Multi-code MC-CDMA Using Joint CDTD and Inter-code Interference Cancellation

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***Abstract*—In multicarrier code division multiple access (MC-CDMA), a good bit error rate (BER) performance can be achieved by using frequency-domain equalization (FDE) based on the minimum mean square error (MMSE) criterion. However, the presence of residual inter-code interference (ICI) after MMSE-FDE and despreading degrades the BER performance of the multi-code MC-CDMA. The use of ICI cancellation can improve the BER performance of the multi-code MC-CDMA. To further improve the BER performance, transmit diversity technique is effective. In this paper, cyclic delay transmit diversity (CDTD) is considered. CDTD can achieve the frequency diversity gain, whereas the well-known space-time transmit diversity (STTD) can achieve the antenna diversity gain owing to the space-time coding and provide a better BER performance than CDTD. It is shown in this paper that the performance degradation of CDTD is mainly due to the residual ICI after MMSE-FDE. In order to achieve a good BER performance, the multi-code MC-CDMA using joint CDTD and ICI cancellation is proposed, and its BER performance is evaluated by computer simulation.**

Keywords-component; MC-CDMA, Inter-code interference cancellation, CDTD, STTD

I. INTRODUCTION

Next generation mobile communication systems are expected to offer broadband services. Wireless broadband channels are severely frequency-selective [1], and the transmission performance significantly degrades. Recently, many researchers have paid much attention to multicarrier code division multiple access (MC-CDMA) [2]-[4]. In MC-CDMA, a large frequency diversity gain can be achieved by using frequency-domain equalization (FDE) based on the minimum mean square error (MMSE) criterion and frequency-domain despreading. Hence, a good bit error rate (BER) performance is obtained in a severe frequency-selective fading channel.

However, if multi-code transmission is used to increase the data rate in the MC-CDMA, the presence of residual inter-code interference (ICI), after MMSE-FDE and despreading, degrades the BER performance as the code multiplexing order increases. The ICI cancellation [5]-[7] can improve the BER performance significantly.

To further improve the BER performance, transmit antenna diversity technique should be considered. Recently, cyclic delay transmit diversity (CDTD) has been proposed [8]. The CDTD can achieve the large frequency diversity gain. Thus, the introduction of CDTD into MC-CDMA can improve the BER performance in a weak frequency-selective (or non-frequency-selective) fading channel. Space-time transmit diversity (STTD) is another well-known transmit diversity technique to achieve an antenna diversity gain owing to the space-time coding [9]. It is shown [10] that STTD provides a better BER performance of the multi-code MC-CDMA than CDTD when the ICI cancellation technique is not considered.

It is shown in this paper that the performance degradation of CDTD is mainly due to the residual ICI after MMSE-FDE and despreading. To improve the BER performance, the multi-code MC-CDMA using joint CDTD and ICI cancellation is proposed. The BER performance is evaluated by computer simulation and is compared between CDTD and STTD. The lower-bound BER is also derived assuming the perfect ICI cancellation (or assuming single-code transmission with maximum ratio combining (MRC)-FDE). We show that the same BER performance as STTD can be achieved when the spreading factor SF equals the number N_c of subcarriers.

II. MULTI-CODE MC-CDMA WITH TRANSMIT DIVERSITY

A. CDTD

Throughout the paper, sample-spaced time representation of the transmitted signals is used. At the transmitter, a binary data sequence is data-modulated and then, serial/parallel (S/P)-converted to U parallel data sequences, each of which is divided into blocks of N_c/SF symbols, where N_c is the number of subcarriers and SF is the spreading factor. A symbol block $\{d_u(m);\ m=0\sim N_c/SF-1$ and $u=0\sim U-1\}$ in the uth sequence is then spread by multiplying it with an orthogonal spreading sequence $\{c_u(k);\ k=0\sim SF-1\}$. The resultant U parallel chip blocks are code-multiplexed and further multiplied by a common scrambling sequence $c_{scr}(k)$. After spreading and code-multiplexing, chip interleaving is used to each chip block in order to make full use of frequency diversity.

The spread chip block $\{S(k);\ k=0\sim N_c-1\}$ is expressed, using the equivalent lowpass representation, as

$$S(k)=\left[\sum_{u=0}^{U-1}d_u(\lfloor k/SF\rfloor)\cdot c_u(k \bmod SF)\right]c_{scr}(k), \tag{1}$$

where $\lfloor x\rfloor$ represents the largest integer smaller than or equal to x. N_c-point inverse fast Fourier transform (IFFT) is applied to $\{S(k);k=0\sim N_c-1\}$ to obtain the MC-CDMA signal block $\{s(t);t=0\sim N_c-1\}$ as

$$s(t)=\sum_{k=0}^{N_c-1}S(k)\exp\left(j\frac{2\pi t}{N_c}k\right). \tag{2}$$

Using CDTD, the same MC-CDMA signal block is simultaneously transmitted from different antennas after adding different cyclic delays [8]. The transmitted signal block from the nth antenna (n=0~N_t–1) is given by

$$\bar{s}_n(t)=\sqrt{\frac{2E_c}{N_t N_c T_c}}s\big((t-n\Delta)\bmod N_c\big), \qquad (3)$$

where E_c and T_c denote the chip energy and chip duration respectively, and $n\Delta$ is the cyclic delay. The transmit signal power is reduced by a factor of N_t to keep the total transmit signal power constant. Finally, the last N_g samples of each MC-CDMA signal block are copied as a cyclic prefix and inserted into the guard interval (GI) placed at the beginning of each MC-CDMA signal block.

The MC-CDMA signal block is transmitted over a frequency-selective fading channel. Block fading is assumed such that the path gains remain constant over one block length of (N_c+N_g) samples. After the removal of the GI, the received MC-CDMA signal {$r(t)$;t=0~N_c–1} is expressed as

$$r(t)=\sum_{n=0}^{N_t-1}\sum_{l=0}^{L-1}h_{n,l}\bar{s}_n\big((t-\tau_l)\bmod N_c\big)+\eta(t), \qquad (4)$$

where $h_{n,l}$ is the lth (l=0~L–1) complex-valued path gain between the nth transmit antenna and the receiver with $\sum_{l=0}^{L-1}E[|h_{n,l}|^2]=1$, where E[.] denotes the ensemble average operation [11]. τ_l is the lth path delay and the maximum delay difference $\tau_{L-1}-\tau_0$ is assumed to be shorter than the GI. $\eta(t)$ is a zero-mean complex Gaussian process with a variance of $2N_0/T_c$; N_0 is the single-sided power spectrum density of the additive white Gaussian noise (AWGN).

At the receiver side, N_c-point FFT is applied to {$r(t)$; t=0~N_c–1} to transform it into the frequency-domain signal {$R(k)$;k=0~N_c–1}, given by

$$R(k)=\sqrt{2E_c/N_tT_c}\,H_{CD}(k)S(k)+\Pi(k), \qquad (5)$$

where $H_{CD}(k)$ and $\Pi(k)$ are the channel gain and the noise due to the AWGN, respectively. $H_{CD}(k)$ is given by

$$H_{CD}(k)=\sum_{n=0}^{N_t-1}\sum_{l=0}^{L-1}h_{n,l}\exp\left(-j2\pi k\frac{n\Delta+\tau_l}{N_c}\right). \qquad (6)$$

One-tap MMSE-FDE is carried out on $R(k)$ as

$$\hat{R}(k)=R(k)W(k), \qquad (7)$$

where $W(k)$ is the FDE weight based on the MMSE criterion per subcarrier [4]. After chip deinterleaving, despreading is performed on {$\hat{R}(k)$;k=0~N_c–1}, giving

$$\hat{d}_u(m)=\frac{1}{SF}\sum_{k=mSF}^{(m+1)SF-1}\hat{R}(k)c_u^*(k\bmod SF)c_{scr}^*(k), \qquad (8)$$

where * represents the complex conjugate operation. $\hat{d}_u(m)$ is the decision variable associated with $d_u(m)$.

B. STTD

In this paper, N_t=2 and 4-antenna STTD's are considered [9], [13]. Below, N_t=2-antenna STTD case is described only. Even and odd chip blocks are respectively denoted by $\{S_e(k);k=0\sim N_c-1\}$ and $\{S_o(k);k=0\sim N_c-1\}$. Space-time (ST) coding is applied to $\{S_e(k)\}$ and $\{S_o(k)\}$ as

$$\begin{pmatrix}\bar{S}_{0,0}(k) & \bar{S}_{1,0}(k)\\ \bar{S}_{0,1}(k) & \bar{S}_{1,1}(k)\end{pmatrix}=\begin{pmatrix}S_e(k) & -S_o^*(k)\\ S_o(k) & S_e^*(k)\end{pmatrix}, \qquad (9)$$

where $\{\bar{S}_{q,n}(k);k=0\sim N_c-1\}$ is the qth STTD-encoded chip block (q=0 and 1) to be transmitted from the nth antenna. N_c-point IFFT is applied to $\{\bar{S}_{q,n}(k);k=0\sim N_c-1\}$ to obtain two consecutive MC-CDMA signal blocks. After the GI insertion, the two consecutive blocks are transmitted over a frequency-selective fading channel and received at a receiver.

At the receiver, after the GI is removed, the two consecutive received MC-CDMA signal blocks are decomposed into the frequency-domain signals $\{R_e(k);k=0\sim N_c-1\}$ and $\{R_o(k);k=0\sim N_c-1\}$ by applying N_c-point FFT. $R_e(k)$ and $R_o(k)$ are given by

$$\begin{cases}R_e(k)=\sqrt{\dfrac{2E_c}{N_tT_c}}H_0(k)S_e(k)+\sqrt{\dfrac{2E_c}{N_tT_c}}H_1(k)S_o(k)+\Pi_e(k),\\ R_o(k)=-\sqrt{\dfrac{2E_c}{N_tT_c}}H_0(k)S_o^*(k)+\sqrt{\dfrac{2E_c}{N_tT_c}}H_1(k)S_e^*(k)+\Pi_o(k),\end{cases} \qquad (10)$$

where $H_{0(\text{or }1)}(k)$ and $\Pi_{e(\text{or }o)}(k)$ are the channel gain between the 0th (or 1st) transmit antenna and the receiver and the noise, respectively. $H_{0(\text{or }1)}(k)$ is given by

$$H_{0(\text{or }1)}(k)=\sum_{l=0}^{L-1}h_{0(\text{or }1),l}\exp\left(-j2\pi k\frac{\tau_l}{N_c}\right). \qquad (11)$$

Joint MMSE-FDE and STTD decoding is carried out as follows [12];

$$\begin{cases}\hat{R}_e(k)=R_e(k)W_0^*(k)+R_o^*(k)W_1(k)\\ \hat{R}_o(k)=R_e(k)W_1^*(k)-R_o^*(k)W_0(k)\end{cases}, \qquad (12)$$

where $W_{0(\text{or }1)}(k)$ is the MMSE-FDE weight combined with STTD decoding. Finally, despreading (see Eq. (8)) is carried out before data demodulation.

C. Residual ICI after MMSE-FDE

The residual ICI after MMSE-FDE, denoted by $\{M_{CD}(k);k=0\sim N_c-1\}$ for CDTD and $\{M_{e(\text{or }o)}(k);k=0\sim N_c-1\}$ for STTD, can be written as [7]

$$\begin{cases}M_{CD}(k)=\sqrt{\dfrac{2E_c}{N_tT_c}}\left\{\hat{H}_{CD}(k)-A_{CD}\left(\left\lfloor\dfrac{k}{SF}\right\rfloor\right)\right\}S(k)\\ M_{e(\text{or }o)}(k)=\sqrt{\dfrac{2E_c}{N_tT_c}}\left\{\hat{H}_{ST}(k)-A_{ST}\left(\left\lfloor\dfrac{k}{SF}\right\rfloor\right)\right\}S_{e(\text{or }o)}(k)\end{cases} \qquad (13)$$

with

$$A_{CD(\text{or }ST)}(m)=\frac{1}{SF}\sum_{k=mSF}^{(m+1)SF-1}\hat{H}_{CD(\text{or }ST)}(k), \qquad (14)$$

where $\hat{H}_{CD}(k)$ is the equivalent channel gain after MMSE-FDE for CDTD. $\hat{H}_{ST}(k)$ is the equivalent channel gain after joint MMSE-FDE and STTD decoding. $\hat{H}_{CD}(k)$ and $\hat{H}_{ST}(k)$ are respectively given by

$$\begin{cases}\hat{H}_{CD}(k)=W(k)H_{CD}(k) & \text{for CDTD}\\ \hat{H}_{ST}(k)=W_0(k)H_0^*(k)+W_1(k)H_1^*(k) & \text{for STTD}\end{cases}. \qquad (15)$$

$|M_{CD}(k)|$ and $|M_e(k)|$ are plotted in Fig. 1 for an L=16-path frequency-selective fading channel with a decay factor α=5 (dB) when N_c=SF=U=256, and E_b/N_0=12 dB. It can be seen that CDTD enhances the frequency-selectivity of the channel, and the large ICI is produced after MMSE-FDE. However, STTD produces less residual ICI since the frequency-selectivity can be suppressed by antenna diversity effect obtained through STTD decoding.

D. Simulated BER performance

We assume QPSK data-modulation, N_c=256 and N_g=32. The channel is assumed to be a frequency-selective block Rayleigh fading channel having a sample-spaced L=16-path exponential power delay profile with a decay factor α=5 dB. Perfect sample timing and ideal channel estimation are assumed.

The simulated BER performance is plotted in Fig. 2 as a function of the average received bit energy-to-AWGN noise power spectrum density ratio E_b/N_0, defined as E_b/N_0=0.5SF(E_c/N_0)(1+N_g/N_c), when SF=U=256 (i.e., full code-multiplexing). The BER performance is plotted for CDTD, STTD and N_t=1 (i.e., no antenna diversity). For CDTD, a cyclic delay of Δ=37-chip is used, which is found by computer simulation to provide the minimum BER. For STTD using N_t=4, space-time coding with a coding rate of 3/4 is used [13]. CDTD can enhance the channel frequency selectivity and hence, improve the BER performance; however, the BER performance using CDTD is inferior to that using STTD which can achieve the antenna diversity gain. This performance degradation of CDTD from STTD is mainly due to the residual ICI. The use of the ICI cancellation technique can be used to improve the BER performance of CDTD.

III. MULTI-CODE MC-CDMA USING JOINT CDTD AND ICI CANCELLATION

An MC-CDMA receiver using joint MMSE-FDE and ICI cancellation is shown in Fig. 3. In this section, the ith iteration process is described.

A. CDTD

MMSE-FDE is carried out as

$$\hat{R}^{(i)}(k) = R(k)W^{(i)}(k), \quad (16)$$

where $W^{(i)}(k)$ is the MMSE-FDE weight taking into account the residual ICI. $W^{(i)}(k)$ is given by

$$W^{(i)}(k) = \frac{H^*_{CD}(k)}{\left(\sum_{u=0}^{U-1}\rho_u^{(i-1)}\left(\left\lfloor\frac{k}{SF}\right\rfloor\right)\right)|H_{CD}(k)|^2 + \left(\frac{1}{N_t}\frac{E_c}{N_0}\right)^{-1}}, \quad (17)$$

where $\rho_u^{(i-1)}(m)$, m=0~N_c/SF, shows the extent to which the residual ICI remains and is given as [7]

$$\rho_u^{(i-1)}(m) = \left\{E\left[|d_u(m)|^2\right] - \left|\tilde{d}_u^{(i-1)}(m)\right|^2\right\} \quad (18)$$

with $\rho_u^{(-1)}(m)=1$ and $\{\tilde{d}_u^{(i-1)}(m); m=0\sim N_c/SF-1\}$ being the soft decision replica of the transmitted symbol block $\{d_u(m)\}$, obtained in the previous iteration. $E[|d_u(m)|^2]$ is the expectation of $|d_u(m)|^2$ for the given received chip block. $\rho_u^{(i-1)}(m)\to 1$ (i.e., $|\tilde{d}_u^{(i-1)}(m)|^2\to 0$) means the residual ICI is kept intact, while $\rho_u^{(i-1)}(m)\to 0$ (i.e., $|\tilde{d}_u^{(i-1)}(m)|^2\to 1$) means the residual ICI is sufficiently cancelled.

ICI cancellation is performed on $\hat{R}^{(i)}(k)$ as

$$\tilde{R}^{(i)}(k) = \hat{R}^{(i)}(k) - \tilde{M}_{CD}^{(i)}(k), \quad (19)$$

where $\tilde{M}_{CD}^{(i)}(k)$ is the replica of $M_{CD}(k)$. $\tilde{M}_{CD}^{(i)}(k)$ is given, from Eq. (13), by

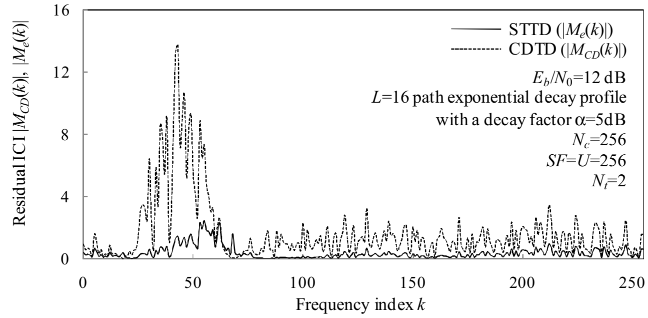

Figure 1 Residual ICI for CDTD and STTD.

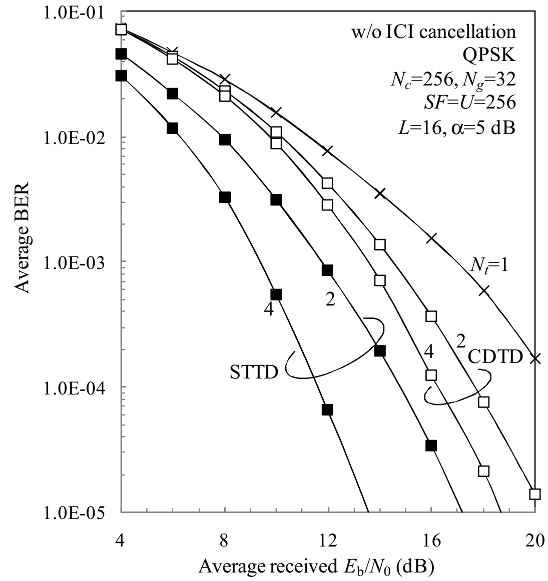

Figure 2 BER performances with CDTD and STTD.

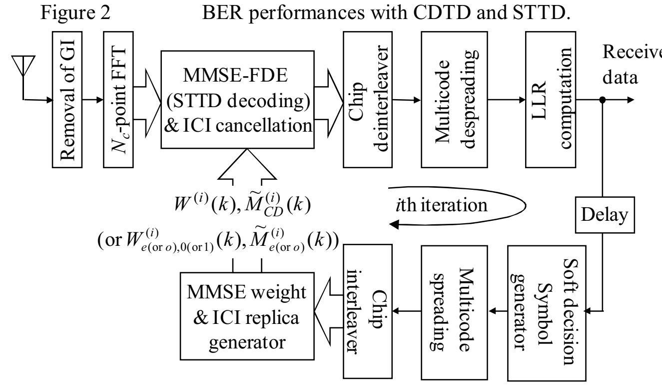

Figure 3 MC-CDMA receiver structure using ICI cancellation.

$$\tilde{M}_{CD}^{(i)}(k)=\sqrt{\frac{2E_c}{N_tT_c}}\left\{\hat{H}_{CD}^{(i)}(k)-A_{CD}^{(i)}\left(\left\lfloor\frac{k}{SF}\right\rfloor\right)\right\}\tilde{S}^{(i-1)}(k)\ , \quad (20)$$

where $\tilde{M}_{CD}^{(0)}(k)=0$, $\hat{H}_{CD}^{(i)}(k)=W^{(i)}(k)H_{CD}(k)$ is the equivalent channel gain after MMSE-FDE, and $\{\tilde{S}^{(i-1)}(k);k=0\sim N_c-1\}$ is the replica of $S(k)$ obtained in the previous iteration. Finally, despreading is carried out on $\{\tilde{R}^{(i)}(k)\}$ to get the decision variable $\{\hat{d}_u^{(i)}(m);m=0\sim N_c/SF-1\}$ associated with $d_u(m)$.

ICI replica generation and MMSE-FDE weight updating for the *i*th iteration are presented below. Using the decision variable $\hat{d}_u^{(i-1)}(m)$ obtained in the previous iteration, the log-likelihood ratio (LLR) $\{L_u(x,m);x=0\sim\log_2K-1 \text{ and } m=0\sim N_c/SF-1\}$ for the *x*th bit in the *m*th symbol $d_u(m)$, where *x*=0~log$_2$*K*−1 and *K* is the modulation level, is computed [14]. For QPSK data modulation, using $L_u(0,m)$ and $L_u(1,m)$, the soft symbol replica is obtained as

$$\tilde{d}_u^{(i-1)}(m)=\frac{1}{\sqrt{2}}\tanh\left(\frac{L_u(0,m)}{2}\right)+j\frac{1}{\sqrt{2}}\tanh\left(\frac{L_u(1,m)}{2}\right)\ . \quad (21)$$

$\tilde{d}_u^{(i-1)}(m)$ is used to compute $\rho_u^{(i-1)}(m)$ of Eq. (18) for updating the MMSE-FDE weight $W^{(i)}(k)$. $E[|d_u(m)|^2]$ is also necessary for the computation of $\rho_u^{(i-1)}(m)$; $E[|d_u(m)|^2]=1$ for QPSK data-modulation. $\tilde{d}_u^{(i-1)}(m)$ is spread, code-multiplexed, and chip-interleaved to generate the chip block replica $\{\tilde{S}^{(i-1)}(k);k=0\sim N_c-1\}$. Re-spreading of erroneous symbol replica results in error propagation over consecutive *SF* chips. To avoid this, chip interleaving is used after spreading (chip interleaving is also effective to achieve the larger frequency diversity). $\tilde{S}^{(i-1)}(k)$ is given by

$$\tilde{S}^{(i-1)}(k)=\left[\sum_{u=0}^{U-1}\tilde{d}_u^{(i-1)}(\lfloor k/SF\rfloor)\cdot c_u(k \bmod SF)\right]c_{scr}(k)\ , \quad (22)$$

which is used for generating the ICI replica of Eq. (20).

B. STTD

ICI cancellation for STTD is briefly explained. Joint MMSE-FDE and STTD decoding is performed using $R_e(k)$ and $R_o(k)$ as

$$\begin{cases}\hat{R}_e^{(i)}(k)=R_e(k)W_{e,0}^{(i)*}(k)+R_o^*(k)W_{e,1}^{(i)}(k)\\ \hat{R}_o^{(i)}(k)=R_e(k)W_{o,1}^{(i)*}(k)-R_o^*(k)W_{o,0}^{(i)}(k)\end{cases}, \quad (23)$$

where $W_{e(\text{or }o),0(\text{or }1)}^{(i)}(k)$ is the MMSE-FDE weight combined with STTD decoding, which takes into account the residual ICI. $W_{e(\text{or }o),0(\text{or }1)}^{(i)}(k)$ is given by

$$W_{e(\text{or }o),0(\text{or }1)}^{(i)}(k)=\frac{H_{0(\text{or }1)}(k)}{\left[\left(\sum_{u=0}^{U-1}\rho_{e(\text{or }o),u}^{(i-1)}\left(\left\lfloor\frac{k}{SF}\right\rfloor\right)\right)\sum_{n=0}^{1}\left|H_n(k)\right|^2+\left(\frac{1}{N_t}\frac{E_c}{N_0}\right)^{-1}\right]}\ , \quad (24)$$

where

$$\rho_{e(\text{or }o),u}^{(i-1)}(m)=\left\{E\left[\left|d_{e(\text{or }o),u}(m)\right|^2\right]-\left|\tilde{d}_{e(\text{or }o),u}^{(i-1)}(m)\right|^2\right\} \quad (25)$$

with $\rho_{e(\text{or }o)}^{(-1)}=1$ and $\{\tilde{d}_{e(\text{or }o),u}^{(i-1)}(m);m=0\sim N_c/SF-1\}$ being the soft decision replica of the transmitted symbol block $\{d_{e(\text{or }o),u}(m)\}$ obtained in the previous iteration. $E[|d_{e(\text{or }o),u}(m)|^2]$ is the expectation of $|d_{e(\text{or }o),u}(m)|^2$ for the given received chip block. ICI cancellation is performed on $\hat{R}_{e(\text{or }o)}^{(i)}(k)$ as

$$\tilde{R}_{e(\text{or }o)}^{(i)}(k)=\hat{R}_{e(\text{or }o)}^{(i)}(k)-\tilde{M}_{e(\text{or }o)}^{(i)}(k)\ , \quad (26)$$

where $\tilde{M}_{e(\text{or }o)}^{(i)}(k)$ is the replica of $M_{e(\text{or }o)}(k)$. $\tilde{M}_{e(\text{or }o)}^{(i)}(k)$ is given, from Eq. (13), as

$$\tilde{M}_{e(\text{or }o)}^{(i)}(k)=\sqrt{\frac{2E_c}{N_tT_c}}\left\{\hat{H}_{e(\text{or }o)}^{(i)}(k)-A_{ST}^{(i)}\left(\left\lfloor\frac{k}{SF}\right\rfloor\right)\right\}\tilde{S}_{e(\text{or }o)}^{(i-1)}(k)\ , \quad (27)$$

where $\tilde{M}_{e(\text{or }o)}^{(0)}(k)=0$, $\hat{H}_{e(\text{or }o)}^{(i)}(k)=W_{e(\text{or }o),0}^{(i)}(k)H_0^*(k)+W_{e(\text{or }o),1}^{(i)}(k)H_1^*(k)$ is the equivalent channel gain after joint MMSE-FDE and STTD decoding, and $\tilde{S}_{e(\text{or }o)}^{(i-1)}(k)$ is the replica of $S_{e(\text{or }o)}(k)$ obtained in the previous iteration. Finally, despreading is carried out on $\{\tilde{R}_{e(\text{or }o)}^{(i)}(k)\}$ to obtain the decision variable $\{\hat{d}_{e(\text{or }o),u}^{(i)}(m);m=0\sim N_c/SF-1\}$ associated with $\{d_{e(\text{or }o),u}(m)\}$.

$\rho_{e(\text{or }o),u}^{(i-1)}(m)$ of Eq. (25) and $\tilde{S}_{e(\text{or }o)}^{(i-1)}(k)$ in Eq. (27) are obtained similar to the CDTD case.

IV. SNR FOR PERFECT ICI CANCELLATION

The lower-bound BER for CDTD and STTD can be derived assuming the perfect ICI cancellation (i.e., $\tilde{d}_u^{(i-1)}(m)=d_u(m)$ and $\tilde{S}^{(i-1)}(k)=S(k)$ for CDTD, while $\hat{d}_{e(\text{or }o),u}^{(i-1)}(m)=d_{e(\text{or }o),u}(m)$ and $\tilde{S}_{e(\text{or }o)}^{(i-1)}(k)=S_{e(\text{or }o)}(k)$ for STTD). Since $\rho_u^{(i)}(m)=0$ and $\rho_{e(o),u}^{(i)}(m)=0$ from Eqs. (18) and (25), the MMSE-FDE weight becomes the maximum ratio combining (MRC) weight, given by [7]

$$\begin{cases}W^{(i)}(k)=H_{CD}^*(k) & \text{for CDTD}\\ W_{e,0(\text{or }1)}^{(i)}(k)=H_{0(\text{or }1)}(k) & \text{for STTD}\end{cases}, \quad (28)$$

where the weight for the even block is only considered for STTD. The SNR after despreading, respectively denoted by $\{\gamma_{CD}(m);m=0\sim N_c/SF-1\}$ for CDTD and $\{\gamma_{ST}(m);m=0\sim N_c/SF-1\}$ and for STTD, is given by

$$\begin{cases}\gamma_{CD}(m)=\frac{2E_s}{N_0}\left(\frac{1}{N_t}\frac{1}{SF}\sum_{k=mSF}^{(m+1)SF-1}\left|H_{CD}(k)\right|^2\right)\\ \gamma_{ST}(m)=\frac{2E_s}{N_0}\left(\frac{1}{N_t}\frac{1}{SF}\sum_{k=mSF}^{(m+1)SF-1}\left(\left|H_0(k)\right|^2+\left|H_1(k)\right|^2\right)\right)\end{cases}, \quad (29)$$

where E_s/N_0 is the signal energy per symbol-to-AWGN power spectrum density ratio.

When $SF=N_c$, substituting Eqs. (6) and (11) into (29) gives

$$\gamma_{CD}(0) = \gamma_{ST}(0) = \frac{2E_s}{N_0}\left(\frac{1}{N_t}\sum_{n=0}^{N_t-1}\sum_{l=0}^{L-1}\left|h_{n,l}\right|^2\right). \quad (30)$$

The same SNR can be achieved for CDTD and STTD when $SF=N_c$. It can be seen from Eq. (30) that the frequency diversity gain achieved by CDTD is equivalent to the antenna diversity gain achieved by STTD.

Assuming QPSK data modulation, the conditional BER for the given set of $\{H_{CD}(k); k=0\sim N_c-1\}$ for CDTD and $\{H_n(k); n=0,1 \text{ and } k=0\sim N_c-1\}$ for STTD is given by

$$p_{CD(\text{or }ST)}\left(\frac{E_s}{N_0}, \{H_{CD(or\,n)}(k)\}\right) = \frac{1}{2}erfc\left(\sqrt{\frac{1}{4}\gamma_{CD(\text{or }ST)}(m)}\right), \quad (31)$$

where $erfc(x) = (2/\sqrt{\pi})\int_x^\infty \exp(-t^2)dt$ is the complimentary error function. The lower-bound BER can be numerically evaluated by averaging Eq. (31) over all realizations of $\{H_{CD}(k); k=0\sim N_c-1\}$ for CDTD and $\{H_n(k); n=0, 1 \text{ and } k=0\sim N_c-1\}$ for STTD.

V. COMPUTER SIMULATION

The simulation parameters are the same as in Fig. 2. The number of iterations of joint MMSE-FDE and ICI cancellation is set to i=2, which provides sufficiently improved BER performance.

The simulated BER performance is plotted in Fig. 4 as a function of the average received E_b/N_0 when $SF=U$=256. The lower-bound BER computed using Eq. (31) is also plotted for comparison. The use of ICI cancellation can improve the BER performance. CDTD shows exactly the same lower-bound BER as STTD as anticipated in Sect. IV. However, CDTD provides a performance slightly inferior to STTD. The reason for this is that the decision variables obtained at the initial iteration (i=0) are more erroneous for CDTD than for STTD as seen from Fig. 2 and hence, the accuracy of the soft symbol replica is worse for CDTD than for STTD. The performance improvement by CDTD and STTD from N_t=1 is due to the antenna diversity gain achieved by space-time encoding/decoding for STTD and due to the additional frequency diversity gain for CDTD.

VI. CONCLUSION

In the multi-code MC-CDMA, STTD gives a better BER performance than CDTD. In this paper, it was shown that the performance degradation of CDTD is due to the residual ICI and the multi-code MC-CDMA using joint CDTD and ICI cancellation was proposed. Assuming the perfect ICI cancellation, the lower-bound BER was derived for CDTD and STTD. For the case of $SF=N_c$, we theoretically showed that the lower-bound BER is the same for CDTD and STTD. We evaluated the BER performances of CDTD and STTD by computer simulation. CDTD with ICI cancellation provides the BER performance very close to STTD.

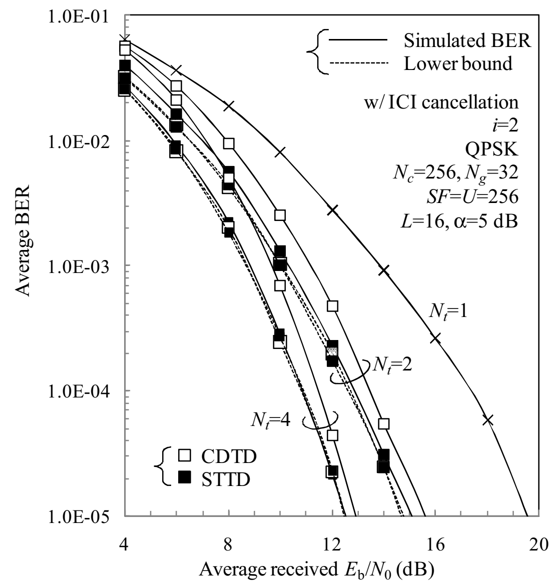

Figure 4 BER performance using joint CDTD or STTD and ICI cancellation.

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