Downlink Space-Time Spreading Using Interference Rejection Codes

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Abstract

In this paper, we will investigate the performance of a Loosely Synchronized LS code based space-time spreading (STS) scheme in comparison to that of classic Walsh code and pseudo-noise code based STS, when communicating over dispersive Nakagami-m multipath channels. Closed-form formulas are derived for characterizing the BER performance as a function of the number of resolvable paths L, and the number of users K. Our numerical results suggest that the employment of LS code based STS scheme is beneficial in a low-user-load and low-dispersion channel scenario, where a near-single-user performance can be achieved without a multiuser detector.

Index Terms

Code-division multiple-access, interference free window, LA codes, LS codes, Gaussian Approximation, Nakagami-m fading.

I. Introduction

The underlying philosophy of Space Time Spreading (STS) is reminiscent the operating principles Space-Time Coding (STC) [1], where multiple replicas of the same symbol are mapped to multiple transmit antennas for the sake of achieving a transmit diversity gain. In the context of STS the information to be transmitted may be mapped to multiple transmitter antennas with the aid of the STS codes proposed in [2], which were graphically illustrated in [3]. In simple terms the STS codes are used for spatially spreading the information to multiple transmit antennas, again, for the sake of achieving spatial diversity.

When we consider a Space-Time Spreading (STS) assisted DS-CDMA scheme communicating over a non-dispersive channel, the employment of orthogonal spreading codes such as Walsh codes and orthogonal Gold codes [3] is ideal for the non-dispersive synchronous downlink (DL) channel, since the channel will not destroy the orthogonality of the codes, when we invoke a matched filter based RAKE receiver at the receiver side. However, classic orthogonal codes – such as for example Walsh codes – will lose their orthogonality, when communicating over a dispersive multipath channel. More specifically, when the RAKE receiver coherently combines the different paths' energy, it will inevitably combine both the multiple access interference (MAI) as well as the multipath interference (MPI) in case of communicating over a dispersive multipath channel. In order to circumvent this problem, the family of Loosely Synchronized (LS) codes [4], [5] has been proposed, which exhibits a so-called Interference Free Window (IFW). More explicitly, these codes are capable of suppressing both the MAI and MPI, provided that these interfering components arrive within the IFW. More specifically, when the dispersive channel's delay spread does not exceed the width of the IFW, we can combine all the paths' energy without imposing any MAI and MPI interference, and hence interference-free CDMA communication becomes possible without the employment of high-complexity multiuser detection. Furthermore, we will demonstrate that even when the channel's delay spread exceeds the width of the IFW, the proposed LS code based STS scheme is capable of outperforming the conventional STS scheme. However, the disadvantage of the LS code based STS scheme advocated is that the number of available LS codes is limited, when aiming for a specific spreading gain G. More explicitly, the number of supported users and the width of the IFW ι must satisfy $K(\iota + 1) \leq G$ [4]. To expound a little further, we can achieve a high IFW width and suppress the interference more effectively, when the number of users supported in the channel is relatively low, because the number of codes exhibiting a high IFW is low. By contrast, as the number of users increases, the IFW width tends to zero, since all the codes having a wide IFW have been activated, and hence the LS code based STS scheme becomes incapable of suppressing the MAI and MPI. Therefore, in this paper we will investigate the performance of an LS code based STS scheme in comparison to that of the STS scheme of [2], when communicating over dispersive Nakagami-m multipath channels. Since LS codes were described in [4], [5], while the philosophy of STS was detailed in [2], [3], here we refrain from their detailed description.

This paper is organized follows. Section II describes the system model used, while Section III illustrates the detection of STS signals. Section IV characterizes the achievable BER performance, while Section V discusses our numerical results. Finally, Section VI offers our conclusions.

A. *Loosely Synchronized Codes*

Loosely Synchronized (LS) codes [4] exploit the properties of the so-called orthogonal complementary sets [4], [6]. To expound further, let us introduce the notation of LS(N, P, W_0) for denoting the family of

LS codes generated by applying a $(P \times P)$-dimensional Walsh-Hadamard (WH) matrix to an orthogonal complementary code set of length N, as it is exemplified in the context of Figure 1. More specifically, we generate a complementary code pair by inserting W_0 number of zeros both in the center and at the beginning of the complementary pair, as shown in Figure 1(a), using the procedure described in [4]. As mentioned above, the polarity of the codes c_0 and s_0 seen in Figure 1(b) during the constitution of the LS codes is determined by the polarity of the components of a Walsh-Hadamard matrix, namely by $(1, 1, 1, 1)$ and $(1, -1, 1, -1)$. Then, the total length of the LS(N, P, W_0) code is given by $L_S = NP + 2W_0$ and later we will demonstrate that the total number of codes available is given by $4P$. The number of these codes having an IFW of W_0 chips is P, which limits the number of users that can be supported without imposing multiuser interference. Hence the number of codes having as long an IFW as possible has to be maximized for a given code length of $L_S = NP + 2W_0$.

Since the construction method of binary LS codes was described in [4], here we refrain from providing an indepth discourse and we will focus our attention on the employment of orthogonal complementary sets [7], [8] for the generation of LS codes.

For a given complementary code pair $\{\mathbf{c}_0, \mathbf{s}_0\}$ of length N, one of the corresponding so-called mate pairs can be written as $\{\mathbf{c}_1, \mathbf{s}_1\}$, where we have:

$$\mathbf{c}_1 = \tilde{\mathbf{s}}_0^*, \tag{1}$$

$$\mathbf{s}_1 = -\tilde{\mathbf{c}}_0^*, \tag{2}$$

and where the superscript * represents conjugation and $\tilde{\mathbf{s}}_0$ denotes the reverse-ordered sequence, while $-\mathbf{s}_0$ is the negated version of $\mathbf{s}_0$, respectively. Note that in Eq.(1) and Eq.(2) additional complex conjugation of the polyphase complementary sequences $\{\mathbf{c}_0, \mathbf{s}_0\}$ is required for deriving the corresponding mate pair $\{\mathbf{c}_1, \mathbf{s}_1\}$ in comparison to binary complementary sequences [4]. Having obtained a complementary pair and its corresponding mate pair, we may employ the construction method of [4] for generating a whole family of LS codes. The LS codes generated exhibit an IFW of length W_0. Hence, we may adopt the choice of $W_0 = N - 1$ in order to minimize the total length of the LS codes generated, while providing as long an IFW as possible.

For example, the LS(N, P, W_0)=LS(4,4,3) codes can be generated based on the complementary pair of [7]:

$$\mathbf{c}_0 = + + + - \quad (3)$$

$$\mathbf{s}_0 = + + - + \,. \quad (4)$$

Upon substituting Eq.(1) and Eq.(2) into both Eq.(3) and Eq.(4), the corresponding mate pair can be obtained as:

$$\mathbf{c}_1 = \tilde{\mathbf{s}}_0^* = + - + + \quad (5)$$

$$\mathbf{s}_1 = -\tilde{\mathbf{c}}_0^* = + - - - \,. \quad (6)$$

The first set of four LS codes can be generated using the first two rows of a $(P \times P) = (4 \times 4)$-dimensional Walsh-Hadamard matrix, namely using $\mathbf{w}_0 = (+1, +1, +1, +1)$ and $\mathbf{w}_1 = (+1, -1, +1, -1)$, as shown in Figure 1(b). Another set of four LS codes can be obtained by exchanging the subscripts 0 and 1. Finally, eight additional LS codes can be generated by applying the same principle, but with the aid of the last two rows of the (4×4)-dimensional Walsh-Hadamard matrix, namely using $\mathbf{w}_2 = (+1, +1, -1, -1)$ and $\mathbf{w}_3 = (+1, -1, -1, +1)$. Hence, the total number of available codes in the family of LS(N, P, W_0) is given by $4P$. More explicitly, there are four sets of P number of LS codes. Each set has four LS codes, and the LS codes in the same set exhibit an IFW length of $[-\iota, +\iota]$, where we have $\iota = \min\{W_0, N-1\}$. The aperiodic auto-correlation and cross-correlation function $\rho_{kk}(\tau)$, $\rho_{jk}(\tau)$ of the codes belonging to the same set will be zero, provided that we have $\tau \leq \iota T_c$. Furthermore, the LS codes belonging to the four different sets are still orthogonal to each other at zero timing offset, namely in a perfectly synchronous environment. However, the LS codes belonging to the four different sets will lose their orthogonality, when they have a non-zero code-offset. All four different codes in the same set of the LS(4,4,3) code family exhibited the same autocorrelation magnitudes, namely that seen in Figure 2(a). It can be observed in Figure 2(a) that the off-peak autocorrelation $R_p[\tau]$ becomes zero for $|\tau| \leq W_0 = 3$. The crosscorrelation magnitudes $|R_{j,k}(\tau)|$ depicted in Figure 2(b) are also zero for $|\tau| \leq W_0 = 3$. Based on the observations made as regards to the aperiodic correlations we may conclude that the LS(4,4,3) codes exhibit an IFW of ± 3 chip durations.

II. Space-Time Spreading Using Interference Rejection Codes

A. Transmitted Signal

As seen in Figure 3, the system considered in this paper consists of U antennas located at the transmitter side. The binary input data stream having a bit duration of T_b is serial-to-parallel (S/P) converted to U parallel sub-streams. The new bit duration of each reduced-rate parallel sub-stream, which we refer to as the symbol duration becomes $T_s = UT_b$. After S/P conversion, the U number of parallel bits which have a U-fold higher bit duration are direct-sequence spread using the STS schemes proposed in [2] with the aid of U number of orthogonal spreading sequences - for example, Walsh codes - having a period of UG, where $G = T_b/T_c$ represents the number of chips per bit and T_c is the chip-duration of the orthogonal spreading sequences.

As described above, based on the recommendations of [2] we have assumed that the number of parallel data sub-streams, the number of orthogonal spreading sequences used by the STS block of Figure 3 and the number of transmission antennas is the same, namely U. This specific STS scheme constitutes a sub-class of the generic family of STS schemes, where the number of parallel data sub-streams, the number of orthogonal spreading sequences required by the STS block and the number of transmission antennas may take different values. However, the study conducted in [2] has shown that the number of orthogonal spreading sequences required by STS is usually higher, than the number of parallel sub-streams. The STS scheme having an equal number of parallel sub-streams, orthogonal STS-related spreading sequences as well as transmission antennas constitutes an attractive scheme, since this STS scheme is capable of providing maximal transmit diversity without requiring extra STS codes [2]. Note that for the specific values of $U = 2, 4, 8$ the above-mentioned attractive STS schemes have been specified in [2]. In this section, we only investigate these attractive STS schemes.

Based on the philosophy of STS as discussed in [2] and referring to Figure 3(a), the transmitted signal of the kth user can be written as:

$$\mathbf{s}_k(t) = \sqrt{\frac{2P}{U^2}}\mathbf{c}(t)\mathbf{B}_U(t) \times \cos(2\pi f_c t), \tag{7}$$

where P represents each user's transmitted power, which is constant for all users, $\mathbf{s}_k(t) = [s_{k1}(t)\ s_{k2}(t)\ ...\ s_{kU}(t)]$ represents the transmitted signal vector of the U transmission antennas, while $\mathbf{c}(t)$ and f_c represent the DS scrambling-based spreading waveform and the carrier frequency, respectively. In Eq.(7) the vector $\mathbf{c}(t) = [c_1(t)\ c_2(t)\ ...\ c_U(t)]$ is constituted by the U number of spreading waveforms assigned for the STS block, where $c_i(t) = \sum_{j=0}^{UG} c_{ij} P_{T_c}(t - jT_c),\ i = 1, 2, \ldots, U$ denotes the individual components of the STS-based spread signals, and $\{c_{ij}\}$ represents a spreading sequence of period UG for each index i, while $P_{T_c}(t)$ is the rectangular chip-waveform spanning the chip-interval $[0, T_c]$. **In this paper, we will consider two different STS schemes. The benchmarker arrangement is the traditional STS scheme of [2], employed for example in W-CDMA.** In this scheme, $c_i(t)$ can be expressed as: $c_i(t) = w_i(t) \otimes \text{PN}(t) = \sum_{j=0}^{UG}(w_{ij} \otimes p_{ij}) P_{T_c}(t - jT_c)$, where $w_i(t) = \sum_{j=0}^{UG} w_{ij} P_{T_c}(t - jT_c)$ denotes the unique user-specific Walsh spreading sequence used by the STS scheme of Figure 3, while $\text{PN}(t) = \sum_{j=0}^{UG} p_{ij} P_{T_c}(t - jT_c)$ is the random cell-specific pseduo-noise scrambling sequence. Hence we have $c_{ij} = w_{ij} \otimes p_{ij}$ and c_{ij} may also be modelled by a random spreading sequence. The employment of the PN scrambling sequence in combination with the Walsh code allows the system to reuse the user-specific Walsh codes in adjacent cells and reduce the MAI. **In contrast to the benchmarker, in our proposed scheme space-time spreading is carried out using the family of interference rejection LS codes.** In this scheme, the spreading signature waveform $c_i(t)$ can be expressed as: $c_i(t) = \text{LS}_i(t) = \sum_{j=0}^{UG} \text{LS}_{ij} P_{T_c}(t - jT_c)$, where $\text{LS}_i(t)$ denotes the ith LS spreading signature waveform. **Moreover, we do not impose PN-code based scrambling on the LS STS codes, because this would destroy their IFW. Hence, for the sake of preserving the IFW of the spreading codes, the STS scheme using LS codes refrains from invoking the conventional scheme's scrambling operation.** It is worth noting that the omission of the PN-code based spreading does not constitute a problem in conjunction with LS codes, since they exhibit an IFW and hence they are more immune to both MAI and MPI than the Walsh codes. It has to be noted however that the owing to the absence of the PN scrambling code the amount of inter-cell interference is expected to be higher, especially, because the adjacent-cell interference is likely to arrive outside the IFW, where the cross-correlation of the LS codes is higher than that of the cell-specific PN scrambling codes. Still considering Eq.(7), $\mathbf{B}_U(t)$ represents the $(U \times U)$-dimensional transmitted data matrix created by mapping U input data bits to the U parallel sub-streams according to the specific design rules of [2], so that the maximum possible transmit diversity is achieved, while using relatively low-complexity signal detection algorithms.

B. Channel Model

The U number of parallel signals $\mathbf{s}_k(t) = [s_{k1}(t)\ s_{k2}(t)\ ...\ s_{kU}(t)]$ are transmitted by the U number of antennas over frequency-selective fading channels, where each parallel signal experiences independent frequency-selective Nakagami-m fading. The complex low-pass equivalent representation of the impulse response experienced by the uth parallel signal of all users is given by [9]:

$$h^u(t) = \sum_{l=1}^{L} h_l^u \delta(t - \tau_l) \exp\left(j\phi_l^u\right), \tag{8}$$

where h_l^u, τ_l and ψ_l^u represent the attenuation, delay and phase-shift of the lth multipath component of the channel, respectively. Without loss of generality, we assume that we have $\tau_l = (l-1)T_c$, while L is the total number of resolvable multipath components and $\delta(t)$ is the Kronecker Delta-function. We assume that the phases $\{\psi_l^u\}$ in Eq.(8) are independent identically distributed (iid) random variables uniformly distributed in the interval $[0, 2\pi)$, while the L multipath attenuations $\{h_l^u\}$ in Eq.(8) are independent Nakagami random variables having a Probability Density Function (PDF) of [10], [11], [12]:

$$\begin{aligned} p(h_l^u) &= M(h_l^u, m_l^{(u)}, \Omega_l^u), \\ M(R, m, \Omega) &= \frac{2m^m R^{2m-1}}{\Gamma(m)\Omega^m} e^{(-m/\Omega)R^2}, \end{aligned} \tag{9}$$

where $\Gamma(\cdot)$ is the gamma function [9], and $m_{kl}^{(u)}$ is the Nakagami-m fading parameter, which characterizes the severity of the fading over the l-th resolvable path [13] between the uth transmission antenna and user

k. Specifically, $m_l^{(u)} = 1$ represents Rayleigh fading, $m_l^{(u)} \to \infty$ corresponds to the conventional Gaussian scenario and $m_l^{(u)} = 1/2$ describes the so-called one-sided Gaussian fading, i.e. the worst-case fading condition. The Rician and log-normal distributions can also be closely approximated by the Nakagami distribution in conjunction with values of $m_l^{(u)} > 1$. The parameter Ω_l^u in Eq.(9) is the second moment of h_l^u, i.e. we have $\Omega_l^u = E[(\alpha_l^u)^2]$. We assume a negative exponentially decaying multipath intensity profile (MIP) given by $\Omega_l^u = \Omega_1^u e^{-\eta(l-1)}, \ \ \eta \geq 0$, where Ω_1^u is the average signal strength corresponding to the first resolvable path and η is the rate of average power decay.

We support K synchronous CDMA users in the system and assume perfect power control. Consequently, when the K users' signals obeying the form of Eq.(7) are transmitted over the frequency-selective fading channels characterized by Eq.(8), the received complex low-pass equivalent signal at a given mobile station can be expressed as:

$$R(t) = \sum_{k=1}^{K} \sum_{l=1}^{L} \sqrt{\frac{2P}{U^2}} \mathbf{c}(t-\tau_l)\mathbf{B}_U(t-\tau_l)\mathbf{h}_l + N(t), \tag{10}$$

where $N(t)$ is the complex-valued low-pass-equivalent AWGN having a double-sided spectral density of N_0, while

$$\mathbf{h}_l = \begin{pmatrix} h_l^1 \exp(j\psi_l^1) \\ h_l^2 \exp(j\psi_l^2) \\ \cdots \\ h_l^U \exp(j\psi_l^U) \end{pmatrix}, \ l = 1, 2, \ldots, L \tag{11}$$

represents the channel's complex impulse response in the context of the kth user and the lth resolvable path, where $\psi_l^u = \phi_l^u - 2\pi f_c \tau_l$. Furthermore, in Eq.(10) we assumed that the signals transmitted by the U number of transmission antennas arrive at the receiver antenna after experiencing the same set of delays. This assumption is justified by the fact that in the frequency band of cellular systems the propagation delay differences among the transmission antenna elements is on the order of nanoseconds, while the multipath delays are on the order of microseconds [2], provided that U is a relatively low number.

C. Receiver Model

Let the first user be the user-of-interest and consider a receiver using space-time de-spreading as well as diversity combining, as shown in Figure 3(b), where the subscript of the reference user's signal has been omitted for notational convenience. The receiver of Figure 3(b) carries out the inverse processing of Figure 3(a), in addition to multipath diversity combining. In Figure 3(b) the received signal is first down-converted using the carrier frequency f_c, and we assumed that the receiver is capable of achieving near-perfect multipath-delay estimation for the reference user. The de-scrambled signal associated with the lth resolvable path is space-time de-spread using the approach of [2] - which will be further discussed in Section III, in order to obtain U separate variables, $\{Z_{1l}, Z_{2l}, \ldots, Z_{Ul}\}$, corresponding to the U parallel data bits $\{b_1, b_2, \ldots, b_U\}$, respectively. Following space-time de-spreading, a decision variable is formed for each parallel transmitted data bit of $\{b_1, b_2, \ldots, b_U\}$ by Equal-Gain (EG) diversity combining the corresponding variables associated with the L number of resolvable paths, which can be written as:

$$Z_u = \sum_{l=1}^{L} Z_{ul},\ u = 1, 2, \ldots, U. \tag{12}$$

Finally, the U number of transmitted data bits $\{b_1, b_2, \ldots, b_U\}$ can be decided based on the decision variables $\{Z_u\}_{u=1}^{U}$ using the conventional decision rule of a BPSK scheme.

Above we have described the transmitter model, the channel model as well as the receiver model of W-CDMA using STS. Let us now describe the detection procedure of the W-CDMA scheme using STS.

III. DETECTION OF SPACE-TIME SPREAD SIGNALS

Let $\mathbf{d}_l = [d_{1l}\ d_{2l}\ \ldots\ d_{Ul}]^T$, $l = 1, 2, \ldots, L$ - where T denotes the vector transpose - represent the correlator's output variable vector in the context of the lth $(l = 1, 2, \ldots, L)$ resolvable path, where

$$d_{ul} = \int_{\tau_l}^{UT_b+\tau_l} R(t)c_u(t-\tau_l)dt. \tag{13}$$

When substituting Eq.(10) into Eq.(13), it can be shown that:

$$d_{ul} = \sqrt{2P}T_b \left[a_{u1}b_{u1}h_l^1 \exp\left(j\psi_l^1\right) + a_{u2}b_{u2}h_l^2 \exp\left(j\psi_l^2\right) + \ldots \right.$$
$$\left. \ldots + a_{uU}b_{uU}h_l^U \exp\left(j\psi_l^U\right)\right] + J_u(l), \;\; u = 1, 2, \ldots, U, \quad (14)$$

where

$$J_u(l) = J_{Su}(l) + J_{Mu}(l) + N_u(l), \; u = 1, 2, \ldots, U \quad (15)$$

and $J_{Su}(l)$ is due to the multipath-induced self-interference of the signal-of-interest inflicted upon the lth path signal, where $J_{Su}(l)$ can be expressed as:

$$J_{Su}(l) = \sum_{j=1, j\neq l}^{L} \sqrt{\frac{2P}{U^2}} \int_{\tau_l}^{UT_b+\tau_l} \mathbf{c}(t-\tau_j)\mathbf{B}_U(t-\tau_j)\mathbf{h}_j$$
$$\times c_u(t-\tau_l)dt, \quad (16)$$

$J_{Mu}(l)$ of Eq.15 represents the multi-user interference inflicted by the signals transmitted simultaneously by the other users, which can be expressed as:

$$J_{Mu}(l) = \sum_{k=2}^{K}\sum_{j=1}^{L} \sqrt{\frac{2P}{U^2}} \int_{\tau_l}^{UT_b+\tau_l} \mathbf{c}(t-\tau_j)\mathbf{B}_U(t-\tau_j)\mathbf{h}_j$$
$$\times c_u(t-\tau_l)dt, \quad (17)$$

and finally $N_u(l)$ of Eq.15 is due to the AWGN, formulated as:

$$N_u(l) = \int_{\tau_l}^{UT_b+\tau_l} N(t)c_u(t-\tau_l)dt, \quad (18)$$

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

Let $\mathbf{J}(l) = [J_1(l) \; J_2(l) \; \ldots \; J_U(l)]^T$. Then, the correlator's output variable vector $\mathbf{d}_l$ can be written as:

$$\mathbf{d}_l = \sqrt{2P}T_b\mathbf{B}_U\mathbf{h}_l + \mathbf{J}(l), \; l = 1, 2, \ldots, L, \quad (19)$$

where $\mathbf{B}_U$ is the reference user's $(U \times U)$-dimensional transmitted data matrix, when we ignore the time dependence, while $\mathbf{h}_l$ is the channel's complex impulse response between the base station and the reference user, as shown in Eq.(11) in the context of the reference user.

Attractive STS schemes have the property [2] of $\mathbf{B}_U \mathbf{h}_l = \mathbf{H}_U \mathbf{b}$, i.e. Equation (19) can be written as:

$$\mathbf{d}_l = \sqrt{2P}T_b \mathbf{H}_U \mathbf{b} + \mathbf{J}(l), \tag{20}$$

where $\mathbf{b} = [b_1\ b_2\ \ldots\ b_U]^T$ represents the U number of transmitted data bits, while $\mathbf{H}_U$ is a $(U \times U)$-dimensional matrix with elements from $\mathbf{h}_l$. Each element of $\mathbf{h}_l$ appears once and only once in a given row and also in a given column of the matrix $\mathbf{H}_U$ [2]. The matrix $\mathbf{H}_U$ can be expressed as:

$$\mathbf{H}_U(l) = \begin{pmatrix} \alpha_{11}(l) & \alpha_{12}(l) & \ldots & \alpha_{1U}(l) \\ \alpha_{21}(l) & \alpha_{22}(l) & \ldots & \alpha_{2U}(l) \\ \vdots & \vdots & \ddots & \vdots \\ \alpha_{U1}(l) & \alpha_{U2}(l) & \ldots & \alpha_{UU}(l) \end{pmatrix}, \tag{21}$$

where $\alpha_{ij}(l)$ takes the form of $d_{ij} h_l^m \exp(j\psi_l^m)$, and $d_{ij} \in \{+1, -1\}$ represents the sign of the (i, j)th element of $\mathbf{H}_U$, while $h_l^m \exp(j\psi_l^m)$ belongs to the mth element of $\mathbf{h}_l$.

It can be shown furthermore with the aid of the analysis provided in [2] that the matrix $\mathbf{H}_U(l)$ has the property of $\text{Re}\left\{\mathbf{H}_U^\dagger(l)\mathbf{H}_U(l)\right\} = \mathbf{h}_l^\dagger \mathbf{h}_l \cdot \mathbf{I}$, where † denotes the complex conjugate transpose and $\mathbf{I}$ represents a $(U \times U)$-dimensional unity matrix. Letting $\mathbf{h}_u(l)$ denote the uth column of $\mathbf{H}_U(l)$, the variable Z_{ul} in Eq.(12) can be formulated as [2]:

$$Z_{ul} = \text{Re}\left\{\mathbf{h}_u^\dagger(l)\mathbf{d}_l\right\} = \sqrt{2P}T_b b_u \sum_{u=1}^{U} |h_l^u|^2 + \text{Re}\left\{\mathbf{h}_u^\dagger(l)\mathbf{J}(l)\right\},\ u = 1, 2, \ldots, U. \tag{22}$$

Finally, according to Eq.(12) the decision variables associated with the U parallel transmitted data bits $\{b_1, b_2, \ldots, b_U\}$ of the reference user can be expressed as:

$$Z_u = \sqrt{2P}T_b b_u \sum_{l=1}^{L} \sum_{u=1}^{U} |h_l^u|^2 + \sum_{l=1}^{L} \text{Re}\left\{\mathbf{h}_u^\dagger(l)\mathbf{J}(l)\right\},\ u = 1, 2, \ldots, U, \tag{23}$$

which shows that the receiver is capable of achieving a diversity order of UL, as indicated by the related sums of the first term.

Above we have analyzed the detection procedure applicable to W-CDMA signals generated using STS. Let us now derive the corresponding BER expression.

IV. BER ANALYSIS

In this section we derive the BER expression of the STS-assisted W-CDMA system by first analyzing the statistics of the variable $Z_u,\ u = 1, 2, \ldots, U$ with the aid of the Gaussian approximation [14]. According to Equation 23, for a given set of complex channel transfer factor estimates $\{h_l^u\}$, Z_u can be approximated as a Gaussian variable having a mean given by:

$$\mathrm{E}\left[Z_u\right] = \sqrt{2P}T_b b_u \sum_{l=1}^{L}\sum_{u=1}^{U} |h_l^u|^2. \tag{24}$$

Based on the assumption that the interferences imposed by the different users, the different paths as well as by the AWGN constitute independent random variables, the variance of Z_u may be expressed as:

$$\begin{aligned}
\mathrm{Var}\left[Z_u\right] &= \mathrm{E}\left[\left(\sum_{l=1}^{L} \mathrm{Re}\left\{\mathbf{h}_u^{\dagger}(l)\mathbf{J}(l)\right\}\right)^2\right] \\
&= \sum_{l=1}^{L} \mathrm{E}\left[\left(\mathrm{Re}\left\{\mathbf{h}_u^{\dagger}(l)\mathbf{J}(l)\right\}\right)^2\right] \\
&= \frac{1}{2}\sum_{l=1}^{L} \mathrm{E}\left[\left(\mathbf{h}_u^{\dagger}(l)\mathbf{J}(l)\right)^2\right].
\end{aligned} \tag{25}$$

Substituting $\mathbf{h}_u(l)$, which is the uth column of $\mathbf{H}_u(l)$ in Eq.(21), and $\mathbf{J}(l)$ having elements given by Eq.(15) into the above equation, it can be shown that for a given set of channel estimates $\{h_l^u\}$, Eq.(25) can be simplified as:

$$\begin{aligned}
\mathrm{Var}\left[Z_u\right] &= \frac{1}{2}\sum_{l=1}^{L}\sum_{u=1}^{U} |h_l^u|^2 \mathrm{E}\left[\left(J_u(l)\right)^2\right] \\
&= \frac{1}{2}\sum_{l=1}^{L}\sum_{u=1}^{U} |h_l^u|^2 \mathrm{Var}\left[J_u(l)\right],
\end{aligned} \tag{26}$$

where $J_u(l)$ is given by Eq.(15). In deriving Eq.(26) we exploited the assumption of $\mathrm{Var}\left[J_1(l)\right] = \mathrm{Var}\left[J_2(l)\right] = \ldots = \mathrm{Var}\left[J_U(l)\right]$.

A. *STS Assisted CDMA Using LS codes*

Having characterized the various sources of interferecne, let us now demonstrate that with the advent of having an IFW, the LS codes are capable of suppressing both the MAI and MPI. More specifically, only the paths which fall outside the IFW will impose MAI and MPI on the decision. We assume having $T_{\text{IFW}} = \iota T_c$, and the jth path will inflict interference upon the lth finger of the RAKE receiver only if we have:

$$|\tau_j - \tau_l| > \iota T_c, \tag{27}$$

which is corresponds to:

$$|j - l| > \iota. \tag{28}$$

Let us first consider the effect of MPI, Similarly to the benchmark of [2], it can be shown for the proposed LS code based system that $J_{Su}(l)$ defined in Eq.(16) is also constituted by U^2 terms and each term takes the form of,

$$\begin{aligned} J_{su}(l) &= \sum_{\substack{j=1 \\ |j-l|>\iota}}^{L} \sqrt{\frac{2P}{U^2}} \int_{\tau_l}^{UT_b+\tau_l} c_m(t-\tau_j) a_{mn} b_{mn}(t-\tau_j) \times \\ &\qquad h_j^n \exp(j\psi_j^n) c_u(t-\tau_l) dt. \end{aligned} \tag{29}$$

If we define the partial auto-correlation coefficient as:

$$\rho_{mm}(j,l) = \frac{1}{UT_b} \int_0^{|\tau_j-\tau_l|} c_m(t) c_m(t - |\tau_j - \tau_l|) dt \tag{30}$$

$$\varrho_{mm}(j,l) = \frac{1}{UT_b} \int_{|\tau_j-\tau_l|}^{UT_b} c_m(t) c_m(t - |\tau_j - \tau_l|) dt, \tag{31}$$

then the integral in Eq.(29) can be expressed as:

$$\int_{\tau_l}^{UT_b+\tau_l} c_m(t-\tau_j) c_u(t-\tau_l) b_{mn}(t-\tau_j) dt = (\rho_{mm}(j,l) b[-1] + \varrho_{mm}(j,l) b[0]) UT_b. \tag{32}$$

Therefore, the corresponding MPI variance of $J_{su}(l), \;\; u = 1, 2, \ldots, U$ can be expressed as:

$$\begin{aligned} \text{Var}[J_{su}(l)] &= \sum_{\substack{j=1 \\ |j-l|>\iota}}^{L} \left\{ 2\Omega_1 e^{(j-1)\eta} [\rho_{mm}^2(j,l) + \varrho_{mm}^2(j,l)] \right\} \times 2E_b T_b \\ &= \sum_{\substack{j=1 \\ |j-l|>\iota}}^{L} \left\{ 2 e^{(j-1)\eta} [\rho_{mm}^2(j,l) + \varrho_{mm}^2(j,l)] \right\} \times 2\Omega_1 E_b T_b. \end{aligned} \tag{33}$$

For convenient formulation and comparison with the benchmarker STS scheme of [2], we define $\Upsilon_S(l) = GU \sum_{\substack{j=1 \\ |j-l|>\iota}}^{L} 2e^{(j-1)\eta}[\rho^2_{mm}(j,l) + \varrho^2_{mm}(j,l)]$, which is the MPI reduction factor for the lth path, owing to the employment of LS codes. Then the MPI variance of Var$[J_{Su}(l)]$, which includes a total of U^2 number of Var$[J_{su}(l)]$ terms, can be approximated as:

$$\text{Var}[J_{Su}(l)]_{\text{II}} = \Upsilon_S(l) 2\Omega_1 E_b T_b U/G. \tag{34}$$

Having characterized the MPI, let us now focus our attention on the effects of MAI. Similarly to Eq.(30) and Eq.(31), we define the partial cross-correlation coefficients as:

$$\begin{aligned} \rho_{um}(j,l) &= \frac{1}{UT_b}\int_0^{|\tau_j-\tau_l|} c_u(t)c_m(t-|\tau_j-\tau_l|)dt & (35)\\ \varrho_{um}(j,l) &= \frac{1}{UT_b}\int_{|\tau_j-\tau_l|}^{UT_b} c_u(t)c_m(t-|\tau_j-\tau_l|)dt. & (36)\end{aligned}$$

Hence the integral in Eq.(17) may be expressed as:

$$\int_{\tau_l}^{UT_b+\tau_l} c_m(t-\tau_j)c_u(t-\tau_l)b_{mn}(t-\tau_j)dt = (\rho_{um}(j,l)b_m[-1] + \varrho_{um}(j,l)b_m[0])UT_b. \tag{37}$$

Similar to the benchmarker of [2], in the LS code based STS scheme the MAI term, namely $J_{Mu}(l)$ defined in Eq.(17) also consists of U^2 terms, and each term takes the form of:

$$\begin{aligned} J_{mu}(l) = \sum_{k=2}^{K}\sum_{\substack{j=1 \\ |j-l|>\iota}}^{L}\sqrt{\frac{2P}{U^2}}\int_{\tau_l}^{UT_b+\tau_l} & c_m(t-\tau_j)a_{mn}b_{mn}(t-\tau_j)\\ & \times h_j^l \exp(j\psi_j^n)c_u(t-\tau_l)dt, \end{aligned} \tag{38}$$

while the variance of the MAI $J_{mu}(l)$ can be expressed as:

$$4\Omega_1 E_b T_b \sum_{\substack{m=1 \\ m\neq u}}^{K}\sum_{\substack{j=1 \\ |j-l|>\iota}}^{L} e^{-(j-1)\eta}[\rho^2_{um} + \varrho^2_{um}]. \tag{39}$$

Similarly to the MPI reduction factor, we define the MAI reduction factor as: $\Upsilon_M(l) = \frac{GU}{K-1}\sum_{\substack{m=1 \\ m\neq u}}^{K}\sum_{\substack{j=1 \\ |j-l|>\iota}}^{L} 2e^{-(j-1)\eta}[\rho^2_{um} + \varrho^2_{um}]$. Then, the MAI variance Var$[J_{Mu}(l)]$ of the proposed LS code based STS scheme, which includes a total of U^2 terms of the form Var$[J_{mu}(l)]$, can be approximated as::

$$\text{Var}[J_{Mu}(l)]_{\text{II}} = \Upsilon_M(l) \times 2(K-1)U\Omega_1 E_b T_b U/G. \tag{40}$$

B. Probability of Bit Error

Having characterized the MAI and MPI variance, let us now quantify the achievable BER performance of the proposed system. Based on analysis [2], [15], for the traditional Walsh-code based STS, the variance can be expressed as:

$$\text{Var}\left[J_u(l)\right] \;=\; 2N_0UT_b + \frac{2UK\Omega_1 E_b T_b[q(L, e^{-\eta(l-1)}]}{G}. \tag{41}$$

By contrast, the corresponding variance of $J_u(l)$ of the LS code based STS scheme can be expressed with the aid of Eq.(15) as:

$$\text{Var}[J_u(l)] = 2N_0UT_b + \frac{\Upsilon_S(l)2\Omega_1 E_b T_b U}{G} + \frac{\Upsilon_M(l)(K-1)2\Omega_1 E_b T_b U}{G}. \tag{42}$$

Let us now assume that the Rake receiver is capable of combining a maximum of L_R paths' energy owing to its complexity limitation. Then the variance of Z_u can expressed as:

$$\text{Var}\left[Z_u\right] = \frac{1}{2}\sum_{l=1}^{L_R}\sum_{u=1}^{U}|h_l^u|^2 \times \text{Var}[J_u(l)], \tag{43}$$

for a given set of channel estimates $\{h_l^u\}$ using Eq.(24). Hence the BER conditioned on h_l^u for $u = 1, 2, \ldots, U$ and $l = 1, 2, \ldots, L_R$ can be written as:

$$P_b(E|\left\{h_l^u\right\}) = Q\left(\sqrt{\frac{\text{E}^2\left[Z_u\right]}{\text{Var}\left[Z_u\right]}}\right) = Q\left(\sqrt{\sum_{l=1}^{L_R}\sum_{u=1}^{U} 2\gamma_{lu}}\right), \tag{44}$$

where $Q(x)$ represents the Gaussian Q-function, which can also be represented in its less conventional form as $Q(x) = \frac{1}{\pi}\int_0^{\pi/2}\exp\left(-\frac{x^2}{2\sin^2\theta}\right)d\theta$, where $x \geq 0$ [13], [16]. Furthermore, $2\gamma_{lu}$ in Eq.(44) represents the output Signal to Interference plus Noise Ratio (SINR) experienced at the lth finger of the RAKE receiver for the uth STS antenna.

In both the Walsh-code aided [2] and in the LS code based STS scheme, γ_{lu} of Eq.(44) is given by:

$$\gamma_{lu} = \frac{(\text{E}[Z_{ul}])^2}{\text{Var}[J_u(l)]} = \overline{\gamma}_{lc} \cdot \frac{(h_l^u)^2}{\Omega_1}. \tag{45}$$

However, in the Walsh-code based STS scheme of [2], $\overline{\gamma}_{lc}$ is given by:

$$\overline{\gamma}_{lc} = \frac{1}{U}\left[\frac{K[q(L,\eta) - e^{-\eta(l-1)}]}{G} + \left(\frac{\Omega_1 E_b}{N_0}\right)^{-1}\right]^{-1}. \tag{46}$$

By contrast, in the LS code based STS scheme, $\overline{\gamma}_{lc}$ can be expressed as:

$$\overline{\gamma}_{lc} = \frac{1}{U}\left[\frac{\Upsilon_S(l)}{G} + \frac{(K-1)\Upsilon_M(l)}{G} + \left(\frac{\Omega_1 E_b}{N_0}\right)^{-1}\right]^{-1}, \tag{47}$$

where the MPI and MAI reduction factors $\Upsilon_S(l),\ \ \Upsilon_M(l)$, respectively, reflect how much interference is suppressed for the lth path with the advent of the IFW, which is mainly determined by the width of the IFW and by the number of resolvable paths, *i.e,* by ι and L.

The average BER, $P_b(E)$ can be obtained by averaging the conditional BER expression of Eq.(44) over the joint PDF of the instantaneous SNR values corresponding to the L_R multipath components and to the U transmit antennas $\{\gamma_{lu}:\ l=1,2,\ldots,L_R;\ u=1,2,\ldots,U\}$. Since the random variables $\{\gamma_{lu}:\ l=1,2,\ldots,L_R;\ u=1,2$ are assumed to be statistically independent, the average BER can be formulated as [17]:

$$P_b(E) = \frac{1}{\pi}\int_0^{\pi/2}\prod_{l=1}^{L_R}\prod_{u=1}^{U} I_{lu}\left(\overline{\gamma}_{lu},\theta\right)d\theta, \tag{48}$$

where we have

$$I_{lu}\left(\overline{\gamma}_{lu},\theta\right) = \int_0^{\infty}\exp\left(-\frac{\gamma_{lu}}{\sin^2\theta}\right)p_{\gamma_{lu}}(\gamma_{lu})d\gamma_{lu}. \tag{49}$$

Since both $\gamma_{lu} = \overline{\gamma}_{lc}\cdot\frac{(h_l^u)^2}{\Omega_1}$ and h_l^u obey the Nakagami-m distribution characterized by Eq.(9), it can be shown that the PDF of γ_{lu} can be formulated as:

$$p_{\gamma_{lu}}(\gamma_{lu}) = \left(\frac{m_l}{\overline{\gamma}_{lu}}\right)^{m_l}\frac{\gamma^{m_l-1}}{\Gamma(m_l)}\exp\left(-\frac{m_l\gamma_{lu}}{\overline{\gamma}_{lu}}\right),\ \ \gamma_{lu}\geq 0, \tag{50}$$

where $\overline{\gamma}_{lu} = \overline{\gamma}_{lc}e^{-\eta(l-1)}$ for $l=1,2,\ldots,L$.

Upon substituting Eq.(50) into Eq.(49) it can be shown that [13]:

$$I_{lu}\left(\overline{\gamma}_{lu},\theta\right) = \left(\frac{m_l^u\sin^2\theta}{\overline{\gamma}_{lu}+m_l^u\sin^2\theta}\right)^{m_l^u}. \tag{51}$$

Finally, upon substituting Eq.(51) into Eq.(48), the average BER of the STS-assisted W-CDMA system using U transmission antennas can be written as:

$$P_b(E) = \frac{1}{\pi}\int_0^{\pi/2}\prod_{l=1}^{L_R}\prod_{u=1}^{U}\left(\frac{m_l^u\sin^2\theta}{\overline{\gamma}_{lu}+m_l^u\sin^2\theta}\right)^{m_l^u}d\theta, \tag{52}$$

which shows that the diversity order achieved is $L_R U$, namely the product of the diversity due to STS and the diversity contributed by the RAKE receiver. Furthermore, if we assume that m_l is independent of u, i.e. that all of the parallel transmitted signals experience an identical Nakagami fading, then Eq.(52) can be expressed as:

$$P_b(E) = \frac{1}{\pi}\int_0^{\pi/2}\prod_{l=1}^{L_R}\left(\frac{m_l \sin^2\theta}{\overline{\gamma}_{lu} + m_l \sin^2\theta}\right)^{Um_l} d\theta. \tag{53}$$

V. Numerical Results

Having characterized the analytical performance of the system, let us now consider the achievable BER performance. Stańczak *et al.* [4] concluded that when using LS codes the width ι of the IFW and the number of users K has to obey:

$$(\iota - 1)K \leq G. \tag{54}$$

Furthermore, LS codes require W_0 number of zero-valued chips, which are inserted in the beginning and center of the code sequence for creating the IFW. In our scenario, the LS(N, P, W_0)=LS(4,32,4) codes having a length of $L_S = NP + 2W_0 = 136$ were invoked, and their effective spreading gain was $L_S = NP = 128$, since the zero-valued chips do not include the spreading gain. For the sake of maintaining the same chip rate and same spectral efficiency for both STS schemes, we set the spreading gain of the traditional STS assisted CDMA system to: $G' = NP + 2W_0 = 136$. Furthermore, for simplicity's sake we assume that all paths have the same Nakagami fading parameter, *i.e.* $m_l = m,\ l = 0, \ldots, L_r - 1$.

We assume that the chip rate is $1.2288M$chip/s, the channel's delay spread is negative exponentially distributed having a uniformly distributed mean-delay in the range of $[0.3, 3]\mu s$ [18], and we assume that both the random and LAS-code based systems have a chip rate of $1.2288M$chips. The number of resolvable paths is $L_p = \lfloor \frac{\tau}{T_c} \rfloor + 1 = 4$, where we have $\tau = 3\mu s$. Both the traditional STS and the LS code based STS schemes supported $K = 32$ users, and the width of the IFW of the LS codes was $\iota = 3$. We can see from Figure 4 that the LS code based STS scheme exhibits a significantly better performance than the traditional Walsh-code based system having that the same diversity order of $L_R U$. The reason that the

LS code based STS scheme outperforms the traditional STS scheme is that the MAI and MPI is reduced, as a benefit of using LAS codes, which was quantified by Eq.(53). Figure 5 characterizes the achievable performance of these two schemes communicating over different fading channels associated with different Nakagami fading parameters. More explicitly, when we have $m = 1$, we model a Rayleigh fading channel, $m = 2$ represents a Rician fading channel, while $m \to \infty$ corresponds to an AWGN channel. From this figure we can observe that the LS code based STS scheme exhibited a better performance than the traditional STS scheme, regardless of the value of m. More specifically, provided that we have $L_p = 4$, the LS code based STS scheme outperformed the traditional STS scheme, when communicating over different Nakagami multipath fading channels.

Figure 6 shows the performance of these two systems for transmission over different dispersive channels having $L = 4 \ldots 12$ resolvable multipath components, but assuming that only $L_r = 3$ of these components were combined by the RAKE receiver owing to its limited affordable complexity. From Figure 6, we may conclude that the LS codes are effective, when the number of resolvable paths is relatively low, for example when we have $L = 4$. When L is increased to 8, the LS code based STS scheme only has a slight gain over the traditional STS scheme, while when L is increased to 12, the LS code based STS scheme performs even worse than the traditional STS scheme. The reason for this performance erosion is that many of the paths will be located outside the IFW, when L_p is high and the auto-correlation as well as cross-correlation of the LS codes outside the IFW is higher than that of the random codes. Hence many of the multipath components arrive outside the IFW when L is high, which inevitably will increase both the MAI and the MPI.

In order to circumvent the performance limitation of the proposed system, we finally introduce the concept of multicarrier LAS DS-CDMA, which allow to extend the IFW duration by a factor of the number subcarriers. Figure 7 demonstrated the achievable performance of Single-carrier (SC) LS codes assisted STS and Multicarrier (MC) LS codes based STS for a single-carrier $3.884M$chips/s system. From this figure we may conclude that the LS code based STS assisted Multicarrier DS-CDMA is capable of achieving the best performance trade-off by selecting the optimum number of subcarriers U_s according to the channel delay dispersion τ_{ch}. For example, From Figure 7, we may conclude that $U_s = 4$ MC LAS DS-CDMA system exhibited the best trade-off in a scenario of $\tau_{ch} = 3\mu s$.

From Figure 8, we can observe that if the system's user load is high, the LS code based STS scheme will have no advantage over the traditional Walsh-code based STS scheme of [2], which is caused by two factors. First, the number of LS codes having an IFW of $\iota = 3$ is limited. For example, when we consider $G = 128$, only 32 LS codes have an IFW of $\iota = 3$, and when the number of users K exceeds 32, the width of the IFW will be reduced to zero, since even the codes having $\iota = 0$ will be required for supporting $K \geq 32$ users. In this scenario, the LS code based STS scheme becomes incapable of effectively suppressing the MAI and MPI. Second, it may be shown that the cross-correlation of LS codes outside the IFW is higher than that of the random codes, hence LS codes may impose increased interferences, when the number of users K is increased. Therefore, the LS code based STS scheme is more effective in low-user-load scenarios, *i.e* when we have $K \leq G/3$.

VI. CONCLUSION

The proposed LS code based STS scheme exhibited a significantly better performance than that of the traditional Walsh-code based STS scheme [2], when the number of users supported does not exceed $G/3$. As the number of resolvable paths L of the channel increases, the LS code based STS scheme only has a slight gain over the traditional STS scheme [2], owing to the fact that many of the paths arrive outside the IFW and also because the auto-correlation as well as cross-correlation of LS codes outside the IFW is higher than that of the random codes. Furthermore, when communicating in a high-user-load scenario, for example when we have $K = G$, the LS code based STS scheme may exhibit a worse performance than the traditional STS scheme of [2].

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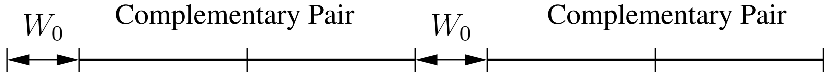
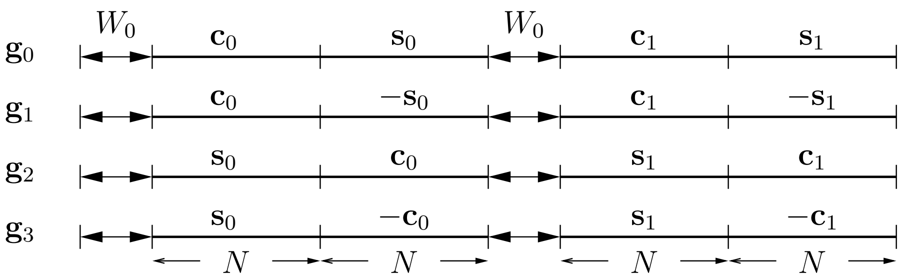
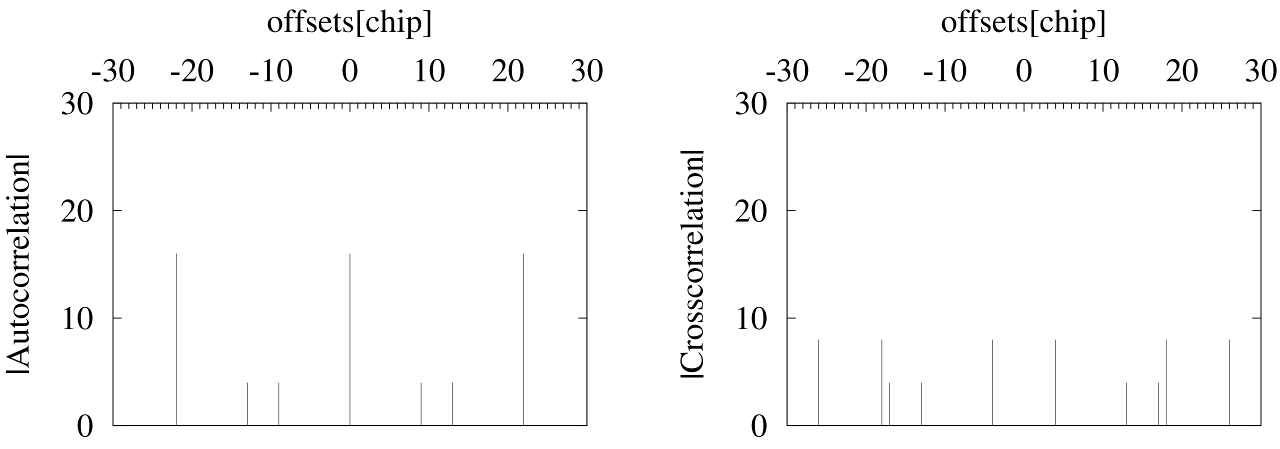


(a) The LS code structure



(b) Generating four LS codes.

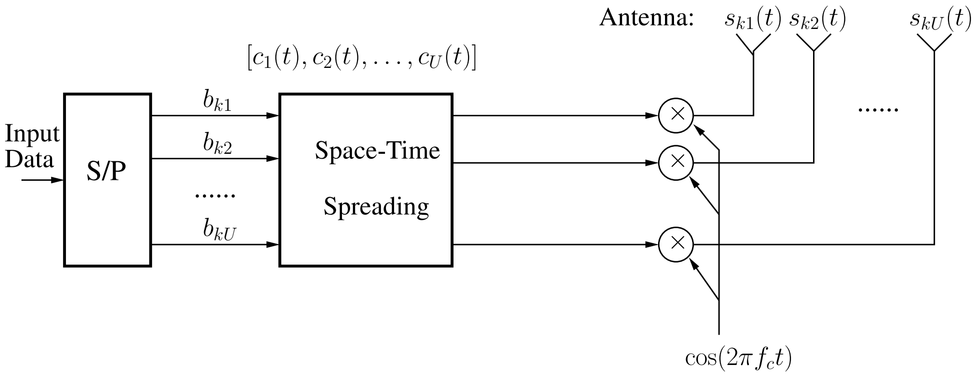
Fig. 1. Generating the LS(N, P, W_0) code using the $(P \times P) = (4 \times 4)$ Walsh-Hadamard matrix components $(1, 1, 1, 1)$ and $(1, -1, 1, -1)$.



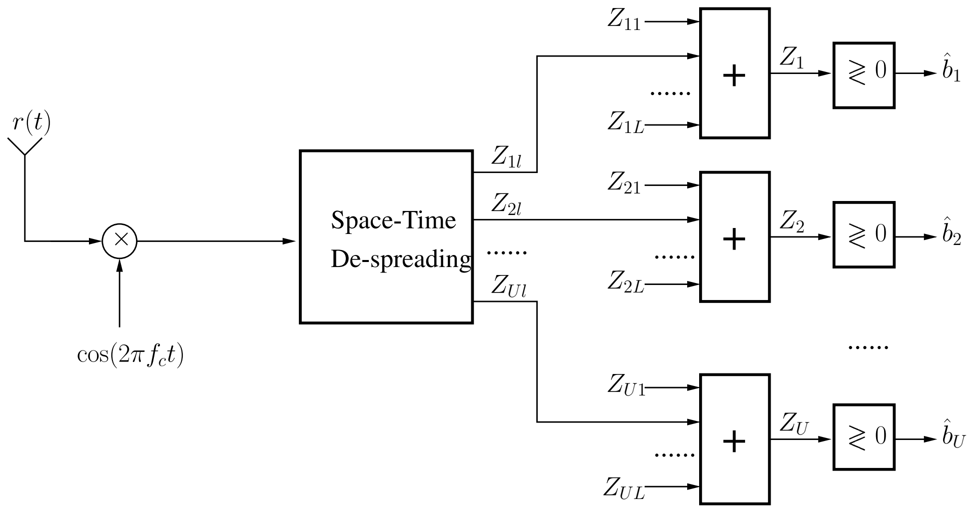
(a) Autocorrelation

(b) Crosscorrelations

Fig. 2. correlation magnitudes of the LS$(4, 4, 3)$ codes. (a) All four codes exhibit the same autocorrelation magnitude. (b) The crosscorrelation magnitudes of $\mathbf{g}_0$ and $\mathbf{g}_2$.



(a) Transmitter



(b) Receiver

Fig. 3. Transmitter and receiver block diagram of the W-CDMA system using space-time spreading.

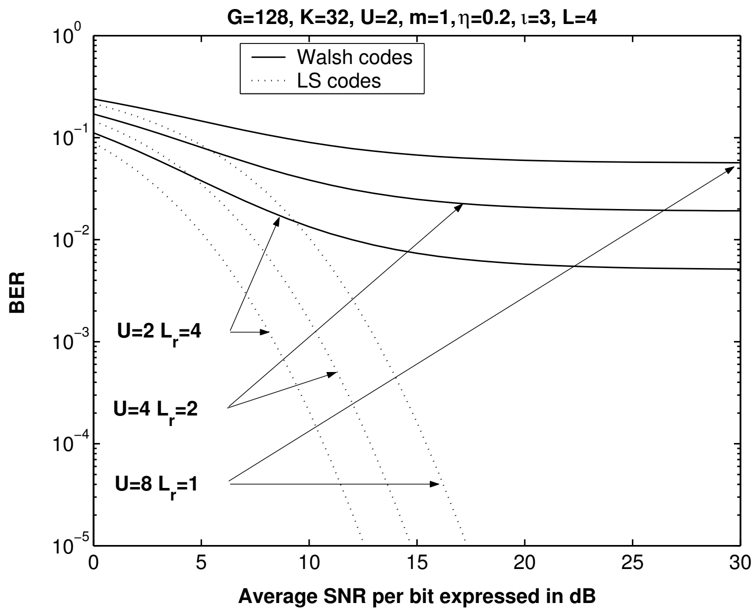


Fig. 4. BER versus SNR per bit, E_b/N_0, performance comparison between the Walsh-code and LS-code based STS transmit diversity schemes having the same diversity order of $L_R \cdot U$, when communicating over a Nakagami-m ($m = 1$) fading multipath ($L = 4$) channel evaluated from Eq.(53) by assuming that the average power decay rate was $\eta = 0.2$. The remaining system parameters are listed at the top of the figure.

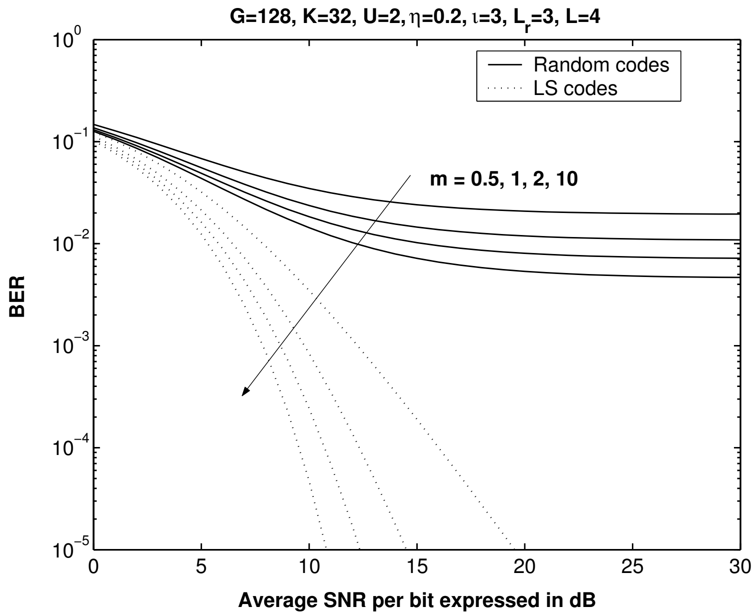


Fig. 5. BER versus SNR per bit, E_b/N_0, performance comparison between the Walsh-code and LS-code based STS transmit diversity schemes, when communicating over various Nakagami-m fading multipath ($L = 4$) channels, where $L_r = 3$ out of the $L = 3$ available paths were combined by the Rake receiver. The remaining system parameters are listed at the top of the figure.

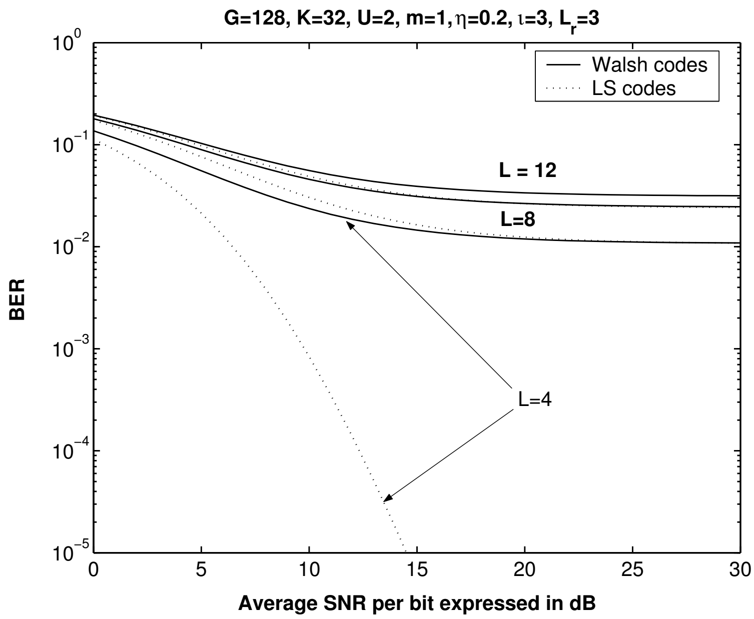


Fig. 6. BER versus SNR per bit, E_b/N_0, performance comparison between the Walsh-code and LS-code based transmit diversity schemes and the conventional RAKE receiver arrangement, when communicating over different dispersive Nakagami-m channels having $L = 4, 8$ and 12 resolvable paths, but only combining $L_R = 3$ of them owing to the maximum complexity limitations.

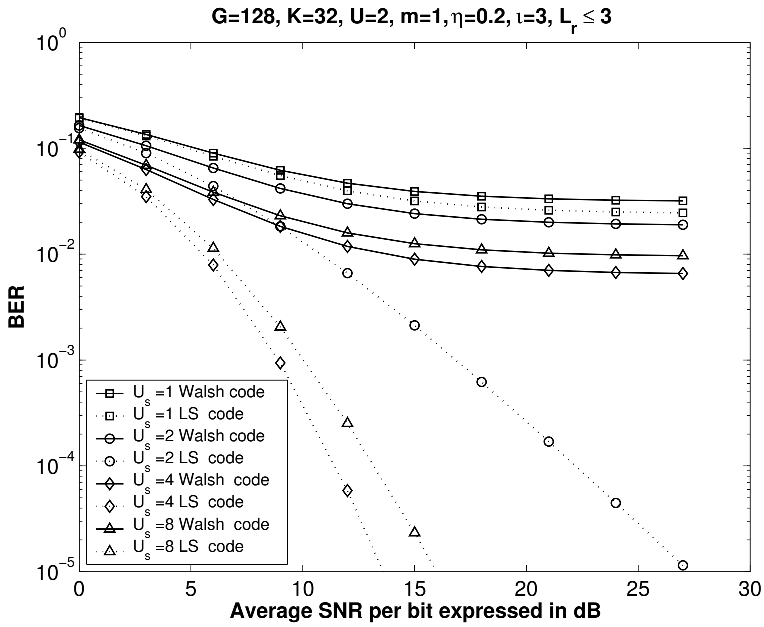


Fig. 7. BER versus SNR per bit, E_b/N_0, performance comparison between the Walsh-code and LS-code based transmit diversity schemes and the conventional RAKE receiver arrangement, when invoking the multicarrier modulation where the number of subcarrier U_s is $1, 2, 4, 8$. respectively. Furthermore, only combining $L_R = 3$ of them owing to the maximum complexity limitations.

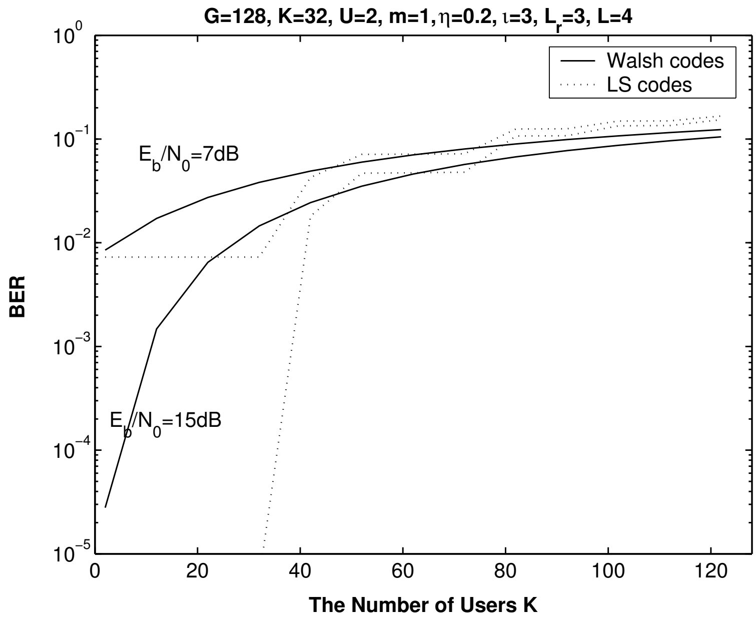


Fig. 8. BER versus SNR per bit, E_b/N_0, performance comparison between the Walsh-code and LS-code based STS transmit diversity schemes as a function of the number of users K. The remaining system parameters are listed at the top of the figure.