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Joint DOA Estimation and Phase Calibration for Synchronous CDMA with Decorrelator*

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Abstract: A joint direction of arrival (DOA) estimation and phase calibration for synchronous CDMA system with decorrelator are presented. Through decorrelating processing DOAs of the desired users can be estimated independently and all other resolved signal interferences are eliminated. Emphasis is directed to applications in which sensor phases may be in error. It is shown that accurate phase calibration in conjunction with their use in high-resolution DOA estimation can be achieved for the decoupled signals.

Key words: DOA estimation, phase calibration, CDMA system, antenna array

Code division multiple access (CDMA) is a spread-spectrum multiple access scheme that is expected to gain a significant share of the cellular market. CDMA has several attractive properties for personal communications. Furthermore, the use of antenna array is expected to improve system capacity, quality, and coverage substantially[1].

The physical foundation of array signal processing is the strict correlation of array element output when receiving the input signals. This requires the strict correlation of complex swing of every array channel processing. Otherwise, it will cause the spectrum estimation error or even failure. In practice, the array channel complex swing is often inconsistent. The main reasons for the inconsistency generally are: ① The gain and phase inconsistence of the amplifier and converter of array channel caused mainly by local oscillator; ② The cross-coupling between array elements; ③ The length and location inconsistency of array elements, etc.

The complex swing inconsistency of array channels includes gain and phase inconsistency. Generally, in hardware configuration it is not very difficult to maintain gain consistency while it is much more difficult to maintain phase consistency particularly for wideband system. Furthermore, between the gain and phase inconsistency the gain inconsistence causes much less spectrum estimation error compared with that of the phase inconsistency[2]. So the main purpose of array calibration is to solve the problem of phase calibration.

For antenna array CDMA system the DOA estimation of the objective signals is a key technique both for beamforming and positioning service. A number of DOA estimation algorithms have been proposed during the last decade[3-5]. However, in CDMA system, the users operate in the same frequency and time channel, and multiple access interference (MAI) arises from other users. The conventional DOA estimation algorithms such as MUSIC and ESPRIT algorithms need that the elements of antenna array outnumber the signal source number. However, generally there are several tens of users and several sub-paths per user within a cell for a typical CDMA system. The cross-correlation from the multipath co-channel signals causes that the conventional DOA estimation algorithms such as MUSIC and ESPRIT algorithms cannot be used directly in CDMA system. However, using multiuser detection technology such as decorrelating processing in antenna array CDMA system not only the near-far resistance is achieved, but also the decoupled signal is narrowband and interferences from all other resolved signals are eliminated. So the DOAs of desired signals can be estimated independently. In this paper we will combine the decorrelating detection technology and Wylie's algorithm[6] to provide a method to realize the joint DOA estimation and phase calibration on-line for synchronous CDMA system.

This paper is organized as follows. In section 1, the decorrelating detector signal model for synchronous CDMA system with antenna array is described. In section 2, the joint DOA estimation and phase

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calibration algorithm for the decoupled signal are presented. Simulation results are shown in section 3. Section 4 presents our conclusion.

1 System Model

In this section, we consider the decorrelating detector[7,8] with antenna array. We only concern a synchronous CDMA channel with slow frequency-nonselective fading. In practice, the channels are generally asynchronous and the detection problem in an asynchronous channel is more complicated than in a synchronous channel. Similarly, in practice, the channel is often subject to multipath fading. If the difference between delays of sub-path signals is smaller than the chip period, the channel is subject to frequency-nonselective fading. Generally the difference between delays of sub-path signals is longer than the chip period, the channel is subject to frequency-selective fading. However, once the detection problem in a synchronous channel with frequency-nonselective fading has been formulated, the extension to the detection problem in an asynchronous channel with frequency-selective fading is straightforward. For simplification and plainness we only consider a synchronous channel with frequency-nonselective fading.

Assume that there are N users impinging on antenna array. The base station receivers employ a linear antenna array of M uniformly spaced receiving elements, separated by a distance d, and is chosen to be $\lambda/2$ in our paper, where λ is the wavelength corresponding to the carrier frequency f. We assume that there are phase perturbations among array channels or elements. In this case the array response vector is as below

$$\tilde{\boldsymbol{a}}(\theta) = \boldsymbol{G}\boldsymbol{a} = [1, e^{j(\beta+\varphi_2)}, \cdots, e^{j[(N-1)\beta+\varphi_N]}]^{T}$$

where $\boldsymbol{a}(\theta) = [1, e^{j\beta}, \cdots, e^{j(N-1)\beta}]]^{T}, \beta = j2\pi d\sin\theta/\lambda$ is the array response vector with perfect condition. $\boldsymbol{G} = \mathrm{diag}[1, e^{j\varphi_2}, \cdots, e^{j\varphi_N}]$ is defined as phase perturbation matrix; φ_n is the nth channel phase perturbation with respect to the reference point of the first sensor location while the gain is regarded as a constant in the paper.

The received signals from the base station antenna array can be written as

$$\boldsymbol{x}(t) = \sum_{i=1}^{N} \sqrt{P_i} b_i(t) c_i(t) e^{j\psi_i} \tilde{\boldsymbol{a}}_i + \boldsymbol{n}(t) \qquad (1)$$

where $P_i, \psi_i, b_i(t), c_i(t), Bw, \tilde{\boldsymbol{a}}_i$ are the signal power, carrier phase, information sequence, spreading sequence, channel bandwidth, and array response vector of the ith user's, respectively; $n(t)$ denotes the additive noise vector of the antenna array. Here we assume $n(t)$ denotes the zero mean spatially and temporally white Gaussian noise vector, i.e.

$$E[\boldsymbol{n}(t)\boldsymbol{n}^{H}(t)] = \sigma^2 \boldsymbol{I}\delta(t-\tau)$$

where $\delta(t)$ denotes the Dirac delta function. For synchronous system only a symbol period T_s needs to be considered. From (1) the received signals from the antenna array can be rewritten as sample matrix form

$$\boldsymbol{x}(i) = \tilde{\boldsymbol{A}}\boldsymbol{B}\boldsymbol{P}\boldsymbol{\Phi}\boldsymbol{C}(i) + n(i) \qquad 0 \leqslant t \leqslant T_s \qquad (2)$$

where

$$\boldsymbol{A} = [\tilde{\boldsymbol{a}}_1, \tilde{\boldsymbol{a}}_2, \cdots, \tilde{\boldsymbol{a}}_N]$$
$$\boldsymbol{B} = \mathrm{diag}[b_1, b_2, \cdots, b_N]$$
$$\boldsymbol{P} = \mathrm{diag}[\sqrt{P_1}, \sqrt{P_2}, \cdots, \sqrt{P_N}]$$
$$\boldsymbol{\Phi} = \mathrm{diag}[e^{j\psi_1}, e^{j\psi_2}, \cdots, e^{j\psi_N}]$$
$$\boldsymbol{C}(t) = [c_1(i), c_2(i), \cdots, c_N(i)]^{T}$$

To extract every resolvable desired user's signals, in conventional detector $N \times L$ correlators need to be used. The performance of the conventional detector depends on the correlation matrix that is defined by

$$\boldsymbol{R}_c = \int_0^{T_s} \boldsymbol{C}(t)\boldsymbol{C}^{T}(t)\,dt$$

From the $N \times L$ correlators with each spreading sequence, the outputs sampled at the bit epochs are given by the matrix

$$\boldsymbol{Y} = \int_0^{T_s} \boldsymbol{x}(t)\boldsymbol{C}^{H}(t)\,dt = \boldsymbol{A}\boldsymbol{B}\boldsymbol{P}\boldsymbol{\Phi}\boldsymbol{R}_c + \boldsymbol{N} \qquad (3)$$

where $\boldsymbol{N} = \int_0^{T_s} \boldsymbol{n}(t)\boldsymbol{C}^{H}(t)\,dt$.

To eliminate the signal interferences and decouple the information of the user's data, the decorrelating detector applies the inverse of correlation matrix to the conventional detector output. The decoupled output matrix from the decorrelator is given by

$$\boldsymbol{Z} = \boldsymbol{Y}\boldsymbol{R}_c^{-1} = \tilde{\boldsymbol{A}}\boldsymbol{B}\boldsymbol{P}\boldsymbol{\Phi} + \boldsymbol{N}_z = [\boldsymbol{z}_1, \boldsymbol{z}_2, \cdots, \boldsymbol{z}_N] \qquad (4)$$

where $\boldsymbol{N}_z = [\boldsymbol{n}_{z_1}, \boldsymbol{n}_{z_2}, \cdots, \boldsymbol{n}_{z_N}]$.

Obviously, each column vector of the matrix $\boldsymbol{Z}$ contains a single resolved desired signal and has not interferences from all the other resolvable signals. So every resolvable sub-path signal is detected independently.

Therefore the decoupled vector of the ith user's can be written as

$$\boldsymbol{z}_i = \tilde{\boldsymbol{a}}_i \sqrt{P_i} e^{j\varphi_i} b_i + \boldsymbol{n}_{z_i} \qquad 0 \leqslant l \leqslant L \qquad (5)$$

where b_i is the symbol of the ith user's signal and $\boldsymbol{n}_{z_i}$ is the corresponding noise vector.

2 Estimation Algorithm

As discussed above for synchronous CDMA system with decorrelator, the decoupled vector of the ith user's can be written as

$$\boldsymbol{z}_i = \tilde{\boldsymbol{a}}_i \sqrt{P_i} e^{j\phi_i} b_i + \boldsymbol{n}_{z_i}$$

For convenience, we drop the subscript of the above equation. Then the decoupled signal model becomes

$$\boldsymbol{z}(k) = \tilde{\boldsymbol{a}}\boldsymbol{s}(k) + \boldsymbol{n}(k) \tag{6}$$

Thus the covariance of the decoupled measurement vector is given by

$$\boldsymbol{R} = \sigma_s^2 \tilde{\boldsymbol{a}}\tilde{\boldsymbol{a}}^H + \sigma_n^2 \boldsymbol{I} \tag{7}$$

where $\boldsymbol{s}(k)$ is the nth snapshot vector of the decoupled signal with its variance $E|s(k)|^2 = \sigma_s^2$ and σ_n^2 is the noise variance with decorrelating transformation.

In our case, after decorrelating processing, from finite observations or snapshots of the decoupled vector of the desired user's, i.e. $\boldsymbol{z}(k)$, we can estimate its covariance matrix $\hat{\boldsymbol{R}}_z$ defined by

$$\hat{\boldsymbol{R}}_z = \frac{1}{K}\sum_{k=1}^{K} \boldsymbol{z}(k)\boldsymbol{z}^H(k) \tag{8}$$

It is well known that for the uniformly spaced line geometry array if the sensors had uniform gain and phase, the true array covariance $\boldsymbol{R}$ will be Toeplitz. However, in the presence of gain or phase distortions (i.e. $\boldsymbol{a} \neq \tilde{\boldsymbol{a}}$), obviously, the covariance matrix $\boldsymbol{R}$ is non-Toeplitz. This class estimation algorithm developed generally seeks to exploit the fundamental Toeplitz property of sample covariance matrix of an unperturbed array in order to achieve self-calibration. Thus the uniformly spaced line geometry assumption is essential for this class estimation algorithm. In this paper, we also seek the same method to achieve the DOA estimation and phase calibration for synchronous CDMA with decorrelator.

The mth and nth element of the covariance matrix of the decoupled signal, $\boldsymbol{R}_z$, denoting by $\boldsymbol{R}_{mm}$, can be described as Ref.[6]

$$R_{m,n} = \sigma_s^2 e^{j[\pi(m-n)\sin\theta + (\varphi_m - \varphi_n)]} + \sigma_n^2 \delta_{mm}$$
$$m, n = 1, \cdots, N \tag{9}$$

According to central limit theorem, we have $\hat{\boldsymbol{R}}_{m,n}$ and its mean and correlation as

$$\hat{R}_{m,n} = \frac{1}{N}\sum_{i=1}^{N} z_m(i) z_n^*(i) \xrightarrow{D} R_{m,n} \qquad n > m \tag{10}$$

$$E\{\hat{R}_{m,n}\} = \sigma_s^2 e^{j[\pi(m-n)\sin\theta + (\varphi_m - \varphi_n)]} + \sigma_n^2 \delta_{mm} \tag{11}$$

$$\begin{aligned} E\{\hat{R}_{m,n}\hat{R}_{k,l}^*\} = {} & \frac{1}{N^2}\sum_{i,p=1}^{N} E\{z_m(i) z_n^*(i) z_k^*(p) z_l(p)\} = \\ & \frac{1}{N^2}\sum_{i=1}^{N} E\{z_m(i) z_n^*(i) z_k^*(p) z_l(p)\} + \\ & \sum_{i=1}^{N}\sum_{\substack{p=1 \\ p\neq i}}^{N} E\{z_m(i) z_n^*(i) z_k^*(p) z_l(p)\} \\ & n > m, l > k \end{aligned} \tag{12}$$

Because the angle and phase information all are included in the superdiagonal elements of the covariance matrix, we assume $m > n, l > k$.

Define $\beta_{\varphi_m} = \pi m \sin\theta + \varphi_m$ and some simplification yield[6]

$$\begin{aligned} \mathrm{cov}\{\hat{R}_{m,n}, \hat{R}_{k,l}^*\} = {} & \frac{1}{M}(\sigma_s^2 e^{j(\beta_{\varphi_m} - \beta_{\varphi_k})} + \sigma_n^2 \delta_{mk}) \times \\ & (\sigma_s^2 e^{j(\beta_{\varphi_l} - \beta_{\varphi_n})} + \sigma_n^2 \delta_{ln}) \end{aligned} \tag{13}$$

Thus, $\hat{R}_{m,n}$ can be modeled as

$$\hat{R}_{m,n} = \sigma_s^2 e^{j(\beta_{\varphi_m} - \beta_{\varphi_n})} + w_{m,n} \qquad n > m \tag{14}$$

where $w_{m,n}$ is a complex, colored, zero-mean (approximately) Gaussian noise with the covariance described in (13). Let

$$v_{m,n} = \frac{e^{-j(\beta_{\varphi_m} - \beta_{\varphi_n})}}{\sigma_s^2} w_{m,n} \tag{15}$$

Thus, $v_{m,n}$ is also a complex, colored, zero-mean (approximately) Gaussian noise like $w_{m,n}$ and its covariance can be given

$$\mathrm{cov}\{v_{m,n}, v_{k,l}\} = \frac{e^{-j(\beta_{\varphi_m} - \beta_{\varphi_n})} e^{-j(\beta_{\varphi_k} - \beta_{\varphi_l})}}{\sigma_s^4} \times$$
$$\mathrm{cov}\{w_{m,n}, w_{k,l}\} = \frac{1}{M}\left(1 + \frac{\sigma_n^2}{\sigma_s^2}\delta_{mk}\right)\left(1 + \frac{\sigma_n^2}{\sigma_s^2}\delta_{ln}\right) \tag{16}$$

Obviously, for sufficiently high SNR i.e. σ_s^2/σ_n^2, or/and large number of snapshots M, the $\mathrm{Re}\{1 + v_{m,n}\}$ is approximately 1 and the $\mathrm{Im}\{1 + v_{m,n}\}$ is the $\mathrm{Im}\{v_{m,n}\}$ according to (15).

Denote $\varepsilon_{m,n}$ as the unwrapped phase of $1 + v_{m,n}$, then $\varepsilon_m \approx \mathrm{Im}\{v_{m,n}\}$. Thus $\varepsilon_{m,n}$ is also a complex, colored, zero-mean (approximately) Gaussian noise like $v_{m,n}$ and its covariance can be given as

$$\mathrm{cov}\{\varepsilon_{m,n}, \varepsilon_{k,l}\} = \frac{1}{2M}\left(1 + \frac{\sigma_n^2}{\sigma_s^2}\delta_{mk}\right)\left(1 + \frac{\sigma_n^2}{\sigma_s^2}\delta_{ln}\right) \tag{17}$$

From (15), Eq.(14) can be written as

$$\hat{R}_{m,n} = \sigma_s^2 e^{j(\beta_{\varphi_m} - \beta_{\varphi_n})}[1 + v_{m,n}] \qquad n > m \tag{18}$$

Thus, the unwrapped phase of $\hat{R}_{m,n}$ can be model as

$$\angle \hat{R}_{m,n} = \pi(m-n)\sin\vartheta(\varphi_m - \varphi_n) + \varepsilon_{m,n} \tag{19}$$

Because of the Hermitian symmetry, we construct a phase vector $\boldsymbol{\gamma} = [\gamma_1 \cdots \gamma_{N(N-1)/2}]$ of the superdiagonal elements of $\hat{\boldsymbol{R}}_z$, where

$$\angle \hat{R}_{m,n} = \gamma_{(r-m)+(m+1)(N-m/2)}$$
$$m = 1, \cdots, N-1;\ r = m+1, \cdots, N \tag{20}$$

From (17) and (20), $\boldsymbol{\gamma}$ has the approximate distribution

$$\gamma_{(r-m)+(m+1)(N-m/2)} \sim N\left\{\angle \hat{R}_{m,n}, \frac{1}{2M}\left(1 + \frac{\sigma_n^2}{\sigma_s^2}\right)^2\right\} \tag{21}$$

Now let $\boldsymbol{\Theta} = [\varphi_2 \cdots \varphi_N, \sin\theta]$ is the unknown to be estimated $M \times 1$ vector, (20) can be written as matrix form

$$\boldsymbol{\gamma} = \boldsymbol{F\Theta} + \boldsymbol{\eta} \tag{22}$$

where $\boldsymbol{\eta}$ is an additive noise vector and its elements

$$\eta_{(r-m)+(m+1)(N-m/2)} = \varepsilon_{m,r}$$
$$m = 1,\cdots,N-1;\ r = m+1,\cdots,N \tag{23}$$

$\boldsymbol{F}$ is a $\frac{1}{2}N(N-1)\times N$ vector with its elements

$$F_{(r-m)+(m+1)(N-m/2),p} = \begin{cases} +1 & \text{if } p = m-1 \\ -1 & \text{if } p = r-1 \\ \pi(m-r) & \text{if } p = N \\ 0 & \text{otherwise} \end{cases}$$

From (22) the vector $\boldsymbol{\gamma}$ has the asymptotic distribution $N(F\Theta,\Sigma)$, where the elements of the covariance matrix Σ are given by

$$\Sigma_{(r-m)+(m-1)(N-m/2),(t-2)+(n-1)(N-n/2)} = \operatorname{cov}\{\varepsilon_{m,n},\varepsilon_{k,l}\} = \frac{1}{M}\left(1+\frac{\sigma_n^2}{\sigma_s^2}\delta_{mt}\right)\left(1+\frac{\sigma_n^2}{\sigma_s^2}\delta_m\right)$$

$m, n = 1,\cdots,N;\ r = n+1,\cdots,N; r = m+1,\cdots,N$

Based on (22), generally, we can estimate the unknown vector $\boldsymbol{\Theta}$ using standard LS approach. That is

$$\underbrace{\min}_{\theta}(\gamma - \boldsymbol{F\Theta})^{\mathrm{T}}(\gamma - \boldsymbol{F\Theta}) \tag{24}$$

However, because of rank$[F'F]$ = rank$[F] = M - 1$[6], (24) has a nonunique solution. Another appropriate linear constraint condition is necessary.

Usually, the phase aberration φ_n is modeled as zero-mean i.i.d random variable that remain constant over the interval of observation. Furthermore, generally, there exists $\frac{1}{N-1}\sum_{n=2}^{N}\varphi_n \approx 0$ for any reasonable large N. Obviously, the formula below will lead to a unique solution.

$$\left.\begin{aligned} &\underbrace{\min}_{\theta}(\gamma - \boldsymbol{F\Theta})^{\mathrm{T}}(\gamma - \boldsymbol{F\Theta}) \\ &\frac{1}{N-1}\sum_{n=2}^{N}\varphi_n = 0 \end{aligned}\right\} \tag{25}$$

Now we have the LS solution without constrain conditions.

$$\underbrace{\min}_{\theta}[(\gamma - \boldsymbol{F\Theta})^{\mathrm{T}}(\gamma - \boldsymbol{F\Theta}) + (\boldsymbol{H}^{\mathrm{T}}\boldsymbol{\Theta})^{\mathrm{T}}(\boldsymbol{H}^{\mathrm{T}}\boldsymbol{\Theta})] \tag{26}$$

where $\boldsymbol{H} = \frac{1}{N}[1,\cdots,1,0]$.

The LS solution of (26) is

$$\hat{\Theta} = (\boldsymbol{F}^{\mathrm{T}}\boldsymbol{F} + \boldsymbol{HH}^{T})^{-1}\boldsymbol{F}^{\mathrm{T}}\gamma \tag{27}$$

Its mean and covariance is

$$\left.\begin{aligned} &E\{\hat{\boldsymbol{\Theta}}\} = \boldsymbol{\Theta} - (\boldsymbol{F}^{\mathrm{T}}\boldsymbol{F} + \boldsymbol{HH}^{\mathrm{T}})^{-1}\boldsymbol{HH}^{\mathrm{T}}\boldsymbol{\Theta} \\ &\operatorname{cov}\{\hat{\boldsymbol{\Theta}}\} = (\boldsymbol{F}^{\mathrm{T}}\boldsymbol{F} + \boldsymbol{HH}^{\mathrm{T}})^{-1}\boldsymbol{F}^{\mathrm{T}}\boldsymbol{\Sigma F\Theta}(\boldsymbol{F}^{\mathrm{T}}\boldsymbol{F} + \boldsymbol{HH}^{\mathrm{T}})^{-1} \end{aligned}\right\} \tag{28}$$

If $\boldsymbol{H}^{\mathrm{T}}\boldsymbol{\Theta} \neq 0$, the estimation is biased and is given by Ref.[6]

$$\left.\begin{aligned} &E\{\hat{\varphi} - \varphi\} = \frac{2S_\varphi}{N(N-1)}[1,\cdots,N-1]^{\mathrm{T}} \\ &E\{\sin\hat{\theta} - \sin\theta\} = \frac{2S_\varphi}{\pi N(N-1)} \end{aligned}\right\} \tag{29}$$

where $S_\varphi = \sum_{n=2}^{N}\varphi_n$ is the true sum of the phases. According to (29), for any reasonably large N the estimation error is very small.

3 Simulation Results

To illustrate the performance of the joint DOA estimation and phase calibration for synchronous CDMA with decorrelator we present simulation results. In the simulation, we employ a 4-element antenna array structured as section 1 for reception of BPSK signals. The carrier phase of signal is assumed to be 0 and the spreading sequence is 31 gold codes. Suppose that there are 10 users within a sector. Assume that the phase perturbations were generated according to $\beta_i = \sigma_\beta x_i$, where $\sigma_\beta = 0.1\pi$, and $x_i \sim N(0,1)$. The values of $\{\beta_i\}$ were kept constant during a simulation run conforming to the constrain condition assumed in section 2.

In our case the snapshot data was generated according to the decoupled observation model (6), i.e.

$$\boldsymbol{z}(k) = \tilde{\boldsymbol{a}}\boldsymbol{s}(k) + \boldsymbol{n}(k)$$

and the covariance matrix of the observations was estimated by a sample average based on 100 snapshots

$$\hat{\boldsymbol{R}}_z = \frac{1}{100}\sum_{k=1}^{100}\boldsymbol{z}(k)\boldsymbol{z}^H(k)$$

In the simulation the user's signals was assumed to impinge on the array from the broadside, and the simulations were repeated for SNR's of 1, 5, 10, 50, 100 dB while the signal-to-noise ratio (SNR, in decibels) is defined as

$$\mathrm{SNR} \overset{\text{def}}{=} 10\log_{10}(s/\sigma^2)$$

To evaluate the joint DOA estimation and phase calibration for the decoupled signal 100 independent trials for different SNR were averaged. The variance of phase and DOA ($\sin\theta$) estimation is calculated and displayed in Fig.1 and Fig.2 respectively. Tab.1 gives the average estimate statistics for five different users selected randomly among the 10 users with SNR 1 dB and 10 dB respectively. Fig.1 and Fig.2 indicate the estimation error decrease with the increasing SNR. Furthermore, from Fig.1, Fig.2, and Tab.1, it is clear that the estimation accuracy is sufficiently high in the viewpoint of practical application even though the number of array element is not large. And because of the decorrelating processing, for signals that may be superposition spatially, their DOAs can be estimated effectively.

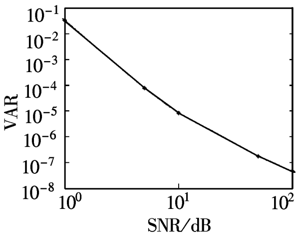

Fig.1 Variance for phase estimation

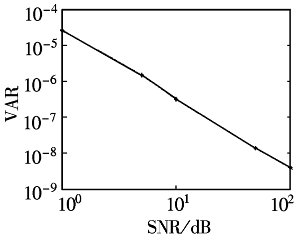

Fig.2 Variance for DOA ($\sin\theta$) estimation

Tab.1 Average estimate statistics

User	SNR(1 dB)		SNR(10 dB)	
User 1	(−17°)	−16.9104	(−17°)	−16.9967
User 2	(−7°)	−7.1286	(−7°)	−7.0062
User 3	(7°)	7.1310	(7°)	6.9971
User 4	(7°)	7.2261	(7°)	7.0031
User 5	(17°)	17.0260	(17°)	17.0012

4 Conclusion

In this paper, a joint DOA estimation and phase calibration for synchronous CDMA with decorrelator are presented. Making full use of the prior knowledge and signature of the spreading sequence of CDMA system that the DOAs of the desired user's can be estimated independently and all the other resolved multipath signal interference is eliminated. Using the sample covariance matrix of the decoupled signals the DOA estimation for the desired user in the presence of unknown phase errors as well as self-calibration of phased array can be achieved. And the phasecalibration is achieved on-line during operation that does not require any a priori knowledge of DOA of a signal. However, reasonable linear constraint condition is necessary for the self-calibration. Simulation results indicate that the algorithm presented has simple, robust quality for the joint DOA estimation and phase calibration of synchronous CDMA systems and may be suitable for the practical application.

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解相关同步 CDMA 系统一种联合 DOA 估计与相位标定算法

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摘 要 提出了解相关同步 CDMA 系统一种联合波达方向(DOA)估计与相位标定算法.通过解相关处理可消除分离出的所有干扰信号,使目标用户的 DOA 可独立地进行估计.文中着重研究了阵列传感器相位存在误差时的估计情况.仿真结果表明可实现精确的相位标定及解藕信号的高分辨 DOA 估计.

关键词 DOA 估计,相位标定,CDMA 系统,阵列天线

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