

AN ITERATIVE BLIND ADAPTIVE RECEIVER FOR DS-SSMA SYSTEMS

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ABSTRACT

Spread-spectrum signals with long signature sequences are considered in this paper. Particularly, a linear receiver proposed in [1] and [2] for direct-sequence spread-spectrum multiple access systems is extended to its iterative form. This extension is most applicable in situations with high processing gain, where it is necessary at times to use a large number of correlators to acquire the PN sequence, e.g. in GPS. Like its noniterative counterpart in [1] and [2], the receiver structure consists of a matched filter followed by a tapped delay line that can be provisioned for use with antenna arrays. Similarly the iterative blind adaptive algorithm (BAA) has the same ability to combine multipaths, suppress multiple-access interference (MAI), and suppress narrowband interference. The iterative BAA demonstrates faster convergence time due to processing more measurements of interference outside the delay spread of the signal.

I. INTRODUCTION

Recently a linear receiver, which is composed of the conventional sequence-matched filter followed by a tapped delay line, has been proposed in [1] and [2] for direct-sequence spread-spectrum multiple-access (DS-SSMA) systems with long signature sequences. This receiver has the ability to suppress multiple-access interference (MAI) and narrowband interference, and combine multipath components. These capabilities are made possible through a suitable choice of weights in the tapped delay line following the conventional matched filter. A simple blind adaptive algorithm (BAA) is developed [1], [2] to obtain the required weights.

We notice that the BAA updates the weights every symbol. In high processing gain situations, e.g., in the Global Positioning System (GPS), this property may not be desirable because the resulting update rate may be too slow. To

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solve this problem, we extend the BAA to its iterative form by processing more measurements of interference outside the delay spread of the signal. Simulation results show that the iterative BAA demonstrates faster convergence time.

This paper is organized as follows: In Section II, we describe the DS-SSMA system model. In Section III, we briefly re-visit the linear receiver and the BAA proposed in [1] and [2]. In Section IV, we extend the BAA to its iterative form. In Section V, we present simulation results for two cases in which the linear receiver is employed to combat narrowband interference and MAI. Finally, we conclude this paper in Section VI.

II. SYSTEM MODEL

In this section, we describe the model of the DS-SSMA system. We assume that there are K simultaneous users in the system.

The k -th user, for $1 \leq k \leq K$, generates a data signal $b_k(t)$ that is given by

$$b_k(t) = \sum_{j=-\infty}^{\infty} b_j^{(k)} p_T(t - jT), \quad (1)$$

where T is the symbol duration, and the unit rectangular pulse $p_T(t)$ is given by $p_T(t) = 1$ for $0 \leq t < T$, and $p_T(t) = 0$, otherwise. The data symbols $b_j^{(k)}$ are such that $|b_j^{(k)}|^2 = 1$.

The k -th user is provided a randomly generated signature sequence $a^{(k)} = (\dots, a_0^{(k)}, a_1^{(k)}, \dots, a_{N-1}^{(k)}, \dots)$. The elements $a_i^{(k)}$ are independent random variables taking values from either $\{-1, +1\}$ or $\{-1, +1, +j, -j\}$ with equal probabilities. The sequence $a^{(k)}$ is used to form the spectral spreading signal that is given by

$$a_k(t) = \sum_{i=-\infty}^{\infty} a_i^{(k)} \psi(t - iT_c), \quad (2)$$

where the chip duration T_c is given by $T_c = T/N$, N is the number of chips per symbol interval, and $\psi(t)$ is the common chip waveform for all signals. The chip waveform $\psi(t)$ is time-limited to the interval $[0, T_c]$, and normalized so that $\int_0^{T_c} |\psi(t)|^2 dt = T_c$.

The transmitted signal for the k -th user, for $1 \leq k \leq K$, can be expressed as $\text{Re}[z_k(t - T_k) \exp(j\omega_c(t - T_k))]$, where

$$z_k(t) = \sqrt{2P_k} a_k(t) b_k(t), \quad (3)$$

P_k is the power for the k -th signal, ω_c is the carrier frequency, and T_k is the delay that models the asynchronous system.

Without loss of generality, we consider the signal from the first user as the communicating signal and the signals from all other users as interfering signals throughout the paper.

We now describe the channel model. We assume that the channel is a multipath fading channel corrupted by narrowband interferers as well as AWGN with two-sided power spectral density $N_0/2$. We assume that there are K_B narrowband interferers. Each narrowband interfering signal is assumed to be a pure tone.

We assume that the signals are received by an antenna array of D elements. The signal vector received by the antenna array in complex baseband notation is given by

$$\mathbf{r}(t) = \mathbf{y}(t) + \mathbf{n}_I(t) + \mathbf{n}_B(t) + \mathbf{n}_W(t), \quad (4)$$

where $\mathbf{n}_W(t)$ represents AWGN. We assume that the AWGN is also spatially white. The signal contribution $\mathbf{y}(t)$ is given by

$$\mathbf{y}(t) = \sum_{\lambda=1}^{L_1} g_{1,\lambda} z_1(t - T_1 - \tau_{1,\lambda}) e^{-j\omega_c(T_1 + \tau_{1,\lambda})} \mathbf{d}_{1,\lambda}. \quad (5)$$

The MAI contribution $\mathbf{n}_I(t)$ is given by

$$\mathbf{n}_I(t) = \sum_{k=2}^K \sum_{\lambda=1}^{L_k} g_{k,\lambda} z_k(t - T_k - \tau_{k,\lambda}) e^{-j\omega_c(T_k + \tau_{k,\lambda})} \mathbf{d}_{k,\lambda}. \quad (6)$$

The narrowband interference contribution $\mathbf{n}_B(t)$ is given by

$$\mathbf{n}_B(t) = \sum_{k=K+1}^{K+K_B} g_k \sqrt{2P_k} e^{j\delta_k t} \mathbf{d}_k. \quad (7)$$

In (5) and (6), L_k is the number of propagation paths from the k -th transmitter, for $1 \leq k \leq K$, to the antenna array. The parameters $\tau_{k,\lambda}$, $g_{k,\lambda}$, and $\mathbf{d}_{k,\lambda}$ represent the delay, the complex gain, and the array response vector associated with the λ -th path of the signal from the k -th transmitter. In (7), the parameters P_k , g_k , δ_k and \mathbf{d}_k represent the power, the complex gain, the frequency deviation from ω_c and the array response vector associated with the tone indexed by k for $K+1 \leq k \leq K+K_B$.

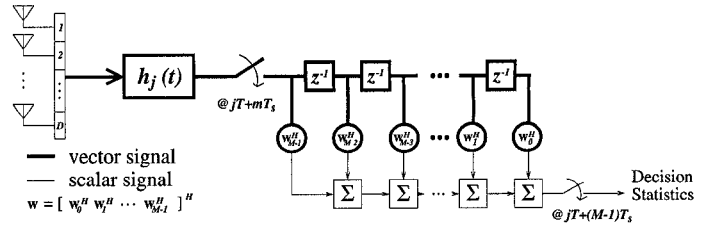


Fig. 1. Linear receiver for the j -th data symbol

III. LINEAR RECEIVER AND BLIND ADAPTIVE ALGORITHM

We assume that we have achieved synchronization with the path of the signal from the desired user that arrives earliest at the antenna array. By relabeling if necessary, we can assume that it is the path indexed by $\lambda = 1$. Hence we may assume $T_1 = 0$, $\tau_{1,1} = 0$, and $\tau_{1,\lambda} > 0$ for $2 \leq \lambda \leq L_1$. We also assume that $\tau_{1,\lambda} \leq \tau_{\max}$ for all λ , where τ_{\max} is a known constant. Without loss of generalization, we may assume the phase angle of $g_{1,1}$ and that of the first element of the array response vector $\mathbf{d}_{1,1}$ to be zero.

The receiver is shown in Fig. 1. To detect the j -th symbol $b_j^{(1)}$, the received signal $\mathbf{r}(t)$ is passed through a matched filter with impulse response $h_j(t)$ where

$$h_j(t) = \left[\sum_{i=0}^{N-1} a_{i+jN}^{(1)} \psi(T - t - iT_c) \right]^*. \quad (8)$$

The output of the matched filter $\tilde{\mathbf{r}}_j(t)$ is sampled once every T_s seconds to give a sample vector of length D , where D is the number of antenna elements and $T_c/T_s = S$ is the number of sample vectors per chip interval. We take M sample vectors, which contain all the received paths of the desired user signal, and concatenate them to give the vector \mathbf{s}_j of length MD . Since $\tau_{1,\lambda} \leq \tau_{\max}$ for all λ , $M \geq \lceil (\tau_{\max} + 2T_c)/T_s \rceil$. The decision statistic for the j -th symbol interval is formed by $\mathbf{w}^H \mathbf{s}_j$, where \mathbf{w} is a weight vector whose components remain to be determined.

To obtain the weights, we need a counterpart of \mathbf{s}_j that contains mainly interference and noise contributions. We obtain this auxiliary received vector by taking another M matched filter output sample vectors at some later time. Again, we concatenate them to give the auxiliary received vector $\hat{\mathbf{s}}_j$ of length MD . We note that the timing difference between each element of $\hat{\mathbf{s}}_j$ and the corresponding element of \mathbf{s}_j has to be an integral multiple of the chip duration, T_c . Fig. 2 depicts the sampling scheme to obtain \mathbf{s}_j and $\hat{\mathbf{s}}_j$ described above.

To develop a blind adaptive algorithm to obtain the optimal weight vector, we consider the constrained minimization of the output energy (OE) [1]. By employing the

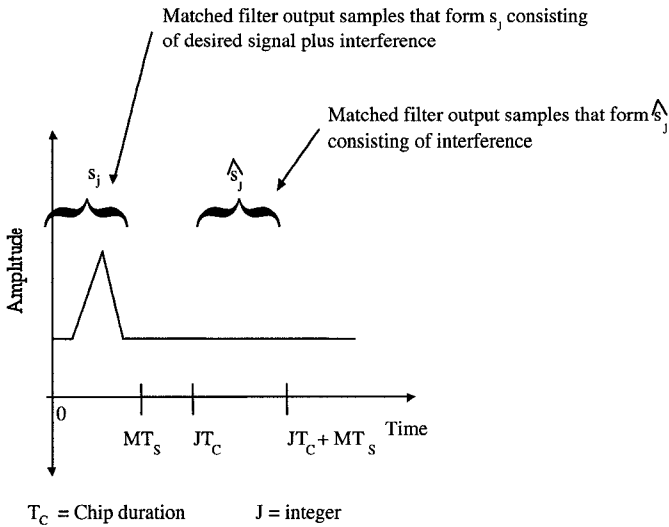


Fig. 2. Sampling scheme for the BAA at the output of each matched filter

method of gradient descent in the direction orthogonal to the expected signal vector \mathbf{z} , we obtain the following adaptive algorithm to solve the constrained minimization of the OE [1]:

Algorithm 1 (BAA) Given any $\mathbf{w}(0)$ such that $\mathbf{z}^H \mathbf{w}(0) \neq 0$, For $j > 0$, get $\mathbf{w}(j)$ by

$$\mathbf{w}(j) = c(j) \left\{ \mathbf{w}(j-1) + \mu [\mathbf{s}_j^H \mathbf{w}(j-1)] [(\hat{\mathbf{s}}_j^H \hat{\mathbf{s}}_j) \mathbf{s}_j - (\hat{\mathbf{s}}_j^H \mathbf{s}_j) \hat{\mathbf{s}}_j] \right\},$$

where $c(j)$ is chosen to stabilize the algorithm.

We can choose $c(j)$ in the BAA as below:

For each $j > 0$, let $w_1(j)$ be the first element of $\mathbf{w}(j)$. If $w_1(j) \neq 0$, choose $c(j)$ such that $\|\mathbf{w}(j)\| = 1$ and $\arg(w_1(j)) = 0$. Otherwise, choose $c(j)$ such that $\|\mathbf{w}(j)\| = 1$ and $\arg(c(j)) = 0$.

The computational complexity of the BAA is of order MD per symbol.

IV. ITERATIVE BLIND ADAPTIVE ALGORITHM

We note that the BAA updates the weight vector, \mathbf{w} , once per symbol interval. This often results in many symbols being processed while the weight vector is slowly converging to the appropriate solution. As a result, the time required for convergence may be too long for systems with large processing gains, such as GPS. This is an undesirable trait of the BAA. To eliminate this shortcoming, we extend the BAA described above to its iterative form—the iterative BAA. The iterative BAA improves convergence time via multiple updates of the weight vector per symbol. It is iterative in that each update of the weight vector corresponds to processing

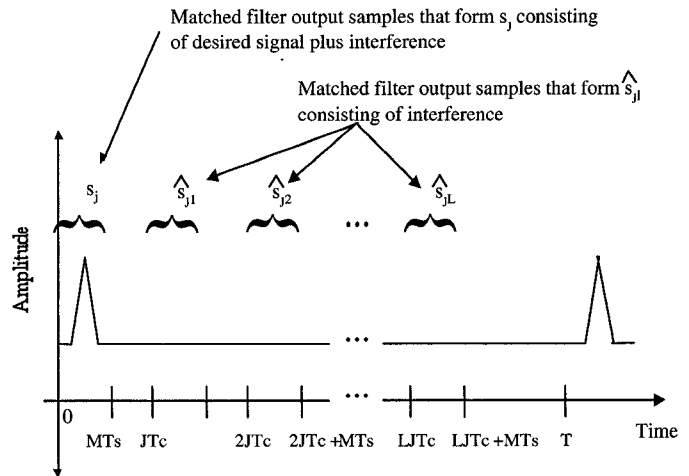


Fig. 3. Sampling scheme for the iterative BAA at the output of each matched filter

the same observed sample vector of signal with different observed vectors of interference. Fig. 3 depicts the corresponding sampling scheme used in the iterative BAA. Following the notation in Fig. 3, we can write the iterative BAA as follows:

Algorithm 2 (Iterative BAA) Given any $\mathbf{w}_L(0)$ such that $\mathbf{z}^H \mathbf{w}_L(0) \neq 0$, For $j > 0$, get $\mathbf{w}_1(j)$ by

$$\mathbf{w}_1(j) = c_1(j) \left\{ \mathbf{w}_L(j-1) + \mu [\mathbf{s}_j^H \mathbf{w}_L(j-1)] [(\hat{\mathbf{s}}_{j,1}^H \hat{\mathbf{s}}_{j,1}) \mathbf{s}_j - (\hat{\mathbf{s}}_{j,1}^H \mathbf{s}_j) \hat{\mathbf{s}}_{j,1}] \right\},$$

and for $l = 2, 3, \dots, L$, get $\mathbf{w}_l(j)$ by

$$\mathbf{w}_l(j) = c_l(j) \left\{ \mathbf{w}_{l-1}(j) + \mu [\mathbf{s}_j^H \mathbf{w}_{l-1}(j)] [(\hat{\mathbf{s}}_{j,l}^H \hat{\mathbf{s}}_{j,l}) \mathbf{s}_j - (\hat{\mathbf{s}}_{j,l}^H \mathbf{s}_j) \hat{\mathbf{s}}_{j,l}] \right\}.$$

As described before, $c_l(j)$ are chosen to stabilize the algorithm.

The computational complexity of the iterative BAA is of order LMD per symbol.

We note that the added interference observations in the iterative BAA not only help to increase the convergence rate, but also may improve the estimation of the interference covariance matrix \mathbf{R}_s as stated in [1]. This, in turn, may improve the performance of the algorithm in terms of obtaining a weight vector closer to the optimal one.

Another way to reduce the convergence time is to divide the long data symbol into shorter sub-symbols. Since the BAA makes no independence assumption on the statistic of the data symbols, we can apply the BAA on the sub-symbol streams. However, this method reduces the processing gain.

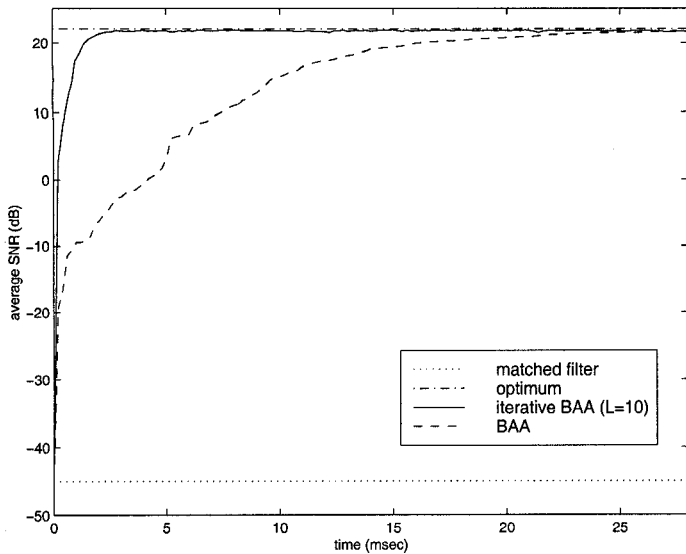


Fig. 4. Antenna example: average SNR versus time

In many cases, this is undesirable. The iterative BAA speeds up the convergence without sacrificing the processing gain.

V. PERFORMANCE

To demonstrate the behavior of the iterative BAA and compare it to the BAA, we consider a DS-SSMA system similar to the one employed in GPS [3]. We employ random sequences with 10MHz chip rate to approximate the P-code used in GPS. Instead of the 50Hz symbol rate used in the GPS standard, we take $T = 0.2\text{ms}$, i.e., we treat consecutive 0.2ms intervals as symbol intervals. This results in a processing gain of 33dB ($N = 2000$). Moreover, we assume the AWGN is at such a level that the output SNR for a single-element antenna would be 15dB if there were no jammer or multiple-access interferer.

We consider two sample cases to test the performance of the iterative BAA. The first case is a 2-user system employing a 5-element linear phased array to cancel MAI, i.e., $K = 2$ and $D = 5$. The received power of the interferer is 80dB stronger than that of the desired user. We sample once per symbol, i.e., $T_s = T_c$ and $M = 1$, and take 10 interference sample vectors for the iterative BAA, i.e., $L = 10$. For comparison, the BAA and iterative BAA use the same step size μ . Fig. 4 shows the output SNR's averaged over 20 simulation runs. The optimal output SNR in this case is 22dB which is 7dB higher than the 15dB level. This extra 7dB is due to the antenna array gain. We see that both the BAA and iterative BAA converge to the 22dB output SNR level. However, the convergence rate of the iterative BAA is about 10 times faster than that of the BAA (2.5msec versus 25msec). The factor 10 is due to the fact that we update 10 times per symbol for the iterative BAA.

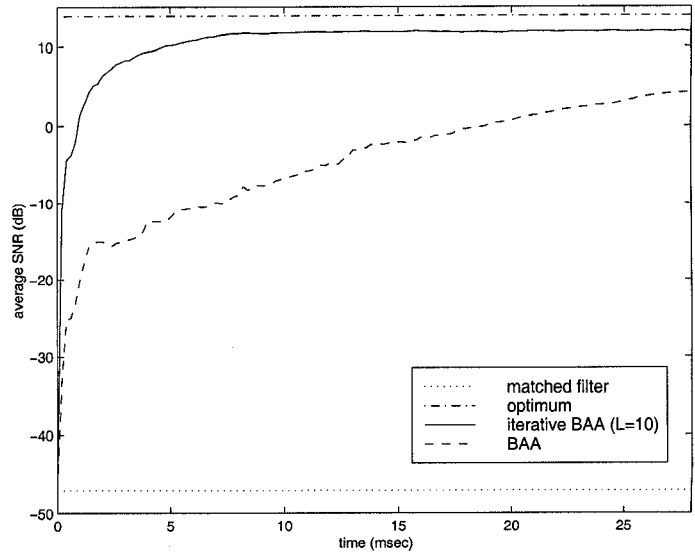


Fig. 5. Tone jammer example: average SNR versus time

The second sample case is a single-user system with a single-element antenna in the presence of a tone jammer at the carrier frequency. The jamming power is 80dB above the user power. To combat the jammer, we employ the sampling scheme with $M = 6$ and $T_s = 0.5T_c$. Again, we take 10 interference sample vectors for the iterative BAA, i.e., $L = 10$, and use the same μ for the BAA and iterative BAA. Fig. 5 shows the output SNR's averaged over 20 runs. In this case, the optimal output SNR is 14dB which is 1dB lower than the 15dB level. This reduction is due to the fact that when we notch out the tone jammer, the effect of AWGN is emphasized. From the figure, we see that the iterative BAA converges to a level of 12dB in 7.5msec. On the other hand, the BAA cannot reach the 12dB level even after 28msec.

The results above indicate that the iterative BAA outperforms the BAA in both cases. We need further investigation of the adaptation parameter μ for both the iterative BAA and the BAA in terms of compromising between faster convergence and better performance.

VI. CONCLUSIONS

In this paper, we have extended the linear receiver proposed in [1] and [2] for direct-sequence spread-spectrum multiple-access systems to its iterative form. The receiver consists of a matched filter followed by a tapped delay line that can be provisioned for use with antenna arrays. We have derived an iterative extension to the blind adaptive algorithm in [1] and [2]. This extension is most applicable in high processing gain situations, e.g., in GPS. Similar to the BAA, the iterative BAA has the same ability to combine multipaths, suppress MAI, and suppress narrowband interference. Sim-

ulation results for the cases with narrowband interference and MAI show that the iterative BAA demonstrates faster convergence time due to processing more measurements of interference outside the delay spread of the signal.

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