Blind Adaptive Multi-User Detection for Multi-Carrier DS-CDMA Systems in Frequency Selective Fading Channels

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Abstract **-** This paper deals with the design of a receiver for a synchronous Multi-Carrier Direct-Sequence Code Division Multiple Access (MC-DS-CDMA) systems, in time-varying frequency selective fading channels. We consider a structure operating in two steps: after channel compensation and time alignment along each carrier, the resulting signals are combined and define the input of the blind adaptive multi-user detector. Our contribution is twofold:
First, we propose a generalization of the blind Least Mean Square (LMS) algorithm on the basis of multiple delayed input signal vectors, which has the advantage to improve the convergence features in high Multiple Access Interference (MAI) environment and time-varying fading scenario.
In addition, we carry out a comparative study between the proposed algorithm and previously published blind LMS and blind Kalman filter algorithms.
Simulation results show that the proposed algorithm can trade-off performance with complexity. In other words, it can have better performances than the existing blind LMS algorithm and less computational complexity than blind Kalman filter algorithm.

I. INTRODUCTION

A great deal of interest has been paid to Code Division Multiple Access (CDMA) communication systems combined with a Multi-Carrier (MC) transmission [1]. Indeed, this approach makes it possible to achieve high bandwidth efficiency, fading resilience and interference suppression capability, which are three crucial issues for future broadband data transmission [2].

When using MC-Direct Sequence (DS)-CDMA system, the data sequence, spread in the time domain by the so-called direct sequence, modulates several carriers [3].

Conventional receiver consists of a correlator for each carrier followed by a Maximal Ratio Combiner (MRC) to counteract fading and narrow band interference [3]. However, this approach cannot eliminate the Multiple Access Interference (MAI) and hence is not "near-far resistant". To avoid this drawback, an alternative approach consists in designing a minimum-mean-square-error receiver. In [5], two structures are proposed: the first one consists in designing a Wiener filter along each carrier, whereas the second is based on a joint structure defined by the concatenation of the filter weights dedicated to each carrier. To reduce the computational cost, adaptive filters can be considered. A comparative study between methods using Normalized Least Mean Square (NLMS), Affine Projection Algorithm (APA) and Recursive Least Square (RLS) is carried out in [8]. However, even if these receivers provide significant results, training data sequences for every active user are required and slow convergence in high MAI environment is observed.

For these reasons, one can pay attention to the so-called "Blind" Adaptive Multi-User Detection (BA-MUD) techniques which require only the knowledge of the spreading waveform and the timing of the desired user. In [9], the authors have developed the first blind LMS-based MUD, based on the minimization of a Mean Output Energy (MOE) criterion. However, this method is not convenient in a dynamic time-varying environment. For this reason, other algorithms have been proposed:

- On the one hand, blind detectors based on RLS [10] and Kalman filter [7] algorithms make it possible to improve the convergence features and tracking capabilities in a dynamic environment, such as in the case of birth or death of interferers.
- On the other hand, Mucchi et al. [11] have proposed a derived version of the pioneering blind LMS-based detector [9], which makes it possible to operate in a time-varying frequency-selective multi-path fading channel. For this purpose, they first complete channel compensation and time alignment on the signal replicas along each independent path and then combine the resulting signals before detection. This so-called pre-detection maximal ratio combining has the advantage to use only one detector for the combined replicas instead of one detector for each signal replica. In addition, according to [11], this yields a remarkable complexity reduction, more reliable decision variable and more robust convergence procedure.

However, the above BA-MUD techniques were only developed for single carrier DS-CDMA systems. Few blind adaptive techniques have been yet proposed for MC-DS-CDMA systems. In [4], Lok et al. propose a method operating in Rayleigh fading channels, which can be viewed as a form of BA-MUD in the frequency-domain. Nevertheless, the authors did not use existing (time-domain) BA-MUD techniques [12] [9] [10] [7] to suppress the MAI.

In this paper, we will focus our attention on synchronous MC-DS-CDMA systems, in a frequency selective fading channel. To mitigate the MAI in high interference environment and a time-varying fading scenario, we propose to consider a receiver structure consisting in pre-detection

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maximal ratio combining followed by a BA-MUD. Our contribution is then twofold. We first develop a generalization of the blind LMS algorithm, on the basis of multiple delayed input signal vectors. Then, we complete a comparative study with previously published blind LMS and blind Kalman filter algorithms.

The remainder of the paper is organized as follows: Section II provides mathematical modelling and description of synchronous MC-DS-CDMA system. In section III, blind adaptive MOE receiver structure with pre-detection maximal ratio combining is presented. Blind adaptive multi-user detectors are introduced in section IV. Simulation results and conclusions are reported in section V.

Notations: Vectors (matrices) are denoted by boldface lower (upper) case letters. Superscripts $(\cdot)^*$ and $(\cdot)^T$ respectively denote the complex conjugate and transpose operations. $\mathbf{I}_L$ is the $L\times L$ identity matrix. $E[\cdot]$ is the statistical expectation. $\operatorname{sgn}(\cdot)$ is the signum operator and $\operatorname{Re}[\cdot]$ is the real part.

II. SYSTEM MODEL

In the following, let us consider the synchronous transmission in MC-DS-CDMA systems with K active users and M carriers.

The transmitted signal at the m^{th} carrier can be expressed in the following manner:

$$s_m(t) = \operatorname{Re}\left[\sum_{n=-\infty}^{+\infty}\sum_{k=1}^{K}\sqrt{2P_k}\, b_k(n) c_k(t-nT_b) e^{j\omega_m t}\right] \tag{1}$$

where P_k is the power of the k^{th} user, $b_k(n)\in\{-1,1\}$ is the n^{th} data bit of the k^{th} user, T_b is the bit duration and ω_m is the m^{th} carrier frequency. In addition, the spreading waveform of the k^{th} user is given by:

$$c_k(t) = \sum_{i=0}^{N-1} a_k(i)\psi(t-iT_c) \tag{2}$$

where T_c is the chip duration, $N = T_b / T_c$ is the processing gain, $a_k(i)\in\{\pm 1/\sqrt{N}\}$, $i = 0,1,\ldots,N-1$ is the normalized spreading sequence and $\psi(t)$ is the chip pulse shape, assigned to one over the interval $[0,T_c]$ and zero otherwise.

For one thing, the transmitted MC signal goes through a frequency selective fading channel. By suitably choosing the number M of carriers and the bandwidth of the chip pulse shape $\psi(t)$ [3], one can assume that each carrier undergoes independent frequency non-selective fading. So, the system will have a diversity gain equal to the number of carriers.

In addition, the transmitted signal on the m^{th} carrier is corrupted by an independent additive white Gaussian noise (AWGN) process $\eta_m(t)$ with power spectral density N_0.

Therefore, the continuous time received signal on the m^{th} carrier in its complex analytic form is given by:

$$r_m(t) = \sum_{n=-\infty}^{+\infty}\sum_{k=1}^{K}\sqrt{2P_k}\,\beta_m(n) b_k(n) c_k(t-nT_b) e^{j\omega_m t} + \eta_m(t) \tag{3}$$

where $\left\{\beta_m(n) = \alpha_m(n) e^{j\theta_m(n)}\right\}_{m=1,2,\ldots,M}$ are independent and identically distributed zero-mean complex Gaussian random variables. It should be noted that they model the overall effects of phase shift $\theta_m(n)$ and fading $\alpha_m(n)$ of the m^{th} carrier during the n^{th} bit interval. In addition, the channel coefficient $\beta_m(n)$ is assumed to be slowly time-varying, i.e. it can be assumed constant within a one-bit interval, and time varying from bit to bit.

In the sequel, our purpose is to retrieve the first user data sequence $b_1(n)$, from the received signal.

III. BLIND ADAPTIVE MOE RECEIVER WITH PRE-DETECTION COMBINING

The receiver we propose in this paper is shown in Fig.1 and operates as follows:

First, the demodulated signal over the m^{th} carrier $x_m(t) = r_m(t) e^{-jw_m t}$ is processed with a chip-matched filter, which consists of an integrator with duration T_c. The samples are then stored for one bit interval, resulting in the following $N\times 1$ vector:

$$\mathbf{x}_m(n) = \sum_{k=1}^{K}\sqrt{2P_k}\,\beta_m(n) b_k(n)\mathbf{a}_k + \boldsymbol{\eta}_m(n) \tag{4}$$

where $\mathbf{a}_k = [a_k(0) \;\; a_k(1) \;\; \cdots \;\; a_k(N-1)]^T$ denotes the normalized spreading vector of the k^{th} user and $\boldsymbol{\eta}_m(n)$ is a vector of independent AWGN samples.

If we assume that estimates of the channel coefficients $\left\{\hat{\beta}_m(n)\right\}_{m=1,2,\ldots,M}$ are available at the receiver, channel compensation along the m^{th} carrier can be then performed as follows:

$$\mathbf{y}_m(n) = \hat{\beta}_m^*(n)\mathbf{x}_m(n) \tag{5}$$

After channel compensation and time alignment, the received vectors $\{\mathbf{y}_m(n)\}_{m=1,2,\ldots,M}$ are coherently combined before detection to obtain the “total” received vector:

$$\mathbf{y}_{tot}(n) = \sum_{m=1}^{M}\mathbf{y}_m(n) \tag{6}$$

Then, a decision variable can be obtained from the total received vector by using a blind adaptive multi-user detector characterized by $\mathbf{w}(n) = [w(1) \;\; w(2)\ldots w(N)]$. It satisfies:

$$z(n) = \mathbf{w}^T(n)\mathbf{y}_{tot}(n) \tag{7}$$

Finally, the symbol decision of the first user can be obtained as follows:

$$\hat{b}_1(n) = \operatorname{sgn}(z(n)) \tag{8}$$

IV. BLIND ADAPTIVE MULTI-USER DETECTORS (BA-MUD)

A. Normalized blind LMS-based MUD

The canonical representation of a linear BA-MUD for user 1 was firstly established in [9] as follows:

$$\mathbf{w}(n)=\mathbf{a}_1+\mathbf{c}(n) \tag{9}$$

subject to the constraint $\mathbf{a}_1^T\mathbf{c}(n)=0$, where $\mathbf{a}_1$ is the normalized spreading vector of the first user and $\mathbf{c}(n)$ is the adaptive part of the detector. By minimising a MOE cost function of the form:

$$\mathrm{MOE}[\mathbf{w}(n)]=E\left[\left|\mathbf{w}^T(n)\mathbf{y}_{tot}(n)\right|^2\right] \tag{10}$$

Honig et al. [9] have proposed a blind LMS based algorithm to update the adaptive part of the detector (9):

$$\mathbf{c}(n)=\mathbf{c}(n-1)-\mu_{LMS}Z(n)[\mathbf{y}_{tot}(n)-Z_{MF}(n)\mathbf{a}_1] \tag{11}$$

where $Z(n)=\mathbf{w}(n-1)^T\mathbf{y}_{tot}(n)$ is the output of the linear detector, $Z_{MF}(n)=\mathbf{a}_1^T\mathbf{y}_{tot}(n)$ is the output of the conventional matched filter and μ_{LMS} is the step size. In order to avoid the gradient noise amplification problem, a normalized version of this algorithm can be written as follows:

$$\mathbf{c}(n)=\mathbf{c}(n-1)-\frac{\mu_N}{\gamma_N+\mathbf{y}_{tot}^T(n)\mathbf{y}_{tot}(n)}Z(n)[\mathbf{y}_{tot}(n)-Z_{MF}(n)\mathbf{a}_1] \tag{12}$$

where $\mu_N\in(0,2)$ is the normalized step size that controls the adaptation speed and γ_N is a small positive regularization constant that insures stability when $\mathbf{y}_{tot}^T(n)\mathbf{y}_{tot}(n)$ is too small.

B. Proposed generalized normalized blind LMS-based MUD

To improve the convergence features in high MAI environment and a time-varying fading scenario, we propose to generalize the algorithm in (12) by using L delayed input signal vectors.

Toward this end, we first define the following received and code matrices of L column each:

$$\mathbf{Y}_{tot}(n)=[\mathbf{y}_{tot}(n)\ \ \mathbf{y}_{tot}(n-1)\ \ \ldots\ \ \mathbf{y}_{tot}(n-L+1)] \tag{13}$$

$$\mathbf{A}_1=[\mathbf{a}_1\ \ \mathbf{a}_1\ \ \ldots\ \ \mathbf{a}_1] \tag{14}$$

In addition, the MOE cost function in (10) is modified to account for L delayed input signal vectors:

$$\mathrm{MOE}[\mathbf{w}(n)]=E\left[\left|\mathbf{w}^T(n)\mathbf{Y}_{tot}(n)\right|^2\right] \tag{15}$$

Taking the unconstraint gradient of the MOE cost function:

$$\begin{aligned}\nabla\mathrm{MOE}[\mathbf{w}(n)]&=\nabla E[(\mathbf{w}^T(n)\mathbf{Y}_{tot}(n))^2]\\&\cong 2[\mathbf{w}^T(n)\mathbf{Y}_{tot}(n)][\mathbf{Y}_{tot}^T(n)]\end{aligned} \tag{16}$$

The projected gradient, orthogonal to $\mathbf{A}_1$ satisfies:

$$\nabla\mathrm{MOE}[\mathbf{w}(n)]\cong 2[\mathbf{w}^T(n)\mathbf{Y}_{tot}(n)][\mathbf{Y}_{tot}^T(n)-(\mathbf{A}_1^T\mathbf{Y}_{tot}(n))\mathbf{A}_1^T] \tag{17}$$

Then, a stochastic gradient algorithm that updates the adaptive part of the detector (9) can be written as follows:

$$\mathbf{c}^T(n)=\mathbf{c}^T(n-1)-\mu_{GLMS}\nabla\mathrm{MOE}[\mathbf{w}(n-1)] \tag{18}$$

where μ_{GLMS} is the step size.

Substituting (17) in (18) and introducing a normalization factor similar to APA [6], a new generalized algorithm for updating $\mathbf{c}(n)$ can be expressed by:

$$\begin{aligned}&\mathbf{c}^T(n)=\mathbf{c}^T(n-1)\\&-\mu_{GN}\mathbf{z}(n)[\gamma_{GN}\mathbf{I}_L+\mathbf{Y}_{tot}^T(n)\mathbf{Y}_{tot}(n)]^{-1}[\mathbf{Y}_{tot}^T(n)-\mathbf{Z}_{MF}(n)\mathbf{A}_1^T]\end{aligned} \tag{19}$$

where $\mathbf{z}(n)=\mathbf{w}^T(n-1)\mathbf{Y}_{tot}(n)$, $\mathbf{Z}_{MF}(n)=\mathbf{A}_1^T\mathbf{Y}_{tot}(n)$, $\mu_{GN}\in(0,2)$ and γ_{GN} is the regularization constant. To insure that the orthogonally condition $\mathbf{a}_1^T\mathbf{c}(n)=0$ is satisfied at each iteration, we replace $\mathbf{c}(n)$ by its orthogonal projection:

$$\mathbf{c}(n)=\mathbf{c}(n)-(\mathbf{c}^T(n)\mathbf{a}_1)\mathbf{a}_1 \tag{20}$$

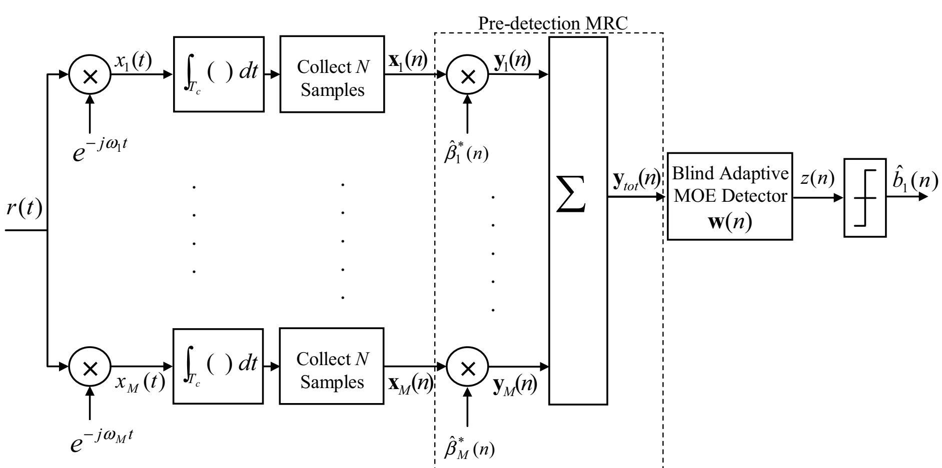

Fig.1: Blind adaptive MC-DS-CDMA receiver structure with pre-detection maximal ratio combining

V. Simulation Results and Conclusion

A. Simulation protocols

We carry out a comparative study between the blind approaches presented in sections IV and Kalman filter based approach presented in [7].

The spreading sequences used are gold codes of length N=31. We use a moderate value for the number of carriers, i.e. M=4. User 1 is assumed to be the desired user with Signal to Noise Ratio (SNR) per transmitted carrier kept constant at 10 dB. High MAI scenario is assumed with 14 multiple-access interfering users, among which 5 users have Interference to Signal Ratio (ISR) of 10 dB each, five users have ISR of 20 dB each, two users have ISR of 30 dB each and another two users have ISR of 40 dB each. The channel parameters $\{\beta_m(n)\}_{m=1,2,\dots,M}$ are generated according to the complex Gaussian distribution with zero-mean and unit-variance.

B. Results, comments and conclusion

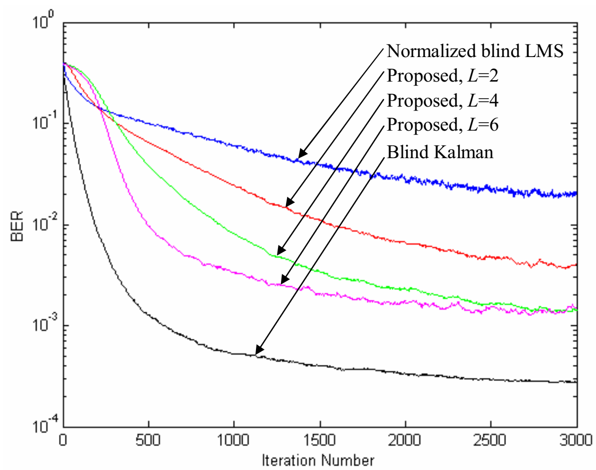

Fig.2: BER vs. number of iterations for the various blind adaptive algorithms.

According to Fig.2, the proposed algorithm with L=2, 4, 6 has better Bit Error Rate (BER) performance and convergence features than the normalized blind LMS algorithm. Increasing the value of L improves the performance but the computational complexity - $O(L^2N)$ - also increases. The blind Kalman algorithm [7] has the best BER performance but the highest computational complexity, of the order of $O(N^2)$.

Fig.3 demonstrates the Signal to Interference-plus-Noise Ratio (SINR) performance for the three algorithms. The SINR improvement of the proposed algorithm is better than the one obtained with the normalized blind LMS algorithm and approaches blind Kalman algorithm one, when L is getting higher.

Therefore, the low complexity interference suppression receiver structure we propose in this paper can trade-off performance with complexity by changing the scalable parameter L.

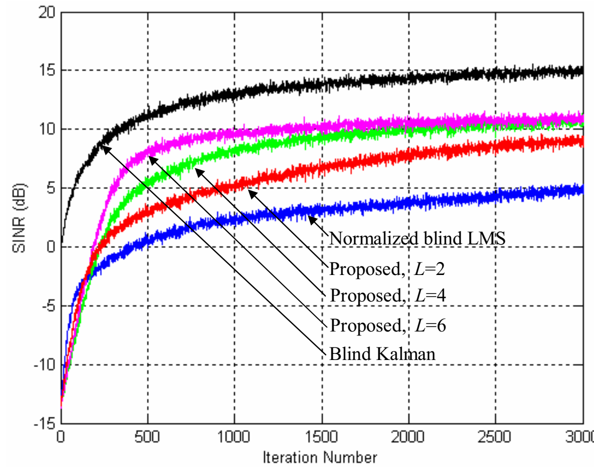

Fig.3: SINR vs. number of iterations for the various blind adaptive algorithms.

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