Two-Dimensional Block-Spread CDMA Relay Using Virtual-Four-Antenna STCDTD

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***Abstract*—Cooperative communication networks can exploit the spatial diversity inherent in a multiuser system by allowing users to cooperate and relay other users' messages to a destination. Avoiding the multiple-access interference (MAI) while taking full advantage of the multiuser cooperative multiple-input–multiple-output (Co-MIMO) is still an interesting problem. In this paper, we propose a relay scheme that combines the source user's two antennas and the relay user's two antennas to form a virtual-four-antenna space–time cyclic delay transmit diversity (STCDTD), and explore 2-D block-spread code-division multiple access (CDMA) to achieve MAI-free uplink transmission. The theoretical error probability of the proposed scheme is described, when a signal is transmitted through a multiple-access Rayleigh fading channel. Computer simulations indicate that the proposed scheme is able to avoid the MAI over a slow fading channel and to improve the bit error rate (BER) significantly.**

***Index Terms*—Cooperative communications, multiple-input–multiple-output, space–time cyclic delay transmit diversity (STCDTD), 2-D block-spread code-division multiple access (CDMA).**

I. Introduction

BROADBAND data services with a peak data rate of around 1 Gb/s will be demanded of future wireless communications [1]–[3]. To support such a high data rate, two important technical issues should be addressed: transmission power and spectrum efficiency.

Transmission power dramatically increases with an increasing data rate. A promising technique that can keep the transmission power the same as in the current cellular systems is the multihop relay [4]. The transmission power of a mobile terminal and the total transmission power, including relays, can be significantly reduced by applying multihop relay techniques [5]–[7]. Multihop relay techniques allow users to cooperate and relay other users' messages to a base station (BS) via diverse channels. If channel state information (CSI) is available, each user is given a fair opportunity to utilize the cooperative relaying channel and then to redistribute the resource or traffic load to improve communication efficiency [8]. Cooperative communications can be operated both between different BSs and between different users [9]–[11]. For the latter, since different users' signals are asynchronously received via different fading channels, multiple-access interference (MAI) occurs. The code-division multiple-access (CDMA) technique can be used to suppress the MAI [12]–[20].

However, the capability of MAI suppression achievable with pure CDMA is not sufficient. In the past, therefore, many improved CDMA systems were proposed. An improved CDMA proposed in [12] partitioned and interleaved the spreading sequences and reduced the MAI by using a soft-iterative multistage receiver for additive white Gaussian noise (AWGN) channel. Considering a multipath frequency-selective channel, an MAI-free CDMA transceiver was first proposed in [13], in which chip-interleaved and zero padding were introduced to keep orthogonality among different users. Owing to the MAI dependent on the correlations among user signals, Lin and Lin [14] separated users by using different users' time-varying channel responses for a chip-interleaved direct-sequence CDMA (DS-CDMA) system since the correlations among different time-varying channel responses were extremely small. As channel estimation was an important issue for chip-interleaved CDMA, Na *et al.* [15] presented a pilot-aided fading-resistant transmission structure and discussed the tradeoff between channel estimation and the number of diversity branches by numerical analysis. To improve further the bit-error-rate (BER) performance of the chip-interleaved DS-CDMA, a switched interleaved DS-CDMA with block-based linear MMSE receivers was proposed [16], which indicated that the best interleaving patterns to suppress MAI were determined by the selection functions of the received signal-to-interference-plus-noise ratio (SINR). However, there were two important aspects that needed to be considered [16]: 1) A feedback channel was necessary for the given scheme to exchange the information on interleaving patterns between the transmitter and the receiver; and 2) the reliable prediction of the CSI was essential. Hence, Cai *et al.* [16] proposed an improved MMSE transceiver using switched interleaving and chipwise precoding for multicarrier CDMA (MC-CDMA) [17], in which the BS and the users were equipped with a codebook of quantized CSI, and the BS determined the optimum interleaver selection function for each user. In addition to the aforementioned papers, there are also some interesting papers that improve the transmit or receive structures of chip-interleaved CDMA. For example, in [18], the authors proposed a block CDMA with a cyclic prefix, and an iterative block decision feedback equalization was used

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to improve the BER performance; whereas in [19], a framework was presented for block-spread CDMA with precoding and special spreading codes to maximize the channel capacity. In [20] and [21], 2-D block-spread CDMA was proposed, consisting of two-level spreading: chip-level spreading and block-level spreading. The chip-level spreading was used to achieve the frequency diversity gain by using frequency-domain equalization (FDE) over a frequency-selective fading channel, and the block-level spreading was used to remove the MAI. The multiple-input–multiple-output (MIMO) technique has been identified as an enabling technique for achieving high-quality broadband data transmissions with limited bandwidth [1]. Point-to-point MIMO is well understood, i.e., space–time transmit diversity (STTD), cyclic delay transmit diversity (CDTD), etc. [22]–[25]. However, the multiuser MIMO network has not yet been fully investigated. This paper will focus on the MAI issue for multiuser cooperative MIMO (co-MIMO), which is the motivation of this paper.

In this paper, it is assumed that two transmit/receive antennas are used at both source and relay terminals. We present a scheme, which combines the source user's two antennas and the relay user's two antennas to form a virtual-four-antenna space–time cyclic delay transmit diversity (STCDTD), exploring a 2-D block-spread CDMA (both single-carrier CDMA (SC-CDMA) and MC-CDMA are considered) relay using the decode-and-forward (DF) strategy to avoid MAI since 2-D block-spread CDMA [20], [26], [27] achieves good BER performance with low-complexity single-user detection. Another important contribution of this paper is that the theoretical BER of the proposed scheme is described, when the signals are transmitted through a multiple-access frequency-selective Rayleigh fading channel.

The remainder of this paper is organized as follows. Section II describes the uplink cooperative relay network model. In Section III, the proposed 2-D block-spread CDMA relay using virtual-four-antenna STCDTD is described. Section IV presents the theoretical BER of the proposed scheme. The achievable BER performances of the proposed relay system are evaluated by computer simulation in Section V. Finally, Section VI offers some concluding remarks.

II. Cooperative Relay Network

The uplink cooperative relay network mode is shown in Fig. 1, where user 1 transmits its information to the BS with cooperation from user 2, and other users are an interference to user 1.

The network protocol used in this paper includes two time slots, as shown in Fig. 2. During the first time slot, the source user broadcasts its signal with small transmit power. The relay user who is located nearby the source user is able to receive the transmission signal from the source user's signal during the first time slot.

The relative locations of users 1 and 2 and the BS are important to ensure the achievable performance in cooperative communications. If users 1 and 2 are close to each other and both are far away from the BS, the relaying is not effective. In contrast, if neither user is far away from the BS, it is

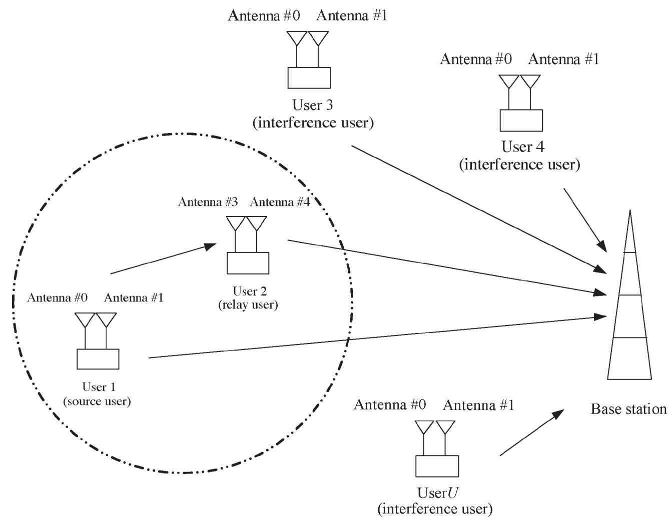

Fig. 1. Uplink cooperative relay network model.

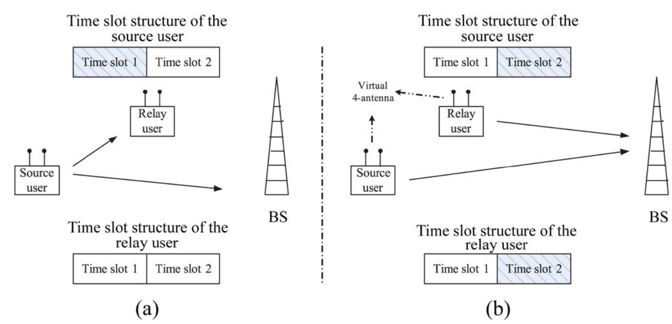

Fig. 2. Network protocol. The processing of the first time slot and the second time slot are shown in the left-hand and right-hand sides, respectively.

impossible to achieve a high delay diversity gain during the second time slot [36]. Due to page limitations, we assume that the user with a good channel state is selected as a relay user. The distance from the source node (user 1) to the BS and the distance from the source node (user 1) to the relay node (user 2) are denoted by $d_{S \to BS}$ and $d_{S \to R}$, respectively. The path-loss exponent is known to be between 3 and 4 in urban areas [28]. When $(d_{S \to BS}/d_{S \to R}) \geq 10$, the received power at the BS approximately becomes $10^{-4} \sim 10^{-3}$ times of the transmit power of the relay node. Therefore, the contribution of the user 1 signal received at the BS can be ignored.

Decoding errors may occur at the relay when DF is used. Hence, some error-control mechanism [29], [30] is necessary, i.e., a hybrid automatic repeat request (HARQ) can be applied to a link between the source user and the relay user. Retransmission is done until the data block is received without error at the relay node. HARQ is one of the most promising technologies to protect communication from a wireless channel more efficiently [31]; specifically, the HARQ is capable of providing the means for alleviating fading impairments and recovering the decoding errors by using time diversity at the expense of delay [32]. Since the HARQ utilizes additional time resources only when an error occurs at the destination, it is more efficient than basic time diversity schemes [33]. Recently, many works are concerned with how to minimize the transmission power and retransmission time delay for collaborative truncated HARQ

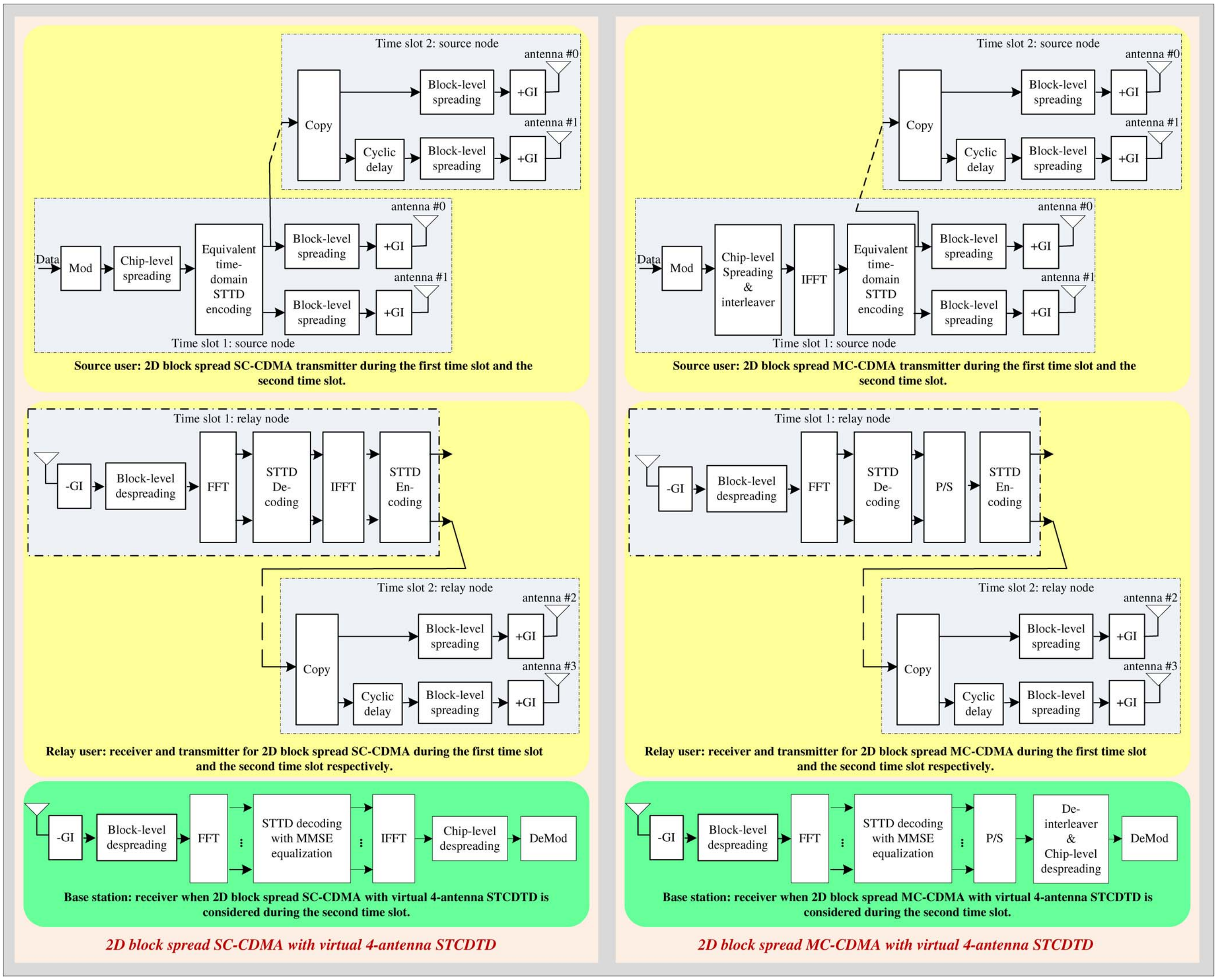

Fig. 3. Uplink 2-D block-spread CDMA using virtual-four-antenna STCDTD, where Mod, DeMod, +GI, and -GI represent modulation, demodulation, add GI, and remove GI, respectively.

scenarios [31], [34], [35]. This paper focuses on investigating the lowest achievable BER performance of multiuser co-MIMO. The investigation of realistic BER performances with HARQ in cooperative communications is left to our future study. During the second time slot, the source user sends its signal to the BS directly, and the relay user relays its received signal using the DF strategy to the BS simultaneously. The relay user's two antennas act as virtual antennas for the source user. At the BS, the two received signals are combined.

The STTD is applied at the first time slot for cooperative relaying. In the second time slot, virtual-four-antenna STCDTD is simultaneously explored at both the source user and the relay user to increase the diversity order.

III. 2-D Block-Spread Code-Division Multiple Access Using Virtual-Four-Antenna Space—Time Cyclic Delay Transmit Diversity

A. System Model

In this paper, the chip-spaced discrete-time signal representation is used. $\lfloor a \rfloor$ is the largest integer smaller than or equal to the real-valued variable a and $\lceil a \rceil$ is the smallest integer larger than or equal to a. Antenna #0 and antenna #1 are the source user's antennas, whereas antenna #2 and antenna #3 are the relay user's antennas. The 2-D block-spread CDMA consists of two-level spreading: chip-level spreading and block-level spreading. The transmitter/receiver structure is shown in Fig. 3.

B. First Time Slot

In the first time slot, the source user broadcasts its signal. The relay user receives and recovers the source user's transmission signal using the DF strategy.

Assume U users are simultaneously transmitting. The data symbol sequence to be transmitted from the uth ($u = 0 \sim U - 1$) user is denoted by $\{d_u(n); n = 0 \sim 2N_c/\mathrm{SF}_f - 1\}$, where SF_f is the chip-level spreading factor, and N_c is the size of fast Fourier transform (FFT) for FDE at the relay receiver. Suppose that the uth user is the source node of interest and the other $(U - 1)$ users create interference to the uth user. The data symbol sequence $\{d_u(n)\}$ is spread by chip-level spreading

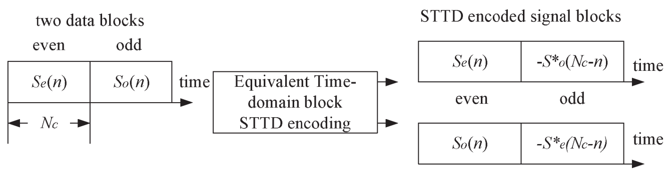

Fig. 4. Equivalent time-domain block STTD encoding.

code $\{c_u^{\mathrm{SF}_f}(t);\, t = 0 \sim \mathrm{SF}_f - 1\}$ with $|c_u^{\mathrm{SF}_f}(t)| = 1$, and is further multiplied by binary scramble sequence $\{c_u^{\mathrm{scr}}(t);\, t = 0 \sim N_c - 1\}$ to make the signal white-noise-like. The resulting SC-CDMA chip sequence $\{s_u^{\mathrm{SC}}(t);\, t = 0 \sim 2N_c - 1\}$ is given as

$$s_u^{\mathrm{SC}}(t) = c_u^{\mathrm{scr}}(t) \cdot d_u\left(\lfloor t/\mathrm{SF}_f \rfloor\right) c_u^{\mathrm{SF}_f}(t \bmod \mathrm{SF}_f). \tag{1}$$

In the case of MC-CDMA, N_c-point inverse FFT (IFFT) is applied to chip sequence $s_u^{\mathrm{SC}}(t)$ to generate MC-CDMA signal $s_u^{\mathrm{MC}}(t)$. To make full use of the channel frequency selectivity, block interleaving of size $\mathrm{SF}_f \times (N_c/\mathrm{SF}_f)$ is performed before the IFFT. Then, the data chips are distributed, with an equal distance of (N_c/SF_f) subcarriers, over N_c subcarriers, and $s_u^{\mathrm{MC}}(t)$ is expressed as

$$s_u^{\mathrm{MC}}(t) = \frac{1}{\sqrt{N_c}} \sum_{n=0}^{N_c/\mathrm{SF}_f - 1} \times \sum_{i=0}^{\mathrm{SF}_f - 1} \left[s_u^{\mathrm{SC}}(n \cdot \mathrm{SF}_f + i) \exp\left\{ j2\pi \frac{t}{N_c} \cdot \left(n + i\frac{N_c}{\mathrm{SF}_f} \right) \right\} \right]. \tag{2}$$

The CDMA signal $\{s_u(t);\, t{=}0{\sim}N_c{-}1\}$ $(s_u(t){=}s_u^{\mathrm{SC}}(t)$ for SC and $s_u(t) = s_u^{\mathrm{MC}}(t)$ for MC) is then divided into two consecutive N_c-chip blocks, even-chip block $\{s_{u,e}(t);\, t = 0 \sim N_c - 1\}$, and odd-chip block $\{s_{u,o}(t);\, t{=}0{\sim}N_c{-}1\}$. $s_{u,e}(t)$ and $s_{u,o}(t)$ are given, respectively, as

$$\begin{cases} s_{u,e}(t) = s_u(t) \\ s_{u,o}(t) = s_u(t + N_c). \end{cases} \tag{3}$$

The equivalent time-domain block STTD encoding for two antennas is shown in Fig. 4. The encoding rule of frequency-domain implementation of Alamouti's two-antenna STTD [22] is shown in Table I. (Its equivalent time-domain STTD encoding is given in Table II [23], [24].) The frequency-domain encoding is carried out as

$$\begin{pmatrix} s_{u,0,0}(t) & s_{u,1,0}(t) \\ s_{u,0,2}(t) & s_{u,1,2}(t) \end{pmatrix} = \frac{1}{2} \begin{pmatrix} s_{u,e}(t) & -s^*_{u,o}\left((N_c - t) \bmod N_c\right) \\ s_{u,o}(t) & s^*_{u,e}\left((N_c - t) \bmod N_c\right) \end{pmatrix} \tag{4}$$

where $\{s_{u,n_t,q}(t);\, t = 0 \sim N_c - 1\}$ is the coded chip block of the qth time block transmitted from the n_tth ($n_t = 0, 1$) antenna. The factor of 1/2 is introduced to keep the total transmit power from four virtual antennas to be the same as the single-transmit-antenna case (no diversity). Finally, block-level spreading is done on the time-domain STTD-encoded N_c-chip block using block-level spreading code $\{c_u^{\mathrm{SF}_t}(t);\, t = 0 \sim \mathrm{SF}_t - 1\}$ with spreading factor SF_t, which is used to avoid

TABLE I
FREQUENCY-DOMAIN STTD ENCODING

Time (in chip-block)	Antenna #0	Antenna #1
Even	$\frac{1}{2}s_{u,e}(k)$	$\frac{1}{2}s_{u,o}(k)$
Odd	$-\frac{1}{2}s^*_{u,e}(k)$	$\frac{1}{2}s^*_{u,o}(k)$

TABLE II
EQUIVALENT TIME-DOMAIN STTD ENCODING

Time (in chip-block)	Antenna #0	Antenna #1
Even	$\frac{1}{2}s_{u,e}(t)$	$\frac{1}{2}s_{u,o}(t)$
Odd	$-\frac{1}{2}s^*_{u,e}((N_c$ $-t) mod\ N_c)$	$\frac{1}{2}s^*_{u,o}((N_c$ $-t) mod\ N_c)$

the MAI. In the block-level spreading, each N_c-chip block is copied SF_t times, and each copy is multiplied by a chip from the orthogonal block-level spreading code $\{c_u^{\mathrm{SF}_t}(t)\}$ as

$$\hat{s}_{u,n_t,q}(t) = \sqrt{\frac{2E_c}{T_c}} s_{u,n_t,q}(t)(t \bmod N_c) c_u^{\mathrm{SF}_t}\left(\lfloor t/N_c \rfloor\right) \tag{5}$$

for $t = 0 \sim N_c \cdot \mathrm{SF}_t - 1$, where E_c is the average chip energy, and T_c is the chip duration. During the first time slot, antennas #0 and #1 of the source user transmit $\hat{s}_{u,0,0}(t)$ and $\hat{s}_{u,1,0}(t)$, respectively.

Before transmission, an N_g-chip guard interval (GI) is inserted at every N_c-chip block to avoid the interblock interference. The GI-inserted signal is transmitted to the relay user through the frequency-selective fading channel.

Assuming that the channel is composed of L chip-spaced independent paths, its impulse response $h_{u,n_t}(\tau)$ is expressed as

$$h_{u,n_t}(\tau) = \sum_{l=0}^{L-1} h_{u,n_t,l}\,\delta(\tau - \tau_{u,n_t,l}) \tag{6}$$

where $h_{u,n_t,l}$ and $\tau_{u,n_t,l}$ are, respectively, the complex-valued path gain and time delay of the lth path for the uth user transmitted through the n_tth ($n_t = 0$,1) antenna. Assume that the channel gain $h_{u,n_t,l}$ stays unchanged during the two block intervals. $\tau_{u,n_t,l}$ is assumed to be $\tau_{u,n_t,l} = \tau_{u,n_t} + l \cdot T_c$, $l = 0 \sim L - 1$, and τ_{u,n_t} is the uth user's transmit timing offset through the n_tth antenna. The maximum time delay of $\{\tau_{u,n_t,l}\}$ is assumed to be shorter than the GI. During the first time slot, the relay node receives the transmitted signal from the source node. The received signal is given as

$$\begin{cases} r_e(t) = \sum\limits_{u=0}^{U-1} \left(\sum\limits_{l=0}^{L-1} h_{u,0,l}\hat{s}_{u,0,e}(t - \tau_{u,0,l} + \sum\limits_{l=0}^{L-1} h_{u,1,l}\hat{s}_{u,1,e}(t - \tau_{u,1,l}) \right) + n_{u,e}(t) \\ r_o(t) = \sum\limits_{u=0}^{U-1} \left(\sum\limits_{l=0}^{L-1} h_{u,0,l}\hat{s}_{u,0,o}(t - \tau_{u,0,l}) + \sum\limits_{l=0}^{L-1} h_{u,1,l}\hat{s}_{u,1,o}(t - \tau_{u,1,l}) \right) + n_{u,o}(t) \end{cases} \tag{7}$$

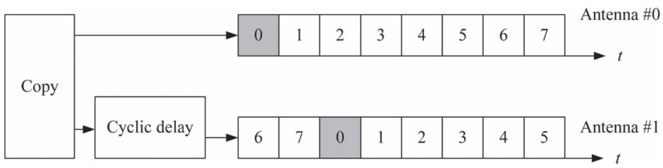

Fig. 5. Cyclic time delay addition ($N_c = 8$ and $\Delta = 2$ as an example).

for $t = 0 \sim N_c \cdot \mathrm{SF}_t - 1$, where $n_{u,e}(t)$ and $n_{u,o}(t)$ are the zero-mean complex-valued AWGN samples with variance $2N_0/T_c$ (N_0 is the AWGN one-sided power spectrum density). The relay node operates block-level despreading and STTD decoding with MMSE-FDE to recover the transmitted signals $\{s_{u,n_t,q}(t);\, t = 0 \sim N_c - 1\}$. (For more details, please check [26].)

C. Second Time Slot

In the second time slot, the source user sends its signal, whereas the relay user recovers and relays its received signal to the BS simultaneously. The relay user's two antennas act as virtual antennas for the source user. The four antennas (two antennas provided by the source user; another two antennas are supported by the relay user) form a virtual-four-antenna STCDTD to serve the source user. To generate an STCDTD signal, STTD encoding is first carried out as (4). The chip blocks are then copied into two streams, respectively, and a cyclic delay is added to form a virtual-four-antenna STCDTD during the second time slot; the cyclic delay is added as shown in Fig. 5, where Δ is the cyclic time delay factor. As the relay user retransmits the signals from the source user, the same block-level spreading code $\{c_u^{\mathrm{SF}_t}(t);\, t = 0 \sim \mathrm{SF}_t - 1\}$ as the source user must be used. After the block-level spreading, the even N_c-chip blocks $\hat{s}_{u,n_t,e}(t)$ and odd N_c-chip blocks $\hat{s}_{u,n_t,o}(t)$ to be transmitted from the n_tth transmit antenna ($n_t = 0$, 1 for source user; $n_t = 2$, 3 for relay user) are given, respectively, as

$$\begin{cases} \hat{s}_{u,0,e}(t) = \sqrt{\frac{2E_c}{T_c}} s_{u,0,0}(t \bmod N_c) c_u^{\mathrm{SF}_t}\left(\lfloor t/N_c \rfloor\right) \\ \hat{s}_{u,1,e}(t) = \sqrt{\frac{2E_c}{T_c}} s_{u,0,0}\left((t-\Delta) \bmod N_c\right) c_u^{\mathrm{SF}_t}\left(\lfloor t/N_c \rfloor\right) \\ \hat{s}_{u,2,e}(t) = \sqrt{\frac{2E_c}{T_c}} s_{u,0,1}(t \bmod N_c) c_u^{\mathrm{SF}_t}\left(\lfloor t/N_c \rfloor\right) \\ \hat{s}_{u,3,e}(t) = \sqrt{\frac{2E_c}{T_c}} s_{u,0,1}\left((t-\Delta) \bmod N_c\right) c_u^{\mathrm{SF}_t}\left(\lfloor t/N_c \rfloor\right) \end{cases} \tag{8}$$

$$\begin{cases} \hat{s}_{u,0,o}(t) = \sqrt{\frac{2E_c}{T_c}} s_{u,0,0}(t \bmod N_c) c_u^{\mathrm{SF}_t}\left(\lfloor t/N_c \rfloor\right) \\ \hat{s}_{u,1,o}(t) = \sqrt{\frac{2E_c}{T_c}} s_{u,0,0}\left((t-\Delta) \bmod N_c\right) c_u^{\mathrm{SF}_t}\left(\lfloor t/N_c \rfloor\right) \\ \hat{s}_{u,2,o}(t) = \sqrt{\frac{2E_c}{T_c}} s_{u,0,1}(t \bmod N_c) c_u^{\mathrm{SF}_t}\left(\lfloor t/N_c \rfloor\right) \\ \hat{s}_{u,3,o}(t) = \sqrt{\frac{2E_c}{T_c}} s_{u,0,1}\left((t-\Delta) \bmod N_c\right) c_u^{\mathrm{SF}_t}\left(\lfloor t/N_c \rfloor\right) \end{cases} \tag{9}$$

for $t = 0 \sim \mathrm{SF}_t \times N_c - 1$. Before transmission, an N_g-chip GI is also inserted every N_c-chip block. Then, the GI-inserted STCDTD-encoded block is transmitted to the BS through the multiple-access frequency-selective fading channel.

The sum of U users' faded signals is received at the BS since we assume that the transmission power of the source user during the first time slot is small enough and that the relay node is nearby the source user. Thereby, the BS only receives the signals during the second time slot. The GI-removed received signals at the even and odd time intervals can be expressed as (10), shown at the bottom of the page, for $t = 0 \sim \mathrm{SF}_t \times N_c - 1$. First, the block-level despreading is applied as

$$\begin{cases} \hat{r}_{u,e}(t) = \frac{1}{\mathrm{SF}_t} \sum\limits_{i=0}^{\mathrm{SF}_t - 1} r_e(t + iN_c) \left\{c_u^{\mathrm{SF}_t}\right\}^* \\ \hat{r}_{u,o}(t) = \frac{1}{\mathrm{SF}_t} \sum\limits_{i=0}^{\mathrm{SF}_t - 1} r_o(t + iN_c) \left\{c_u^{\mathrm{SF}_t}\right\}^* \end{cases} \tag{11}$$

for $t = 0 \sim N_c - 1$. Note that, if it is slow fading, the path gains stay almost constant over at least SF_t consecutive blocks; thus, the MAI can be perfectly removed by the block-level despreading. The even and odd received chip blocks of the uth user, i.e., $\{\hat{r}_{u,e}(t);\, t = 0 \sim N_c - 1\}$ and $\{\hat{r}_{u,o}(t);\, t = 0 \sim N_c - 1\}$, are decomposed by an N_c-point FFT into frequency domain as $\{\hat{R}_{u,e}(t);\, t = 0 \sim N_c - 1\}$ and $\{\hat{R}_{u,o}(t);\, t = 0 \sim N_c - 1\}$, respectively, given as (12) at the bottom of the next page, where $H_{u,n_t}(k)$ represents the N_c-point Fourier transform of the channel gain between the receive antenna and the n_tth transmit antenna ($n_t = 0$, 1, 2, 3) and $\Pi_{u,e(o)}(k)$ represents the zero-mean noise component having the variance $2N_0/T_c$ at the kth frequency in the received even (or odd) blocks.

The frequency-domain STTD decoding is carried out as [23], [24]

$$\begin{cases} \tilde{S}_{u,e}(k) = W_{u,0}^* \hat{R}_{u,e}(k) + W_{u,2}(k) \hat{R}_{u,o}^*(k) \\ \tilde{S}_{u,o}(k) = W_{u,2}^* \hat{R}_{u,e}(k) - W_{u,0}(k) \hat{R}_{u,o}^*(k) \end{cases} \tag{13}$$

$$\begin{cases} r_e(t) = \sum\limits_{u=0}^{U-1} \left(\sum\limits_{l=0}^{L-1} h_{u,0,l} \hat{s}_{u,0,e}(t - \tau_{u,0,l}) + \sum\limits_{l=0}^{L-1} h_{u,1,l} \hat{s}_{u,1,e}(t - \tau_{u,1,l}) \right. \\ \qquad\qquad \left. + \sum\limits_{l=0}^{L-1} h_{u,2,l} \hat{s}_{u,2,e}(t - \tau_{u,2,l}) + \sum\limits_{l=0}^{L-1} h_{u,3,l} \hat{s}_{u,3,e}(t - \tau_{u,3,l}) \right) + n_{u,e}(t) \\ r_o(t) = \sum\limits_{u=0}^{U-1} \left(\sum\limits_{l=0}^{L-1} h_{u,0,l} \hat{s}_{u,0,o}(t - \tau_{u,0,l}) + \sum\limits_{l=0}^{L-1} h_{u,1,l} \hat{s}_{u,1,o}(t - \tau_{u,1,l}) \right. \\ \qquad\qquad \left. + \sum\limits_{l=0}^{L-1} h_{u,2,l} \hat{s}_{u,2,o}(t - \tau_{u,2,l}) + \sum\limits_{l=0}^{L-1} h_{u,3,l} \hat{s}_{u,3,o}(t - \tau_{u,3,l}) \right) + n_{u,o}(t) \end{cases} \tag{10}$$

where $W_{u,0}(k)$and $W_{u,2}(k)$ are MMSE weights written as

$$\begin{cases} W_{u,0}(k) = \frac{\bar{H}_{u,0}(k)}{|H_{u,0}(k)|^2+|H_{u,2}(k)|^2+\left(\frac{1}{4}\cdot\mathrm{SF}_t\cdot\frac{E_c}{N_0}\right)^{-1}} \\ W_{u,2}(k) = \frac{\bar{H}_{u,2}(k)}{|H_{u,0}(k)|^2+|H_{u,2}(k)|^2+\left(\frac{1}{4}\cdot\mathrm{SF}_t\cdot\frac{E_c}{N_0}\right)^{-1}} \end{cases} \tag{14}$$

with

$$\begin{cases} \bar{H}_{u,0}(k) = H_{u,0}(k) + H_{u,1}(k)\cdot\exp\left(-j2\pi k\frac{\Delta}{N_c}\right) \\ \bar{H}_{u,2}(k) = H_{u,2}(k) + H_{u,3}(k)\cdot\exp\left(-j2\pi k\frac{\Delta}{N_c}\right) \end{cases}. \tag{15}$$

For SC-CDMA, N_c-point IFFT is applied to $\{\tilde{S}_{u,e}(k);\ k = 0 \sim N_c-1\}$ and $\{\tilde{S}_{u,o}(k);\ k=0\sim N_c-1\}$, and then the soft-decision time-domain chip blocks are obtained as

$$\begin{cases} \tilde{s}^{\mathrm{SC}}_{u,e}(t) = \frac{1}{\sqrt{N_c}}\sum_{k=0}^{N_c-1}\tilde{S}_{u,e}(k)\cdot\exp\left(j2\pi t\frac{k}{N_c}\right) \\ \tilde{s}^{\mathrm{SC}}_{u,o}(t) = \frac{1}{\sqrt{N_c}}\sum_{k=0}^{N_c-1}\tilde{S}_{u,o}(k)\cdot\exp\left(j2\pi t\frac{k}{N_c}\right) \end{cases} \tag{16}$$

for $t=0\sim N_c-1$. For MC-CDMA, the soft-decision transmitted chip block is directly obtained from the frequency-domain deinterleaver as

$$\begin{cases} \tilde{s}^{\mathrm{MC}}_{u,e}(t) = \tilde{S}_{u,e}\left((t \bmod \mathrm{SF}_f)\cdot(N_c/\mathrm{SF}_f)+\lfloor t/\mathrm{SF}_f\rfloor\right) \\ \tilde{s}^{\mathrm{MC}}_{u,o}(t) = \tilde{S}_{u,o}\left((t \bmod \mathrm{SF}_f)\cdot(N_c/\mathrm{SF}_f)+\lfloor t/\mathrm{SF}_f\rfloor\right) \end{cases}. \tag{17}$$

Finally, chip-level despreading is carried out on $\{\tilde{s}_{u,e(o)}(t);\ t=0\sim N_c-1\}$ ($\tilde{s}_{u,e(o)}(t) = \tilde{s}^{\mathrm{SC}}_u(t)$ for SC, and $\tilde{s}_{u,e(o)}(t) = \tilde{s}^{\mathrm{MC}}_u(t)$ for MC), thereby

$$\hat{d}_u(n) = \frac{1}{\mathrm{SF}_f}\sum_{t=n\cdot\mathrm{SF}_f}^{(n+1)\cdot\mathrm{SF}_f-1}\tilde{s}_{u,e(o)}(t)\left[c_u^{\mathrm{SF}_f}(t)\cdot c_u^{\mathrm{scr}}(t)\right]^* \tag{18}$$

which is the decision variable for data demodulation, associated with $d_u(n)$, $n = 0\sim 2N_c/\mathrm{SF}_f - 1$.

IV. Theoretical Analysis

For theoretical analysis, we assume that the fading channels of different antennas and different users are independent to each other.

A. Decision Variable

To achieve the theoretical BER of the proposed system, it is important to analyze the received signal exhaustively. It should be noted that, since virtual-four-antenna STCDTD diversity is considered, the following discussion is based on both even and odd time intervals, which is different from traditional single-antenna case.

First, substituting (10) into (11), the GI-removed received signal is expressed as (19), shown at the bottom of the page, for $t=0\sim N_c-1$. For convenience, we combine multiantenna signals together, and (19) is rewritten as (20), shown at the bottom of the next page. Since block fading is assumed, the received signal is derived as (21), which is also shown at the bottom of the next page.

The first ($\mu_{u,e}$and$\mu_{u,o}$), second ($\mu_{\mathrm{MAI},e}$ and $\mu_{\mathrm{MAI},o}$), and third ($\mu_{\mathrm{noise},e}$ and $\mu_{\mathrm{noise},o}$) terms in (21) are the desired signal; the MAI from other users and noise, respectively, as in (22) and (23), shown at the bottom of the next page; and

$$\begin{cases} \mu_{\mathrm{noise},e} = \frac{1}{\mathrm{SF}_t}\sum_{i=0}^{\mathrm{SF}_t-1} n_{u,e}(t+iN_c)\left\{c_u^{\mathrm{SF}_t}\right\}^* \\ \mu_{\mathrm{noise},o} = \frac{1}{\mathrm{SF}_t}\sum_{i=0}^{\mathrm{SF}_t-1} n_{u,o}(t+iN_c)\left\{c_u^{\mathrm{SF}_t}\right\}^* \end{cases}. \tag{24}$$

What should be noted is that (21) reduces to (11) when the fading is slow enough. In this case, since the path gains stay constant over at least consecutive blocks, the MAI can be perfectly removed by the block-level despreading. It should be noted that (21) is developed for the theoretical analysis of BER; (11) is directly used for block-level dispreading at the receiver.

$$\begin{cases} \hat{R}_{u,e}(t) = \sqrt{\frac{2E_c}{T_c}}\left(H_{u,0}(k) + H_{u,1}(k)\ \exp\left(-j2\pi k\frac{\Delta}{N_c}\right)\right)S_{u,0,0}(k) \\ \qquad + \sqrt{\frac{2E_c}{T_c}}\left(H_{u,2}(k) + H_{u,3}(k)\ \exp\left(-j2\pi k\frac{\Delta}{N_c}\right)\right)S_{u,0,2}(k) + \Pi_{u,e}(k) \\ \hat{R}_{u,o}(t) = \sqrt{\frac{2E_c}{T_c}}\left(H_{u,0}(k) + H_{u,1}(k)\cdot\ \exp\left(-j2\pi k\frac{\Delta}{N_c}\right)\right)S_{u,1,0}(k) \\ \qquad + \sqrt{\frac{2E_c}{T_c}}\left(H_{u,2}(k) + H_{u,3}(k)\ \exp\left(-j2\pi k\frac{\Delta}{N_c}\right)\right)S_{u,1,2}(k) + \Pi_{u,o}(k) \end{cases} \tag{12}$$

$$\begin{cases} \hat{r}_{u,e}(t) = \frac{1}{\mathrm{SF}_t}\sum_{i=0}^{\mathrm{SF}_t-1}\left[\sum_{u=0}^{U-1}\left(\sum_{l=0}^{L-1}h_{u,0,l}\hat{s}_{u,0,e}(t+iN_c-\tau_{u,0,l}) + \sum_{l=0}^{L-1}h_{u,1,l}\hat{s}_{u,1,e}(t+iN_c-\tau_{u,1,l})\right.\right. \\ \qquad \left.\left. +\sum_{l=0}^{L-1}h_{u,2,l}\hat{s}_{u,0,e}(t+iN_c-\tau_{u,2,l}) + \sum_{l=0}^{L-1}h_{u,3,l}\hat{s}_{u,1,e}(t+iN_c-\tau_{u,3,l})\right) + n_{u,e}(t+iN_c)\right]\left\{c_u^{\mathrm{SF}_t}\right\}^* \\ \hat{r}_{u,o}(t) = \frac{1}{\mathrm{SF}_t}\sum_{i=0}^{\mathrm{SF}_t-1}\left[\sum_{u=0}^{U-1}\left(\sum_{l=0}^{L-1}h_{u,0,l}\hat{s}_{u,0,o}(t+iN_c-\tau_{u,0,l}) + \sum_{l=0}^{L-1}h_{u,1,l}\hat{s}_{u,1,o}(t+iN_c-\tau_{u,1,l})\right.\right. \\ \qquad \left.\left. +\sum_{l=0}^{L-1}h_{u,2,l}\hat{s}_{u,0,o}(t+iN_c-\tau_{u,2,l}) + \sum_{l=0}^{L-1}h_{u,3,l}\hat{s}_{u,1,o}(t+iN_c-\tau_{u,3,l})\right) + n_{u,o}(t+iN_c)\right]\left\{c_u^{\mathrm{SF}_t}\right\}^* \end{cases} \tag{19}$$

According to (21), the frequency-domain representation of the received signal for the uth user can be written as

$$\begin{cases} \hat{R}_{u,e}(t) = \sqrt{\frac{2E_c}{T_c}} \sum\limits_{n_t=0}^{3} \left[H_{u,n_t}(k) \cdot S_{u,n_t,e}(k)\right] \\ \qquad + \sqrt{\frac{2E_c}{T_c}} \sum\limits_{\substack{u'=0 \\ u' \neq u}}^{U-1} \sum\limits_{n_t=0}^{3} \left[Z_{u',n_t,e}(k) \cdot S_{u',n_t,e}(k)\right] \\ \qquad + \Pi_{u,e}(k) \\ \hat{R}_{u,o}(t) = \sqrt{\frac{2E_c}{T_c}} \sum\limits_{n_t=0}^{3} \left[H_{u,n_t}(k) \cdot S_{u,n_t,o}(k)\right] \\ \qquad + \sqrt{\frac{2E_c}{T_c}} \sum\limits_{\substack{u'=0 \\ u' \neq u}}^{U-1} \sum\limits_{n_t=0}^{3} \left[Z_{u',n_t,o}(k) \cdot S_{u',n_t,e}(k)\right] \\ \qquad + \Pi_{u,o}(k) \end{cases} \tag{25}$$

for $k = 0 \sim N_c - 1$, where u' denotes an interference user, and the first, second, and third terms represent the desired signal, MAI, and AWGN components, respectively, with

$$\begin{cases} S_{u,n_t,e}(k) = \frac{1}{\sqrt{N_c}} \sum\limits_{t=0}^{N_c-1} S_{u,n_t,e}(t) \exp\left(-j2\pi k \frac{t}{N_c}\right) \\ H_{u,n_t}(k) = \frac{1}{\mathrm{SF}_t} \sum\limits_{i=0}^{\mathrm{SF}_t} \sum\limits_{l=0}^{L-1} h_{u,\,n_t,\,l} \exp\left(-j2\pi k \frac{\tau_{u,\,n_t,\,l}}{N_c}\right) \\ Z_{u',n_t}(k) = \frac{1}{\mathrm{SF}_t} \sum\limits_{i=0}^{\mathrm{SF}_t} \left[c_{u'}^{\mathrm{SF}_t} \left\{c_u^{\mathrm{SF}_t}\right\}^* \right. \\ \qquad\qquad \left. \times \sum\limits_{l=0}^{L-1} h_{u',n_t,l} \exp\left(-j2\pi k \frac{\tau_{u',n_t,l}}{N_c}\right)\right] \\ \Pi_u(k) = \frac{1}{\sqrt{N_c}} \sum\limits_{t=0}^{N_c-1} \left[\frac{1}{\mathrm{SF}_t} \cdot \sum\limits_{i=0}^{\mathrm{SF}_t-1} n_{u,\,e}(t+iN_c) \left\{c_u^{\mathrm{SF}_t}\right\}^*\right] \\ \qquad \times \exp\left(-j2\pi k \frac{t}{N_c}\right) \end{cases} \tag{26}$$

where $Z_{b'\_u'}(k)$ is the frequency-domain MAI after block-level despreading.

$$\begin{cases} \hat{r}_{u,e}(t) = \frac{1}{\mathrm{SF}_t} \cdot \sqrt{\frac{2E_c}{T_c}} \sum\limits_{i=0}^{\mathrm{SF}_t-1} \times \left[\sum\limits_{u=0}^{U-1} \sum\limits_{n_t=0}^{3} \left(\sum\limits_{l=0}^{L-1} h_{u,n_t,l} s_{u,n_t,e}(t+iN_c-\tau_{u,n_t,l}) c_u^{\mathrm{SF}_t}\left(\lfloor t+iN_c-\tau_{u,n_t,l}/N_c\rfloor\right)\right) + n_{u,e}(t+iN_c)\right] \left\{c_u^{\mathrm{SF}_t}\right\}^* \\ \hat{r}_{u,o}(t) = \frac{1}{\mathrm{SF}_t} \cdot \sqrt{\frac{2E_c}{T_c}} \sum\limits_{i=0}^{\mathrm{SF}_t-1} \times \left[\sum\limits_{u=0}^{U-1} \sum\limits_{n_t=0}^{3} \left(\sum\limits_{l=0}^{L-1} h_{u,n_t,l} s_{u,n_t,o}(t+iN_c-\tau_{u,n_t,l}) c_u^{\mathrm{SF}_t}\left(\lfloor t+iN_c-\tau_{u,n_t,l}/N_c\rfloor\right)\right) + n_{u,o}(t+iN_c)\right] \left\{c_u^{\mathrm{SF}_t}\right\}^* \end{cases} \tag{20}$$

$$\begin{cases} \hat{r}_{u,e}(t) = \frac{1}{\mathrm{SF}_t} \cdot \sqrt{\frac{2E_c}{T_c}} \sum\limits_{i=0}^{\mathrm{SF}_t-1} \sum\limits_{n_t=0}^{3} \left(\sum\limits_{l=0}^{L-1} h_{u,n_t,l} s_{u,n_t,e}(t+iN_c-\tau_{u,n_t,l}) c_u^{\mathrm{SF}_t}\left(\lfloor t+iN_c-\tau_{u,n_t,l}/N_c\rfloor\right)\right) \left\{c_u^{\mathrm{SF}_t}\right\}^* \\ \qquad + \frac{1}{\mathrm{SF}_t} \cdot \sqrt{\frac{2E_c}{T_c}} \sum\limits_{i=0}^{\mathrm{SF}_t-1} \left[\sum\limits_{\substack{u'=0 \\ u' \neq u}}^{U-1} \sum\limits_{n_t=0}^{3} \left(\sum\limits_{l=0}^{L-1} h_{u',n_t,l} s_{u',n_t,e}(t+iN_c-\tau_{u,n_t,l}) c_{u'}^{\mathrm{SF}_t}\left(\lfloor t+iN_c-\tau_{u',n_t,l}/N_c\rfloor\right)\right)\right] \left\{c_u^{\mathrm{SF}_t}\right\}^* + \frac{1}{\mathrm{SF}_t} \sum\limits_{i=0}^{\mathrm{SF}_t-1} n_{u,e}(t+iN_c) \left\{c_u^{\mathrm{SF}_t}\right\}^* \\ \quad = \mu_{u,e} + \mu_{\mathrm{MAI},\,e} + \mu_{\mathrm{noise},\,e} \\ \hat{r}_{u,o}(t) = \frac{1}{\mathrm{SF}_t} \cdot \sqrt{\frac{2E_c}{T_c}} \sum\limits_{i=0}^{\mathrm{SF}_t-1} \sum\limits_{n_t=0}^{3} \left(\sum\limits_{l=0}^{L-1} h_{u,n_t,l} s_{u,n_t,o}(t+iN_c-\tau_{u,n_t,l}) c_u^{\mathrm{SF}_t}\left(\lfloor t+iN_c-\tau_{u,n_t,l}/N_c\rfloor\right)\right) \left\{c_u^{\mathrm{SF}_t}\right\}^* \\ \qquad + \frac{1}{\mathrm{SF}_t} \cdot \sqrt{\frac{2E_c}{T_c}} \sum\limits_{i=0}^{\mathrm{SF}_t-1} \left[\sum\limits_{\substack{u'=0 \\ u' \neq u}}^{U-1} \sum\limits_{n_t=0}^{3} \left(\sum\limits_{l=0}^{L-1} h_{u',n_t,l} s_{u',n_t,o}(t+iN_c-\tau_{u,n_t,l}) c_{u'}^{\mathrm{SF}_t}\left(\lfloor t+iN_c-\tau_{u',n_t,l}/N_c\rfloor\right)\right)\right] \left\{c_u^{\mathrm{SF}_t}\right\}^* + \frac{1}{\mathrm{SF}_t} \sum\limits_{i=0}^{\mathrm{SF}_t-1} n_{u,o}(t+iN_c) \left\{c_u^{\mathrm{SF}_t}\right\}^* \\ \quad = \mu_{u,o} + \mu_{\mathrm{MAI},\,e} + \mu_{\mathrm{noise},\,o}. \end{cases} \tag{21}$$

$$\begin{cases} \mu_{u,e} = \frac{1}{\mathrm{SF}_t} \cdot \sqrt{\frac{2E_c}{T_c}} \sum\limits_{i=0}^{\mathrm{SF}_t-1} \sum\limits_{n_t=0}^{3} \left(\sum\limits_{l=0}^{L-1} h_{u,n_t,l} s_{u,n_t,e}(t+iN_c-\tau_{u,n_t,l}) c_u^{\mathrm{SF}_t}\left(\lfloor t+iN_c-\tau_{u,n_t,l}/N_c\rfloor\right)\right) \left\{c_u^{\mathrm{SF}_t}\right\}^* \\ \mu_{u,o} = \frac{1}{\mathrm{SF}_t} \cdot \sqrt{\frac{2E_c}{T_c}} \sum\limits_{i=0}^{\mathrm{SF}_t-1} \sum\limits_{n_t=0}^{3} \left(\sum\limits_{l=0}^{L-1} h_{u,n_t,l} s_{u,n_t,o}(t+iN_c-\tau_{u,n_t,l}) c_u^{\mathrm{SF}_t}\left(\lfloor t+iN_c-\tau_{u,n_t,l}/N_c\rfloor\right)\right) \left\{c_u^{\mathrm{SF}_t}\right\}^* \end{cases} \tag{22}$$

$$\begin{cases} \mu_{\mathrm{MAI},\,e} = \frac{1}{\mathrm{SF}_t} \cdot \sqrt{\frac{2E_c}{T_c}} \sum\limits_{i=0}^{\mathrm{SF}_t-1} \left[\sum\limits_{\substack{u'=0 \\ u' \neq u}}^{U-1} \sum\limits_{n_t=0}^{3} \left(\sum\limits_{l=0}^{L-1} h_{u',n_t,l} s_{u',n_t,e}(t+iN_c-\tau_{u,n_t,l}) c_{u'}^{\mathrm{SF}_t}\left(\lfloor t+iN_c-\tau_{u',n_t,l}/N_c\rfloor\right)\right)\right] \left\{c_u^{\mathrm{SF}_t}\right\}^* \\ \mu_{\mathrm{MAI},o} = \frac{1}{\mathrm{SF}_t} \cdot \sqrt{\frac{2E_c}{T_c}} \sum\limits_{i=0}^{\mathrm{SF}_t-1} \left[\sum\limits_{\substack{u'=0 \\ u' \neq u}}^{U-1} \sum\limits_{n_t=0}^{3} \left(\sum\limits_{l=0}^{L-1} h_{u',n_t,l} s_{u',n_t,o}(t+iN_c-\tau_{u,n_t,l}) c_{u'}^{\mathrm{SF}_t}\left(\lfloor t+iN_c-\tau_{u',n_t,l}/N_c\rfloor\right)\right)\right] \left\{c_u^{\mathrm{SF}_t}\right\}^* \end{cases} \tag{23}$$

According to (13) and (14), FDE is applied to $\hat{R}_{u,e}(t)$ and $\hat{R}_{u,o}(t)$, followed by IFFT to get the time-domain signal shown as (16). The decision variable can be also expressed as

$$\begin{cases} \hat{d}_{u,e}(n)=\lambda_{u,e}(n)+\zeta_{\mathrm{MANI},e}(n)+\zeta_{\mathrm{MAI},e}(n)+\zeta_{\mathrm{noise},e}(n) \\ \hat{d}_{u,o}(n)=\lambda_{u,o}(n)+\zeta_{\mathrm{MANI},o}(n)+\zeta_{\mathrm{MAI},o}(n)+\zeta_{\mathrm{noise},o}(n) \end{cases} \tag{27}$$

where $\lambda_{u,e(o)}(n)$, $\zeta_{\mathrm{MANI},e(o)}(n)$, $\zeta_{\mathrm{MAI},e(o)}(n)$, and $\zeta_{\mathrm{noise},e(o)}(n)$ are the equivalent channel gain, multiple-antenna interference (MANI), MAI, and noise components, respectively.

Substituting (4), (5), and (12) into (13), $\lambda_{u,e(o)}(n)$ can be thus expressed as (28) and (29), shown at the bottom of page, for SC-CDMA and MC-CDMA, respectively.

Similarly, the MANI for SC-CDMA and MC-CDMA are given as (30) and (31), respectively, shown at the bottom of the page.

Equations (32) and (33) show the MAI for SC-CDMA and MC-CDMA, respectively, as shown at the bottom of the next page.

For noise, the signals are given as (34) and (35), which are shown at the bottom of the next page.

B. SINR

According to (27), the SINR of the uth user at odd time intervals and even time intervals are capable of deriving as

$$\begin{cases} \gamma_{u,e}(n) = \frac{|\lambda_{u,e}(n)|^2}{|\zeta_{\mathrm{MAI},e}(n)+\zeta_{\mathrm{MANI},e}(n)+\zeta_{\mathrm{noise},e}(n)|^2} \\ \gamma_{u,o}(n) = \frac{|\lambda_{u,o}(n)|^2}{|\zeta_{\mathrm{MAI},o}(n)+\zeta_{\mathrm{MANI},o}(n)+\zeta_{\mathrm{noise},o}(n)|^2}. \end{cases} \tag{36}$$

As all signal modulation points are equiprobable during odd and even time intervals, considering that binary phase-shift keying (PSK) is used, the conditional BER $P_{u,b}(\gamma_{u,o}(n), \gamma_{u,e}(n))$ is achieved as

$$\begin{aligned} &P_{u,b}\left(\gamma_{u,o}(n), \gamma_{u,e}(n)\right) \\ &= \frac{1}{N_c}\sum_{n=0}^{N_c-1}\left\{\frac{1}{2}Q\left(\sqrt{2\gamma_{u,o}(n)}\right)+\frac{1}{2}Q\left(\sqrt{2\gamma_{u,e}(n)}\right)\right\} \end{aligned} \tag{37}$$

where Q function is related to $Q(x) = 1/2\mathrm{erfc}(x/\sqrt{2})$, $\mathrm{erfc}(x)$ is the complementary error function.

$$\begin{cases} \lambda_{u,e}(n) = \frac{1}{\mathrm{SF}_f}\sum_{t=n\cdot\mathrm{SF}_f}^{(n+1)\mathrm{SF}_f-1}\left\{C_u^{\mathrm{SF}_f}(t)C_u^{\mathrm{scr}}(t)\right\}^* \frac{1}{\sqrt{N_c}}\sqrt{\frac{E_c}{T_c}} \\ \qquad \times \sum_{k=0}^{N_c-1}\left[W_{u,0}^*(k)\bar{H}_{u,0}(k)\cdot S_{u,e}(k) + W_{u,2}(k)\bar{H}_{u,2}^*\cdot S_{u,o}(k)\right]\cdot\exp\left(j\frac{2\pi tk}{N_c}\right) \\ \lambda_{u,o}(n) = \frac{1}{\mathrm{SF}_f}\sum_{t=n\cdot\mathrm{SF}_f}^{(n+1)\mathrm{SF}_f-1}\left\{C_u^{\mathrm{SF}_f}(t)C_u^{\mathrm{scr}}(t)\right\}^* \frac{1}{\sqrt{N_c}}\sqrt{\frac{E_c}{T_c}} \\ \qquad \times \sum_{k=0}^{N_c-1}\left[W_{u,2}^*(k)\bar{H}_{u,2}(k)\cdot S_{u,e}(k) + W_{u,0}(k)\bar{H}_{u,0}^*\cdot S_{u,o}(k)\right]\cdot\exp\left(j\frac{2\pi tk}{N_c}\right) \end{cases} \tag{28}$$

$$\begin{cases} \lambda_{u,e}(n) = \frac{1}{\mathrm{SF}_f}\sqrt{\frac{E_c}{T_c}}\sum_{i=0}^{\mathrm{SF}_f-1}\left[W_{u,0}^*\left(n+i\cdot\frac{N_c}{\mathrm{SF}_f}\right)\bar{H}_{u,0}\left(n+i\cdot\frac{N_c}{\mathrm{SF}_f}\right)\cdot S_{u,e}(k)\right. \\ \qquad\qquad \left.+W_{u,2}\left(n+i\cdot\frac{N_c}{\mathrm{SF}_f}\right)\bar{H}_{u,2}^*\left(n+i\cdot\frac{N_c}{\mathrm{SF}_f}\right)\cdot S_{u,o}(k)\right] \\ \lambda_{u,o}(n) = \frac{1}{\mathrm{SF}_f}\sqrt{\frac{E_c}{T_c}}\sum_{i=0}^{\mathrm{SF}_f-1}\left[W_{u,2}^*\left(n+i\cdot\frac{N_c}{\mathrm{SF}_f}\right)\bar{H}_{u,2}\left(n+i\cdot\frac{N_c}{\mathrm{SF}_f}\right)\cdot S_{u,e}(k)\right. \\ \qquad\qquad \left.+W_{u,0}\left(n+i\cdot\frac{N_c}{\mathrm{SF}_f}\right)\times\bar{H}_{u,0}^*\left(n+i\cdot\frac{N_c}{\mathrm{SF}_f}\right)\cdot S_{u,o}(k)\right] \end{cases} \tag{29}$$

$$\begin{cases} \zeta_{\mathrm{MANI},e}(n)=\frac{1}{\mathrm{SF}_f}\sum_{t=n\cdot\mathrm{SF}_f}^{(n+1)\mathrm{SF}_f-1}\left\{C_u^{\mathrm{SF}_f}(t)C_u^{\mathrm{scr}}(t)\right\}^*\times\frac{1}{\sqrt{N_c}}\sqrt{\frac{E_c}{T_c}}\sum_{k=0}^{N_c-1}\left[W_{u,0}^*(k)\bar{H}_{u,2}(k)\cdot S_{u,e}(k)-W_{u,2}(k)\bar{H}_{u,0}^*\cdot S_{u,o}(k)\right]\cdot\exp\left(j\frac{2\pi tk}{N_c}\right) \\ \zeta_{\mathrm{MANI},o}(n)=\frac{1}{\mathrm{SF}_f}\sum_{t=n\cdot\mathrm{SF}_f}^{(n+1)\mathrm{SF}_f-1}\left\{C_u^{\mathrm{SF}_f}(t)C_u^{\mathrm{scr}}(t)\right\}^*\times\frac{1}{\sqrt{N_c}}\sqrt{\frac{E_c}{T_c}}\sum_{k=0}^{N_c-1}\left[W_{u,2}^*(k)\bar{H}_{u,0}(k)\cdot S_{u,e}(k)-W_{u,0}(k)\bar{H}_{u,2}^*\cdot S_{u,o}(k)\right]\cdot\exp\left(j\frac{2\pi tk}{N_c}\right) \end{cases} \tag{30}$$

$$\begin{cases} \zeta_{\mathrm{MANI},e}(n)=\frac{1}{\mathrm{SF}_f}\sqrt{\frac{E_c}{T_c}}\sum_{i=0}^{\mathrm{SF}_f-1}\left[W_{u,0}^*\left(n+i\cdot\frac{N_c}{\mathrm{SF}_f}\right)\times\bar{H}_{u,2}\left(n+i\cdot\frac{N_c}{\mathrm{SF}_f}\right)\cdot S_{u,e}(k)-W_{u,2}\left(n+i\cdot\frac{N_c}{\mathrm{SF}_f}\right)\times\bar{H}_{u,0}^*\left(n+i\cdot\frac{N_c}{\mathrm{SF}_f}\right)\cdot S_{u,o}(k)\right] \\ \zeta_{\mathrm{MANI},o}(n)=\frac{1}{\mathrm{SF}_f}\sqrt{\frac{E_c}{T_c}}\sum_{i=0}^{\mathrm{SF}_f-1}\left[W_{u,2}^*\left(n+i\cdot\frac{N_c}{\mathrm{SF}_f}\right)\times\bar{H}_{u,0}\left(n+i\cdot\frac{N_c}{\mathrm{SF}_f}\right)\cdot S_{u,e}(k)-W_{u,0}\left(n+i\cdot\frac{N_c}{\mathrm{SF}_f}\right)\bar{H}_{u,2}^*\left(n+i\cdot\frac{N_c}{\mathrm{SF}_f}\right)\cdot S_{u,o}(k)\right] \end{cases} \tag{31}$$

Equation (37) can be viewed as conditional error probability, where the condition is that $\gamma_{u,o}(n)$ and $\gamma_{u,e}(n)$ are constant. To obtain the error probability when $\gamma_{u,o}(n)$ and $\gamma_{u,e}(n)$ are random, the average BER $P_{u,b}$ is given as

$$P_{u,b} = \iint P_{u,b}\left(\gamma_{u,o}(n), \gamma_{u,e}(n)\right) \times p\left(\gamma_{u,o}(n), \gamma_{u,e}(n)\right) d\,\gamma_{u,o}(n) d\,\gamma_{u,e}(n) \quad (38)$$

where $p(\gamma_{u,o}(n), \gamma_{u,e}(n))$ is the joint probability density function (pdf) of $\gamma_{u,o}(n)$ and $\gamma_{u,e}(n)$. If we can obtain the pdf of $p(\gamma_{u,o}(n), \gamma_{u,e}(n))$, it is possible to derive the accurate theoretical BER expression. However, it is extremely difficult to achieve the distributions of SINR due to the complexity equation. A reasonable simplification is to assume that the pdf of $p(\gamma_{u,o}(n), \gamma_{u,e}(n))$ is not random but constant to simplify the complexity of the calculations. Thus, we define the average

$$\begin{cases} \zeta_{\mathrm{MAI},e}(n) = \frac{1}{\mathrm{SF}_f} \sum\limits_{t=n\cdot\mathrm{SF}_f}^{(n+1)\mathrm{SF}_f-1} \left\{C_u^{\mathrm{SF}_f}(t) C_u^{\mathrm{scr}}(t)\right\}^* \times \frac{1}{\sqrt{N_c}} \sum\limits_{k=0}^{N_c-1} \left\{ W_{u,0}^*(k) \times \left[\sqrt{\frac{2E_c}{T_c}} \sum\limits_{\substack{u'=0\\u'\neq u}}^{U-1} Z_{u',e}(k) \times S_{u',e}(k)\right] \right. \\ \qquad \left. + W_{u,2}(k) \cdot \left[\sqrt{\frac{2E_c}{T_c}} \left(\sum\limits_{\substack{u'=0\\u'\neq u}}^{U-1} Z_{u',o}(k) \times S_{u',o}(k)\right)^*\right]\right\} \exp\left(j\frac{2\pi t k}{N_c}\right) \\ \zeta_{\mathrm{MAI},o}(n) = \frac{1}{\mathrm{SF}_f} \sum\limits_{t=n\cdot\mathrm{SF}_f}^{(n+1)\mathrm{SF}_f-1} \left\{C_u^{\mathrm{SF}_f}(t) C_u^{\mathrm{scr}}(t)\right\}^* \times \frac{1}{\sqrt{N_c}} \sum\limits_{k=0}^{N_c-1} \left\{ W_{u,2}^*(k) \times \left[\sqrt{\frac{2E_c}{T_c}} \sum\limits_{\substack{u'=0\\u'\neq u}}^{U-1} Z_{u',e}(k) \times S_{u',e}(k)\right] \right. \\ \qquad \left. - W_{u,0}(k) \cdot \left[\sqrt{\frac{2E_c}{T_c}} \left(\sum\limits_{\substack{u'=0\\u'\neq u}}^{U-1} Z_{u',o}(k) \times S_{u',o}(k)\right)^*\right]\right\} \exp\left(j\frac{2\pi t k}{N_c}\right) \end{cases} \quad (32)$$

$$\begin{cases} \zeta_{\mathrm{MAI},e}(n) = \frac{1}{\mathrm{SF}_f} \sum\limits_{i=0}^{\mathrm{SF}_f-1} \left\{C_u^{\mathrm{SF}_f}(t) C_u^{\mathrm{scr}}(t)\right\}^* \times \left\{ W_{u,0}^*\left(n+i\cdot\frac{N_c}{\mathrm{SF}_f}\right) \times \left[\sqrt{\frac{2E_c}{T_c}} \sum\limits_{\substack{u'=0\\u'\neq u}}^{U-1} Z_{u',e}\left(n+i\cdot\frac{N_c}{\mathrm{SF}_f}\right) \cdot S_{u',e}\left(n+i\cdot\left(\frac{N_c}{\mathrm{SF}_f}\right)\right)\right] \right. \\ \qquad \left. + W_{u,2}\left(n+i\cdot\frac{N_c}{\mathrm{SF}_f}\right) \times \left[\sqrt{\frac{2E_c}{T_c}} \left(\sum\limits_{\substack{u'=0\\u'\neq u}}^{U-1} Z_{u',o}\left(n+i\cdot\frac{N_c}{\mathrm{SF}_f}\right) \cdot S_{u',o}(n+i\cdot\left(\frac{N_c}{\mathrm{SF}_f}\right)\right)^*\right]\right\} \\ \zeta_{\mathrm{MAI},o}(n) = \frac{1}{\mathrm{SF}_f} \sum\limits_{i=0}^{\mathrm{SF}_f-1} \left\{C_u^{\mathrm{SF}_f}(t) C_u^{\mathrm{scr}}(t)\right\}^* \times \left\{ W_{u,2}^*\left(n+i\cdot\frac{N_c}{\mathrm{SF}_f}\right) \times \left[\sqrt{\frac{2E_c}{T_c}} \sum\limits_{\substack{u'=0\\u'\neq u}}^{U-1} Z_{u',e}\left(n+i\cdot\frac{N_c}{\mathrm{SF}_f}\right) \cdot S_{u',e}\left(n+i\cdot\left(\frac{N_c}{\mathrm{SF}_f}\right)\right)\right] \right. \\ \qquad \left. - W_{u,0}\left(n+i\cdot\frac{N_c}{\mathrm{SF}_f}\right) \times \left[\sqrt{\frac{2E_c}{T_c}} \left(\sum\limits_{\substack{u'=0\\u'\neq u}}^{U-1} Z_{u',o}\left(n+i\cdot\frac{N_c}{\mathrm{SF}_f}\right) \cdot S_{u',o}(n+i\cdot\left(\frac{N_c}{\mathrm{SF}_f}\right)\right)^*\right]\right\} \end{cases} \quad (33)$$

$$\begin{cases} \zeta_{\mathrm{noise},e}(n) = \frac{1}{\mathrm{SF}_f} \sum\limits_{t=n\cdot\mathrm{SF}_f}^{(n+1)\mathrm{SF}_f-1} \left\{C_u^{\mathrm{SF}_f}(t) C_u^{\mathrm{scr}}(t)\right\}^* \\ \qquad \times \frac{1}{\sqrt{N_c}} \sum\limits_{k=0}^{N_c-1} \left\{W_{u,0}^*(k) \times \Pi_{u,e}(k) + W_{u,2}(k) \times (\Pi_{u,o}(k))^*\right\} \exp\left(j\frac{2\pi t k}{N_c}\right) \\ \zeta_{\mathrm{noise},e}(n) = \frac{1}{\mathrm{SF}_f} \sum\limits_{t=n\cdot\mathrm{SF}_f}^{(n+1)\mathrm{SF}_f-1} \left\{C_u^{\mathrm{SF}_f}(t) C_u^{\mathrm{scr}}(t)\right\}^* \\ \qquad \times \frac{1}{\sqrt{N_c}} \sum\limits_{k=0}^{N_c-1} \left\{W_{u,2}^*(k) \times \Pi_{u,e}(k) - W_{u,0}(k) \times (\Pi_{u,o}(k))^*\right\} \exp\left(j\frac{2\pi t k}{N_c}\right) \end{cases} \quad (34)$$

$$\begin{cases} \zeta_{\mathrm{noise},e}(n) = \frac{1}{\mathrm{SF}_f} \sum\limits_{i=0}^{\mathrm{SF}_f-1} \left\{C_u^{\mathrm{SF}_f}(t) C_u^{\mathrm{scr}}(t)\right\}^* \\ \qquad \times \left\{W_{u,0}^*\left(n+i\cdot\frac{N_c}{\mathrm{SF}_f}\right) \times \Pi_{u,e}\left(n+i\cdot\frac{N_c}{\mathrm{SF}_f}\right) + W_{u,2}\left(n+i\cdot\frac{N_c}{\mathrm{SF}_f}\right) \times \Pi_{u,o}\left(n+i\cdot\frac{N_c}{\mathrm{SF}_f}\right)^*\right\} \\ \zeta_{\mathrm{noise},o}(n) = \frac{1}{\mathrm{SF}_f} \sum\limits_{i=0}^{\mathrm{SF}_f-1} \left\{C_u^{\mathrm{SF}_f}(t) C_u^{\mathrm{scr}}(t)\right\}^* \\ \qquad \times \left\{W_{u,2}^*\left(n+i\cdot\frac{N_c}{\mathrm{SF}_f}\right) \times \Pi_{u,e}\left(n+i\cdot\frac{N_c}{\mathrm{SF}_f}\right) - W_{u,0}\left(n+i\cdot\frac{N_c}{\mathrm{SF}_f}\right) \times \Pi_{u,o}\left(n+i\cdot\frac{N_c}{\mathrm{SF}_f}\right)^*\right\} \end{cases} \quad (35)$$

SINR of the uth user at odd time intervals and even time intervals as

$$\begin{cases} \bar{\gamma}_{u,e}(n) = \frac{\mathrm{E}\left[|\lambda_{u,e}(n)|^2\right]}{\mathrm{E}\left[|\zeta_{\mathrm{MAI},e}(n)+\zeta_{\mathrm{MANI},e}(n)+\zeta_{\mathrm{noise},e}(n)|^2\right]} \\ \bar{\gamma}_{u,o}(n) = \frac{\mathrm{E}\left[|\lambda_{u,o}(n)|^2\right]}{\mathrm{E}\left[|\zeta_{\mathrm{MAI},o}(n)+\zeta_{\mathrm{MANI},o}(n)+\zeta_{\mathrm{noise},o}(n)|^2\right]} \end{cases} \quad (39)$$

where $E[|\lambda_{u,e(o)}(n)|^2]$ represents the average signal power, and $E[|\zeta_{\mathrm{MAI},e(o)}(n) + \zeta_{\mathrm{MANI},e(o)}(n) + \zeta_{\mathrm{noise},e(o)}(n)|^2]$ is the average interference-plus-noise power. We can directly substitute $\bar{\gamma}_{u,o}(n)$ and $\bar{\gamma}_{u,e}(n)$ into (37), and then calculate the average BER. If we can confirm the distributions of the above random variables, the pdfs can be obtained.

In Section IV-A, it is found that $\lambda_{u,e(o)}(n)$, $\zeta_{\mathrm{MAI},e(o)}(n)$, $\zeta_{\mathrm{MANI},e(o)}(n)$, and $\zeta_{\mathrm{noise},e(o)}(n)$ are all sum signals; hence, according to the central limit theorem, they are approximated as complex zero-mean Gaussian variables. The variances of $\lambda_{u,e(o)}(n)$, $\zeta_{\mathrm{MAI},e(o)}(n)$, $\zeta_{\mathrm{MANI},e(o)}(n)$, and $\zeta_{\mathrm{noise},e(o)}(n)$ are denoted by $\sigma^2_{\lambda,e(o)}(n)$, $\sigma^2_{\mathrm{MAI},e(o)}(n)$, $\sigma^2_{\mathrm{MANI},e(o)}(n)$, and $\sigma^2_{\mathrm{noise},e(o)}(n)$, respectively. Here, we only illustrate the variances of MAI $\sigma^2_{\mathrm{MAI},e(o)}(n)$ (as shown in (40) and (41) in the following) and noise $\sigma^2_{\mathrm{noise},e(o)}(n)$ [as shown in (44) and (45) at the bottom of the page], the $\sigma^2_{\lambda,e(o)}(n)$ and $\sigma^2_{\mathrm{MANI},e(o)}(n)$ can be similarly achieved

$$\begin{cases} \sigma^2_{\mathrm{MAI},e}(n) = \frac{1}{\mathrm{SF}_f^2} \cdot \frac{2E_c}{T_c} \cdot \frac{\sigma_z^2}{N_c} \sum_{k=0}^{N_c-1} \left\{ \left|W^*_{u,0}(k)\right|^2 + |W_{u,2}(k)|^2 \right\} \\ \sigma^2_{\mathrm{MAI},o}(n) = \frac{1}{\mathrm{SF}_f^2} \cdot \frac{2E_c}{T_c} \cdot \frac{\sigma_z^2}{N_c} \sum_{k=0}^{N_c-1} \left\{ \left|W^*_{u,2}(k)\right|^2 + |W_{u,0}(k)|^2 \right\} \end{cases}$$

$$\mathrm{SC-CDMA} \quad (40)$$

$$\begin{cases} \sigma^2_{\mathrm{MAI},e}(n) = \frac{1}{\mathrm{SF}_f^2} \cdot \frac{2E_c}{T_c} \cdot \sigma_z^2 \sum_{i=0}^{\mathrm{SF}_f-1} \left\{ \left|W^*_{u,0}\left(n+i\cdot\frac{N_c}{\mathrm{SF}_f}\right)\right|^2 + \left|W_{u,2}\left(n+i\cdot\frac{N_c}{\mathrm{SF}_f}\right)\right|^2 \right\} \\ \sigma^2_{\mathrm{MAI},o}(n) = \frac{1}{\mathrm{SF}_f^2} \cdot \frac{2E_c}{T_c} \cdot \sigma_z^2 \sum_{i=0}^{\mathrm{SF}_f-1} \left\{ \left|W^*_{u,2}\left(n+i\cdot\frac{N_c}{\mathrm{SF}_f}\right)\right|^2 + \left|W_{u,0}\left(n+i\cdot\frac{N_c}{\mathrm{SF}_f}\right)\right|^2 \right\} \end{cases}$$

$$\mathrm{MC-CDMA} \quad (41)$$

where σ_z^2 is the variance of $Z_{b'\_u'}(k)$ defined in (26), given as

$$\sigma_z^2 = E\left[\left|\sum_{\substack{u'=0\\ \neq u}}^{U-1} S_{u'}(k) Z_{u'}(k)\right|^2\right]. \quad (42)$$

Since the channel gains are zero-mean complex Gaussian processes, $\bar{H}_{u,0(2)}(k)$ and $Z_{b'\_u'}(k)$ are also zero-mean complex Gaussian variables. Considering the use of Jakes model [38], each path consists of many irresolvable paths with the same time delay arriving from all directions uniformly; thereby, we can get

$$\sigma_z^2 = \sum_{\substack{u'=0\\ \neq u}}^{U-1} \frac{1}{(\mathrm{SF}_t)^2} \sum_{i=0}^{\mathrm{SF}_t-1} \sum_{i'=0}^{\mathrm{SF}_t-1} \left\{ J_0\left(2\pi|i-i'|f_D^{(u')T}\right) \times \left[C_{b'\_u'}^{\mathrm{SF}_{t\max}}(i) C_{b\_u}^{\mathrm{SF}_{t\max}}(i') \left\{ C_{b'\_u'}^{\mathrm{SF}_{b'\_u',t}}(i) C_{b\_u}^{\mathrm{SF}_{b\_u,t}}(i') \right\}^* \right] \right\} \quad (43)$$

where $f_D^{(u')}$ is the maximum Doppler frequency of the u'th user, and $J_0(\cdot)$ is the zeroth-order Bessel function of the first kind. In the case of block fading (i.e., $f_D^{(u')} = 0$), since block-level spreading codes are orthogonal to each other, we have $\sum_{i=0}^{SF_t-1} \{c_{u'}^{SF_t}\} \cdot \{c_u^{SF_t}\}^* = 0$ ($u \neq u'$); therefore, the second term of (21) becomes zero, and the MAI from other users can be removed.

Equations (44) and (45) show the variances of noise signals. Since the MAI, MANI, and noise can be viewed as being independent and identically distributed, the average power of interference plus noise can be given as

$$\mathrm{E}\left[\left|\zeta_{\mathrm{MAI},e(o)}(n)+\zeta_{\mathrm{MANI},e(o)}(n)+\zeta_{\mathrm{noise},e(o)}(n)\right|^2\right] = \mathrm{E}\left[\left|\zeta_{\mathrm{MAI},e(o)}(n)\right|^2+\left|\zeta_{\mathrm{MANI},e(o)}(n)\right|^2+\left|\zeta_{\mathrm{noise},e(o)}(n)\right|^2\right]. \quad (46)$$

Set $X_{u,e(o)} = |\zeta_{\mathrm{MAI},e(o)}(n)|^2 + |\zeta_{\mathrm{MANI},e(o)}(n)|^2 + |\zeta_{\mathrm{noise},e(o)}(n)|^2$, which can be viewed as a chi-square distribution (also chi-square or χ^2-distribution) with three degrees of freedom. Regarding the Chi-square distributions, it is always assumed that

$$\begin{cases} \sigma^2_{\mathrm{noise},e}(n) = \frac{1}{\mathrm{SF}_f^2} \cdot \frac{2N_0}{T_c} \cdot \frac{1}{N_c} \sum_{k=0}^{N_c-1} \left\{ \left|W^*_{u,0}(k)\right|^2 + |W_{u,2}(k)|^2 \right\} \\ \sigma^2_{\mathrm{noise},o}(n) = \frac{1}{\mathrm{SF}_f^2} \cdot \frac{2N_0}{T_c} \cdot \frac{1}{N_c} \sum_{k=0}^{N_c-1} \left\{ \left|W^*_{u,2}(k)\right|^2 + |W_{u,0}(k)|^2 \right\} \end{cases}, \quad \mathrm{SC-CDMA} \quad (44)$$

$$\begin{cases} \sigma^2_{\mathrm{noise},e}(n) = \frac{1}{\mathrm{SF}_f^2} \cdot \frac{2N_0}{T_c} \sum_{i=0}^{\mathrm{SF}_f-1} \left\{ \left|W^*_{u,0}\left(n+i\cdot\frac{N_c}{\mathrm{SF}_f}\right)\right|^2 + \left|W_{u,2}\left(n+i\cdot\frac{N_c}{\mathrm{SF}_f}\right)\right|^2 \right\} \\ \sigma^2_{\mathrm{noise},o}(n) = \frac{1}{\mathrm{SF}_f^2} \cdot \frac{2N_0}{T_c} \sum_{i=0}^{\mathrm{SF}_f-1} \left\{ \left|W^*_{u,2}\left(n+i\cdot\frac{N_c}{\mathrm{SF}_f}\right)\right|^2 + \left|W_{u,0}\left(n+i\cdot\frac{N_c}{\mathrm{SF}_f}\right)\right|^2 \right\} \end{cases}, \quad \mathrm{MC-CDMA} \quad (45)$$

TABLE III
SIMULATION CONDITION

Transmitter	Modulation	QPSK
	Chip-block length (no. of FFT points)	$N_c = 256$
	GI length	$N_g = 32$
	Spreading codes	Walsh sequences
	No. of transmit antennas	$N_t = 2$
	No. of relay user	1
	Cyclic delay	16
	Spreading factor	$SF(= SF_f \cdot SF_t) = 16$
Channel	Type of fading	Rayleigh
	Power delay profile	$L = 16$-path uniform
	Maximum Doppler frequency	$f_D T_c(N_c + N_g) = 0.001$ *or* 0.01
	Pass loss exponent	$\alpha = 4$
Receiver	No. of receive antennas	$N_r = 1$
	Channel estimation	Ideal
	Equalization	MMSE-FDE

each term has a common variance, the pdf of variable $X_{u,\,e(o)}$ is thereby given as

$$p(\chi) = \begin{cases} \frac{1}{2^{1.5}\Gamma(1.5)(\bar{\sigma})^3}\chi^{0.5}\exp\left(-\frac{\chi}{2\bar{\sigma}^2}\right), & \chi > 0 \\ 0, & \text{otherwise} \end{cases} \tag{47}$$

where $\bar{\sigma}((\sigma^2_{\mathrm{MAI},\,e(o)}(n) + \sigma^2_{\mathrm{MANI},\,e(o)}(n) + \sigma^2_{\mathrm{noise},\,e(o)}(n))/3)$ is the average variance that is used, instead of the common variance. Equation (47) is only an approximate expression because of the approximate calculation of the average variance. We can use the feature of chi-square random variables that their distributions approach that of Gaussian random variables when the number of degrees of freedom becomes large. Thus, to simplify the processing of calculations, we viewed $X_{u,\,e(o)}$ as complex Gaussian random variables instead of chi-square random variables, and the variance of $X_{u,\,e(o)}$ is equal to

$$\begin{cases} \sigma^2_{X_{u,\,e}}(n) = \sigma^2_{\mathrm{MAI},\,e}(n) + \sigma^2_{\mathrm{MANI},\,e}(n) + \sigma^2_{\mathrm{noise},\,e}(n) \\ \sigma^2_{X_{u,\,o}}(n) = \sigma^2_{\mathrm{MAI},o}(n) + \sigma^2_{\mathrm{MANI},\,e}(n) + \sigma^2_{\mathrm{noise},\,o}(n). \end{cases} \tag{48}$$

In this paper, only two source antennas and two relay antennas are considered, which form a virtual-four-antenna STCDTD system. Of course, more than two transmit antennas can be used; however, it is quite difficult if not impossible to theoretically analyze the achievable BER performance for the case of more than four antennas. A BER analysis of 2-D block-spread SC-CDMA using virtual STCDTD with more than four antennas is left as a future study.

In this paper, it is assumed that slow transmit power control (TPC) is used during the transmission from the user node and its relay node to BS to compensate for the path loss and shadowing loss. The shadowing loss is known to follow the lognormal distribution. Its standard deviation depends on the propagation scenario. According to [37], it is 8 dB for both suburb macro and urban macro scenarios, whereas it is 10 and 4 dB for a non-line-of-sight urban micro scenario and a line-of-sight urban micro scenario, respectively. Therefore, the transmit power (accordingly, the interference power) would be different for different user locations when slow TPC is used. Therefore, operating the proposed scheme in different propagation scenarios is another attractive topic.

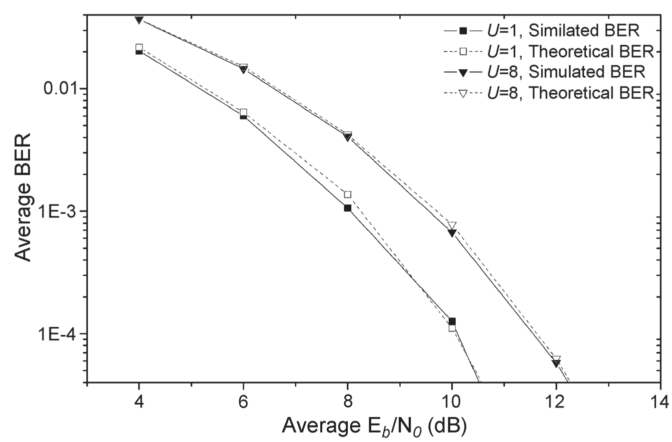

Fig. 6. BER performances of 2-D block-spread SC-CDMA using virtual-four-antenna STCDTD between simulation results and theoretical results.

V. SIMULATION RESULTS

The simulation condition is shown in Table III. Assume that the spreading factor of $\mathrm{SF}(= \mathrm{SF}_f \cdot \mathrm{SF}_t) = 16$, and $\mathrm{SF}_t = U$, which is sufficient for the block-level despreading to remove the MAI. The most popular channel model is the space channel model (SCM), which was first proposed in 3GPP TR 25.996 Release 8 and further modified in ETSI TR.125.996 [37]. For the SCM, as it considers six resolvable paths, with each consisting of 20 unresolvable paths, it seems to be applicable to the frequency bandwidth narrower than or equal to 5 MHz. Hence, for theoretical analysis, we use more general and tractable Jakes channel model [38] instead of the SCM. It is assumed that the Jakes channel model with a 16-path uniform power delay profile (each path consisting of 32 unresolvable paths) and the normalized maximum Doppler frequency of $f_D T_c(N_c + N_g) = 0.001$ or 0.01, which corresponds to a moving speed of 60 or 6 km/h at 2-GHz carrier frequency for a data rate of 32 Msymbol/s.

The theoretical and simulated average BER performances of 2-D block-spread SC-CDMA using virtual-four-antenna STCDTD are plotted in Fig. 6. The theoretical average BER was obtained using the Monte Carlo numerical method. First, the conditional average BER for the multipath channel gains

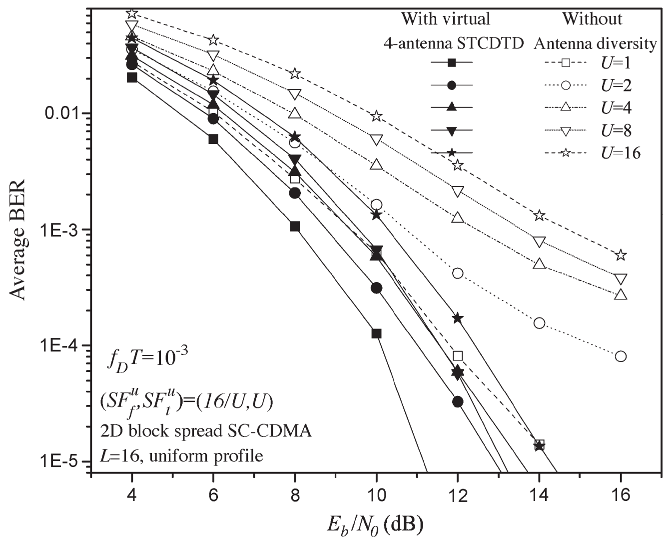

Fig. 7. BER performances of 2-D block-spread SC-CDMA using virtual-four-antenna STCDTD.

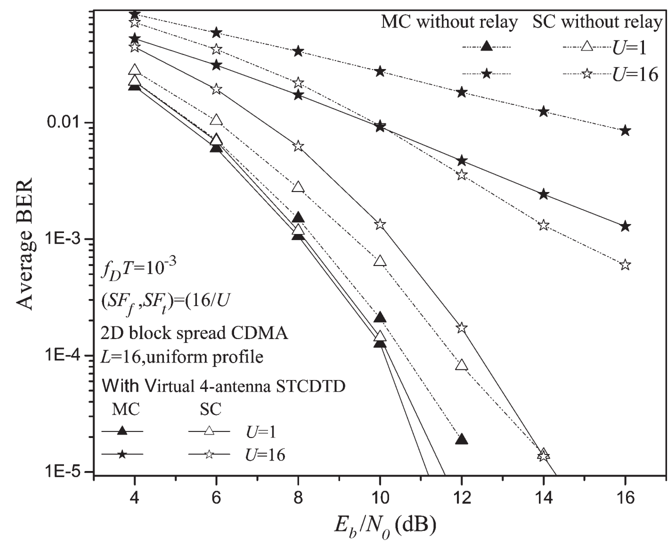

Fig. 8. BER performances comparison between 2-D block-spread MC-CDMA and SC-CDMA when virtual-four-antenna STCDTD is used.

was computed using (6), (15), and (26). Then, the average BER was obtained by repeating the above computation a sufficient number of times by changing the multipath channel gains. To obtain the simulated average BER, the average BER was measured by transmitting a large number of chip blocks. Owing to compare the transmitted and received digital data, the simulated results could be achieved. It is found that the theoretical results agree fairly well with the simulated results. Hence, for convenience, we only show the results obtained by the simulation method.

Performance comparisons between 2-D block-spread CDMA using virtual-four-antenna STCDTD and 2-D block-spread CDMA without relay/antenna diversity are shown in Figs. 7–9.

In Fig. 7, it is found that the use of virtual-four-antenna STCDTD can significantly improve the BER performance. When $U = 1$ (no interference user), the use of virtual-four-antenna STCDTD reduces, by about 1.3 dB, the required E_b/N_0 for achieving $\mathrm{BER} = 10^{-3}$. The BER performance improvement becomes more significant as U increases. When $U = 16$ (the number of interference users is 15), the E_b/N_0 reduction can be as much as 4.3 dB. As U increases, SF_f becomes smaller; therefore, the available frequency diversity gain by MMSE-FDE becomes smaller. As a consequence, the relative diversity gain achievable by antenna becomes stronger.

According to Fig. 7, it is seen that the BER floor exists when no antenna diversity is used; however, the use of antenna diversity can sufficiently suppress the floor and achieve a good BER performance. This is because antenna diversity takes full advantages of the frequency-selective Rayleigh fading channel by using FDE.

Taking SC-CDMA with antenna diversity during an even time interval as an example, the soft-decision output from IFFT after FDE is given as $\hat{d}_{u,e}(n) = \lambda_{u,e}(n) + \zeta_{\mathrm{MANI},e}(n) + \zeta_{\mathrm{MAI},e}(n) + \zeta_{\mathrm{noise},e}(n)$ [according to (27)]. (The definition for the variables of the soft-decision output equation is given at the bottom of the next page.)

From the definition, it can be understood that the desired signal at subcarrier k is given by $[W_{u,0}^*(k)\bar{H}_{u,0}(k) \cdot S_{u,e}(k) + W_{u,2}(k)\bar{H}_{u,2}^* \cdot S_{u,o}(k)]$. In addition, assume that $[W_{u,0}^*(k)\bar{H}_{u,0}(k) \cdot S_{u,e}(k) + W_{u,2}(k)\bar{H}_{u,2}^* \cdot S_{u,o}(k)] = A(k) \cdot S_{u,e}(k)$, where $A(k)$ is the equivalent gain. By computer calculation, it was found that $A(k) > 1$; therefore, the received signal power is increased by $|A(k)|^2$. Moreover, as the MANI at subcarrier k is given as $[W_{u,0}^*(k)\bar{H}_{u,2}(k) \cdot S_{u,e}(k) - W_{u,2}(k)\bar{H}_{u,0}^* \cdot S_{u,o}(k)]$ with the assumption $[W_{u,0}^*(k)\bar{H}_{u,2}(k) \cdot S_{u,e}(k) - W_{u,2}(k)\bar{H}_{u,0}^* \cdot S_{u,o}(k)] = B(k) \cdot S_{u,e}(k)$, where $B(k)$ is the equivalent gain. By computer calculation, it is found that $B(k) < 1$, which reduces the MANI. Similarly, it can be proven that the equivalent gain of MAI is also smaller than 1. Hence, the SINR of the proposed scheme is larger than the case without antenna diversity since both the desired signal and the MAI are multiplied by the same factor $W_{u,0}^*(k)$ for the case without multiple antennas.

The comparison between MC-CDMA and SC-CDMA is shown in Fig. 8. For both cases of with and without virtual-four-antenna STCDTD, the BER performances degrade as U increases. This is because the chip-level spreading factor SF_f decreases for the given total spreading factor ($\mathrm{SF} = \mathrm{SF}_f \cdot \mathrm{SF}_t = 16$). Hence, the residual ISI, which is present after FDE, increases in the case of SC-CDMA. In the case of MC-CDMA, since the diversity gain is obtained from the SF_f subcarriers, the smaller SF_f is, the worse the BER performance becomes.

Consider one extreme case of $U = 1$ ($\mathrm{SF}_f = 16, \mathrm{SF}_t = 1$). It is shown in Fig. 8 that without relay, MC-CDMA has better BER performance than SC-CDMA due to the interleaving of MC-CDMA. However, when using virtual-four-antenna STCDTD, SC-CDMA obtains almost the same BER performances as MC-CDMA.

Take another extreme case of $U = 16$ ($\mathrm{SF}_f = 1, \mathrm{SF}_t = 16$) as an example. In this case, no ISI suppression is achieved (spreading is used to remove the MAI only) for SC-CDMA. Since the chip-level spreading $\mathrm{SF}_f = 1$, no frequency diversity gain is obtained for MC-CDMA. The bandwidth of SC-CDMA is wider than that of MC-CDMA when two systems suppose the same data rate; thus, SC-CDMA provides better BER performance.

However, which CDMA is selected by system designers should be determined not only by the BER performances but also by other factors. For example, for the downlink transmission (from the BS to the user terminal), MC-CDMA is more flexible in frequency–time resource-block allocation than SC-CDMA. However, for the uplink transmission (from user terminal to BS), SC-CDMA provides more advantages due to its lower peak-to-average power ratio.

The 2-D block-spread CDMA assumes the block slow fading to remove the MAI. However, if this assumption does not hold (i.e., a fast fading environment), the MAI remains, and the achievable BER performance degrades. How the normalized maximum Doppler frequency f_DT impacts the achievable BER performances is shown in Fig. 9 when $U = 8$. Almost the same BER performance can be achieved if f_DT is below 10^{-3} when SC-CDMA is used. STCDTD provides about 4.3-dB gain in the required E_b/N_0 for BER $= 10^{-3}$. However, when $f_DT = 10^{-2}$, the path gains vary over $\mathrm{SF}_t = 8$ consecutive chip blocks; hence, the orthogonality among different users is distorted. As a consequence, the MAI remains, thereby significantly degrading the BER performance. In such a fast fading environment, the BER performance with STCDTD is degraded.

Similar results can be derived in the case of MC-CDMA. It is seen that, when $U = 8$, SC-CDMA always have a better BER performance than MC-CDMA.

A performance comparison is made between 2-D block-spread CDMA using relay virtual-four antennas, and the source user directly adopts 2-D block-spread CDMA with four-antenna STTD/CDTD without any relay scheme, as shown in Figs. 10 and 11.

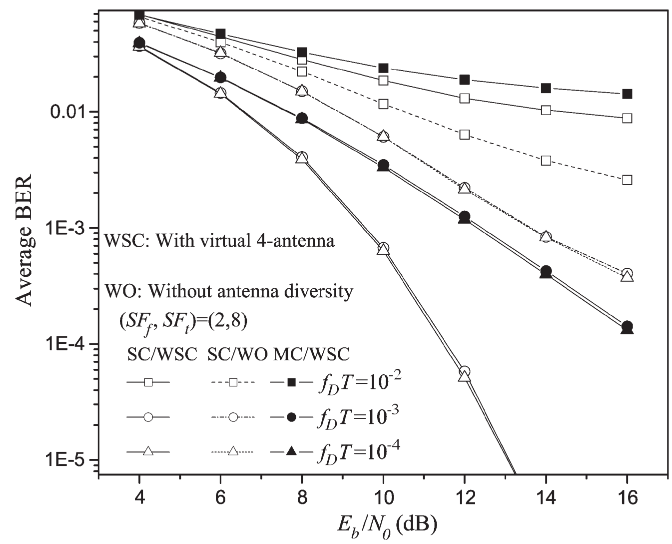

Fig. 9. Impact of f_DT on 2-D block-spread SC-CDMA and MC-CDMA using virtual-four-antenna STCDTD.

It is shown in Fig. 10 that the use of four-antenna STTD has the best BER performance, followed by relay-virtual-four-antenna STCDTD, whereas the use of four-antenna CDTD only provides only a slight performance improvement. As four-antenna STTD can take full advantage of the transmit diversity, it provides better BER performance. However, the difference in the required E_b/N_0 for BER $= 10^{-3}$ between relay-virtual-four-antenna STCDTD and four-antenna STTD is as small as about 0.2 dB when $U = 1$. However, it should be noted that

$$\begin{aligned}
\lambda_{u,e}(n) &= \frac{1}{\mathrm{SF}_f}\sum_{t=n\cdot \mathrm{SF}_f}^{(n+1)\mathrm{SF}_f-1}\left\{C_u^{\mathrm{SF}_f}(t)C_u^{\mathrm{scr}}(t)\right\}^* \\
&\quad\times\frac{1}{\sqrt{N_c}}\sqrt{\frac{E_c}{T_c}}\sum_{k=0}^{N_c-1}\left[W_{u,0}^*(k)\bar{H}_{u,0}^*(k)\cdot S_{u,e}(k)+W_{u,2}(k)\bar{H}_{u,2}^*\cdot S_{u,o}(k)\right]\cdot\exp\left(j\frac{2\pi tk}{N_c}\right) \\
\zeta_{\mathrm{MANI},e}(n) &= \frac{1}{\mathrm{SF}_f}\sum_{t=n\cdot \mathrm{SF}_f}^{(n+1)\mathrm{SF}_f-1}\left\{C_u^{\mathrm{SF}_f}(t)C_u^{\mathrm{scr}}(t)\right\}^* \\
&\quad\times\frac{1}{\sqrt{N_c}}\sqrt{\frac{E_c}{T_c}}\sum_{k=0}^{N_c-1}\left[W_{u,0}^*(k)\bar{H}_{u,2}(k)\cdot S_{u,e}(k)-W_{u,2}(k)\bar{H}_{u,0}^*\cdot S_{u,o}(k)\right]\cdot\exp\left(j\frac{2\pi tk}{N_c}\right) \\
\zeta_{\mathrm{MAI},e}(n) &= \frac{1}{\mathrm{SF}_f}\sum_{t=n\cdot \mathrm{SF}_f}^{(n+1)\mathrm{SF}_f-1}\left\{C_u^{\mathrm{SF}_f}(t)C_u^{\mathrm{scr}}(t)\right\}^* \\
&\quad\times\frac{1}{\sqrt{N_c}}\sum_{k=0}^{N_c-1}\left\{W_{u,0}^*(k)\times\left[\sqrt{\frac{2E_c}{T_c}}\sum_{\substack{u'=0\\u'\neq u}}^{U-1}Z_{u',e}(k)\times S_{u',e}(k)\right]\right. \\
&\qquad\left.+W_{u,2}(k)\times\left[\sqrt{\frac{2E_c}{T_c}}\left(\sum_{\substack{u'=0\\u'\neq u}}^{U-1}Z_{u',o}(k)\times S_{u',o}(k)\right)^*\right]\right\}\exp\left(j\frac{2\pi tk}{N_c}\right)
\end{aligned}$$

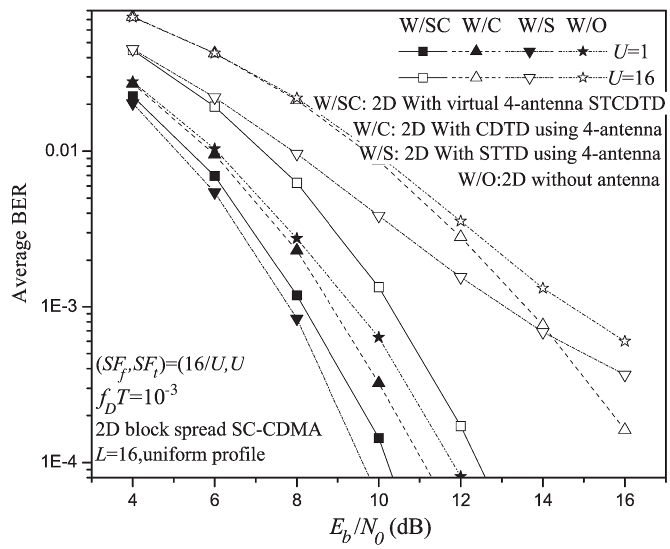

Fig. 10. BER performance comparisons among different antenna diversity methods for 2-D block-spread SC-CDMA.

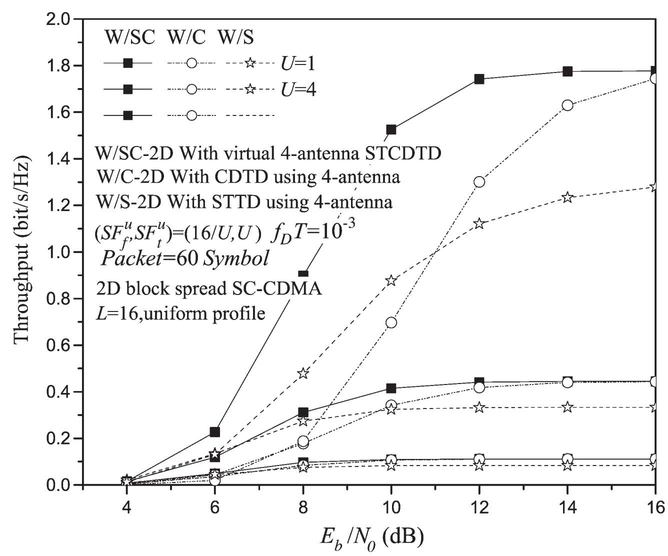

Fig. 11. Throughput comparison among different antenna diversity methods.

relay-virtual-four-antenna STCDTD has a much better throughput than four-antenna STTD. Here, the total throughput S is defined as

$$S = \frac{1}{\text{SF}} \cdot U \cdot R \cdot (\log_2 M)(1 - \text{PER}) \left(\frac{1}{1 + N_g/N_c} \right) \quad (49)$$

where U is the number of users, R is the transmit diversity coding rate ($R = 1$ for both four-antenna CDTD and relay-virtual-four-antenna STCDTD, whereas $R = 3/4$ for four-antennas STTD), M is the modulation level ($M = 4$ for quadrature PSK), and PER is the packet error rate; the throughput is shown in Fig. 11. The maximum throughput (achievable with PER $=$ 0) when SF $= 16$ and $U = 16$ becomes $S = 1.78$ bit/s/Hz. For a high-E_b/N_0 region (above 12 dB), relay-virtual-four-antenna STCDTD and four-antenna CDTD can achieve the maximum throughput of 1.78 bit/s/Hz (when $U = 16$); however, four-antenna STTD can only achieve a throughput of 1.335 bit/s/Hz.

In Figs. 10 and 11, it is clearly confirmed that the virtual-four-antenna STCDTD relay provides a low BER and a high throughput.

VI. Conclusion

In this paper, a 2-D block-spread CDMA using relay-virtual-four-antenna STCDTD has been proposed, and the theoretical BER has been discussed in detail. Computer simulations showed that the proposed scheme is capable of removing the MAI when the channel is slow fading, meanwhile obtaining both the frequency diversity gain and antenna diversity gain, and therefore significantly improving the BER performances. While the decoding error at the relay is a critical and important issue for the proposed scheme, investigating a realistic BER performance with HARQ in cooperative communications is left for future studies.

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