Performance of an LMS Type Receiver for Interference Suppression in Long Code DS-CDMA on Rayleigh Fading Channels*

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Abstract

We consider interference suppression for direct sequence code division multiple access systems when *long codes* are used for spreading. In particular, we evaluate the performance of a Least Mean Square (LMS) type adaptive receiver on AWGN and correlated Rayleigh fading channels. The receiver uses a chip matched filter followed by an adaptive equalizer structure which performs despreading and interference suppression. The receiver requires only the knowledge of the long code and timing of the desired user. Probability of error performance of the long code LMS (LC-LMS) receiver is estimated for varying degrees of near-far ratio (NFR), mobile user speed (equivalently, the Doppler bandwidth), and long code period (expressed in number of information bits). Performance results indicate that the given adaptive receiver outperforms the conventional linear correlation receiver by a considerable margin in high NFR and low mobility (i.e., small Doppler bandwidth) scenarios.

1 Introduction

Direct sequence code division multiple access (DS-CDMA) has recently received considerable attention as an attractive alternative to traditional multiple access techniques for use in mobile cellular as well as other multiuser systems. The main drawback associated with DS-CDMA is the *near-far problem*, whereby a weak signal from a distant user is overwhelmed by a strong signal from a nearer interferer. One way to deal with inequality in the received signal powers is to use transmitter power control [1]. If the transmitter power is updated at a moderate rate, power control can at best account for differences in received power due to propagation losses and most shadowing effects. In order to overcome multipath fading via power control, mobile transmitters would have to adjust their powers at rates on the order of a magnitude greater than the Doppler bandwidth.

Another approach considered to overcome received power inequality is to use *near-far resistant detectors* at the base station [2]. Many researchers have investigated this vastly rich topic, and proposed several optimum/suboptimum receiver structures and analyzed their performance in comparison with the conventional linear correlation receiver in both non-fading and fading scenarios [3]–[16]. Some of these proposed receivers [3],[5],[6],[8] need exact knowledge of the propagation delays and phase information of all the users, which are unknown *a priori*. Hence such receivers will be sensitive to the delay/phase estimation errors, particularly in a near-far scenario [10]. On the other hand, adaptive multiuser detection techniques are more attractive because of their lower sensitivity to estimation errors [11],[12],[13],[14]. Most of these adaptive receivers require transmission of a training sequence, before the transmission of actual data bits, to enable the receiver filter to adapt to the optimum coefficients. A blind adaptation algorithm which does not require training bits is addressed in [15].

We note that many of the interference mitigation schemes treated in the literature mentioned above are developed under the periodic sequence assumption, in which the signature sequences employed to spectrally spread the data symbols are periodic with period equal to one data symbol duration. Such sequences are known as *short sequences* (or short codes). Although analytically friendly, short sequences are not preferred in applications where low probability of intercept is a desired feature (e.g., as in military communications). Spreading sequences whose periods span multiple symbols are known as *long sequences* (or long codes). Long code transmissions are less vulnerable to intercept than short codes. The use of certain types of multiuser detection techniques becomes cumbersome for long codes, as in that case, the cross correlations vary at the data rate [2]. Recently in [16], Wong *et al* developed a linear receiver based on minimum-mean-squared error (MMSE) principles for long code DS-CDMA.

In this paper, we study an LMS-type adaptive receiver capable of DS-CDMA multiple access interfer-

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ence (MAI) suppression when *long codes* are used. The considered long code LMS (LC-LMS) receiver uses a chip matched filter followed by an adaptive equalizer structure which performs despreading and interference suppression. The receiver can be conceptually viewed as a bank of L_o adaptive chip-interval spaced equalizers, each adapting its coefficients to a different *code segment* of the long code, where L_o is the period of the long code expressed in number of bits. Consequently, the number of bits required for training will be L_o times that of the short code. However, the receiver can be realized using a single adaptive equalizer structure using additional memory to store the coefficients of the code segments which are not operated on during a given bit interval. We evaluate the performance of the above LC-LMS receiver in AWGN and correlated Rayleigh fading channels at varying near-far ratios using detailed simulations. We also evaluate the performance dependence of the LC-LMS algorithm on user mobility speeds and the periodicity of the long code. We show that the considered LC-LMS adaptive receiver outperforms the conventional linear correlation receiver by a considerable margin in high NFR and low mobility (i.e., small Doppler bandwidth) scenarios.

The rest of the paper is organized as follows. Section 2 provides a description of the system model, including the correlated Rayleigh fading channel and the LC-LMS receiver. In Section 3, the performance results are discussed. Section 4 provides the conclusions.

2 System Description

A standard model for an asynchronous binary phase shift keyed DS-CDMA system is considered in this work. Assume there are $K+1$ users in the system. The k^{th} user, $0 \leq k \leq K$, generates a stream of equiprobable data bits $b_j^{(k)}$, $b_j \in \{+1, -1\}$. Each user is assigned a randomly generated *long code sequence* for spreading the data bits. The k^{th} user is provided a long sequence $c^{(k)}$, given by

$$c^{(k)} = \left(..., c_0^{(k)}, c_1^{(k)}, ..., c_{N-1}^{(k)}, ...\right). \qquad (1)$$

where $c_i^{(k)} \in \{+1, -1\}$, and N is the number of chips per bit interval (also known as the processing gain). The period of the code sequence, $c^{(k)}$, is assumed to span L_o bits, where L_o is an integer. It is noted that

$$\begin{aligned} L_o &> 1 \quad \text{for long codes} \\ &= 1 \quad \text{for short codes.} \end{aligned} \qquad (2)$$

Because the period of the long code is NL_o chips, a given chip within a code period will repeat every L_oN chips (see Fig. 1). In other words,

$$c_i = c_{jNL_o+i}, \quad \text{where } j \text{ is an integer.} \qquad (3)$$

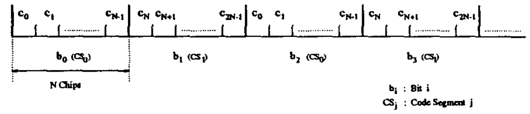

Figure 1: Long Code Spreading for $L_o = 2$.

The sequence $c^{(k)}$ is used to generate the spectrally spread signal given by

$$a_k(t) = \sum_{i=-\infty}^{\infty} b_{\lfloor i/N \rfloor}^{(k)} c_i^k g(t - iT_c) \qquad (4)$$

where T_c is the chip interval and $g(t)$ is the chip waveform. We assume $g(t)$ to be a rectangular pulse. The bit duration is given by $T_b = NT_c$.

The transmitted signal for the k^{th} user, $0 \leq k \leq K$, can be expressed as

$$s_k(t) = \text{Re}\left[\sqrt{(2P_k)} a_k(t) e^{j(\omega_c t + \theta_k)}\right] \qquad (5)$$

where P_k is the k^{th} user's transmit signal power, and $\omega_c = 2\pi f_c$ is the carrier frequency. We consider the zeroth user as the desired user and the rest as interferers.

2.1 Channel Model

The channel is assumed to experience multiplicative (i.e., frequency non-selective) multipath fading and AWGN. The received signal, $r(t)$, at the base station is given by

$$r(t) = \sum_{i=0}^{K} \sqrt{2P_k} \alpha_k(t) a_k(t - \tau_k) \cos(\omega_c t + \phi_k) + n(t), \qquad (6)$$

where $\alpha_k(t)$ represents the multiplicative fading term for the k^{th} user, $\tau_k \in [0, T_c), 0 \leq k \leq K$ is the k^{th} user's delay, ϕ_k models the phase angle of the k^{th} user, and $n(t)$ is AWGN with power spectral density $\eta_0/2$.

The fading term $\alpha(t)$ is considered to follow a Rayleigh distribution with a Doppler spectrum of the form [17]

$$\alpha(f) = \begin{cases} \frac{1}{2\pi f_d \sqrt{1 - (\frac{f}{f_d})^2}} & |f| < f_d \\ 0 & \text{elsewhere,} \end{cases} \qquad (7)$$

so that the underlying Gaussian processes have normalized correlation functions given by $J_0(2\pi f_d \tau)$, where $J_0(.)$ is the Bessel function of the first kind of order zero, $f_d = v/\lambda$ is the Doppler bandwidth, v is the mobile user speed, λ is the wavelength, and τ is the time delay between the specified correlated samples.

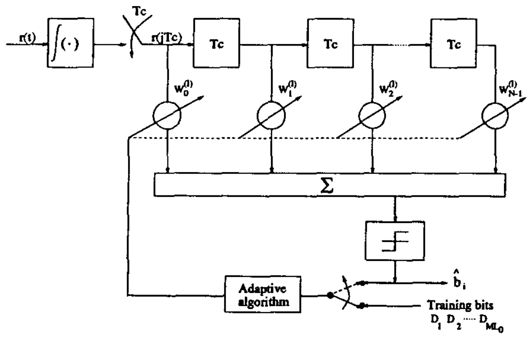

Figure 2: Long Code LMS Receiver

2.2 Long Code LMS Receiver

Since the chip sequence (or code segment) in a bit interval repeats every L_o bits, the LMS algorithm must adapt by considering those code segments which are the same. Conceptually then, the LMS algorithm for long codes can be viewed as a parallel bank of equalizers, each adapting to a different code segment comprising the long code. Thus, for the long code, L_o such equalizer blocks are needed. For example, if $L_o = 2$, the first equalizer will adapt to code segment CS_0, and the second equalizer will adapt to code segment CS_1. The practical realization of the LC-LMS receiver, however, need not have L_o separate equalizers. Since the input samples arrive serially, a single adaptive equalizer block with additional memory to store the filter coefficients corresponding to those code segments which are not operated on during a given a bit interval is adequate. Figure 2 illustrates such a LC-LMS receiver which uses a chip matched filter followed by an adaptive chip interval spaced equalizer structure.

An implication of the above LC-LMS receiver structure is that the number of training bits must be increased in order to ensure convergence on all L_o code segments. Here, we assume that the transmitter will initially send ML_o training bits before commencing transmission of actual data bits, where M is the number of training bits needed for convergence in a short code ($L_o = 1$) system.

The received signal, $r(t)$, after conversion to baseband, is passed through a chip matched filter and sampled at the end of every chip interval, i.e.,

$$r(jT_c) = \int_{(j-1)T_c}^{jT_c} r(t)dt. \tag{8}$$

The received chip sequence $r(jT_c)$ is shifted through the N-length adaptive filter taps at the chip rate. Let $\mathbf{R}^{(l)} = \left(r_0^{(l)}, r_1^{(l)}, \ldots. r_{N-1}^{(l)}\right)$ represent the tap vector and $\mathbf{W}^{(l)} = \left(w_0^{(l)}, w_1^{(l)}, \ldots., w_{N-1}^{(l)}\right)$ represent the tap

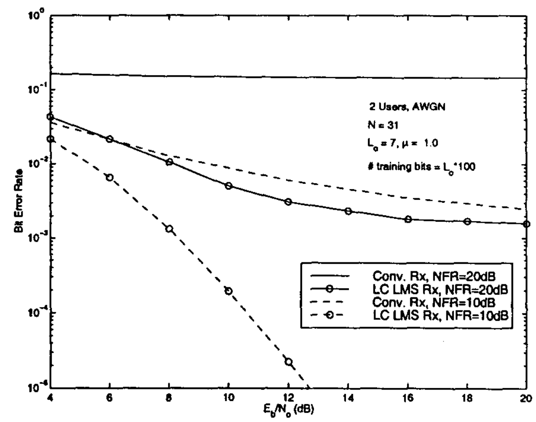

Figure 3: BER performance of Long Code LMS *vs* Conventional receiver on AWGN channel. $N = 31$. $L_o = 7$. # training bits = L_o*100.

weight vector (i.e., filter coefficients), corresponding to the l^{th} code segment, $0 \leq l \leq L_o - 1$. The filter coefficients can be iteratively obtained using the LMS algorithm given by

$$\mathbf{W}_{m+1}^{(l)} = \mathbf{W}_m^{(l)} + \frac{\mu e_m^{(l)} \mathbf{R}_m^{(l)}}{\mathbf{T}(\mathbf{R}_m^{(l)})\mathbf{R}_m^{(l)}} \tag{9}$$

where $\mathbf{R}_m^{(l)}$ and $\mathbf{W}_m^{(l)}$ are the tap vector and tap weight vector, respectively, of the m^{th} iteration of l^{th} code segment. $\mathbf{T}(.)$ is the transpose operation and μ is the adaptation step size. The error signal, $e_m^{(l)}$, is computed as the difference between the training sequence bit value for the m^{th} iteration and $\mathbf{T}(\mathbf{W}_m^{(l)})\mathbf{R}_m$, i.e.,

$$e_m^{(l)} = D_m^{(l)} - \mathbf{T}(\mathbf{W}_m^{(l)})\mathbf{R}_m^{(l)}. \tag{10}$$

Once the training ends, the data bit estimates $\hat{b}_i$ are obtained by sampling and hard quantizing the filter output, once every bit interval. If the elements of the vector $\mathbf{W}^{(l)}$ are taken to be the elements of the spreading sequence without adaptation, the receiver becomes a conventional correlation receiver.

3 Performance Results

In order to show the improved performance offered by the adaptive LC-LMS receiver relative to conventional techniques, several numerical examples are generated through simulations for a two-user DS-CDMA system. The two-user example is simple and yet illustrative of the salient features of the receiver performance, particularly in a near-far scenario. The near-far scenario is characterized by the near-far-ratio (NFR) parameter, which is defined as the ratio (or dB difference) of the

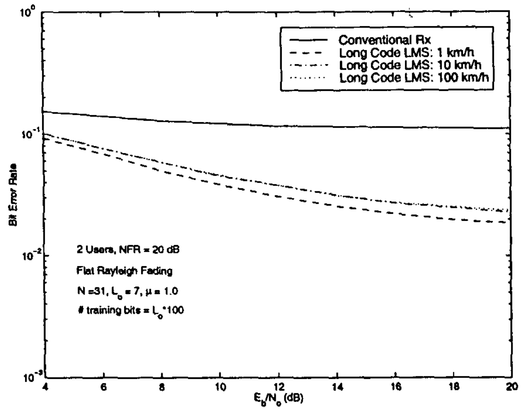

Figure 4: Performance of LC-LMS *vs* Conventional Rx on Rayleigh fading at different user speeds. $N = 31$. NFR = 20 dB. $L_o = 7$. # training bits = L_o*100.

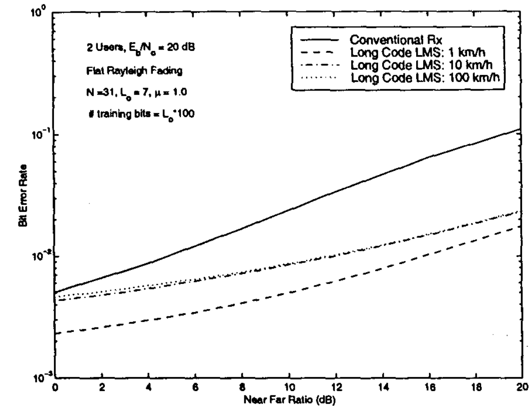

Figure 5: Performance of LC-LMS Rx as a function of NFR. Rayleigh fading. $N = 31$. $E_b/N_o = 20$ dB. $L_o = 7$. # training bits = L_o*100.

interfering users' transmit powers to the desired user's transmit power. If $P_0 = P_1 = P_2 = P_3 = ... = P_K$, then $P_1/P_0 = P_2/P_0 = ... = P_K/P_0 = 1$, meaning 0 dB near-far-ratio.

Figure 3 shows a comparison of the bit error performance of the adaptive LC-LMS receiver and the conventional receiver for different values of E_b/N_o on an AWGN channel. A processing gain of $N = 31$ and a long code period of $L_o = 7$ are used. The data rate is taken to be 10 kbps (i.e., $T_b = 0.1$ msec)[1]. The NFR values considered are 10 dB and 20 dB. As the near-far-ratio increases, both conventional as well as LC-LMS receivers degrade in bit error performance, as expected. However, for a given NFR, the LC-LMS receiver shows significant improvement in bit error rate (BER) compared to conventional receiver, illustrating the LC-LMS receiver's interference suppression capability. For example, at $E_b/N_o = 20$ dB, the LC-LMS receiver shows two orders of magnitude improvement in BER when NFR = 20 dB, and much more improvement in BER when NFR = 10 dB, clearly demonstrating the LC-LMS receiver's near-far resistance.

Figures 4 shows the performance of both LC-LMS and conventional receivers on a flat Rayleigh fading channel when the period of the long code $L_o = 7$ and the near-far ratio is 20 dB. The curves are parameterized by the mobile user speeds. The various speeds considered are 1 km/h, 10 km/h, and 100 km/h. At 900 MHz carrier frequency, these speeds correspond to 0.83 Hz, 8.3 Hz, and 83 Hz of Doppler frequency, respectively. The channel changes rather slowly at low Doppler frequencies, but quite rapidly at high Doppler values.

[1]Note that for a given channel coherence time, the number of bits in a long code period (i.e., L_o) can be proportionately larger if a higher data rate is considered in place of 10 kbps.

From Figure 4, it can be seen that LC-LMS performs best when the user speed is 1 km/h compared to 10 km/h and 100 km/h user speeds. This is because, when the channel fades are very slow, the cyclostationary property of the CDMA channel is not altered significantly over a long code period (for $L_o = 7$, in this case), and this enables the LC-LMS receiver to suppress the interference effectively. It is noted that the LC-LMS performance for 10 and 10 km/f are almost same, while still retaining the near-far resistance advantage compared to the conventional receiver, even at high speeds.

In Figure 5, the effect of increasing the near-far-ratio on the bit error performance is illustrated at an E_b/N_o value of 20 dB and $L_o = 7$. The conventional receiver performance degrades from 0.005 BER for NFR = 0 dB (equal power users) to 0.1 BER for NFR = 20 dB. However, the LC-LMS receiver provides increasing near-far-resistance compared to the conventional receiver for increasing NFR values. For example, the BER improves from 0.005 for the conventional receiver to 0.0045 for the LC-LMS receiver (a small improvement) when NFR = 0 dB at 10 and 100 km/hr speeds, whereas when NFR = 20 dB, the BER improves from 0.1 to 0.03 (a larger improvement). For a 1 km/h speed, however, the performance improvement is more uniform at different NFR values (i.e., about half an order improvement at both NFR = 0 and 20 dB).

Finally, the effect of increasing the long code period (L_o) at different user speeds is shown in Figure 6 for $E_b/N_o = 20$ dB and NFR = 10 dB. As observed earlier, the LC-LMS receiver performs better at a 1 km/h speed than at 10 and 100 km/h. We note that by increasing the value of L_o we are increasing the number of training bits, which may not be desired in certain

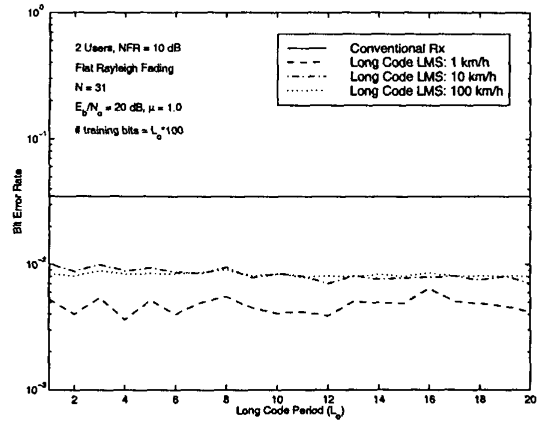

Figure 6: Performance of LC-LMS Rx as a function of L_o. Rayleigh fading at different user speeds. $N = 31$. $E_b/N_o = 20$ dB. NFR = 10 dB. # training bits = L_o*100.

applications. However, when low probability of intercept along with near-far resistance is of paramount importance, irrespective of the length of the training sequences being used, then the LC-LMS receiver appears to be a good choice. An ad-hoc network where users may randomly enter and leave the network without tight power control (e.g., military communications) is one such example.

4 Conclusion

An interference suppression scheme for DS-CDMA systems employing long codes was presented. The bit error performance of the long code LMS-type adaptive receiver on AWGN and correlated Rayleigh fading channels was evaluated for varying degrees of near-far ratio, user mobility and long code period. We showed that the LC-LMS receiver outperformed the conventional receiver by a significant margin in high NFR and low mobility scenarios. It was also seen that the LC-LMS receiver retained a reasonable near-far resistance advantage compared to the conventional receiver at high user speeds as well. A drawback in this receiver is the large number of training bits needed by the receiver to learn the channel. Further research work is being carried out to investigate other LC-LMS receiver algorithms which would require a lesser number of training bits.

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