INTERFERENCE SUPPRESSION FOR UPLINK MULTIUSER CDMA*

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***Abstract*:** In this paper, techniques are proposed for interference suppression for the uplink multiuser CDMA systems. The uplink multiuser CDMA communication system model is described in the form of space-time domain through antenna array and multipath expression. Linear prediction technique is used to perform channel estimation. Based on the estimated channel state information, constrained minimum power digital filters are used to suppress interference. The pre-filtered multipath signals from all antennas are combined to get the final decision statistic. The correlation between the successive periods is considered to further improve the performances of the proposed scheme. Simulations demonstrate the effectiveness of the proposed scheme.

1. INTRODUCTION

Inter-symbol interference (ISI) and multiple access interference (MAI) are two major problems in wideband code division multiple access (CDMA) [1], [2]. Different advanced signal processing techniques have proposed to combat these problems. Multiuser detection [2] and space-time processing [3] are the two main category techniques to combat interference and multipath channel distortion.

In this paper, we consider the problem of interference suppression for uplink multiuser CDMA systems.

In what follows, the superscript $(\cdot)^T$, $(\cdot)^H$, $(\cdot)^*$, respectively, denotes the transpose, Hermitian transpose, the complex conjugate.

2. COMMUNICATION SYSTEM MODEL

The uplink in an asynchronous DS-CDMA cellular mobile radio network with K active users is considered. The transmitted baseband signal of the kth user is

$$x_k(t) = A_k \sum_{i=0}^{M-1} b_k(i) s_k(t - iT - q_k), \quad k = 1,2,\cdots,K \qquad (1)$$

where M, T, $b_k(i) \in \{+1,-1\}$ are, the number of data symbols per user per frame, the symbol interval, the ith transmitted symbol by the kth user. A_k, $s_k(t)$, q_k $(0 \le q_k < T)$ are, the amplitude, the normalized signaling waveform, the delay of the kth user's signal. The signaling waveforms are assumed to be DS-CDMA spread-spectrum signals of the form

$$s_k(t) = \sum_{n=0}^{N-1} c_k(n) p_{T_c}(t - nT_c), \; 0 \le t \le T \qquad (2)$$

where N, $\{c_k(n)\}_{n=0}^{N-1}$, $1/T_C$ and $p_{T_c}(t)$ are the spreading processing gain, the binary (± 1) spreading code assigned to the kth user, the chip rate of the spread-spectrum signal $\{s_k(t)\}$ and the normalized chip waveform (unit-height rectangular pulse) of duration T_c.

At the base station receiver, a uniform linear antenna array of P elements is employed. Assume that the channel can be modeled as a tapped delay line with L complex coefficients as the number of resolvable multipaths. The baseband multipath channel between the kth user's transmitter and the base station receiver can be modeled as a single-input multiple-output channel. The total received signal can be expressed as

$$\mathbf{y}(t) = \sum_{i=0}^{M-1}\sum_{k=1}^{K} A_k b_k(i) \sum_{l=1}^{L} \mathbf{a}_{kl}(\mathbf{O}_{kl}) g_{kl} s_k(t - iT - t_{kl}) + \mathbf{n}(t) \qquad (3)$$

where $\mathbf{y} = [y_1, y_2, \cdots, y_P]^T$, $\mathbf{n}(t)$ $(0 \le t \le T)$, g_{kl}, t_{kl}, $\mathbf{a}_{kl} = [a_{kl,1}, a_{kl,2}, \cdots, a_{kl,P}]^T$, $\mathbf{O}_{kl} = [O_{kl,1}, O_{kl,2}, \cdots, O_{kl,P}]^T$,

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L, denote, the vector of the received signals, the white Gaussian noise vector of antenna elements with variance of $s^2\mathbf{I}_{PL}$, the complex gain of the lth path of the kth user's signal, the delay of the lth path of the kth user's signal, the array response vector corresponding to the lth path of the kth user's signal, the angles of arrival (AOA) of the lth path of the kth user's signal, the number of paths in each user's channel.

3. CHANNEL ESTIMATION AND INTERFERENCE SUPPRESSION

Assume that the kth user is the user of interest, the receiver knows this user's spreading waveform $s_k(t)$ and its multipath delays $t_{k1}, t_{k2}, \cdots, t_{kL}$, the multipath spread of any user signal is limited to at most d symbol intervals, where d is a positive integer. That is, $t_{kl} \le dT$, $1 \le k \le K$, $1 \le l \le L$. The received signal at the pth antenna after chip-matched filtering can be expressed as

$$\mathbf{y}_p(i) = [y_{p,0}(i), y_{p,1}(i), \cdots, y_{p,\overline{N}-1}(i)]^T \quad (4)$$

where $\overline{N} = N + \lceil (t_{kL} - t_{k1})/T_c \rceil$ is used to capture the desired user's signal from all paths, and the samples are

$$y_{p,n}(i) = \int_{iT+t_{k1}+nT_c}^{iT+t_{k1}+(n+1)T_c} r_p(t) p_{T_c}(t - iT - t_{k1} - nT_c) dt \quad (5)$$

Using (3), the signal vector (4) can be rewritten as

$$\mathbf{y}_p(i) = A_k b_k(i) \sum_{l=1}^{L} a_{kl,p}(O_{kl,p}) g_{kl} \mathbf{s}_{kl}^{[0]} + \bar{\mathbf{i}}_P(i) + \mathbf{n}_p(i) \quad (6)$$

where $\mathbf{n}_p(i) = s\mathbf{I}_{\overline{N}}$ is the white Gaussian noise vector of the pth antenna element with variance of s^2, and $\bar{\mathbf{i}}_p(i)$ consists of the interfering signals given by

$$\bar{\mathbf{i}}_p(i) = \sum_{\substack{j=-d \\ j\neq 0}}^{d} A_k b_k(i+j) \sum_{l=1}^{L} a_{kl,p}(O_{kl,p}) g_{kl} \mathbf{s}_{kl}^{[j]} + \sum_{j=-d}^{d} \sum_{k'\neq k} A_{k'} b_{k'}(i+j) \sum_{l=1}^{L} a_{k'l,p} g_{k'l} \mathbf{s}_{k'l}^{[j]} \quad (7)$$

In (7), the vector $\mathbf{s}_{k'l}^{[j]}$ is the discretized version of the delayed signature waveform of user k', $s_{k'l}(t - jT - t_{k'l})$ with its nth element given by

$$s_{k'l}^{[j]}(n) = \int_{t_{k1}+nT_c}^{t_{k1}+(n+1)T_c} s_{k'l}(t - jT - t_{k'l}) p_{T_c}(t - t_{k1} - nT_c) dt$$
$$k' = 1,2,\cdots,K; \quad l = 1,2,\cdots,L; \quad j = -d,\cdots,d \quad (8)$$

On the right-hand side of (7), the first term represents the interference caused by the previous and subsequent symbols of the desired user, i.e., ISI; the second term represents the interference caused by the other user's signals, i.e., MAI.

From (6), we can see that the signals of the desired user depend only on the channel array response $a_{kl,p}(O_{kl,p}) g_{kl} \mathbf{s}_{kl}^{[j]}$. So, we can construct following linear prediction problem

$$e_1(i) = \mathbf{y}^{[0]}{}_p(i) - \mathbf{P}\mathbf{y}^{[else]}{}_p(i) \quad (9)$$

where $\mathbf{y}^{[0]}{}_p(i)$ is corresponding to $\mathbf{s}_{kl}^{[0]}$, $\mathbf{y}^{[else]}{}_p(i)$ is corresponding to $\mathbf{s}_{kl}^{[j]}$ $(j \neq 0)$. Assume that the symbols $b_k(i)$ are uncorrelated in time and that $\mathbf{b}_1(i), \cdots, \mathbf{b}_K(i)$ are mutually uncorrelated with variances $A_1, \cdots, A_K$. We define

$$\mathbf{y}^{[0]}{}_p(i) = \mathbf{H}_0 b_k(i) + \tilde{\mathbf{H}}_0 \tilde{\mathbf{b}}_k(i')$$
$$i' = i + j, \quad j \neq 0, \quad -d \le j \le d \quad (10)$$

where $\tilde{\mathbf{b}}_k(i')$ contains all symbol components in $[\mathbf{b}_1^H(i), \cdots, \mathbf{b}_K^H(i)]^H$ except for $b_k(i)$. $\mathbf{H}_0$ is the column vector of the channel matrix $[\mathbf{a}_{kl,p}(\mathbf{O}_{kl,p}) g_{kl} s_k]$ corresponding to $b_k(i)$, whereas all other columns of $[\mathbf{a}_{kl}(\mathbf{O}_{kl}) g_{kl} s_k]$ comprise $\tilde{\mathbf{H}}_0$. The optimal linear prediction matrix $\mathbf{P}$ gives [4]

$$e_1(i) = \mathbf{H}_0 b_k(i) \quad (11)$$

Let

$$\boldsymbol{e}_2(i) = \mathbf{y}^{[0]}{}_p(i) - \boldsymbol{e}_1(i) \qquad (12)$$

Then, we have

$$e_2(i) = \tilde{\mathbf{H}}_0 \tilde{\mathbf{b}}_k(i') \qquad (13)$$

Thus, the channel matrix vector space can be separated into two subspaces by linear prediction.

The solution of $\mathbf{P}$ and the linear prediction error can also be represented by the data correlation. Re-express the linear prediction problem as

$$e_1(i) = [\mathbf{I} - \mathbf{P}] \begin{bmatrix} \mathbf{y}^{[0]}{}_p(i) \\ \mathbf{y}^{[else]}{}_p(i) \end{bmatrix} \qquad (14)$$

and let

$$\mathbf{R} = E\left\{ \begin{bmatrix} \mathbf{y}^{[0]}{}_p(i) \\ \mathbf{y}^{[else]}{}_p(i) \end{bmatrix} [\mathbf{y}^{[0]}{}_p(i) \quad \mathbf{y}^{[else]}{}_p(i)] \right\} = \begin{bmatrix} \mathbf{R}_{11} & \mathbf{R}_{12} \\ \mathbf{R}_{21} & \mathbf{R}_{22} \end{bmatrix} \qquad (15)$$

Then, it is well known that the optimal solution for the linear prediction problem of (14) is [4]

$$\mathbf{P} = \mathbf{R}_{12}\mathbf{R}_{22}^{+} \qquad (16)$$

$$E\{e_1(i)e_1^H(i)\} = A_k \mathbf{H}_0 \mathbf{H}_0^H = \mathbf{R}_{11} - \mathbf{R}_{12}\mathbf{R}_{22}^{+}\mathbf{R}_{21} \qquad (17)$$

where $\mathbf{R}_{22}^{+}$ denotes the pseudo-inverse of $\mathbf{R}_{22}$.

Using the above linear prediction algorithm, we can estimate the channel response $\mathbf{H}_{0,p}$ corresponding to $b_k(i)$ for the pth antenna element.

Digital filters can be used to suppress interference. Let $\mathbf{w}_k$ be the filter weight vector for user k, then, the minimum power criterion can be used to optimize the filter weight vector as

$$\mathbf{w}_k = \arg\min_{\mathbf{w}_k} \arg\min \mathbf{w}_k^H \mathbf{R}_{yy} \mathbf{w}_k \qquad (18)$$

Under the constraint $\mathbf{w}_k^H(i)\overline{\mathbf{H}}_0(i) = 1$, the optimum coefficients are

$$\mathbf{w}_k(i) = \mathbf{R}_{yy}^{-1}(i)\overline{\mathbf{H}}_0(i)\left(\overline{\mathbf{H}}_0{}^H(i)\mathbf{R}_{yy}^{-1}(i)\overline{\mathbf{H}}_0(i)\right)^{-1} \qquad (19)$$

where $\mathbf{R}_{yy}(i) = E\{\mathbf{y}(i)\mathbf{y}^H(i)\}$ is the autocovariance of the received signal vector $\mathbf{y}(i) = [\mathbf{y}_1(i), \mathbf{y}_2(i), \cdots, \mathbf{y}_P(i)]^T$, and $\overline{\mathbf{H}}_0(i) = [\mathbf{H}_{0,1}(i), \mathbf{H}_{0,2}(i), \cdots, \mathbf{H}_{0,P}(i)]^T$ is the channel response vector for all antennas corresponding to $b_k(i)$.

The weighted outputs of the ith symbol of k user is

$$z_k(i) = \mathbf{w}_k^H(i)\mathbf{y}(i) \qquad (20)$$

The final decision rules applied for the demodulation of the estimated outputs is made by

$$\hat{b}_k(i) = \mathrm{sgn}\{\Re(\hat{z}_k(i))\} \qquad (21)$$

where $\mathrm{sgn}\{\}$ and $\Re(\cdot)$ denote sigmum and real part.

The linear prediction method proposed above can be directly used to estimate each channel response vector $\overline{\mathbf{H}}_0(i) = [\mathbf{H}_{0,1}(i), \mathbf{H}_{0,2}(i), \cdots, \mathbf{H}_{0,P}(i)]^T$, which is referred to as the DLPM method. In the following, we will propose another recursive method by considering the correlation between the successive periods, which is referred to as the RCM method.

From (20), we have

$$\hat{z}_k(i+1) = \hat{\mathbf{w}}_k^H(i)\mathbf{y}(i+1) \qquad (22)$$

From (20), and other denotations, we have

$$\overline{\mathbf{H}}_0(i) = E\{z_k^*(i)\mathbf{y}(i)\} \qquad (23)$$

Considering the correlation between the successive periods, we have following recursive estimates expression

$$\hat{\overline{\mathbf{H}}}_0(i+1) = (1-e)\hat{\overline{\mathbf{H}}}_0(i) + e\left(\hat{z}_k^*(i+1)\mathbf{y}(i+1)\right) \qquad (24)$$

$$\hat{\mathbf{R}}_{yy}(i+1) = (1-e)\hat{\mathbf{R}}_{yy}(i) + e\mathbf{y}(i+1)\mathbf{y}^H(i+1) \qquad (25)$$

To make the computation efficient, the inverse matrix of the estimated $\hat{\mathbf{R}}_{yy}(i)$ can be recursively calculated as [4]

$$\hat{\mathbf{R}}_{yy}{}^{-1}(i+1) = \frac{1}{m}\left[(1-e)\hat{\mathbf{R}}_{yy}{}^{-1}(i) - e\frac{\hat{\mathbf{R}}_{yy}{}^{-1}(i)\mathbf{y}(i+1)^*\mathbf{y}(i+1)^H\hat{\mathbf{R}}_{yy}{}^{-1}(i)}{m + \mathbf{y}(i+1)^H\hat{\mathbf{R}}_{yy}{}^{-1}(i)\mathbf{y}(i+1)^*}\right] \qquad (26)$$

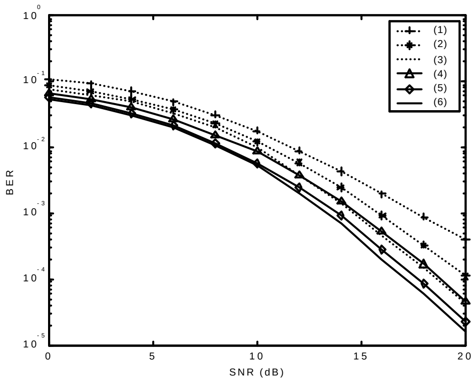

Figure 1. Average simulated and theoretical BER versus SNR performance for different NFR and different Doppler frequencies: (1)(2)(3) are for $D_{fb} = 0.005$ Hz/s and $\text{NFR} = 20$ dB;
(4)(5)(6) are for $D_{fb} = 0.5$ Hz/s and $\text{NFR} = 0$ dB;
(1)(4) are for DLPM method; (2)(5) are for RCM method;
(3)(6) are for theoretical.

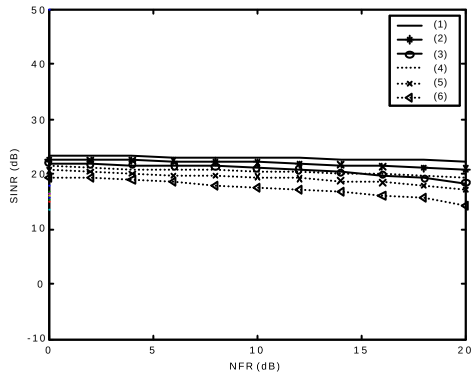

Figure 2. Average theoretical and simulated SINR versus NFR performance for different SNR and different Doppler frequencies: (1)(2)(3) are for $D_{fb} = 0.5$ Hz/s and $\text{SNR} = 20$ dB;
(4)(5)(6) are for $D_{fb} = 0.005$ Hz/s and $\text{SNR} = 5$ dB;
(1)(4) are for theoretical; (2)(5) are for RCM method; (3)(6) are for DLPM method.

The weighting factor, $e \in [0.5,1]$, is determined by the SNR and Doppler frequency as following rules: the value of e needs to be increased and closed to 1 when the signal-to-noise ratio (SNR) becomes infinite and Doppler frequency is very high. m is a forgetting factor.

4. SIMULATION STUDIES

The simulated CDMA system is an asynchronous system adopting $N = 31$ Gold codes as spreading sequences, with users $K = 15$, antenna elements $P = 3$, multipath diversity order $L = 5$, the number of symbols per frame is $M = 300$, the chip pulse is a raised cosine pulse with roll-off factor 0.5, the initial delay q_k of each user is uniformly generated on $[0,3T_c]$, the delay of each path t_{kl} is uniformly generated on $[0,5T_c]$. The modulation scheme is BPSK. The average simulated (estimated channels and noise variance) and theoretical (known channels and noise variance) bit error rate (BER) versus signal-to-noise ratio (SNR) performance is shown in Fig.1 for different near-far ratio (NFR) and different Doppler frequencies. The average theoretical (known channels and noise variance) and simulated (estimated channels and noise variance) output signal-to-interference-noise ratio (SINR) versus near-far ratio (NFR) performance is shown in Fig.2 for different SNR and different Doppler frequencies. From Fig.1 and Fig.2, we can see that the RCM method can further improve the BER-SNR and SINR-NFR performances of the system than the DLPM method. (All the results of Fig.1, and Fig.2 are obtained by averaging by 100 runs. The simulated methods include the DLPM method and the RCM method).

5. CONCLUSIONS

Techniques are proposed for interference suppression for the uplink multiuser CDMA systems. Simulations are used to demonstrate the effectiveness of the proposed scheme.

6. REFERENCES

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