0}^{L-1} h_l \exp\left(-j2\pi k \frac{\tau_l}{N_c}\right) \\ \Pi(k) = \dfrac{1}{\sqrt{N_c}} \displaystyle\sum_{t=0}^{N_c-1} \eta(t) \exp\left(-j2\pi k \frac{t}{N_c}\right) \end{cases} . \quad (6)$$

2.2 FDE and Residual ICI

The structure of IFDSIC is illustrated in Fig. 2. At the initial iteration ($i = 0$), all C data sequences are detected by the conventional MMSE-FDE. At the $i = 1$st iteration onwards, soft symbol replicas are generated and joint MMSE-FDE and ICI cancellation is carried out in an iterative fashion. MMSE-FDE for the ith iteration is carried out as [10], [11]

$$\hat{R}_i(k) = w_i(k)R(k) = S(k)\hat{H}_i(k) + \hat{\Pi}_i(k) \quad (7)$$

with

$$\begin{cases} \hat{H}_i(k) = w_i(k)H(k) \\ \hat{\Pi}_i(k) = w_i(k)\Pi(k) \end{cases} , \quad (8)$$

where $w_i(k)$ is the MMSE weight and $\hat{H}_i(k)$ and $\hat{\Pi}_i(k)$ are the equivalent channel gain and the noise component, respectively. In the following, we derive the residual ICI and its frequency-domain representation necessary for implementing the IFDSIC.

First, we consider DS-CDMA. The time-domain signal after MMSE-FDE is given by applying N_c-point IFFT to $\{\hat{R}_i(k); k = 0 \sim N_c - 1\}$ as

$$\begin{aligned} \hat{r}_i(t) &= \frac{1}{\sqrt{N_c}} \sum_{k=0}^{N_c-1} \hat{R}_i(k) \exp\left(j2\pi t \frac{k}{N_c}\right) \\ &= A_i^{DS} s(t) + \mu^{DS}_{ICI,i}(t) + \mu^{DS}_{noise,i}(t), \end{aligned} \quad (9)$$

where

$$\begin{cases} A_i^{DS} = \dfrac{1}{N_c} \displaystyle\sum_{k=0}^{N_c-1} \hat{H}_i(k) \\ \mu^{DS}_{ICI,i}(t) = \dfrac{1}{N_c} \displaystyle\sum_{k=0}^{N_c-1} \hat{H}_i(k) \left[\sum_{\substack{\tau=0 \\ \neq t}}^{N_c-1} s(\tau) \exp\left(j2\pi k \frac{t-\tau}{N_c}\right) \right] . \\ \mu^{DS}_{noise,i}(t) = \dfrac{1}{\sqrt{N_c}} \displaystyle\sum_{k=0}^{N_c-1} \hat{\Pi}_i(k) \exp\left(j2\pi k \frac{t}{N_c}\right) \end{cases} \quad (10)$$

The first term of Eq. (9) is the desired signal, the second is the residual ICI and the third is the noise. For IFDSIC, the frequency-domain representation of the residual

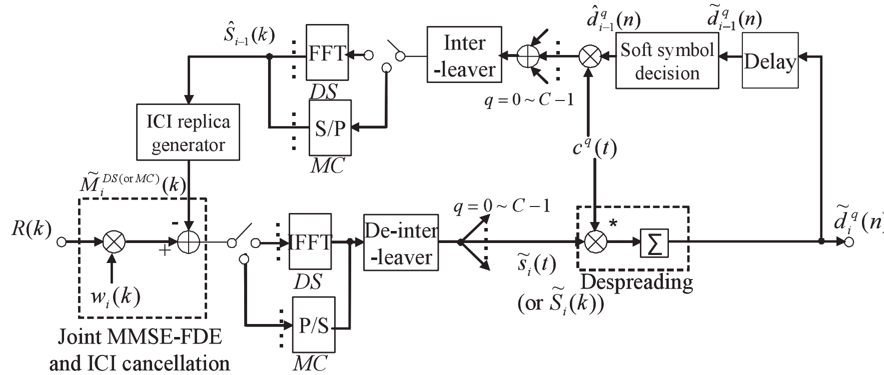


Fig. 2 IFDSIC structure for DS- and MC-CDMA.

ICI $\mu^{DS}_{ICI,i}(t)$ is necessary. The kth frequency component $M_i^{DS}(k)$ of the residual ICI before despreading can be expressed from Eq. (10) as

$$\begin{aligned} M_i^{DS}(k) &= \frac{1}{\sqrt{N_c}} \sum_{t=0}^{N_c-1} \mu^{DS}_{ICI,i}(t) \exp\left(-j2\pi k \frac{t}{N_c}\right) \\ &= \left(\hat{H}_i(k) - A_i^{DS}\right) S(k). \end{aligned} \tag{11}$$

Next, we consider MC-CDMA. Since despreading is carried out on the frequency-domain signal, we obtain an expression for the despreader output. From Eq. (7), we have

$$\begin{aligned} \tilde{d}^q_i(n) &= \frac{1}{SF} \sum_{k=nSF}^{(n+1)SF-1} \hat{R}_i(k) \cdot \{c^q(k)\}^* \\ &= \sqrt{\frac{2E_s}{T_c SF}} A_i^{MC}(n) d^q(n) + \mu^{MC}_{ICI,i}(n) + \mu^{MC}_{noise,i}(n), \end{aligned} \tag{12}$$

where

$$\begin{cases} A_i^{MC}(n) = \dfrac{1}{SF} \displaystyle\sum_{k=nSF}^{(n+1)SF-1} \hat{H}_i(k) \\ \mu^{MC}_{ICI,i}(n) = \sqrt{\dfrac{2E_s}{T_c SF}} \dfrac{1}{SF} \displaystyle\sum_{k=nSF}^{(n+1)SF-1} \hat{H}_i(k) \\ \qquad \times \left[\displaystyle\sum_{\substack{q'=0 \\ \neq q}}^{C-1} d^{q'}\left(\left\lfloor \frac{k}{SF} \right\rfloor\right) c^{q'}(k) \{c^q(k)\}^* \right] \\ \mu^{MC}_{noise,i}(n) = \dfrac{1}{SF} \displaystyle\sum_{k=nSF}^{(n+1)SF-1} \hat{\Pi}_i(k) \{c^q(k)\}^* \end{cases}. \tag{13}$$

The first term of Eq. (12) is the desired symbol, the second is the residual ICI, and the third is the noise. For IFDSIC, the frequency-domain representation of the residual ICI $\mu^{MC}_{ICI,i}(n)$ is necessary. Since $\mu^{MC}_{ICI,i}(n)$ can be rewritten using Eq. (13) as

$$\mu^{MC}_{ICI,i}(n) = \frac{1}{SF} \sum_{k=nSF}^{(n+1)SF-1} \left\{ \left(\hat{H}_i(k) - A_i^{MC}(\lfloor k/SF \rfloor)\right) S(k) \right\} \{c^q(k)\}^*, \tag{14}$$

the kth frequency component $M_i^{MC}(k)$ of the residual ICI before despreading can be expressed as

$$M_i^{MC}(k) = \left(\hat{H}_i(k) - A_i^{MC}(\lfloor k/SF \rfloor)\right) S(k). \tag{15}$$

2.3 ICI Cancellation and Despreading

ICI cancellation is performed on $\hat{R}_i(k)$ for both DS- and MC-CDMA as

$$\tilde{R}_i(k) = \hat{R}_i(k) - \tilde{M}_i^{DS(\text{or } MC)}(k), \tag{16}$$

where $\tilde{M}_i^{DS}(k)$ is given by Eq. (11) for DS-CDMA and $\tilde{M}_i^{MC}(k)$ by Eq. (15) for MC-CDMA. However, the transmitted multicode CDMA signal $s(t)$ is unknown to the receiver, and hence, $S(k)$ is also unknown. Therefore, the replica of the transmitted multicode CDMA signal is generated by feeding back the $(i-1)$th iteration result. The kth frequency component of the multicode CDMA signal replica is denoted by $\hat{S}_{i-1}(k)$. Instead of Eqs. (11) and (15), we use the following:

$$\begin{cases} \tilde{M}_i^{DS}(k) = \left(\hat{H}_i(k) - A_i^{DS}\right) \hat{S}_{i-1}(k) \\ \tilde{M}_i^{MC}(k) = \left(\hat{H}_i(k) - A_i^{MC}(\lfloor k/SF \rfloor)\right) \hat{S}_{i-1}(k) \end{cases}, \tag{17}$$

where $\tilde{M}_0^{DS}(k) = \tilde{M}_0^{MC}(k) = 0$ for all k.

In DS-CDMA, after the residual ICI cancellation, N_c-point IFFT is applied to transform $\{\tilde{R}_i(k); k = 0 \sim N_c - 1\}$ into the time-domain chip sequence $\{\tilde{r}_i(t); t = 0 \sim N_c - 1\}$. Then, despreading is carried out. On the other hand, in MC-CDMA, despreading is carried out directly using $\{\tilde{R}_i(k); k = 0 \sim N_c - 1\}$. The despreading operation is expressed as

$$\tilde{d}^q_i(n) = \begin{cases} \dfrac{1}{SF} \displaystyle\sum_{t=nSF}^{(n+1)SF-1} \tilde{r}_i(t) \cdot \{c^q(t)\}^* & \text{for DS-CDMA} \\ \dfrac{1}{SF} \displaystyle\sum_{k=nSF}^{(n+1)SF-1} \tilde{R}_i(k) \cdot \{c^q(k)\}^* & \text{for MC-CDMA} \end{cases}, \tag{18}$$

which is the decision variable associated with $d^q(n)$ after the ith iteration.

3. ICI Replica Generation

The ICI replica generation for $i(\geq 1)$th iteration is described below. The soft symbol replica $\hat{d}^q_{i-1}(n)$ for the qth parallel symbol stream is generated by using the decision variable $\tilde{d}^q_{i-1}(n)$ obtained at the $(i-1)$th iteration stage. Then, the LLR $\lambda^q_m(n)$ of the mth bit $b^q_{m,n}$ in the nth symbol $d^q(n)$ ($m = 0 \sim \log_2 M - 1$, where M is the modulation level) is computed using $\tilde{d}^q_{i-1}(n)$ as (see Appendix A)

$$\lambda^q_m(n) = \ln\left(\frac{p(b^q_{m,n}=1)}{p(b^q_{m,n}=0)}\right) \approx \begin{cases} \dfrac{1}{2(\hat{\sigma}^{DS}_{i-1})^2}\left(\left|\tilde{d}^q_{i-1}(n) - \sqrt{\dfrac{2E_s}{T_c SF}}A^{DS}_{i-1}d^{\min}_{b^q_{m,n}=0}\right|^2 - \left|\tilde{d}^q_{i-1}(n) - \sqrt{\dfrac{2E_s}{T_c SF}}A^{DS}_{i-1}d^{\min}_{b^q_{m,n}=1}\right|^2\right) & \text{for DS-CDMA} \\ \dfrac{1}{2\left(\hat{\sigma}^{MC}_{i-1}(n)\right)^2}\left(\left|\tilde{d}^q_{i-1}(n) - \sqrt{\dfrac{2E_s}{T_c SF}}A^{MC}_{i-1}(n)d^{\min}_{b^q_{m,n}=0}\right|^2 - \left|\tilde{d}^q_{i-1}(n) - \sqrt{\dfrac{2E_s}{T_c SF}}A^{MC}_{i-1}(n)d^{\min}_{b^q_{m,n}=1}\right|^2\right) & \text{for MC-CDMA} \end{cases}, \quad (19)$$

where $p(b^q_{m,n} = 1)$ (or $p(b^q_{m,n} = 0)$) is the probability of $b^q_{m,n} = 1$ (or 0) and $d^{\min}_{b^q_{m,n}=0}$ (or $d^{\min}_{b^q_{m,n}=1}$) is the most probable symbol whose mth bit is 0 (or 1), for which the Euclidean distance from $\tilde{d}^q_{i-1}(n)$ is minimum. $2(\hat{\sigma}^{DS}_{i-1})^2$ (or $2\left(\hat{\sigma}^{MC}_{i-1}(n)\right)^2$) is the variance of the interference plus noise and $\hat{H}_{i-1}(k)$ is the equivalent channel gain, given by (see Appendices B and C)

$$\begin{cases} 2\left(\hat{\sigma}^{DS}_{i-1}\right)^2 = \dfrac{1}{SF^2}\cdot\dfrac{2E_s}{T_c} \times \left[\left(\sum_{q'=0}^{C-1}\rho^{q'}_{i-2}\right)\left\{\dfrac{1}{N_c}\sum_{k=0}^{N_c-1}\left|\hat{H}_{i-1}(k)\right|^2 - \left|A^{DS}_{i-1}\right|^2\right\} + \left(\dfrac{E_s}{N_0 SF}\right)^{-1}\left(\dfrac{1}{N_c}\sum_{k=0}^{N_c-1}|w_{i-1}(k)|^2\right)\right] & \text{for DS-CDMA} \\ 2\left(\hat{\sigma}^{MC}_{i-1}(n)\right)^2 = \dfrac{1}{SF^2}\cdot\dfrac{2E_s}{T_c} \times \left[\left(\sum_{\substack{q'=0\\ \neq q}}^{C-1}\rho^{q'}_{i-2}(n)\right)\left\{\dfrac{1}{SF}\sum_{k=nSF}^{(n+1)SF-1}\left|\hat{H}_{i-1}(k)\right|^2 - \left|A^{MC}_{i-1}(n)\right|^2\right\} + \left(\dfrac{E_s}{N_0 SF}\right)^{-1}\left(\dfrac{1}{SF}\sum_{k=nSF}^{(n+1)SF-1}|w_{i-1}(k)|^2\right)\right] & \text{for MC-CDMA} \end{cases}, \quad (20)$$

where $w_{i-1}(k)$ and $\rho^{q'}_{i-1}$ (or $\rho^{q'}_{i-1}(n)$) are the MMSE weight and the residual ICI, respectively (which will be derived in Sect. 4).

The soft decision symbol $\hat{d}^q_{i-1}(n)$ can be obtained from [13]

$$\hat{d}^q_{i-1}(n) = \sum_{d\in D} d_{b^q_{m,n}} \prod_{b^q_{m,n}\in d} p(b^q_{m,n}), \quad (21)$$

where $d_{b^q_{m,n}}$ is the candidate symbol (that has $b^q_{m,n}$ as the mth bit) in the signal space D. Since $p(b^q_{m,n} = 1) + p(b^q_{m,n} = 0) = 1$, $p(b^q_{m,n} = 1)$ and $p(b^q_{m,n} = 0)$ are given by

$$\begin{cases} p(b^q_{m,n}=1) = \dfrac{\exp[\lambda^q_m(n)]}{1+\exp[\lambda^q_m(n)]} \\ p(b^q_{m,n}=0) = \dfrac{1}{1+\exp[\lambda^q_m(n)]} \end{cases}. \quad (22)$$

From Eq. (21), $\hat{d}^q_{i-1}(n)$ for QPSK can be written as

$$\begin{aligned} \hat{d}^q_{i-1}(n) = &\left(\frac{1}{\sqrt{2}} + j\frac{1}{\sqrt{2}}\right)p\left(b^q_{0,n}=1\right)p\left(b^q_{1,n}=1\right) \\ &+\left(\frac{1}{\sqrt{2}} - j\frac{1}{\sqrt{2}}\right)p\left(b^q_{0,n}=1\right)p\left(b^q_{1,n}=0\right) \\ &+\left(-\frac{1}{\sqrt{2}} + j\frac{1}{\sqrt{2}}\right)p\left(b^q_{0,n}=0\right)p\left(b^q_{1,n}=1\right) \\ &+\left(-\frac{1}{\sqrt{2}} - j\frac{1}{\sqrt{2}}\right)p\left(b^q_{0,n}=0\right)p\left(b^q_{1,n}=0\right). \end{aligned} \quad (23)$$

Similar to QPSK, we get $\hat{d}^q_{i-1}(n)$ for 16QAM as

$$\begin{aligned} &\hat{d}^q_{i-1}(n) \\ &= \left(\frac{3}{\sqrt{10}}+j\frac{3}{\sqrt{10}}\right)p\left(b^q_{0,n}=1\right)p\left(b^q_{1,n}=1\right)p\left(b^q_{2,n}=1\right)p\left(b^q_{3,n}=1\right) \\ &+\left(\frac{3}{\sqrt{10}}+j\frac{1}{\sqrt{10}}\right)p\left(b^q_{0,n}=1\right)p\left(b^q_{1,n}=1\right)p\left(b^q_{2,n}=1\right)p\left(b^q_{3,n}=0\right) \\ &+\left(\frac{3}{\sqrt{10}}-j\frac{1}{\sqrt{10}}\right)p\left(b^q_{0,n}=1\right)p\left(b^q_{1,n}=1\right)p\left(b^q_{2,n}=0\right)p\left(b^q_{3,n}=0\right) \\ &+\left(\frac{3}{\sqrt{10}}-j\frac{3}{\sqrt{10}}\right)p\left(b^q_{0,n}=1\right)p\left(b^q_{1,n}=1\right)p\left(b^q_{2,n}=0\right)p\left(b^q_{3,n}=1\right) \\ &+\cdots. \end{aligned} \quad (24)$$

The substitution of Eq. (22) into Eqs. (23) and (24) gives

$$\hat{d}^q_{i-1}(n) = \begin{cases} \dfrac{1}{\sqrt{2}}\tanh\left(\dfrac{\lambda^q_0(n)}{2}\right) + j\dfrac{1}{\sqrt{2}}\tanh\left(\dfrac{\lambda^q_1(n)}{2}\right) & \text{for QPSK} \\ \dfrac{1}{\sqrt{10}}\tanh\left(\dfrac{\lambda^q_0(n)}{2}\right)\left\{2+\tanh\left(\dfrac{\lambda^q_1(n)}{2}\right)\right\} + j\dfrac{1}{\sqrt{10}}\tanh\left(\dfrac{\lambda^q_2(n)}{2}\right)\left\{2+\tanh\left(\dfrac{\lambda^q_3(n)}{2}\right)\right\} & \text{for 16QAM} \end{cases}. \quad (25)$$

$\hat{d}^q_{i-1}(n)$ is re-spread to obtain the soft replica $\hat{s}_{i-1}(t)$ of the transmitted multicode DS-CDMA signal. Then, $\hat{s}_{i-1}(t)$ is decomposed into N_c frequency components $\{\hat{S}_{i-1}(k); k = 0 \sim (N_c-1)\}$ by applying N_c-point FFT. The kth frequency component $\hat{S}_{i-1}(k)$ of $\hat{s}_{i-1}(t)$ is given by

$$\hat{S}_{i-1}(k) = \begin{cases} \frac{1}{\sqrt{N_c}}\sum_{t=0}^{N_c-1}\hat{s}_{i-1}(t)\exp\left(-j2\pi k\frac{t}{N_c}\right) \\ = \sqrt{\frac{2E_s}{T_c SFN_c}}\sum_{t=0}^{N_c-1}\left\{\sum_{q=0}^{C-1}\hat{d}_{i-1}^{q}\left(\left\lfloor\frac{t}{SF}\right\rfloor\right)\cdot c^q(t)\right\} \\ \times\exp\left(-j2\pi k\frac{t}{N_c}\right) \qquad \text{for DS-CDMA} \\ \sqrt{\frac{2E_s}{T_c SF}}\sum_{q=0}^{C-1}\hat{d}_{i-1}^{q}\left(\left\lfloor\frac{k}{SF}\right\rfloor\right)\cdot c^q(k) \\ \qquad \text{for MC-CDMA} \end{cases}. \tag{26}$$

Using Eqs. (17) and (26), we obtain the residual ICI replica $\tilde{M}_i^{DS}(k)$ and $\tilde{M}_i^{MC}(k)$ to be used for the ICI cancellation.

4. Derivation of MMSE Weights

First, we derive the MMSE weight for DS-CDMA and then for MC-CDMA.
(1) DS-CDMA
The equalization error $e_i(k)$ at the ith iteration is defined as the difference between $\tilde{R}_i(k)$ and the kth frequency signal component $A_i^{DS}S(k)$, which is given by the first term of Eq. (9). Using Eqs. (16) and (17), $e_i(k)$ is given by

$$\begin{aligned} e_i(k) &= \tilde{R}_i(k) - A_i^{DS}S(k) \\ &= \left(w_i(k)H(k) - A_i^{DS}\right)\left(S(k) - \hat{S}_{i-1}(k)\right) - w_i(k)\Pi(k). \end{aligned} \tag{27}$$

Note that although $A_i^{DS}S(k)$ is used as the reference in Eq. (27), an arbitrary coefficient α can be used instead of A_i^{DS} (i.e., $\alpha S(k)$), resulting in the same MMSE weight. Since $\Pi(k)$ is a zero-mean complex-valued Gaussian noise having variance $2N_0/T_c$, the mean square error (MSE) is given, using Eqs. (1), (6) and (26), by

$$\begin{aligned} E\left[|e_i(k)|^2\right] &= \frac{2E_s}{SFT_c}\left(\sum_{q'=0}^{C-1}\rho_{i-1}^{q'}\right)\left|w_i(k)H(k) - A_i^{DS}\right|^2 \\ &\quad + \frac{2N_0}{T_c}|w_i(k)|^2, \end{aligned} \tag{28}$$

where

$$\rho_{i-1}^{q'} = \begin{cases} 1, & i = 0 \\ \frac{1}{N}\sum_{n=0}^{N-1}\left\{\left|\bar{d}_{i-1}^{q'}(n)\right|^2 - \left|\hat{d}_{i-1}^{q'}(n)\right|^2\right\}, & i \geq 1 \end{cases} \tag{29}$$

with $\bar{d}_{i-1}^{q}(n)$ being the hard decision result obtained from $\tilde{d}_{i-1}^{q}(n)$.

The set of MMSE weights $\{w_i(k); k = 0 \sim (N_c-1)\}$ is the one that satisfies $\partial E[|e_i(k)|^2]/\partial w_i(k) = 0$ for all k. From Eq. (28), we have

$$\begin{aligned} &\frac{\partial E[|e_i(k)|^2]}{\partial w_i(k)} \\ &= \frac{\partial E[|e_i(k)|^2]}{\partial \mathrm{Re}[w_i(k)]} + j\frac{\partial E[|e_i(k)|^2]}{\partial \mathrm{Im}[w_i(k)]} \\ &= \left(\sum_{q'=0}^{C-1}\rho_{i-1}^{q'}\right)\left[w_i(k)\,|H(k)|^2 - A_i^{DS}H^*(k)\right] + w_i(k)\left(\frac{E_s}{SFN_0}\right)^{-1}, \end{aligned} \tag{30}$$

from which, we get the following MMSE weight:

$$w_i(k) = \frac{H^*(k)}{|H(k)|^2 + \left(\frac{E_s}{SFN_0}\sum_{q'=0}^{C-1}\rho_{i-1}^{q'}\right)^{-1}}. \tag{31}$$

Note that $w_0(k)$ is equal to the conventional MMSE weight.
(2) MC-CDMA
Using Eqs. (16) and (17), the equalization error $e_i(k)$ between $\tilde{R}_i(k)$ and the kth frequency desired signal component $A_i^{MC}(\lfloor k/SF\rfloor)S(k)$ is given as

$$\begin{aligned} e_i(k) &= \tilde{R}_i(k) - A_i^{MC}(\lfloor k/SF\rfloor)S(k) \\ &= \left(w_i(k)H(k) - A_i^{MC}(\lfloor k/SF\rfloor)\right)\left(S(k) - \hat{S}_{i-1}(k)\right) \\ &\quad + w_i(k)\Pi(k). \end{aligned} \tag{32}$$

Therefore, the MSE is given, using Eqs. (1), (6) and (26), as

$$\begin{aligned} &E\left[|e_i(k)|^2\right] \\ &= \frac{2E_s}{T_c SF}\left(\sum_{q'=0}^{C-1}\rho_{i-1}^{q'}\left(\left\lfloor\frac{k}{SF}\right\rfloor\right)\right)\left|w_i(k)H(k) - A_i^{MC}\left(\left\lfloor\frac{k}{SF}\right\rfloor\right)\right|^2 \\ &\quad + \frac{2N_0}{T_c}|w_i(k)|^2. \end{aligned} \tag{33}$$

Similar to DS-CDMA, we get the following weight for MC-CDMA:

$$w_i(k) = \frac{H^*(k)}{|H(k)|^2 + \left(\frac{E_s}{N_0 SF}\sum_{q'=0}^{C-1}\rho_{i-1}^{q'}(\lfloor k/SF\rfloor)\right)^{-1}}. \tag{34}$$

where

$$\rho_{i-1}^{q'}(n) = \begin{cases} 1, & i = 0 \\ \left|\bar{d}_{i-1}^{q'}(n)\right|^2 - \left|\hat{d}_{i-1}^{q'}(n)\right|^2, & i \geq 1 \end{cases}. \tag{35}$$

In the proposed IFDSIC, the ICI cancellation is performed in the frequency-domain after MMSE-FDE. However, the ICI cancellation can be carried out also in the time-domain without sacrificing the BER performance at all. For the time-domain SIC, the replica of time-domain ICI $\mu_{ICI,i}^{DS}(t)$ shown in Eq. (10) is required. It can be shown from Eq. (10) that the number of multiply operations necessary to generate the replica of $\{\mu_{ICI,i}^{DS}(t); t = 0 \sim N_c - 1\}$ is N_c^3. However, the number of multiply operations required in the proposed IFDSIC to generate the frequency-domain replica $\{\tilde{M}_i^{DS}(k);$

$k = 0 \sim N_c - 1\}$ is only $(N_c \log_2 N_c + N_c)$ (this includes $N_c \log_2 N_c$ multiply operations required to obtain $\hat{S}_{i-1}(k)$). Therefore, the proposed IFDSIC is much less computationally complex compared to the time-domain ICI cancellation scheme.

5. Simulation Results

Table 1 shows the computer simulation conditions. A chip-spaced 16-path ($L = 16$) frequency-selective block Rayleigh fading channel having uniform power delay profile are assumed. We assume an FFT block size of $N_c = 256$ chips, a GI length of $N_g = 32$ chips. We assume ideal channel estimation. In MC-CDMA, frequency interleaving is used to obtain the frequency diversity gain similar to DS-CDMA.

Figure 3 plots the average BER performance of IFDSIC for QPSK modulation as a function of the average received signal energy per bit-to-the AWGN power spectrum density ratio E_b/N_0 $(= (1/\log_2 M)SF(1 + N_g/N_c)(E_s/(SFN_0)))$. For comparison, the lower bound BER performance is also plotted. Without ICI cancellation ($i = 0$), the BER performance is severely degraded. However, it can be seen that the proposed IFDSIC can significantly improve the BER performance. When $SF = 16$, the BER performances improve as the number of iterations increases. However, DS-CDMA provides much better performance than MC-CDMA, although almost the same frequency diversity gain can be obtained in both DS- and MC-CDMA. This is because the accuracy of the generated ICI replica is worse in MC-CDMA than in DS-CDMA. In DS-CDMA, the ICI replica is generated using $(N\times C)$ decision variables while it is generated using only C decision variables in MC-CDMA (see Eq. (26)). (In Ref. [10], a similar trend in BER performance to our case is seen, although the interference replica and MMSE weight generations for the interference cancellation are different from our proposed scheme.) When $C = 4$ (16), at the 1st (2nd) iteration, the E_b/N_0 reduction for BER $= 10^{-3}$ from the no ICI cancellation case ($i = 0$) is 3.1 (4.7) dB for DS-CDMA and 2.3 (3.2) dB for MC-CDMA. On the other hand, when $SF = 256$, almost the same BER performance can be achieved for DS- and MC-CDMA because the accuracy of the ICI replica is almost the same in both DS- and MC-CDMA. When $C = 64$ (256), even at the $i =$ 1st (2nd) iteration, the E_b/N_0 reduction from the no ICI cancellation case ($i = 0$) is as much as 3.1 (4.7) dB for BER $= 10^{-3}$ and the BER performance gets close to the theoretical lower bound by about 0.6 (1.1) dB.

The BER performance is plotted in Fig. 4 for 16QAM. For 16QAM, the Euclidean distance between different symbols becomes shorter and hence, decision errors due to the residual ICI are more likely than for QPSK. However, IFDSIC is very effective to improve the BER performance even for 16QAM. Similar to the case of QPSK, MC-CDMA performs worse than DS-CDMA when $SF = 16$ due to the replica generation accuracy. When BER $= 10^{-3}$ and for full load case ($SF = C = 16$), only a slight performance improvement can be observed for MC-CDMA, while as much as a 5.5 dB E_b/N_0 reduction is obtained for DS-CDMA. However, when $SF = 256$, almost the same BER performance can be obtained for both DS- and MC-CDMA. When $C = 64$ (256), an E_b/N_0 reduction of as much as 6.2 (7.2) dB can be achieved for BER $= 10^{-3}$ in both DS- and MC-CDMA.

Table 1 Simulation conditions.

Transmitter	Modulation	QPSK, 16QAM
	FFT block length	N_c=256 chips
	GI length	N_g=32 chips
	Spreading factor	SF=16, 256
Channel	Fading	Frequency-selective block Rayleigh fading
	Number of paths	L=16
	Power delay profile	Uniform
Receiver	Frequency-domain equalization	MMSE
	Channel estimation	Ideal

6. Conclusion

In this paper, we proposed an iterative frequency-domain soft interference cancellation (IFDSIC) scheme for multi-code DS- and MC-CDMA systems. Joint MMSE-FDE and ICI cancellation is carried out in an iterative fashion. The MMSE and cancellation weights were derived taking into account the residual ICI. Both weights are updated in each iteration. The BER performance, with the proposed IFDSIC in a frequency-selective Rayleigh fading, was evaluated by computer simulation. It was found that in a small spreading factor case ($SF = 16$), since the accuracy of the generated ICI replica is worse in MC-CDMA than in DS-CDMA, the E_b/N_0 reduction from the no cancellation case is as much as about 4.7 (5.5) dB and 3.2 (2.4) dB when $C = 16$ and QPSK data modulation (16QAM) for achieving BER $= 10^{-3}$ in DS- and MC-CDMA, respectively. However, when $SF = C = 256$ and QPSK data modulation (16QAM) is used, the E_b/N_0 reduction from the no cancellation case is as much as about 5.8 (7.2) dB for achieving BER $= 10^{-3}$ and the performance approaches the theoretical lower bound by about 0.7 (2.4) dB in both DS- and MC-CDMA.

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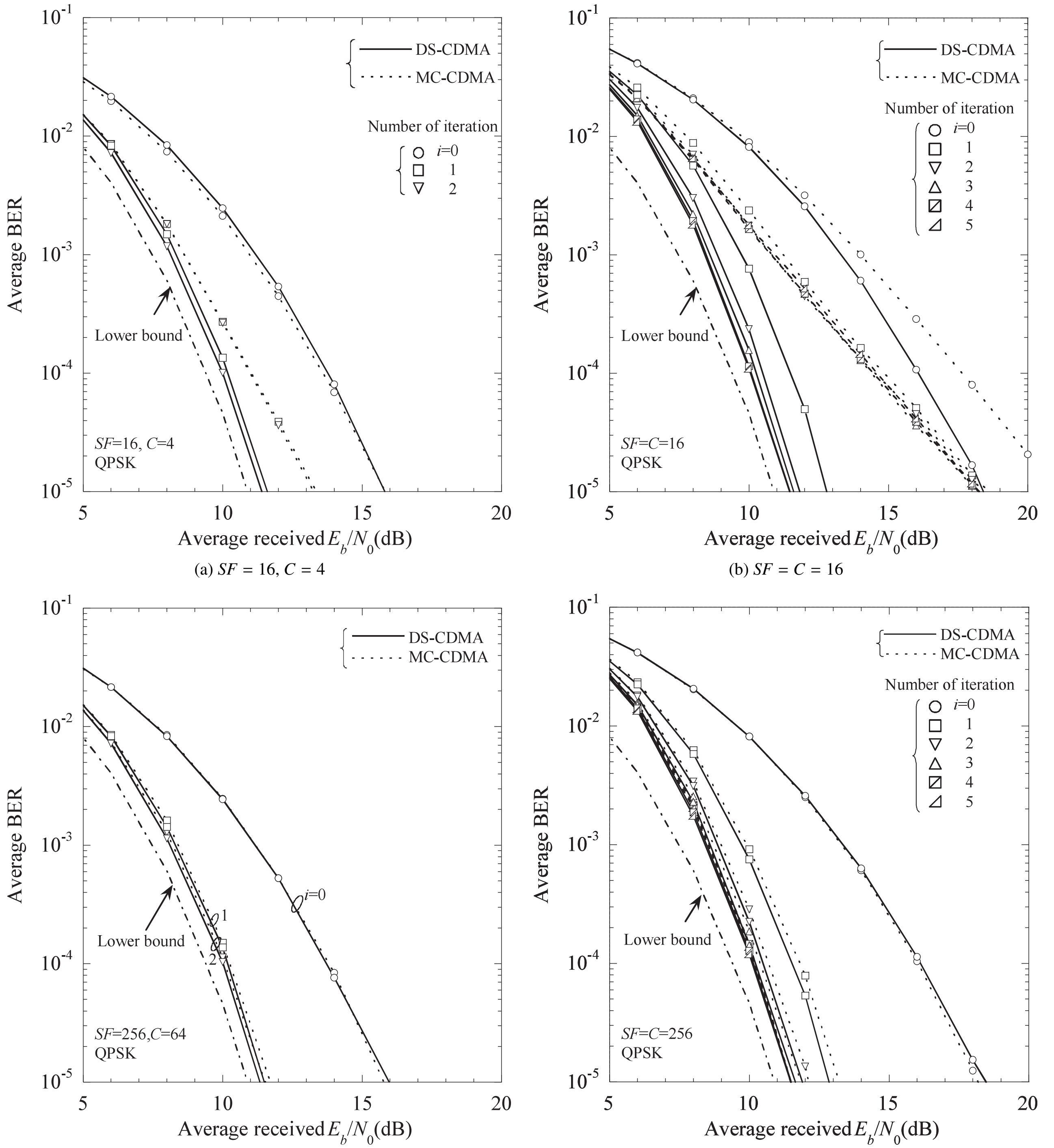

(a) $SF = 16$, $C = 4$ (b) $SF = C = 16$

(c) $SF = 256$, $C = 64$ (d) $SF = C = 256$

Fig. 3 BER performance of IFDSIC for QPSK modulation.

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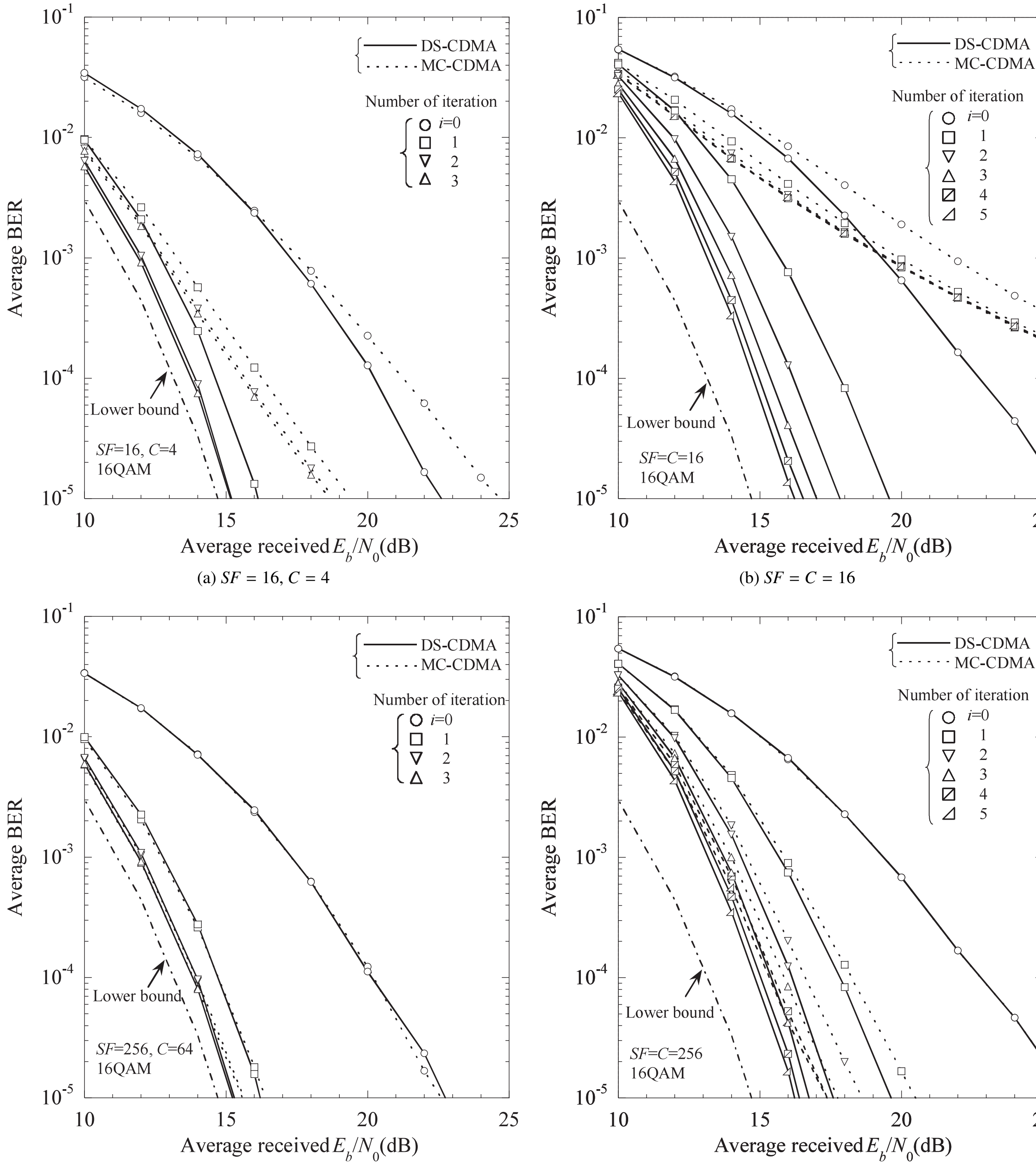

(a) SF = 16, C = 4 (b) SF = C = 16

(c) SF = 256, C = 64 (d) SF = C = 256

Fig. 4 BER performance of IFDSIC for 16QAM.

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Appendix A: Log-Likelihood Ratio (LLR)

Form Eqs. (5) and (16), the kth frequency component after joint MMSE-FDE and ICI cancellation is given by

$$\tilde{R}_i(k) = \hat{H}_i(k)S(k) - M_i^{DS\,(\text{or } MC)}(k) + w_i(k)\Pi(k). \quad (\text{A}\cdot 1)$$

The decision variable $\tilde{d}_i^q(n)$ obtained after despreading can be expressed as

$$\tilde{d}_i^q(n) = \begin{cases} \dfrac{1}{SF}\displaystyle\sum_{t=nSF}^{(n+1)SF-1} \tilde{r}_i(t)\cdot\{c^q(t)\}^* \\ = \sqrt{\dfrac{2E_s}{T_cSF}}A_i^{DS}d^q(n) + \tilde{\mu}_{ICI,i}^{DS}(n) + \tilde{\mu}_{noise,i}^{DS}(n) \\ \qquad \text{for DS-CDMA} \\ \dfrac{1}{SF}\displaystyle\sum_{k=nSF}^{(n+1)SF-1} \tilde{R}_i(k)\cdot\{c^q(k)\}^* \\ = \sqrt{\dfrac{2E_s}{T_cSF}}A_i^{MC}(n)d^q(n) + \tilde{\mu}_{ICI,i}^{MC}(n) + \tilde{\mu}_{noise,i}^{MC}(n) \\ \qquad \text{for MC-CDMA} \end{cases}, \quad (\text{A}\cdot 2)$$

where the 1st term is the desired signal component, the 2nd and 3rd terms are the residual ICI and noise components, respectively. A_i^{DS} and $A_i^{MC}(n)$ are the equivalent channel gains and are given by Eqs. (10) and (13). $\tilde{\mu}_{ICI,i}^{DS\,(\text{or } MC)}(n)$ and $\tilde{\mu}_{noise,i}^{DS\,(\text{or } MC)}(n)$ are respectively given by

$$\begin{cases} \tilde{\mu}_{ICI,i}^{DS}(n) \\ = \dfrac{1}{SF}\displaystyle\sum_{t=nSF}^{(n+1)SF-1}\{c^q(t)\}^* \\ \times\dfrac{1}{\sqrt{N_c}}\displaystyle\sum_{k=0}^{N_c-1}\left(\hat{H}_i(k) - A_i^{DS}\right)\left(S(k) - \hat{S}_{i-1}(k)\right)\exp\left(j2\pi k\dfrac{t}{N_c}\right) \\ \qquad \text{for DS-CDMA} \\ \tilde{\mu}_{ICI,i}^{MC}(n) \\ = \dfrac{1}{SF}\sqrt{\dfrac{2E_s}{T_cSF}}\displaystyle\sum_{k=nSF}^{(n+1)SF-1}\{c^q(k)\}^* \\ \times\left(\hat{H}_i(k) - A_i^{MC}(n)\right)\displaystyle\sum_{\substack{q'=0\\ \neq q}}^{C-1}\left(d^{q'}(n) - \hat{d}_{i-1}^{q'}(n)\right)c^{q'}(k) \\ \qquad \text{for MC-CDMA} \end{cases} \quad (\text{A}\cdot 3)$$

and

$$\begin{cases} \tilde{\mu}_{noise,i}^{DS}(n) = \dfrac{1}{SF}\displaystyle\sum_{t=nSF}^{(n+1)SF-1}\{c^q(t)\}^* \\ \qquad\times\dfrac{1}{\sqrt{N_c}}\displaystyle\sum_{k=0}^{N_c-1}w_i(k)\Pi(k)\exp\left(j2\pi k\dfrac{t}{N_c}\right) \\ \qquad \text{for DS-CDMA} \\ \tilde{\mu}_{noise,i}^{MC}(n) = \dfrac{1}{SF}\displaystyle\sum_{k=nSF}^{(n+1)SF-1}\{c^q(k)\}^*\times w_i(k)\Pi(k) \\ \qquad \text{for MC-CDMA} \end{cases}. \quad (\text{A}\cdot 4)$$

Approximating the noise plus ICI as a zero-mean complex Gaussian variable with variance $2(\hat{\sigma}_i^{DS})^2$ (or $2\left(\hat{\sigma}_i^{MC}(n)\right)^2$), the LLR for $b_{m,n}^q$ is given by [14]

$$\lambda_m^q(n) = \ln\left(\frac{p(b_{m,n}^q = 1)}{p(b_{m,n}^q = 0)}\right) \approx$$

$$\begin{cases} \ln\dfrac{\displaystyle\max_{\{d:b_{m,n}^q=1\}}\left[\dfrac{1}{\sqrt{2\pi(\hat{\sigma}_i^{DS})^2}}\exp\left(-\dfrac{1}{2(\hat{\sigma}_i^{DS})^2}\left|\tilde{d}_i^q(n) - \sqrt{\dfrac{2E_s}{T_cSF}}A_i^{DS}d\right|^2\right)\right]}{\displaystyle\max_{\{d:b_{m,n}^q=0\}}\left[\dfrac{1}{\sqrt{2\pi(\hat{\sigma}_i^{DS})^2}}\exp\left(-\dfrac{1}{2(\hat{\sigma}_i^{DS})^2}\left|\tilde{d}_i^q(n) - \sqrt{\dfrac{2E_s}{T_cSF}}A_i^{DS}d\right|^2\right)\right]} \\ \qquad \text{for DS-CDMA} \\ \ln\dfrac{\displaystyle\max_{\{d:b_{m,n}^q=1\}}\left[\dfrac{1}{\sqrt{2\pi\left(\hat{\sigma}_i^{MC}(n)\right)^2}}\times\exp\left(-\dfrac{1}{2\left(\hat{\sigma}_i^{MC}(n)\right)^2}\left|\tilde{d}_i^q(n) - \sqrt{\dfrac{2E_s}{T_cSF}}A_i^{MC}(n)d\right|^2\right)\right]}{\displaystyle\max_{\{d:b_{m,n}^q=0\}}\left[\dfrac{1}{\sqrt{2\pi\left(\hat{\sigma}_i^{MC}(n)\right)^2}}\times\exp\left(-\dfrac{1}{2\left(\hat{\sigma}_i^{MC}(n)\right)^2}\left|\tilde{d}_i^q(n) - \sqrt{\dfrac{2E_s}{T_cSF}}A_i^{MC}(n)d\right|^2\right)\right]} \\ \qquad \text{for MC-CDMA} \end{cases}, \quad (\text{A}\cdot 5)$$

where $\{d : b_{m,n}^q = 1 \text{ or } 0\}$ denotes a set of symbols whose mth bit is 1 or 0, respectively. From [15], we approximate the denominator and numerator of Eq. (A·5) as

$$\begin{cases} \max_{\{d:b^q_{m,n}=1\text{ (or 0)}\}} \left[\frac{1}{\sqrt{2\pi\left(\hat{\sigma}^{DS}_i\right)^2}} \exp\left(-\frac{1}{2\left(\hat{\sigma}^{DS}_i\right)^2} \left| \tilde{d}^q_i(n) - \sqrt{\frac{2E_s}{T_cSF}} A^{DS}_i d \right|^2 \right) \right] \\ = \frac{1}{\sqrt{2\pi\left(\hat{\sigma}^{DS}_i\right)^2}} \\ \quad \times \exp\left(-\frac{1}{2\left(\hat{\sigma}^{DS}_i\right)^2} \left| \tilde{d}^q_i(n) - \sqrt{\frac{2E_s}{T_cSF}} A^{DS}_i d^{\min}_{b^q_{m,n}=1\text{ (or 0)}} \right|^2 \right) \quad \text{for DS-CDMA} \\ \max_{\{d:b^q_{m,n}=1\text{ (or 0)}\}} \left[\frac{1}{\sqrt{2\pi\left(\hat{\sigma}^{MC}_i(n)\right)^2}} \times \exp\left(-\frac{1}{2\left(\hat{\sigma}^{MC}_i(n)\right)^2} \left| \tilde{d}^q_i(n) - \sqrt{\frac{2E_s}{T_cSF}} A^{MC}_i(n) d \right|^2 \right) \right] \\ = \frac{1}{\sqrt{2\pi\left(\hat{\sigma}^{MC}_i(n)\right)^2}} \\ \quad \times \exp\left(-\frac{1}{2\left(\hat{\sigma}^{MC}_i(n)\right)^2} \left| \tilde{d}^q_i(n) - \sqrt{\frac{2E_s}{T_cSF}} A^{MC}_i(n) d^{\min}_{b^q_{m,n}=1\text{ (or 0)}} \right|^2 \right) \quad \text{for MC-CDMA} \end{cases}, \tag{A·6}$$

where $d^{\min}_{b^q_{m,n}=1\text{ (or 0)}}$ is the most probable symbol whose mth bit is 1 (or 0). As a consequence, we can rewrite Eq. (A·5) as

$$\lambda^q_m(n) \approx \begin{cases} \frac{1}{2(\hat{\sigma}^{DS}_i)^2} \left(\left| \tilde{d}^q_i(n) - \sqrt{\frac{2E_s}{T_cSF}} A^{DS}_i d^{\min}_{b^q_{m,n}=0} \right|^2 - \left| \tilde{d}^q_i(n) - \sqrt{\frac{2E_s}{T_cSF}} A^{DS}_i d^{\min}_{b^q_{m,n}=1} \right|^2 \right) \quad \text{for DS-CDMA} \\ \frac{1}{2\left(\hat{\sigma}^{MC}_i(n)\right)^2} \left(\left| \tilde{d}^q_i(n) - \sqrt{\frac{2E_s}{T_cSF}} A^{MC}_i(n) d^{\min}_{b^q_{m,n}=0} \right|^2 - \left| \tilde{d}^q_i(n) - \sqrt{\frac{2E_s}{T_cSF}} A^{MC}_i(n) d^{\min}_{b^q_{m,n}=1} \right|^2 \right) \quad \text{for MC-CDMA} \end{cases}. \tag{A·7}$$

Appendix B: Derivation of $2(\hat{\sigma}^{DS}_i)^2$ for DS-CDMA

(1) $2(\hat{\sigma}^{DS}_{ICI,i})^2$

Since the scramble sequence is white-noise like and hence, $E[c^q(t)\{c^q(\tau)\}^*] = \delta(t-\tau)$ (here, $\delta(t)$ is the delta function). From Eq. (A·3), we have

$$\begin{aligned} 2\left(\hat{\sigma}^{DS}_{ICI,i}\right)^2 &= E\left[\left|\tilde{\mu}^{DS}_{ICI,i}(n)\right|^2\right] \\ &= \frac{1}{N_cSF^2} \sum_{t=nSF}^{(n+1)SF-1} \sum_{k=0}^{N_c-1} \sum_{k'=0}^{N_c-1} \left(\hat{H}_i(k) - A^{DS}_i\right)\left(\hat{H}_i(k') - A^{DS}_i\right)^* \\ &\quad \times E\left[\left(S(k) - \hat{S}_{i-1}(k)\right)\left(S(k') - \hat{S}_{i-1}(k')\right)^*\right] \exp\left(j2\pi(k-k')\frac{t}{N_c}\right). \end{aligned} \tag{A·8}$$

The residual ICI component $\left\{\hat{H}_i(k) - A^{DS}_i\right\}\left\{S(k) - \hat{S}_{i-1}(k)\right\}$ at the kth frequency is assumed to be zero-mean variable. Eq.(A·8) becomes

$$\begin{aligned} 2\left(\hat{\sigma}^{DS}_{ICI,i}\right)^2 &= \frac{1}{N_cSF} \sum_{k=0}^{N_c-1} \left|\hat{H}_i(k) - A^{DS}_i\right|^2 E\left[\left|S(k) - \hat{S}_{i-1}(k)\right|^2\right] \\ &= \frac{1}{SF^2} \frac{2E_s}{T_c} \left(\sum_{q=0}^{C-1} \rho^q_{i-1}\right) \left[\frac{1}{N_c} \sum_{k=0}^{N_c-1} \left|\hat{H}_i(k)\right|^2 - \left|A^{DS}_i\right|^2\right], \end{aligned} \tag{A·9}$$

where

$$\sum_{q=0}^{C-1} \rho^q_{i-1} = \left(\frac{2E_s}{T_cSF}\right)^{-1} E\left[\left|S(k) - \hat{S}_{i-1}(k)\right|^2\right]. \tag{A·10}$$

Substitution of Eqs. (1), (6) and (26) into Eq. (A·10) gives

$$\begin{aligned} \sum_{q=0}^{C-1} \rho^q_{i-1} &= \frac{1}{N_c} E\left[\left| \sum_{t=0}^{N_c-1} \sum_{q'=0}^{C-1} \left(d^q\left(\left\lfloor \frac{t}{SF} \right\rfloor\right) - \hat{d}^q_{i-1}\left(\left\lfloor \frac{t}{SF} \right\rfloor\right)\right) c^{q'}(t) \times \exp\left(-j2\pi k\frac{t}{N_c}\right) \right|^2\right] \\ &= \frac{1}{N} \sum_{n=0}^{N-1} \sum_{q=0}^{C-1} E\left[\left|d^q(n) - \hat{d}^q_{i-1}(n)\right|^2\right], \end{aligned} \tag{A·11}$$

where $N = N_c/SF$. In Eq. (A·11), $\hat{d}^q_{i-1}(n)$ is given by Eq. (21) and is an expectation of $d^q(n)$, i.e., $\hat{d}^q_{i-1}(n) = E[d^q(n)]$. Therefore, Eq. (A·11) becomes

$$\sum_{q=0}^{C-1} \rho^q_{i-1} = \frac{1}{N} \sum_{n=0}^{N-1} \sum_{q=0}^{C-1} \left\{|d^q(n)|^2 - \left|\hat{d}^q_{i-1}(n)\right|^2\right\}. \tag{A·12}$$

Since $d^q(n)$ is unknown to the receiver, we take a heuristic approach and replace $d^q(n)$ by the hard decision result $\bar{d}^q_{i-1}(n)$ obtained from $\tilde{d}^q_{i-1}(n)$. For the case of QPSK data modulation, since we always have $|d^q(n)| = 1$, no approximation error exists. Eq. (A·12) is replaced by

$$\sum_{q=0}^{C-1} \rho^q_{i-1} \approx \frac{1}{N} \sum_{n=0}^{N-1} \sum_{q=0}^{C-1} \left\{\left|\bar{d}^q_{i-1}(n)\right|^2 - \left|\hat{d}^q_{i-1}(n)\right|^2\right\}. \tag{A·13}$$

(2) $2(\hat{\sigma}^{DS}_{noise,i})^2$

From Eq. (A·4), $2(\hat{\sigma}^{DS}_{noise,i})^2$ is given by

$$2(\hat{\sigma}^{DS}_{noise,i})^2 = E\left[\left|\tilde{\mu}^{DS}_{noise,i}(n)\right|^2\right] = \frac{1}{N_c SF^2}\sum_{t=nSF}^{(n+1)SF-1} \times \sum_{k=0}^{N_c-1}\sum_{k=0}^{N_c-1} E\left[w_i(k)\Pi(k)\{w_i(k')\Pi(k')\}^*\right] \times \exp\left(j2\pi(k-k')\frac{t}{N_c}\right). \quad (A\cdot 14)$$

Since $\{\Pi(k); k = 0 \sim (N_c-1)\}$ are i.i.d. zero-mean complex-valued Gaussian variables with variance $2(N_0/T_c)$, we have

$$2(\hat{\sigma}^{DS}_{noise,i})^2 = \frac{1}{SF}\frac{2N_0}{T_c}\left(\frac{1}{N_c}\sum_{k=0}^{N_c-1}|w_i(k)|^2\right). \quad (A\cdot 15)$$

(3) $2\hat{\sigma}_i^2$
From Eqs. (A·9) and (A·15), $2\hat{\sigma}_i^2$ is given by

$$2\left(\hat{\sigma}^{DS}_i\right)^2 = 2\left(\hat{\sigma}^{DS}_{ICI,i}\right)^2 + 2(\hat{\sigma}^{DS}_{noise,i})^2 = \frac{1}{SF^2}\frac{2E_s}{T_c}\left[\begin{array}{l}\left(\sum_{q'=0}^{C-1}\rho^{q'}_{i-1}\right)\left\{\frac{1}{N_c}\sum_{k=0}^{N_c-1}\left|\hat{H}_i(k)\right|^2-\left|A^{DS}_i\right|^2\right\}\\ +\left(\frac{E_s}{N_0 SF}\right)^{-1}\left(\frac{1}{N_c}\sum_{k=0}^{N_c-1}|w_i(k)|^2\right)\end{array}\right]. \quad (A\cdot 16)$$

Appendix C: Derivation of $2\left(\hat{\sigma}^{MC}_i(n)\right)^2$ for MC-CDMA

(1) $2\left(\hat{\sigma}^{MC}_{ICI,i}(n)\right)^2$
From Eq. (A·3), $2\left(\hat{\sigma}^{MC}_{ICI,i}(n)\right)^2$ is given by

$$2\left(\hat{\sigma}^{MC}_{ICI,i}(n)\right)^2 = E\left[\left|\tilde{\mu}^{MC}_{ICI,i}(n)\right|^2\right] = \frac{1}{SF^2}\frac{2E_s}{T_c SF} \times \sum_{k=nSF}^{(n+1)SF-1}\sum_{k'=nSF}^{(n+1)SF-1}\left(\hat{H}_i(k)-A^{MC}_i(n)\right)\left(\hat{H}_i(k')-A^{MC}_i(n)\right)^* \times \sum_{\substack{q'=0\\ \neq q}}^{C-1} c^{q'}(k)\{c^q(k)\}^*\left\{c^{q'}(k')\right\}^* c^q(k')\rho^{q'}_{i-1}(n), \quad (A\cdot 17)$$

where

$$\rho^{q'}_{i-1}(n) = E\left[\left|d^{q'}(n) - \hat{d}^{q'}_{i-1}(n)\right|^2\right]. \quad (A\cdot 18)$$

Since $\hat{d}^{q'}_i(n)$ is an expectation of $d^{q'}(n)$, Eq. (A·18) can be rewritten as

$$\rho^{q'}_{i-1}(n) \approx \left|\bar{d}^{q'}_{i-1}(n)\right|^2 - \left|\hat{d}^{q'}_{i-1}(n)\right|^2. \quad (A\cdot 19)$$

Eq. (A·17) becomes

$$2\left(\hat{\sigma}^{MC}_{ICI,i}(n)\right)^2 = \frac{1}{SF^2}\frac{2E_s}{T_c SF}\left(\sum_{\substack{q'=0\\ \neq q}}^{C-1}\rho^{q'}_{i-1}(n)\right)\sum_{k=nSF}^{(n+1)SF-1}\left|\hat{H}_i(k)-A^{MC}_i(n)\right|^2 + \frac{1}{SF^2}\frac{2E_s}{T_c SF} \times \sum_{k=nSF}^{(n+1)SF-1}\sum_{\substack{k'=nSF\\ \neq k}}^{(n+1)SF-1}\left(\hat{H}_i(k)-A^{MC}_i(n)\right)\left(\hat{H}_i(k')-A^{MC}_i(n)\right)^* \times \sum_{\substack{q'=0\\ \neq q}}^{C-1} c^{q'}(k)\{c^q(k)\}^*\left\{c^{q'}(k')\right\}^* c^q(k')\rho^{q'}_{i-1}(n). \quad (A\cdot 20)$$

The second term tends to 0 for large C and SF, according to the law of large numbers. Finally, $2\hat{\sigma}^2_{ICI,i}(n)$ is given by

$$2\left(\hat{\sigma}^{MC}_{ICI,i}(n)\right)^2 \approx \frac{1}{SF^2}\frac{2E_s}{T_c SF}\sum_{k=nSF}^{(n+1)SF-1}\left|\hat{H}_i(k)-A^{MC}_i(n)\right|^2\left(\sum_{\substack{q'=0\\ \neq q}}^{C-1}\rho^{q'}_{i-1}(n)\right) = \frac{1}{SF}\frac{2E_s}{T_c SF}\left(\sum_{\substack{q'=0\\ \neq q}}^{C-1}\rho^{q'}_{i-1}(n)\right) \times \left(\frac{1}{SF}\sum_{k=nSF}^{(n+1)SF-1}\left|\hat{H}_i(k)\right|^2 - \left|A^{MC}_i(n)\right|^2\right). \quad (A\cdot 21)$$

(2) $2\left(\hat{\sigma}^{MC}_{noise,i}(n)\right)^2$
From Eq. (A·4), $2\left(\hat{\sigma}^{MC}_{noise,i}(n)\right)^2$ is given by

$$2\left(\hat{\sigma}^{MC}_{noise,i}(n)\right)^2 = \frac{1}{SF^2}\sum_{k=nSF}^{(n+1)SF-1}\sum_{k'=nSF}^{(n+1)SF-1} E\left[w_i(k)\Pi(k)\{w_i(k')\Pi(k')\}^*\right]. \quad (A\cdot 22)$$

Since $\{\Pi(k); k = 0 \sim (N_c-1)\}$ are i.i.d. zero-mean complex-valued Gaussian variables with variance $2(N_0/T_c)$, we have

$$2\left(\hat{\sigma}^{MC}_{noise,i}(n)\right)^2 = \frac{1}{SF}\frac{2N_0}{T_c}\left(\frac{1}{SF}\sum_{k=nSF}^{(n+1)SF-1}|w_i(k)|^2\right). \quad (A\cdot 23)$$

(3) $2\left(\hat{\sigma}^{MC}_i(n)\right)^2$
From Eqs. (A·21) and (A·23), $2\left(\hat{\sigma}^{MC}_i(n)\right)^2$ is given by

$$2\left(\hat{\sigma}_i^{MC}(n)\right)^2 = 2\left(\hat{\sigma}_{ICI,i}^{MC}(n)\right)^2 + 2\left(\hat{\sigma}_{noise,i}^{MC}(n)\right)^2$$
$$= \frac{1}{SF^2}\cdot\frac{2E_s}{T_c}$$
$$\times\left[\begin{array}{l}\left(\sum_{\substack{q'=0\\ \neq q}}^{C-1}\rho_{i-1}^{q'}(n)\right)\left\{\frac{1}{SF}\sum_{k=nSF}^{(n+1)SF-1}\left|\hat{H}_i(k)\right|^2-\left|A_i^{MC}(n)\right|^2\right\}\\ +\left(\frac{E_s}{N_0SF}\right)^{-1}\left(\frac{1}{SF}\sum_{k=nSF}^{(n+1)SF-1}|w_i(k)|^2\right)\end{array}\right]. \quad (A\cdot 24)$$

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