IEICE TRANS. COMMUN., VOL.E90–B, NO.5 MAY 2007

PAPER

Frequency-Domain Multi-Stage Soft Interference Cancellation for DS-CDMA Uplink Signal Transmission

Koichi ISHIHARA[†a)], *Member*, **Kazuaki TAKEDA**[†], *Student Member*, *and* **Fumiyuki ADACHI**[†], *Member*

SUMMARY It is well-known that, in DS-CDMA downlink signal transmission, frequency-domain equalization (FDE) based on minimum mean square error (MMSE) criterion can replace rake combining to achieve much improved bit error rate (BER) performance in severe frequency-selective fading channel. However, in uplink signal transmission, as each user's signal goes through a different channel, a severe multi-user interference (MUI) is produced and the uplink BER performance severely degrades compared to the downlink. When a small spreading factor is used, the uplink BER performance further degrades due to inter-chip interference (ICI). In this paper, we propose a frequency-domain multi-stage soft interference cancellation scheme for the DS-CDMA uplink and the achievable BER performance is evaluated by computer simulation. The BER performance comparison of the proposed cancellation technique and the multi-user detection (MUD) is also presented.

key words: *DS-CDMA, frequency-domain equalization (FDE), MUI cancellation, multi-user detection (MUD)*

1. Introduction

Wideband direct sequence code division multiple access (DS-CDMA) with coherent rake combining has been adopted in the 3rd generation mobile communication systems for data transmissions rates up to a few Mbps transmissions [1]. In the next generation mobile communication systems, however, much higher speed data transmission (e.g., close to 1 Gbps) is required. For such high-speed data transmission, the channel becomes severely frequency-selective [2] and the transmission performance significantly degrades due to a large inter-path interference (IPI) even if coherent rake combining is used. Recently, orthogonal frequency division multiplexing (OFDM) and multi-carrier (MC)-CDMA [3]–[5] have been attracting much attention. OFDM uses many low rate orthogonal subcarriers for parallel transmission and applies frequency-domain equalization (FDE) to overcome the frequency-selective channel. MC-CDMA is a combination of OFDM and CDMA, and spreads the data modulated symbol over a number of subcarriers (frequency-domain spreading) to achieve the frequency-diversity gain. However, OFDM and MC-CDMA signals have large peak-to-average power ratio (PAPR). This is a serious problem for the uplink (mobile-to-base) signal transmission since a linear transmit power amplifier with large peak power is required at a mobile station (MS). On the other hand, DS-CDMA does not have such a PAPR problem in general. However, when rake combining is used for the reception of DS-CDMA signals, the transmission performance significantly degrades due to severe IPI. Therefore, a proper equalization technique instead of rake combining is necessary to improve the DS-CDMA transmission performance.

Recently, it was shown [6]–[8] that the use of FDE based on minimum mean square error (MMSE) criterion can significantly improve the bit error rate (BER) performance of DS-CDMA downlink and give similar performance to MC-CDMA. However, in the uplink, each user's signal goes through a different channel and thus, the orthogonality between users is lost; the uplink BER performance severely degrades due to multi-user interference (MUI) [9]–[13] even if FDE is applied. Furthermore, when a small spreading factor is used, the BER performance degrades due to inter-chip interference (ICI) [8]. Many works on the MUI and ICI cancellation are found, e.g., [11], [14]. In [14], MUI cancellation is repeated together with time-domain equalization and decoding. In [11], MUI cancellation and inter-symbol interference (ISI) cancellation are carried out separately; parallel interference cancellation (PIC) is used to cancel MUI and frequency-domain decision-feedback equalization (DFE) is used to cancel ISI.

In this paper, we propose a frequency-domain multi-stage soft PIC (FD-MS-SPIC) scheme and a frequency-domain multi-stage soft successive interference cancellation (FD-MS-SSIC) scheme to simultaneously suppress both MUI and ICI. The remainder of this paper is organized as follows. The DS-CDMA uplink signal transmission is presented in Sect. 2. The FD-MS-SPIC and SSIC are described in Sect. 3. In Sect. 4, the MMSE weight and cancellation weight are derived. Section 5 presents the computer simulation results for the uplink BER performance in a frequency-selective Rayleigh fading channel. The paper is concluded in Sect. 6.

2. DS-CDMA Uplink Signal Transmission

The DS-CDMA uplink transmitter/receiver structure is illustrated in Fig. 1. Throughout the paper, the chip-spaced discrete-time signal representation is used.

We assume that U users are transmitting their signals to the base station (BS). We consider the transmission of one block of N_c chips, where N_c denotes the block length for fast Fourier transform (FFT). At the uth user MS transmit-

Manuscript received December 12, 2005.
Manuscript revised July 26, 2006.

[†]The authors are with the Dept. of Electrical and Communication Engineering, Graduate School of Engineering, Tohoku University, Sendai-shi, 980-8579 Japan.
a) E-mail: ishihara@mobile.ecei.tohoku.ac.jp

DOI: 10.1093/ietcom/e90–b.5.1152

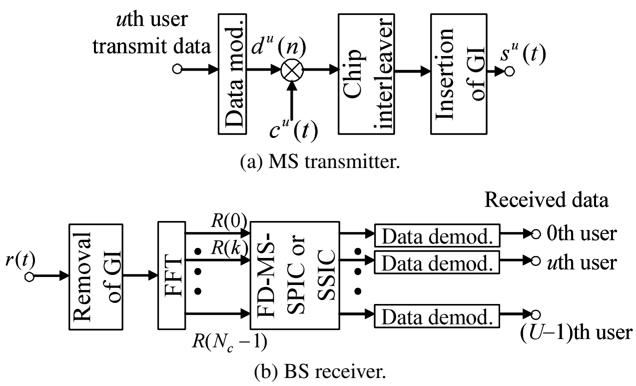


Fig. 1 Uplink transmitter/receiver structure.

ter, u=0~$(U-1)$, a binary data sequence is transformed into data-modulated symbol sequence $d^u(n)$, n=0~$(N_c/SF-1)$, and then spread by multiplying it with a user-specific long pseudo noise (PN) sequence $c^u(t)$, where SF is the spreading factor. The resultant DS-CDMA signal $s^u(t)$, t=0~(N_c-1), can be expressed using the equivalent baseband representation as

$$s^u(t) = \sqrt{\frac{2E_c}{T_c}} d^u\left(\left\lfloor \frac{t}{SF} \right\rfloor\right) \cdot c^u(t), \tag{1}$$

where E_c and T_c represent the chip energy and the chip length, respectively and $\lfloor x \rfloor$ represents the largest integer smaller than or equal to x. A user-specific random chip interleaver is used in order to reduce the negative effect of error propagation due to decision feedback used for the interference replica generation. The last N_g chips of $s^u(t)$ are copied and inserted as a cyclic-prefix into the guard interval (GI) at the beginning of the block [8].

U users' transmitted signals go through different fading channels and are received at the BS receiver. Assuming that each user's fading channel has L independent propagation paths with chip-spaced distinct time delays $\{\tau_l^u; l = 0\sim(L-1)\}$, the impulse response $h^u(t)$ of the uth user's channel can be expressed as [15]

$$h^u(t) = \sum_{l=0}^{L-1} h_l^u \delta(t - \tau_l^u), \tag{2}$$

where h_l^u is the lth path gain with $\sum_{l=0}^{L-1} E[|h_l^u|^2] = 1$ ($E[.]$ is the ensemble mean). In this paper, we assume that the users' transmit timings are asynchronous but they are kept within the GI and we also assume perfect chip timing.

The received signal $r(t)$ can be expressed as

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

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 block fading, where 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 given by

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

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

$$\begin{cases} S^u(k) = \displaystyle\sum_{t=0}^{N_c-1} s^u(t) \exp\left(-j2\pi k \frac{t}{N_c}\right) \\ H^u(k) = \displaystyle\sum_{l=0}^{L-1} h_l^u \exp\left(-j2\pi k \frac{\tau_l^u}{N_c}\right) \\ \Pi(k) = \displaystyle\sum_{t=0}^{N_c-1} \eta(t) \exp\left(-j2\pi k \frac{t}{N_c}\right) \end{cases} . \tag{5}$$

Then, FD-MS-SPIC or SSIC is performed to obtain a sequence of decision variables for data demodulation.

3. FD-MS-SPIC and SSIC

The operation principle is described first for FD-MS-SPIC and then for FD-MS-SSIC. We assume quadrature-phase shift keying (QPSK) data modulation. The cancellation structure at the ith stage is illustrated in Figs. 2 and 3 for FD-MS-SPIC and SSIC, respectively.

3.1 FD-MS-SPIC

First, the uth user's signal replica $\{\bar{S}_{i-1}^u(k); k = 0\sim N_c-1\}$ in the frequency-domain is generated. The replica generation for the uth user is illustrated in Fig. 4. The decision variable for the nth symbol $d^u(n)$ obtained after the $(i-1)$th stage is denoted by $\tilde{d}_{i-1}^u(n)$. The soft symbol replica $\bar{d}_{i-1}^u(n)$ of the uth user is generated by using the decision variable $\tilde{d}_{i-1}^u(n)$.

The log-likelihood ratio (LLR) $\lambda_m^u(n)$ of the mth bit $b_{m,n}^u$ in the nth symbol $d^u(n)$ is computed using the decision variable $\tilde{d}_{i-1}^u(n)$, where m=0~$\log_2 M - 1$ with M being the modulation level. Approximating the residual interference (MUI+ICI) plus noise as a zero-mean complex Gaussian noise variable with variance $2\hat{\sigma}_{i-1}^2$ (see Sect. 4.3), the LLR for $b_{m,n}^u$ is given by [16]

$$\begin{aligned} \lambda_m^u(n) &= \ln\left(\frac{p(b_{m,n}^u = 1)}{p(b_{m,n}^u = 0)}\right) \\ &\approx \ln \frac{\displaystyle\max_{\{d^u: b_{m,n}^u = 1\}} \frac{1}{\sqrt{2\pi\hat{\sigma}_{i-1}^2}} \exp\left(-\frac{1}{2\hat{\sigma}_{i-1}^2}\left|\tilde{d}_{i-1}^u(n) - \sqrt{\frac{2E_c}{T_c}} A_{i-1}^u d^u\right|^2\right)}{\displaystyle\max_{\{d^u: b_{m,n}^u = 0\}} \frac{1}{\sqrt{2\pi\hat{\sigma}_{i-1}^2}} \exp\left(-\frac{1}{2\hat{\sigma}_{i-1}^2}\left|\tilde{d}_{i-1}^u(n) - \sqrt{\frac{2E_c}{T_c}} A_{i-1}^u d^u\right|^2\right)}, \end{aligned} \tag{6}$$

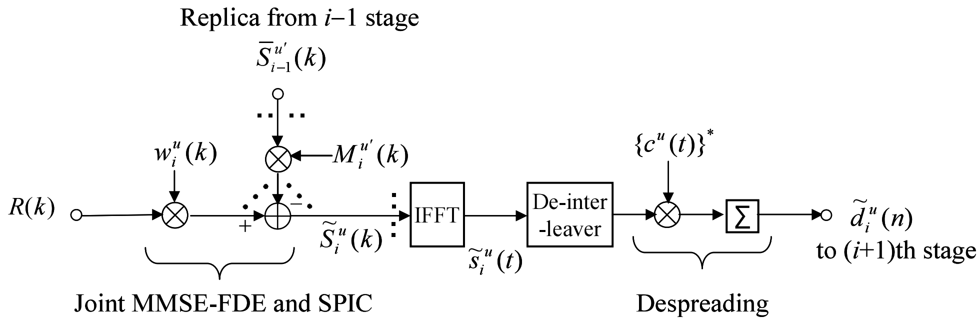


Fig. 2 FD-MS-SPIC structure for the uth user's signal detection at the ith stage.

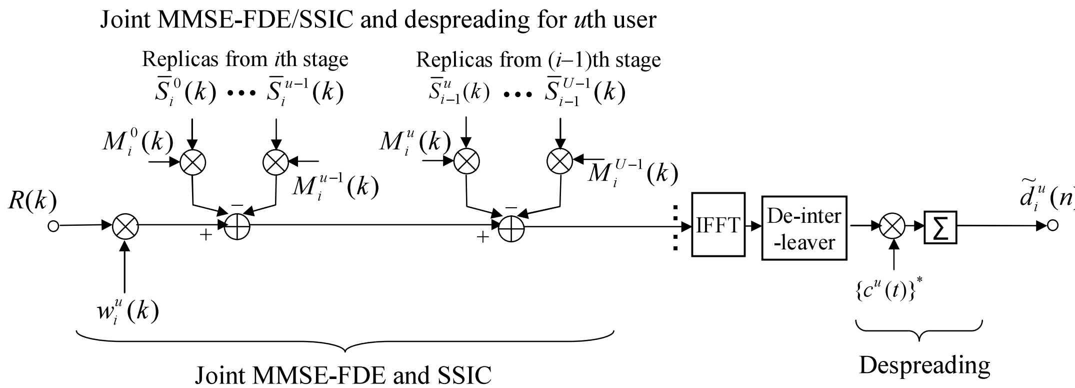


Fig. 3 FD-MS-SSIC structure for the uth user's signal detection at the ith stage.

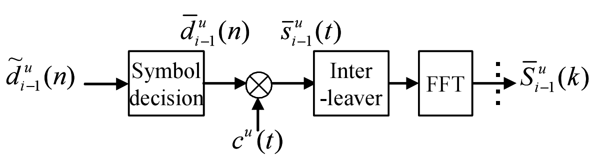


Fig. 4 Replica generator for uth user at the ith stage.

where $p(b^u_{m,n} = 1)$ (or $p(b^u_{m,n} = 0)$) is the probability of $b^u_{m,n} = 1$ (or 0), $\{d^u : b^u_{m,n} = 1 \text{ (or 0)}\}$ denotes a set of symbols whose mth bit is 1 (or 0), and

$$A^u_{i-1} = \frac{1}{N_c}\sum_{k=0}^{N_c-1}\hat{H}^u_{i-1}(k) \tag{7}$$

with

$$\hat{H}^u_{i-1}(k) = w^u_{i-1}(k)H^u(k), \tag{8}$$

where $w^u_{i-1}(k)$ is the MMSE weight, which is derived in Sect. 4.2. The denominator and numerator of Eq. (6) are given as

$$\begin{aligned}&\max_{\{d^u:b^u_{m,n}=1\text{ (or 0)}\}}\\&\left[\frac{1}{\sqrt{2\pi\hat{\sigma}^2_{i-1}}}\exp\left(-\frac{1}{2\hat{\sigma}^2_{i-1}}\left|\tilde{d}^u_{i-1}(n)-\sqrt{\frac{2E_c}{T_c}}A^u_{i-1}d^u\right|^2\right)\right]\\&=\frac{1}{\sqrt{2\pi\hat{\sigma}^2_{i-1}}}\exp\left(-\frac{1}{2\hat{\sigma}^2_{i-1}}\left|\tilde{d}^u_{i-1}(n)-\sqrt{\frac{2E_c}{T_c}}A^u_{i-1}d^{\min}_{b^u_{m,n}=1\text{ (or 0)}}\right|^2\right),\end{aligned} \tag{9}$$

where $d^{\min}_{b^u_{m,n}=1\ (or\ 0)}$ is the most probable symbol whose mth bit is 1 (or 0), for which the Euclidean distance from $\tilde{d}^u_{i-1}(n)$ is minimal. Therefore, Eq. (6) can be approximated as

$$\begin{aligned}\lambda^u_m(n) \approx \frac{1}{2\hat{\sigma}^2_{i-1}}&\left(\left|\tilde{d}^u_{i-1}(n)-\sqrt{\frac{2E_c}{T_c}}A^u_{i-1}d^{\min}_{b^u_{m,n}=0}\right|^2\right.\\&\left.-\left|\tilde{d}^u_{i-1}(n)-\sqrt{\frac{2E_c}{T_c}}A^u_{i-1}d^{\min}_{b^u_{m,n}=1}\right|^2\right).\end{aligned} \tag{10}$$

The soft decision symbol $\bar{d}^u_{i-1}(n)$ can be obtained from [14]

$$\bar{d}^u_{i-1}(n) = \sum_{d\in D} d_{b^u_{m,n}} \prod_{b^u_{m,n}\in d} p(b^u_{m,n}), \tag{11}$$

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

$$\begin{cases} p(b^u_{m,n}=1) = \dfrac{\exp[\lambda^u_m(n)]}{1+\exp[\lambda^u_m(n)]} \\ p(b^u_{m,n}=0) = \dfrac{1}{1+\exp[\lambda^u_m(n)]} \end{cases}. \tag{12}$$

Assuming QPSK data modulation in this paper, the soft symbol replica $\bar{d}^u_{i-1}(n)$ is given as

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

$$+\left(-\frac{1}{\sqrt{2}}+j\frac{1}{\sqrt{2}}\right)p\left(b^u_{0,n}=0\right)p\left(b^u_{1,n}=1\right)$$
$$+\left(-\frac{1}{\sqrt{2}}-j\frac{1}{\sqrt{2}}\right)p\left(b^u_{0,n}=0\right)p\left(b^u_{1,n}=0\right). \tag{13}$$

Substitution of Eq. (12) into Eq. (13) gives

$$\bar{d}^u_{i-1}(n)=\frac{1}{\sqrt{2}}\left[\tanh\left(\frac{\lambda^u_0(n)}{2}\right)+j\tanh\left(\frac{\lambda^u_1(n)}{2}\right)\right]. \tag{14}$$

However, in this paper, to avoid the negative effect of the error propagation from the replica generation, we take a heuristic approach; the parameter β_{i-1} is introduced as

$$\bar{d}^u_{i-1}(n)\approx\frac{1}{\sqrt{2}}\left[\tanh\left(\beta_{i-1}\frac{\lambda^u_0(n)}{2}\right)+j\tanh\left(\beta_{i-1}\frac{\lambda^u_1(n)}{2}\right)\right]. \tag{15}$$

The soft symbol replica $\bar{d}^u_{i-1}(n)$ is re-spread as

$$\bar{s}^u_{i-1}(t)=\sqrt{\frac{2E_c}{T_c}}\bar{d}^u_{i-1}\left(\left\lfloor\frac{t}{SF}\right\rfloor\right)\cdot c^u(t) \tag{16}$$

and then decomposed into N_c frequency components $\{\bar{S}^u_{i-1}(k);\, k=0\sim(N_c-1)\}$ by applying N_c-point FFT. $\bar{S}^u_{i-1}(k)$ is given by

$$\bar{S}^u_{i-1}(k)=\sum_{t=0}^{N_c-1}\bar{s}^u_{i-1}\exp\left(-j2\pi k\frac{t}{N_c}\right), \tag{17}$$

where $\bar{s}^u_{-1}(t)=0$. Then, joint MMSE-FDE and SPIC is carried out, as in Fig. 2, as

$$\tilde{S}^u_i(k)=w^u_i(k)R(k)-M^u_i(k)\bar{S}^u_{i-1}(k)-\sum_{\substack{u'=0\\ \neq u}}^{U-1}M^{u'}_i(k)\bar{S}^{u'}_{i-1}(k), \tag{18}$$

where $M^{u'}_i(k)$ is the cancellation weight (it will be derived in Sect. 4.1), the second term is the ICI cancellation component and the third one is the MUI cancellation component. N_c-point inverse FFT (IFFT) is performed on $\{\tilde{S}^u_i(k);\, k=0\sim(N_c-1)\}$ to recover the uth user's transmitted signal $\tilde{s}^u_i(t)$, which corresponds to Eq. (1), and despreading is carried out to obtain the decision variable $\tilde{d}^u_i(n)$ as

$$\tilde{d}^u_i(n)=\frac{1}{SF}\sum_{t=nSF}^{(n+1)SF-1}\tilde{s}^u_i(t)\cdot\{c^u(t)\}^*. \tag{19}$$

The above procedure is carried out for all U users.

3.2 FD-MS-SSIC

In FD-MS-SSIC, a series of joint MMSE-FDE and SSIC, despreading, symbol decision, replica generation is carried out for all users, according to the ranking of users' received signal powers. The uth user's received signal power P^u is given by

$$P^u=\sum_{k=0}^{N_c-1}|H^u(k)|^2. \tag{20}$$

$\{P^u;\, u=0\sim(U-1)\}$ are compared and users are ranked according to $\{P^u\}$ in the descending order. In this paper, we assume $P^0\geq P^1\geq\ldots\geq P^u\geq\ldots\geq P^{U-1}$ without loss of generality. The replica generation is the same as for SPIC. The soft symbol replicas $\{\bar{d}^{u'}_i(n);\, u'=0\sim(u-1)\}$ and $\{\bar{d}^{u'}_{i-1}(n);\, u'=u\sim(U-1)\}$ are respectively generated from $\{\tilde{d}^{u'}_i(n);\, u'=0\sim(u-1)\}$ and $\{\tilde{d}^{u'}_{i-1}(n);\, u'=u\sim(U-1)\}$ using Eqs. (10) and (15). The symbol replica $\bar{d}^{u'}_i(n)$ (or $\bar{d}^{u'}_{i-1}(n)$) is re-spread and decomposed into N_c frequency components $\{\bar{S}^{u'}_i(k);\, k=0\sim(N_c-1)\}$ (or $\{\bar{S}^{u'}_{i-1}(k);\, k=0\sim(N_c-1)\}$) by applying N_c-point FFT as in Eq. (17).

Joint MMSE-FDE and SSIC for the uth user at the ith stage is carried out, as in Fig. 3, as

$$\begin{aligned}\tilde{S}^u_i(k)&=w^u_i(k)R(k)-M^u_i(k)\bar{S}^u_{i-1}(k)\\&\quad-\left\{\sum_{u'=0}^{u-1}M^{u'}_i(k)\bar{S}^{u'}_i(k)+\sum_{u'=u+1}^{U-1}M^{u'}_i(k)\bar{S}^{u'}_{i-1}(k)\right\},\end{aligned} \tag{21}$$

where the second and third terms are respectively the ICI and MUI components to be cancelled.

Similar to SPIC, N_c-point IFFT is performed on $\{\tilde{S}^u_i(k);\, k=0\sim(N_c-1)\}$ to recover the uth user's transmitted signal $\tilde{s}^u_i(t)$, which corresponds to Eq. (1), and despreading is carried out to obtain the decision variable $\tilde{d}^u_i(n)$ as in Eq. (19).

4. Derivation of MMSE Weight, Cancellation Weight and Interference Plus Noise Variance

4.1 Cancellation Weight

Substitution of Eq. (4) into Eq. (18) gives

$$\begin{aligned}\tilde{S}^u_i(k)&=\hat{H}^u_i(k)S^u(k)+\sum_{\substack{u'=0\\ \neq u}}^{U-1}w^u_i(k)H^{u'}(k)S^{u'}(k)\\&\quad-M^{u'}_i(k)\bar{S}^{u'}_{i-1}(k)-\sum_{\substack{u'=0\\ \neq u}}^{U-1}M^{u'}_i(k)\bar{S}^{u'}_{i-1}(k)+w^u_i(k)\Pi(k)\end{aligned} \tag{22}$$

for SPIC. Substitution of Eq. (4) into Eq. (21) gives

$$\begin{aligned}\tilde{S}^u_i(k)&=\hat{H}^u_i(k)S^u(k)+\sum_{\substack{u'=0\\ \neq u}}^{U-1}w^u_i(k)H^{u'}(k)S^{u'}(k)\\&\quad-M^u_i(k)\bar{S}^u_{i-1}(k)\\&\quad-\left\{\sum_{u'=0}^{u-1}M^{u'}_i(k)\bar{S}^{u'}_i(k)+\sum_{u'=u+1}^{U-1}M^{u'}_i(k)\bar{S}^{u'}_{i-1}(k)\right\}\\&\quad+w^u_i(k)\Pi(k)\end{aligned} \tag{23}$$

for SSIC. The uth user's time-domain signal $\tilde{s}^u_i(t)$ after joint MMSE-FDE and SPIC or SSIC is obtained, by applying N_c-point IFFT to $\{\tilde{S}^u_i(k);\, k=0\sim(N_c-1)\}$, as

$$\tilde{s}^u_i(t)=A^u_i s^u(t)+\mu_{ICI,i}(t)+\mu_{MUI,i}(t)+\tilde{\eta}_i(t), \tag{24}$$

where $\mu_{ICI,i}(t)$ and $\mu_{MUI,i}(t)$ are the residual ICI and MUI components, respectively. They are given by

$$\mu_{ICI,i}(t) = \frac{1}{N_c}\sum_{k=0}^{N_c-1}\left\{\left(\hat{H}_i^u(k) - A_i^u\right)S^u(k) - M_i^u(k)\bar{S}_{i-1}^u(k)\right\} \times \exp\left(j2\pi k\frac{t}{N_c}\right) \tag{25}$$

and

$$\mu_{MUI,i}(t) = \begin{cases} \dfrac{1}{N_c}\displaystyle\sum_{k=0}^{N_c-1}\left\{\sum_{\substack{u'=0\\ \neq u}}^{U-1}\begin{pmatrix} w_i^u(k)H^{u'}(k)S^{u'}(k) \\ -M_i^{u'}(k)\bar{S}_{i-1}^{u'}(k)\end{pmatrix}\right\} \\ \times\exp\left(j2\pi k\frac{t}{N_c}\right) & \text{for SPIC} \\ \dfrac{1}{N_c}\displaystyle\sum_{k=0}^{N_c-1}\left\{\begin{array}{l}\displaystyle\sum_{u'=0}^{u-1}\begin{pmatrix} w_i^u(k)H^{u'}(k)S^{u'}(k) \\ -M_i^{u'}(k)\bar{S}_{i}^{u'}(k)\end{pmatrix} \\ +\displaystyle\sum_{u'=u+1}^{U-1}\begin{pmatrix} w_i^u(k)H^{u'}(k)S^{u'}(k) \\ -M_i^{u'}(k)\bar{S}_{i-1}^{u'}(k)\end{pmatrix}\end{array}\right\} \\ \times\exp\left(j2\pi k\frac{t}{N_c}\right) & \text{for SSIC}\end{cases}. \tag{26}$$

$\tilde{\eta}_i(t)$ is the noise component, given by

$$\tilde{\eta}_i(t) = \frac{1}{N_c}\sum_{k=0}^{N_c-1}\left(w_i^u(k)\Pi(k)\right)\exp\left(j2\pi k\frac{t}{N_c}\right). \tag{27}$$

If $M_i^u(k)$ can be chosen such that $\mu_{ICI,i}(t) = \mu_{MUI,i}(t) = 0$ for all t, then the ICI and MUI cancellation is perfect. Assuming that the replica generation in the $(i-1)$th or ith stage is perfect (i.e., $S^u(k) = \bar{S}_{i-1}^u(k)$ or $S^u(k) = \bar{S}_i^u(k)$), the cancellation weight $M_i^{u'}(k)$ can be obtained, from Eqs. (25) and (26), as

$$M_i^{u'}(k) = \begin{cases} \hat{H}_i^u(k) - A_i^u & \text{if } u' = u \\ w_i^u(k)H^{u'}(k) & \text{otherwise}\end{cases}, \tag{28}$$

which is used for both SPIC and SSIC in this paper.

4.2 MMSE Weight

The equalization error $e_i^u(k)$ is defined as the difference between $\tilde{S}_i^u(k)$ and the uth user's kth frequency signal component $A_i^u S^u(k)$, given by the first term of Eq. (24). Therefore, $e_i^u(k)$ is given by

$$e_i^u(k) = \tilde{S}_i^u(k) - A_i^u S^u(k)$$

$$= \begin{cases} w_i^u(k)\left[R(k) - \displaystyle\sum_{u'=0}^{U-1}H^{u'}(k)\bar{S}_{i-1}^{u'}(k)\right] \\ -A_i^u\left(S^u(k) - \bar{S}_{i-1}^u(k)\right) & \text{for SPIC} \\ w_i^u(k)\left[R(k) - \displaystyle\sum_{u'=0}^{u-1}H^{u'}(k)\bar{S}_i^{u'}(k) - \sum_{u'=u}^{U-1}H^{u'}(k)\bar{S}_{i-1}^{u'}(k)\right] \\ -A_i^u\left(S^u(k) - \bar{S}_{i-1}^u(k)\right) & \text{for SSIC}\end{cases}. \tag{29}$$

Since we assume a white-noise like user-specific long PN sequence $c^u(t)$, we have $E\left[S^u(k)\{S^{u'}(k)\}^*\right] = E\left[S^u(k)\{\bar{S}_{i-1}^{u'}(k)\}^*\right] = E\left[\bar{S}_{i-1}^u(k)\{\bar{S}_{i-1}^{u'}(k)\}^*\right] = 0$ if $u \neq u'$, $\{\Pi(k);\, k = 0\sim(N_c-1)\}$ are identical independent distributed (i.i.d.) zero-mean complex-valued Gaussian variables having variance $2(N_0/T_c)N_c$. Therefore, the MSE $E[|e_i^u(k)|^2]$ for SPIC becomes

$$E[|e_i^u(k)|^2] = \frac{2E_c}{T_c}N_c\left[\left|w_i^u(k)\right|^2\sum_{u'=0}^{U-1}\rho_{i-1}^{u'}\left|H^{u'}(k)\right|^2 + \rho_{i-1}^u\left|A_i^u\right|^2 - 2\mathrm{Re}\left[w_i^u(k)H^u(k)\rho_{i-1}^u A_i^u\right] + \left(\frac{E_c}{N_0}\right)^{-1}\left|w_i^u(k)\right|^2\right], \tag{30}$$

where

$$\rho_{i-1}^{u'} = \left(\frac{2E_c}{T_c}\right)^{-1}\frac{1}{N_c}E\left[\left|S^{u'}(k) - \bar{S}_{i-1}^{u'}(k)\right|^2\right] = \left(\frac{2E_c}{T_c}\right)^{-1}\frac{1}{N_c}\sum_{t=0}^{N_c-1}E\left[\left|s^{u'}(t) - \bar{s}_{i-1}^{u'}(t)\right|^2\right]. \tag{31}$$

Since $\bar{s}_{i-1}^{u'}(t)$ is the expectation of $s^{u'}(t)$ (i.e., $\bar{s}_{i-1}^{u'}(t) = E[s^{u'}(t)]$) and $E[|s^{u'}(t)|^2] = 2E_c/T_c$, we have

$$\rho_{i-1}^{u'} = \left(\frac{2E_c}{T_c}\right)^{-1}\frac{1}{N_c}\sum_{t=0}^{N_c-1}\left\{E\left[\left|s^{u'}(t)\right|^2\right] - \left|\bar{s}_{i-1}^{u'}(t)\right|^2\right\} = \frac{1}{N_c}\sum_{t=0}^{N_c-1}\left\{1 - \left(\frac{2E_c}{T_c}\right)^{-1}\left|\bar{s}_{i-1}^{u'}(t)\right|^2\right\}. \tag{32}$$

The set of MMSE weights $\{w_i^u(k);\, k = 0\sim(N_c-1)\}$ satisfies $\partial E[|e_i(k)|^2]/\partial w_i^u(k) = 0$ for all k. Since

$$\frac{\partial E[|e_i(k)|^2]}{\partial w_i^u(k)} = \frac{2E_c}{T_c}N_c\left[2w_i^u(k)\left\{\sum_{u'=0}^{U-1}\rho_{i-1}^{u'}\left|H^{u'}(k)\right|^2 + \left(\frac{E_c}{N_0}\right)^{-1}\right\} - 2\rho_{i-1}^u A_i^u H^{u*}(k)\right], \tag{33}$$

we have

$$w_i^u(k) = \frac{H^{u*}(k)}{\displaystyle\sum_{u'=0}^{U-1}\rho_{i-1}^{u'}\left|H^{u'}(k)\right|^2 + \left(\frac{E_c}{N_0}\right)^{-1}}. \tag{34}$$

At the initial stage (i=0) of SPIC, Eq. (34) becomes the MMSE weight derived in Ref. [13].

Similar to SPIC, we can show that $w_i^u(k)$ for SSIC is given by

$$w_i^u(k) = \frac{H^{u*}(k)}{\displaystyle\sum_{u'=0}^{u-1}\rho_i^{u'}|H^{u'}(k)|^2 + \sum_{u'=u}^{U-1}\rho_{i-1}^{u'}|H^{u'}(k)|^2 + \left(\frac{E_c}{N_0}\right)^{-1}}. \tag{35}$$

4.3 Interference Plus Noise Variance

The variance $2\hat{\sigma}_i^2$ of interference (MUI+ICI) plus noise is necessary to compute the LLR for soft symbol replica generation (see Sect. 3). The residual MUI can be approximated as a zero-mean complex-valued Gaussian variable according to the central limit theorem [17] since it is a contribution from a number of different users' signals. The residual ICI can also be approximated as another zero-mean complex-valued Gaussian variable since it is a contribution from a large number of chips (i.e., $N_c - 1$ chips). Therefore, the sum of interference (MUI+ICI) and noise can be treated as a new zero-mean complex-valued Gaussian noise. The variance $2\hat{\sigma}_i^2$ is given by

$$2\hat{\sigma}_i^2 = 2\hat{\sigma}_{MUI,i}^2 + 2\hat{\sigma}_{ICI,i}^2 + 2\hat{\sigma}_{noise,i}^2, \tag{36}$$

where $2\hat{\sigma}_{MUI,i}^2$, $2\hat{\sigma}_{ICI,i}^2$ and $2\hat{\sigma}_{noise,i}^2$ are the variances of MUI, ICI and noise, respectively.

From Eq. (26), $2\hat{\sigma}_{MUI,i}^2$ is given by

$$\begin{aligned}2\hat{\sigma}_{MUI,i}^2 &= E\left[\left|\frac{1}{SF}\sum_{t=nSF}^{(n+1)SF-1}\mu_{MUI,i}(t)\{c^u(t)\}^*\right|^2\right]\\ &= \frac{1}{SFN_c}\left(\frac{2E_c}{T_c}\right)\sum_{\substack{u'=0\\ \neq u}}^{U-1}\rho_{i-1}^{u'}\sum_{k=0}^{N_c-1}\left|w_i^u(k)H^{u'}(k)\right|^2,\end{aligned} \tag{37}$$

since we are assuming a white-noise like PN sequence and therefore, $E[c^u(t)\{c^u(\tau)\}^*] = \delta(t-\tau)$ ($\delta(t)$ is the delta function). Here, $\rho_{i-1}^{u'}$ is given by Eq. (32). $2\hat{\sigma}_{ICI,i}^2$ is obtained, from Eq. (25), as

$$\begin{aligned}2\hat{\sigma}_{ICI,i}^2 &= E\left[\left|\frac{1}{SF}\sum_{t=nSF}^{(n+1)SF-1}\mu_{ICI,i}(t)\{c^u(t)\}^*\right|^2\right]\\ &= \frac{1}{SF}\frac{2E_c}{T_c}\rho_{i-1}^u\left[\frac{1}{N_c}\sum_{k=0}^{N_c-1}\left|\hat{H}_i^u(k)\right|^2-\left|A_i^u\right|^2\right],\end{aligned} \tag{38}$$

since $E[c^u(t)\{c^u(\tau)\}^*] = \delta(t-\tau)$. Finally, $2\hat{\sigma}_{noise,i}^2$ is given by [8]

$$2\hat{\sigma}_{noise,i}^2 = \frac{1}{SF}\frac{2N_0}{T_c}\left(\sum_{k=0}^{N_c-1}|w_i^u(k)|^2\right). \tag{39}$$

As a consequence, we have

$$\begin{aligned}2\hat{\sigma}_i^2 &= 2\hat{\sigma}_{MUI,i}^2 + 2\hat{\sigma}_{ICI,i}^2 + 2\hat{\sigma}_{noise,i}^2\\ &= \frac{1}{SF}\cdot\frac{2E_c}{T_c}\left[\frac{1}{N_c}\sum_{\substack{u'=0\\ \neq u}}^{U-1}\left\{\rho_{i-1}^{u'}\sum_{k=0}^{N_c-1}\left|w_i^u(k)H^{u'}(k)\right|^2\right\}\right.\\ &\quad+\rho_{i-1}^u\left\{\frac{1}{N_c}\sum_{k=0}^{N_c-1}\left|\hat{H}_i^u(k)\right|^2-\left|A_i^u\right|^2\right\}\\ &\quad\left.+\left(\frac{E_c}{N_0}\right)^{-1}\left(\frac{1}{N_c}\sum_{k=0}^{N_c-1}|w_i^u(k)|^2\right)\right].\end{aligned} \tag{40}$$

In a similar manner to SPIC, we can derive $2\hat{\sigma}_i^2$ as

$$\begin{aligned}2\hat{\sigma}_i^2 &= \frac{1}{SF}\cdot\frac{2E_c}{T_c}\left[\frac{1}{N_c}\sum_{u'=0}^{u-1}\left\{\rho_i^{u'}\sum_{k=0}^{N_c-1}\left|w_i^u(k)H^{u'}(k)\right|^2\right\}\right.\\ &\quad+\frac{1}{N_c}\sum_{u'=u+1}^{U-1}\left\{\rho_{i-1}^{u'}\sum_{k=0}^{N_c-1}\left|w_i^u(k)H^{u'}(k)\right|^2\right\}\\ &\quad+\rho_{i-1}^u\left\{\frac{1}{N_c}\sum_{k=0}^{N_c-1}\left|\hat{H}_i^u(k)\right|^2-\left|A_i^u\right|^2\right\}\\ &\quad\left.+\left(\frac{E_c}{N_0}\right)^{-1}\left(\frac{1}{N_c}\sum_{k=0}^{N_c-1}|w_i^u(k)|^2\right)\right]\\ &\qquad\qquad\text{for SSIC.}\end{aligned} \tag{41}$$

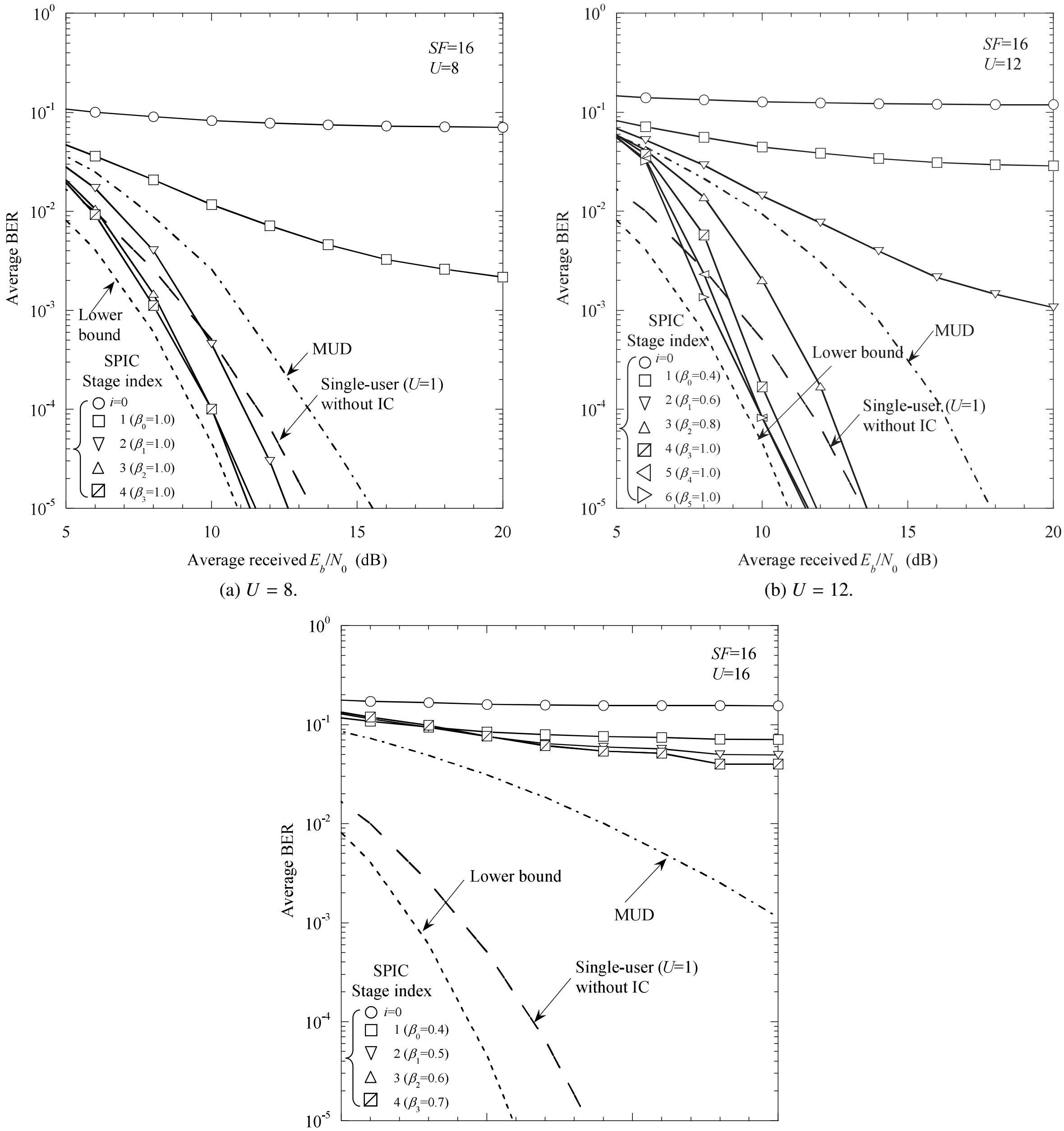
5. Computer Simulation Results

Table 1 shows the computer simulation conditions. We assume an FFT block length of N_c=256 chips, a GI length of N_g=32 chips, and QPSK data-modulation. A chip-spaced 16-path (L=16) frequency-selective block Rayleigh fading channel having a uniform power delay profile (i.e., $E[|h_l^u|^2] = 1/L$ for all l) are assumed. The path gains stay unchanged, but vary block by block. In [18], how the variation of the path gains during a block affects the BER performance is discussed. According to [18], the BER performance is almost insensitive to the fading rate if the normalized maximum Doppler frequency $f_D(N_c + N_g)T_c$ is smaller than 0.01 (this corresponds to a terminal moving speed of 750 km/h for a chip rate $1/T_c$ of 100 Mchip/s and a carrier frequency of 5 GHz for an FFT size of N_c=256 chips and a GI length of N_g=32 chips). Therefore, we assume block fading.

In the next generation mobile communication systems, much higher speed data transmissions than the present systems are required. To achieve high speed data transmissions

Table 1 Simulation conditions.

Transmitter	Modulation	QPSK
	FFT block length	N_c=256 chips
	GI length	N_g=32 chips
	Spreading sequence	Long PN sequence
	Spreading factor	SF=16
	Number of users	U=8~16
Channel	Fading	Frequency-selective block Rayleigh fading
	Number of paths	L=16 paths
	Power delay profile	Uniform
Receiver	Frequency-domain equalization	MMSE
	Channel estimation	Ideal

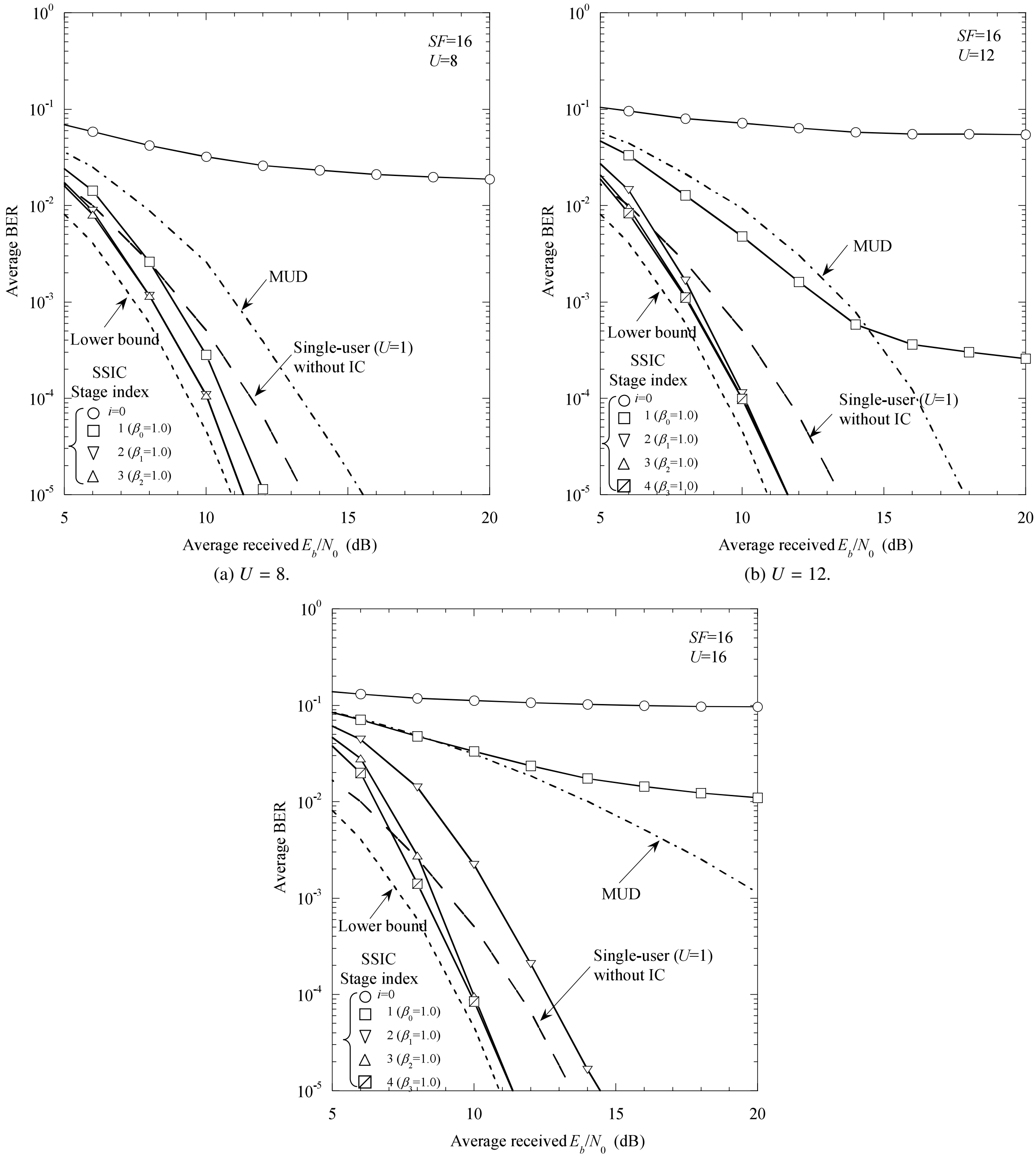


(a) U = 8. (b) U = 12. (c) U = 16.

Fig. 5 Uplink BER performance using FD-MS-SPIC.

for the given chip rate (or bandwidth), the spreading factor *SF* must be small. In Ref. [19], *SF*=16 is used in uplink packet data transmission. In the following computer simulation, only *SF*=16 is considered. Users' transmit timings are asynchronous but they are assumed to be kept within the GI and to be perfect chip timing by transmit timing control. We assume ideal channel estimation.

For the soft symbol replica generation, the parameter β_i was introduced in Eq. (15). As $\beta_i \to 0$, the soft decision chip replica becomes zero and the BER performance approaches that without interference cancellation (IC). On the other hand, as $\beta_i \to 1$, the accuracy of chip replica may deteriorate due to the error propagation. Therefore, there exists an optimum value in β_i, which was found by computer simulation in this paper. We use a random interleaver in order to reduce the negative effect of error propagation due to decision feedback. We have found by computer simulation that the interleaver size to sufficiently improve the BER performance is *SF*-by-N_c chips. With chip interleaver, the decision error is scattered over an interval of *SF*-by-N_c chips and more reliable ICI replica can be generated (without chip interleaver, re-spreading of an erroneous symbol



(a) U = 8. (b) U = 12. (c) U = 16.

Fig. 6 Uplink BER performance using FD-MS-SSIC.

replica produces the error propagation over consecutive *SF* chips).

Figures 5 and 6 show the average BER performance as a function of the average received signal energy per bit-to-the AWGN power spectrum density ratio E_b/N_0 (=$0.5SF(1+N_g/N_c)(E_c/N_0)$). The optimum parameter β_i found by the computer simulation is shown in the figures. Also plotted in the figures for comparison are the theoretical lower bound BER performance [8], the single-user (U=1) performance without IC (therefore, the residual ICI remains intact) and the BER performance with multi-user detection (MUD) [12]. The lower bound BER performance is obtained by assuming maximum ratio combining (MRC)-FDE and neglecting the ICI for the single-user case (U=1). The reason for assuming MRC-FDE instead of MMSE-FDE is that MRC-FDE achieves the highest signal-to-noise power ratio (SNR).

In this paper, we have assumed the same average received signal power for all users, but they have different *instantaneous* received signal powers. In SSIC, users are ranked according to their *instantaneous* received signal powers. It was found by computer simulation that when

$i \leq 3$, the BER performance with ranking is better than that without ranking; however, when i=4, almost the same BER performance is achieved with and without ranking. In Fig. 6, the BER performance of SSIC with ranking is plotted.

Without IC (i.e., FD-MS-PIC with i=0), the BER performance is severely degraded due to a severe MUI. However, it can be seen that the proposed FD-MS-SPIC/SSIC scheme can significantly improve the BER performance under multi-user environment.

When $U = 8$ (see Figs. 5(a) and 6(a)), even with i=3rd (2nd) stage for SPIC (SSIC), the average BER performance can be significantly improved and is better than with MUD. Furthermore, the BER performance with FD-MS-SPIC/SSIC is better than that of single-user case (without IC). This is because SPIC and SSIC can sufficiently suppress the MUI and ICI while the ICI remains intact in the single-user case. The E_b/N_0 degradation for BER=10^{-3} from the lower bound performance becomes as small as 0.6 dB.

For full load condition of U/SF=1 (i.e., U=16), SPIC cannot improve the BER performance because of the negative effect of the error propagation due to decision feedback. On the other hand, SSIC with i=4 SSIC can achieve a BER performance close to the lower bound. In SSIC, the interference cancellation is more reliable since it is done in the descending order of the received signal power.

As the number of users increases, the MUI becomes severer, and hence the number of stages necessary to achieve a good BER performance increases. For FD-MS-SSIC, a sufficient number of iterations for improving the BER performance is found, from Fig. 6, to be i=2, 3, and 4 for U=8, 12, and 16, respectively. On the other hand, for FD-MS-SPIC, it is found, from Fig. 5, to be i=4 and 6 for U=8 and 12, respectively; however, when U=16, SPIC cannot provide a good BER performance because of the negative effect of the error propagation due to decision feedback.

In this paper, we have assumed the same *average* received signal power for all users; this is equivalent to the use of slow transmit power control (TPC). We have also examined the BER performance assuming the same *instantaneous* received signal power for all users (equivalent to the use of fast TPC) and found that although all users are ranked the same, SIC provides better BER performance than PIC (for the sake of brevity, the BER performance with fast TPC are not shown here). A possible reason for this is as follows. In SIC, MUI replica of an interfering user is subtracted one by one from the received signal. However, in PIC, the MUI replicas of all interfering users are subtracted at once from the received signal, thereby large error propagation due to feedback is sometimes produced.

6. Conclusion

In this paper, we proposed a frequency-domain multi-stage soft parallel interference cancellation (FD-MS-SPIC) scheme and a frequency-domain multi-stage soft successive interference cancellation (FD-MS-SSIC) scheme in DS-CDMA uplink signal transmission. We have derived the MMSE weight and IC weight taking into account the residual inter-chip interference (ICI) and multi-user interference (MUI). The MMSE weight, cancellation weight, and the soft interference replica generation weight are updated in each cancellation stage. The BER performances with the proposed FD-MS-SPIC/SSIC scheme in a frequency-selective Rayleigh fading were evaluated by computer simulation. When the channel is lightly loaded (e.g., U/SF=0.5), both SPIC and SSIC work well and a performance close to the lower bound can be achieved at the i=3rd stage for SPIC and the i=2nd stage for SSIC. The average BER performance with the proposed scheme is better than with multi-user detection (MUD) and that of the single-user case (without IC). This is because SPIC/SSIC can sufficiently suppress the MUI and ICI. However, when the channel is fully loaded (U/SF=1), SPIC cannot work, but SSIC with the i=4th stage provides a good BER performance.

In this paper, by assuming ideal channel estimation, we have shown that the proposed IC schemes can significantly improve the DS-CDMA uplink performance in a multi-user environment. However, the BER performance with IC is sensitive to the channel estimation error. The impact of the channel estimation error is left as an important future study.

References

[1] F. Adachi, M. Sawahashi, and H. Suda, "Wideband DS-CDMA for next generation mobile communication systems," IEEE Commun. Mag., vol.36, no.9, pp.56–69, Sept. 1998.
[2] W.C. Jakes Jr., ed., Microwave mobile communications, Wiley, New York, 1974.
[3] M. Okada, S. Hara, and N. Morinaga, "Bit error performance of orthogonal multicarrier modulation radio transmission system," IEICE Trans. Commun., vol.E76-B, no.2, pp.113–119, Feb. 1993.
[4] S. Hara and R. Prasad, "Overview of multicarrier CDMA," IEEE Commun. Mag., vol.35, no.12, pp.126–133, Dec. 1997.
[5] L. Hanzo, W. Webb, and T. Keller, Single- and multi-carrier quadrature amplitude modulation, John Wiley & Sons, 2000.
[6] F. Adachi, D. Garg, S. Takaoka, and K. Takeda, "Broadband CDMA techniques," IEEE Wireless Commun. Mag., vol.2, no.2, pp.8–18, April 2005.
[7] F.W. Vook, T.A. Thomas, and K.L. Baum, "Cyclic-prefix CDMA with antenna diversity," Proc. IEEE VTC'02-Spring, pp.1002–1006, May 2002.
[8] F. Adachi and K. Takeda, "Bit error rate analysis of DS-CDMA with joint frequency-domain equalization and antenna diversity combining," IEICE Trans. Commun., vol.E87-B, no.10, pp.2991–3002, Oct. 2004.
[9] S. Moshavi, "Multi-user detection for DS-CDMA communications," IEEE Commun. Mag., vol.34, no.10, pp.124–136, Oct. 1996.
[10] F. Berggren and S.B. Slimane, "Linear successive interference cancellation in DS-CDMA systems," Wirel. Commun. Mob. Comput., vol.3, no.7, pp.847–859, Nov. 2003.
[11] S. Tomasin and N. Benvenuto, "Equalization and multiuser interference cancellation in CDMA systems," Proc. 6th International Symposium on WPMC, vol.1, pp.10–14, Yokosuka, Japan, Oct. 2003.
[12] S. Tsumura, S. Hara, and Y. Hara, "Performance comparison of MC-CDMA and cyclically prefixed DS-CDMA in an uplink channel," Proc. 60th IEEE Veh. Technol. Conf. (VTC), Los Angeles, CA, Sept. 2004.
[13] K. Takeda and F. Adachi, "Performance evaluation of multi-rate DS-CDMA using frequency-domain equalization in a frequency-selective fading channel," IEICE Trans. Commun., vol.E88-B, no.3,

pp.1191–1201, March 2005.
[14] X. Wang and H.V. Poor, "Iterative (turbo) soft interference cancellation and decoding for coded CDMA," IEEE Trans. Commun., vol.47, no.7, pp.1046–1061, July 1999.
[15] T.S. Rappaport, Wireless Communications, Prentice Hall, 1996.
[16] A. Stefanov and T. Duman, "Turbo coded modulation for wireless communications with antenna diversity," Proc. IEEE VTC'99 Fall, pp.1565–1569, Netherlands, Sept. 1999.
[17] J.G. Proakis, Digital communications, 3rd ed., McGraw-Hill, 1995.
[18] K. Takeda, T. Itagaki, and F. Adachi, "Joint use of frequency-domain equalization and transmit/receive antenna diversity for single-carrier transmissions," IEICE Trans. Commun., vol.E87-B, no.7, pp.1946–1953, July 2004.
[19] 3GPP TR25.804, "Feasibility study on uplink enhancements for UTRA TDD," version 6.0.0.

Koichi Ishihara received his B.E. and M.E. degrees in communications engineering from Tohoku University, Sendai, Japan, in 2004 and 2006, respectively. Since April 2006, he has been with NTT Network Innovation Laboratories, Nippon Telegraph & Telephone Corporation (NTT), Yokosuka, Japan. His research interests include equalization, interference cancellation, and multiple access techniques for broadband wireless communication systems. He is a member of IEEE.

Kazuaki Takeda received his B.E. and M.S. degrees in communications engineering from Tohoku University, Sendai, Japan, in 2003 and 2004. Currently he is a PhD student at the Department of Electrical and Communications Engineering, Graduate School of Engineering, Tohoku University. His research interests include equalization, interference cancellation, transmit/receive diversity, and multiple access techniques. He was a recipient of the 2003 IEICE RCS (Radio Communication Systems) Active Research Award and 2004 Inose Scientific Encouragement Prize.

Fumiyuki Adachi received the B.S. and Dr.Eng. degrees in electrical engineering from Tohoku University, Sendai, Japan, in 1973 and 1984, respectively. In April 1973, he joined the Electrical Communications Laboratories of Nippon Telegraph & Telephone Corporation (now NTT) and conducted various types of research related to digital cellular mobile communications. From July 1992 to December 1999, he was with NTT Mobile Communications Network, Inc. (now NTT DoCoMo, Inc.), where he led a research group on wideband/broadband CDMA wireless access for IMT-2000 and beyond. Since January 2000, he has been with Tohoku University, Sendai, Japan, where he is a Professor of Electrical and Communication Engineering at the Graduate School of Engineering. His research interests are in CDMA wireless access techniques, equalization, transmit/receive antenna diversity, MIMO, adaptive transmission, and channel coding, with particular application to broadband wireless communications systems. From October 1984 to September 1985, he was a United Kingdom SERC Visiting Research Fellow in the Department of Electrical Engineering and Electronics at Liverpool University. Dr. Adachi served as a Guest Editor of IEEE JSAC on Broadband Wireless Techniques, October 1999, Wideband CDMA I, August 2000, Wideband CDMA II, Jan. 2001, and Next Generation CDMA Technologies, Jan. 2006. He is an IEEE Fellow and was a co-recipient of the IEEE Vehicular Technology Transactions Best Paper of the Year Award 1980 and again 1990 and also a recipient of Avant Garde award 2000. He was a recipient of IEICE Achievement Award 2002 and a co-recipient of the IEICE Transactions Best Paper of the Year Award 1996 and again 1998. He was a recipient of Thomson Scientific Research Front Award 2004.