# Transmit Diversity for DS-CDMA/MMSE-FDE with Frequency-domain ICI Cancellation

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***Abstract*—Frequency-domain equalization (FDE) based on the minimum mean square error (MMSE) criterion can provide a better bit error rate (BER) performance than rake combining. However, the residual inter-chip interference (ICI) is produced after MMSE-FDE and degrades the BER performance. Recently, we showed that frequency-domain ICI cancellation can bring the BER performance close to the theoretical lower bound. To further improve the BER performance, transmit antenna diversity technique is effective. Cyclic delay transmit diversity (CDTD) can increase the number of equivalent paths and hence achieve a large frequency diversity gain. Space-time transmit diversity (STTD) can obtain antenna diversity gain due to the space-time coding and achieve a better BER performance than CDTD. In this paper, we point out that the performance difference between CDTD and STTD is mainly due to the residual ICI. We theoretically show that the introduction of ICI cancellation into DS-CDMA/MMSE-FDE gives almost the same BER performance for CDTD and STTD. This is confirmed by computer simulation.**

***Index Terms*—DS-CDMA, MMSE-FDE, CDTD, STTD, ICI cancellation**

## I. Introduction

With the growing market of mobile wireless communications, high-rate data services are demanded. Wireless channels for such high-speed data transmissions are severely frequency-selective [1], and this degrades the transmission performance significantly. In the present mobile communication systems, direct sequence code division multiple access (DS-CDMA) has been adopted to exploit a moderate channel frequency-selectivity by using rake combining [2], [3]. However, for much higher-rate data transmissions than the present systems, the wireless channels become severely frequency-selective and hence, the bit error rate (BER) performance of rake combining significantly deteriorates due to the strong inter-path interference (IPI). Therefore, an advanced equalization technique is necessary.

Frequency-domain equalization (FDE) based on the minimum mean square error (MMSE) criterion can improve the BER performance of DS-CDMA transmissions [4]-[7]. However, the residual inter-chip interference (ICI) is present after MMSE-FDE and this limits the BER performance improvement. The frequency-domain interference cancellation was proposed for DS-CDMA uplink in [8]. Recently, we proposed a frequency-domain ICI cancellation for the DS-CDMA downlink and showed that frequency-domain ICI cancellation can bring the BER performance very close to the theoretical lower-bound [9].

To further improve the BER performance, transmit antenna diversity technique is effective. Recently, cyclic delay transmit diversity (CDTD) has been proposed for multi-carrier transmissions [10]. CDTD can increase the number of equivalent paths by transmitting the same data block from different antennas after adding different cyclic delays, and hence can achieve a large frequency diversity gain. CDTD can be applied to DS-CDMA/MMSE-FDE to improve the BER performance in a weak frequency-selective fading channel [10]. Space-time transmit diversity (STTD) is another attractive transmit diversity technique that can obtain the antenna diversity gain due to the space-time coding [11]. It was shown that the application of STTD to DS-CDMA/MMSE-FDE gives a better BER performance than CDTD [12].

In this paper, we first point out that the performance degradation of CDTD from STTD is mainly due to the residual ICI after MMSE-FDE. Then, we theoretically show that the introduction of ICI cancellation to DS-CDMA/MMSE-FDE gives almost the same BER performance for CDTD and STTD.

## II. CDTD

### *A. Transmitted signal*

DS-CDMA transmitter using CDTD is illustrated in Fig. 1. Throughout the paper, chip-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. The $u$th data modulated symbol sequence $\{d_u(m); m=0\sim N_c/SF-1\}$, $u=0\sim U-1$, is then spread by multiplying it with an orthogonal spreading sequence $c_u(t)$, where $SF$ is a spreading factor and $N_c$ is an FFT window size. The resultant $U$ chip sequences are code-multiplexed and further multiplied by a common scramble sequence $c_{scr}(t)$ to make the resultant multicode DS-CDMA signal like white-noise. The spread signal chip block $\{s(t);\ t=0\sim N_c-1\}$ to be transmitted can be expressed, using the equivalent lowpass representation, as

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

where $\lfloor x \rfloor$ represents the largest integer smaller than or equal to $x$. In CDTD, the same chip block is simultaneously transmitted from different antennas after adding different cyclic delays [10]. $N_t$ copies of the chip block $\{s(t);\ t=0\sim N_c-1\}$ are generated and then, cyclic delay $n\Delta$ is added before the transmission from the $n$th antenna ($n=0\sim N_t-1$). The transmitted chip block from the $n$th antenna is given by

$$\bar{s}_n(t)=\sqrt{2E_c/N_tT_c}\,s\left((t-n\Delta) \bmod N_c\right) \qquad ,(2)$$

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

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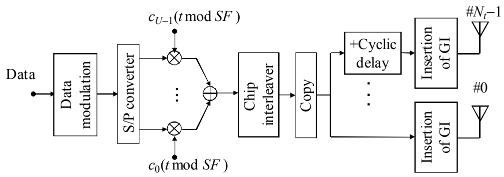

Figure 1 Transmitter structure using CDTD.

### B. FDE

The GI-inserted chip block is transmitted over a frequency-selective fading channel. The received chip block after the removal of the GI, $\{r(t);t=0\sim N_c-1\}$, is expressed as

$$r(t)=\sum_{n=0}^{N_t-1}\sum_{l=0}^{L-1}h_{n,l}\bar{s}_n\left((t-\tau_l)\bmod N_c\right)+\eta(t) \quad , (3)$$

where $h_{n,l}$ is the $l$th ($l$=0~$L$–1) complex-valued path gain between the $n$th transmit antenna and the receiver with $\sum_{l=0}^{L-1}E[|h_{n,l}|^2]=1$ ($E$[.] denotes the ensemble average operation) [13]. We assume block fading so that the path gains remain constant over one block length of $(N_c+N_g)$ chips. $\tau_l$ is the $l$th path delay and the maximum time delay $\tau_{L-1}$ 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).

$N_c$-point fast Fourier transform (FFT) is applied to $\{r(t);t=0\sim N_c-1\}$ to transform it into the frequency-domain signal $\{R(k);k=0\sim N_c-1\}$. $R(k)$ is given by

$$R(k)=\sqrt{2E_c/N_tT_c}\,H_{CD}(k)S(k)+\Pi(k) \quad , (4)$$

where $H_{CD}(k)$, $S(k)$ and $\Pi(k)$ are the channel gain, the $k$th frequency component of $\{s(t);t=0\sim N_c-1\}$ and the noise due to the AWGN, respectively. $H_{CD}(k)$ and $S(k)$ are given by

$$\begin{cases} S(k)=\sum_{t=0}^{N_c-1}s(t)\exp\left(-j2\pi k\dfrac{t}{N_c}\right) \\ H_{CD}(k)=\sum_{n=0}^{N_t-1}\sum_{l=0}^{L-1}h_{n,l}\exp\left(-j2\pi k\dfrac{n\Delta+\tau_l}{N_c}\right) \end{cases} \quad . (5)$$

One-tap MMSE-FDE is carried out on $R(k)$ as [7]

$$\hat{R}(k)=R(k)W(k) \quad , (6)$$

where $W(k)$ is the MMSE-FDE weight [7]. After MMSE-FDE, $\hat{R}(k)$ is transformed by applying $N_c$-point inverse FFT (IFFT) into the time-domain chip block $\{\tilde{r}(t);t=0\sim N_c-1\}$ as

$$\tilde{r}(t)=\frac{1}{N_c}\sum_{k=0}^{N_c-1}\hat{R}(k)\exp\left(j2\pi t\frac{k}{N_c}\right) \quad . (7)$$

Finally, despreading is performed on $\tilde{r}(t)$, giving

$$\hat{d}_u(m)=\frac{1}{SF}\sum_{t=mSF}^{(m+1)SF-1}\tilde{r}(t)c_u^*(t\bmod SF)c_{scr}^*(t) \quad , (8)$$

which is the decision variable associated with $d_u(m)$.

## III. STTD

In this paper, STTD of $N_t$=2 is considered. Even and odd chip blocks are respectively denoted by $\{s_e(t);t=0\sim N_c-1\}$ and $\{s_o(t);t=0\sim N_c-1\}$. Time-domain space-time coding is applied to $\{s_e(t)\}$ and $\{s_o(t)\}$ [14]. After inserting the GI, two consecutive chip blocks are transmitted over a frequency-selective fading channel and received at a receiver. After the GI is removed, the two consecutive received chip blocks $\{r_e(t);t=0\sim N_c-1\}$ and $\{r_o(t);t=0\sim N_c-1\}$ 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

$$\begin{cases} R_e(k)=\sqrt{\dfrac{2E_c}{N_tT_c}}H_0(k)S_e(k)+\sqrt{\dfrac{2E_c}{N_tT_c}}H_1(k)S_o(k)+\Pi_e(k) \\ R_o(k)=-\sqrt{\dfrac{2E_c}{N_tT_c}}H_0(k)S_o^*(k)+\sqrt{\dfrac{2E_c}{N_tT_c}}H_1(k)S_e^*(k)+\Pi_o(k) \end{cases} \quad , (9)$$

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. $H_{0(\text{or }1)}(k)$ is given by

$$H_{0(\text{or }1)}(k)=\sum_{l=0}^{L-1}h_{0(\text{or }1),l}\exp\left(-j2\pi k\frac{\tau_l}{N_c}\right) \quad . (10)$$

Joint MMSE-FDE and STTD decoding is carried out to obtain $\{\hat{R}_e(k);k=0\sim N_c-1\}$ and $\{\hat{R}_o(k);k=0\sim N_c-1\}$ [14]. $N_c$-point IFFT is applied to obtain the time-domain chip blocks, $\{\tilde{r}_e(t);t=0\sim N_c-1\}$ and $\{\tilde{r}_o(t);t=0\sim N_c-1\}$. Finally, despreading is done before data demodulation.

## IV. 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 [9]

$$\begin{cases} M_{CD}(k)=\sqrt{\dfrac{2E_c}{N_tT_c}}\left\{\hat{H}_{CD}(k)-\left(\dfrac{1}{N_c}\sum_{k'=0}^{N_c-1}\hat{H}_{CD}(k')\right)\right\}S(k) \\ M_{e(\text{or }o)}(k)=\sqrt{\dfrac{2E_c}{N_tT_c}}\left\{\hat{H}_{ST}(k)-\left(\dfrac{1}{N_c}\sum_{k'=0}^{N_c-1}\hat{H}_{ST}(k')\right)\right\}S_{e(\text{or }o)}(k) \end{cases} \quad ,(11)$$

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

$$\begin{cases} \hat{H}_{CD}(k)=W(k)H_{CD}(k) \quad \text{for CDTD} \\ \hat{H}_{ST}(k)=W_0(k)H_0^*(k)+W_1(k)H_1^*(k) \quad \text{for STTD} \end{cases} \quad . (12)$$

$M_{CD}(k)$ and $M_e(k)$ are shown in Fig. 2 for an $L$=16 path frequency-selective fading channel when $N_c$=256 and

$E_b/N_0$=10dB. CDTD enhances the frequency-selectivity of the channel and hence, 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.

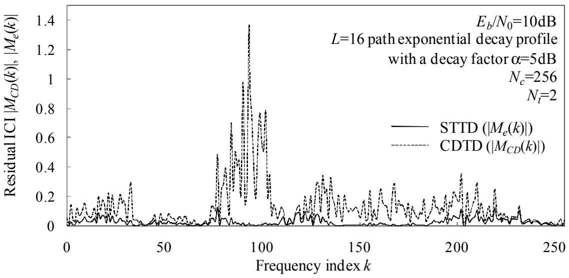

Figure 2 Frequency spectrum of residual ICI.

## V. ICI CANCELLATION

To improve the BER performance using CDTD and STTD, frequency-domain ICI cancellation is introduced [9]. A DS-CDMA receiver using joint MMSE-FDE and frequency-domain ICI cancellation is shown in Fig. 3. In this section, the process in the $i$th iteration is described.

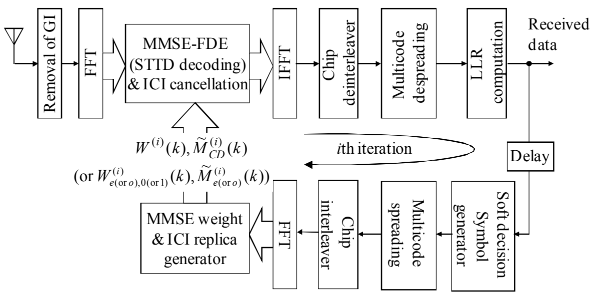

Figure 3 DS-CDMA receiver using joint MMSE-FDE and ICI cancellation.

### A. Joint MMSE-FDE and ICI cancellation

#### 1) CDTD

MMSE-FDE is carried out as

$$\hat{R}^{(i)}(k) = R(k)W^{(i)}(k) \quad , (13)$$

where $W^{(i)}(k)$ is the MMSE-FDE weight, taking into account the residual ICI, at the $i$th iteration. $W^{(i)}(k)$ is given by [9]

$$W^{(i)}(k) = \frac{H^*_{CD}(k)}{\rho^{(i-1)}|H_{CD}(k)|^2 + \left(\frac{1}{N_t}\frac{E_c}{N_0}\right)^{-1}} \quad , (14)$$

where $\rho^{(i-1)}$ (=0~$U$) indicates the extent to which the residual ICI remains and is given as

$$\rho^{(i-1)} = \frac{SF}{N_c}\sum_{m=0}^{N_c/SF-1}\sum_{u=0}^{U-1}\left\{E\left[|d_u(m)|^2\right] - \left|\widetilde{d}_u^{(i-1)}(m)\right|^2\right\} \quad (15)$$

with $\rho^{(-1)} = U$. $\{\widetilde{d}_u^{(i-1)}(m); m=0\sim N_c/SF-1\}$ is the soft decision replica of the transmitted symbol block $\{d_u(m)\}$ and $E[|d_u(m)|^2]$ is the expectation of $|d_u(m)|^2$ for the given received chip block.

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

$$\widetilde{R}^{(i)}(k) = \hat{R}^{(i)}(k) - \widetilde{M}^{(i)}_{CD}(k) \quad , (16)$$

where $\widetilde{M}^{(i)}_{CD}(k)$ is the replica of $M_{CD}(k)$, and is given by

$$\widetilde{M}^{(i)}_{CD}(k) = \sqrt{\frac{2E_c}{N_tT_c}}\left\{\hat{H}^{(i)}_{CD}(k) - \left(\frac{1}{N_c}\sum_{k'=0}^{N_c-1}\hat{H}^{(i)}_{CD}(k')\right)\right\}\widetilde{S}^{(i-1)}(k) \quad , (17)$$

where $\widetilde{M}^{(0)}_{CD}(k) = 0$ and $\hat{H}^{(i)}_{CD}(k) = W^{(i)}(k)H_{CD}(k)$ is the equivalent channel gain after MMSE-FDE at the $i$th iteration and $\{\widetilde{S}^{(i-1)}(k); k=0\sim N_c-1\}$ is the replica of $S(k)$. After ICI cancellation is done, $N_c$-point IFFT is applied to obtain the equalized chip block $\{\widetilde{r}^{(i)}(t); t=0\sim N_c-1\}$. Finally, despreading is carried out on $\{\widetilde{r}^{(i)}(t)\}$ to get the decision variable $\{\hat{d}_u^{(i)}(m); m=0\sim N_c/SF-1\}$ associated with $d_u(m)$.

#### 2) STTD

Joint MMSE-FDE and STTD decoding is performed using $R_e(k)$ and $R_o(k)$ as

$$\begin{cases}\hat{R}_e^{(i)}(k) = R_e(k)W^{(i)*}_{e,0}(k) + R^*_o(k)W^{(i)}_{e,1}(k)\\ \hat{R}_o^{(i)}(k) = R_e(k)W^{(i)*}_{o,1}(k) - R^*_o(k)W^{(i)}_{o,0}(k)\end{cases} \quad , (18)$$

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

$$W^{(i)}_{e(\text{or }o),\,0(\text{or }1)}(k) = \frac{H_{0(\text{or }1)}(k)}{\rho^{(i-1)}_{e(\text{or }o)}\left(|H_0(k)|^2 + |H_1(k)|^2\right) + \left(\frac{1}{N_t}\frac{E_c}{N_0}\right)^{-1}} \quad (19)$$

where $\rho^{(i-1)}_{e(\text{or }o)}$ indicates the extent to which the residual ICI remains and is given as

$$\rho^{(i-1)}_{e(\text{or }o)} = \frac{SF}{N_c}\sum_{m=0}^{N_c/SF-1}\sum_{u=0}^{U-1}\left\{E\left[|d_{e(\text{or }o),u}(m)|^2\right] - \left|\widetilde{d}^{(i-1)}_{e(\text{or }o),u}(m)\right|^2\right\} \quad (20)$$

where $\rho^{(-1)}_{e(\text{or }o)} = U$. $\{\widetilde{d}^{(i-1)}_{e(\text{or }o),u}(m)\ ; m=0\sim N_c/SF-1\}$ is the replica of the transmitted symbol block $\{d_{e(\text{or }o),u}(m)\}$ and $E[|d_{e(\text{or }o),u}(m)|^2]$ is the expectation of $|d_{e(\text{or }o),u}(m)|^2$ for the given received chip block. ICI cancellation is done for $\hat{R}^{(i)}_{e(\text{or }o)}(k)$ as

$$\widetilde{R}^{(i)}_{e(\text{or }o)}(k) = \hat{R}^{(i)}_{e(\text{or }o)}(k) - \widetilde{M}^{(i)}_{e(\text{or }o)}(k) \quad , (21)$$

where $\widetilde{M}^{(i)}_{e(\text{or }o)}(k)$ is the replica of $M_{e(\text{or }o)}(k)$, and is given from Eq. (11), by

$$\widetilde{M}^{(i)}_{e(\text{or }o)}(k) = \sqrt{\frac{2E_c}{N_tT_c}}\left\{\hat{H}^{(i)}_{e(\text{or }o)}(k) - \left(\frac{1}{N_c}\sum_{k'=0}^{N_c-1}\hat{H}^{(i)}_{e(\text{or }o)}(k')\right)\right\}\widetilde{S}^{(i-1)}_{e(\text{or }o)}(k) \quad , (22)$$

where $\widetilde{M}^{(0)}_{e(\text{or }o)}(k) = 0$ and

$\hat{H}^{(i)}_{e(\text{or } o)}(k)=W^{(i)}_{e(\text{or } o),0}(k)H^*_0(k)+W^{(i)}_{e(\text{or } o),1}(k)H^*_1(k)$ is the equivalent channel gain after joint MMSE-FDE and STTD decoding at the $i$th iteration. $\tilde{S}^{(i-1)}_{e(\text{or } o)}(k)$ is the replica of $S_{e(\text{or } o)}(k)$. After ICI cancellation, $N_c$-point IFFT is applied to obtain the chip block $\{\tilde{r}^{(i)}_{e(\text{or } o)}(t);t=0\sim N_c-1\}$. Finally, despreading is carried out to obtain the decision variable $\{\hat{d}^{(i)}_{e(\text{or } o),u}(m);m=0\sim N_c/SF-1\}$ associated with $\{d_{e(\text{or } o),u}(m)\}$.

### B. ICI replica generation

For the case of CDTD, using the decision variable $\hat{d}^{(i-1)}_u(m)$, the log-likelihood ratio (LLR) $\{L_u(x,m);x=0\sim\log_2 K-1 \text{ and } m=0\sim N_c/SF-1\}$ for the $x$th bit in the $m$th symbol $d_u(m)$ ($m$=0~$N_c/SF$−1), where $x$=0~$\log_2 K$−1 and $K$ is the modulation level, is computed [15]. For quaternary phase shift keying (QPSK) data-modulation, using $L_u(0,m)$ and $L_u(1,m)$, the soft symbol replica is obtained as

$$\tilde{d}^{(i-1)}_u(m)=\frac{1}{\sqrt{2}}\tanh\left(\frac{L_u(0,m)}{2}\right)+j\frac{1}{\sqrt{2}}\tanh\left(\frac{L_u(1,m)}{2}\right) \quad (23)$$

$E[|d_u(m)|^2]$ is also necessary for updating the MMSE-FDE weight and is obtained as $E[|d_u(m)|^2]=1$ for QPSK data-modulation. Using $\tilde{d}^{(i-1)}_u(m)$ and $E[|d_u(m)|^2]$, $\rho^{(i-1)}$ of Eq. (15) can be computed for updating the MMSE-FDE weight $W^{(i)}(k)$ for the $i$th iteration.

Then, $\tilde{d}^{(i-1)}_u(m)$ is spread and code-multiplexed to generate the chip block replica $\{\tilde{s}^{(i-1)}(t);t=0\sim N_c-1\}$. After applying $N_c$-point FFT to $\{\tilde{s}^{(i-1)}(t)\}$, the frequency-domain signal replica $\{\tilde{S}^{(i-1)}(k);k=0\sim N_c-1\}$ is obtained, which is used for generating the ICI replica of Eq. (17).

For the case of STTD, the above process is done for two consecutive blocks to compute $\rho^{(i-1)}_{e(\text{or } o)}$ of Eq. (20) and $\tilde{S}^{(i-1)}_{e(\text{or } o)}(k)$ in Eq. (22).

## VI. SNR FOR PERFECT ICI CANCELLATION

The BER lower-bound achievable with perfect ICI cancellation for CDTD and STTD can be derived using the signal-to-noise power ratio (SNR). Below, the SNR expression is derived assuming the perfect ICI cancellation (i.e., $\tilde{d}^{(i-1)}_u(m)=d_u(m)$ and $\tilde{S}^{(i-1)}(k)=S(k)$ for CDTD, while $\hat{d}^{(i-1)}_{e(\text{or } o),u}(m)=d_{e(\text{or } o),u}(m)$ and $\tilde{S}^{(i-1)}_{e(\text{or } o)}(k)=S_{e(\text{or } o)}(k)$ for STTD). Since $\rho^{(i)}=0$ and $\rho^{(i)}_{e(\text{or } o)}=0$ from Eqs. (15) and (20), MMSE-FDE weight is the same as the maximum ratio combining (MRC) weight given by [16]

$$\begin{cases}W^{(i)}(k)=H^*_{CD}(k) & \text{for CDTD}\\ W^{(i)}_{e,0(\text{or }1)}(k)=H_{0(\text{or }1)}(k) & \text{for STTD}\end{cases}, \quad (24)$$

where the weight for the even block is only considered for STTD. Assuming the perfect ICI cancellation and MRC weight, the SNR after despreading, denoted by $\gamma_{CD}$ and $\gamma_{ST}$ respectively for CDTD and for STTD, is given by [16]

$$\begin{cases}\gamma_{CD}=\frac{2E_s}{N_0}\left(\frac{1}{N_t}\frac{1}{N_c}\sum_{k=0}^{N_c-1}|H_{CD}(k)|^2\right) & \text{for CDTD}\\ \gamma_{ST}=\frac{2E_s}{N_0}\left(\frac{1}{N_t}\frac{1}{N_c}\sum_{k=0}^{N_c-1}\left(|H_0(k)|^2+|H_1(k)|^2\right)\right) & \text{for STTD}\end{cases}, \quad (25)$$

where $E_s/N_0$ is the signal energy per symbol-to-AWGN power spectrum density ratio. Substituting Eqs. (5) and (10) into Eq. (25) gives

$$\gamma_{CD}=\gamma_{ST}=\frac{2E_s}{N_0}\left(\frac{1}{N_t}\sum_{n=0}^{N_t-1}\sum_{l=0}^{L-1}|h_{n,l}|^2\right) \text{ for CDTD and STTD}. \quad (26)$$

The same SNR can be achieved for CDTD and STTD if the residual ICI is perfectly cancelled. This means 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 $\{h_{n,l}$; $n$=0~$N_t$−1 and $l$=0~$L$−1$\}$ is given by

$$p_b\left(\frac{E_s}{N_0},\{h_{n,l}\}\right)=\frac{1}{2}erfc\left[\sqrt{\frac{1}{4}\gamma_{CD(\text{or } ST)}}\right], \quad (27)$$

where $erfc[x]=(2/\sqrt{\pi})\int_x^\infty \exp(-t^2)dt$ is the complimentary error function. The BER lower-bound can be numerically evaluated by averaging Eq. (27) over all realizations of $\{h_{n,l}$; $n$=0~$N_t$−1 and $l$=0~$L$−1$\}$.

## VII. COMPUTER SIMULATION

We assume QPSK data-modulation, $N_c$=256 chips, $N_g$=32 chips. The channel is assumed to be a frequency-selective block Rayleigh fading channel having a chip-spaced $L$=16-path exponential power delay profile with a decay factor α=5 dB (e.g., the weak channel frequency-selectivity case). Perfect chip timing and ideal channel estimation are assumed. The number of iterations of joint MMSE-FDE and ICI cancellation is set to $i$=3, which provides sufficiently improved BER performance.

The simulated BER performance of DS-CDMA with MMSE-FDE is plotted in Fig. 4 as a function of the average received bit energy-to-AWGN noise power spectrum density ratio $E_b/N_0$, defined as $E_b/N_0$=0.5$SF$($E_c/N_0$)(1+$N_g/N_c$). The BER performance is plotted for CDTD, STTD and $N_t$=1 (i.e., no antenna diversity). For CDTD, a cyclic delay of Δ=32-chip is used [12]. For STTD using $N_t$=4, space-time coding with a coding rate of 3/4 is used [17]. We assume the spreading factor $SF$=16 and $U$=$SF$, i.e., full code-multiplexing.

We first discuss the case without ICI cancellation. The BER performance without ICI cancellation is plotted in Fig. 4 (a). CDTD can enhance the channel frequency selectivity and hence, improve the BER performance; however, the BER performance with CDTD is still inferior to that using STTD which can achieve the antenna diversity gain.

Fig. 4 (b) shows the simulated BER performance with ICI cancellation. The BER lower-bound computed using Eqs. (26) and (27) is also plotted for comparison. The use of ICI cancellation can significantly improve the BER performance. Since the frequency diversity gain achievable by FDE is smaller, the performance improvement 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 achieved by the enhancement of the frequency-selectivity for CDTD. Also seen is that CDTD provides a performance only slightly inferior to STTD. The reason for this performance difference is that the decision variables obtained at $i$=0 are more erroneous for CDTD than for STTD as seen from Fig. 4 (a) and hence, the accuracy of the soft symbol replica is worse for CDTD than for STTD. In a high $E_b/N_0$ region, however, both CDTD and STTD give almost the same BER performance as the lower-bound. With STTD (CDTD) using $N_t$=4, the $E_b/N_0$ reduction from the single transmit antenna case ($N_t$=1) is 6 (5.5) dB for BER=$10^{-4}$.

## VIII. Conclusion

In DS-CDMA/MMSE-FDE, STTD provides a better BER performance than CDTD. In this paper, we pointed out that this performance degradation of CDTD from STTD results from the residual ICI after MMSE-FDE. Then, we showed that the introduction of ICI cancellation gives almost the same BER performance for CDTD and STTD. We have shown that, with perfect ICI cancellation, CDTD can achieve the additional frequency diversity gain, which is equivalent to the antenna diversity gain achieved by STTD. This was confirmed by computer simulation. When $N_t$=4, ICI cancellation can reduce the required $E_b/N_0$ for BER=$10^{-4}$ by as much as 6 dB for STTD and 5.5 dB for CDTD compared to $N_t$=1.

STTD using more than three transmit antennas reduces the transmission data rate, while the transmission data rate of CDTD does not depend on the number of transmit antennas. Throughput comparison between CDTD and STTD is left for an interesting future work.

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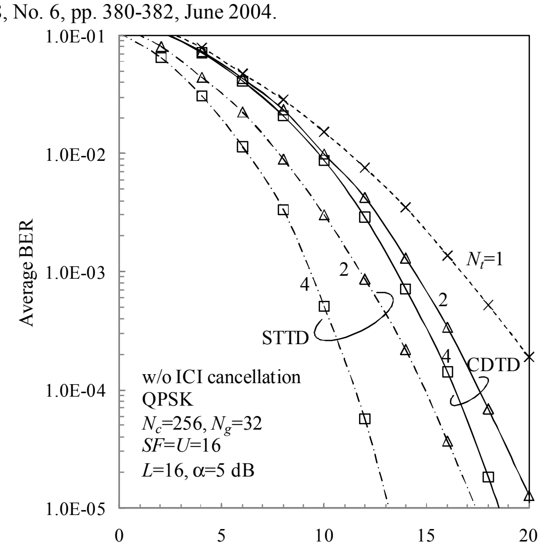
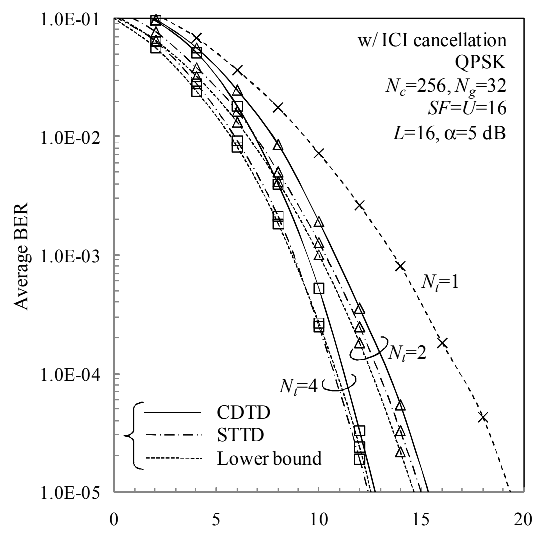



(a) *Without ICI cancellation.*




(b) *With ICI cancellation.*

Figure 4 Simulated BER performance of DS-CDMA/MMSE-FDE.