A Multidimensional Adaptive Linear Receiver for CDMA Transmission Corrupted by SD-MTI

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Abstract

This paper presents an extension of the single antenna adaptive linear receiver (ALR) for CDMA transmission. We call the proposed multiuser receiver Multidimensional Adaptive Linear Receiver (MALR). Specifically, we investigate the spatially-diverse multi-tone interference (SD-MTI) excision performance of the MALR. We quantify the BER performance enhancement achieved by the MALR relative to the ALR for the case of relatively large frequency MTI occupation.

1. Introduction

In recent years the application of antenna arrays for digital mobile communications has received increasing attention in the communication research community. In particular, several researchers, including [5] and [2], have indicated the use of antenna arrays for the reception of code division multiple access (CDMA) signals corrupted by multipath propagation, additive white Gaussian noise (AWGN), and multi-access interference (MAI). However, current research in this area has focused on the use of antenna arrays for the cancellation of MAI and not on the excision of spatially-diverse multi-tone interferences (SD-MTI) which might be present in CDMA transmissions.

In this paper, we consider the extension of the single antenna multi-user adaptive linear receiver (ALR) first introduced in [1]. Our interest here is in the BER performance of the extended system for CDMA transmission corrupted by SD-MTI and limited spreading gain. These spatially distributed interferences might have an aggregate power contribution and a large frequency occupancy so as to severely corrupt the CDMA signal. The ALR has been shown to be robust to MAI, multipath propagation, and monotone interference. In addition, it has been shown to be near-far

resistant and with complexity independent of the number of users. Moreover, chip synchronization is not needed [1]. We extend this single antenna system to a system that contains an adaptive antenna array and call it Multidimensional Adaptive Linear Receiver (MALR). We provide several representative BER simulation examples that show the enhanced performance of the MALR relative to the ALR.

2. MALR

This section presents the MALR structure and its relevance to the adaptive linear multiuser receiver. Generally classified as a minimum mean square error (MMSE) receiver [3], the ALR consists of a training-sequence-based adaptive linear MMSE filter and a decision device, see Figure 1. The training sequence is a symbol-length signal locally generated at the transmitter and receiver with the same PN code.

Figure 1. Adaptive Linear Receiver.

A more detailed diagram of the ALR system is shown in Figure 2. Notice that this system is structurally similar to the conventional LMS adaptive filter implementation. However, for this system, we process a block of chip samples greater than the spreading gain, L , and compute one output symbol every symbol period. The output for the k -th user is given by

where K is the number of users, $2N + 1$ is the duration of transmission, $a_k(i)$ is the i -th symbol for the k -th user, L is the signature length, $c_k(l)$ is the l -th signature chip, P is the number of channel paths, $g_{k,p}$ is the p -th complex attenuation relative to the k -th user, $\varphi_k(t)$ is the chip pulse for the k -th user, $\tau_{k,p}$ is the p -th path delay relative to the k -th user, τ_k is the relative delay for the k -th user, and $\psi_{k,p} = 2\pi \frac{d}{\lambda} \sin(\phi_k + \varepsilon_{k,p})$, where $\varepsilon_{k,p}$ is a small angle variation from the direction of arrival angle ϕ_k ; d and λ are the sensor separation and signal wavelength, respectively. The SD-MTI signal is given by

$$i^{(m)}(t) = \sum_{k_i=1}^{K_i} \sum_{l=0}^{L_{k_i}} \beta_{k_i,l} e^{-j\varphi_{k_i,l}} \sum_{q=1}^{Q_{k_i}} A_{k_i,q} \cos(2\pi f_{k_i,q}(t - \tau_{k_i,l}) + \phi_{k_i,q}) e^{j(m-1)\psi_{k_i,l}} \quad (4)$$

where L_{k_i} is the number of paths for the k_i interferer, $\beta_{k_i,l}$ is the l -th attenuation factor for the k_i interferer, $\varphi_{k_i,l}$ are uniformly distributed phases in the range $(0, 2\pi)$, Q_{k_i} is the number of sinusoids for the k_i interferer, $A_{k_i,q}$ is the q -th amplitude of the k_i interferer, $f_{k_i,q}$ is the q -th equally-spaced frequency of the k_i interferer, $\tau_{k_i,l}$ is the l -th delay path for the k_i interferer, $\phi_{k_i,q}$ is the q -th phase shift of the k_i interferer, and $\psi_{k_i,l} = 2\pi \frac{d}{\lambda_{k_i}} \sin(\theta_{k_i} + \varepsilon_{k_i,l})$, where $\varepsilon_{k_i,l}$ are small uniformly distributed angle variations in the range $(\theta_{k_i} + \varepsilon_{k_i}, \theta_{k_i} - \varepsilon_{k_i})$. These angle variation, which we call angle spread, result from the different paths in the channel. Figure 5 shows a coarse visual representation of the spatio-spectral allocation of two SD-MTI signals located at -35° and at 45° with 20% and 10% bandwidth occupancy, respectively. We let $Q = 10$, interference-to-signal power ratio (ISPR) = 5 dB and an angle spread of 4° .

Figure 5. Two MTIs with angle spread.

4. MMSE Analysis

In this section, we provide the MMSE analysis for the MALR. Without loss of generality we assume that the signal of interest (SOI) is user number one. To this end, we first observe that the correlation of the symbol for user number one $a_1^*(nT)$ and the ℓ -th element of the $2M + 1$ received vector $\mathbf{r}^{(m)}$ is given by

$$\begin{aligned} z_{1,n}^{(m)}(\ell T_f) &= E\{a_1^*(nT) \mathbf{r}^{(m)}(nT - \ell T_f)\} \\ &= \sigma_{a,1}^2(n) f_1^{(m)}(\ell T_f). \end{aligned} \quad (5)$$

where $\sigma_{a,1}^2(n) = E\{|a_1(nT)|^2\}$ and $f_1^{(m)}(\ell T_f) = \sum_{l=1}^L c_k(l) \sum_{p=1}^P g_{k,p} \varphi_k(t - \tau_{k,p}) e^{j(m-1)\psi_{k,p}}$. Therefore, the $2M + 1$ elements of $\mathbf{z}_{1,n}^{(m)}$ are basically the scaled and distorted samples of the transmitted signature waveform. We call $\mathbf{z}_{k,n}^{(m)}$ the *windowed signature* for the k -th user at time n . In vector notation, the received signal for user number one at the m -th antenna element is given by

$$\mathbf{r}^{(m)}(n) = \sum_{i=-N}^N z_{1,i}^{(m)} a_1(i) + \mathbf{v}^{(m)}(n) + \mathbf{i}^{(m)}(n) \quad (6)$$

where $\mathbf{v}^{(m)}$ is the AWGN vector and $\mathbf{i}^{(m)}$ is the interference vector. At time $i = n$, this vector is given by $\mathbf{r}^{(m)}(n) = \mathbf{z}_{1,n}^{(m)} a_1(n) + \mathbf{v}^{(m)}(n) + \mathbf{i}^{(m)}(n)$. Then the output of the combiner is

$$\begin{aligned} Y(n) &= \sum_{m=1}^M \mathbf{w}^{(m)T}(n) \mathbf{r}^{(m)}(n) \\ &= \mathbf{W}^T(n) \mathbf{r}(n) \end{aligned} \quad (7)$$

where $\mathbf{W}^T(n) = [\mathbf{w}^{(1)T}(n) \dots \mathbf{w}^{(M)T}(n)]$ and $\mathbf{r}^T(n) = [\mathbf{r}^{(1)T}(n) \dots \mathbf{r}^{(M)T}(n)]$. Thus, the mean-squared error at time n is calculated as follows

$$\begin{aligned} E\{|e(n)|^2\} &= E\{|Y(n) - a_1(n)|^2\} \\ &= \mathbf{W}^T E\{\mathbf{r} \mathbf{r}^H\} \mathbf{W}^* - \mathbf{W}^T E\{\mathbf{r} a_1^*(n)\} \\ &\quad - E\{a_1(n) \mathbf{r}^H\} \mathbf{W}^* + \sigma_{a,1}^2(n). \end{aligned} \quad (8)$$

The partial derivative with respect to $\mathbf{W}$ results in

$$\frac{\partial E\{|e(n)|^2\}}{\partial \mathbf{W}} = \Omega \mathbf{W}^* - \Lambda. \quad (9)$$

It follows that the optimum weight vector is given by the well known Wiener solution, $\mathbf{W}_{opt}^* = \Omega^{-1} \Lambda$, where Ω is given as follows

$$\Omega = \begin{bmatrix} \mathbf{z}_{1,n}^{(1)} \mathbf{z}_{1,n}^{(1)H} & \dots & \mathbf{z}_{1,n}^{(1)} \mathbf{z}_{1,n}^{(M)H} \\ \vdots & \ddots & \vdots \\ \mathbf{z}_{1,n}^{(M)} \mathbf{z}_{1,n}^{(1)H} & \dots & \mathbf{z}_{1,n}^{(M)} \mathbf{z}_{1,n}^{(M)H} \end{bmatrix} + \begin{bmatrix} E\{\nu^{(1)} \nu^{(1)H}\} & \dots & E\{\nu^{(1)} \nu^{(M)H}\} \\ \vdots & \ddots & \vdots \\ E\{\nu^{(M)} \nu^{(1)H}\} & \dots & E\{\nu^{(M)} \nu^{(M)H}\} \end{bmatrix} \quad (10)$$

ν is the vector composed of the noise and the interference samples. The vector Λ is given by

$$\Lambda = \begin{bmatrix} \mathbf{z}_{1,n}^{(1)} \sigma_{a,1}^2(n) \\ \vdots \\ \mathbf{z}_{1,n}^{(M)} \sigma_{a,1}^2(n) \end{bmatrix} \quad (11)$$

We substitute the result in (9) into the deterministic weight update equation [4]

$$\begin{aligned} \mathbf{W}_{n+1}^* &= \mathbf{W}_n^* - \frac{1}{2} \mu \nabla J(\mathbf{W}_n^*) \\ &= \mathbf{W}_n^* - \frac{1}{2} \mu (\Omega \mathbf{W}_n^* - \Lambda) \end{aligned} \quad (12)$$

where μ is the step size and ∇ is the vector gradient operator. The substitution of (10) and (11) into (12) results in

$$\begin{aligned} \mathbf{W}_{n+1}^* &= \mathbf{W}_n^* - \frac{1}{2} \mu [\tilde{\mathbf{z}}_{1,n} e^*(n) a_1(n) + \mathbf{R} \mathbf{W}_n^*] \\ &= \left(\mathbf{I} - \frac{1}{2} \mu \mathbf{R} \right) \mathbf{W}_n^* - \frac{1}{2} \mu \tilde{\mathbf{z}}_{1,n} e^*(n) a_1(n) \end{aligned} \quad (13)$$

where $\tilde{\mathbf{z}}_{1,n}^T = [\mathbf{z}_{1,n}^{(1)T} \dots \mathbf{z}_{1,n}^{(M)T}]$, $\mathbf{R}$ is the block-correlation matrix of the received signal and $\mathbf{I}$ is the identity matrix. The size of $(\mathbf{I} - \frac{1}{2} \mu \mathbf{R})$ is $M(2M+1) \times M(2M+1)$. Because the second order statistics are not known apriori, the above weight update is replaced by the LMS weight update

$$\mathbf{W}_{n+1}^* = \mathbf{W}_n^* - \mu e^*(n) \mathbf{r}_n \quad (14)$$

where we let μ absorbed the $\frac{1}{2}$.

5. Simulations

In this section, we provide representative BER performance comparisons between the MALR and the ALR systems. We assume that all the users have equal received power and spreading gain equal to 31. We use gold sequences to generate the different PN codes. Each user's signal is corrupted by different multipath channels of length

46, AWGN and 2 SD-MTI, and where necessary we averaged over 10 BER that result from 10 different sets of multipath channel realizations. We also let the angle spread equal to 2° . The interferences occupy 20% and 10% of the signal bandwidth and are located at 0° and 35° , respectively. The interference parameters are $Q = 10$, $L = 46$, ISPR equal to 5 dB, and an angle spread of 2° . Figure 6 shows the simulation scenario for 11 users and 2 SD-MTI. We let the angle separation between users equal to 10° . Notice that the SOI is located at 10° from the array normal and that we use an array of four elements with $d/\lambda = 0.5$, and each adaptive filter containing $M = 93$ weights. We chose the step size so as to approximate MMSE convergence.

Figure 7 shows the BER performance for a single user in an AWGN channel for both the ALR and the MALR. Notice that for this case we obtained the theoretical gain of 6 dB for 4 antennas. Figure 8 shows the effect of the multipath channel and the MTIs on the BER performance for one user. The BER for the ALR degrades approximately, from the multipath case, by 4 dB at 10^{-2} when the interferences are introduced. In the other hand, notice that the introduction of the SD-MTIs does not degrade the BER performance of the MALR. Conclusively, this result implies that the MALR is able to exploit the spatial domain to null the two spatially distributed interferences. Figure 9 shows the effect of multipath and SD-MTI on the BER performance for 11 users. As in the previous case, the BER performance for the MALR does not degrade severely by the introduction of the SD-MTI and the MAI. A comparison of Figures 8 and 9 indicates that the major contribution for the BER degradation of the MALR is due to the multipath channel. Moreover, notice that the difference in BER performance between the ALR and the MALR exceeds the theoretical gain of 6 dBs. Finally, Figure 10 shows the BER vs. ISPR. For this simulation we fixed the SNR at 10 dB, incremented the interference power from 0 dB to 10 dB, and used one channel realization for each of the 11 users. Notice that the BER degradation for the ALR is more pronounced than that of the MALR. This reflects the robustness of the MARL to SD-MTI signals.

Figure 6. Representative simulation example.

Figure 7. BER performance for a single user in an AWGN channel.

Figure 10. BER vs. ISPR for 11 users corrupted by multipath and 2 SD-MTIs.

Figure 8. BER performance for 1 user corrupted by multipath and 2 SD-MTIs.

Figure 9. BER performance for 11 users corrupted by multipath and 2 SD-MTIs.

6. Conclusion

In this paper we provided a multidimensional adaptive linear receiver for CDMA transmission corrupted by spatially-diverse multi-tone interference. This multi-user receiver incorporates the processing mechanism and attributes of the single antenna ALR, but more importantly it incorporates spatial domain processing. In addition to showing that the multidimensional receiver archives the the Winner solution, we also showed the enhanced BER performance of the MALR and robustness to SD-MTIs relative to the ALR.

References

- [1] P. Rapajic and B. Vucetic. Adaptive receiver structures for asynchronous CDMA systems. *IEEE Journal on Sel. Areas in Comm.*, 12(4):685–697, May 1994.
- [2] J. Thompson, P. Grant, and B. Mulgrew. Smart antenna arrays for CDMA systems. *IEEE Personal Communications*, 3(5):16–25, Oct. 1996.
- [3] S. Verdu. *Multisuser Detection*. Cambridge University Press, New York, NY, 1998.
- [4] B. Widrow and S. D. Stearns. *Adaptive Signal Processing*. Prentice-Hall, Inc., New Jersey, 1985.
- [5] J. Winters. Smart antennas for wireless systems. *IEEE Personal Communications*, 5(1):23–27, Feb 1998.