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Constrained least squares algorithm for channel vector estimation in 2-D RAKE receiver

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Abstract: Based on the fact that the variation of the direction of arrival (DOA) is slower than that of the channel fading, the steering vector of the desired signal is estimated firstly using a subspace decomposition method and then a constrained condition is configured. Traffic signals are further employed to estimate the channel vector based on the constrained least squares criterion. We use the iterative least squares with projection (ILSP) algorithm initialized by the pilot to get the estimation. The accuracy of channel estimation and symbol detection can be progressively increased through the iteration procedure of the ILSP algorithm. Simulation results demonstrate that the proposed algorithm improves the system performance effectively compared with the conventional 2-D RAKE receiver.

Key words: 2-D RAKE receiver; channel estimation; subspace decomposition; constrained least squares

It is well known that the use of antenna arrays can considerably increase the linking capacity of mobile radio systems. Various algorithms of space-time signal processing can be used to enhance the system's performance. In CDMA system, the 2-D RAKE receiver is a typical space-time processing method. The essence of the 2-D RAKE receiver is to calculate a vector to weigh the observed signals to optimize the system performance. The calculation of this vector is usually determined by the channel vector[1]. Thus, an accurate estimation of the channel vector is of great importance in the 2-D RAKE receiver.

In the conventional 2-D RAKE receiver, the channel vector estimation is ordinarily based on a least squares (LS) criterion when a continuous pilot is available[2]. The estimation accuracy is determined by two factors: the time window and the pilot power. The wider the time window or the stronger the pilot power, the more accurate is the estimation. However, the time-varying channel restricts the window length and the strong pilot power incurs high capacity loss. Therefore the channel vector estimation accuracy is limited by these two bottlenecks.

To combat this problem, we propose an improved channel vector estimation algorithm. In a typical mobile communications scenario, the variation of the direction of arrival (DOA) is slower than that of the fading[3]. We firstly exploit this character to estimate the steering vector in a relatively wider time window and then use the estimated result to set a constraint condition. Traffic signals are further employed to estimate the channel vector based on a constrained LS criterion. The solution can be obtained using the iterative LS with projection (ILSP) algorithm[4]. In this improved algorithm, the pilot is only used to provide the initial estimate for the ILSP algorithm. High estimation accuracy can be achieved through iteration even if the pilot power is very weak. Therefore two bottlenecks that the conventional LS algorithm suffers are mitigated.

1 System Model

The up-link of a mobile communication system is considered in this paper. The transmitter structure of the mobile station is shown in Fig.1. The base station is equipped with M-element d-spaced uniform linear array. Assume that there are K users in the section and the k-th user has L_k paths. Also assume that the plane wave from the l-th path of the k-th user impinges upon the array at an angle θ_{kl} and its corresponding delay is $\tau_{kl}(l=1,2,\cdots,L_k)$. The observed base-band signals can be written as

$$\boldsymbol{x}(t)=\sum_{k=1}^{K}\Big[A_k^{\mathrm{t}}\sum_{l=1}^{L_k}\rho_{kl}\boldsymbol{a}(\theta_{kl})\sum_{n=-\infty}^{\infty}b_k^{\mathrm{t}}(n)c_k^{\mathrm{t}}(t-nT_{\mathrm{b}}-\tau_{kl})+A_k^{\mathrm{p}}\sum_{l=1}^{L_k}\rho_{kl}\boldsymbol{a}(\theta_{kl})\sum_{n=-\infty}^{\infty}b_k^{\mathrm{p}}(n)c_k^{\mathrm{p}}(t-nT_{\mathrm{b}}-\tau_{kl})\Big]+\boldsymbol{n}(t) \quad (1)$$

where $b_k^{\mathrm{t}}(n)$ and $b_k^{\mathrm{p}}(n)$ are the n-th symbol of the k-th user for traffic channel and pilot channel, respectively. In this paper, we assume the symbol is BPSK modulated. $c_k^{\mathrm{t}}(t)$ and $c_k^{\mathrm{p}}(t)$ are the spreading

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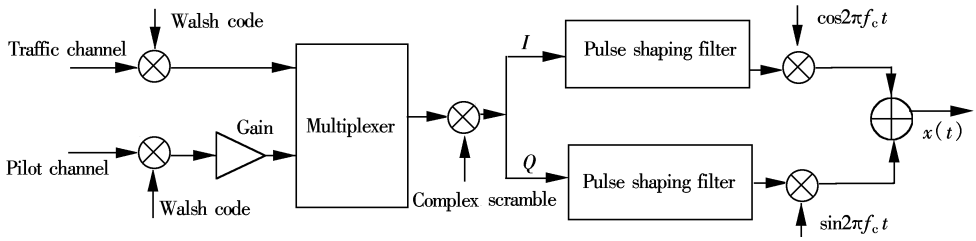

Fig. 1 Transmitter structure of mobile station in up-link

waveforms of the k-th user for the traffic channel and pilot channel, respectively. A_k^{t} is the amplitude of the traffic signal and A_k^{p} is the pilot amplitude. Without loss of generality, we assume $A_k^{t} = 1$ and $A_k^{p} = \beta$. Then β^2 is the pilot-to-traffic power ratio. T_b is the symbol interval and $\boldsymbol{n}(t)$ is the $M \times 1$ additive Gaussian noise vector. ρ_{kl} is the channel complex fading of the l-th path for the k-th user and $\boldsymbol{a}(\theta_{kl})$ is its corresponding steering vector. We define the channel vector as

$$\boldsymbol{h}_{kl} = \rho_{kl}\boldsymbol{a}(\theta_{kl}) \tag{2}$$

An algorithm of the 2-D RAKE receiver is presented in Ref.[2]. Its basic procedure is described in Fig.2.

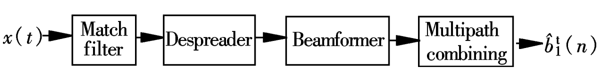

Fig. 2 Principle diagram of the 2-D RAKE receiver

Assuming the first user is the desired user, the dispread signals for the m-th path of its pilot channel and traffic channel can be written as[2]

$$\boldsymbol{r}_{1m}^{p}(n) = \beta\boldsymbol{h}_{1m}b_1^{p}(n) + \boldsymbol{i}_{1m}^{p}(n) \tag{3}$$

$$\boldsymbol{r}_{1m}^{t}(n) = \boldsymbol{h}_{1m}b_1^{t}(n) + \boldsymbol{i}_{1m}^{t}(n) \tag{4}$$

If the maximal ratio combining (MRC) algorithm[5] is used, the output of the 2-D RAKE receiver is

$$\hat{b}_1^{t}(n) = \operatorname{sgn}\left[\operatorname{Re}\left(\sum_{m=1}^{L_1}\hat{\boldsymbol{h}}_{1m}^{H}\boldsymbol{r}_{1m}^{t}(n)\right)\right] \tag{5}$$

where $\hat{\boldsymbol{h}}_{1m}$ is the estimate of $\boldsymbol{h}_{1m}$ and $(\cdot)^{H}$ denotes the conjugate transpose. Eq.(5) shows that the estimation of the channel vector is important to the 2-D RAKE receiver. In a conventional 2-D RAKE receiver, the pilot is used to estimate the channel vector based on an LS criterion, which can be described as

$$\hat{\boldsymbol{h}}_{1m} = \arg\min_{\hat{h}_{1m}}\sum_{n=1}^{J}\left\|\boldsymbol{r}_{1m}^{p}(n) - \beta\hat{\boldsymbol{h}}_{1m}b_1^{p}(n)\right\|^2 \tag{6}$$

The solution is

$$\hat{\boldsymbol{h}}_{1m} = \frac{1}{J\beta}\sum_{n=1}^{J}\boldsymbol{r}_{1m}^{p}(n)[b_1^{p}(n)]^{*} \tag{7}$$

It is well known that the estimation accuracy of $\boldsymbol{h}_{1m}$ is limited by J and β. The selection of J is limited by the condition that the channel vector must be almost invariable during the J-symbol interval. According to Eq.(2), the channel vector is determined by the steering vector and the fading. We know that the variation of fading is faster than that of the steering vector in the practical mobile communications scenario. As a result, J is mainly determined by the speed of fading variation. This limits the accuracy of channel vector estimation.

2 Improved Algorithm for Channel Vector Estimation

Assuming that θ_{1m} is almost invariable during the observation period, we can derive the following equation from Eqs.(2) to (4):

$$\boldsymbol{a}_{1m}(\theta_{1m})\boldsymbol{a}_{1m}^{H}(\theta_{1m}) = \frac{\boldsymbol{R}_{1m}^{t} - \boldsymbol{R}_{1m}^{p}}{(1-\beta^2)E(|\rho_{1m}|^2)} \tag{8}$$

where $\boldsymbol{R}_{1m}^{t}$ and $\boldsymbol{R}_{1m}^{p}$ are the covariance matrices of $\boldsymbol{r}_{1m}^{t}$ and $\boldsymbol{r}_{1m}^{p}$, respectively. They can be estimated by time averaging as

$$\hat{\boldsymbol{R}}_{1m}^{t} = \frac{1}{N}\sum_{n=1}^{N}[\boldsymbol{r}_{1m}^{t}(n)(\boldsymbol{r}_{1m}^{t}(n))^{H}] \tag{9}$$

$$\hat{\boldsymbol{R}}_{1m}^{p} = \frac{1}{N}\sum_{n=1}^{N}[\boldsymbol{r}_{1m}^{p}(n)(\boldsymbol{r}_{1m}^{p}(n))^{H}] \tag{10}$$

where N is determined by the speed of the DOA variation. Because the variation of the DOA is slower than that of the fading, N can be selected to be bigger than J mentioned in Eq.(6). Assuming $\boldsymbol{e}_{1m}$ is the eigenvector corresponding to the largest eigenvalue of $\boldsymbol{R}_{1m}^{t} - \boldsymbol{R}_{1m}^{p}$, and according to Eq.(8), we know that $\boldsymbol{a}_{1m} \in \operatorname{span}(\boldsymbol{e}_{1m})$. Eq.(2) shows that $\boldsymbol{h}_{1m}$ is in the same space as $\boldsymbol{a}_{1m}$. Thus $\boldsymbol{h}_{1m} \in \operatorname{span}(\boldsymbol{e}_{1m})$. We now get the following equation[6]:

$$P_{\boldsymbol{e}_{1m}}^{\perp}\boldsymbol{h}_{1m} = 0 \tag{11}$$

where $P_{\boldsymbol{e}_{1m}}^{\perp}$ is the projector on the orthogonal complement of the column space of $\boldsymbol{e}_{1m}$. Because $\boldsymbol{R}_{1m}^{t}$ and $\boldsymbol{R}_{1m}^{p}$ are estimated in a wider time window, more accurate steering vector estimation can be achieved. Furthermore, to overcome the limit of the pilot power, we take Eq.(11) as a constraint condition and employ the traffic signals to estimate the channel vector based on constrained LS criteria. In this circumstance, the pilot is only used to get the initial estimate.

$$\hat{\boldsymbol{h}}_{1m} = \arg\min_{\hat{\boldsymbol{h}}_{1m}} \sum_{n=1}^{J} \| \boldsymbol{r}_{1m}^{\mathrm{t}}(n) - \hat{\boldsymbol{h}}_{1m} \hat{b}_1^{\mathrm{t}}(n) \|^2 \quad \text{s.t. } P_{\boldsymbol{e}_{1m}}^{\perp} \hat{\boldsymbol{h}}_{1m} = 0 \tag{12}$$

The initial condition is

$$\hat{\boldsymbol{h}}_{1m} = \arg\min_{\hat{\boldsymbol{h}}_{1m}} \sum_{n=1}^{J} \| \boldsymbol{r}_{1m}^{\mathrm{p}}(n) - \beta \hat{\boldsymbol{h}}_{1m} b_1^{\mathrm{p}}(n) \|^2 \quad \text{s.t. } P_{\boldsymbol{e}_{1m}}^{\perp} \hat{\boldsymbol{h}}_{1m} = 0 \tag{13}$$

where $\hat{b}_1^{\mathrm{t}}(n)$ is the estimate of $b_1^{\mathrm{t}}(n)$. The ILSP algorithm is employed to solve this problem. Its detailed procedure is illustrated as follows:

Step 1 Let $k = 0$, for $m = 1, \cdots, L_1$,

$$^{(0)}\hat{\boldsymbol{h}}_{1m} = \frac{\boldsymbol{e}_{1m} \boldsymbol{e}_{1m}^{\mathrm{H}} \sum_{n=1}^{J} \boldsymbol{r}_{1m}^{\mathrm{p}}(n) b_1^{\mathrm{p}}(n)}{J\beta} \tag{14}$$

Step 2 Set $k = k + 1$; for $n = 1, \cdots, J$; $m = 1, \cdots, L_1$,

$$^{(k)}\hat{b}_1^{\mathrm{t}}(n) = \mathrm{sgn}\left[\mathrm{Re}\left(\sum_{m=1}^{L_1} {}^{(k-1)}\hat{\boldsymbol{h}}_{1m}^{\mathrm{H}} \boldsymbol{r}_{1m}^{\mathrm{t}}(n) \right) \right] \tag{15}$$

$$^{(k)}\hat{\boldsymbol{h}}_{1m} = \frac{\boldsymbol{e}_{1m} \boldsymbol{e}_{1m}^{\mathrm{H}} \sum_{n=1}^{J} \boldsymbol{r}_{1m}^{\mathrm{t}}(n) {}^{(k)}\hat{b}_1^{\mathrm{t}}(n)}{J} \tag{16}$$

Step 3 Repeat step 2 until $k > 1$ and $^{(k)}b_1^{\mathrm{t}}(n) = {}^{(k-1)}b_1^{\mathrm{t}}(n)$ $(n = 1, 2, \cdots, J)$.

Comparing Eq.(14) and Eq.(7), we find that the estimated result by the pilot must be projected onto $\boldsymbol{e}_{1m}$. As mentioned above, $\boldsymbol{e}_{1m}$ is estimated in a wider time window. This algorithm also shows that the pilot is only used to get the initial estimate of the channel vector. The estimated traffic signals are feedback to re-estimate the channel. The estimation accuracy can be gradually improved through further iterations. Thus, this algorithm can mitigate the two bottlenecks mentioned in the introduction.

3 Simulation

The system parameters are: the carrier frequency is 2 GHz, the chip rate is 3.84 Mchip/s and the spread factor is 64. The 6-element half-wavelength spaced uniform linear array is equipped in the base station. The BPSK signals of the desired user reach the array through the Rayleigh fading channel. Assume that there are 4 paths impinging the array with injection angle {40°, 10°, 60°, 80°}, average power {1, 0.72, 0.25, 0.03} and delay chips {0, 2, 3, 5}, respectively. The vehicular velocity is 100 km/h and the maximal Doppler shift is 185 Hz. During our simulations, the DOA varies 0.001° per symbol. The time window length is 32-symbol interval in the fading estimation and 1 000-symbol interval in the steering vector estimation respectively.

The normalized mean square error (NMSE) of the channel vector estimation and symbol error rate (SER) using different algorithms are given in Fig.3 and Fig.4, respectively, where the pilot-to-traffic power ratio (PTR) is 0.25. According to the figures, the improved algorithm outperforms the conventional LS algorithm under various signal-to-interference-plus-noise ratio (SINR) conditions. The reason is that the improved algorithm employs both the pilot and traffic information, and exploits the character that DOA varies more slowly than the fading at the same time.

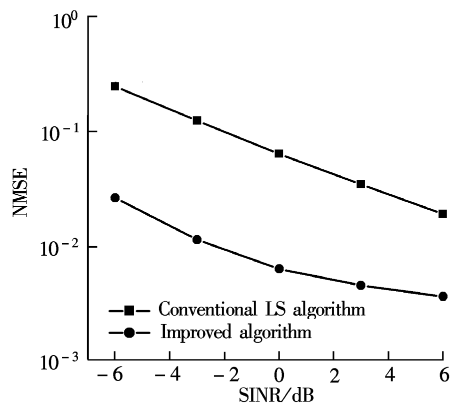

Fig.3 NMSE for different SINR with PTR = 0.25

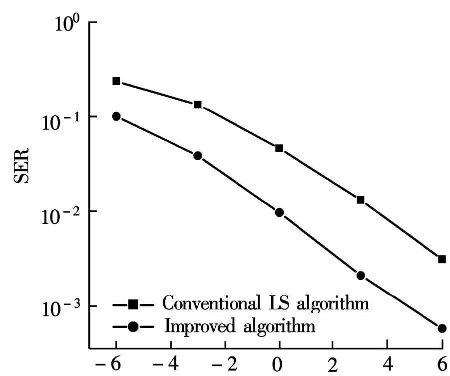

Fig.4 SER for different SINR with PTR = 0.25

Fig.5 compares the NMSE performance between two algorithms in different PTR when the SINR is −3 dB. The figure shows that the estimation error increases with the decrease of the pilot power in the conventional LS algorithm. In the improved algorithm, however, the pilot is only used to get the initial estimate. It is the iterative procedure that progressively increases the accuracy of channel vector estimation, so the estimation error is nearly constant.

Fig.6 illustrates the average iteration number for different SINR when the PTR is 0.25. It can be observed from the figure that the unmber of iterations decreases when the SINR increases. It is because the initial estimate becomes more accurate when the SINR increases. This will accelerate the convergence speed of the algorithm. Fig.6 also shows that the iteration number is small even when the SINR is − 6 dB, which demonstrates the good convergence property of this algorithm.

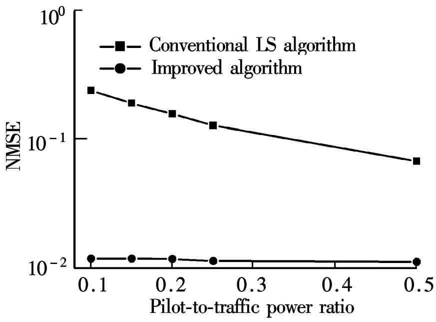

Fig.5 NMSE for different PTR with SINR = − 3 dB

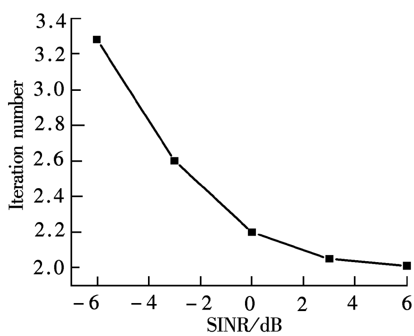

Fig.6 Iteration number for different SINR with PTR = 0.25

4 Conclusion

This paper proposes an improved algorithm to estimate the steering vector in a wider time window based on the fact that the DOA varies more slowly than fading. The traffic signals, as well as the pilot, are then used to estimate the channel vector based on the constrained LS criterion. The accuracy of channel vector estimation can increase through iteration. The proposed algorithm effectively mitigates the pilot power and time window problem from which the conventional LS algorithm suffers.

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二维 RAKE 接收机中一种约束最小二乘信道矢量估计算法

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摘 要 针对移动通信中目标方位变化比信道衰落变化慢的特点,首先利用子空间分解法估计出目标信号的导向矢量,并用估计出的导向矢量构造一个约束条件,然后用业务信号在约束最小二乘准则下估计信道矢量,本文用带投影的迭代最小二乘(ILSP)算法求解约束最小二乘问题,导频用于ILSP 算法的初始化,通过 ILSP 算法的迭代过程,可以逐步提高信道矢量估计和符号检测的精度.仿真结果表明:与传统二维 RAKE 接收机相比,本文提出的算法使系统的性能得到了有效改善.

关键词 二维 RAKE 接收机; 信道估计; 子空间分解; 约束最小二乘

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