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Transmit/Receive Antenna Diversity for Frequency-interleaved Single-carrier Spread Spectrum Multi-access with Frequency-domain Equalization

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Abstract— **Frequency-domain equalization (FDE) based on the minimum mean square error (MMSE) criterion can significantly improve the downlink bit error rate (BER) performance of DS-CDMA in a frequency-selective fading channel. However, the uplink BER performance degrades due to a strong multi-user interference (MUI). We have proposed frequency-interleaved single-carrier spread spectrum multi-access (SC-SSMA) with MMSE-FDE, in which the subcarrier components of each user's signal are interleaved onto a wider bandwidth. Since frequency-interleaving patterns assigned to different users are orthogonal to each other, frequency-interleaved SC-SSMA can avoid the MUI completely while achieving frequency diversity gain with MMSE-FDE. In this paper, transmit/receive antenna diversity is applied to frequency-interleaved SC-SSMA with MMSE-FDE to further improve the BER performance. Its conditional BER is derived for the given set of channel gains. The average BER performance is numerically evaluated using the derived conditional BER and is confirmed by computer simulation.**

Keywords-component; Frequency-interleaving, MMSE-FDE, space-time transmit diversity, receive antenna diversity

I. INTRODUCTION

With the growing mobile wireless communication market, there has been tremendous demand for high-speed data transmissions [1]. A mobile communication channel is composed of many distinct propagation paths having different time delays, resulting in a frequency-selective fading channel [2]. Direct sequence code division multiple access (DS-CDMA) using a coherent rake combining is adopted in the present cellular mobile communication systems for data transmissions of around a few Mbps [3]. However, the wireless channel for the data transmissions higher than few tens of Mbps becomes severely frequency-selective and the BER performance with rake combining degrades due to a strong inter-path interference. Hence, the use of some advanced channel equalization technique is indispensable.

It was shown [4~7] that frequency-domain equalization based on the minimum mean square error criterion (MMSE-FDE) can replace rake combining to significantly improve the downlink BER performance of DS-CDMA. However, for DS--CDMA uplink transmission, different user's signal goes through different propagation channels and hence, the BER floor is produced due to strong multi-user interference (MUI) even if MMSE-FDE is applied [8]. Quite recently, chip repetition DS-CDMA has been proposed that uses comb-like spectrum to avoid the spectrum overlapping among different users [9, 10]. We have also proposed frequency-interleaved single-carrier spread spectrum multi-access (SC-SSMA) with MMSE-FDE to avoid the MUI completely and to improve the uplink performance [11]. In frequency-interleaved SC-SSMA, the subcarrier components of each user's SS signal are interleaved, using orthogonal interleaving patterns, onto a wider bandwidth. Since, orthogonal frequency-interleaving patterns are assigned to different users, frequency-interleaved SC-SSMA can avoid MUI completely while making full use of the frequency diversity gain owing to MMSE-FDE.

In this paper, to further improve the uplink BER performance, transmit/receive antenna diversity technique based on space-time transmit diversity (STTD) [12], [13] is applied to frequency-interleaved SC-SSMA with MMSE-FDE. The conditional BER is derived for the given set of channel gains. The average BER performance is numerically evaluated using the derived conditional BER and is confirmed by computer simulation.

II. TRANSMIT/RECEIVE ANTENNA DIVERSITY FOR FREQUENCY-INTERLEAVED SC-SSMA WITH MMSE-FDE

A. Transmit signal

Fig. 1 shows the uplink transmitter/receiver for frequency-interleaved SC-SSMA with MMSE-FDE. Throughout this paper, the chip-spaced time representation of transmit signals is used. We assume that U users are transmitting their data to the base station. At the uth user's (u=0~U−1) mobile transmitter, a binary data sequence is transformed into a data modulated symbol sequence and is divided into a sequence of blocks of N_c/SF_t symbols each, where N_c is the FFT window size and SF_t is the spreading factor. Without loss of generality, the transmission of the data symbol sequence {$d_{e(o)}^{(u)}(q)$; q=0~N_c/SF_t-1} in one even (odd) block is considered. The symbol sequence is spread with a spreading sequence {$c^{(u)}(t)$; t=..,−1,0,1,..} of spreading factor SF_t. The resulting chip sequence in the even (odd) block is expressed, using vector representation, as $\mathbf{s}_{e(o)}^{(u)}=[s_{e(o)}^{(u)}(0),...,s_{e(o)}^{(u)}(t),\ldots,s_{e(o)}^{(u)}(N_c-1)]^{\mathrm{T}}$, where the superscript T denotes the transposition. $s_{e(o)}^{(u)}(t)$ can be expressed, using the equivalent lowpass representation, as

$$s_{e(o)}^{(u)}(t)=\sqrt{E_s/(T_c\,SF_t)}\,d_{e(o)}^{(u)}\left(\lfloor t/SF_t\rfloor\right)c(t)\quad ,(1)$$

where E_s and T_c denote the symbol energy and the chip duration, respectively.

$\mathbf{s}_{e(o)}^{(u)}$ is decomposed by N_c-point FFT into frequency-domain signal $\mathbf{S}_{e(o)}^{(u)}=[S_{e(o)}^{(u)}(0),...,S_{e(o)}^{(u)}(k),\ldots,S_{e(o)}^{(u)}(N_c-1)]^{\mathrm{T}}$, where $S_{e(o)}^{(u)}(k)$ is the kth subcarrier component, given by

$$S_{e(o)}^{(u)}(k)=\sum_{t=0}^{N_c-1}s_{e(o)}^{(u)}(t)\exp\left(-j2\pi k\frac{t}{N_c}\right)\quad .(2)$$

$\mathbf{S}_e^{(u)}$ and $\mathbf{S}_o^{(u)}$ are ST encoded as shown in Fig. 2:

$$\begin{pmatrix}\mathbf{S}_{0,0}^{(u)} & \mathbf{S}_{0,1}^{(u)}\\ \mathbf{S}_{1,0}^{(u)} & \mathbf{S}_{1,1}^{(u)}\end{pmatrix}=\begin{pmatrix}\mathbf{S}_e^{(u)} & \mathbf{S}_o^{(u)}\\ -\mathbf{S}_o^{(u)*} & \mathbf{S}_e^{(u)*}\end{pmatrix}\ ,(3)$$

where $\mathbf{S}_{p,n}^{(u)}$ is the frequency-domain STTD-encoded signal in the pth block (p=0,1) to be transmitted from the nth antenna (n=0,1). The frequency-domain ST-encoded signals are interleaved onto SF_f times wider bandwidth of N_cSF_f frequency components. The resulting frequency-interleaved signals can be represented as

$$\begin{aligned}\hat{\mathbf{S}}_{p,n}^{(u)}&=[\hat{S}_{p,n}^{(u)}(0),...,\hat{S}_{p,n}^{(u)}(k'),\ldots,\hat{S}_{p,n}^{(u)}(N_cSF_f-1)]^{\mathrm{T}}\\&=\mathbf{Q}^{(u)}\mathbf{S}_{p,n}^{(u)}\end{aligned}\quad ,(4)$$

where $\mathbf{Q}^{(u)}$ is the (N_cSF_f)-by-N_c frequency-interleaving matrix.

Interleaving patterns are determined so that different user's subcarrier components do not overlap with each other. $\mathbf{Q}^{(u)}$ must satisfy

$$\{\mathbf{Q}^{(u)}\}^T\mathbf{Q}^{(u')}=\begin{cases}\mathbf{I} & \text{if } u=u'\\ \mathbf{0} & \text{otherwise}\end{cases}\ ,\ (5)$$

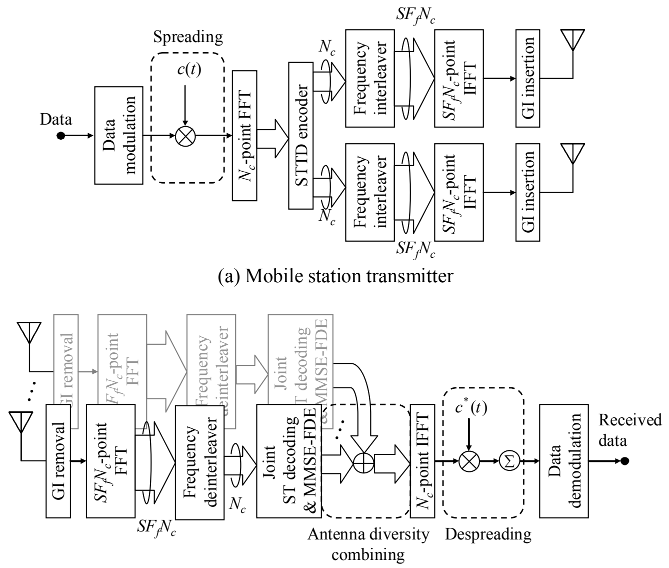
where $\mathbf{I}$ is an $N_c\times N_c$ identity matrix. Below, an example is shown for the case of SF_f=2 and N_c=2 for multiplexing two users (u=0 and 1). The following interleaving matrices, $\mathbf{Q}^{(0)}$ and $\mathbf{Q}^{(1)}$, can be used:

$$\mathbf{Q}^{(0)}=\begin{pmatrix}0&0\\1&0\\0&1\\0&0\end{pmatrix},\quad \mathbf{Q}^{(1)}=\begin{pmatrix}0&1\\0&0\\0&0\\1&0\end{pmatrix}\ .\ (6)$$

We can see that the positions of "1" for $\mathbf{Q}^{(0)}$ and $\mathbf{Q}^{(1)}$ are not overlapping and therefore $\mathbf{Q}^{(0)}$ and $\mathbf{Q}^{(1)}$ are orthogonal to each other.

Finally, N_cSF_f-point IFFT is applied to obtain the STTD-encoded and frequency-interleaved SC-SSMA signal $\{\tilde{s}_{p,n}^{(u)}(t')$; t'=0~(N_cSF_f–1)} which can be expressed as

$$\tilde{s}_{p,n}^{(u)}(t')=\frac{1}{N_c}\sum_{k'=0}^{N_cSF_f-1}\hat{S}_{p,n}^{(u)}(k')\exp\left(j2\pi k'\frac{t'}{N_cSF_f}\right)\quad .(7)$$



(a) Mobile station transmitter

(b) Base station receiver

Figure 1. Transmission system model.

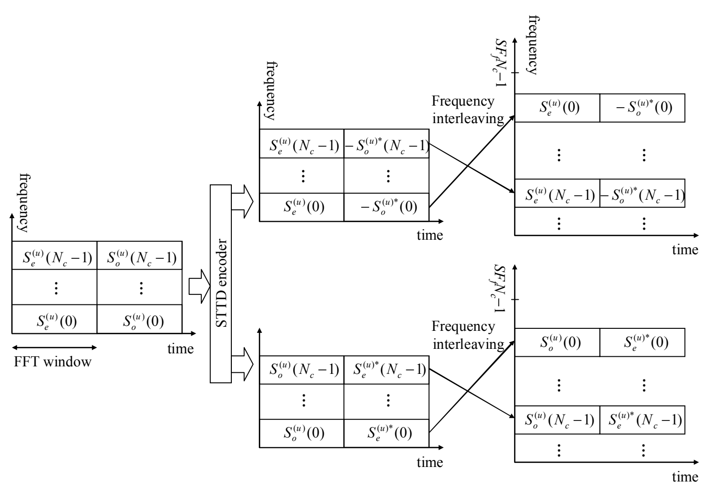


Figure 2. Frequency-domain ST encoding and frequency-interleaving.

B. Received signal representation

We assume a block fading so that the path gains stay constant over two consecutive block length of 2(N_cSF_f+N_g) chips. Assuming that the channel has L independent propagation paths with T_c-spaced distinct time delays {$\tau_l^{(u)}$; l=0~L–1}, the multipath channel impulse response $h_{n,m}^{(u)}(t')$ between the nth transmit antenna (n=0, 1) of the uth user and the mth receive antenna (m=0~M–1) is expressed as [14]

$$h_{n,m}^{(u)}(t')=\sum_{l=0}^{L-1}h_{n,m,l}^{(u)}\delta(t'-\tau_l^{(u)})\ ,\ (8)$$

where $h_{n,m,l}^{(u)}$ is the lth path gain with $\sum_{l=0}^{L-1}E[|h_{n,m,l}^{(u)}|^2]=1$ (E[.] denotes the ensemble average operation). It is assumed that the maximum time delay is shorter than the GI length for all users.

The pth (p=0, 1) received signal block $r_{p,m}(t')$, $t'=-N_g\sim(N_cSF_f-1)$, at the mth receive antenna of the base station is expressed as

$$r_{p,m}(t')=\sum_{u=0}^{U-1}\sum_{n=0}^{1}\sum_{l=0}^{L-1}h_{n,m,l}^{(u)}\tilde{s}_{p,n}^{(u)}(t'-\tau_l^{(u)})+\eta_{p,m}(t'),\ (9)$$

where $\eta_{p,m}(t')$ is a zero-mean complex Gaussian noise process, in the pth block, having a variance of $2N_0SF_f/T_c$ with N_0 being the single-sided power spectrum density of the additive white Gaussian noise (AWGN).

C. Frequency-deinterleaving and joint STTD decoding and MMSE-FDE

The received signal $r_{p,m}(t')$ is decomposed by N_cSF_f-point FFT into N_cSF_f frequency components $\mathbf{R}_{p,m}=[R_{p,m}(0),...,R_{p,m}(k'),...,R_{p,m}(N_cSF_f-1)]^T$. $R_{p,m}(k')$ is given by

$$R_{p,m}(k')=\sum_{t'=0}^{N_cSF_f-1}r_{p,m}(t')\exp\left(-j2\pi k'\frac{t'}{N_cSF_f}\right).\ (10)$$

Without loss of generality, the detection of only the 0th (u=0) user's data sequence is considered. $\mathbf{R}_{p,m}$ is deinterleaved to pick up the 0th user's frequency-domain signal $\tilde{\mathbf{R}}_{p,m}^{(0)}=[\tilde{R}_{p,m}^{(0)}(0),...,\tilde{R}_{p,m}^{(0)}(k),...,\tilde{R}_{p,m}^{(0)}(N_c-1)]^{\mathrm{T}}$ as [11]

$$\begin{aligned}\tilde{\mathbf{R}}_{p,m}^{(0)}&=\mathbf{Q}^{(0)^T}\mathbf{R}_{p,m}\\&=SF_f\sum_{u=0}^{U-1}\sum_{n=0}^{1}\left\{\mathbf{Q}^{(0)^T}\mathbf{H}_{n,m}^{(u)}\mathbf{Q}^{(u)}\right\}\mathbf{S}_{p,n}^{(u)}+\mathbf{Q}^{(0)^T}\mathbf{\Pi}_{p,m}\end{aligned},\ (11)$$

where $\mathbf{H}_{n,m}^{(u)}=diag\left(H_{n,m}^{(u)}(0),...,H_{n,m}^{(u)}(k'),...,H_{n,m}^{(u)}(N_cSF_f-1)\right)$ is (N_cSF_f)-by-(N_cSF_f) channel gain matrix and $\mathbf{\Pi}_{p,m}$ is (N_cSF_f)-by-1 noise vector. Since the frequency-interleaving matrices $\mathbf{Q}^{(u)}$'s are orthogonal (see Eq. (5)) and $\mathbf{H}_{n,m}^{(u)}$ is the diagonal matrix, $\mathbf{Q}^{(0)^T}\mathbf{H}_{n,m}^{(u)}\mathbf{Q}^{(u)}$ satisfies

$$\mathbf{Q}^{(0)^{\mathrm{T}}}\mathbf{H}_{n,m}^{(u)}\mathbf{Q}^{(u)}=\begin{cases}\mathbf{Q}^{(0)^{\mathrm{T}}}\mathbf{H}_{n,m}^{(0)}\mathbf{Q}^{(0)} & \text{if } u=0\\ \mathbf{0} & \text{otherwise}\end{cases}.\ (12)$$

Hence, we obtain

$$\tilde{\mathbf{R}}_{p,m}^{(0)}=SF_f\sum_{n=0}^{1}\tilde{\mathbf{H}}_{n,m}^{(0)}\mathbf{S}_{p,n}^{(0)}+\tilde{\mathbf{\Pi}}_{p,m},(13)$$

where $\tilde{\mathbf{H}}_{n,m}^{(0)}=diag\left(\tilde{H}_{n,m}^{(0)}(0),...,\tilde{H}_{n,m}^{(u)}(k),...,\tilde{H}_{n,m}^{(0)}(N_c-1)\right)$ and $\tilde{\mathbf{\Pi}}_{p,m}^{(0)}=\left[\tilde{\Pi}_{p,m}^{(0)}(0),...,\tilde{\Pi}_{p,m}^{(0)}(k),...,\tilde{\Pi}_{p,m}^{(0)}(N_c-1)\right]^T$ are the frequency-deinterleaved channel gain and the noise, given by

$$\begin{cases}\tilde{\mathbf{H}}_{n,m}^{(0)}=\mathbf{Q}^{(0)^T}\mathbf{H}_{n,m}^{(0)}\mathbf{Q}^{(0)}\\ \tilde{\mathbf{\Pi}}_{p,m}^{(0)}=\mathbf{Q}^{(0)^T}\mathbf{\Pi}_{p,m}\end{cases}.\ (14)$$

From Eqs. (3) and (13), $\tilde{R}_{p,m}^{(0)}(k)$ is given for p=0 and 1, as

$$\begin{cases}\tilde{R}_{0,m}^{(0)}(k)=SF_f\tilde{H}_{0,m}^{(0)}(k)S_e^{(0)}(k)+SF_f\tilde{H}_{1,m}^{(0)}(k)S_o^{(0)}(k)+\tilde{\Pi}_{0,m}^{(0)}(k)\\ \tilde{R}_{1,m}^{(0)}(k)=-SF_f\tilde{H}_{0,m}^{(0)}(k)S_o^{(0)*}(k)+SF_f\tilde{H}_{1,m}^{(0)}(k)S_e^{(0)*}(k)+\tilde{\Pi}_{1,m}^{(0)}(k)\end{cases}.\ (15)$$

It is understood from Eq. (15) that $S_e^{(0)}(k)$ and $S_o^{(0)}(k)$ are extracted without MUI. Joint MMSE-FDE, STTD-decoding and receive antenna diversity combining is carried out to obtain $\tilde{S}_e^{(0)}(k)$ and $\tilde{S}_o^{(0)}(k)$ as

$$\begin{cases}\tilde{S}_e^{(0)}(k)=\sum_{m=0}^{M-1}\left\{w_{0,m}^{(0)*}(k)\tilde{R}_{0,m}^{(0)}(k)+w_{1,m}^{(0)}(k)\tilde{R}_{1,m}^{(0)*}(k)\right\}\\ \qquad=SF_f\hat{H}^{(0)}(k)S_e^{(0)}(k)+\hat{\Pi}_e^{(0)}(k)\\ \tilde{S}_o^{(0)}(k)=\sum_{m=0}^{M-1}\left\{w_{1,m}^{(0)*}(k)\tilde{R}_{0,m}^{(0)}(k)-w_{0,m}^{(0)}(k)\tilde{R}_{1,m}^{(0)*}(k)\right\}\\ \qquad=SF_f\hat{H}^{(0)}(k)S_o^{(0)}(k)+\hat{\Pi}_o^{(0)}(k)\end{cases},(16)$$

where $w_{n,m}^{(u)}(k)$, $\hat{H}^{(0)}(k)$ and $\hat{\Pi}_{e(o)}^{(0)}(k)$ are respectively the MMSE-FDE weight, the equivalent channel gain and the noise after joint MMSE-FDE, STTD-decoding and receive antenna diversity. $w_{n,m}^{(u)}(k)$ and $\hat{H}^{(0)}(k)$ are given by

$$\begin{cases}w_{n,m}^{(0)}(k)=\dfrac{\tilde{H}_{n,m}^{(0)}(k)}{\sum_{n=0}^{1}\sum_{m=0}^{M-1}\left|\tilde{H}_{n,m}^{(0)}(k)\right|^2+\left(\frac{1}{2}\frac{1}{SF_t}\frac{E_s}{N_0}\right)^{-1}}\\ \hat{H}^{(0)}(k)=\sum_{n=0}^{1}\sum_{m=0}^{M-1}\left\{w_{n,m}^{(0)}(k)\right\}^*\tilde{H}_{n,m}^{(0)}(k)\end{cases}.(17)$$

N_c-point IFFT is applied to $\tilde{S}_{e(o)}^{(0)}(k)$ to recover the time-domain chip block:

$$\tilde{s}_{e(o)}^{(0)}(t)=\frac{1}{N_c}\sum_{k=0}^{N_c-1}\tilde{S}_{e(o)}^{(0)}(k)\exp\left(j2\pi t\frac{k}{N_c}\right)\ (18)$$

Finally, despreading is performed on $\tilde{s}_{e(o)}^{(0)}(t)$ to obtain the 0th user's decision variable associated with $d_{e(o)}^{(0)}(q)$:

$$\begin{aligned}\tilde{d}_{e(o)}^{(0)}(q)&=\frac{1}{SF_t}\sum_{t=qSF_t}^{(q+1)SF_t-1}\tilde{s}_{e(o)}^{(0)}(t)c^{(0)*}(t)\\&=\sqrt{\frac{E_s}{T_cSF_t}}\left(\frac{SF_f}{N_c}\sum_{k=0}^{N_c-1}\hat{H}^{(0)}(k)\right)d_{e(o)}^{(0)}(q)+\mu^{(0)}(q)+\hat{\eta}^{(0)}(q)\end{aligned},\ (19)$$

where $\mu^{(0)}(q)$ and $\hat{\eta}_{e(o)}^{(0)}(q)$ are the residual inter-chip interference (ICI) and the noise after joint MMSE-FDE, STTD-decoding and receive antenna diversity combining, respectively.

III. BER ANALYSIS

From Eq. (19), the decision variable $\tilde{d}^{(0)}_{e(o)}(q)$ is a random variable with a mean of $\sqrt{\frac{E_s}{T_c SF_t}}\left(\frac{SF_f}{N_c}\sum_{k=0}^{N_c-1}\hat{H}^{(0)}(k)\right)d^{(0)}_{e(o)}(q)$. The residual ICI $\mu^{(0)}(q)$ can be approximated as a zero-mean complex-valued Gaussian variable according to the central limit theorem [15] since it is a contribution from a large number of chips (i.e., N_c–1 chips). Hence, the sum of $\mu^{(0)}(q)$ and $\hat{\eta}^{(0)}_{e(o)}(q)$ can be treated as a new zero-mean Gaussian noise ν. Its variance is given by

$$2\sigma_\nu^2 = 2\sigma_{ICI}^2 + 2\sigma_{noise}^2 \,, \quad (20)$$

where $2\sigma_{ICI}^2$ and $2\sigma_{noise}^2$ are a variances of $\mu^{(0)}(q)$ and $\hat{\eta}^{(0)}_{e(o)}(q)$, respectively. Since $E\left[s^{(0)}_{e(o)}(t)s^{(0)*}_{e(o)}(\tau)\right]=\left(E_s/T_c SF_t\right)\delta(t-\tau)$ and $E[c(t)c^*(\tau)]=\delta(t-\tau)$, σ_{ICI}^2 is given by [6]

$$\sigma_{ICI}^2=\frac{1}{2}\frac{SF_f^2}{SF_t}\frac{E_s}{T_c SF_t}\left[\frac{1}{N_c}\sum_{k=0}^{N_c-1}|\hat{H}^{(0)}(k)|^2-\left|\frac{1}{N_c}\sum_{k=0}^{N_c-1}\hat{H}^{(0)}(k)\right|^2\right] .(21)$$

$\tilde{\Pi}^{(0)}_{p,m}(k)$ is a zero-mean Gaussian noise with a variance of $2N_0 SF_f(SF_f N_c)/T_c$ and hence, σ_{noise}^2 is given by [6]

$$\sigma_{noise}^2=\frac{SF_f^2}{SF_t}\frac{N_0}{T_c}\left[\frac{1}{N_c}\sum_{k=0}^{N_c-1}\sum_{n=0}^{1}\sum_{m=0}^{M-1}|w^{(0)}_{n,m}(k)|^2\right] . (22)$$

Therefore, σ_ν^2 is given, from Eqs. (20)~(22), by

$$\sigma_\nu^2=\frac{SF_f^2}{SF_t}\frac{N_0}{T_c}\left[\begin{array}{l}\frac{1}{N_c}\sum_{k=0}^{N_c-1}\sum_{n=0}^{1}\sum_{m=0}^{M-1}|w^{(0)}_{n,m}(k)|^2+\left(\frac{1}{2}\frac{1}{SF_t}\frac{E_s}{N_0}\right)\times\\ \left(\frac{1}{N_c}\sum_{k=0}^{N_c-1}|\hat{H}^{(0)}(k)|^2-\left|\frac{1}{N_c}\sum_{k=0}^{N_c-1}\hat{H}^{(0)}(k)\right|^2\right)\end{array}\right] .(23)$$

Assuming quaternary phase shift keying (QPSK) modulation, the conditional BER for the given set of $\{H^{(0)}_{n,m}(k)$; n=0, 1, m=0~M–1 and k=0~$N_c SF_f$–1} is given by

$$p_b\left(\frac{E_s}{N_0},\{H^{(0)}(k)\}\right)=\frac{1}{2}erfc\left[\sqrt{\frac{1}{4}\gamma\left(\frac{E_s}{N_0},\{H^{(0)}_{n,m}(k)\}\right)}\right] , (24)$$

where $erfc[x]=(2/\sqrt{\pi})\int_x^\infty \exp(-t^2)dt$ is the complimentary error function and $\gamma\left(E_s/N_0,\{H^{(0)}_{n,m}(k)\}\right)$ is an instantaneous signal-to-interference plus noise power ratio (SINR), defined as

$$\gamma\left(\frac{E_s}{N_0},\{H^{(0)}_{n,m}(k)\}\right)=\frac{\frac{E_s}{N_0}\left|\frac{1}{N_c}\sum_{k=0}^{N_c-1}\hat{H}^{(0)}(k)\right|^2}{\left[\begin{array}{l}\frac{1}{N_c}\sum_{k=0}^{N_c-1}\sum_{n=0}^{1}\sum_{m=0}^{M-1}\left|w^{(0)}_{n,m}(k)\right|^2+\left(\frac{1}{2}\frac{1}{SF_t}\frac{E_s}{N_0}\right)\times\\ \left\{\frac{1}{N_c}\sum_{k=0}^{N_c-1}\left|\hat{H}^{(0)}(k)\right|^2-\left|\frac{1}{N_c}\sum_{k=0}^{N_c-1}\hat{H}^{(0)}(k)\right|^2\right\}\end{array}\right]} .(25)$$

The theoretical BER can be numerically obtained by averaging Eq. (24) over all { $H^{(0)}_{n,m}(k)$; n=0, 1, m=0~M–1 and k=0~$N_c SF_f$–1}.

IV. THEORETICAL AND SIMULATION RESULT

We assume an FFT window size of $N_c SF_f$=1024, a GI length of N_g=32 chips, the overall spreading factor SF=$SF_t SF_f$=16, the number M=2 of receive antennas, and QPSK data modulation. The channel is assumed to be a frequency-selective block Rayleigh fading channel having a chip-spaced L-path uniform power delay profile (i.e., $E[|h_{n,m,l}|^2]=1/L$ for all n, m and l). The time delay $\tau_l^{(u)}$, u=0~U–1, of lth path is $\tau_l^{(u)}=l+\bar{\tau}^{(u)}$ chips, where $\bar{\tau}^{(u)}$ is the uth user's time delay. We assume that the maximum time delay $\tau_{L-1}^{(u)}$ is shorter than the GI length for all u. Perfect chip timing and ideal channel estimation are assumed. MMSE weight is used for FDE. We assume three types of frequency-interleaving patterns as shown in Fig. 3; equal-space, block and random interleaving patterns.

The numerical evaluation of the theoretical average BER is done by Monte-Carlo numerical computation method as follows. The set of path gains { $h^{(0)}_{n,m,l}$; n=0, 1, m=0~M–1 and l=0~L–1} is generated to obtain { $H^{(0)}_{n,m}(k')$; n=0, 1, m=0~M–1 and k'=0~$N_c SF_f$–1}. Then, the frequency-deinterleaved channel gain $\tilde{H}^{(0)}_{n,m}(k)$ is obtained according to frequency-interleaving patterns, using Eq. (14), for computing $w^{(0)}_{n,m}(k)$ and $\hat{H}^{(0)}(k)$ using Eq. (17). The conditional BER for the given average received E_s/N_0 is computed using Eq. (24). This is repeated a sufficient number of times to obtain the theoretical average BER. When U>1, the BER performance is averaged over all users.

The theoretical average BER performance of frequency-interleaved SC-SSMA with joint MMSE-FDE, STTD and receive antenna diversity (M=2) is plotted in Fig. 4 as a function of the average received bit energy-to-AWGN noise power spectrum density ratio E_b/N_0, given by E_b/N_0=0.5(E_s/N_0)[1+N_g/($N_c SF_f$)]. The number of users is U=16. As many as SF_f users can be multiplexed without MUI, hence, (SF_t, SF_f) is set as (SF_t, SF_f)=(1,16) for U=16 [11]. For comparison, the theoretical average BER performance without STTD and receive antenna diversity, the BER performance of the conventional DS-CDMA, and that of U=1 are also plotted for (SF_t, SF_f)=(16, 1) and N_cSF_t=1024. Frequency-interleaved SC-SSMA provides much better BER performance than the conventional DS-CDMA since all users' frequency components do not overlap at all and the orthogonality among users is maintained. We can see that the equal-space and random interleavings are superior to the block interleaving (the block interleaving pattern provides the smallest frequency diversity gain).

Joint use of STTD and receive antenna diversity is very effective to improve the uplink BER performance of frequency-interleaved SC-SSMA (however, for the conventional DS-CDMA, the performance improvement is very slight since large MUI is produced). With antenna diversity, the E_b/N_0 reduction from the case without antenna diversity is as much as 8.7 dB for BER=10^{-4} when equal-space frequency-interleaving is used. The E_b/N_0 degradation from the performance of U=1 is only 0.7 dB (It is 4.4 dB for the case of no antenna diversity).

The computer simulation results are also plotted in Fig. 4 to confirm the validity of the theoretical analysis. In the simulation, a PN sequence with a repetition period of 4095 chips is used as the spreading sequence. A fairly good agreement between theoretical and computer simulation results is seen.

V. CONCLUSION

In this paper, STTD and receive antenna diversity have been introduced to frequency-interleaved SC-SSMA uplink with MMSE-FDE. Frequency-interleaving is used to avoid the MUI while MMSE-FDE is used to achieve the frequency-diversity gain. We have also derived the conditional BER for the given set of channel gains. A fairly good agreement between the theoretical and computer simulated BER performances was confirmed. Joint use of STTD and receive antenna diversity is very effective to improve the uplink BER performance of frequency-interleaved SC-SSMA with MMSE-FDE. With antenna diversity, the E_b/N_0 reduction from the case without antenna diversity is as much as 8.7 dB for BER=10^{-4} and the E_b/N_0 degradation from the case of U=1 is only 0.7 dB when equal-space frequency-interleaving is used.

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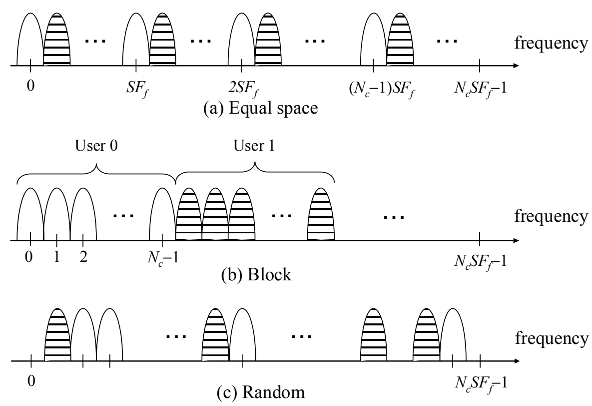

Figure 3. Frequency-interleaving pattern.

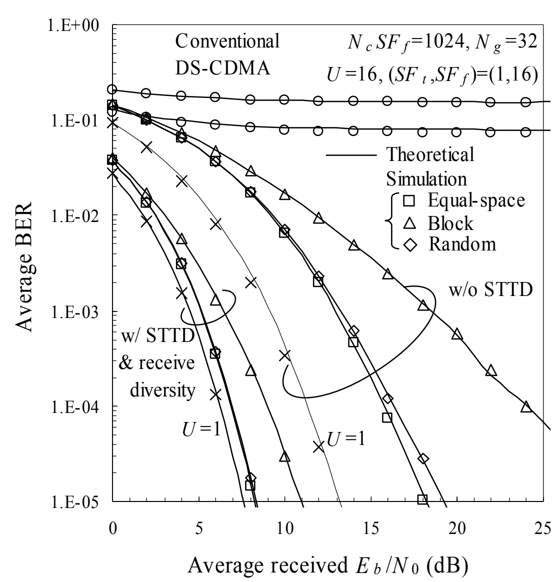

Figure 4. Average BER performance of SC-SS using STTD and receive antenna diversity when SF=16, U=16.