On the Uplink Performance of LAS-CDMA

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***Abstract*— In this paper Large Area Synchronized (LAS)-CDMA is investigated, which exhibits a significantly better performance than the traditional random code based DS-CDMA system. Closed-form formulas are derived for characterizing the BER performance of LAS-CDMA as a function of the number of resolvable paths L_p, the maximum delay difference τ_{max} and the number of users K when communicating over a Nakagami-m fading channel. Moreover, the limited number of available LAS codes having a certain interference free window width suggests that the employment of LAS-CDMA is beneficial in a low-user-load scenario, where a near-single-user performance can be achieved without a multiuser detector. In this paper Large Area Synchronized (LAS)-CDMA is investigated, which exhibits a significantly better performance than the traditional random code based DS-CDMA system. Closed-form formulas are derived for characterizing the BER performance of LAS-CDMA as a function of the number of resolvable paths L_p, the maximum delay difference τ_{max} and the number of users K when communicating over a Nakagami-m fading channel. Moreover, the limited number of available LAS codes having a certain interference free window width suggests that the employment of LAS-CDMA is beneficial in a low-user-load 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, LAS codes, Gaussian Approximation.**

I. INTRODUCTION

IN Direct Sequence Code Division Multiple Access (DS-CDMA) systems, the spreading sequences characterize the associated Inter Symbol Interference (ISI) as well as the Multiple Access Interference (MAI) properties [1]. Traditional spreading sequences, such as m-sequences [1], Gold codes [1] and Kasami codes [1] exhibit non-zero off-peak auto-correlations and cross-correlations, which results in a high MAI in case of asynchronous uplink transmissions. Another family of orthogonal codes is constituted by Walsh codes [1] and orthogonal Gold codes, which retain their orthogonality only in case of perfect synchronization, but they also exhibit non-zero off-peak auto-correlations and cross-correlations in asynchronous scenarios. Consequently, these correlation properties limit the achievable performance in asynchronous scenarios. Hence traditional DS-CDMA cellular systems are interference limited and suffer from the so-called 'near-far' effects, unless complex interference cancellers [1] or multi-user detectors [1] are employed for combating these adverse effects. This results in costly and 'power-hungry' implementations. All these limitations are imposed by the imperfect correlation properties of the spreading sequences employed.

Hence, considerable research efforts have been invested in designing spreading sequences, which exhibit zero correlation values, when the relative delay-induced code offset is in the so-called Zero Correlation Zone (ZCZ) or Interference Free Window (IFW) of the spreading code [2]. The attractive family of Large Area Synchronized (LAS) CDMA spreading sequences is constituted by the combination of the so-called Large Area (LA) codes [3], [4] and Loosely Synchronous (LS) codes [5]. The resultant LAS codes exhibit an IFW, where the off-peak aperiodic autocorrelation values as well as the aperiodic cross-correlation values become zero, resulting in zero ISI and zero MAI, provided that the time-offset of the codes is within the IFW. In order to ensure that the relative time-offsets between the codes are within the IFW, the mobiles are expected to operate in a quasi-synchronous manner. More specifically, interference-free CDMA communications become possible, when the total time-offset expressed in terms of the number of chip intervals, which is the sum of the time-offset of the mobiles plus the maximum channel-induced delay spread is within the designed IFW. In case of high transmission-delay differences accurate timing-advance control has to be used [6], as it was also advocated in the GSM system [7]. Provided that these conditions are satisfied, a major benefit of the LAS codes is that they are capable of achieving a near-single-user performance without multi-user detectors.

The disadvantage of LAS codes is that the number of codes having an IFW is limited. For example, we will show that when we consider a spreading factor of 151, we only have 32 LAS codes exhibiting an IFW of width $3T_c$, where T_c is the chip duration. Furthermore, the auto-correlation and cross-correlation function of LAS codes typically exhibits a higher value outside the IFW than that of traditional random codes. More explicitly, when the LAS-CDMA system operates in an asynchronous manner, such as for example the third-generation W-CDMA system [1], [8], it will encounter more serious MAI and Multipath Interference (MPI) than traditional DS-CDMA.

Against this backdrop, the novel contribution of this paper is that we analytically investigate the performance of LAS-CDMA in a quasi-synchronous uplink scenario when communicating over a Nakagami-m channel and characterize its BER performance as a function of the number of resolvable paths L_p, the maximum delay difference τ_{max}, the number of

Manuscript received June 7, 2004; revised December 14, 2004; accepted February 16, 2005. The associate editor coordinating the review of this paper and approving it for publication was A. Conti. The work reported in this paper has formed part of the Wireless Enablers Work Area of the Core 2 Research Programme of the Virtual Centre of Excellence in Mobile & Personal Communications, Mobile VCE (www.mobilevce.com) whose funding support, including that of EPSRC, is gratefully acknowledged. Fully detailed technical reports on this research are available to Industrial Members of Mobile VCE.

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Digital Object Identifier 10.1109/TWC.2006.05027.

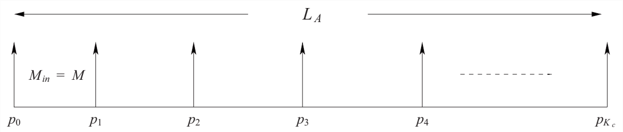

Fig. 1. Stylized pulse-positions in the LA(L_A, M, K_c) code having K_c number of binary ± 1 pulses, and exhibiting a minimum spacing of M chip durations between non-zero pulses, while having a total code length of L_A chips.

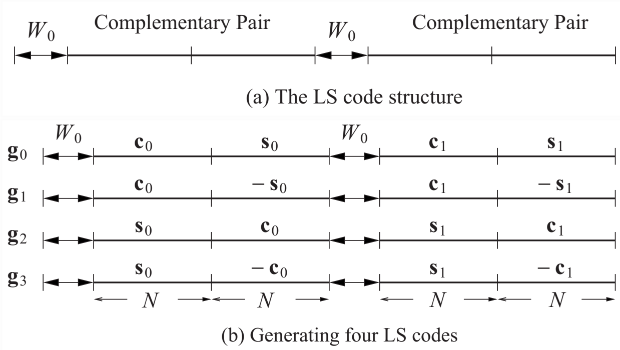

Fig. 2. Generating the LS(N, P, W_0) code using the $(P \times P) = (4 \times 4)$ Walsh-Hadamard matrix components $(1, 1, 1, 1)$ and $(1, -1, 1, -1)$.

users K and the Nakagami fading parameter m. Furthermore, we will comparatively study LAS-CDMA and traditional DS-CDMA systems.

This paper is organized as follows. Section II will introduce the family of LAS codes, while Section III will describe the LAS-CDMA system model. In Section IV we will characterize the BER performance of LAS-CDMA, and in Section V we will discuss our findings. Finally, in Section VI we will offer our conclusion.

II. Generation of LAS-Codes

A. LA Codes

LA codes [3], [4] belong to a family of ternary codes having elements of ± 1 or 0. Their maximum correlation magnitude is unity and they also exhibit an IFW. Let us denote the family of the K_c number of orthogonal ternary codes employing K_c number of binary ± 1 pulses by LA(L_A, M, K_c), which exhibit a minimum spacing of M chip durations between non-zero pulses, while having a total code length of L_A chips, as shown in Figure 1. All the codes corresponding to an LA code family share the same legitimate pulse positions. However, a specific drawback of this family of sequences is their relatively low duty ratio, quantifying the density of the non-zero pulses, since this limits the number of codes available and hence the number of users supported. Li [3] characterized the pulse positions of the LA code family, while Choi and Hanzo [9] further improved the achievable duty ratio of the LA codes. In the LAS-CDMA 2000 system [10], the LA codes used constitute a modified version of the LA(L_A, M, K_c)=LA(2552, 136, 17) code, where the $K_c = 17$ non-zero pulse positions, p_k, $k = 0, \cdots, 16$, are given by:

$$\{p_k\} = \{0, 136, 274, 414, 556, 700, 846, 994, 1144, 1296, 1450, 1606, 1764, 1924, 2086, 2250, 2416, 2552\}.$$

For a specific procedure concerning the design of LA-code based LAS codes, please refer to [9], where the associated correlation properties and the IFW width of the codes were also characterized.

B. Loosely Synchronized Codes

Apart from the LA codes of Section II-A, there exists another specific family of spreading codes, which also exhibits an IFW. Specifically, Loosely Synchronized (LS) codes [5] exploit the properties of the so-called orthogonal complementary sets [5], [11]. 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 2. More specifically, we generate a complementary code pair inserting W_0 number of zeros both in the center and at the beginning of the complementary pair, as shown in Figure 2(a), using the procedure described in [5]. As mentioned above, the polarity of the codes c_0 and s_0 seen in Figure **??** 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 $L_S = NP + 2W_0$.

Since the construction method of binary LS codes was described in [5], here we refrain from providing an indepth discourse and we will focus our attention on the employment of orthogonal complementary sets [12], [13] 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:

$$\mathbf{c}_1 = \tilde{\mathbf{s}}_0^*, \quad (1)$$

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

and where $\tilde{\mathbf{s}}_0$ denotes the reverse-ordered sequence and $-\mathbf{s}_0$ is the negated version of $\mathbf{s}_0$, respectively. Note that in Equation 1 and Equation 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 [5]. Having obtained a complementary pair and its corresponding mate pair, we may employ the construction method of [5] 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

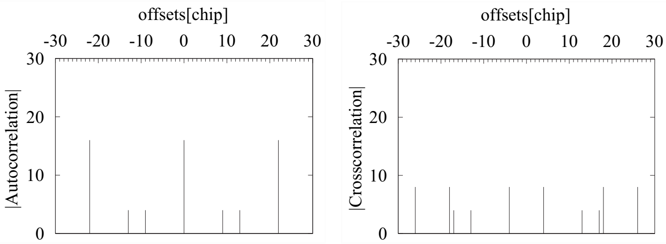

Fig. 3. Correlation magnitudes of the LS$(4,4,3)$ codes: (a) all four codes exhibit the same autocorrelation magnitude; (b) the crosscorrelation magnitudes of $\mathbf{g}_0$ and $\mathbf{g}_2$.

generated based on the complementary pair of [12]:

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

$$\mathbf{s}_0 \quad = + + - + . \tag{4}$$

Upon substituting Equation 1 and Equation 2 into Equation 3 and Equation 4, the corresponding mate pair can be obtained as:

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

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

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 2(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) codes 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 $\rho_p(\tau)$, $\rho_{jk}(\tau)$ of the codes belonging to the same set will be zero for $|\tau| \leq W_0 = 3$, as seen in Figure 3. Furthermore, the LS codes belonging to the four different sets are still orthogonal to each other at zero 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.

In this section we have demonstrated that the family of LS(N, P, W_0) codes can be constructed for almost any arbitrary code-length related parameter N by employing binary sequences. Having discussed the construction of LA and LS codes, let us now consider how LS codes are implanted at the non-zero pulse-positions of the LA codes for the sake of generating LAS codes.

C. Seeding LS Codes in LA Codes to Generate LAS codes

We observed in Section II-A that the main problems associated with applying LA codes in practical CDMA systems are related to their low duty ratio and to the resultant small number of available codes. A specific family of LAS codes [5] mitigates this problem by combining the LA codes of Section II-A and the LS codes of Section II-B. More specifically, LS codes are inserted between the non-zero pulses of the LA code sequence of Figure 1, in an effort to generate an increased number of spreading codes having an increased duty ratio, while maintaining attractive correlation properties. For example, in the LAS-2000 system [10] , the LS spreading codes are inserted into the LA code's zero space, as shown in Figure 4.

Let us denote the combined code generated from the LA(L_A,M,K) and LS(N,P,W_0) codes as LAS$(L_A, M, K_c; N, P, W_0)$, which is generated by employing the so-called absolute encoding method [3], [9]. For the sake of preserving the original IFW size of the constituent LS(N,P,W_0) code when combined with an LA(L_A,M,K) code employing the absolute encoding scheme, the length of the LS code - including W_0 number of trailing zeros - should not exceed the minimum pulse spacing M of the LA code, requiring that we have:

$$PN + 2W_0 \leq M. \tag{7}$$

In the LAS-CDMA 2000 system [10], a modified version of the LA(2552,136,17) and LS(4,32,4) codes was combined for the sake of generating the LAS$(L_A = 2552, L_S = M = 136, K_c = 17; N = 4, P = 32, W_0 = 4)$ code. As we mentioned in Section II-B, from the total set of $4P = 128$ LS codes, only 32 LS codes exhibited an IFW width of $\iota = 3$. Different permutations of the LA codes were employed in the different cells for the sake of mitigating the inter-cell interferences imposed [1]

III. LAS-CDMA System Model

A. Channel Model

The DS-CDMA signal experiences independent frequency-selective Nakagami-m fading. The complex low-pass equivalent representation of the Channel Impulse Response (CIR) encountered by the kth user is given by [14]:

$$h_k(t) = \sum_{l=0}^{L_p-1} h_{kl}\delta(t - lT_c)\exp(j\theta_{kl}), \tag{8}$$

where h_{kl} represents the Nakagami-distributed fading envelope, lT_c is the relative delay of the lth path of user k with respect to the main path, while L_p is the total number of resolvable multipath components. Furthermore, θ_{kl} is the uniformly distributed phase-shift of the lth multipath component of the channel and $\delta(t)$ is the Kronecker Delta-function. More explicitly, the L multipath attenuations $\{h_{kl}\}$ are independent Nakagami distributed random variables having a Probability Density Function (PDF) of [15]–[17]:

$$\begin{aligned} p(h_{kl}) &= M(h_{kl}, m_{kl}, \Omega_{kl}), \\ M(R, m, \Omega) &= \frac{2m^m R^{2m-1}}{\Gamma(m)\Omega^m} e^{(-m/\Omega)R^2}, \end{aligned} \tag{9}$$

[1]For example, the LA code [1,1,1,1,. . . ,-1,1-1] may be used in cell 1, wile the LA code [1,-1,1,-1,. . . ,1,1 1] can be used in cell 2. This specific permutation of the code-allocation mitigates the inter-cell interference, because using the same codes in different cells would result in an increased inter-cell interference due to their high cross-correlation.

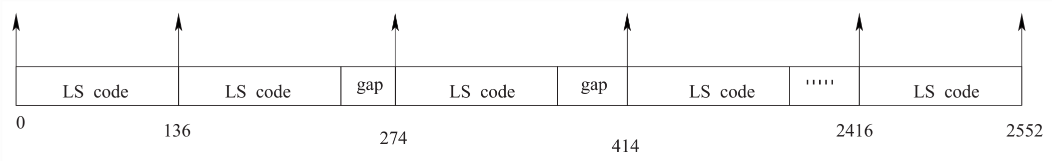

Fig. 4. LAS($L_A, M, K_c; N, P, W_0$)=LAS(2552,136,17;4,32,4) spreading, inserting the LS codes of Fig. 2 into the zero-space of the LA codes seen Fig. 1. The gap seen in the figure indicates that the $M = 136$-chip LS code does not always fill the spacing between the consecutive pulse of the constituent LA code.

where $\Gamma(\cdot)$ is the gamma function [14], and m_{kl} is the Nakagami-m fading parameter, which characterizes the severity of the fading for the l-th resolvable path of user k [18] . Specifically, $m_{kl} = 1$ represents Rayleigh fading, $m_{kl} \rightarrow \infty$ corresponds to the conventional Gaussian scenario and $m_{kl} = 1/2$ 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 $m_{kl} > 1$. The parameter Ω_{kl} in Equation 9 is the second moment of h_{kl}, i.e. we have $\Omega_{kl} = E[(h_{kl})^2]$. We assume a negative exponentially decaying Multipath Intensity Profile (MIP) given by $\Omega_{kl} = \Omega_{k0}e^{-\eta l}, \quad \eta \geq 0, l = 0, \ldots, L_p - 1$, where Ω_{kl} is the average signal strength corresponding to the first resolvable path and η is the rate of average power decay.

B. System Model

We support K asynchronous CDMA users in the system and each user is assigned a unique signature waveform $\mathbf{c}_k(t) = \sum_{i=0}^{G-1} c_{ki}\psi_{T_c}(t-iT_c)$, where G is the spreading gain and $\psi_{T_c}(t)$ is the rectangular chip waveform, which is defined over the interval $[0, T_c)$. Consequently, when the K users' signals are transmitted over the frequency-selective fading channel considered, the complex low-pass equivalent signal received at a given base station can be expressed as:

$$R(t) = \sum_{k=1}^{K}\sum_{l=0}^{L_p-1} \sqrt{2P}\mathbf{c}_k(t - lT_c - \tau_k) b_k(t - lT_c - \tau_k)h_{kl}\exp(j\theta_{kl}) + N(t), \quad (10)$$

where b_k is the transmitted bit of k user, while $N(t)$ is the complex-valued low-pass-equivalent AWGN having a double-sided spectral density of N_0, τ_k is the propagation delay of user k, while τ_k is assumed to be a random variable uniformly distributed in the range of $[0, \tau_{max}]$, and L_p is the total number of resolvable paths.

IV. BER Analysis

A. Random Spreading Code Based CDMA

Let the kth user be the user-of-interest and consider a receiver using de-spreading as well as multipath diversity combining. The conventional matched filter based RAKE receiver using MRC can be invoked for detection, where we assume that the RAKE receiver combines a total of L_r number of diversity paths, which may be more or possibly less than the actual number of resolvable components available at the current chip-rate. The value of L_r is typically restricted by the affordable receiver complexity.

Let us assume that we have achieved perfect time synchronization and that perfect estimates of the channel tap weights as well as phases are available. Then, after appropriately delaying the individual matched filter outputs, in order to coherently combine the L number of path signals with the aid of the RAKE combiner, the output Z_{kl} of the RAKE receiver's lth finger sampled at $t = T + lT_c + \tau_k$, can be expressed as:

$$Z_{kl} = D_{kl} + I_{kl}, \quad (11)$$

where D_{kl} represents the desired direct component, which can be expressed as:

$$D_{kl} = \sqrt{2P}T_s b_k[0]h_{kl}^2. \quad (12)$$

In Equation 12 $b_k[0]$ is the first bit transmitted by the kth user, where we have $b_k[0] \in \{+1, -1\}$. Hence, the interference plus noise term I_{kl} in Equation 11 can be expressed as:

$$I_{kl} = I_{kl}[S] + I_{kl}[M] + N_k, \quad (13)$$

where $I_{kl}[S]$ represents the multipath interference imposed by the user-of-interest, which can be expressed as:

$$I_{kl}[S] = \sqrt{2P}T_s h_{kl} \sum_{\substack{l_p=0 \\ l_p \neq l}}^{L_p-1} \frac{h_{kl_p}\cos\theta_{kl_p}}{T_s} \quad (14)$$
$$\times \int_0^{T_s} b_k[t - (l_p - l)T_c] \cdot c_k[t - (l_p - l)T_c]c_k[t]dt.$$

Furthermore, $I_{kl}[M]$ represents the multiuser interference inflicted by the $K - 1$ number of interfering signals, which can be expressed as:

$$I_{kl}[M] = \sqrt{2P}T_s h_{kl} \sum_{\substack{k'=1 \\ k' \neq k}}^{K} \sum_{l_p=0}^{L_p-1} \frac{h_{k'l_p}\cos\theta_{k'l_p}}{T_s}$$
$$\cdot \int_0^{T_s} b_{k'}[t - (l_p - l)T_c - (\tau_{k'} - \tau_k)] \times c_{k'}[t - (l_p - l)T_c - (\tau_{k'} - \tau_k)]c_k[t]dt. \quad (15)$$

In Equation 14 and Equation 15 the $\cos(\cdot)$ terms are contributed by the phase differences between the incoming carrier and the locally generated carrier used in the demodulation. Finally, the noise term in Equation 13 can be expressed as:

$$N_{kl} = h_{kl}\int_0^{T_s} n(t)c_k[t]\cos(2\pi f_c t + \theta_{kl})dt, \quad (16)$$

which is a Gaussian random variable having a zero mean and a variance of $N_0 T_s h_{kl}^2$, where $\{h_{kl}\}$ represents the path attenuations.

The MRC's decision variable Z_k, which is given by the sum of all the RAKE fingers' outputs, can be expressed as:

$$Z_k = \sum_{l=0}^{L_r-1} Z_{kl}. \quad (17)$$

Having obtained the decision variables of the MRC's output samples, let us now analyze the BER performance of the proposed LAS-CDMA system and benchmark it against a random code based CDMA system using hard-detection by invoking the often-used classic Gaussian approximation. We employ the standard Gaussian approximation and hence model both the multiuser interference and the self-interference terms of Equation 13 as an AWGN process having a zero mean and a variance equal to the corresponding variances. Consequently, for a given set of channel amplitudes $\{h_{kl}\}$ – according to the analysis of the previous sections – for the random spreading codes and BPSK modulation considered, the RAKE fingers' output signal Z_{kl} is a Gaussian distributed random variable having a mean of D_{kl}.

Let us first consider the random code based DS-CDMA system. The variance of the lth RAKE finger's output samples Z_{kl} for a given set of channel amplitudes $\{h_{kl}\}$ may be approximated as [18], [19]:

$$\begin{aligned}\sigma_{kl}^2 &= 2PT_s^2 \cdot \cdot \Omega_0 h_{kl}^2 \quad (18)\\ &\cdot \left[\frac{Kq(L_p,\eta)}{3G} + \frac{q(L_p,\eta)-1}{2G} + \left(\frac{2\Omega_0 E_b}{N_0}\right)^{-1}\right]\end{aligned}$$

where $E_b = PT_s$ is the energy per bit and we have $q(L_p,\eta) = \sum_{l=0}^{L_p-1} e^{-\eta l}$. Furthermore, the MRC's output sample Z_k can be approximated by an AWGN variable having a mean value of $\mathrm{E}[Z_k] = \sum_{l=0}^{L_r-1} D_{kl}$ and a variance of $\mathrm{Var}[Z_k] = \sum_{l=0}^{L_r-1} \sigma_{kl}^2$ [16], [19], where we have:

$$\mathrm{E}[Z_k] = \sum_{l=0}^{L_r-1} \sqrt{2P} T_s b_k[0] h_{kl}^2, \quad (19)$$

$$\begin{aligned}\mathrm{Var}[Z_k] &= 2PT_s^2 \cdot \Omega_0 \sum_{l=0}^{L_r-1} h_{kl}^2 \quad (20)\\ &\left[\frac{Kq(L_p,\eta)}{3G} + \frac{q(L_p,\eta)-1}{2G} + \left(\frac{2\Omega_0 E_b}{N_0}\right)^{-1}\right]\end{aligned}$$

Therefore, the BER using BPSK modulation conditioned on a set of fading attenuations $\{h_{kl}, l = 0, 1, \ldots, L_r - 1\}$ can be expressed as:

$$P_b(\gamma) = Q\left(\sqrt{\frac{(\mathrm{E}[Z_k])^2}{\mathrm{Var}[Z_k]}}\right) = Q\left(\sqrt{\sum_{l=0}^{L_r-1} 2\gamma_l}\right), \quad (21)$$

where $Q(x)$ represents the Gaussian Q-function, which can also be represented in its less conventional form as [18], [19] $Q(x) = \frac{1}{\pi}\int_0^{\pi/2} \exp\left(-\frac{x^2}{2\sin^2\theta}\right) d\theta$, where $x \geq 0$. Furthermore, $2\gamma_l$ in Equation 21 represents the output Signal to Interference plus Noise Ratio (SINR) at the lth finger of the RAKE receiver and γ_l is given by:

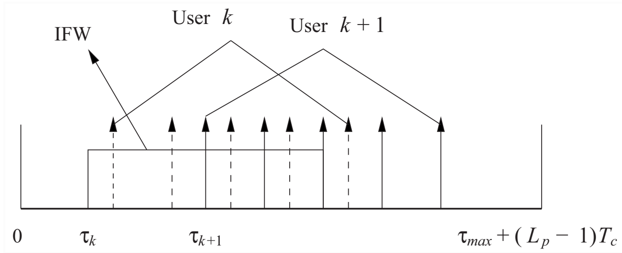

Fig. 5. Illustration of the interference suppression capability of the LAS codes for the first finger of the RAKE receiver, when the width of the IFW is $\iota = 2$.

$$\gamma_l = \gamma_c \cdot \frac{h_{kl}^2}{\Omega_0}. \quad (22)$$

Let us now substitute Equation 22 into Equation 21 and Equation 19 as well as Equation 20 also into Equation 21. We can see then that the expressions under the square-root functions must be equal, which allows us to express γ_c as follows:

$$\gamma_c = \left[\frac{q(L_p,\eta)-1}{G} + \frac{2Kq(L_p,\eta)}{3G} + \left(\frac{\Omega E_b}{N_0}\right)^{-1}\right]^{-1}. \quad (23)$$

Therefore, for a random code based DS-CDMA system we may argue that all paths of the interfering users will impose MAI on the reference user, and all multipath components of the reference user will additionally inflict MPI upon the reference user's desired signal.

B. LAS Spreading Code Based CDMA

Let us now consider the interference suppression achieved with the aid of the IFW of LAS codes. In this scenario, only the paths arriving from the interfering users outside the IFW will inflict MAI upon the reference user, and this user's own delayed paths arriving outside the IFW will additionally impose MPI. Observe in Figure 5, for example that both the kth user and the $(k+1)$st user encounter five multipath components. When we considered the kth user's first RAKE receiver finger and an IFW of $\iota = 2$, only those two paths of the kth user will impose MPI, which fall outside the code's IFW. Similarly, the $(k+1)st$ user imposed only one interfering path on the kth user's first RAKE finger's decision variable, since four of the five path fall within the IFW. Hence, we will investigate the interference suppression capability of the IFW, as shown in a stylized fashion in Figure 5.

As argued above, in the context of LAS-CDMA, not all the $(L_p - 1)$ paths will impose MPI. Specifically, for the lth finger's output sample Z_{kl} of the RAKE receiver only the rays arriving outside the IFW will impose MPI, where we have:

$$|l_p - l| > \iota. \quad (24)$$

Hence, for the lth finger of the RAKE receiver, the corresponding MPI term $I_{kl}[S]$ can be expressed as:

$$I_{kl}[S] = \sqrt{2P}T_s h_{kl} \sum_{\substack{l_p=0 \\ |l_p-l|>l}}^{L_p-1} \frac{h_{kl_p}\cos\theta_{kl_p}}{T_s} \int_0^{T_s} b_k[t-(l_p-l)T_c]\cdot c[t-(l_p-l)T_c]c[t]dt. \quad (25)$$

We introduce the notation $\xi = l_p - l$ for the path length difference between the l_pth and the lth path, and when the delay spread exceeds ι, the integral seen in Equation 25 can be expressed as:

$$\frac{1}{T_s}\int_0^{T_s} b_k[t-\xi T_c]\cdot c[t-\xi T_c]c[t]dt = \rho_{kk}(\xi)b_k[-1] + \varrho_{kk}(\xi)b_k[0], \quad (26)$$

where we define the partial auto-correlation $\rho_{kk}(\xi)$ and $\varrho_{kk}(\xi)$ as:

$$\rho_{kk}(\xi) = \frac{1}{T_s}\int_0^{|\xi|T_c} c_k[t]c_k[t-|\xi|T_c]dt, \quad (27)$$

$$\varrho_{kk}(\xi) = \frac{1}{T_s}\int_{|\xi|T_c}^{T_s} c_k[t]c_k[t-|\xi|T_c]dt, \quad (28)$$

and we have $E\{h_{kl_p}^2\} = \Omega_0 e^{-\eta l_p}$. Hence the lth finger's MPI variance can be expressed as:

$$\text{Var}\{I_{kl}[S]\} = 2P\cdot T_s^2 h_{kl}^2 \cdot \sum_{\substack{l_p=0 \\ |l_p-l|>\iota}}^{L_p-1} \Omega_0 e^{-\eta l_p}[\rho_{kk}^2(\xi) + \varrho_{kk}^2(\xi)]. \quad (29)$$

Similarly, for the multiuser interference term $I_{kl}[M]$ of Equation 13, only the rays arriving outside the IFW will impose MAI in Equation 15. Therefore, only the paths satisfying

$$|(l_p - l)T_c + (\tau_{k'} - \tau_k)| > \iota T_c, \quad (30)$$

will impose MAI. Let us now define $\xi = Int\{(l_p - l) + (\tau_{k'} - \tau_k)/T_c\}$, where $Int\{x\}$ denotes the integer part of the arbitrary value x, while $x - Int\{x\}$ represents the non-integer part of the value x, hence we have:

$$\xi = Int\{(l_p - l)T_c + (\tau_{k'} - \tau_k)/T_c\}, \quad (31)$$

$$\tau_c = (l_p - l)T_c + (\tau_{k'} - \tau_k) - \xi T_c. \quad (32)$$

Furthermore, we define the partial cross-correlation of the spreading codes $\rho_{k'k}(\xi)$ and $\varrho_{k'k}(\xi)$ as:

$$\rho_{k'k}(\xi) = \frac{1}{T_s}\int_0^{|\xi|T_c} c_k(t)c_{k'}(t-|\xi|T_c)dt, \quad (33)$$

$$\varrho_{k'k}(\xi) = \frac{1}{T_s}\int_{|\xi|T_c}^{T_s} c_k(t)c_{k'}(t-|\xi|T_c)dt. \quad (34)$$

According to [20], the integral in Equation 15 can be expressed as:

$$\begin{aligned}\frac{1}{T_s}\cdot \int_0^{T_s} & b_{k'}[t-(l_p-l)T_c-(\tau_{k'}-\tau_k)] \quad (35)\\ \cdot\ & c_{k'}[t-(l_p-l)T_c-(\tau_{k'}-\tau_k)]c[t]dt \\ =\ & b_{k'}[-1]\{\rho_{k'k}(\xi)\hat{R}_\psi(\tau_c) + \rho_{k'k}(\xi+1)R_\psi(\tau_c)\} \\ +\ & b_{k'}[0]\{\varrho_{k'k}(\xi)\hat{R}_\psi(\tau_c) + \varrho_{k'k}(\xi+1)R_\psi(\tau_c)\},\end{aligned}$$

where $\hat{R}_\psi(\tau_c)$ and $R_\psi(\tau_c)$ are the partial autocorrelation functions of the chips waveform, which are defined as [20]:

$$R_\psi(\tau_c) = \int_0^{\tau_c} \psi_{T_c}(t)\psi_{T_c}(t+T_c-\tau_c)dt, \quad (36)$$

$$\hat{R}_\psi(\tau_c) = \int_{\tau_c}^{T_c} \psi_{T_c}(t)\psi_{T_c}(t-\tau_c)dt. \quad (37)$$

For a rectangular chip waveform we have $R_\psi(\tau_c) = \tau_c$ and $\hat{R}_\psi(\tau_c) = (T_c - \tau_c)$.

For convenience, we define the asynchronous partial cross-correlations $\hat{\rho}_{k'k}(\tau_c)$ and $\hat{\rho}_{k'k}(\tau_c)$ as:

$$\hat{\rho}_{k'k}(\tau_c) = \rho_{k'k}(\xi)\hat{R}_\psi(\tau_c) + \rho_{k'k}(\xi+1)R_\psi(\tau_c), \quad (38)$$

$$\hat{\varrho}_{k'k}(\tau_c) = \varrho_{k'k}(\xi)\hat{R}_\psi(\tau_c) + \varrho_{k'k}(\xi+1)R_\psi(\tau_c), \quad (39)$$

and we assume that τ_k and $\tau_{k'}$ are random variables uniformly distributed in $[0, \tau_{max}]$. Therefore, the average variance of the MAI imposed by the k'th user at the lth finger of the RAKE receiver can expressed as:

$$V_{k'k}(l) = 2PT_s h_{kl}^2 \int_0^{\tau_{max}}\int_0^{\tau_{max}} \frac{1}{\tau_{max}^2}\frac{1}{\tau_{max}} \times \sum_{\substack{l_p=0 \\ |(l_p-l)T_c+(\tau_{k'}-\tau_k)|>\iota T_c}}^{L_p-1} \Omega_0 e^{-\eta l_p}\left[\hat{\rho}_{k'k}^2(\tau_c) + \hat{\varrho}_{k'k}^2(\tau_c)\right] d\tau_k d\tau_{k'}. \quad (40)$$

Finally, the variance of the MAI term $I_{kl}[M]$ of Equation 13 can be expressed as:

$$\text{Var}(I_{kl}[M]) = \sum_{\substack{k'=1 \\ k'\neq k}}^{K} V_{k'k}(l). \quad (41)$$

For the sake of convenient comparison to the random code based system, we will introduce the MPI and MAI interference reduction factors of $\Upsilon_S(l)$ and $\Upsilon_M(l)$, respectively, which are defined as:

$$\Upsilon_S(l) = \frac{2G\cdot \text{Var}\{I_{kl}[S]\}}{2PT_s^2\Omega_0 h_{kl}^2}, \quad (42)$$

$$\Upsilon_M(l) = \frac{3G\cdot \text{Var}\{I_{kl}[M]\}}{2PT_s^2\Omega_0 h_{kl}^2}. \quad (43)$$

Hence, upon substituting Equation 29 into Equation 42, the corresponding MPI reduction factor $\Upsilon_S(l)$ can be expressed as:

$$\Upsilon_S(l) = 2G\cdot \sum_{\substack{l_p=0 \\ |l_p-l|>\iota}}^{L_p-1} e^{-\eta l_p}[\rho_{kk}^2(\xi) + \varrho_{kk}^2(\xi)]. \quad (44)$$

Similarly, upon substituting Equation 41 into Equation 43, the corresponding MUI reduction factor $\Upsilon_M(l)$ can be expressed as:

$$\Upsilon_M(l) = \frac{3G}{K}\sum_{\substack{k'=1 \\ k'\neq k}}^{K} \int_0^{\tau_{max}}\int_0^{\tau_{max}} \frac{1}{\tau_{max}}\frac{1}{\tau_{max}} \times \sum_{\substack{l_p=0 \\ |(l_p-l)T_c+(\tau_{k'}-\tau_k)|>\iota T_c}}^{L_p-1} e^{-\eta l_p}(\hat{\rho}_{k'k}^2(\tau_c) + \hat{\varrho}_{k'k}^2(\tau_c)) d\tau_k d\tau_{k'}. \quad (45)$$

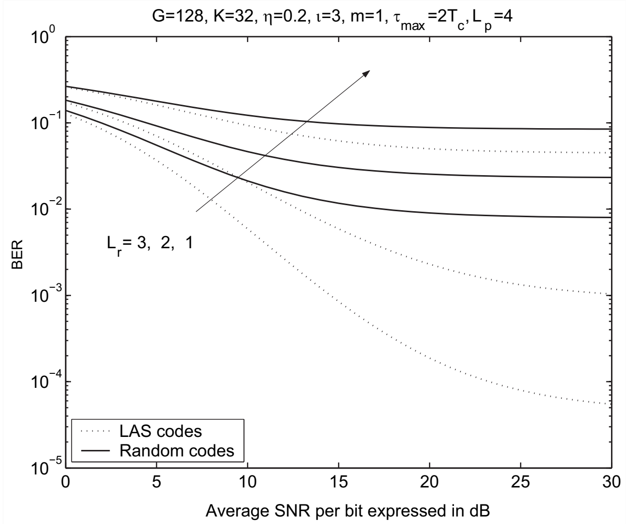

Fig. 6. BER versus channel SNR performance comparison of random code based classic CDMA and LAS CDMA based on Equation 53, when communicating over a Nakagami-m channel having $m = 1$. $K = 32$ users were supported, and the negative exponential MIP decay factor was $\eta = 0.2$. The IFW width was $\iota = 3$, and the maximum delay difference was $\tau_{max} = 2T_c$. The number of resolvable paths was $L_p = 4$ and the RAKE receiver combined $L_r = 1, 2, 3$ paths, respectively.

Having obtained $\Upsilon_S(l)$ and $\Upsilon_M(l)$, now we are ready to calculate the γ_c and the average bit error probability, which we synonymously refer to as the BER $P_b(E)$.

For the LAS-CDMA system, given a fading attenuation set of $\{h_{kl}, l = 0, 1, \ldots, L_r - 1\}$, the BER is given by:

$$P_b(\gamma) = Q\left(\sqrt{\sum_{l=0}^{L_r-1} 2\gamma_l}\right), \tag{46}$$

where γ_l may be expressed as:

$$\gamma_l = \gamma_c \cdot \frac{h_{kl}^2}{\Omega_0}. \tag{47}$$

Following a similar approach to that used in the context of the random code based CDMA system, the corresponding γ_c expression can be formulated as:

$$\gamma_c = \left[\frac{\Upsilon_S(l)}{G} + \frac{2K\Upsilon_M(l)}{3G} + \left(\frac{\Omega E_b}{N_0}\right)^{-1}\right]^{-1}. \tag{48}$$

C. Bit Error Probability Analysis

The average BER, $P_b(E)$ can be obtained by the weighted averaging of the conditional BER expression of Equation 21 and Equation 46 over the joint PDF of the instantaneous SNR values corresponding to the L_r multipath components $\{\gamma_l : l = 1, 2, \ldots, L_r\}$. Since the random variables $\{\gamma_l : l = 1, 2, \ldots, L_r\}$ are assumed to be statistically independent, the average BER expression of Equation 21 and Equation 46 can be formulated as [21]:

$$P_b(E) = \frac{1}{\pi}\int_0^{\pi/2} \prod_{l=0}^{L_r-1} I_l(\overline{\gamma}_l, \theta)\, d\theta, \tag{49}$$

where we have

$$I_l(\overline{\gamma}_l, \theta) = \int_0^{\infty} \exp\left(-\frac{\gamma_l}{\sin^2\theta}\right) p_{\gamma_l}(\gamma_l) d\gamma_l. \tag{50}$$

Since $\gamma_l = \gamma_c \cdot \frac{(h_l)^2}{\Omega_0}$ and h_l obeys the Nakagami-m distribution characterized by Equation 9, it can be shown that the PDF of γ_l can be formulated as:

$$p_{\gamma_l}(\gamma_l) = \left(\frac{m_l}{\overline{\gamma}_l}\right)^{m_l} \frac{\gamma^{m_l-1}}{\Gamma(m_l)} \exp\left(-\frac{m_l\gamma_l}{\overline{\gamma}_l}\right), \quad \gamma_l \geq 0, \tag{51}$$

where $\overline{\gamma}_l = \gamma_c e^{-\eta l}$ for $l = 0, 1, \ldots, L_r - 1$.

Upon substituting (51) into (50) it can be shown that we have [18]:

$$I_l(\overline{\gamma}_l, \theta) = \left(\frac{m_l \sin^2\theta}{\overline{\gamma}_l + m_l \sin^2\theta}\right)^{m_l}. \tag{52}$$

Finally, upon substituting (52) into (49), the average BER of the both the random and LAS-code based CDMA system can be written as:

$$P_b(E) = \frac{1}{\pi}\int_0^{\pi/2} \prod_{l=0}^{L_r-1} \left(\frac{m_l \sin^2\theta}{\overline{\gamma}_l + m_l \sin^2\theta}\right)^{m_l} d\theta. \tag{53}$$

V. Performance of LAS DS-CDMA

In our investigations we compared a traditional and a LAS-code based CDMA system, both of which have the same chip-rate and bandwidth. However, both their effective spreading gain as well as their correlation functions are different. Therefore these two systems are affected differently by the MAI and MPI and hence their expected performance will differ. In the LAS-CDMA 2000 system, the LA(2552,136,17) and LS(4,32,4) codes are combined, as seen in Figure 4. More explicitly, the total length of the LS(N, P, W_0) =LS(4,32,4) code is $L_s = NP + 2W_0 = 136$ chips, which is incorporated into the LA(L_A, M, K)=LA(2552,136,17) code, again, as seen in Figure 4, for the sake of creating the LAS$(L_A, M, K_c; N, P, W_0)$ = LAS(2552,136,17;4,32,4) code. Since this LAS code has certain zero-valued gaps after inserting the LS code into the LA code as portrayed in Figure 4, as well as the $2W_0$ number of zeros constituting the IFW, its spreading factor may be calculated by simply noting that each bit to be transmitted is spread by one of the constituent LS(4,32,4) codes. Although the length of this LS code is $L_S = NP + 2W_0 = 136$, the effective spreading gain of the LAS code is identical to that of its constituent LS codes, namely $G_{\text{LAS}} = G_{\text{LS}} = 128$. By contrast, a traditional random code based CDMA system having the same $L_A = 2552$ chips would have a higher spreading gain, since it does not have any zero-valued gaps, nor has an IFW. Hence the corresponding spreading gain becomes $G_{\text{Random}} = 2552/17 = 151$, since in Figure 4 $K = 17$ bits are mapped to $L_A = 2552$ chips. This spreading gain difference was taken into account in our results, again, assuming that the LAS-code and random code based systems considered have the same bandwidth. We will compare these two system's performance based on these two different effective spreading gains. For simplicity's sake, we assume that all paths have the same Nakagami fading parameter, *i.e.* we have $m_l = m,\ l = 0, \ldots, L_r - 1$.

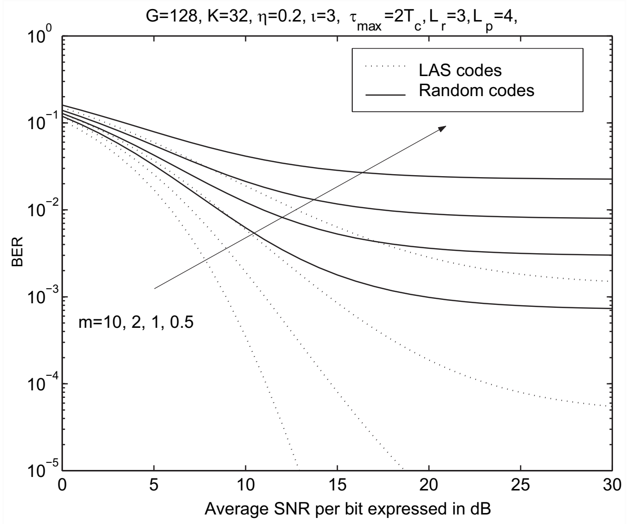

Fig. 7. BER versus channel SNR performance comparison of random code based classic CDMA and LAS CDMA based on Equation 53, when communicating over different Nakagami fading channels. $K = 32$ users were supported and the negative exponential MIP decay factor was $\eta = 0.2$. The IFW width was $\iota = 3$, and the maximum delay difference was $\tau_{max} = 2T_c$. The number of resolvable paths was $L_p = 4$ and the RAKE receiver combined $L_r = 3$ paths.

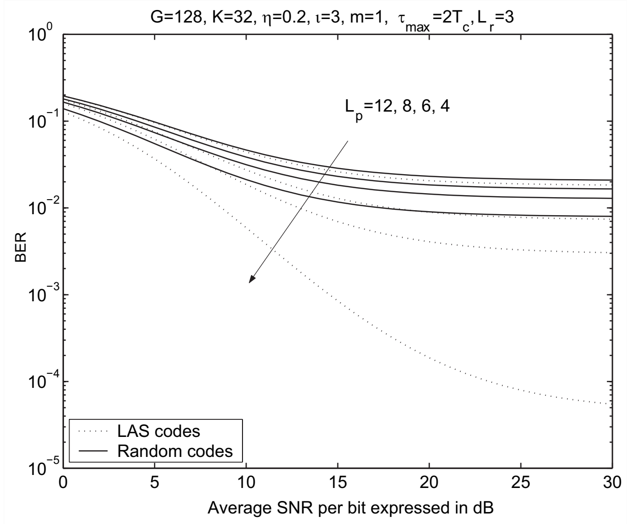

Fig. 8. BER versus channel SNR performance comparison of random code based classic CDMA and LAS CDMA based on Equation 53, when communicating over a dispersive Rayleigh-fading channel having $m = 1$. The number of resolvable paths was $L_p = 4, 6, 8, 12$, respectively. $K = 32$ users were supported, and the negative exponential MIP decay factor was $\eta = 0.2$. The IFW width was $\iota = 3$ and the maximum delay difference was $\tau_{max} = 2T_c$. The RAKE receiver combined $L_r = 3$ paths.

In Figure 6, we assumed that the LAS-CDMA system operated in a quasi-synchronous scenario, which can be achieved for example with the aid of a Global Positioning System (GPS) assisted synchronization protocol. We assume that we have a maximum propagation delay of $\tau_{max} = 2T_c$. The channel's delay spread is negative exponentially distributed in the range of $[0.3, 3]\mu s$ [22], 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 $L_p = \lfloor \frac{\tau_{ch}}{T_c} \rfloor + 1 = 4$, where $\tau_{ch} = 3\mu s$. Both the random code based CDMA system and the LAS-CDMA system supported $K = 32$ users, and the width of the IFW was $3T_c$ for the LAS-CDMA system, *i.e* we had $\iota = 3$. From Figure 6, we can observe that the LAS-CDMA system has a significantly better BER performance than the traditional DS-CDMA system, when communicating over a quasi-synchronous channel, provided that both these two systems combine the same $L_r = 1, 2, 3 \leq L_p$ number of resolvable paths, respectively. The reason that the LAS-CDMA scheme outperforms the traditional DS-CDMA system is that the MAI and MPI is reduced, as a benefit of using LAS codes.

Figure 7 exhibits the performance of these two systems 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 $m \rightarrow \infty$ corresponds to an AWGN channel. We can observe from Figure 7 that the LAS-CDMA system exhibited a significantly better BER performance than the traditional DS-CDMA system, regardless of the value of m. More specifically, provided that all these uplink users are in a quasi-synchronous state, $i.e$ we have $\tau_{max} = 2T_c$ and $L_p = 4$, the LAS-CDMA scheme outperformed the traditional DS-CDMA system, when communicating over different Nakagami multipath fading channels.

Figure 8 shows the performance of these two systems for transmission over different dispersive channels having $L_p = 4 \ldots 12$ resolvable multipath components and assuming that $L_r = 3$ of these components were combined by the RAKE receiver. We can observe from Figure 8 that when the channel became more dispersive, the LAS-CDMA system's performance was significantly degraded and its gain over the traditional DS-CDMA system was eroded. Nonetheless, the LAS-CDMA scheme still outperformed the traditional DS-CDMA system, provided that the users were in a quasi-synchronous state, *i.e* when we had $\tau_{max} = 2T_c$. However, when L_p was increased to 12, the LAS-CDMA system retained only a moderate gain over the traditional DS-CDMA arrangement even if it operated in a quasi-synchronous scenario. 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 LS codes outside the IFW is higher than that of the random codes. Hence, when L_p is high, LAS-CDMA inevitably encounters serious MAI and MPI. However, for a high-chip-rate system we may consider the employment of MC DS-CDMA [1], which is capable of ensuring that each subcarrier encounters only $L_p = 4$ resolvable paths. In this scenario, LAS MC DS-CDMA may be expected to retain its ability to effectively suppress both the MAI and MPI.

In Figure 9 we can observe that as the maximum propagation delay τ_{max} increases, the performance of LAS-CDMA significantly degrade. When we have $\tau_{max} \geq 10T_c$, the LAS-CDMA system's performance becomes even worse than that of traditional DS-CDMA. This is because the insertion of zeros in the LAS codes reduces the effective spreading gain of the LAS-CDMA system and when the propagation delay τ_{max} increases, its MAI suppression capability will be inevitably reduced. Hence we may conclude that LAS-CDMA systems

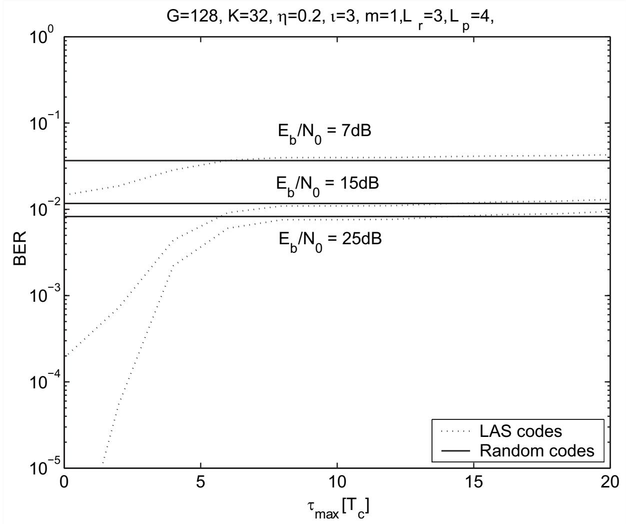

Fig. 9. BER performance comparison of random code based classic CDMA and LAS CDMA based on Equation 53 as a function of the maximum delay difference τ_{max}, when communicating over a dispersive Rayleigh-fading channel having $m = 1$. $K = 32$ users were supported, and the negative exponential MIP decay factor was $\eta = 0.2$. The IFW width was $\iota = 3$. The number of resolvable path was $L_p = 4$ and the RAKE receiver combined $L_r = 3$ paths.

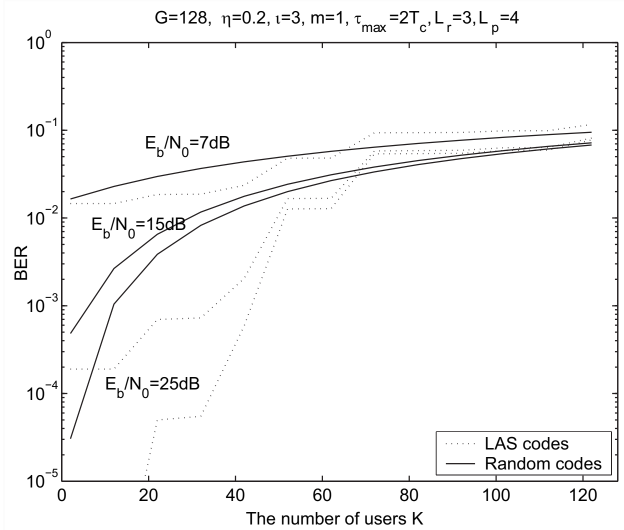

Fig. 10. BER performance comparison of random code based classic CDMA and LAS CDMA based on Equation 53 as a function of the number of users K supported, when communicating over a dispersive Rayleigh fading channel having $m = 1$. The negative exponential MIP decay factor was $\eta = 0.2$, and the number of resolvable paths was $L_p = 4$. The RAKE receiver combined $L_r = 3$ paths.

are suitable for operating in a quasi-synchronous CDMA environment.

Stańczak *et al.* [5] concluded that the width ι of the IFW and the number of users K will obey $K(\iota + 1) \leq G$. In the LAS-CDMA 2000 system [10], when the number of users K exceeds 32, or more explicitly, when all the four different sets of LS codes mentioned in Section II-B are employed, the width of the IFW will be reduced to zero for the codes belonging to different sets associated with different rows of the corresponding WH matrix. For example, when we have $K = 128$, all the 128 users' signals are orthogonal to each other in case of perfect synchronization. However, the width of the IFW of LS codes belonging different LAS code sets generated using different rows of the WH matrix becomes zero. In this scenario, serious MAI and MPI will be incurred, when the inter-set orthogonality of the LAS-codes generated with the aid of different rows of the WH matrix is destroyed by the multipath channel. Therefore, the performance of LAS-CDMA will significantly degrades, when the number of users supported becomes $K > 32$. In Figure 10, we can observe that for $K \geq 60$ the LAS-CDMA system will have no advantage in comparison to the traditional DS-CDMA scheme, or even may perform worse than the traditional DS-CDMA arrangement, although all users operate in a quasi-synchronous manner. Hence, we may conclude that the employment of LAS-CDMA is beneficial in low-user-load scenarios, where the delay spread is also limited.

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 11 demonstrated the achievable performance of Single-carrier LAS DS-CDMA and MC LAS CDMA for a single-carrier $3.84M$chips/s system. From this figure we may conclude that the MC LAS DS-CDMA is capable of achieving the best performance trade-off by selecting the optimum number of subcarriers U according to the channel's delay dispersion τ_{ch} and the delay difference τ_{max}. For example, we may conclude from Figure 11 that the $U = 8$ MC LAS DS-CDMA system exhibited the best trade-off in a scenario of $\tau_{ch} = 3\mu s$ and $\tau_{max} = 5\mu s$

VI. Conclusion

In conclusion, LAS-CDMA was investigated, which exhibited a significantly better performance than the traditional random code based DS-CDMA system in a relatively low-chip-rate scenario, provided that all users operate in a quasi-synchronous manner. As the chip-rate increases, the number of resolvable paths also increases, which will impose a performance degradation. Hence, as suggested in Figure 8, LAS-CDMA may be expected to have a moderate performance gain over the traditional DS-CDMA system, when L_p is in excess of 12, but as the maximum delay difference τ_{max} increases, the performance of LAS-CDMA degrades significantly. Furthermore, the limited number of available LAS codes having a certain IFW width suggests that the employment of LAS-CDMA is beneficial in a low-user-load scenario. Our further research will investigate the performance of multi-carrier LAS-CDMA.

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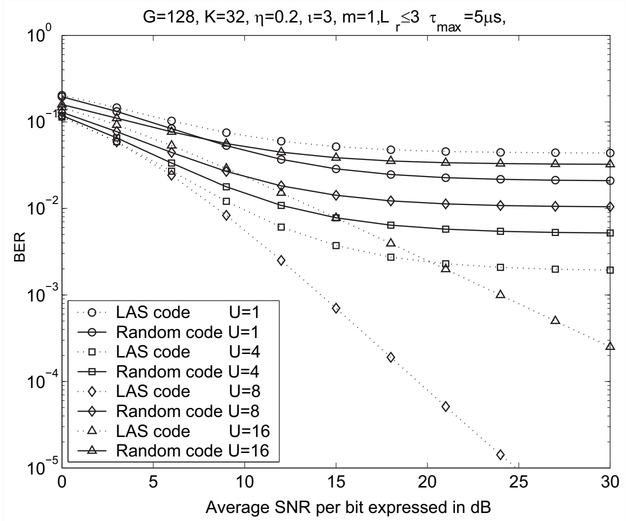

Fig. 11. BER versus E_b/N_0 performance comparison of two MC DS-CDMAs based on Equation 53 when we considered LAS codes and random spreading codes. The RAKE receiver was configured for combining $L_r \leq 3$ paths' energy. The channel dispersion was $\tau_{ch} = 3\mu s$, Hence the number of resolvable multipath components for $U = 1, 4, 8, 16$ was $L_p = 12, 3, 2, 1$, and the dispersion of the propagation environment considered was $\tau_{max} = 5\mu s$.

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