

An online Calibration Algorithm using binary spreading code for the CDMA-based Adaptive Antenna Array

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Abstract

In this paper, an iterative subspace-based calibration algorithm for a CDMA-based antenna array in the presence of unknown gain and phase error is presented. The algorithm does not depend on the array geometry and does not require a prior knowledge of the Directions Of Arrival (DOA) of the signals. The method requires the code sequence of a reference user only. The proposed algorithm is based on the subspace method and root finding approach, and it provides estimates of the calibration vector, the DOA and the channel impulse response, by using the code sequence of a reference user. The performance of the proposed algorithm was investigated by means of computer simulations and was verified using field data measured through a custom-built W-CDMA test-bed. The data show that experimental results match well with the theoretical calibration algorithm. Also, the study propose an efficient algorithm using the simulated annealing technique. This algorithm overcomes the requirement of initial guessing in the subspace-based approach.

Key Words : W-CDMA, Array Antenna, Subspace, Calibration, Channel Estimation

1. Introduction

Spatial processing techniques employing multiple antenna arrays have been proposed to improve the spectrum efficiency. The goal of the space-time processing is to combine spatial and temporal information. For the downlink beam-forming, accurate channel estimation such as DOA's and time delays are essential. Many

high-resolution DOA estimation algorithms, however, require a perfect knowledge of the array manifold, which is not possible in practice. The gain and phase responses of an antenna vary according to temperature and humidity changes from day to day [1], and therefore online calibration is preferable in wireless cellular communications. Many space-time algorithms for CDMA system have been proposed for channel estimation and for the optimum receiver by examining the covariance structure of the input signal. However, not many space-time algorithms for the downlink in Frequency Division Duplexing (FDD) have been proposed. This is partially

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because the channel information obtained in uplink can not be directly used for downlink beamforming in FDD. Especially, if the array of antenna is not well calibrated, then finding an optimum downlink beamforming algorithm gets more difficult.

In this study, the subspace-based online calibration algorithms for CDMA-based antenna array [2], which is not dependent on the geometry of the antenna array was verified experiment and data were collected. The data show that the algorithm based on the signal model described in [3] performs well under multiple access interferences. However, the convergence of the subspace-based online calibration algorithm is depends on a good guess of the initial values. Thus, this investigation proposes another method based on the simulated annealing technique [4], [5] which can avoid local minima. To verify the performance of the new algorithm, computer simulations were performed with varying parameters.

2. Subspace-based Approach

2.1 Signal Model

Assume that an antenna array is composed of M elements and that K_u users are in a cell. Suppose that the received continuous-time signals $z_m(t)$, ($m=1, \dots, M$) at the receiver front-end, are down-converted (by the IQ mixing at each antenna) to baseband and then converted to discrete-time signals by sampling the outputs of the integrator which integrates receiving signals over a subinterval $T_a = T_c/Q$ where T_c is the chip duration and Q is the oversampling factor. The multipath channel and the receiver front-end including the frequency offset are shown in Figure 1. According to this configuration, the obtained

complex sequence with an unknown complex antenna gain d_m can be expressed as:

$$z_m(q) = d_m \sum_{k=1}^{K_u} \sum_{l=1}^{L_k} \sqrt{P_{k,l}} \exp(j\phi_{k,l}^{k,l}) \beta_{k,l}(q) y_{k,l}(q) + n(q)$$

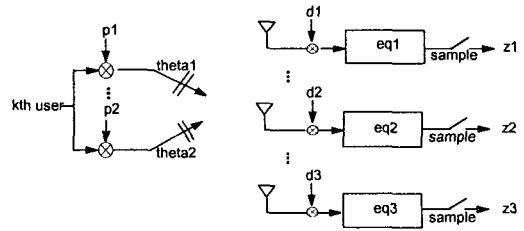


Fig. 1. The channel and receiver front-end

where $P_{k,l}$ and $\beta_{k,l}(q)$ are the received power and the envelope of the path fading, L_k is the number of multi-paths from the kth user, and $\phi_{k,l}^{k,l}$ is the phase delay due to the signals coming from the angle of $\theta_{k,l}$ (for the lth path from the kth user). The term $n(q)$ represents additive white Gaussian noise with zero-mean and covariance σ_n^2 at the receiver. The term $y_{k,l}(q)$ represents the chip matched filter output defined as

$$y_{k,l}(q) = \frac{1}{T_a} \int_{(q-1)T_a}^{qT_a} r_k(t - \tau_{k,l}) dt,$$

where $\tau_{k,l}$ is the time delay of the lth path from the kth user. The transmitted signal from the kth user is denoted as

$$r_k(t) = \sum_{i=-\infty}^{\infty} b_k(i) c_k(t - iT),$$

where $b_k(i)$ is the bit sequence, $c_k(t)$ is the spreading code waveform of the kth user. Here, $c_k(t) = 0$ for $t \notin [0, T]$ where T is the bit period. Let the received vector during one bit interval, $z_m(i) \in C^{QN}$ and the noise vector, $n_m(i) \in C^{QN}$, be defined as

$$z_m(i) = [z_m(iQN+1), \dots, z_m(iQN+QN)]^H,$$

$$n_m(i) = [n_m(iQN+1), \dots, n_m(iQN+QN)]^H,$$

where N represents the processing gain defined as $N=T/(T_c)$, and the superscript H, the complex conjugate transpose.

Since the system is assumed to be asynchronous, the observation vector will contain the end of the previous symbol and the beginning of the current symbol for each user. By adopting the asynchronous CDMA model in [2] and [3], the contribution $x_{m^{k,l}}(i)$ of the lth path from the kth user to the received vector,

$$z_m(i) = \sum_{k=1}^K \sum_{l=1}^{L_k} x_{m^{k,l}}(i) + n_m(i)$$

can be expressed as

$$x_{m^{k,l}}(i) = d_m \exp(j\phi_{m^{k,l}}) [u_{k,l}^R, u_{k,l}^L] \begin{bmatrix} \gamma_{k,l}(i-1) \\ \gamma_{k,l}(i) \end{bmatrix}$$

where $\gamma_{k,l}(i-1)$ and $\gamma_{k,l}(i)$ are complex constants which involve power, fading and transmitted bits. The vector pair $[u_{k,l}^R, u_{k,l}^L]$ has the form

$$u_{k,l}^R = U_k^R h_{k,l}, \quad u_{k,l}^L = U_k^L h_{k,l},$$

where $h_{k,l}$ is a vector of non-integer time delay and

$$U_k^R = [p_k^R(0) \dots p_k^R(N-1)],$$

$$U_k^L = [p_k^L(0) \dots p_k^L(N-1)].$$

The vector $p_k^R(\tau_{k,l})$ and $p_k^L(\tau_{k,l})$ are vectors of code sequence, which can be written as

$$p_k^R(\tau_{k,l}) = [0, \dots, 0, c_k(N-\tau_{k,l}), \dots, c_k(N-1)]^H$$

$$p_k^L(\tau_{k,l}) = [c_k(0), \dots, c_k(N-\tau_{k,l}-1), 0, \dots, 0]^H$$

where $c_k(t)$ is the spreading code of the kth user and the superscript H represents the complex conjugate transpose. The integer $\tau_{k,l}$ is time delay such that $\tau_{k,l} \in \{0, \dots, N-1\}$. Here, note that non-integer time delay can be expressed by choosing appropriate values in vector $h_{k,l}$.

After forming the matrix $Z = [z_1(i), \dots, z_M(i)]$, and stacking the row vectors of Z into an MN x 1 single composite snapshot vector $z(i)$, we obtain

$$z(i) = As(i) + n(i),$$

where

$$A = [a_{1,1}^R(\theta_{1,1}, h_{1,1}, d), \dots, a_{K,L_k}^L(\theta_{K,L_k}, h_{K,L_k}, d)],$$

$$[a_{k,l}^R(\cdot), a_{k,l}^L(\cdot)] = [u_{k,l}^R, u_{k,l}^L] \otimes (b(\theta_{k,l}) \odot d),$$

$$b(\theta_{k,l}) = [e^{j\phi^{k,l}}, e^{j\phi^{k,l}}, \dots, e^{j\phi^{k,l}}]^H$$

$$d = [d_1, \dots, d_M]^H,$$

$$s(i) = [\gamma_{1,1}(i-1), \gamma_{1,1}(i), \dots, \gamma_{K,L_k}(i-1), \gamma_{K,L_k}(i)]^H.$$

Here, \odot represents element-by-element multiplication, and \otimes the Kronecker product.

2.2 Algorithm

The algorithm is composed of two parts, one for estimating channel (including Direction of Arrival (DOA) θ , and impulse response h) and the other for calibration vector d. Without loss of generality, we assume that the first antenna element is the reference and

we consider only the relative magnitude of h. We also assume that the code sequence of reference user is known and the reference user transmits L multi-path signals.

We set the channel estimation problem as a minimization problem of which criterion is to

$$\min_{\theta} \mathbf{h}^H \mathbf{Q}_1(\theta) \mathbf{h} \quad \text{such that} \quad \|\mathbf{h}\|_2 = 1$$

where the superscript H is the complex conjugate operation and $\mathbf{Q}_1(\theta)$ is a matrix composed of array response and code sequence of reference user, which can be formulated as follows:

$$\mathbf{Q}_1(\theta) = \mathbf{B}_{L^u}(\theta) \mathbf{P}_S \mathbf{B}_L(\theta) + \mathbf{B}_{R^u}(\theta) \mathbf{P}_S \mathbf{B}_R(\theta). \quad (1)$$

Here, the $\mathbf{B}_L(\theta)$ and $\mathbf{B}_R(\theta)$ are the matrices of array response vector and code sequence vector, and the \mathbf{P}_S is an orthogonal projector onto subspace of the covariance matrix of measurement vector. Apparently, we can estimate $\theta_{l=1}$ by finding the local minima of minimum eigenvalues of $\mathbf{Q}_1(\theta)$ and the $\mathbf{h}_{l=1}$ by finding the eigenvector associated with the minimum eigenvalues of $\mathbf{Q}_1(\theta)$.

Next, we set the estimation of \mathbf{d} as a constraint minimization problem of which criterion is to

$$\min_{\tilde{\mathbf{d}}} \tilde{\mathbf{d}}^H \mathbf{Q}_2 \tilde{\mathbf{d}} \quad \text{such that} \quad \tilde{\mathbf{d}}^H \mathbf{w} = 1$$

where $\mathbf{w} = [1 \dots 0]$ and

$$\mathbf{Q}_2 = \sum_{l=1}^L \mathbf{B}_{L^u}(h_l, \theta_l) \mathbf{P}_S \mathbf{B}_L(h_l, \theta_l) + \mathbf{B}_{R^u}(h_l, \theta_l) \mathbf{P}_S \mathbf{B}_R(h_l, \theta_l). \quad (2)$$

Here, the $\mathbf{B}_L(h_l, \theta_l)$ and $\mathbf{B}_R(h_l, \theta_l)$ are the matrices of array response vector and code sequence vectors obtained from h_l and θ_l . Consequently, we can find the vector $\tilde{\mathbf{d}}$ by solving $\tilde{\mathbf{d}} = \mathbf{Q}_2^{-1} \mathbf{w} / \mathbf{w}^H \mathbf{Q}_2^{-1} \mathbf{w}$. We repeat estimation procedures using (1) and (2) until a cost function defined using (2) is less than a certain threshold.

2.3 Data Measurements

The data measurements were performed with a carrier frequency of 1.95[GHz]. At the receiver, a uniform circular array with eight elements was used. Each antenna was a dipole antenna and the space between two adjacent of them was half the wavelength. The received signal was down-converted using 70[MHz] IF and then sampled with 19.2[MHz]. The transmitter module was composed of a single dipole antenna, a signal generator and a power amplifier. In generating signal, the BPSK modulation was used and the chip rate was set at 3.84[Mcps]. The gold code of length 31 was used for PN sequences. To generate multi-users, different a Gold code was assigned to each user. A photograph of the antenna and receiver modules are shown in Figure 2.

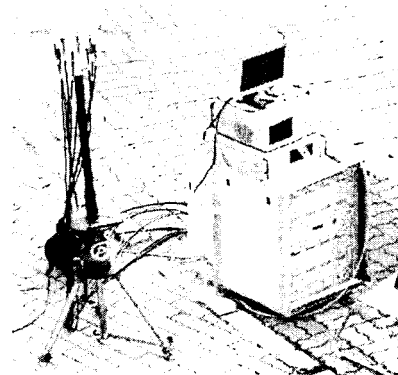


Fig. 2. Antennas and receiver module

To initially calibrate the receiver module, the receiver was manually adjusted by injecting a continuous-time signal. This operation calibrated all channels. Then, to obtain initial array response of antenna and receiver module, data were collected and measured by changing the transmitter location from -60 to 60 degree. Figure 3 illustrates the locations of the calibration sources and the measured array angle responses.

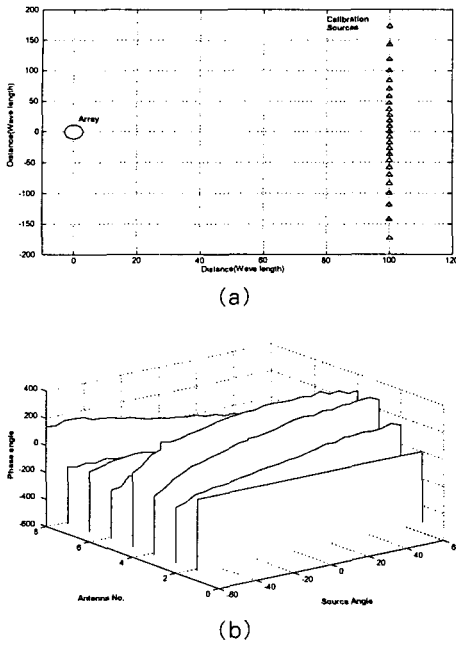


Fig. 3. (a) The calibration source and (b) The measured angle response of array

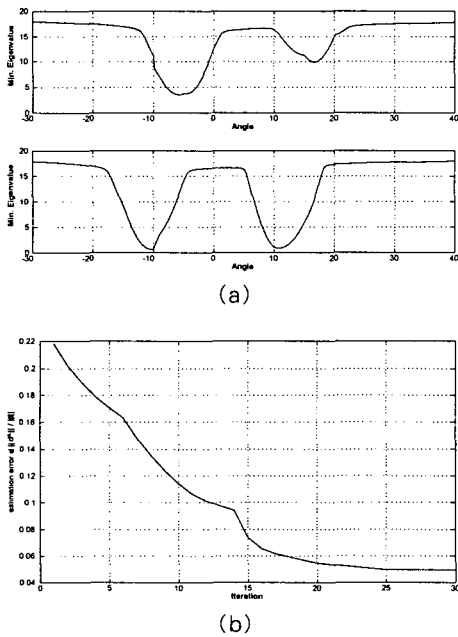


Fig. 4. (a) The estimated minimum eigenvalues at initial and last iteration (b)The normalized calibration error value vs. iterations

2.4 Experimental Results

During the experiment, reference user were selected as the signals measured at the angle of -10 and 10 degree using code number one. Then the time delays was set to be six and ten chips. As for multiple access interference signals, they were measured at the angle of $5, 12, 0, -5, -15$ and -30 degrees, of which code numbers and time delays were set arbitrary. The estimated minimum eigenvalues according to the DOA obtained at the initial and the last iteration are shown in Figure 4. Also, Figure 4 shows the normalized calibration error defined as

$$Error = \frac{\|d_j - d_t\|_2}{\|d_t\|_2}$$

where d_j and d_t are the gain-phase vector at the j th iteration and the true gain-phase vector,

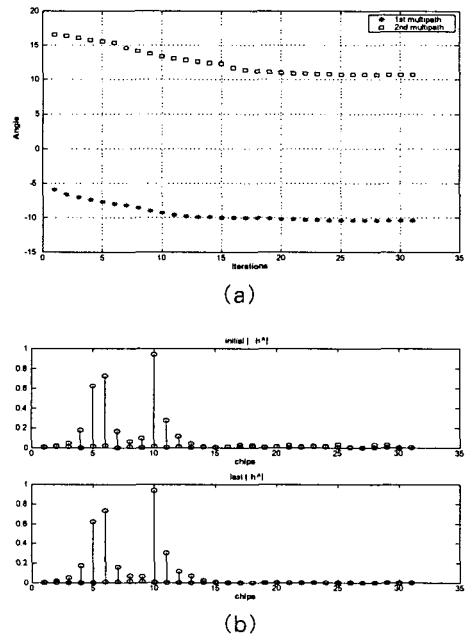


Fig. 5. (a) The estimated DOA vs. iterations (b) The estimated channel impulse response at initial and last iteration

respectively. The estimated DOA and channel impulse response according to the iterations are shown in Figure 5. As shown in these figures, it can be observed that the calibration algorithm performs well under multiple access interference.

However, the subspace-based approach has an initial value problem. With an inappropriate initial value, the iteration may converge to a local minimum.

3. Optimization Approach

3.1 Algorithm

We assume that antenna circular array is composed of M elements, the processing gain of code sequence is N and the same symbol, $s \in \{-1, 1\}$ are on the code all the time.

Suppose the signal is collected for one bit interval and formed into a vector. Then, we can write the signal as follows:

$$z = \sum_{l=1}^L (C h_l \otimes (b(\theta_l) \odot d)) s_l + n,$$

$$z = A(\theta_1, \dots, \theta_L, h_1, \dots, h_L, d) s + n, \quad (3)$$

where C is the code matrix, b is the array response vector, A is the function of DOA, channel impulse response h and calibration vector d , and n is the measurement noise vector. The matrix A is reformed as follows:

$$A = [C h_1 \otimes (b(\theta_1) \odot d) \dots C h_L \otimes (b(\theta_L) \odot d)]$$

$$A = \bar{A} H$$

where,

$$\bar{A} = [(C \otimes I_M)(b(\theta_1) \odot d) \dots (C \otimes I_M)(b(\theta_L) \odot d)]$$

$$H = \begin{bmatrix} h_1 & 0 & \dots & 0 \\ 0 & h_2 & 0 & 0 \\ \vdots & 0 & \ddots & \vdots \\ 0 & 0 & \dots & h_L \end{bmatrix}$$

Here, I_M is the $M \times M$ identity matrix. Then, (3) can be rewritten by

$$z = \bar{A} H s + n. \quad (4)$$

The least squares solution of (4) is given by

$$\hat{H} s = \bar{A}^+ z,$$

where D^+ is the pseudo inverse of the matrix D . The $\hat{H} s$ is the function of DOA θ and calibration vector d . The elements of the calibration vector is assumed to be normalized as $|d_i| = 1$. Thus, the d can be represented as $d_i = e^{j\phi_i}$. Our choice of the elements of the state vector are as:

$$x = [\theta_1 \dots \theta_L \quad \phi_1 \dots \phi_M]^T.$$

Then, the cost function $J(x)$ is

$$J(x) = \|z - \bar{A} \bar{A}^+ z\|_F,$$

where the subscript F is the Frobenious norm. To minimize the cost function, which has several minima as shown in Figure 6, we use the simulated annealing technique.

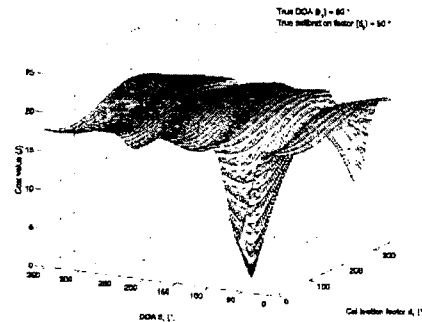


Fig. 6. The cost function w.r.t θ and ϕ

The simulated annealing technique is used to obtain the state vector which is close enough to

the global minimum. The initial state x_0 can be chosen arbitrary anywhere in the state space. This can be done by using the constraint condition,

$$\Gamma \subset R^{M+L}: 0^\circ < \theta < 360^\circ, 0^\circ < \phi < 360^\circ.$$

Because each element of the calibration vector can be normalized, we only consider the phase ϕ . The simulated annealing technique is summarized below:

Procedure: Simulated Annealing

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Choose  $x_0$ 
Compute  $J_0 = J(x_0)$  using (5)
IF  $|J(x_0)| < \epsilon$ , then break;  $\epsilon$  is a threshold.
End If
While  $(|J(x_0)| > \epsilon)$ 
    For  $k = 1, 2, \dots$ 
        I : Generate a random direction,  $\Delta r u$ ;
             $\Delta r$  is a step size and  $u$  is a
            standard normal variate.
            Set  $x^* = x_0 + \Delta r u$ 
            If  $x^* \in \Gamma$  then  $J_1 = J(x^*)$ 
                and  $\Delta J = J_1 + J_0$ 
            Else goto step I
            End If
            If  $J_1 < J_0$  then) set  $x^* = x_0$ 
                and  $J_0 = J_1$ .
                If  $|J_1| < \epsilon$  then, break.
                Else goto step I
                End If
            Else set  $P = \exp(-T\Delta J)$ ;  $T$  is a
            positive parameter.
                Generate a uniform variate,
                 $V$  on  $[0, 1]$ 
                If  $V > P$  then goto step I
                Else set  $x_0 = x^*$  and  $J_0 = J_1$ .
                    Goto step I
                End If
    End If

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End If
End For
Change  $\Delta r$  and  $T$ 
End While

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The simulated annealing technique always finds the global minimum starting with any initial state even when the cost function $J(x)$ has many local minima [6].

3.2 Simulation Results

To test the algorithm, we consider an uniform circular array with eight antennas separated by half a wavelength ($M=8$). We assume that the number of multipath signals of the reference user is known ($L=2$). We use 10 observation symbols with no over-sampling. The DOA's of the reference user are $60^\circ, 250^\circ$. The true phase of calibration vector are $20^\circ, 70^\circ, 200^\circ, 300^\circ, 120^\circ, 60^\circ, 150^\circ, 260^\circ$. The signal-to-noise ratio is assumed to be 20[dB]. In the simulated annealing phase, the parameters are listed in Table I.

Table 1. Parameters in the Simulated Annealing

Iteration	ϵ	Calibration ϕ		DOA: θ	
		T1	Δr	T2	Δr
1	3.0	12	5°	0.5	5°
2	0.5	12	5°	1.0	5°
3	0.2	12	2.5°	7.0	2.5°

We start the iteration with the initial state $\theta_{0,i} = 180^\circ$ ($i=1,2$) and $\phi_{0,i} = 180^\circ$ ($i=1,2$) ($i=1,\dots,8$). The initial calibration error $Error = \frac{\|d_i - \hat{d}_i\|_2}{\|d_i\|_2}$ and initial DOA estimation error are 1.3567 and $120^\circ, -70^\circ$, respectively. Figure 7 shows the values of the cost function of the proposed algorithm along the iteration steps. The total size

of the iterations is 62073. Accepted size of iterations is 39212. The converged state values are listed in

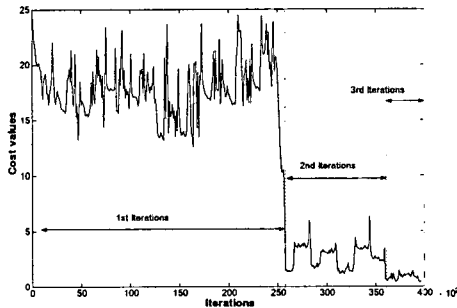


Fig. 7. Cost Value of the proposed Algorithm

Table 2. Converged State Values

Iteration	J	DOA: θ	Error
1	2.68	[56.34[°] 251.11[°]]	0.101
2	0.39	[58.89[°] 250.50[°]]	0.056
3	0.16	[59.41[°] 250.40[°]]	0.027

4. Conclusions

In this paper, we have proposed algorithms for calibration which can be used in asynchronous CDMA systems. The algorithm requires the code sequence of any reference user which is already available at the base station. By using the sequence, the algorithm provides estimates of the calibration vector, the DOA's and the channel impulse response. Throughout the computer simulations and experiments on the field measured data, this study demonstrated that the proposed algorithm is proven to work even when the array antenna is not properly calibrated. Furthermore, an efficient algorithm which is independent of a good guess of the initial state have also been presented.

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Biography

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