비동기 CDMA 시스템 기반의 배열 안테나용 온라인 보정 알고리즘

정회원 이 종 현

On-line Calibration algorithm for Asynchronous CDMA-based antenna arrays

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요 약

본 논문에서는 비동기 CDMA 시스템에서 발생하는 배열 안테나 소자의 오차 보정(calibration) 문제를 다룬다. 수신기의 주파수 오프셋 오차가 있을 경우에 적용 가능한 새로운 순차적인 오차 보정 알고리즘을 제안한다. 제안된 알고리즘은 비선형 배열 안테나에도 적용가능하며, 사전에 알고 있는 신호를 사용하여 보정하는 것이 아니라, 셀 내의 임의의 사용자 PN code 만을 사용하여 온라인으로 오차 보정을 수행한다. 제안된 알고리즘은 채널과 주파수 오차를 추정하는 부분과 배열안테나 보정을 추정하는 부분으로 구성되어 있으며, 컴퓨터 모의 실험과 W-CDMA용 스마트 안테나 테스트 베드에서 얻어진 실측 데이터를 이용한 실험을 통하여 그 성능이 검증되었다.

Keyword: Asynchronous CDMA, Antenna Array, Calibration, PN code, Frequency Offset

ABSTRACT

In this paper, the calibration problem of an asynchronous CDMA-based antenna array is studied. A new iterative calibration algorithm for antenna array in the presence of frequency offset error is presented. The algorithm is applicable to a non-linear array and does not require a prior knowledge of the (direction of arrivals) DOAs of the signals of any user, and it only requires the code sequence of a reference user. The algorithm is based on the two step procedures, one for estimating both channel and frequency offset and the other for estimating the unknown array gain and phase. Consequently, estimates of the DOAs, the multi-path impulse response of the reference signal sources, and the carrier frequency offset as well as the calibration of antenna array are provided. The performance of the proposed algorithm is investigated by means of computer simulations and is verified by using field data measured through a custom-built W-CDMA test-bed.

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논문번호 : 030091-0306, 접수일자 : 2003년 3월 6일

I. Introduction

It is well known that space-time processing techniques employing multiple antennas can be used to increase spectrum efficiency and capacity for future mobile communications. Since the goal of the space-time processing is to combine spatial and temporal information effectively, accurate channel estimation of DOAs and time delays is essential. An efficient space-time channel estimation algorithm was proposed in [1]. Many high-resolution DOA estimation algorithms, however, require perfect knowledge of the array manifold, which is not possible in practice.

Various array calibration methods with or without known source directions have been proposed. Since the amplitude and phase of an antenna varies according to temperature and humidity changes from day to day [2], and thus online calibration is preferable in wireless mobile communications. However, deploying known signals at a known location may not be tolerable since it would increase multiple access interference and decreases the channel capacity as well. To cope with this problem, a calibration algorithm for CDMA-based antenna array was reported in [3]. The algorithm proposed in [3] can be applied to asynchronous CDMA system, and its performance was proven to be effective via computer and field measurement data.

In this parer, the calibration problem is studied even when the receiver undergoes carrier frequency error. In general, most digital communication systems involve frequency offset errors caused by either unstable oscillator or Doppler effect. It is well known that this error causes degradation of the system performance. To solve this problem, many algorithms have been reported in [4] and [5] in which training sequences is used to obtain the frequency offset.

To cope with calibration problem including frequency offset, a new iterative calibration algorithm which estimates channel parameters and the carrier offset as well as calibration vector is presented. The algorithm is based on date model made by incorporating the frequency offset term with the model for an asynchronous CDMA-based antenna array in [6] and [7]. The algorithm is based on the two step procedures, one for estimating both channel and frequency offset and the other for estimating the unknown array gain and phase. The algorithm utilizes the code sequence of any reference user and is applicable to the situation where the total number of signals is less than M times N, where M is the number of array and N is processing gain of code sequence. The algorithm is applicable to a non-uniform array and does not require a prior knowledge of the channel information such as multi-path delays and DOAs of the signals. To verify the performance of the algorithm, computer simulations and experiments on field measured data have been performed.

II. Data Formulation

Assume that antenna array is composed of M elements and K_a users are in a cell. Suppose that the received signals at the array are sampled at chip rate T_c and have the identical frequency offset, Δf for all K_a users. The multipath channel and the receiver front-end including frequency offset is shown in Figure 1.

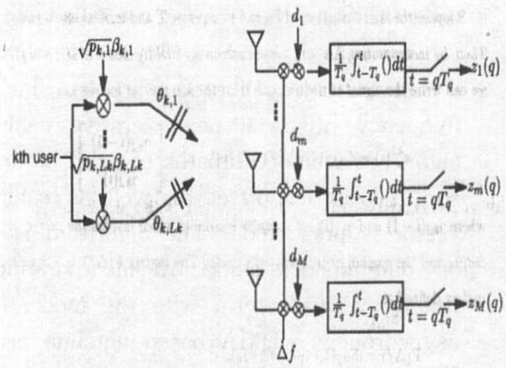


Figure 1　The channel and receiver front end including frequency offset

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

$$z_m(q) = d_m \exp(j2\pi\Delta f q T_c) \sum_{k=1}^{K_s} \sum_{l=1}^{L_k} \sqrt{P_{k,l}} \exp(j\phi_m^{k,l}) \beta_{k,l}(q) y_{k,l}(q) + n(q)$$

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 $\phi_m^{k,l}$ 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 of the transmitted signal from the kth user.

Suppose the signal is collected for one bit interval T and formed into a vector. Then, by incorporating Δf with the asynchronous CDMA model in [6] and [7], we can write the signal of the lth path from the kth user as follows:

where $\gamma_{k,l}(i-1)$ and $\gamma_{k,l}(i)$ are complex constants which involve the power, the fading and the symbol of transmitted signals. The matrix $\boldsymbol{F}(\Delta f)$ is a diagonal matrix defined as

$$\boldsymbol{F}(\Delta f) = diag(1, \exp(j2\pi\Delta f T_c), \cdots, \exp(j2\pi\Delta f(N-1)T_c))$$

where N represents the processing gain defined as N = T/Tc.

The vector pair $[\boldsymbol{u}_{k,l}^R, \boldsymbol{u}_{k,l}^L]$ has the form

$$\boldsymbol{u}_{k,l}^R = U_k^R \boldsymbol{h}_{k,l}, \quad \boldsymbol{u}_{k,l}^L = U_k^L \boldsymbol{h}_{k,l},$$

where $\mathbf{h}_{k,l}$ is a vector of non-integer time delay and

$$U_k^R = [\boldsymbol{p}_k^R(0) \ldots \boldsymbol{p}_k^R(N-1)], \quad U_k^L = [\boldsymbol{p}_k^L(0) \ldots \boldsymbol{p}_k^L(N-1)].$$

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

$$\boldsymbol{p}_k^R(\tau_{k,l}) = [0, \ldots, 0, \ c_k(N-\tau_{k,l}), \ldots, c_k(N-1)]^H$$

$$\boldsymbol{p}_k^L(\tau_{k,l}) = [c_k(0), \ldots, c_k(N-\tau_{k,l}-1), \ 0, \ldots, 0]^H$$

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 $\tau_{k,l} \in \{0, \cdots, N-1\}$. Here, note that non-integer time delay can be expressed by choosing appropriate values in vector $\boldsymbol{h}_{k,l}$.

After forming the matrix $Z = [z_1(i), \cdots, z_M(i)]$, 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

$$A = [\, a^R_{1,1}(\Delta f, \theta_{1,1}, h_{1,1}, d), a^L_{1,1}(\Delta f, \theta_{1,1}, h_{1,1}, d), \ldots, a^L_{K,L_K}(\Delta f, \theta_{K_a,L_K}, h_{K_a,L_K}, d)],$$

$$[\, a^R_{k,l}(\cdot), a^L_{k,l}(\cdot)] = F(\Delta f)[\, u^R_{k,l}, u^L_{k,l}] \otimes (\, b(\theta_{k,l}) \odot d),$$

$$b(\theta_{k,l}) = [\, e^{j\phi^{k,l_1}}, e^{j\phi^{k,l_2}}, \ldots, e^{j\phi^{k,l_M}}]^H,$$

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

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

Ⅲ. Estimation of Frequency offset, channel and calibration vector

Without loss of generality, let us assume that the user using code vector $p_1(\tau)$ is reference user and **h** is the vector of time delay to be estimated. By using the received signal vector z, we perform the eigen-decomposition of R_{zz}:

$$R_{zz} = E\{\, zz^H\} = VDV^H = [\, E_S, E_N]D[\, E_S, E_N]^H$$

where E_S and E_N span the signal space and the noise space, respectively. Then, we define MUSIC-like cost function as follows:

$$J_1(\Delta f, h, \theta, d) = \| E^H_N\, v^R_1(\Delta f, \theta, d) \|^2_2 + \| E^H_N\, v^L_1(\Delta f, \theta, d) \|^2_2, \tag{1}$$

where

$$v^R_1(\Delta f, \theta, d) = F(\Delta f) u^R_1 \otimes (\, b(\theta) \odot d),$$
$$v^L_1(\Delta f, \theta, d) = F(\Delta f) u^L_1 \otimes (\, b(\theta) \odot d),$$

The vector u^R_1 and u^L_1 can be written as

$$u^R_1 = U^R_1 h, \quad u^L_1 = U^L_1 h,$$

where

$$U^R_1 = [\, p^R_1(0) \ldots p^R_1(N-1)],\ U^L_1 = [\, p^L_1(0) \ldots p^L_1(N-1)].$$

For simplicity, let us assume that estimate of **d** is available. Then, in order to find the channel parameters and frequency offset, a multi-dimensional search would be needed in the space of **h**, Δf and θ. Instead, by using the ''Mixed Product Rule'' of kronecker product of $(AC \otimes BD) = (A \otimes B)(C \otimes D)$ [8], we rewrite $v^R_1(\cdot)$ and $v^L_1(\cdot)$ as

$$v^R_1(\Delta f, \theta, d) = C^R_1(\Delta f) B(\theta) h,$$
$$v^L_1(\Delta f, \theta, d) = C^L_1(\Delta f) B(\theta) h$$

where

$$C^R_1(\Delta f) = [\, F(\Delta f) U^R_1] \otimes I_M,$$
$$C^L_1(\Delta f) = [\, F(\Delta f) U^L_1] \otimes I_M. \tag{2}$$

The matrix I_M is the identity matrix of size M. The matrix $B(\theta)$ which contains both angle and calibration information can be written as:

$$B(\theta) = I_N \otimes (b(\theta) \odot d). \quad (3)$$

Then the cost function $J_1(\Delta f, h, \theta, d)$ can be written, with respect to **h** and θ as

$$J_1(\Delta f, \boldsymbol{h}, \theta, \boldsymbol{d}) = \boldsymbol{h}^H \boldsymbol{Q}_1(\theta, \Delta f)\boldsymbol{h}, \quad (4)$$

where

$$\boldsymbol{Q}_1(\theta, \Delta f) = B^H(\theta)[C_1^R(\Delta f)^H E_N E_N^H C_1^R(\Delta f) + C_1^L(\Delta f)^H E_N E_N^H C_1^L(\Delta f)]B(\theta).$$

Here note that since transmitted multi-path fading is unknown, we can only determine **h** to within a complex constant. Thus, by introducing the constraint of $\| h \|_2 = 1$, the solution for **h** is given by

$$h = \nu_{\min}(\boldsymbol{Q}_1(\theta, \Delta f)), \quad (5)$$

where $\nu_{\min}$ denotes the eigenvector of $\boldsymbol{Q}_1(\theta, \Delta f)$ associated with the minimum eigenvalue $\lambda_{\min}(\boldsymbol{Q}_1(\theta, \Delta f))$.

By substituting (5) to (4), we get the function of angle parameter θ,

$$\theta = \arg \min_\theta \lambda_{\min}(\boldsymbol{Q}_1(\theta, \Delta f)). \quad (6)$$

Here, note that $\boldsymbol{Q}_1(\theta, \Delta f)$ is a function of θ and Δf. Thus in order to find **h**, we need to find the eigenvector corresponding the minimum eigenvalue by searching the space of θ and Δf. However, by using the fact that θ and Δf are the irrelevant parameters, a suboptimal algorithm is possible, which finds the one parameter (say θ) by fixing the other parameter (say Δf) or vice versa.

In short, the proposed algorithm can be described as follows:

(1) First, with initial calibration vector **d** and frequency offset Δf, we obtain estimates of $\boldsymbol{h}_1$ and θ by finding the values minimizing the J_1 in (4).

(2) Next by using the obtained channel parameters, we find channel offset Δf by finding the root of the polynomial obtained by replacing $\exp(j2\pi\Delta f T_c)$ by z.

(3) Finally, we find the calibration vector **d** which minimizes a certain cost function J_3, which described in next section, with the obtained channel parameters and the frequency offset.

The above steps are iterated until the cost function converges J_3 to the minimum value. By assuming that the first user is the reference user, we shall summarize the estimation procedure below

3.1 Estimation of channel parameters and frequency offset

Procedure 1

With estimated (Δf) and **d**,

(A) Formulate $\boldsymbol{Q}_1(\theta_i, \Delta f)$ using (2) and (3) for each discrete $0 \le \theta_i \le \pi$, $(1 \le i \le I_\theta)$;

(B) Find and record the minimum eigenvalue $\lambda_{\min}^{(i)}$ of $\boldsymbol{Q}_1(\theta_i, \Delta f)$;

(C) Plot $\{\lambda_{\min}^{(i)}\}_{i=1}^{I_\theta}$ with respect to $\{\theta_i\}_{i=1}^{I_\theta}$, and select L_1 local minima $\{\theta_{1,i}\}_{i=1}^{L_1}$; Compute the $\{\boldsymbol{h}_{1,i}\}_{i=1}^{L_1}$ corresponding $\{\theta_i\}_{i=1}^{I_\theta}$ from the plot;

(D) Using the estimated $\{\theta_{1,i}\}_{i=1}^{L_1}$, and $\{\boldsymbol{h}_{1,i}\}_{i=1}^{L_1}$ formulate a cost function J_2

$$J_2(\Delta f, \mathbf{h}_1, \theta, \mathbf{d}) = \sum_{i=1}^{L_1}(\| E_N^H \ \hat{\mathbf{v}}_1^{\ R}(\theta_{1,i}, \mathbf{h}_{1,i}, \Delta f) \|_2^2 + \| E_N^H \ \hat{\mathbf{v}}_1^{\ L}(\theta_{1,i}, \mathbf{h}_{1,i}, \Delta f) \|_2^2) \quad (7)$$

where

$$\hat{\mathbf{v}}_1^{\ R}(\theta_{1,i}, \mathbf{h}_{1,i}, \Delta f) = [\mathbf{F}(\Delta f) U_1^R \mathbf{h}_{1,i}] \otimes (\mathbf{b}(\theta_{1,i}) \odot \mathbf{d}),$$

$$\hat{\mathbf{v}}_1^{\ L}(\theta_{1,i}, \mathbf{h}_{1,i}, \Delta f) = [\mathbf{F}(\Delta f) U_1^L \mathbf{h}_{1,i}] \otimes (\mathbf{b}(\theta_{1,i}) \odot \mathbf{d}),$$

(E) By replacing $\exp(j2\pi\Delta f T_s)$ in matrix $\mathbf{F}(\Delta f)$ by z, reformulate the $J_2(\cdot)$ in (7) into a polynomial function. Find the roots of $J_2(\cdot)$ and choose a root, z_o which is closest to unit circle. Then the frequency offset estimate can be obtained as follows:

$$\Delta f = \frac{1}{2\pi T_c} \arg(z_o)$$

End of Procedure 1

3.2 Estimation of calibration vector d

With the estimates $\mathbf{h}_{1,i}, \Delta f$ and $\theta_{1,i}$ obtained in the section 3.1, we can find the frequency compensated code vectors $\hat{\mathbf{u}}_i^{\ R}$ and $\hat{\mathbf{u}}_i^{\ L}$, $(i=1,\ldots,L_1)$ by using the following equation

$$\hat{\mathbf{u}}_i^{\ R} = \mathbf{F}(\Delta f) U_1^R \mathbf{h}_{1,i}, \quad \hat{\mathbf{u}}_i^{\ L} = \mathbf{F}(\Delta f) U_1^L \mathbf{h}_{1,i}. \quad (8)$$

By substituting the vectors of $\hat{\mathbf{u}}_i^{\ R}$, $\hat{\mathbf{u}}_i^{\ L}$ and $\theta_{1,i}, (i=1,\ldots,L_1)$ into equation (7), we define a cost function J_3 defined as follows

$$J_3(\mathbf{d}, \mathbf{h}_1, \theta, \Delta f) = \sum_{i=1}^{L_1}(\| E_N^H \ \hat{\mathbf{v}}_1^{\ R}(\mathbf{d}) \|_2^2 + \| E_N^H \ \hat{\mathbf{v}}_1^{\ L}(\mathbf{d}) \|_2^2)$$

where

$$\hat{\mathbf{v}}_1^{\ R}(\mathbf{d}) = \hat{\mathbf{u}}_i^{\ R} \otimes (\mathbf{b}(\theta_{1,i}) \odot \mathbf{d}),$$

$$\hat{\mathbf{v}}_1^{\ L}(\mathbf{d}) = \hat{\mathbf{u}}_i^{\ L} \otimes (\mathbf{b}(\theta_{1,i}) \odot \mathbf{d}),$$

This function represents the current cost value obtained with the current estimates of channel parameters and calibration vector **d**. Note that $\hat{v}_1^{\ R}(d)$ and $\hat{v}_1^{\ L}(d)$ can be rewritten as

$$\hat{\mathbf{v}}_1^{\ R}(\mathbf{d}) = \hat{U}_R \hat{B}_i \mathbf{d}, \quad \hat{\mathbf{v}}_1^{\ L}(\mathbf{d}) = \hat{U}_L \hat{B}_i \mathbf{d},$$

where

$$\hat{U}_i^{\ R} = diag([u_1^R, u_1^R, \ldots, u_1^R \ \cdots \ u_N^R, u_N^R, \ldots, u_N^R])$$
$$\hat{U}_i^{\ L} = diag([u_1^L, u_1^L, \ldots, u_1^L \ \cdots \ u_N^L, u_N^L, \ldots, u_N^L]) \quad (9)$$

$$\hat{B}_i = [\ \tilde{\mathbf{b}}^H(\theta_{1,i}), \ldots, \ \tilde{\mathbf{b}}^H(\theta_{1,i})]^H.$$

Here, $\tilde{\mathbf{b}}(\theta_{1,i}) = diag(\mathbf{b}(\theta_{1,i}))$, $\mathbf{u}_j^R$ and $\mathbf{u}_j^L, (j=1,\ldots,N)$ are the jth elements of $\hat{\mathbf{u}}_i^{\ R}$ and $\hat{\mathbf{u}}_i^{\ L}$, respectively. Then the cost function J_3 can be simplified as

$$J_3(\mathbf{d}; \Delta f, \mathbf{h}_1, \theta) = \mathbf{d}^H \ \mathbf{Q}_2 \mathbf{d},$$

where

$$\mathbf{Q}_2 = \sum_{i=1}^{L_1} B_i^H [\ \hat{U}_i^{\ RH} E_N E_N^H \hat{U}_i^{\ R} + \hat{U}_i^{\ LH} E_N E_N^H \hat{U}_i^{\ L}] B_i.$$

Here the first antenna element is assumed to be reference one since relative magnitude and phase responses of antenna elements are needed. Thus the minimization of $J_3(\cdot)$ with respect to **d** can be written as constraint minimization problem which is described as :

$$\boldsymbol{d}^H \boldsymbol{w} = 1, \qquad (10)$$

where $\boldsymbol{w} = [1, 0, \cdots, 0]^H$

With this constraint, we shall summarize the procedure for estimating **d** below:

Procedure 2.

(A) Formulate $\boldsymbol{Q}_2$ using (8), (9) and $\{\theta_{1,i}\}_{i=1}^{L_1}$;

(B) Find $\boldsymbol{d}_{new}$ which satisfies the equation in (10);

(C) Compute the following equation

$$J_{new} = \boldsymbol{d}_{new^H} \boldsymbol{Q}_2 \boldsymbol{d}_{new};$$

If $J_2 - J_{new} >$ *Threshold*, then go to Procedure 1, and repeat Procedure 2;

If $J_2 - J_{new} \le$ *Threshold*, then terminate;

End of Procedure 2

Ⅳ. Experimental Results

4.1 Computer Simulation Results

In computer simulation, the first element is set to be reference one such that $\boldsymbol{d}(1) = 1$. An uniform circular array with six antennas separated by half a wavelength is used. The random BPSK modulated data streams, and the Gold codes with the processing gain of N = 31 is adopted for simulation. It is assumed that 15 users (K = 15) produce two multi-path signals ($L_k = 2$). For simplicity, azimuthal angle is considered only. The DOAs of the reference user are assumed to be [40˚, 85˚] and the delays be [19.3, 25.3] chips. The DOAs and delays of the rest of users are randomly generated between [0, 180] and [0, 31], respectively. The frequency offset Δf is assumed to be 0.1.

It is assumed that the number of multi-path signals of the reference user is known. The 400 observation symbols, which are obtained without over-sampling, are used for estimation procedure. The signal-to-noise ratio (SNR) is assumed to be 20 dB, and the total power of an interfering user is twice of that of the reference user. The gain-phase $\alpha_m e^{j\psi_m}$ of each antenna is chosen as

$$\alpha_m = 1 + \sqrt{12}\sigma_\alpha u_m, \qquad \psi_m = \sqrt{12}\sigma_\psi u_p$$

where u_m and u_p are uniformly distributed in [-0.5, 0.5] and $\sigma_\alpha = 0.2$ $\sigma_\psi = 20˚$.

By using the parameters of described above, we plot the minimum eigenvalues. The result obtained when the frequency offset is not compensated, is shown in Figure 2 and Figure 3.

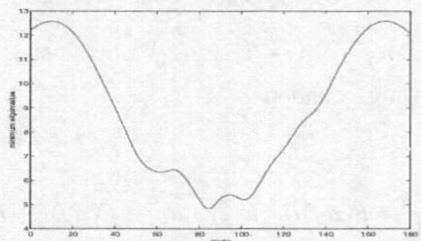


Figure 2. Minimum eigenvalues obtained using un-calibrated array of un-compensated frequency offset

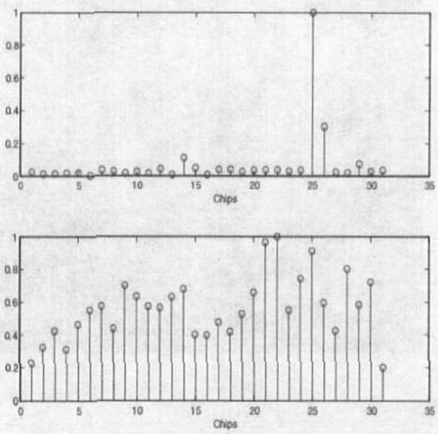


Figure 3. time delays obtained usin un-calibrated array of un-compensate frequency offset

The result obtained when the frequency offset is compensated but the array is not calibrated, is shown in Figure 4. The result obtained when the array is calibrated and the frequency offset is compensated, is shown in Figure 5 and Figure 6.

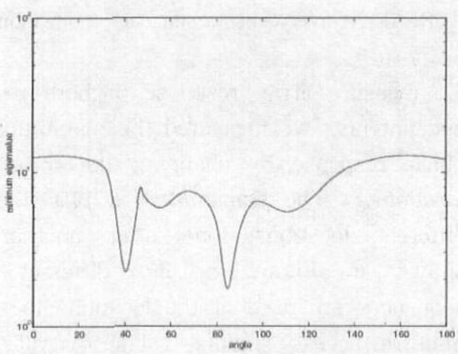


Figure 4. Minimum eigenvalues obtained usin un-calibrated array of compensated frequenc offset

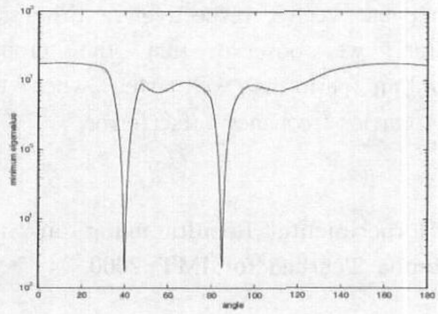


Figure 5. Minimum eigenvalues obtained using calibrated array of compensated frequency offset

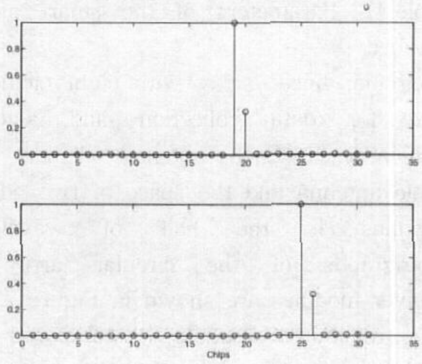


Figure 6. time delays obtained using calibrated array of frequency compensation

In these figures, we observe that when frequency offset is not compensated, we cannot estimate channel parameters in both space and time as well as the calibration vector **d**.

To measure the calibration vector estimation error, we use the normalized calibration error defined as follows:

$$Error = \frac{\| \boldsymbol{d}_j - \boldsymbol{d}_t \|_2}{\| \boldsymbol{d}_t \|_2},$$

where $\boldsymbol{d}_j$ and $\boldsymbol{d}_t$ are the gain-phase vector at the *j*th iteration and the true gain-phase vector, respectively. From these figures, we observe that the proposed algorithm performs well even when there exits carrier frequency offset error.

4.2 Experimental Results using an Smart Antenna Testbed for IMT-2000

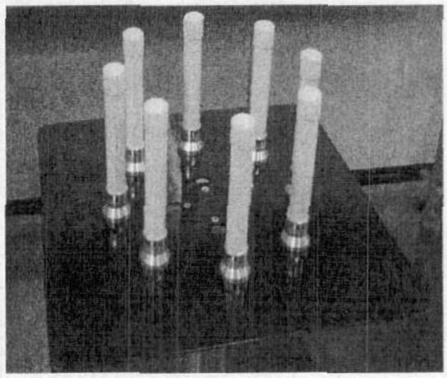
To verify our algorithm, we use field *measurement data obtained from a smart* antenna test-bed. The parameters of the test-bed are described in the table as follows:

Table 1: Parameters of the smart antenna test-bed

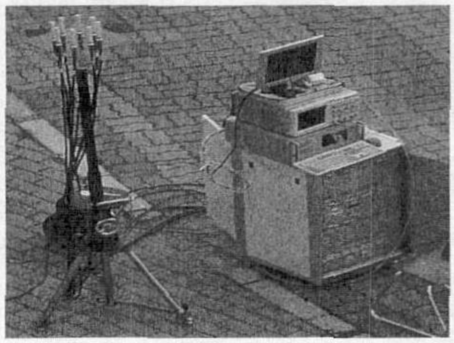
A circular linear array with eight element is used for data collection and algorithm verification. Each antenna element is a dipole antenna and the space of two adjacent antennas is the half of wavelength. Photographs of the circular array and receiver modules are shown in Figure 7.

The transmitter module is composed of a single dipole antenna, a signal generator and a power amplifier. In generating signal, BPSK modulation is used and chip rate is set to be 3.84 Mcps. The gold code of length 31 is used for PN sequence.

Parameters	Value
Number of antennas/channels	8
Carrier frequency	1950 MHz
IF frequency	70 MHz
Chip rate	3.84 Mcps
Modulation scheme	BPSK
Pulse shaping filter	raised cosine with & roll-off factor 0.65
Processing gain	31
A/D sampling rate	19.2 Msps
A/D Resolution	8 bits
Oversampling factor	1



(a)



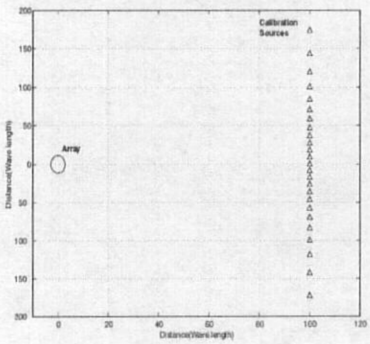
(b)

Figure 7 RF/IF Receiver Modules

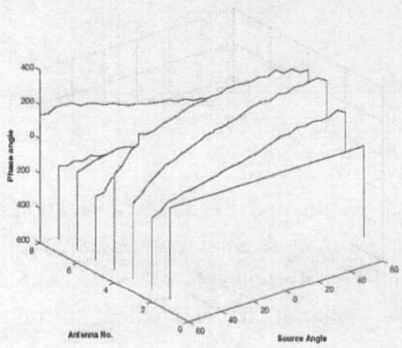
(a)Eight-element uniform circular array antenna.

(b)RF/IF receiver console and data acquisition unit

To measure array response of both receiver and antenna, we measured the magnitude and phase response by changing the transmitter locations. The transmitter is placed at 25 different locations (one after another) as shown in Figure 8 and different code sequences are used at the locations to obtain multiple access signals. The overall angle range is from -60° degree to 60° degree and angle difference each location is 5° degree. The location of receiver is expressed as "circle" in the Figure. To obtain array response of the intermediate angle between 0° and 5°, we use interpolation technique and finally obtain array response table of 0.1° resolution. The measured array response is also shown in Figure 8.



(a)



(b)

Figure 8 Locations of Calibration Sources and Antenna Response

(a) Locations of Calibration Sources.

(b) Measured Array Response.

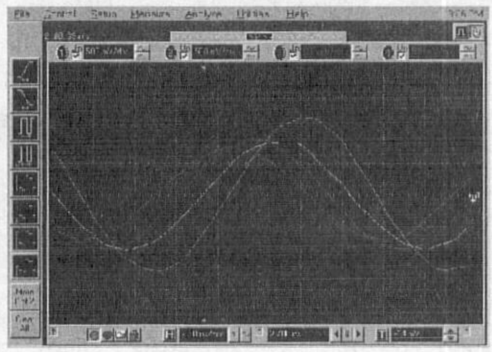
To test the proposed calibration algorithm with data obtained from an uncalibrated antenna, the following steps are performed.

[1] Insert the 70 MHz IF signal into the receiver with the receiver cable, which is originally connected to the antenna array.

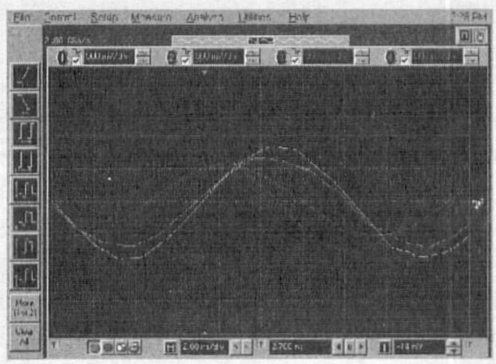
[2] And then change the phase of each channel in the receiver to give uncertainty in the receiver and cable .

[3] Finally the uncalibrated data is obtained by connecting the cable to the array antenna and collecting the data.

The resulting phase difference measured from oscilloscope is shown in Figure 9. The measured angle calibration value is listed in Table 2



(a)



(b)

Figure 9 Captured Calibration Values from receiver

(a) Phase angle of antenna 1, 2, 3 and 4.

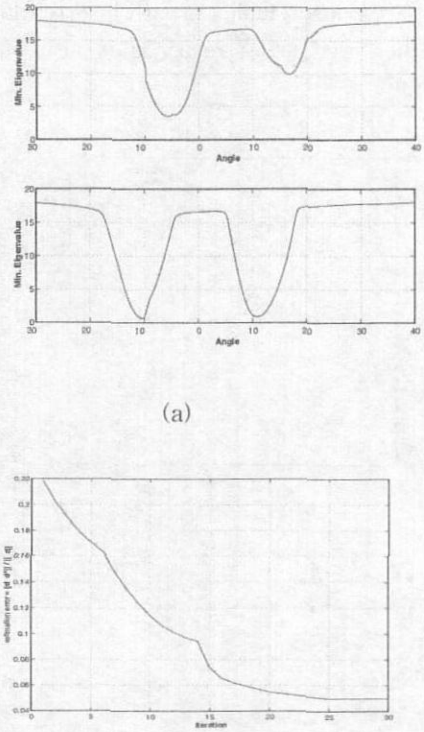
(b) Phase angle of antenna 1, 5, 6 and 7.

Table 2 The measured calibration value

Antenna No.	angle (degree)	Antenna No.	angle (degree)
1	0	5	-5
2	30	6	-70
3	-28	7	-65
4	-55	8	50

In experiment, signals obtained using code sequence number 1 are assumed to be reference signals and DOAs of the signals are measured as -10° and 10°, respectively. To generate interference signal, the signals coming from DOAs of $5^\circ, 12^\circ, 0^\circ, -5^\circ, -15^\circ, -30^\circ$ are used and the power of each signal is set to be the same as the reference signal. To generate frequency offset error, we arbitrary shift the phase of the received data, since the frequency offset is already compensated via oscillator in the receiver. The frequency offset Δf, consequently, is set to be 0.01.

The estimated minimum eigenvalues according to DOA obtained at initial and last iteration is shown in Figure 10, Also, estimation error of the calibration vector is shown in the figure. The estimated DOA and channel impulse response according to iterations is shown in Figure 11. In Figure 12, the roots of estimated polynomial are plotted, in which frequency offset can be estimated clearly.

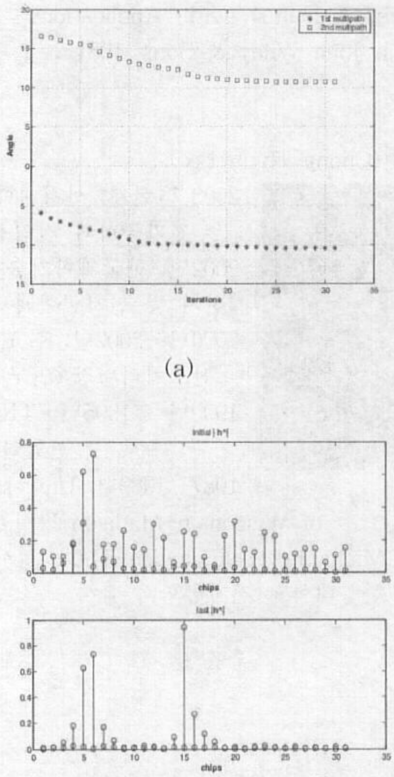


(a)

(b)

Figure 10 Estimation Results vs. iterations.
(a) The Estimated minimum eigenvalues at initial and last iteration.
(b) The calibration estimation error vs. iterations.

In order to observe the effect of on the estimation performance, we examine validity of proposed algorithm by plotting the minimum eigenvalues according to the offset errors. Since the frequency offset is already compensated via oscillator in the receiver, we arbitrary shift the phase of the received data. The minimum eigenvalues obtained from uncalibrated array are shown in Figure 13. In this figure, we can observe that when the offset error is very big, estimation of DOA and impulse response would be very difficult. In these figures, the performance of the proposed algorithm is clearly observed.



(a)

(b)

Figure 11 DOA and Time Delay Estimation Results vs. iterations.

(a) The Estimated DOA vs. iterations.

(b) The Estimated channel impulse response at initial and last iteration.

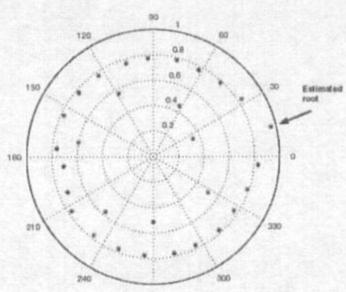


Figure 12 Roots of polynomial.

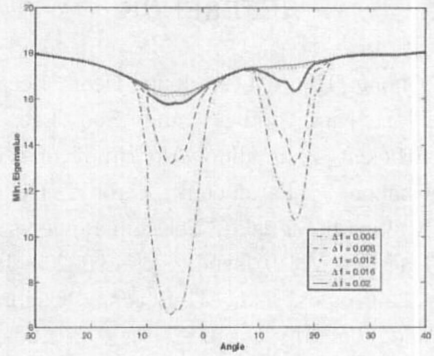


Figure 13 Minimum Eigenvalues with different frequency offset.

V Conclusions

In this paper, a calibration algorithm, which can be used in asynchronous CDMA system with frequency offset error, is presented. The proposed algorithm is not dependent on the array geometry and requires the code sequence of any arbitrarily selected reference user only (which is already available at the base station). By using the code sequence, the algorithm provides us with estimates of the frequency offset, the channel estimates such as DOA and delay of multipath signals. Throughout the computer simulations and experiments on the field measured data, it is shown that if the frequency offset is not properly compensated, the array cannot be calibrated properly. The proposed algorithm, however, is proven to work even when the carrier frequency offset error exist. The sensitivity analysis and Cramer Rao Bound evaluation of the proposed estimator will be future research topics.

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