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Performance of Broadband Multicarrier DS-CDMA Using Space–Time Spreading-Assisted Transmit Diversity

Lie-Liang Yang, *Senior Member, IEEE,* and Lajos Hanzo, *Fellow, IEEE*

***Abstract*—In this contribution multicarrier direct-sequence code-division multiple-access (MC DS-CDMA) using space–time spreading (STS)-assisted transmit diversity is investigated in the context of broadband communications over frequency-selective Rayleigh-fading channels. We consider the issue of parameter design for the sake of achieving high-efficiency communications in various dispersive environments. Furthermore, in contrast to conventional MC DS-CDMA schemes employing time (T)-domain spreading only, in this contribution we also investigate broadband MC DS-CDMA schemes employing both T-domain and frequency (F)-domain spreading, i.e., employing TF-domain spreading. The bit-error rate (BER) performance of STS-assisted broadband MC DS-CDMA is investigated for downlink transmissions associated with the correlation based single-user detector and the decorrelating multiuser detector. Our study demonstrated that when appropriately selecting the system parameters, broadband MC DS-CDMA using STS-assisted transmit diversity constitutes a promising downlink transmission scheme. This scheme is capable of supporting ubiquitous communications over diverse communication environments without BER performance degradation.**

***Index Terms*—Broadband system, code-division multiple access (CDMA), frequency-domain spreading, frequency-selective fading, multicarrier direct-sequence code-division multiple-access (MC-DS-CDMA), multicarrier CDMA, multicarrier modulation, multiuser detection, space–time spreading (STS), transmit diversity.**

I. Introduction

ONE of the basic requirements in broadband mobile wireless systems is that of supporting the expected high bit rates required by wireless Internet services and for delivering high-speed multimedia services. However, the achievable capacity and data rate of wireless communication systems are limited by the time-varying characteristics of the dispersive fading channel. An efficient technique of combating the time-varying effects of wireless channels is employing diversity. In recent years, space–time coding has received much attention as an effective transmit diversity technique used for combating fading in wireless communications [1]–[3]. Inspired by space–time codes, in [4] an attractive transmit diversity scheme based on space–time spreading (STS) has been proposed for employment in code-division multiple-access (CDMA) systems [5]. An STS scheme designed for supporting two transmission antennas and one receiver antenna has been included also in the cdma2000 wideband CDMA (W-CDMA) standard [5], [6]. In [4], the performance of single-carrier CDMA systems using STS has been investigated, when the channel is modeled either as a flat or as a frequency-selective Rayleigh-fading channel in the absence of multiuser interference (MUI).

In this contribution, we investigate the issues of parameter design and bit-error rate (BER) performance of broadband multicarrier direct-sequence CDMA (MC DS-CDMA) using STS-assisted transmit diversity, when communicating over frequency-selective Rayleigh-fading channels. The reason of considering broadband MC DS-CDMA is because it constitutes a generalized multiple-access scheme [5], [7]–[10] which exhibits a high grade of design high flexibility. MC DS-CDMA possesses a range of parameters that can be adjusted for satisfying the required design tradeoffs. Our objective in the context of parameter design is to configure the broadband MC DS-CDMA scheme for achieving high-efficiency communications in various propagation environments characterized by different grade of dispersion. In this contribution, specifically, synchronous downlink (base-to-mobile) transmission of the user signals is considered and the BER performance is evaluated for a range of parameter values. Furthermore, in contrast to the family of conventional MC DS-CDMA schemes employing time (T)-domain spreading only, in this contribution we also investigate the broadband MC DS-CDMA scheme's performance, when employing both T-domain and frequency (F)-domain spreading, i.e., employing TF-domain spreading. A typical advantage of using TF-domain spreading in MC DS-CDMA is that the maximum number of users supported is determined by the product of the T-domain spreading factor and the F-domain spreading factor. Therefore, the MC DS-CDMA system using TF-domain spreading is capable of supporting a significantly higher number of users, than in case of using solely T-domain spreading, since in this case the maximum

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The authors are with the School of Electronics and Computer Sciences, University of Southampton, SO17 1BJ, U.K. (e-mail: lly@ecs.soton.ac.uk; lh@ecs.soton.ac.uk).

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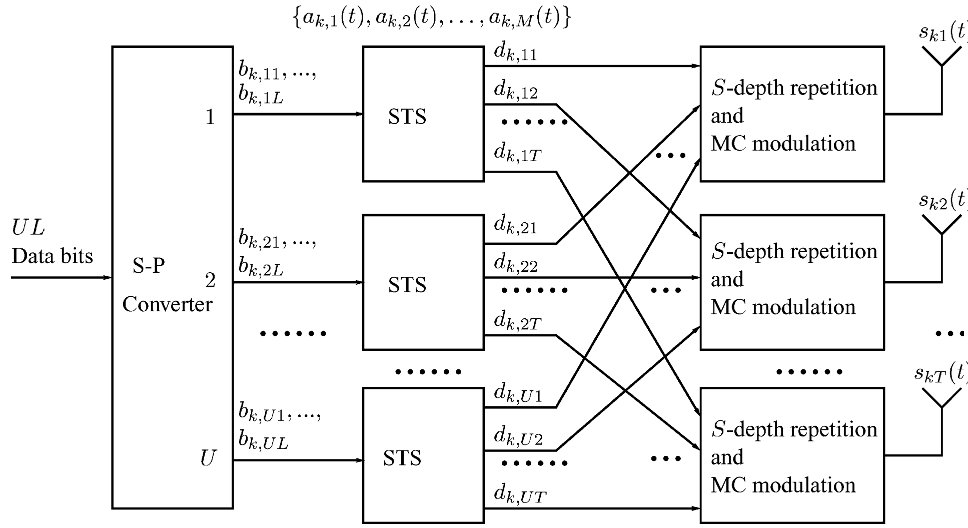

Fig. 1. The transmitter schematic of the MC DS-CDMA system using space-time spreading.

number of users supported is determined by the T-domain spreading factor alone. The performance of the STS-assisted broadband MC DS-CDMA scheme using TF-domain spreading is investigated in conjunction with both the correlation based single-user detector and the decorrelating multiuser detector [5], [11]. Our study shows that a high number of users can be supported by broadband MC DS-CDMA using TF-domain spreading, without having to impose tradeoffs in terms of the achievable diversity order.

The remainder of this contribution is organized as follows. In the next section, the STS-assisted broadband MC DS-CDMA scheme is described and the required parameter values are investigated. In Section III, we derive the achievable BER and characterize the BER performance of the broadband MC DS-CDMA scheme using STS. Section IV considers the achievable system capacity improvement of broadband MC DS-CDMA using STS and TF-domain spreading. Finally, our conclusions are offered in Section V.

II. System Description

A. Transmitter Model

The system considered in this paper is an orthogonal MC DS-CDMA scheme [5], [8] using $U \cdot S$ number of subcarriers, T number of transmitter antennas, and one receiver antenna. Furthermore, in this paper, a synchronous MC DS- CDMA scheme is investigated, where the K user signals are transmitted synchronously. The transmitter schematic of the kth user is shown in Fig. 1, where real-valued data symbols using binary phase-shift keying (BPSK) modulation and real-valued spreading [4], [5] were considered. Fig. 2 shows the frequency arrangement of the $U \cdot S$ subcarriers. As shown in Fig. 1, at the transmitter side a block of $U \cdot L$ data bits each having a bit duration of T_b is serial-to-parallel (S-P) converted to U parallel subblocks. Each parallel subblock has L data bits, which are space–time spread using the schemes of [4] with the aid of M orthogonal spreading codes [5]

$$\{a_{k,1}(t), a_{k,2}(t), \ldots, a_{k,M}(t)\}, \qquad k = 1, 2, \ldots, K$$

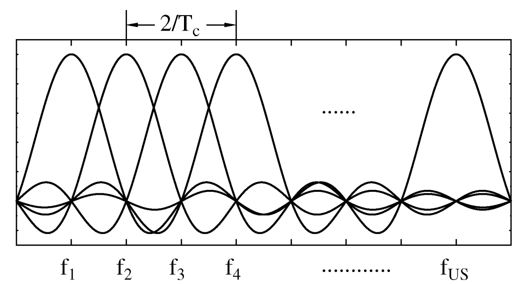

Fig. 2. Spectrum of orthogonal MC DS-CDMA signals having a minimum subcarrier spacing of $1/T_c$, where the zero-to-zero bandwidth of each DS spread signal is $2/T_c$.

and mapped to T transmitter antennas. The symbol duration of the STS signals is ULT_b, and the discrete period of the orthogonal codes is $ULT_b/T_c = ULN$, where $N = T_b/T_c$ and T_c represents the chip duration of the orthogonal spreading codes. The orthogonal codes take the form of

$$a_{k,i}(t) = \sum_{j=0}^{ULN-1} a_{k,i}[j] P_{T_c}(t - jT_c)$$

where $a_{k,i}[j] \in \{+1, -1\}$ and they obey the relationship of

$$\sum_{l=0}^{ULN-1} a_{i,m}[l] a_{j,n}[l] = 0$$

whenever $i \neq j$ or $m \neq n$. Furthermore, $P_{T_c}(t)$ represents the chip impulse waveform defined over the interval of $[0, T_c)$. Since the total number of orthogonal codes having a discrete period of ULN is ULN and since each user requires M orthogonal codes for STS, the maximum number of users supported by these orthogonal codes is ULN/M. As seen in Fig. 1, following STS, each STS block generates T parallel signals to be mapped to the T transmitter antennas. For each transmitter antenna, the specific U STS signals generated by the U STS blocks are then repeated S times, so that each STS signal is transmitted on S subcarriers. The corresponding S number of subcarreirs are selected for guaranteeing that the same STS signal is transmitted by the specific S subcarriers having the maximum possible frequency spacings, so that they experience independent fading and hence achieve maximum frequency diversity for a given S value. Specifically, let $\{f_1, f_2, \ldots, f_{US}\}$ be the subcarrier frequencies, which are arranged according to Fig. 2. These subcarrier frequencies can be written in the form of a matrix as

$$\{f_i\} = \begin{pmatrix} f_1 & f_{U+1} & \cdots & f_{(S-1)U+1} \\ f_2 & f_{U+2} & \cdots & f_{(S-1)U+2} \\ \vdots & \vdots & \ddots & \vdots \\ f_U & f_{2U} & \cdots & f_{SU} \end{pmatrix}. \quad (1)$$

Then, an STS signal will be transmitted using the subcarrier frequencies from the same row of (1). Finally, as shown in Fig. 1, the inverse fast Fourier transform (IFFT) is invoked for carrying out multicarrier modulation, and the IFFT block's output signal is transmitted using one of the transmitter antennas.

The general form of the kth user's transmitted baseband signal corresponding to the T transmitter antennas can be expressed as

$$\boldsymbol{s}_k(t) = \text{Re}\left\{\sqrt{\frac{2E_b}{UT_b}\frac{1}{SMT}}\boldsymbol{D}_k\boldsymbol{P}\boldsymbol{w}\right\} \quad (2)$$

where E_b/UT_b represents the transmitted power per subcarrier expressed as $LE_b/ULT_b = E_b/UT_b$, the factor S in the denominator is due to the S-depth repetition, while the factor of MT represents STS using M orthogonal codes and T transmitter antennas. In (2), $\boldsymbol{s}_k(t) = [s_{k1}(t)\ s_{k2}(t)\ \ldots\ s_{kT}(t)]^T$—where the superscript T denotes the vector or matrix transpose—represents the transmitted signal vector of the T transmitter antennas, $\boldsymbol{P}$ represents the S-depth repetition operation, which is a $U \times US$ matrix expressed as $\boldsymbol{P} = [\boldsymbol{I}_U\ \boldsymbol{I}_U\ \ldots\ \boldsymbol{I}_U]$ with $\boldsymbol{I}_U$ being a unity matrix of rank U. Furthermore, in (2), $\boldsymbol{D}_k = \{d_{k,ij}\}$ is a $U \times T$-dimensional matrix representing the outputs of the STS. The STS invoked in matrix $\boldsymbol{D}_k$ can be expressed as $\boldsymbol{D}_k = [\boldsymbol{C}_k\boldsymbol{B}_k]^T$, where $\boldsymbol{C}_k$ is a $U \times UM$-dimensional matrix constituted by orthogonal codes, which can be expressed as

$$\boldsymbol{C}_k^T = \begin{pmatrix} a_{k,1}(t) & 0 & \ldots & 0 \\ a_{k,2}(t) & 0 & \ldots & 0 \\ \vdots & \vdots & \ddots & \vdots \\ a_{k,M}(t) & 0 & \ldots & 0 \\ 0 & a_{k,1}(t) & \ldots & 0 \\ 0 & a_{k,2}(t) & \ldots & 0 \\ \vdots & \vdots & \ddots & \vdots \\ 0 & a_{k,M}(t) & \ldots & 0 \\ \vdots & \vdots & \ddots & \vdots \\ 0 & 0 & \vdots & a_{k,1}(t) \\ 0 & 0 & \vdots & a_{k,2}(t) \\ \vdots & \vdots & \ddots & \vdots \\ 0 & 0 & \vdots & a_{k,M}(t) \end{pmatrix}. \quad (3)$$

In (2), $\boldsymbol{B}_k$ is a $UM \times T$ matrix mapped from the U subblock data bits, according to the requirements of the STS [4]. Specifically, the matrix $\boldsymbol{B}_k$ can be expressed as

$$\boldsymbol{B}_k = \left[\boldsymbol{B}_{k1}^T\ \boldsymbol{B}_{k2}^T\ \ldots\ \boldsymbol{B}_{kU}^T\right]^T \quad (4)$$

where $\boldsymbol{B}_{ku}$ for $u = 1, 2, \ldots, U$ are $M \times T$-dimensional matrices, which obey the structure of

$$\boldsymbol{B}_{ku} = \begin{pmatrix} a_{11}b'_{k,11} & a_{12}b'_{k,12} & \cdots & a_{1L}b'_{k,1T} \\ a_{21}b'_{k,21} & a_{22}b'_{k,22} & \cdots & a_{2L}b'_{k,2T} \\ \vdots & \vdots & \ddots & \vdots \\ a_{M1}b'_{k,M1} & a_{U2}b'_{k,M2} & \cdots & a_{ML}b'_{k,MT} \end{pmatrix}, \quad u = 1, 2, \ldots, U \quad (5)$$

where a_{ij} represents the sign of the element at the ith row and the jth column, which is determined by the STS design rule, while $b'_{k,ij}$ in $\boldsymbol{B}_{ku}$ is the data bit assigned to the (i,j)th element, which is one of the L input data bits $\{b_{k,u1}, b_{k,u2}, \ldots, b_{kL}\}$ of user k. For example, for $L = M = T = 2$, the corresponding $\boldsymbol{B}_{ku}$ matrices are given by [4]

$$\begin{pmatrix} b_{k,u1} & b_{k,u2} \\ b_{k,u2} & -b_{k,u1} \end{pmatrix}, \qquad u = 1, 2, \ldots, U. \quad (6)$$

Finally, in (2), $\boldsymbol{w}$ represents the multicarrier modulated vector of length US, which can be expressed as

$$\boldsymbol{w} = [\exp(j2\pi f_1 t)\ \exp(j2\pi f_2 t)\ \ldots\ \exp(j2\pi f_{SU} t)]^T.$$

Equation (2) represents the general form of the transmitted signals using STS, regardless of the values of L, M, and T. However, the study conducted in [4] has shown that STS schemes using $L = M = T$, i.e., those having an equal number of data bits, orthogonal STS-related spreading sequences, as well as transmission antennas constitute attractive schemes, since they are capable of providing maximal transmit diversity without requiring extra STS codes. Therefore, in this contribution, we only investigate these attractive STS schemes, and our results are mainly based on MC DS-CDMA systems using two or four transmitter antennas.

As an example, for the case of $L = M = T = 2$, the MC DS-CDMA signals transmitted by antenna 1 and 2 can be simply expressed as

$$\boldsymbol{s}_k(t) = \begin{pmatrix} s_{k1}(t) \\ s_{k2}(t) \end{pmatrix} = \sqrt{\frac{2E_b}{4UST_b}} \begin{pmatrix} \sum_{u=1}^{U}\sum_{s=1}^{S}[a_{k,1}b_{k,u1} + a_{k,2}b_{k,u2}] \\ \times \cos\left(2\pi f_{(s-1)U+u}t\right) \\ \sum_{u=1}^{U}\sum_{s=1}^{S}[a_{k,1}b_{k,u2} - a_{k,2}b_{k,u1}] \\ \times \cos\left(2\pi f_{(s-1)U+u}t\right) \end{pmatrix}. \quad (7)$$

Note that in (7), the explicit notation indicating the time dependence of $a_{k,i}(t)$ has been omitted for notational convenience, since in this contribution only synchronous transmissions are considered.

B. Channel Model and System Parameter Design

The channels are assumed to be slowly varying frequency-selective Rayleigh-fading channels and the delay spreads are assumed to be limited to the range of $[T_m, T_M]$, where T_m corresponds to the environments having the shortest delay spread considered, for example, in an indoor environment, while T_M is associated with an environment having the highest possible delay spread, as in an urban area. Below, we impose some limitations on the set of parameters used by STS-assisted MC DS-CDMA, in order to ensure that MC DS-CDMA operate

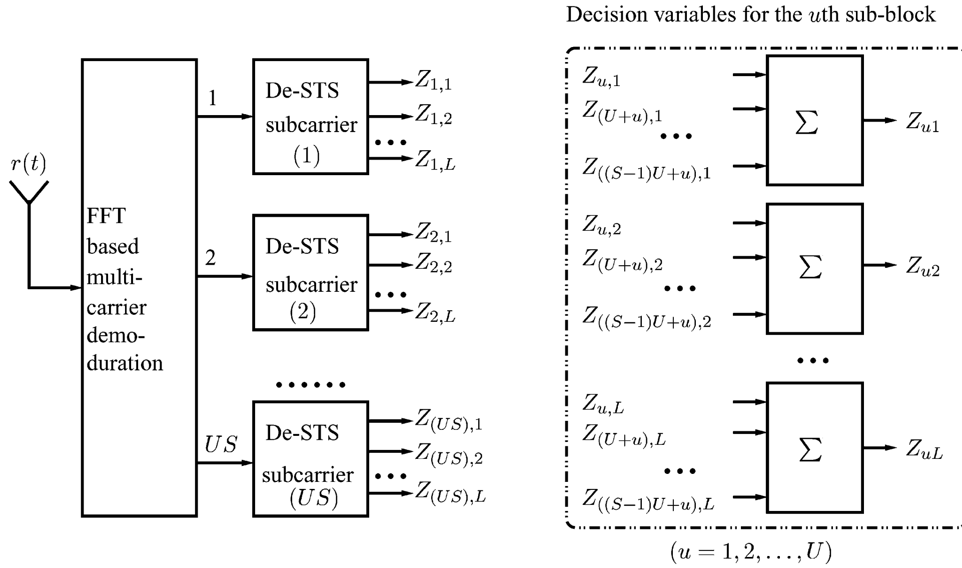

Fig. 3. The receiver schematic of the MC DS-CDMA system using space–time spreading.

efficiently in different dispersive environments having a delay spread in the range of $[T_m, T_M]$.

- In order to ensure that STS maintains the required frequency diversity order in *different communication environments*, we configure the system such that each subcarrier signal is guaranteed to experience flat fading. The required frequency diversity is attained by combining the independently faded subcarrier signals, with the aid of F-domain repetition. Since the delay spread experienced in different communication environments is assumed to be limited to the range of $[T_m, T_M]$, the flat-fading condition of each subcarrier in these different communication environments is satisfied, provided that $T_c > T_M$.
- In order to achieve the highest possible grade of frequency diversity for a given number of combined subcarrier signals, the subcarrier signals combined must experience independent fading. This implies that the F-domain spacing between the specific subcarriers that are combined must be higher than the maximum coherence bandwidth of $(\Delta f)_{cM} \approx 1/T_m$ [12]. Let U be the number of subblocks after the S-P conversion stage of Fig. 1. Then, according to Fig. 2 and (1), the above condition is satisfied, if $(U/T_c) \geq (1/T_m)$, i.e., $U \geq (T_c/T_m)$.

According to the above design philosophy, it can be shown that an MC DS-CDMA system having $T_c > T_M$ and $U \geq T_c/T_m$ is capable of achieving a constant frequency-selective diversity order, provided that the delay spread of the wireless channels encountered falls in the range of $[T_m, T_M]$.

Assuming that K user signals in the form of (2) are transmitted synchronously over Rayleigh-fading channels, the received complex low-pass equivalent signal can be expressed as

$$R(t) = \sum_{k=1}^{K}\sum_{g=1}^{T_x}\sqrt{\frac{2E_b}{UT_b}\frac{1}{SM_xT_x}}(\boldsymbol{D}_k\boldsymbol{P})_g\boldsymbol{H}\boldsymbol{w} + N(t), \quad (8)$$

where $(\boldsymbol{X})_g$ represents the gth row of the matrix $\boldsymbol{X}$, $N(t)$ is the complex valued low-pass-equivalent additive white Gaussian noise (AWGN) having a double-sided spectral density of N_0, while

$$\boldsymbol{H} = \text{diag}\{h_{1g}\exp(j\psi_{1g}), h_{2g}\exp(j\psi_{2g}), \ldots, \\ h_{(US)g}\exp(j\psi_{(US)g})\}, \qquad g = 1, 2, \ldots, T$$

is a diagonal matrix of rank US, which represents the channel's complex impulse response in the context of the gth antenna. The coefficients h_{ig}, $i = 1, 2, \ldots, US$; $g = 1, 2, \ldots, T$ in $\boldsymbol{H}$ are independent and identically distributed (i.i.d.) random variables obeying the Rayleigh distribution, which can be expressed as

$$f_{h_{ig}}(y) = \frac{2y}{\Omega}\exp\left(-\frac{y^2}{\Omega}\right), \qquad y \geq 0, \quad (9)$$

where $\Omega = \text{E}[(h_{ig})^2]$. Furthermore, the phases ψ_{ig}, $i = 1, 2, \ldots, US$; $g = 1, 2, \ldots, T$ are introduced by the fading channels and are uniformly distributed in the interval $[0, 2\pi)$.

Specifically, for the case of $L = M = T = 2$, the received complex low-pass equivalent signal can be expressed as

$$R(t) = \sum_{k=1}^{K}\sum_{u=1}^{U}\sum_{s=1}^{S}\sqrt{\frac{2E_b}{4UST_b}} \\ \times\big(h_{(us)1}\exp\big(j\psi_{(us)1}\big)[a_{k,1}(t)b_{k,u1} + a_{k,2}(t)b_{k,u2}] \\ + h_{(us)2}\exp\big(j\psi_{(us)2}\big)[a_{k,1}(t)b_{k,u2} - a_{k,2}(t)b_{k,u1}]\big) \\ \times\exp\left(j2\pi f_{(s-1)U+u}t\right) + N(t). \quad (10)$$

C. Receiver Model

Let the first user be the user-of-interest and consider a receiver employing fast Fourier transform (FFT)-based multicarrier demodulation, space–time despreading, as well as diversity combining, as shown in Fig. 3. The receiver of Fig. 3 essentially carries out the inverse operations of those seen in Fig. 1. In Fig. 3, the received signal is first demodulated using FFT-based multicarrier demodulation, obtaining US number of parallel streams corresponding to the signals transmitted on US subcarriers. Then, each stream is space–time despread using the approach of [4], in order to obtain L separate variables, $\{Z_{u,l}, Z_{u,2}, \ldots, Z_{u,L}\}_{u=1}^{US}$, corresponding to the L data bits transmitted on the uth stream, where $u = 1, 2, \ldots, US$, respectively. Following space–time despreading, a decision variable is formed for each of the transmitted data bits $\{b_{u1}, b_{u2}, \ldots, b_{uL}\}_{u=1}^{U}$, which can be expressed as

$$Z_{ui} = \sum_{s=1}^{S} Z_{((s-1)U+u),i}, \quad u = 1, 2, \ldots, U;\ i = 1, 2, \ldots, L. \tag{11}$$

Finally, the UL number of transmitted data bits can be decided based on the decision variables $\{Z_{ui}, u = 1, 2, \ldots, U; i = 1, 2, \ldots, L\}$ using the conventional decision rule of a BPSK scheme. Let us now investigate the achievable BER performance.

III. Bit-Error Rate Analysis

In this section, we derive the BER expression of the broadband MC DS-CDMA system using STS, which was described in Section II. As an example, we derive the BER expression in detail for STS-based MC DS-CDMA using the parameters of $L = M = T = 2$. The generalized BER expression of MC DS-CDMA using the set of attractive STS schemes, i.e., using $L = M = T = 2, 4, 8$, etc., is then derived from the case of $L = M = T = 2$ without providing the detailed derivations, since the extension is relatively straightforward.

For the case of $L = M = T = 2$, the analysis can be commenced from (10). Let $y_{u,1}$, $y_{u,2}$—where $u = 1, 2, \ldots, US$—represent the correlator's output variables corresponding to the first two data bits transmitted on the uth subcarrier, where

$$y_{u,1} = \int_0^{2UT_b} R(t)a_{1,1}(t) \exp(-j2\pi f_u t)\, dt \tag{12}$$

$$y_{u,2} = \int_0^{2UT_b} R(t)a_{1,2}(t) \exp(-j2\pi f_u t)\, dt. \tag{13}$$

Since orthogonal multicarrier signals, orthogonal STS codes, synchronous transmission of the K user signals, as well as slowly flat-fading of each subcarrier are assumed, there is no interference between the different users and the different subcarrier signals. Therefore, when substituting (10) into (12) and (13), we have

$$y_{u,1} = \sqrt{\frac{2UE_bT_b}{S}} [h_{u1} \exp(j\psi_{u1}) b_{1,u1} + h_{u2} \exp(j\psi_{u2}) b_{1,u2}] + N_{u,1} \tag{14}$$

$$y_{u,2} = \sqrt{\frac{2UE_bT_b}{S}} [h_{u1} \exp(j\psi_{u1}) b_{1,u2} - h_{u2} \exp(j\psi_{u2}) b_{1,u1}] + N_{u,2} \tag{15}$$

where $N_{u,i}$, $i = 1, 2$ is due to the AWGN expressed as

$$N_{u,i} = \int_0^{2UT_b} N(t)a_{1,i}(t) \exp(-j2\pi f_u t) dt$$

which is a complex Gaussian distributed variable having zero mean and a variance of $2UN_0T_b$.

Assuming that the receiver has perfect knowledge of the fading parameters of $h_{ui}\exp(j\psi_{ui})$,$i = 1, 2$, the decision variables corresponding to the data bits $b_{1,ui}$,$i = 1, 2$ associated with the uth subcarrier can be expressed as

$$\begin{aligned} Z_{u,1} &= \mathrm{Re}\{y_{u,1}h_{u1}\exp(-j\psi_{u1}) - y_{u,2}h_{u2}\exp(-j\psi_{u2})\} \\ &= \sqrt{\frac{2UE_bT_b}{S}} \left[h_{u1}^2 + h_{u2}^2\right] b_{1,u1} \\ &\quad + \mathrm{Re}\{N_{u,1}h_{u1}\exp(-j\psi_{u1}) - N_{u,2}h_{u2}\exp(-j\psi_{u2})\} \end{aligned} \tag{16}$$

$$\begin{aligned} Z_{u,2} &= \mathrm{Re}\{y_{u,1}h_{u2}\exp(-j\psi_{u2}) + y_{u,1}h_{u1}\exp(-j\psi_{u1})\} \\ &= \sqrt{\frac{2UE_bT_b}{S}} \left[h_{u1}^2 + h_{u2}^2\right] b_{1,u2} \\ &\quad + \mathrm{Re}\{N_{u,1}h_{u2}\exp(-j\psi_{u2}) + N_{u,2}h_{u1}\exp(-j\psi_{u1})\} \end{aligned} \tag{17}$$

for $u = 1, 2, \ldots, US$.

Finally, after combining the replicas of the same signal transmitted on the S subcarriers, the decision variables corresponding to the two bits in the uth subblock can be expressed as

$$Z_{u1} = \sum_{s=1}^{S} Z_{((s-1)U+u),1} \tag{18}$$

$$Z_{u2} = \sum_{s=1}^{S} Z_{((s-1)U+u),2} \tag{19}$$

for $u = 1, 2, \ldots, U$.

Since $Z_{((s-1)U+u),1}$ of (16) and $Z_{((s-1)U+u),2}$ of (17) are i.i.d. random variables for different s values associated with a mean of

$$\sqrt{(2UE_bT_b/S)}[h_{((s-1)U+u)1}^2 + h_{((s-1)U+u)2}^2]$$

and variance of

$$UN_0T_b[h_{((s-1)U+u)1}^2 + h_{((s-1)U+u)2}^2]$$

conditioned on $h_{((s-1)U+u)1}$ and $h_{((s-1)U+u)2}$, it is well recognized [4] that the conditional BER can be expressed as

$$P_b\left(E|\{h_{((s-1)U+u)1}, h_{((s-1)U+u)2}\}\right) = Q\left(\sqrt{2 \cdot \sum_{s=1}^{S}\sum_{l=1}^{2} \gamma_{sl}}\right) \tag{20}$$

where $Q(x)$ represents the Gaussian Q-function, $\gamma_{sl} = \overline{\gamma} \times ((h_{((s-1)U+u),l})^2/\Omega)$, and $\overline{\gamma} = E_b\Omega/SN_0$. Finally, with the aid of (9) and [12, p. 781], it can be readily shown that the average BER can be expressed as

$$P_b = \left[\frac{1-\mu}{2}\right]^{2S} \sum_{k=0}^{2S-1} \binom{2S-1+k}{k} \left[\frac{1+\mu}{2}\right]^k \tag{21}$$

where $\mu = \sqrt{\overline{\gamma}/(1+\overline{\gamma})}$.

For the general case of $L = M = T = 2, 4, 8$, etc., the decision variable Z_{u1} corresponding to the first bit in the subblock u can also be expressed as in (18), but with $Z_{((s-1)U+u),1}$ given by

$$Z_{((s-1)U+u),1} = \sqrt{\frac{2UE_bT_b}{S}} \left(\sum_{l=1}^{T} h_{((s-1)U+u)l}^2\right) b_{1,u1} + \mathrm{Re}\left\{N'_{((s-1)U+u),1}\right\}, \quad s = 1, 2, \ldots, S \tag{22}$$

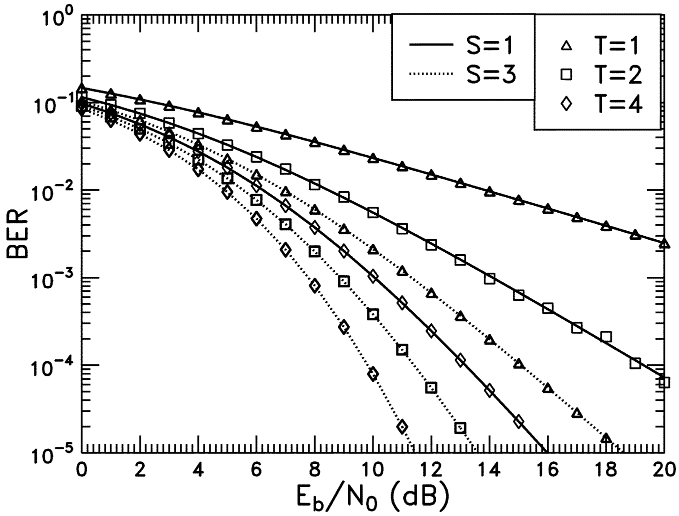

Fig. 4. Numerical (lines) and simulated (markers) BER versus the SNR per bit, E_b/N_0, performance for the broadband MC DS-CDMA using STS based transmit diversity, when communicating over frequency-selective Rayleigh-fading channels evaluated from (23).

where $\mathrm{Re}\{N'_{((s-1)U+u),1}\}$ is an AWGN process having a zero mean and a variance of $UN_0T_b\sum_{l=1}^{T} h^2_{((s-1)U+u)l}$. The average BER for the case of $L = M = T = 2, 4, 8$, etc., can be expressed as

$$P_b = \left[\frac{1-\mu}{2}\right]^{TS}\sum_{k=0}^{TS-1}\binom{TS-1+k}{k}\left[\frac{1+\mu}{2}\right]^k. \quad (23)$$

Note that according to (18) and (22), by using STS and F-domain subcarrier repetition, the diversity order achieved is TS, provided that T transmitter antennas and S-depth F-domain repetition schemes were used. The resultant diversity order of TS can also be seen in (23).

Fig. 4 shows both the numerical and simulation based BER results, which are drawn using lines and markers, respectively, for $S = 1$,$T = 1, 2, 4$, as well as for $S = 3$,$T = 1, 2, 4$. From the results we observe that at a BER of 0.01, using two transmitter antennas rather than one yields a gain of approximately 5.0 dB. Furthermore, when $T = 4$ transmitter antennas and an repetition depth of $S = 3$ are considered instead of $T = 1$, $S = 1$, the diversity gain achieved is approximately 9.0 dB.

The BER performance of Fig. 4 can be achieved, provided that the number of orthogonal STS codes is sufficiently high for supporting the K number of users without reuse. Based on our arguments in Sections II and III, the number of orthogonal STS codes is given by $ULT_b/T_c = ULN$. Since each user requires M orthogonal STS codes, the maximum number of users supported by the ULN orthogonal STS codes is given by $\mathcal{K}_{\max} = ULN/M$. In other words, if the number of users supported by the synchronous broadband MC DS-CDMA system designed according to the philosophy of this paper obeys $K \leq \mathcal{K}_{\max}$, the BER performance of Fig. 4 can be achieved for transmissions over frequency-selective Rayleigh-fading channels. For the attractive STS schemes using $L = M = T = 2, 4, 8$, etc., the maximum number of users supported by the orthogonal STS codes is $\mathcal{K}_{\max} = UN$. It can be seen that in this case the number of users supported is independent of the number of transmitter antennas, which emphasizes the advantages of the STS schemes using the specific values of $L = M = T = 2, 4, 8$, etc. [4].

Let $N = T_b/T_c$ and the total number of subcarriers US be constants. Then, the maximum number of users supported by the broadband MC DS-CDMA system depends only on the repetition depth S. Specifically, the maximum number of users supported decreases, when increasing the subcarrier repetition depth S, i.e. when increasing the frequency diversity order. Consequently, the maximum number of users supported and the frequency diversity gain achieved have to obey a tradeoff. This is not a desirable result. We would like to achieve the maximum transmit diversity gain as well as the required frequency diversity gain without having to accept any other tradeoff, i.e., without decreasing the total number of users supported by the system. Let us now consider this issue in the next section.

IV. Capacity Extension Using Time–Frequency-Domain Spreading

A. System Description

In this section, we propose and investigate an MC DS-CDMA, which employs both time (T)-domain and frequency (F)-domain spreading, i.e., it employs TF-domain spreading. The transmitter schematic of the STS-assisted broadband MC DS-CDMA using TF-domain spreading is similar to that seen in Fig. 1, except that the S-depth subcarrier repetition of Fig. 1 is now replaced by the F-domain spreading associated with an orthogonal spreading code of length S. Specifically, let $\{c_k[0], c_k[1], \ldots, c_k[S-1]\}$ be the kth user's orthogonal code in discrete form, which will be used for the F-domain spreading. By contrast, the kth user's T-domain orthogonal codes used for STS have been expressed in Section II as $a_{k,i}(t)$, $i = 1, 2, \ldots, M$ in continuous form. In the broadband MC DS-CDMA system using TF-domain spreading, the U subblock signals of user k after STS are now further spread over the F-domain using the above F-domain spreading codes. The signals transmitted from the T transmitter antennas can be expressed as

$$\boldsymbol{s}_k(t) = \mathrm{Re}\left\{\sqrt{\frac{2E_b}{UT_b}\frac{1}{SMT}}\boldsymbol{D}_k\boldsymbol{Q}\boldsymbol{w}\right\} \quad (24)$$

where all variables have the same interpretations as those in (2), except that the repetition matrix $\boldsymbol{P}$ of (2) is now replaced by $\boldsymbol{Q}$ in (24), which represents the F-domain spreading and is a $U \times US$-dimensional matrix expressed as $\boldsymbol{Q} = [\boldsymbol{C}_k[0]\ \boldsymbol{C}_k[1]\ \ldots\ \boldsymbol{C}_k[S-1]]$, where $\boldsymbol{C}_k[s]$,$s = 0, 1, \ldots, S-1$ are diagonal matrices of rank U, which can be expressed as $\boldsymbol{C}_k[s] = \mathrm{diag}\{c_k[s], c_k[s], \ldots, c_k[s]\}$.

Specifically, for $L = M = T = 2$, the TF-domain spread MC DS-CDMA signals transmitted by antenna 1 and 2 can be simply expressed as

$$\boldsymbol{s}_k(t) = \begin{pmatrix} s_{k1}(t) \\ s_{k2}(t) \end{pmatrix} = \sqrt{\frac{2E_b}{4UST_b}} \begin{pmatrix} \sum_{u=1}^{U}\sum_{s=1}^{S}[a_{k,1}b_{k,u1}+a_{k,2}b_{k,u2}] \\ \times c_k[s-1]\cos\left(2\pi f_{(s-1)U+u}t\right) \\ \sum_{u=1}^{U}\sum_{s=1}^{S}[a_{k,1}b_{k,u2}-a_{k,2}b_{k,u1}] \\ \times c_k[s-1]\cos\left(2\pi f_{(s-1)U+u}t\right) \end{pmatrix}. \quad (25)$$

The total number of users supported by the broadband MC DS-CDMA system using TF-domain spreading and the assignment of orthogonal codes to the users is analyzed as follows. According to our analysis in Sections II and III, the total number of orthogonal codes that can be used for STS is ULN and the maximum number of users supported by these orthogonal codes is $\mathcal{K}_{\max} = ULN/M$. By contrast, the total number of orthogonal codes that can be used for F-domain spreading is S. This implies that even if S number of users share the same set of STS codes, these S user signals might be distinguishable with the aid of the associated S number of F-domain spreading codes. Explicitly, the total number of users supported is $S\mathcal{K}_{\max} = USLN/M$. Therefore, the orthogonal spreading codes can be assigned as follows. If the number of users is in the range of $0 \leq K \leq \mathcal{K}_{\max}$, these users will be assigned the required orthogonal STS codes and the same F-domain orthogonal spreading code. However, when the number of users is in the range of $s\mathcal{K}_{\max} \leq K \leq (s+1)\mathcal{K}_{\max}, s = 1, 2 \ldots, S-1$, then the same set of M STS orthogonal codes must be assigned to s or $(s+1)$ users, but these s or $(s+1)$ users are assigned different F-domain spreading codes. These s or $(s+1)$ users employing the same set of M STS codes are identified by their corresponding F-domain spreading codes. Since the subcarrier signals across which F-domain spreading takes place encounter independent fading, the orthogonality of the F-domain spreading codes cannot be retained. Hence, MUI is inevitably introduced, which degrades the BER performance, when increasing the number of users sharing the same set of M STS orthogonal codes.

Let $1 \leq K' \leq S$ be the number of users sharing the same set of M STS orthogonal codes. We also assume that any set of M STS orthogonal codes is shared by the same K' number of users. Then, when the $K'\mathcal{K}_{\max}$ signals expressed in the form of (24) are transmitted over frequency-selective fading channels, the received complex low-pass equivalent signal can be expressed as

$$R(t) = \sum_{k=1}^{K'\mathcal{K}_{\max}} \sum_{g=1}^{T} \sqrt{\frac{2E_b}{UT_b}\frac{1}{SMT}} (\boldsymbol{D}_k\boldsymbol{Q})_g \boldsymbol{Hw} + N(t) \quad (26)$$

where $(X)_g$ and $N(t)$ have the same interpretation as in (8). Specifically, for the case of $L = M = T = 2$, $R(t)$ can be expressed as

$$\begin{aligned} R(t) = & \sum_{k=1}^{K'\mathcal{K}_{\max}} \sum_{u=1}^{U}\sum_{s=1}^{S} \sqrt{\frac{2E_b}{4UST_b}} \\ & \times \left(h_{(us)1} \exp\left(j\psi_{(us)1}\right) [a_{k,1}(t)b_{k,u1} + a_{k,2}(t)b_{k,u2}] \right. \\ & \quad \times c_k[s-1] + h_{(us)2} \exp\left(j\psi_{(us)2}\right) \\ & \quad \left. \times \, [a_{k,1}(t)b_{k,u2} - a_{k,2}(t)b_{k,u1}]\, c_k[s-1]\right) \\ & \times \exp\left(j2\pi f_{(s-1)U+u}t\right) + N(t). \end{aligned} \quad (27)$$

The receiver schematic of the MC DS-CDMA scheme using TF-domain spreading is similar to that of Fig. 3, except that the de-repetition operation seen in Fig. 3 is now replaced by the F-domain despreading. The signals at the output of the De-STS' block of Fig. 3 can be detected by invoking a range of single- or multiuser detection schemes.

B. Signal Detection

Following the derivations in Section III, for the general case of $L = M = T = 2, 4, 8$, etc., the decision variable $Z_{((s-1)U+u),1}$ in terms of the first data bit in the subblock u and the subcarrier $(s-1)U + u$ can now be expressed as

$$\begin{aligned} Z_{((s-1)U+u),1} = & \sqrt{\frac{2UE_bT_b}{S}} \sum_{k=1}^{K'} \left(\sum_{l=1}^{T} h^2_{((s-1)U+u)l}\right) \\ & \times c_k[s-1]b_{k,u1} + \mathrm{Re}\left\{N'_{((s-1)U+u),1}\right\}, \\ & s = 1, 2, \ldots, S \end{aligned} \quad (28)$$

where $\mathrm{Re}\{N'_{u,1}\}$ is an AWGN process having zero mean and a variance of $UN_0T_b\sum_{l=1}^{T} h^2_{ul}$.

Let

$$\boldsymbol{z}_{u1} = \left[Z_{u,1}\ Z_{(U+u),1}\ \ldots\ Z_{((S-1)U+u),1}\right]^T \quad (29)$$

$$\boldsymbol{A} = \mathrm{diag}\left\{\sum_{l=1}^{T_x} h^2_{ul},\ \sum_{l=1}^{T_x} h^2_{(U+u)l},\ \ldots,\ \sum_{l=1}^{T_x} h^2_{((S-1)U+u)l}\right\} \quad (30)$$

$$\boldsymbol{C} = \begin{pmatrix} c_1[0] & c_2[0] & \ldots & c_{K'}[0] \\ c_1[1] & c_2[1] & \ldots & c_{K'}[1] \\ \vdots & \vdots & \ddots & \vdots \\ c_1[S-1] & c_2[S-1] & \ldots & c_{K'}[S-1] \end{pmatrix} \quad (31)$$

$$\boldsymbol{b} = [b_{1,u1}b_{2,u1}\ldots b_{K',u1}]^T \quad (32)$$

$$\boldsymbol{n} = \mathrm{Re}\left[N'_{u,1}N'_{(U+u),1}\cdots N_{((S-1)U+u),1}\right]^T. \quad (33)$$

Then (28) can be written in a matrix form as

$$\boldsymbol{z}_{u1} = \sqrt{\frac{2UE_bT_b}{S}}\boldsymbol{ACb} + \boldsymbol{n}. \quad (34)$$

Detection of the MC DS-CDMA signals of (34) is similar to the detection of the conventional MC-CDMA signals using solely F-domain spreading, where single-user detectors [5], [13] or multiuser detectors [5], [14], [15] can be employed. In this paper, as examples, we investigate two detection algorithms, namely, the single-user correlation detector and the multiuser decorrelating detector [5], [11]. In the context of the correlation detector, let $\boldsymbol{z} = [Z_{u1}\ Z_{u2}\ \ldots\ Z_{uK'}]^T$ represent the decision variables. Then, these decision variables are obtained by multiplying both sides of (34) with $\boldsymbol{C}^T$, which can be expressed as

$$\boldsymbol{z} = \sqrt{\frac{2UE_bT_b}{S}}\left(\sum_{s=1}^{S}\sum_{l=1}^{T_x} h^2_{((s-1)U+u)l}\right)\boldsymbol{Rb} + \boldsymbol{C}^T\boldsymbol{n} \quad (35)$$

where

$$\boldsymbol{R} = \begin{pmatrix} 1 & \rho_{12} & \ldots & \rho_{1K'} \\ \rho_{21} & 1 & \ldots & \rho_{2K'} \\ \vdots & \vdots & \ddots & \vdots \\ \rho_{K'1} & \rho_{K'2} & \ldots & 1 \end{pmatrix} \quad (36)$$

is the correlation matrix among the K' user signals, while ρ_{ij} represents the correlation factor between user i and user j, which can be expressed as

$$\rho_{ij} = \frac{\sum_{s=1}^{S}\left(c_i[s-1]c_j[s-1]\sum_{l=1}^{T} h^2_{((s-1)U+u)l}\right)}{\sum_{s=1}^{S}\sum_{l=1}^{T} h^2_{((s-1)U+u)l}}. \quad (37)$$

Equation (35) suggests that the diversity gain contributed both by the transmit diversity and frequency diversity can be retained, since we have a double sum of the components h^2 corresponding to the transmit and frequency diversity orders of S and T, respectively. However, MUI is introduced by the channel's time-varying characteristics. Finally, the corresponding data bits, $b_{k,u1}$, $k = 1, 2, \ldots, K'$, are decided according to $\hat{b}_{k,u1} = \text{sgn}((\boldsymbol{z})_k)$ for $k = 1, 2, \ldots, K'$, where $(\boldsymbol{z})_k$ represents the kth row of $\boldsymbol{z}$, while $\text{sgn}(\cdot)$ is the sign function [11].

Note that the correlation factors $\{\rho_{ij}\}$ in (37) are time-variant due to the time-varying nature of the channel's fading envelope. However, since $\sum_{s=1}^{S} c_i[s-1]c_j[s-1] = 0$, it can be shown that $\rho_{ij} = 0$, provided that the sum of $\sum_{l=1}^{T} h^2_{((s-1)U+u)l}$ is identical for different values of s. Moreover, it can be shown that the correlation factors $\{\rho_{ij}\}$ are contributed by the differences of the sums $\sum_{l=1}^{T} h^2_{((s-1)U+u)l}$ experienced according to the different values of s, while the common part of $\sum_{l=1}^{T} h^2_{((s-1)U+u)l}$ in terms of the different values of s can be successfully removed due to the orthogonality of the F-domain spreading codes. Specifically, let

$$\sum_{l=1}^{T} h^2_{((s-1)U+u)l} = A_h + \Delta_s$$

where A_h represents the average value of $\sum_{l=1}^{T} h^2_{((s-1)U+u)l}$ in terms of s, while

$$\Delta_s = \sum_{l=1}^{T} h^2_{((s-1)U+u)l} - A_h.$$

Then, (37) can be written as

$$\rho_{ij} = \frac{\sum_{s=1}^{S} (c_i[s-1]c_j[s-1]\Delta_s)}{\sum_{s=1}^{S} (A_h + \Delta_s)}. \tag{38}$$

Fig. 5 shows the probability density function (pdf) of the correlation factor ρ_{12} between the signals of user 1 and user 2 for a two-user MC DS-CDMA system employing STS-based transmit diversity, when communicating over frequency-selective Rayleigh-fading channels. The curves show that the correlation factor using $T = 1, 2.$ and 4 transmitter antennas is symmetrically distributed around $\rho_{12} = 0$. An important observation is that the correlation factor's value is predominantly distributed in the vicinity of $\rho_{12} = 0$ and becomes similar to a truncated Gaussian random variable distributed within $[-1, 1]$ having a relatively low variance, when increasing the number of transmitter antennas. **This observation implies that STS using several transmitter antennas has two-fold importance. First, it is capable of providing spatial diversity. Second, it is capable of suppressing the MUI imposed by the F-domain despreading**.

In the context of the decorrelating detector, the decision variables associated with $\{b_{k,u1}\}_{k=1}^{K'}$ are obtained by multiplying both sides of (35) with the inverse of $\boldsymbol{R}$, i.e., with $\boldsymbol{R}^{-1}$, which can be expressed as

$$\boldsymbol{R}^{-1}\boldsymbol{z} = \sqrt{\frac{2UE_bT_b}{S}} \left(\sum_{s=1}^{S} \sum_{l=1}^{T} h^2_{((s-1)U+u)l} \right) \boldsymbol{b} + \boldsymbol{R}^{-1}\boldsymbol{C}^T\boldsymbol{n} \tag{39}$$

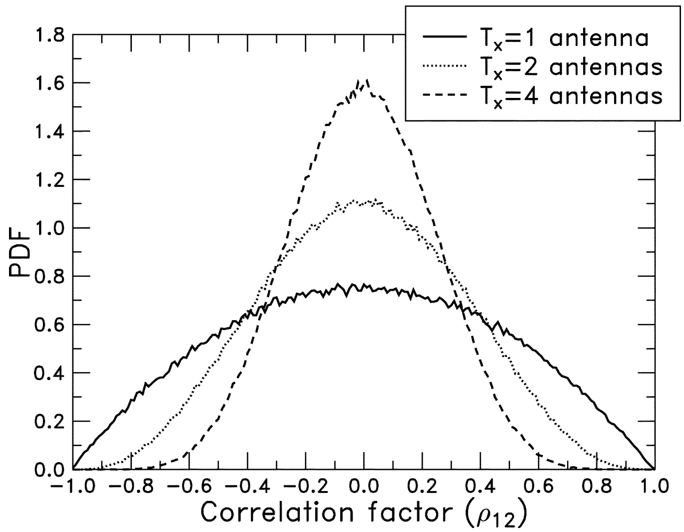

Fig. 5. Simulated pdf of the correlation factor ρ_{12} for a two-user MC DS-CDMA using $T = 1, 2$, or 4 transmitter antennas and STS, when communicating over frequency-selective Rayleigh-fading channels.

and the corresponding data bits $b_{k,u1}$, $k = 1, 2, \ldots, K'$ are decided according to $\hat{b}_{k,u1} = \text{sgn}((\boldsymbol{R}^{-1}\boldsymbol{z})_k)$, $k = 1, 2, \ldots, K'$. Equation (39) shows that each user's data can be decided independently of the other users' data and the diversity order achieved is TS. Let us now provide a range of simulation results for the STS-assisted broadband MC DS-CDMA system using TF-domain spreading.

C. BER Performance

The BER versus SNR per bit, E_b/N_0, performance of both the correlation detector and that of the decorrelating detector is shownin Figs. 6 and 7 for the TF-domain spread broadband MC DS-CDMA systems. In both figures, we considered $T = 2$ transmitter antennas and supporting $K = \mathcal{K}_{\max}$, $2\mathcal{K}_{\max}$, $3\mathcal{K}_{\max}$, and $4\mathcal{K}_{\max}$ users corresponding to $K' = 1, 2, 3, 4$, where $\mathcal{K}_{\max}$ represented the maximum number of users supported by the T-domain orthogonal spreading codes without imposing MUI. The difference between Fig. 6 and Fig. 7 is that in Fig. 6, the length of the F-domain spreading codes was $S = 4$, while in Fig. 7, it was $S = 8$. As expected, we observe in both figures that the BER performance is significantly improved, when the correlation detector is replaced by the decorrelating detector. For both the correlation detector and the decorrelating detector, the BER performance degrades, when increasing the number of users sharing the same T-domain spreading code, i.e., when increasing the value of K'. However, the BER increase due to increasing the value of K' is significantly lower for the decorrelating detector, than that of the correlation detector. Furthermore, upon comparing Fig. 6 to Fig. 7, we observe that the BER performance of the decorrelating detector is closer to the BER performance without MUI, when using $S = 8$ (Fig. 7) instead of $S = 4$ (Fig. 6).

In Fig. 8 we investigated the BER performance of both the correlation detector and that of the decorrelating detector for various numbers of transmitter antennas, namely, for $T = 1, 2, 4$ and for $S = 8, 4, 2$ F-domain spreading codes, while maintaining a constant TS value of eight. In our experiments, we assumed that $K' = 2$, i.e., that each set of T-domain STS

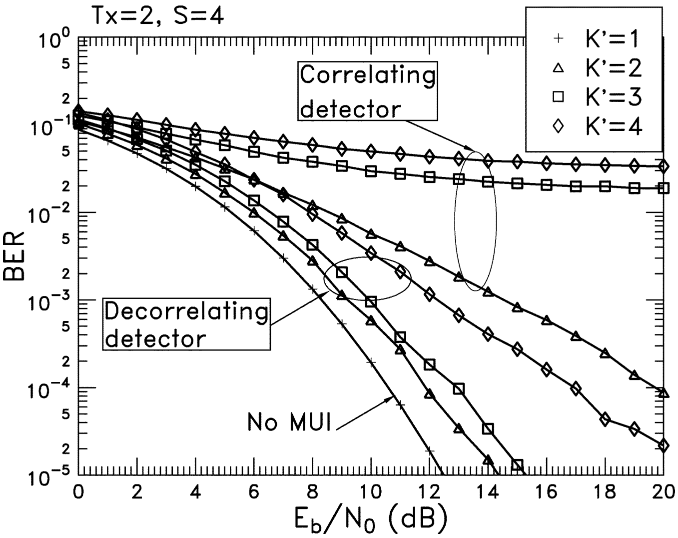

Fig. 6. Simulation-based BER versus the SNR per bit, E_b/N_0, performance of both the correlation detector and decorrelating detectors for STS-based broadband MC DS-CDMA using TF-domain spreading, when communicating over frequency-selective Rayleigh fading channels.

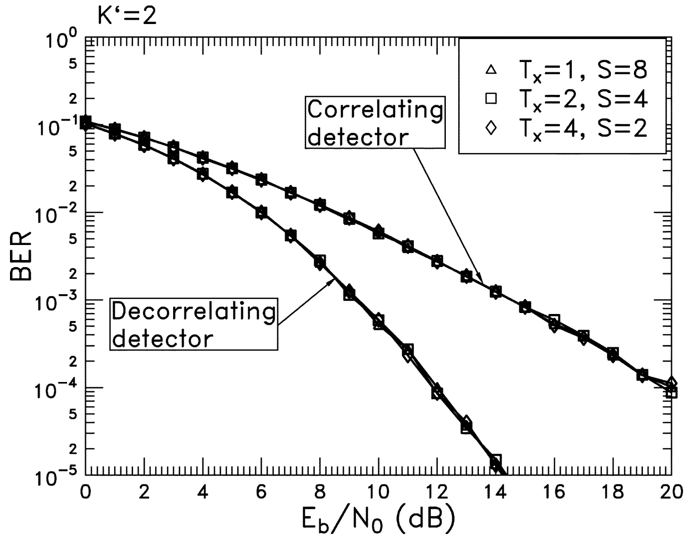

Fig. 8. Simulation-based BER versus the SNR per bit, E_b/N_0, performance of both the correlation detector and decorrelating detector for STS based broadband MC DS-CDMA using TF-domain spreading, when communicating over frequency-selective Rayleigh-fading channels.

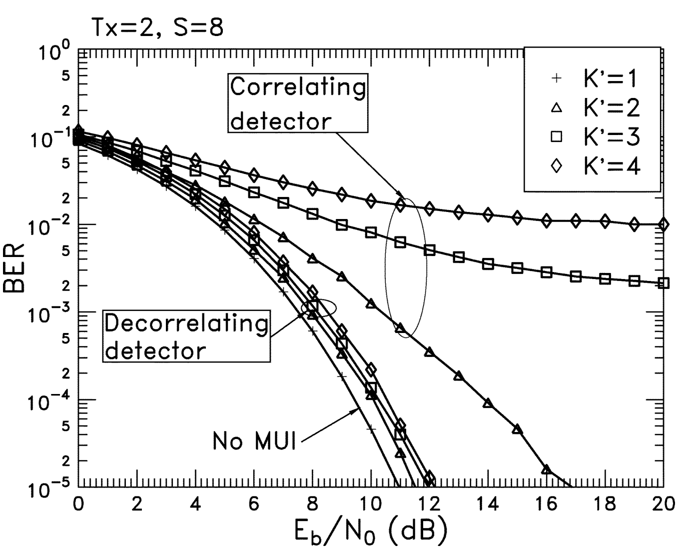

Fig. 7. Simulation-based BER versus the SNR per bit, E_b/N_0, performance of both the correlation detector and decorrelating detector for STS-based broadband MC DS-CDMA using TF-domain spreading, when communicating over frequency-selective Rayleigh-fading channels.

codes was shared by two users. Let the maximum number of users supported by the T-domain codes be $\mathcal{K}_{\max}$, while using the parameters of $(T = 4, S = 2)$. Then, for a broadband MC DS-CDMA system having a constant system bandwidth, the maximum number of users, $\mathcal{K}_{\max}$ supported by the T-domain codes and using the parameters of $(T = 1, S = 8)$ or $(T = 2, S = 4)$ is $\mathcal{K}_{\max}/4$ or $\mathcal{K}_{\max}/2$, respectively. In other words, there is a maximum of two, four, or eight users sharing the same set of orthogonal STS codes, corresponding to the cases of $(T = 4, S = 2)$, $(T = 2, S = 4)$, or $(T = 1, S = 8)$, respectively. From the results we infer the following observa-tions. a) All schemes achieve the same total diversity gain. b) The number of transmitter antennas has the same effect on the BER performance as the length of the F-domain spreading codes, i.e., the same BER can be maintained, regardless of what values T and S take, provided that the product TS remains a constant. c) The decorrelating detector significantly outperforms the correlation detector. The gain achieved at the BER of 10^{-3} by using multiuser detection instead of single-user detection is about 5–6 dB. d) The maximum number of users $K'_{\max}$ sharing the same set of STS orthogonal codes is two, four, or eight, when we use the parameters $(T = 4, S = 2)$, $(T = 2, S = 4)$, or $(T = 1, S = 8)$. Furthermore, since, according to Figs. 6 and 7, the BER performance degrades upon increasing the number of users sharing the same set of STS orthogonal codes, consequently, for a fully loaded system using the maximum values of $K'_{\max}$, we can surmise that MC DS-CDMA employing $(T = 4, S = 2)$ outperforms the scheme using the parameter combinations of both $(T = 2, S = 4)$ and $(T = 1, S = 8)$. Furthermore, the MC DS-CDMA system using the parameters $(T = 2, S = 4)$ outperforms that employing $(T = 1, S = 8)$. The above arguments suggest that **the best broadband MC DS-CDMA system will only use transmit diversity and no frequency diversity at all, i.e., use the parameters** $(T = 8, S = 1)$, **which simultaneously suggests that no multiuser detection is required**.

V. Conclusion

In this contribution, we have investigated the performance of a broadband MC DS-CDMA using STS-assisted transmit diversity, when frequency-selective Rayleigh-fading channels are considered. The issue of parameter design has been investigated motivated by the objective of ensuring that the same broadband MC DS-CDMA scheme is equipped to provide efficient communications in various fading channels having different grade of frequency selectivity. The BER performance of the broadband MC DS-CDMA system using STS has been evaluated both analytically and by simulation. Furthermore, we have considered the capacity extension achievable by STS assisted broadband MC DS-CDMA with the aid of TF-domain spreading. The corresponding BER performance has been investigated in the context of the single-user correlation detector and the decorrelating multiuser detector for transmissions over frequency-selective Rayleigh-fading channels. In summary, the broadband MC DS-CDMA using STS-assisted transmit diversity has the following characteristics.

- By appropriately selecting the system parameters, the same broadband MC DS-CDMA system using STS-assisted transmit diversity is rendered capable of achieving a similar BER performance in various communication environments characterized by different grades of frequency selectivity.
- It is capable of mitigating the peak-to-average power fluctuation experienced, since with the advent of DS spreading of the subcarriers we require only a decreased number of subcarriers.
- Using multiple transmit antennas in MC DS-CDMA, the system becomes capable of providing transmit diversity and, simultaneously, suppressing the multiuser interference. The associated STS-based transmit diversity scheme used in MC DS-CDMA can be designed for maintaining a constant diversity gain in various fading channels having a different grade of frequency selectivity.
- When the STS-assisted broadband MC DS-CDMA scheme employs both T-domain and F-domain spreading, the maximum number of users supported becomes significantly higher than that employing solely T-domain spreading. The higher number of users supported is achieved without any tradeoffs imposed on the achievable diversity order.

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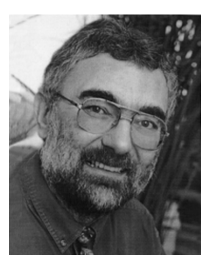
Lie-Liang Yang (M'98–SM'02) received the B.Eng. degree in communication engineering from Shanghai TieDao University, Shanghai, China, in 1988 and the M.S. and Ph.D. degrees in communications and electronics from Northern Jiaotong University, Beijing, China in 1991 and 1997, respectively.

From June 1997 to December 1997, he was a Visiting Scientist at the Institute of Radio Engineering and Electronics, Academy of Sciences of the Czech Republic, Prague. Since December 1997, he has been with the Communications Research Group, School of Electronics and Computer Science, University of Southampton, Southampton, U.K, where he is a Member of the Academic Staff. His research has covered a wide range of areas in telecommunications, which include error control coding, modulation and demodulation, spread-spectrum communications and multiuser detection, synchronization, smart antennas, adaptive wireless systems, as well as wideband, broadband, and ultrawideband code-division multiple access (CDMA). He has published over 90 papers as the first author in journals and conference proceedings, coauthored one book, and published several book chapters.

Dr. Yang was awarded the Royal Society Sino-British Fellowship in 1997 and the EPSRC Research Fellowship in 1998.



Lajos Hanzo (M'91–SM'92–F'04) received the master degree in electronics in 1976 and the doctorate in 1983.

During his career in telecommunications, he has held various research and academic posts in Hungary, Germany, and the U.K. Since 1986, he has been with the Department of Electronics and Computer Science, University of Southampton, Southampton, UK, where he holds the chair in telecommunications. He coauthored 11 books totalling 8000 pages on mobile radio communications, published in excess of 500 research papers, organized and chaired conference sessions, presented overview lectures, and has been awarded a number of distinctions. Currently he heads an academic research team, working on a range of research projects in the field of wireless multimedia communications sponsored by industry, the Engineering and Physical Sciences Research Council (EPSRC) U.K., the European IST Programme, and the Mobile Virtual Centre of Excellence (VCE), U.K. He is an enthusiastic supporter of industrial and academic liaison and he offers a range of industrial courses. For further information on research in progress and associated publications please refer to http://www-mobile.ecs.soton.ac.uk

Prof. Hanzo is an IEEE Distinguished Lecturer of both the Communications as well as the Vehicular Technology Societies and a Fellow of the IEE.