
An LMS-Based Beamforming Algorithm for Antenna Array Applied to an MC-CDMA System

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Abstract

In this paper, we propose an adaptive beamforming algorithm, called MC-LMS, for adaptive antenna applied to an MC-CDMA system. New method for updating the weight vector based on the MSE criterion is derived. Computer simulations show that MC-LMS algorithm is capable of rejecting co-channel interference that affects the MC-CDMA system. Thus, the BER performance of the MC-CDMA system is much higher than that of the MC-CDMA system without using adaptive antenna and that of the DS-CDMA system with adaptive antenna in multi-path Rayleigh fading channel.

I. Introduction

Multi-path fading effects and co-channel interference are the main limitations facing wireless communication systems. Therefore, finding approaches to overcome these limitations is greatly important for these systems to be able to accommodate a huge number of users. Recently, an approach, called the multi-carrier CDMA (MC-CDMA), [1]- [2], has been proposed to overcome the limitation in the capacity of the CDMA systems due to inter-chip interference and multi-user interference in high data rate communications systems [1]. The MC-CDMA is a combination of the CDMA scheme and the orthogonal frequency division multiplexing (OFDM) technique. Therefore, inter-chip interference could be eliminated and the complexity of the receiver could be reduced. Another key advantage of using OFDM is that the modulation and demodulation can be achieved in the discrete-domain by using a discrete Fourier transform (DFT) [3], which is efficiently implemented by using the fast Fourier transform (FFT). Thus, there would be no increase in the transmitter and receiver complexities of the MC-CDMA system.

It is also well known that adaptive antenna arrays, [4]-[6], have been introduced in the literature for TDMA and CDMA systems to mitigate rapid dispersive fading and to suppress co-channel interference so as to improve the performances of such systems. For the MC-CDMA with co-channel interference, adaptive antenna arrays are desirable. Unfortunately, there are up to now not many adaptive beamforming algorithms

for antenna arrays applied to the MC-CDMA system. There have been several publications [7], [8] investigating the performance of MC-CDMA system using antenna arrays. Nonetheless, due to the use of the inverse FFT (IFFT) and FFT processing in transmitter and receiver, respectively, new adaptive beamforming techniques are required for the MC-CDMA system.

In this paper, we propose an adaptive beamforming algorithm, called MC-LMS, for antenna array applied to the MC-CDMA system. In the proposed algorithm, the weight vector is first updated by short training process. Then a decision-directed technique is used for updating the weight vector so that the capacity could be more improved. Both of the processes are performed in the time domain. Simulation results show that the MC-LMS algorithm is able to extract the desired signal while it compresses other undesirable co-channel interferers. As a result, the MC-CDMA system employing MC-LMS algorithm has a very high BER performance compared with the MC-CDMA system without using antenna arrays and the DS-CDMA using an adaptive antenna array in multi-path Rayleigh fading channel.

II. MC-CDMA system with an adaptive antenna array

1. The transmitter of the MC-CDMA system:

In the transmitter the signal symbols of each user are first multiplied by the corresponding PN-

sequence assigned to each user. The spread signal of the m^{th} user can be written as:

$$y_m(t) = \sum_{i=-\infty}^{\infty} \sum_{j=0}^{N-1} b_m(i) c_{m,j}(i) p_c(t - jT_c - iT) \quad (1)$$

where $b_m(i) = \pm 1$ is the i^{th} signal symbol of the m^{th} user. $c_{m,j}(i) = \pm 1$ is the j^{th} spreading chip of the m^{th} user corresponding to the i^{th} signal symbol. $p_c(t)$ is the unit-magnitude rectangular pulse given by:

$$p_c(t) = \begin{cases} 1 & 0 \leq t < T_c \\ 0 & \text{otherwise} \end{cases} \quad (2)$$

T_c is the chip duration. $T = NT_c$ is the signal symbol duration. And N is the processing gain.

Without loss of generality, let us consider the n^{th} signal symbol, the spread signal $y_m(t)$ may be expressed in the vector form as follows:

$$\underline{y}_m(n) = [y_{m,0}(n) \ y_{m,1}(n) \ \dots \ y_{m,N-1}(n)] \quad (3)$$

where $y_{m,j}(n) = b_m(n) c_{m,j}(n)$, $j = 0, 1, \dots, N-1$.

This frequency-domain signal vector is then serial-to-parallel converted and is transformed in the time-domain signal by the IFFT as follows:

$$x_{m,l}(n) = \sum_{k=0}^{N-1} y_{m,k}(n) e^{j2\pi k l / N} \quad (4)$$

$$l = 0, 1, \dots, N-1$$

where $y_{m,k}(n)$ is the symbol of the m^{th} user carried by the k^{th} sub-carrier of the n^{th} block or n^{th} signal symbol. $x_{m,l}(n)$ is the l^{th} time-domain sample of the n^{th} block. In this system the number of sub-carrier is equal to the processing gain N . Equation (4) can be rewritten in the vector form as follows:

$$\underline{x}_m(n) = \underline{y}_m(n) \mathbf{F}^H(n) \quad (5)$$

where $\underline{x}_m(n) = [x_{m,0}(n) \ x_{m,1}(n) \ \dots \ x_{m,N-1}(n)]$, is the signal vector of the m^{th} user in the time-domain. And

$$\mathbf{F}(n) = \begin{bmatrix} 1 & 1 & \dots & 1 \\ 1 & e^{-j2\pi(1)(1)/N} & \dots & e^{-j2\pi(1)(N-1)/N} \\ \vdots & \vdots & \ddots & \vdots \\ 1 & e^{-j2\pi(N-1)(1)/N} & \dots & e^{-j2\pi(N-1)(N-1)/N} \end{bmatrix}$$

represents the FFT operation matrix. H denotes the Hermitian transpose.

After that the time-domain signal vector, $\underline{x}_m(n)$ is parallel-to-serial converted and is up converted to transmit to the base station.

2. The receiver of the MC-CDMA system:

In the receiver, the received signal is first down converted to get the base-band signal. Then the desired signal is obtained through the use of the smart antenna. Let us make the following assumptions:

- There are M signals impinging at an array of K elements, $K \geq M$.
- The channel is frequency selective fading.
- The number of sub-carrier N is chosen appropriately so that sub-carriers independently experience frequency nonselective fading.
- Perfect phase and timing synchronizations are achieved.

Under the above assumptions, the signal matrix, $V(n)$, received at the array antenna is represented by:

$$V(n) = A(\theta)Z(n) + G(n) \quad (6)$$

where,

$$V(n) = \begin{bmatrix} v_{0,0}(n) & v_{0,1}(n) & \dots & v_{0,N-1}(n) \\ v_{1,0}(n) & v_{1,1}(n) & \dots & v_{1,N-1}(n) \\ \vdots & \vdots & \ddots & \vdots \\ v_{K-1,0}(n) & v_{K-1,1}(n) & \dots & v_{K-1,N-1}(n) \end{bmatrix}$$

$$Z(n) = \begin{bmatrix} \beta_{0,0} x_{0,0}(n) & \dots & \beta_{0,N-1} x_{0,N-1}(n) \\ \beta_{1,0} x_{1,0}(n) & \dots & \beta_{1,N-1} x_{1,N-1}(n) \\ \vdots & \ddots & \vdots \\ \beta_{M-1,0} x_{M-1,0}(n) & \dots & \beta_{M-1,N-1} x_{M-1,N-1}(n) \end{bmatrix}$$

$$A(\theta) = \begin{bmatrix} a_0(\theta_0) & a_0(\theta_1) & \dots & a_0(\theta_{M-1}) \\ a_1(\theta_0) & a_1(\theta_1) & \dots & a_1(\theta_{M-1}) \\ \vdots & \vdots & \ddots & \vdots \\ a_{K-1}(\theta_0) & a_{K-1}(\theta_1) & \dots & a_{K-1}(\theta_{M-1}) \end{bmatrix}$$

$A(\theta)$ is the array response matrix for M users.

$G(n)$ is the $K \times N$ matrix of additive white Gaussian noises with zero mean and power spectral density $N_0/2$.

$Z(n)$ is the matrix of the signals of M users at the antenna array.

$\beta_{m,k}$, $m = 0, 1, \dots, M-1$, $k = 0, 1, \dots, N-1$, is a complex coefficient associated with the k^{th} snapshot of the m^{th} user, which is resulted from multi-path fading effects and is given by:

$$\beta_{m,k} = \alpha_{m,k} \exp(-j\phi_{m,k}) = \alpha_{m,k}^l + j\alpha_{m,k}^o \quad (7)$$

If we assume that $\alpha_{m,k}^I$ and $\alpha_{m,k}^Q$ are independent identically distributed (i.i.d) zero mean Gaussian random variables with variance $b_{m,k}$, the magnitude $\alpha_{m,k} = |\beta_{m,k}|$ has a Rayleigh distribution with the probability density function (PDF) given by [3]:

$$p_{\alpha_{m,k}}(x) = \frac{x}{b_{m,k}} \exp\left(-\frac{x^2}{2b_{m,k}}\right) \quad (8)$$

The desired signal vector of the m^{th} user, $\underline{s}_m(n)$, achieved by passing $V(n)$ through an adaptive beamformer with the weight vector $\underline{w}_m(n)$, is given by

$$\underline{s}_m(n) = \underline{w}_m^H(n)V(n) \quad (12)$$

where,

$$\underline{s}_m(n) = [s_{m,0}(n) \ s_{m,1}(n) \ \dots \ s_{m,N-1}(n)]$$

$$\underline{w}_m(n) = [w_{m,0}(n) \ w_{m,1}(n) \ \dots \ w_{m,K-1}(n)]^T$$

and T denotes the transpose.

The weighted signal vector $\underline{s}_m(n)$ is then converted into the frequency-domain by the FFT operation, thus yielding:

$$\tilde{\underline{y}}_m(n) = [\tilde{y}_{m,0}(n) \ \tilde{y}_{m,1}(n) \ \dots \ \tilde{y}_{m,N-1}(n)]$$

$$= \underline{s}_m(n)\mathbf{F}(n) = \underline{w}_m^H(n)V(n)\mathbf{F}(n) \quad (13)$$

This signal vector is converted to serial data and fed into a decision block. Based on the finite symbol property of the transmitted signal, the original spread signal from the transmitter could be recovered. Let us denote the recovered signal vector at the output of the decision block as $\underline{y}_{dm}(n)$. Note that this recovered signal vector corresponds to the transmitted spread signal vector $\underline{y}_m(n)$. The recovered signal vector, $\underline{y}_{dm}(n)$, is then multiplied by the spreading sequence assigned to the m^{th} user, $c_{m,j}(n)$, so that we can get the original signal symbol $b_m(n)$.

III. The proposed MC-LMS algorithm

1. Training period:

In this period only training signal is transmitted so as to train the weight vector of the array antenna. Note that, in the frequency-domain, both the received training signal and the received information signal are given by equation (13). In the time-domain, the training signal vector $\underline{x}_{tm}(n)$ and the received training signal vector $\tilde{\underline{x}}_{tm}(n)$ can be found from the corresponding frequency-domain

training signal vector $\underline{y}_m(n)$ and the received training signal vector $\tilde{\underline{y}}_{tm}(n)$ as follows:

$$\underline{x}_{tm}(n) = [x_{tm,0}(n) \ x_{tm,1}(n) \ \dots \ x_{tm,N-1}(n)]$$

$$= \underline{y}_m(n)\mathbf{F}^H(n) \quad (14)$$

and

$$\tilde{\underline{x}}_{tm}(n) = \tilde{\underline{y}}_{tm}(n)\mathbf{F}^H(n) = \underline{w}_m^H V(n)\mathbf{F}(n)\mathbf{F}^H(n) \quad (15)$$

where, $\mathbf{F}(n)$ is the FFT operation matrix.

Now let

$$U_m(n) = [\underline{u}_{m,0}(n) \ \underline{u}_{m,1}(n) \ \dots \ \underline{u}_{m,N-1}(n)]$$

$$= V(n)\mathbf{F}(n)\mathbf{F}^H(n) \quad (16)$$

where $\underline{u}_{m,i}(n) = [u_{m,i}^0(n) \ u_{m,i}^1(n) \ \dots \ u_{m,i}^{K-1}(n)]^T$, $i = 0, 1, \dots, N-1$

Then the received training signal vector in equation (15) can be rewritten as:

$$\tilde{\underline{x}}_{tm}(n) = \underline{w}_m^H U_m(n)$$

$$= [\underline{w}_m^H \underline{u}_{m,0}(n) \ \underline{w}_m^H \underline{u}_{m,1}(n) \ \dots \ \underline{w}_m^H \underline{u}_{m,N-1}(n)] \quad (17)$$

Based on the MSE criterion, we propose a new cost function for the MC-LMS algorithm applied to the MC-CDMA system as follows:

$$J(n) = \sum_{i=0}^{N-1} E \left[\left| \underline{w}_m^H(n) \underline{u}_{m,i}(n) - x_{tm,i}(n) \right|^2 \right] \quad (18)$$

The gradient of equation (18) with respect to the weight vector $\underline{w}_m(n)$ is given by:

$$\nabla J(n) = \sum_{i=0}^{N-1} [2R_{m,i}(n)\underline{w}_m(n) - 2\underline{p}_{m,i}(n)] \quad (19)$$

where,

$R_{m,i}(n) = E[\underline{u}_{m,i}(n)\underline{u}_{m,i}^H(n)]$ is the correlation matrix of the i^{th} received training symbol of the n^{th} block. $\underline{p}_{m,i}(n) = E[\underline{u}_{m,i}(n)x_{tm,i}^*(n)]$ is the cross-correlation vector between the i^{th} received training symbol and the i^{th} training symbol at the receiver.

An adaptive solution that minimizes the cost function is [6]:

$$\underline{w}_m(n+1) = \underline{w}_m(n) - \frac{1}{2}\mu\nabla J(n)$$

$$= \underline{w}_m(n) - \mu \sum_{i=0}^{N-1} [R_{m,i}(n)\underline{w}_m(n) - \underline{p}_{m,i}(n)] \quad (20)$$

From equation (20) we see that in order to update the weight vector we must know the correlation matrix $R_{m,i}(n)$ and the cross-correlation vector $\underline{p}_{m,i}(n)$. The simplest choice of estimators for $R_{m,i}(n)$ and $\underline{p}_{m,i}(n)$ is to use instantaneous estimates, that is:

$$R_{m,i}(n) = \underline{u}_{m,i}(n) \underline{u}_{m,i}^H(n) \quad (21)$$

$$\underline{p}_{m,i}(n) = \underline{u}_{m,i}(n) x_{m,i}^*(n) \quad (22)$$

Substituting equations (21) and (22) into equation (20) yields:

$$\underline{w}_m(n+1) = \underline{w}_m(n) - \mu \sum_{i=0}^{N-1} \underline{u}_{m,i}(n) e_{m,i}^*(n) \quad (23)$$

where $e_{m,i}(n) = \underline{w}_m^H(n) \underline{u}_{m,i}(n) - x_{m,i}(n)$ is the instantaneous error between the i^{th} received training symbol at the array output and the i^{th} training symbol of the n^{th} block or n^{th} signal symbol in the time-domain.

Thus in the training period the MC-LMS algorithm is summarized as follows:

1. Initialize

$$\hat{\underline{w}}_m(0) = 0, n = 0.$$

2. Update the weight vector, $n = n + 1$

- Receive a new training signal matrix $V(n)$.

- Calculate $U_m(n)$ as

$$U_m(n) = \begin{bmatrix} \underline{u}_{m,0}(n) & \underline{u}_{m,1}(n) & \dots & \underline{u}_{m,N-1}(n) \end{bmatrix} \\ = V(n) \mathbf{F}(n) \mathbf{F}^H(n)$$

- Calculate

$$e_{m,i}(n) = \underline{w}_m^H(n) \underline{u}_{m,i}(n) - x_{m,i}(n) \\ i = 0, 1, \dots, N-1$$

- Update the weight vector

$$\underline{w}_m(n) = \underline{w}_m(n-1) - \mu \sum_{i=0}^{N-1} \underline{u}_{m,i}(n) e_{m,i}^*(n).$$

3. Iterate step 2) until the weight vector converges

2. Communication period:

In this period, training signal is not used any more. A decision-directed approach is applied for updating the weight vector. The process for updating the weight vector is almost the same as that in the training period, except for the followings:

-The received information signal matrix $U_m(n)$ now replaces the received training signal matrix $U_m(n)$ and is still calculated by using equation (16).

-The time-domain reference signal is now created from the output of the decision block $\underline{y}_{dm}(n)$ and is given by:

$$\underline{x}_{dm}(n) = \begin{bmatrix} x_{dm,0}(n) & x_{dm,1}(n) & \dots & x_{dm,N-1}(n) \end{bmatrix} \\ = \underline{y}_{dm}(n) \mathbf{F}^H(n) \quad (26)$$

-The received information signal in the time domain is now given by:

$$\tilde{\underline{x}}_m(n) = \tilde{\underline{y}}_m(n) \mathbf{F}^H(n) = \underline{w}_m^H V(n) \mathbf{F}(n) \mathbf{F}^H(n) \quad (27)$$

Or, equivalently

$$\tilde{\underline{x}}_m(n) = \underline{w}_m^H U_m(n) \\ = \begin{bmatrix} \underline{w}_m^H \underline{u}_{m,0}(n) & \underline{w}_m^H \underline{u}_{m,1}(n) & \dots & \underline{w}_m^H \underline{u}_{m,N-1}(n) \end{bmatrix} \quad (28)$$

During this period, the proposed algorithm work as follows:

4. Receive a new information signal matrix $V(n)$, $n = n + 1$

Calculate the output of the array

$$\underline{s}_m(n) = \hat{\underline{w}}_m^H(n-1) V(n)$$

Convert into the frequency domain

$$\tilde{\underline{y}}_m(n) = \underline{s}_m(n) \mathbf{F}(n) = \underline{w}_m^H(n) V(n) \mathbf{F}(n)$$

Make decision to retrieve the original spread signal, denoted as $\underline{y}_{dm}(n)$.

Despread $\underline{y}_{dm}(n)$ to get the original signal symbol $b_m(n)$.

5. Update the weight vector,

Find the reference signal by converting $\underline{y}_{dm}(n)$

into the time domain as in equation (26)

$$\underline{x}_{dm}(n) = \begin{bmatrix} x_{dm,0}(n) & \dots & x_{dm,N-1}(n) \end{bmatrix} = \underline{y}_{dm}(n) \mathbf{F}^H(n)$$

Calculate

$$U_m(n) = \begin{bmatrix} \underline{u}_{m,0}(n) & \dots & \underline{u}_{m,N-1}(n) \end{bmatrix} = V(n) \mathbf{F}(n) \mathbf{F}^H(n)$$

Calculate

$$e_{m,i}(n) = \underline{w}_m^H(n) \underline{u}_{m,i}(n) - x_{dm,i}(n) \\ i = 0, 1, \dots, N-1$$

Update the weight vector

$$\underline{w}_m(n) = \underline{w}_m(n-1) - \mu \sum_{i=0}^{N-1} \underline{u}_{m,i}(n) e_{m,i}^*(n).$$

6. Repeat from step 4) until the end of the communication process.

IV. Simulation results and discussion

In the simulation, the channel is assumed to be multi-path Rayleigh fading and Rician fading with AWGN. The data symbols at the output of the spreader is binary; that is, $y_{m,k} = 1$ for bit 1 and $y_{m,k} = -1$ for bit 0. The number of sub-carriers for one block and the processing gain is the same and equal to $N = 64$. The spreading signal $c_m(t)$ is a Gold sequence. The antenna array is linear and the

distance between adjacent elements is half of the wavelength, $\lambda/2$. The number of users is 6. The mobile is moving at a velocity that results in a maximum Doppler shift of 66.67 Hz. The number of multi-path is assumed to be 30.

Fig. 1 shows the BER performances for the MC-CDMA system in multi-path Rayleigh fading channel with AWGN when an adaptive antenna array is applied. The number of elements is changed. As can be seen from the figure, when MC-LMS algorithm is employed, BER performances of the system are greatly improved in comparison with the case in which no array is used. In addition, the improvement in BER performances of the system is proportional to the number of array elements exploited.

Fig. 2 illustrates the comparison of the BER performances for the MC-CDMA system with MC-LMS algorithm and the DS-CDMA system using the conventional LMS algorithm presented in [6] in multi-path Rayleigh fading channel with AWGN. As shown in Fig. 9, with the aid of MC-LMS, BER performances of the MC-CDMA system are much higher than those of the DS-CDMA system exploiting smart antenna.

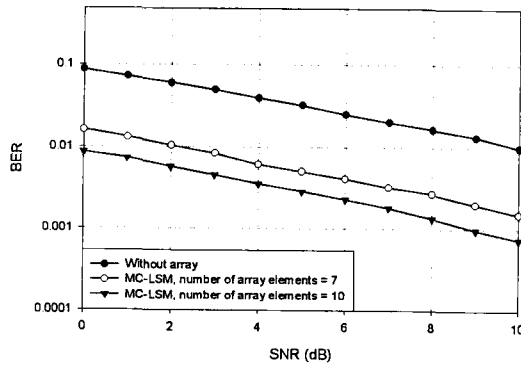


Fig.1: BER performance versus SNR for the MC-CDMA system in Rayleigh fading channel with the MC-LMS algorithm when the number of elements is varied.

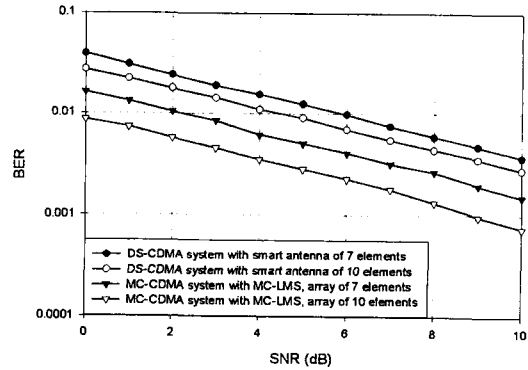


Fig. 2: Comparison of BER performances versus SNR for the MC-CDMA system with the MC-LMS algorithm and the DS-CDMA system using smart antenna in Rayleigh fading channel when the number of elements is varied.

V. Conclusion

In this paper, we for the first time propose the MC-LMS adaptive beamforming algorithm for the MC-CDMA system with array antenna. The proposed algorithm has a very good capability of suppressing co-channel interference, resulting in a MC-CDMA system with very high BER performances in multi-path Rayleigh fading channel with AWGN compared with the DS-CDMA system using smart antenna. The MC-LMS algorithm initially requires a very short training period, saving capacity of the system. Furthermore, the weight vector is updated in a similar manner as the conventional beamforming algorithms for TDMA and CDMA systems. Consequently, the proposed algorithm is a good candidate for combating co-channel interference in future wireless communication networks.

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