PAPER

Joint Space-Time Transmit Diversity and Minimum Mean Square Error Equalization for MC-CDMA with Antenna Diversity Reception

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SUMMARY In this paper, the space time transmit diversity (STTD) decoding combined with minimum mean square error (MMSE) equalization is presented for MC-CDMA downlink and uplink in the presence of multiple receive antennas. The equalization weights that minimize the MSE for each subcarrier are derived. From computer simulation, it was found that the BER performance of STTD decoding combined with MMSE equalization and M_r -antenna diversity reception using the weights derived in this paper provides the same diversity order as $2M_r$ -antenna receive diversity with MMSE equalization but with 3 dB performance can also be improved with STTD, but the error floor still exists. However, with 2-receive antennas in addition to 2-antenna STTD, the BER floor can be reduced to around 10^{-5} even for the uplink.

key words: mobile radio, MC-CDMA, MMSE equalization, space-time coding, transmit diversity

1. Introduction

Recently, the combination of multicarrier (MC) modulation based on orthogonal frequency division multiplexing and code division multiple access (CDMA), called MC-CDMA [1], [2], has gained a lot of attention because of its ability to allow high data rate transmission in a harsh mobile environment and has emerged as the most promising candidate for the forth generation of mobile communication systems [3]. In MC-CDMA, each user's data-modulated symbol to be transmitted is spread over a number of subcarriers using an orthogonal spreading code defined in the frequencydomain. Since the received signal suffers from frequency selective multipath fading, the orthogonality among different users' signals is partially lost, producing a large multiuser interference (MUI). However, the orthogonality property can be partially restored while achieving the frequency diversity effect by using the minimum mean square error (MMSE) equalization per subcarrier [4] and hence, a better bit error rate (BER) performance can be achieved.

Multiple antennas can be used to reduce the adverse effect of multipath fading. Receive diversity has been successfully used in practical systems. However, recently, transmit antenna diversity has been gaining much attention since the use of transmit diversity at a base station alleviates

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the complexity problem of mobile receivers [5]. Spacetime transmit diversity (STTD) [6] offers a way to introduce a degree of space diversity without going to the complexity of closed-loop transmit diversity solutions. In MC-CDMA, STTD decoding can be performed in conjunction with MMSE equalization to further improve the BER performance. In [7], STTD is applied to downlink MC-CDMA and it is shown that MC-CDMA with STTD has a better performance than without transmit diversity in multipath fading channels. Four detection schemes: maximum ratio combiner, orthogonal restoration combiner, MMSE combiner and MMSE multiuser detector are compared to show that MMSE combiner is the best tradeoff between performance and complexity [7]. However in [7], the uplink MC-CDMA case has not been treated.

In this paper, we derive the weights for joint STTD decoding and equalization based on MMSE (henceforth, referred to as joint STTD and MMSE equalization) for downlink and uplink MC-CDMA with multiple receive antennas. In the downlink case, all the users' data are spread using orthogonal codes and transmitted synchronously from the base station. At the receiver in the mobile terminal, the orthogonality among the users is partially lost, however frequency domain equalization can be utilized to restore the orthogonality. The orthogonality can be completely restored by the zero-forcing equalization, but it cannot take advantage of the frequency diversity effect that is present in a frequencyselective channel. MMSE equalization, on the other hand, restores orthogonality to a certain extent and also improves the performance owing to frequency diversity effect.

The uplink scenario is a little different. In the uplink case, users located at geographically different locations transmit asynchronously. Even when orthogonal codes are used, the orthogonality is completely destroyed as the users' channel gains and transmit timings are different. Hence for the uplink case, using orthogonal codes is not necessary; only the long scramble codes are used. If two antennas are available for diversity reception in the downlink case at the mobile terminal, these two antennas can be used for transmit diversity in the uplink case. Hence, in this paper we apply STTD for both the downlink and the uplink cases and combine the STTD decoding with the MMSE equalization and antenna diversity reception. In this paper, the weights for joint STTD and MMSE equalization are derived and the average BER performance is evaluated by computer simula-

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tions for various conditions.

The remainder of this paper is organized as follows. The transmission system model for MC-CDMA with joint STTD and MMSE equalization is presented in Sect. 2. The weights for joint STTD and MMSE equalization with multiple receive antennas are derived in Sect. 3. In Sect. 4, the simulation results for the downlink and uplink cases are presented and discussed. Section 5 concludes the paper.

2. Transmission System Model with Joint STTD and MMSE Equalization

Figure 1 shows the transmission system model with joint STTD and MMSE equalization. We consider MC-CDMA having N_c subcarriers with a carrier spacing of $1/T_s$ to transmit U users' data in parallel. Without loss of generality, we consider the two MC-CDMA signaling intervals, i.e., $0 \le t < 2T$ with $T = T_s + T_g$, where T_s and T_g are the effective symbol length and the guard interval (GI), respectively. Throughout the paper, discrete-time representation of the MC-CDMA signal is used.

2.1 Transmit Signal

At the transmitter, each user's data is spread using the frequency-domain orthogonal spreading code with a spreading factor *SF*. Let $d^u(n)$ be the *u*th user's *n*th data-modulated symbol with $|d^u(n)| = 1$. The STTD encoder encodes the

modulated symbols { $d^u(n)$; $n = 0 \sim 2(N_c/SF) - 1$ } for transmission during the signaling interval $0 \le t < 2T$ from the two antennas over the N_c subcarriers, where N_c/SF is an integer. The serial-to-parallel (S/P) converter converts the STTD encoded data into N_c/SF parallel data streams, each of which is copied SF times and multiplied by the orthogonal spreading code { $c^u(k)$; $k = 0 \sim SF - 1$ } for the *u*th user. It is then further multiplied by a long scramble sequence {pn(k); $k = \cdots, -1, 0, 1, \cdots$ }. The STTD encoded data waveform $y^u_{m_t}(n, t)$, to be transmitted from the m_t th antenna, $m_t=0,1$, over the $q \cdot SF$ th $\sim ((q + 1) \cdot SF - 1)$ th subcarriers, $q = 0 \sim N_c/SF - 1$, is as shown in Table 1.

Uplink is considered first. The *u*th user's *k*th subcarrier component $\{x_{m_t}^u(k,t); k = 0 \sim N_c - 1\}$, transmitted on the m_t th antenna during the time interval $t = 0 \sim T$ and $t = T \sim 2T$, may be expressed using the equivalent baseband representation as

Table 1 STTD encoded data waveform $y_{m_l}^u(n, t)$ transmitted over the $q \cdot SF$ th ~ $((q + 1) \cdot SF - 1)$ th subcarriers.

time antenna	$m_t=0$	$m_t=1$
t=0~T	$d_e^u = d^u(2q)$	$d_o^u = d^u (2q+1)$
t=T~2T	$-d_o^{u^*} = -d^{u^*}(2q+1)$	$d_e^{u^*} = d^{u^*}(2q)$



(b) Receiver

Fig. 1 Transmission system model.

$$x_{m_{t}}^{u}(k,t) = \sqrt{\frac{A}{SF}} c^{u}(k \mod SF) pn(k) y_{m_{t}}^{u} \left(\left\lfloor \frac{k}{SF} \right\rfloor, t \right)$$
for uplink, (1)

where *A* represents the total transmit power of each user, $y_{m_t}^u(n, t)$ is as shown in Table 1 for the respective signaling interval and antenna, and $\lfloor a \rfloor$ denotes the largest integer smaller than or equal to *a*. In Table 1 and henceforth, (.)* denotes the complex conjugate operation. N_c -point inverse fast Fourier transform (IFFT) is applied to the sequence $\{x_{m_t}^u(k, t); k = 0 \sim N_c - 1\}$ to generate the MC-CDMA signal $\{s_{m_t}^u(i, t); i = 0 \sim N_c - 1\}$:

$$s_{m_t}^u(i,t) = \frac{1}{N_c} \sum_{k=0}^{N_c-1} x_{m_t}^u(k,t) \exp\left(j2\pi k \frac{i}{N_c}\right),$$
(2)

where *i* represents the sample position in the signaling intervals $t = 0 \sim T$ and $t = T \sim 2T$. After insertion of the N_g -sample GI, the resultant MC-CDMA signal $\{\tilde{s}_{m_t}^u(i,t); i = -N_g \sim N_c - 1\}$ is transmitted over a propagation channel, where

$$\tilde{s}_{m_{c}}^{u}(i,t) = s_{m_{c}}^{u}(i \bmod N_{c},t).$$
(3)

In the downlink case, the U users' symbols spread on each subcarrier are added and then multiplied by the long scramble sequence. Hence the resultant *k*th subcarrier component $x_{m_t}(k, t)$ is given by

$$x_{m_t}(k,t) = \sqrt{\frac{A}{SF}} \sum_{u=0}^{U-1} c^u(k \mod SF) pn(k) y_{m_t}^u \left(\left\lfloor \frac{k}{SF} \right\rfloor, t \right)$$
for downlink. (4)

The resulting MC-CDMA signal $\{s_{m_t}(i, t); i = 0 \sim N_c - 1\}$ and the GI-inserted MC-CDMA signal $\{\tilde{s}_{m_t}(i, t); i = -N_g \sim N_c - 1\}$ are given by Eqs. (2) and (3), respectively, with $\{x_{m_t}^u(k, t)\}$ replaced by $\{x_{m_t}(k, t)\}$. The IFFT sampling period is taken to be $\Delta T = T_s/N_c$, such that $T_g = N_g \Delta T$ and $T = T_s + T_g = T_s(1 + N_g/N_c)$.

2.2 Channel Model

L independent propagation paths with distinct time delays $\{\tau_l^u\}$ are assumed. For the uplink case, the discrete time impulse response $\xi_m^u(\tau)$ of the multipath channel between the m_t th transmit antenna and the m_r th receive antenna for the *u*th user may be expressed as

$$\xi_{m_{l},m_{r}}^{u}(\tau) = \sum_{l=0}^{L-1} \xi_{m_{l},m_{r},l}^{u} \delta(\tau - \tau_{l}^{u})$$
(5)

with $\sum_{l=0}^{L-1} E[|\xi_{m_l,m_r,l}^u|^2] = 1$, where $\delta(t)$ is the delta function and E[.] denotes ensemble average. It is assumed that the channel impulse response remains the same for the two signaling intervals $t = 0 \sim T$ and $t = T \sim 2T$. The time delays $\{\tau_l^u\}$ are assumed to be multiples of the FFT sampling period ΔT .

For the downlink case, all users' signals go through the same channel and therefore, the superscript u representing the *u*th user is omitted from Eq. (5).

2.3 Received Signal

Uplink is considered first. We assume M_r -antenna receive diversity. The received MC-CDMA signal is sampled at the rate of $\Delta T^{-1} = N_c/T_s$ to obtain { $\tilde{r}_{m_r}(i, t)$; $i = -N_g \sim N_c - 1$ }, which is expressed as

$$\tilde{r}_{m_r}(i,t) = \sum_{m_i=0}^{1} \sum_{u=0}^{U-1} \sum_{l=0}^{L-1} \xi^u_{m_i,m_r,l} \tilde{s}^u_{m_l}(i - \tau^u_l / \Delta T, t) + \eta_{m_r}(i,t), \quad (6)$$

where $\eta_{m_r}(i, t)$ represents the additive white Gaussian noise (AWGN) process at the sampling instant *i* within the signaling interval *t* for the m_r th receive antenna and has zero mean and variance $2N_0/T_s$, with N_0 representing the single sided AWGN power spectrum density. It is assumed for simplicity that even for uplink, all the users are synchronized, but the impact of transmit timing asynchronism is evaluated by computer simulations in Sect. 4. On the other hand, the received MC-CDMA signal for the downlink can be given by Eq. (6) with $\xi^u_{m_r,m_r}(\tau)$ and $\{\tau^u_l\}$ replaced by $\xi_{m_t,m_r}(\tau)$ and $\{\tau_l\}$, respectively.

The N_g -sample GI is removed and the N_c -point FFT is applied to decompose the received MC-CDMA signal into the N_c -subcarrier components { $r_{m_r}(k, t)$; $k = 0 \sim N_c - 1$ }:

$$r_{m_r}(k,t) = \sum_{i=0}^{N_c-1} \tilde{r}_{m_r}(i,t) \exp\left(-j2\pi k \frac{i}{N_c}\right).$$
 (7)

Denoting the *u*th user's channel gain at the *k*th subcarrier for the m_t th transmit antenna and m_r th receive antenna by $H^u_{m_t,m_r}(k)$, the *k*th subcarrier component $r_{m_r}(k, 0)$ received in the signaling interval $t = 0 \sim T$ by the m_r th antenna may be represented as

$$r_{m_r}(k,0) = \sum_{m_t=0}^{1} \sum_{u=0}^{U-1} H^u_{m_t,m_r}(k) x^u_{m_t}(k,0) + \eta_{m_r}(k,0)$$
for uplink, (8a)

where $x_{m_t}^u(k, t)$ is as defined in Eq. (1). Likewise, the signal received during the signaling interval $t = T \sim 2T$ can be expressed as

$$r_{m_r}(k,T) = \sum_{m_t=0}^{1} \sum_{u=0}^{U-1} H^u_{m_t,m_r}(k) x^u_{m_t}(k,T) + \eta_{m_r}(k,T)$$
for uplink. (8b)

In Eqs. (8a) and (8b), $\{H_{m_t,m_r}^u(k); k = 0 \sim N_c - 1\}$ and $\{\eta_{m_r}(k,t); k = 0 \sim N_c - 1\}$ are respectively the fast Fourier transforms of the channel impulse response $\xi_{m_t,m_r}^u(\tau)$ and the AWGN process $\eta_{m_r}(i, t)$. They are given by

$$\begin{cases} H^{u}_{m_{r},m_{r}}(k) = \sum_{l=0}^{L-1} \xi^{u}_{m_{r},m_{r},l} \exp\left(-j2\pi k \frac{\tau^{u}_{l}/\Delta T}{N_{c}}\right) \\ \eta_{m_{r}}(k,t) = \sum_{i=0}^{N_{c}-1} \eta_{m_{r}}(i,t) \exp\left(-j2\pi k \frac{i}{N_{c}}\right) \end{cases}$$
(9)

For the downlink, the *k*th subcarrier components $r_{m_r}(k, 0)$ and $r_{m_r}(k, T)$ received in the signaling intervals $t = 0 \sim T$ and $t = T \sim 2T$ may be represented as

$$r_{m_r}(k,t) = \sum_{m_t=0}^{1} H_{m_t,m_r}(k) x_{m_t}(k,t) + \eta_{m_r}(k,t),$$

for downlink (10)

where $H_{m_t,m_r}(k)$ is the channel gain of the downlink at the *k*th subcarrier for the m_t th transmit antenna and m_r th receive antenna and $x_{m_t}(k, t)$ is as defined in Eq. (4).

The STTD decoding combined with MMSE equalization is carried out using the equalization weights as derived in Sect. 3 and the resulting soft samples { $\tilde{x}_{m_r}(k, t)$; $k = qSF \sim (q + 1)SF - 1$ } received by M_r -antennas are added and then despread, i.e., multiplied by the scramble sequence {pn(k)} and the orthogonal spreading code { $c^u(k \mod SF)$; $k = qSF \sim (q + 1)SF - 1$ } and summed, to obtain the decision variable $\tilde{d}^u(q, t)$ for the *q*th data-modulated symbol of the *u*th user transmitted in the signaling interval *t* given as

$$\tilde{d}^{u}(q,t) = \sum_{k=qSF}^{(q+1)SF-1} \sum_{m_{r}=0}^{M_{r}-1} \tilde{x}_{m_{r}}(k,t) c^{u*}(k \mod SF) pn^{*}(k).$$
(11)

The recovered modulated symbols for the two signaling intervals $t = 0 \sim T$ and $t = T \sim 2T$ are then aligned to obtain the *n*th data-modulated symbol $\hat{d}^u(n)$ as

$$\hat{d}^{u}(n) = \begin{cases} \tilde{d}^{u}(q,0), & \text{if } n = 2q \\ \tilde{d}^{u}(q,T), & \text{if } n = 2q + 1 \end{cases}$$
(12)

for $n = 0 \sim 2(N_c/SF) - 1$.

3. Equalization Weights for Joint STTD Decoding and MMSE Equalization

The STTD decoding is performed on each subcarrier and then the frequency-domain equalization is carried out using MMSE equalization. In this paper, we derive the equalization weights for joint STTD decoding and MMSE equalization. The STTD decoding is carried out as follows:

$$\begin{cases} \tilde{x}_{m_r}(k,0) = w_{0,m_r}^*(k)r_{m_r}(k,0) + w_{1,m_r}(k)r_{m_r}^*(k,T) \\ \tilde{x}_{m_r}(k,T) = w_{1,m_r}^*(k)r_{m_r}(k,0) - w_{0,m_r}(k)r_{m_r}^*(k,T) \end{cases},$$
(13)

where $w_{0,m_r}(k)$ and $w_{1,m_r}(k)$ are the equalization weights for the m_r th receive antenna. The equalization weights for both signaling intervals $t = 0 \sim T$ and $t = T \sim 2T$ are the same since we are assuming that the channel impulse response remains the same over the two signaling intervals; hence we use the first signaling interval to derive the equalization weights. Without loss of generality, we assume that the 0th user is the desired user. The instantaneous error $\varepsilon(k)$ of the *k*th subcarrier soft sample in the signaling interval $t = 0 \sim T$ is defined as

$$\varepsilon(k) = \tilde{x}(k,0) - x(k,0), \tag{14}$$

where

$$\tilde{x}(k,0) = \sum_{m_r=0}^{M_r-1} \tilde{x}_{m_r}(k,0)$$
(15)

and

$$x(k, 0) = \begin{cases} \sqrt{\frac{A}{SF}} \sum_{u=0}^{U-1} c^{u}(k \mod SF) pn(k) d_{e}^{u} \left(\left\lfloor \frac{k}{SF} \right\rfloor, 0 \right) \\ \text{for downlink} \\ \sqrt{\frac{A}{SF}} c^{0}(k \mod SF) pn(k) d_{e}^{0} \left(\left\lfloor \frac{k}{SF} \right\rfloor, 0 \right) \\ \text{for uplink} \end{cases}$$
(16)

is the symbol transmitted during the signaling interval $t = 0 \sim T$ from the 0th antenna $(m_t=0)$. We want to find the set of equalization weights $w_{0,m_r}(k)$ and $w_{1,m_r}(k)$ that minimizes the mean square error (MSE) $E[|\varepsilon(k)|^2]$. It is assumed that the U data-modulated symbols $\{d^u(n); u = 0 \sim U - 1\}$ are zero-mean and independent random variables.

3.1 Downlink

For the downlink, the channel gain for all users' transmitted symbols are the same and the MSE for the given set of $\{H_{m_t,m_r}(k); m_t = 0, 1; m_r = 0 \sim M_r\}$ becomes

$$\begin{split} E\left[|\varepsilon(k)|^{2}\right] &= E\left[|x(k,0)|^{2}\right] + E\left[|\tilde{x}(k,0)|^{2}\right] \\ &- 2\operatorname{Re}\left[E\left[\tilde{x}_{m_{r}}(k,0)x^{*}(k,0)\right]\right] = U\frac{A}{SF} \\ &\cdot \left\{\left(\sum_{m_{r}=0}^{M_{r}-1}w_{0,m_{r}}^{*}(k)H_{0,m_{r}}(k)\right)^{2} + \left(\sum_{m_{r}=0}^{M_{r}-1}w_{1,m_{r}}(k)H_{1,m_{r}}^{*}(k)\right)^{2} \\ &+ \left(\sum_{m_{r}=0}^{M_{r}-1}w_{0,m_{r}}^{*}(k)H_{1,m_{r}}(k)\right)^{2} + \left(\sum_{m_{r}=0}^{M_{r}-1}w_{1,m_{r}}(k)H_{0,m_{r}}^{*}(k)\right)^{2}\right\} \\ &+ U\frac{A}{SF} + 2\frac{N_{0}}{T_{s}}\sum_{m_{r}=0}^{M_{r}-1}\left(\left|w_{0,m_{r}}(k)\right|^{2} + \left|w_{1,m_{r}}(k)\right|^{2}\right) \\ &- 2U\frac{A}{SF}\operatorname{Re}\left[\sum_{m_{r}=0}^{M_{r}-1}w_{0,m_{r}}^{*}(k)H_{0,m_{r}}(k) + \sum_{m_{r}=0}^{M_{r}-1}w_{1,m_{r}}(k)H_{1,m_{r}}^{*}(k)\right] \end{split}$$

$$(17)$$

Expectation is taken over the noise variables and the scrambling codes for the given values of $\{H_{m_r,m_r}(k)\}$. Letting

$$\begin{cases} w_{m_t,m_r}(k) = a_{m_t,m_r} + jb_{m_t,m_r} \\ H_{m_t,m_r}(k) = u_{m_t,m_r} + jv_{m_t,m_r} \end{cases},$$
(18)

the MSE is given by

 $E[|\varepsilon(k)|^2]$

$$= U \frac{A}{SF} \left\{ \left(\sum_{m_r=0}^{M_r-1} a_{0,m_r} u_{0,m_r} + b_{0,m_r} v_{0,m_r} \right)^2 + \left(\sum_{m_r=0}^{M_r-1} a_{0,m_r} v_{0,m_r} - b_{0,m_r} u_{0,m_r} \right)^2 + \left(\sum_{m_r=0}^{M_r-1} a_{1,m_r} u_{1,m_r} + b_{1,m_r} v_{1,m_r} \right)^2 + \left(\sum_{m_r=0}^{M_r-1} a_{1,m_r} v_{1,m_r} - b_{1,m_r} u_{1,m_r} \right)^2 + \left(\sum_{m_r=0}^{M_r-1} a_{0,m_r} u_{1,m_r} + b_{0,m_r} v_{1,m_r} \right)^2 + \left(\sum_{m_r=0}^{M_r-1} a_{0,m_r} v_{1,m_r} - b_{0,m_r} u_{1,m_r} \right)^2 + \left(\sum_{m_r=0}^{M_r-1} a_{1,m_r} u_{0,m_r} + b_{1,m_r} v_{0,m_r} \right)^2 + \left(\sum_{m_r=0}^{M_r-1} a_{1,m_r} v_{0,m_r} - b_{1,m_r} u_{0,m_r} \right)^2 + \left(\sum_{m_r=0}^{M_r-1} a_{1,m_r} v_{0,m_r} - b_{1,m_r} u_{0,m_r} \right)^2 + U \frac{A}{SF} + \frac{2N_0}{T_s} \sum_{m_r=0}^{M_r-1} \left(a_{0,m_r}^2 + b_{0,m_r}^2 + a_{1,m_r}^2 + b_{1,m_r}^2 \right) + 2U \frac{A}{SF} \sum_{m_r=0}^{M_r-1} [(a_{0,m_r} u_{0,m_r} + b_{0,m_r} v_{0,m_r} + a_{1,m_r} u_{1,m_r} + b_{1,m_r} v_{1,m_r})].$$

From $\partial \boldsymbol{E}[|\boldsymbol{\varepsilon}(k)|^2] / \partial w_{0,m_r}(k) = 0$, i.e., $\partial \boldsymbol{E}[|\boldsymbol{\varepsilon}(k)|^2] / \partial a_{0,m_r} = 0$ and $\partial \boldsymbol{E}[|\boldsymbol{\varepsilon}(k)|^2] / \partial b_{0,m_r} = 0$, we get

$$H_{0,m_r}(k) \sum_{m_r=0}^{M_r-1} \left(w_{0,m_r}(k) H_{0,m_r}^*(k) \right) + H_{1,m_r}(k) \sum_{m_r=0}^{M_r-1} \left(w_{0,m_r}(k) H_{1,m_r}^*(k) \right) + \frac{2N_0}{T_s} \frac{SF}{UA} w_{0,m_r}(k) - H_{0,m_r}(k) = 0.$$
(20)

After some manipulations, we obtain

$$w_{0,m_r}(k) = \frac{H_{0,m_r}(k)}{\sum_{m_r=0}^{1} \sum_{m_r=0}^{M_r-1} \left| H_{m_r,m_r}(k) \right|^2 + \left(\frac{U}{SF} \frac{T_s A}{2N_0} \right)^{-1}}.$$
 (21)

In a similar manner, from $\partial E[|\varepsilon(k)|^2] / \partial w_{1,m_r}(k) = 0$ we obtain

$$w_{1,m_r}(k) = \frac{H_{1,m_r}(k)}{\sum_{m_r=0}^{1} \sum_{m_r=0}^{M_r-1} \left| H_{m_r,m_r}(k) \right|^2 + \left(\frac{U}{SF} \frac{T_s A}{2N_0} \right)^{-1}}.$$
 (22)

3.2 Uplink

For the uplink, on the other hand, the channel gain is different for different users, i.e., $H^u_{0,m_r}(k) \neq H^{u'}_{0,m_r}(k)$ and $H^u_{1,m_r}(k) \neq H^{u'}_{1,m_r}(k)$ if $u \neq u'$. Slow transmit power control is assumed so that all users' signals are received at the base station with the same average power. Hence, the MSE is as follows:

$$\begin{split} E\left[|\varepsilon(k)|^{2}\right] &= E\left[|x(k,0)|^{2}\right] + E\left[\left|\tilde{x}_{m_{r}}(k,0)\right|^{2}\right] \\ &-2\operatorname{Re}[\tilde{x}_{m_{r}}(k,0)x^{*}(k,0)] \\ &= \frac{A}{SF}\left\{\sum_{u=0}^{U-1}\left|\sum_{m_{r}=0}^{M_{r}-1}\left(w_{0,m_{r}}^{*}(k)H_{0,m_{r}}^{u}(k) + w_{1,m_{r}}(k)H_{1,m_{r}}^{u*}(k)\right)\right|^{2} \\ &+ \sum_{u=0}^{U-1}\left|\sum_{m_{r}=0}^{M_{r}-1}\left(w_{0,m_{r}}^{*}(k)H_{1,m_{r}}^{u}(k) + w_{1,m_{r}}(k)H_{0,m_{r}}^{u*}(k)\right)\right|^{2}\right\} \\ &+ \frac{A}{SF} + \frac{N_{0}}{T_{s}}\left(\sum_{m_{r}=0}^{M_{r}-1}\left(\left|w_{0,m_{r}}(k)\right|^{2} + \left|w_{1,m_{r}}(k)\right|^{2}\right)\right) \\ &- 2\frac{A}{SF}\operatorname{Re}\left[\sum_{m_{r}=0}^{M_{r}-1}w_{0,m_{r}}^{*}(k)H_{0,m_{r}}^{0}(k) + \sum_{m_{r}=0}^{M_{r}-1}w_{1,m_{r}}(k)H_{1,m_{r}}^{0*}(k)\right]. \end{split}$$

$$\tag{23}$$

Expectation is taken over the noise variables and the scrambling codes for the given values of $\{H_{m_r,m_r}^u(k)\}$. Letting

$$\begin{cases} w_{m_t,m_r}(k) = a_{m_t,m_r} + jb_{m_t,m_r} \\ H^u_{m_t,m_r}(k) = g^u_{m_t,m_r} + jh^u_{m_t,m_r} \end{cases},$$
(24)

the MSE is given by

(19)

$$\begin{split} E[|\varepsilon(k)|^2] \\ &= \frac{A}{SF} \sum_{u=0}^{U-1} \left\{ \left(\sum_{m_r=0}^{M_r-1} \left(a_{0,m_r} g_{0,m_r}^u - b_{0,m_r} h_{0,m_r}^u \right) \right)^2 \\ &\quad + \left(\sum_{m_r=0}^{M_r-1} \left(a_{0,m_r} h_{0,m_r}^u + b_{0,m_r} g_{0,m_r}^u \right) \right)^2 \\ &\quad + \left(\sum_{m_r=0}^{M_r-1} \left(a_{1,m_r} g_{1,m_r}^u - b_{1,m_r} h_{1,m_r}^u \right) \right)^2 \\ &\quad + \left(\sum_{m_r=0}^{M_r-1} \left(a_{1,m_r} h_{1,m_r}^u + b_{1,m_r} g_{1,m_r}^u \right) \right)^2 \\ &\quad + \left(\sum_{m_r=0}^{M_r-1} \left(a_{0,m_r} g_{1,m_r}^u - b_{0,m_r} h_{1,m_r}^u \right) \right)^2 \\ &\quad + \left(\sum_{m_r=0}^{M_r-1} \left(a_{0,m_r} g_{1,m_r}^u - b_{0,m_r} h_{1,m_r}^u \right) \right)^2 \\ &\quad + \left(\sum_{m_r=0}^{M_r-1} \left(a_{0,m_r} h_{1,m_r}^u + b_{0,m_r} g_{1,m_r}^u \right) \right)^2 \end{split}$$

$$+ \left\{ \sum_{m_r=0}^{M_r-1} \left(a_{1,m_r} h_{0,m_r}^u + b_{1,m_r} g_{0,m_r}^u \right) \right\}^2 \right\}$$
$$+ \frac{A}{SF} + \frac{2N_0}{T_s} \sum_{m_r=0}^{M_r-1} \left(a_{0,m_r}^2 + b_{0,m_r}^2 + a_{1,m_r}^2 + b_{1,m_r}^2 \right) - 2 \frac{A}{SF}$$
$$\cdot \sum_{m_r=0}^{M_r-1} \left[\left(a_{0,m_r} g_{0,m_r}^0 + b_{0,m_r} h_{0,m_r}^0 + a_{1,m_r} g_{1,m_r}^0 + b_{1,m_r} h_{1,m_r}^0 \right) \right].$$
(25)

From $\partial \boldsymbol{E}[|\boldsymbol{\varepsilon}(k)|^2] / \partial w_{0,m_r}(k) = 0$, i.e., $\partial \boldsymbol{E}[|\boldsymbol{\varepsilon}(k)|^2] / \partial a_{0,m_r} = 0$ and $\partial \boldsymbol{E}[|\boldsymbol{\varepsilon}(k)|^2] / \partial b_{0,m_r} = 0$, we get

$$\frac{2A}{SF} \sum_{u=0}^{U-1} H_{0,m_r}^{u*}(k) \sum_{m'_r=0}^{M_r-1} w_{0,m'_r}(k) H_{0,m'_r}^{u}(k) + \sum_{u=0}^{U-1} H_{1,m_r}^{u*}(k) \sum_{m'_r=0}^{M_r-1} w_{0,m'_r}(k) H_{1,m'_r}^{u}(k) + \frac{4N_0}{T_s} w_{0,m_r}(k) - 2\frac{A}{SF} H_{0,m_r}^0(k) = 0.$$
(26)

It is very difficult if not impossible to solve Eq. (26). Hence, in this paper, we find a suboptimal solution. In Eq. (26), we take the expectation over phase difference between $\{H_{m_t,m_r}(k)\}$ for the given $|H_{m_t,m_r}(k)|$ i.e., $E[H_{m_t,m_r}(k) \cdot H_{m_t,m'_r}(k)] = 0$ if $m_r \neq m'_r$, and hence Eq. (26) simplifies to

$$\frac{2A}{SF} \sum_{u=0}^{U-1} \left| H_{0,m_r}^u(k) \right|^2 w_{0,m_r}(k) + \sum_{u=0}^{U-1} \left| H_{1,m_r}^u(k) \right|^2 w_{0,m_r}(k) + \frac{4N_0}{T_s} w_{0,m_r}(k) - 2\frac{A}{SF} H_{0,m_r}^0(k) = 0.$$
(27)

From (27) we get $w_{0,m_r}(k)$ as

$$w_{0,m_r}(k) = \frac{H_{0,m_r}^0(k)}{\sum_{m_t=0}^{1} \sum_{u=0}^{U-1} \left| H_{m_t,m_r}^u(k) \right|^2 + \left(\frac{1}{SF} \frac{T_s A}{2N_0}\right)^{-1}}.$$
 (28)

In a similar manner, from $\partial E[|\varepsilon(k)|^2]/\partial w_{1,m_r}(k) = 0$ we obtain

$$w_{1,m_r}(k) = \frac{H_{1,m_r}^0(k)}{\sum_{m_t=0}^1 \sum_{u=0}^{U-1} \left| H_{m_t,m_r}^u(k) \right|^2 + \left(\frac{1}{SF} \frac{T_s A}{2N_0}\right)^{-1}}.$$
 (29)

With the weights derived in Eqs. (21), (22), (28) and (29), joint STTD and MMSE equalization is performed as in Eq. (13).

4. Simulation Results

Table 2 shows the computer simulation conditions. We

assume MC-CDMA using $N_c=256$ subcarriers with a carrier spacing of $1/T_s$, GI of $T_g = T_s/8$ (i.e., $N_g=32$), and ideal coherent BPSK data-modulation. IFFT and FFT sampling period ΔT is $\Delta T = T_s/256$. A frequency-selective Rayleigh fading channel having L=16-path uniform power delay profile with $\tau_l = 2l\Delta T$ for downlink ($\tau_l^u = 2l\Delta T$, $u = 0 \sim U - 1$, for uplink), and the normalized maximum Doppler frequency $f_D T = 0.01$ is assumed. Uncorrelated, time-varying Rayleigh faded paths are generated using Dent's model [8]. The estimations of the channel gains, AWGN power spectrum density and the number of users are assumed to be ideal.

4.1 Downlink

Figure 2 plots the downlink BER performance as a function of the average received signal energy per bit-to-the AWGN power spectrum density ratio (E_b/N_0) , $E_b/N_0 = (AT_s/N_0)(1 + T_g/T_s)$, for SF=32 with the number U of communicating users as a parameter. For reference, the BER with MMSE equalization but for no diversity and that for

Table 2Simulation conditions.

Data modulation		Coherent BPSK
MC-CDMA	No. of subcarriers	N _c =256
	Effective symbol length	<i>T_s</i> =256∆ <i>T</i>
	GI	$T_g=32\Delta T$ $(T_g/T_s=1/8)$
	Spreading factor	SF=1~256
Channel model	Rayleigh fading (L=16)	
	$\tau_l = 2l\Delta T$	
		$f_D T = 0.01$







Fig.3 Downlink BER performance for U = SF.

2-antenna receive diversity combined with MMSE equalization as in [9] are also plotted. The BER performance worsens with the increase in the number of users due to increasing MUI. However, it is seen that the STTD achieves the performance of 2-antenna receive diversity but with a penalty of 3 dB.

In the downlink, all users' signal transmissions are synchronous and orthogonal spreading codes can be used to reduce the MUI. This suggests that as many as SF users can be multiplexed on the downlink without significant performance degradation. Figure 3 plots the BER as a function of the average received E_b/N_0 with SF as a parameter when the number U of users is the same as SF. SF = U = 1 corresponds to the well-known orthogonal frequency division multiplexing (OFDM). It is interesting to note that the BER performance of MC-CDMA with SF > 1 with U = SF is seen to be better than that of SF=1 in spite of the increase in MUI. This is because as the value of SF increases the increase in frequency diversity effect becomes large enough to offset the increase in MUI and also provide additional improvement [10]. With STTD, the average received E_b/N_0 for a BER of 10^{-4} is 8 dB less for SF=256 than that for SF=1. The no diversity and 2-antenna receive diversity curves are also plotted. It is again seen that the STTD achieves the performance of 2-antenna receive diversity but with a 3 dB penalty.

4.2 Uplink

In the uplink case, the orthogonality among the users cannot be maintained in a frequency selective channel. Hence, a BER error floor occurs when more than one user are in communication at the same time. The weights for joint STTD and MMSE equalization were derived in Sect. 3 for the case



when the users are synchronous. However, in actuality, the users' transmitting timings are asynchronous. It was found from our preliminary simulation that the BER performances for the synchronous and asynchronous cases are almost identical. Figure 4 plots the uplink BER performance as a function of the average received E_b/N_0 for SF=32 with U as a parameter for the asynchronous case when the users' timing are uniformly distributed over $0 \sim T$. For reference, the BER performances for no diversity with MMSE equalization and 2-antenna receive diversity combined with MMSE equalization are also plotted. Using STTD with MMSE equalization improves the performance, however the

BER floor still exists.

In [1] it is shown that MRC equalization provides a better BER performance than equal gain combining (EGC) equalization and hence, we do not consider EGC equalization here. The BER performance is compared for MMSE equalization and MRC equalization in Fig. 5. It can be seen from Fig. 5 that MMSE equalization gives better performance than MRC equalization, but gives almost the same performance as MRC equalization for U > 4 (U > 8) with STTD (without STTD). When many users exit, the interference can be approximated by a Gaussian variable according to the central limit theorem [11]. Sum of the interference approximated by a Gaussian variable and the Gaussian noise is also Gaussian. This condition can be treated as a single user condition with increased noise and hence the MRC equalization would be the best. This is confirmed by the fact that the term $\sum_{m_t=0}^{1} \sum_{u=0}^{U-1} |H_{m_t,m_r}^u(k)|^2$ in the denominators of Eqs. (28) and (29) approaches the same constant for all subcarriers when U is large, according to the "Law of large numbers" [11]. Hence the weights for joint STTD and MMSE equalization are in fact the same as those for MRC

(BS, MT) = (1, 1)

(2, 1)

Downlink

Uplink

 10^{0}

10

 10^{-2}

10-3

10-4

10

0

Fig. 7

Average BER

SF=32

(2,

5

10

15

Average received E_1/N_0 [dB]

BER performance with STTD and $M_r=2$.

20

25

30

U=8

Fig. 5 Uplink BER performance MRC and MMSE equalization.

No transmit diversity

STTD



 10^{0}

10

Average BER

10

 10^{-4}

 10^{-5}

Fig. 6

0

5

SF=U=32

4.3 Combined Effect of STTD and Diversity Reception

10

Average received $E_{\rm h}/N_{\rm o}$ per rx. antenna [dB]

Downlink BER performance for U = SF=32.

15

20

25

So far, we have treated STTD and antenna receive diversity separately. Below, we evaluate the combined effect of STTD and receive antenna diversity. Figure 6 plots the average BER performance as a function of the average received E_b/N_0 per antenna with M_r as a parameter for downlink when SF=U=32. For reference, the average BER performance for no transmit diversity is also plotted. It is seen that



Figure 7 plots the average BER performance as a function of the average received E_b/N_0 per antenna for both uplink and downlink when SF=32 and U=8. Three cases are considered: (BS, MT) = (1, 1), i.e., 1-antenna at both BS and MT, (BS, MT) = (2, 1) and (BS, MT) = (2, 2). With (BS, MT) = (2,1) the downlink performance improves due to STTD gain and the uplink performance improves due to receive diversity gain. With (BS, MT) = (2, 2), a large improvement is attained for both downlink and uplink because of the combined effect of STTD and receive diversity. The uplink error floor is reduced to around 10^{-5} . The result encourages the use of two antennas at the mobile terminal for receive diversity on the downlink and for STTD on the uplink.

5. Conclusion

In this paper, the STTD decoding combined with MMSE equalization was presented for MC-CDMA and the equalization weights were derived where the weights minimize the MSE for each subcarrier. It was found that the equaliza-



tion weights for downlink and uplink are different. From computer simulation, it was found that the BER performance of STTD decoding combined with MMSE equalization and M_r -antenna diversity reception using the weights derived in this paper provides the same diversity order as $2M_r$ -antenna receive diversity with MMSE equalization but with 3 dB performance penalty and is always better than that with no diversity. The uplink BER performance can also be improved with STTD, but the error floor still exists. However, with 2-receive antennas in addition to 2-antenna STTD, the BER floor can be reduced to around 10^{-5} even for uplink MC-CDMA.

In this paper, the estimations of the channel gains, AWGN power spectrum density and the number of users are assumed to be ideal. The estimation errors would result in performance degradation. It is interesting to see how much the performance would degrade in an MC-CDMA system with STTD when a practical channel estimator as in [12] is used. It is a practically important future study. Also interesting would be to study the performance with channel coding and higher level modulation. OFDM has a higher coding gain than MC-CDMA in a frequency selective channel. So it is important to study MC-CDMA with STTD and channel coding employing higher level modulation. We considered a single user MC-CDMA receiver based on persubcarrier MMSE equalization. Joint-subcarrier multiuser MC-CDMA receiver can be used to improve the uplink BER performance [13]. The study of the multiuser MC-CDMA receiver in the presence of STTD is left as an interesting future study.

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