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An improved channel estimation with multipath search for MIMO-OFDM systems

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Abstract: This paper addresses the problem of channel estimation for broadband MIMO-OFDM systems. An improved channel estimator with multipath time delay detection and channel gain estimation is proposed. In the algorithm, we used the correlation of the channel taps and a well-designed adjustment scheme to increase the accuracy of the time delay detection. The most attractive advantage is that the complicated matrix calculation is replaced by search steps which can acquire the channel order and estimate the channel parameters without significantly increasing the complexity of the system. Computer simulation showed that the proposed algorithm can track the time delays adaptively and, consequently, improve the channel estimation performance.

Key words: Channel estimation, MIMO, OFDM, Time-delay estimation
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INTRODUCTION

Multimedia wireless services require reliable high-bit-rate transmission over mobile radio channels. One of the challenges in designing such systems is the mitigation of fading propagation effects within the prescribed bandwidth and power limitations. Multiple-input and multiple-output (MIMO) techniques can be implemented to obtain a capacity gain in rich scattering environments without increasing the bandwidth or transmit power and/or to obtain the diversity gain to combat signal fading (Zheng and Tse, 2003). While orthogonal frequency division multiplexing (OFDM) is an effective technique for mitigating the effects of delay spread in a frequency selective fading channel. Therefore, growing attention has been paid to the research and development of combinations of MIMO techniques and OFDM (MIMO-OFDM) (Stuber *et al*., 2004; Ebeling *et al*., 2004).

Coherent detection requires knowledge of the channel at the receiver, which can be obtained by pilot-symbol-aided approaches (Li *et al*., 1999; Yang *et al*., 2001; Chen and Chang, 2003). In (Li *et al*., 1999), discrete Fourier transform (DFT) approach was used to perform robust channel estimation for OFDM systems with transmit diversity. The algorithm also showed that significant-tap catching can further improve the estimation performance. However, in a multipath channel with non-sample-spaced time delays, the energy leakage problem will bring difficulties in catching the significant taps and possibly result in an error floor. A parametric channel model based channel estimation algorithm was proposed by Yang *et al*.(2001) to reduce the error floor, where complicated matrix calculations were used to estimate the channel order and the multipath time delays. Chen and Letaief (2002) extended the algorithm to space-time coded OFDM systems. The parametric channel model is reasonable for high-speed data transmission in large cells and has been applied to the GSM system and the DVB-T system to improve estimator performance. The corresponding estimator can perform quite close to the case with ideal channel state information, but the high computational complexity makes it difficult to use in practice. In this paper, we focus on reduced complexity channel estimation which can also perform well in non-sample-spaced multipath channels.

We developed an improved channel estimator with multipath search for MIMO-OFDM systems. The algorithm, based on a parametric channel model, jointly estimates the channel order, the time delays and the channel gains. First, we extend a significant-tap approach (Han *et al.*, 2002) to MIMO-OFDM systems. The method utilizes the correlation of channel taps to obtain more robustness to the energy leakage. Then, utilizing the slowly time-varying nature of time delays, we propose a scheme to bound the channel order and adjust the multipath time delays by smoothing over symbols. For the channel environments where the channel order is unknown or time varying, this adaptive adjustment scheme can track the channel order and the time delays in real-time. Compared with the method proposed by Yang *et al.*(2001), the present work does not have complicated matrix calculation and is readily hardware-implementable.

SYSTEM DESCRIPTION

System model

A MIMO OFDM system with N_{t} transmit and N_{r} receive antennas is shown in Fig.1. At the transmission time n, a binary data block $\{c[n,k]: k=0, 1, \ldots\}$ is transformed into N_{t} different signals $\{x_i[n,k]: k=0,1,\ldots\}$ for $i=1, \ldots, N_{\mathrm{t}}$. Each of these signals forms an OFDM block. The transmit antennas simultaneously transmit OFDM signals modulated by $x_i[n,k]$. The channel is assumed to be quasi-stationary, which means it does not change within one OFDM symbol but varies for different symbols. Then, DFT of the received signal at the receive antenna $j\in\{1, \ldots, N_{\mathrm{r}}\}$ can be given by

$$y_j[n,k]=\sum_{i=1}^{N_{\mathrm{t}}} x_i[n,k]H_{ji}[n,k]+v_j[n,k], \tag{1}$$

where $k=0,\ldots,K-1$; K is the number of tones of an OFDM block; $H_{ji}[n,k]$ is the channel frequency response for the kth tone at time n, corresponding to the ith transmit antenna and the jth receive antenna; and $v_j[n,k]$ denotes the additive white Gaussian noise with zero mean and variance of σ_n^2 on the jth receive antenna.

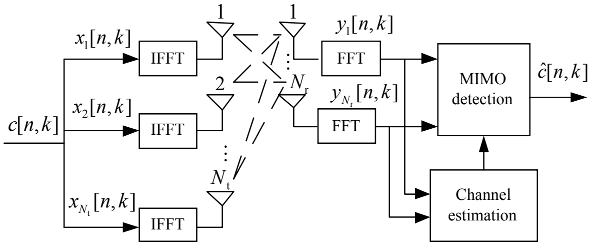


Fig.1 Simplified block diagram of MIMO-OFDM system

Channel model

The channel impulse response (CIR) corresponding to the transmitter i and receiver j can be modeled by

$$c_{ij}(t,\tau)=\sum_{l}\gamma_{ij}^{(l)}(t)\delta(\tau-\tau_{ji}^{(l)}), \tag{2}$$

where $\tau_{ji}^{(l)}$ is the delay of the lth path; and $\gamma_{ij}^{(l)}(t)$ is the lth path complex amplitude. Due to the motion of the vehicle, $\gamma_{ij}^{(l)}(t)$'s are modeled to be wide-sense stationary (WSS) narrowband complex Gaussian random process, which are statistically independent and identically distributed (i.i.d.) for different l, i or j.

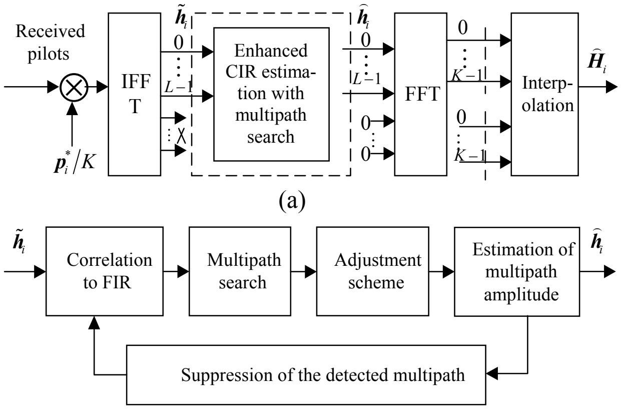
Notice that both the transmitted OFDM signal and the received signal are frequency limited. Then the channel time response will be

$$h_{ij}(t,\tau)=c_{ij}(t,\tau)\otimes g(\tau)=\sum_{l}\gamma_{ij}^{(l)}(t)g(\tau-\tau_{ji}^{(l)}), \tag{3}$$

where $\otimes$ denotes convolution; $g(\tau)$ is a shaping pulse, the frequency response of which is usually a raised-cosine Nyquist filter. If the delay $\tau_{ji}^{(l)}$ is an integer, all the energy from $\tau_{ji}^{(l)}$ is mapped to the tap $h(t,\tau_{ji}^{(l)})$. However, for non-sample-spaced case, i.e., $\tau_{ji}^{(l)}$ is not an integer, the energy from $\tau_{ji}^{(l)}$ will leak to all the other taps (Beek *et al.*, 1995). Fortunately, most of the energy is kept in the neighborhood of the original pulse location.

CHANNEL ESTIMATION

The proposed channel estimator is depicted in Fig.2a. First, the channel frequency response is estimated by using the pilot symbols. Then the channel gains over the entire frequency-time grid can be obtained by interpolation. To improve the performance, we developed an enhanced CIR estimation algorithm with multipath search embedded in the first part. The diagram of the enhanced algorithm is shown in Fig.2b. It is an iterative algorithm which combines the estimation of time delays and channel gains. Since the estimator works independently for each receive antenna, in the following, the receive antenna index j will be omitted.



(a)

(b)

Fig.2 The block diagrams of the proposed estimator. (a) The overall estimator block diagram; (b) The joint estimation block diagram

Joint time delay detection and channel estimation

At each transmit antenna, pilot tones are inserted in all subcarriers of a particular OFDM symbol to form an OFDM training symbol, and the training symbols are transmitted at an appropriate regular rate determined by the time varying nature of the wireless channel. Let $p_i[n,k]$ for k=0,…,K−1 denote the training symbol for the ith transmit antenna, which satisfy (Li, 2002)

$$p_i[n,k] = p_1[n,k] W_K^{-L(i-1)k}, \tag{4}$$

where $W_K=\exp(-j(2\pi/K))$; and L is the maximum multipath delay with $L \le K/N_t$. Then the channels corresponding to the same receive antenna and the different transmit antennas can be separated without interantenna interference. Let $y[n,k]$ for k=0,…,K−1 denote the received training symbol at a receive antenna. The temporal CIR corresponding to the ith transmit antenna is given by (Li, 2002)

$$\tilde{\boldsymbol{h}}_i = \boldsymbol{q}_i / K, \tag{5}$$

where $\boldsymbol{q}_i=(q_i[n,0],\ldots,q_i[n,L-1])^{\mathrm{T}}$, and $q_i[n,l] = \sum_{k=0}^{K-1} y[n,k] p_i^*[n,k] W_K^{-kl}$.

From Eqs.(1), (3) and (5), $\tilde{\boldsymbol{h}}_i$ can be expressed as

$$\tilde{\boldsymbol{h}}_i = \boldsymbol{G}\boldsymbol{c}_i + \boldsymbol{w}_i, \tag{6}$$

where $\boldsymbol{c}_i$ is the ideal CIR; $\boldsymbol{w}_i$ represents the effect of the additive white Gaussian noise; and $\boldsymbol{G}$ consists of the shaping pulse filter coefficients described by

$$\boldsymbol{G} = \begin{bmatrix} g(0\cdot\beta) & \cdots & g(0-L\beta+1) \\ \vdots & & \vdots \\ g((L-1)\beta) & \cdots & g((L-1)\beta-L\beta+1) \end{bmatrix}_{L\times L\beta}, \tag{7}$$

where β is the interpolation factor.

The correlation of $\tilde{\boldsymbol{h}}_i$ and $\boldsymbol{G}$ is

$$\boldsymbol{s}_i = \boldsymbol{G}^{\mathrm{H}}\tilde{\boldsymbol{h}}_i = \boldsymbol{G}^{\mathrm{H}}\boldsymbol{G}\boldsymbol{c}_i + \boldsymbol{G}^{\mathrm{H}}\boldsymbol{w}_i, \tag{8}$$

where $\boldsymbol{s}_i=(s_i[n,0], \ldots, s_i[n, L\beta-1])^{\mathrm{T}}$. The correlation of the channel taps can concentrate the leaky energy so that the taps nearest to the locations of the time delays will attain peaks of $\boldsymbol{s}_i$. Then more accurate time-delay estimation will be achieved.

From $\boldsymbol{s}_i$, we search the paths successively in order to suppress the inter-path interferences. The index of the strongest tap can be obtained by

$$l_i^m = \arg\max_l (s_i[n,l]). \tag{9}$$

A well-designed adjustment scheme described in the next subsection will check whether the tap is a real path. If the tap is checked to be a real path, its amplitude can be estimated by

$$\hat{c}_i[n,l_i^m] = s_i[n,l_i^m] \big/ r_g[l_i^m, l_i^m], \tag{10}$$

where $r_g[u,v]$ is the (u,v)th element of $\boldsymbol{G}^{\mathrm{H}}\boldsymbol{G}$ known to the receiver in advance.

For searching the next strongest path, we suppress the detected path as follows

$$\boldsymbol{s}_i = \boldsymbol{s}_i - \hat{c}_i[n,l_i^m] r_g[l_i^m], \tag{11}$$

where $r_g[l]$ is the lth column of $\boldsymbol{G}^{\mathrm{H}}\boldsymbol{G}$.

If there are other undetected paths indicated by the adjustment scheme, the algorithm goes on detecting the next tap with Eqs.(9)~(11). Otherwise, stop searching and compute the estimate of the channel time response $\hat{\boldsymbol{h}}_i$ as

$$\hat{\boldsymbol{h}}_i = \boldsymbol{G}\hat{\boldsymbol{c}}_i. \tag{12}$$

The corresponding estimate of the channel frequency response is

$$\hat{H}_i[n,k] = \sum_{l=0}^{L-1} \hat{h}_i[n,l] W_K^{kl}, \quad k = 0,\ldots,K-1. \tag{13}$$

With the channel estimation on the pilot symbols, the channel parameters corresponding to the regular symbols can be obtained by interpolation according to the time correlation of channels as described by

$$\hat{H}_i[n,k] = \sum_{m\in M} \alpha(n;m)\hat{H}_i[m,k], \quad k = 0,\ldots,K-1 \tag{14}$$

where m indexes the locations of the training symbols; M denotes the set that contains m; and $\alpha(n;m)$ are the coefficients of the interpolation filter. The optimal filter in the minimum mean square error sense is a Wiener filter (Hoeher *et al.*, 1997). Other lower complexity interpolation filters are proposed in (Moon and Choi, 2000).

Adjustment scheme

We propose an adjustment scheme to bound the channel order and check the detected taps by smoothing over symbols. The lower and upper bound of the channel order are estimated adaptively according to the detected time delays. We assume that all the detected taps under the lower bound of the channel order are real paths, whose gains can be estimated by the current training symbol. The other detected taps during the lower bound and upper bound of the channel order are possibly pseudo paths, which should be checked by comparison with the detection results of the previous training symbol. This well-designed adjustment scheme can make full use of the slow time-varying nature of time delays and the randomicity of additive white Gaussian noise, and then achieves more robustness to the interferences.

Adjustment Scheme: The lower and upper bound of the channel order is denoted by T_- and T_+. Executing the loop of joint time delay detection and channel estimation, we obtain the indexes of the detected taps from Eq.(9) and put them into a set $\boldsymbol{I}_\mathrm{d}$. Let $\boldsymbol{I}_\mathrm{r}$ denote the indexes of the real paths checked from $\boldsymbol{I}_\mathrm{d}$. In order to track the slow time-varying paths, we define an extension set $\tilde{\boldsymbol{I}}_\mathrm{d}$ of $\boldsymbol{I}_\mathrm{d}$ and an extension set $\tilde{\boldsymbol{I}}_\mathrm{r}$ of $\boldsymbol{I}_\mathrm{r}$. The extension set contains the original set and one or several taps around each original element. For example, $\boldsymbol{I}_\mathrm{d} = \{6\} \rightarrow \tilde{\boldsymbol{I}}_\mathrm{d} = \{5,6,7\}$. The following steps are implemented at the time when dealing with the jth training symbol.

(1) For j=1.

Given the initial values of the channel order bound, for example $T_-^{(1)} = 1$, $T_+^{(1)} = 8$, since the number of the dominant paths in large cells is typically two to six. Obtain $\boldsymbol{I}_\mathrm{d}^{(1)}$ from the estimation loop. All the detected taps are chosen as real paths, so we get that $\boldsymbol{I}_\mathrm{r}^{(1)} = \boldsymbol{I}_\mathrm{d}^{(1)}$.

(2) For j=2,3,…

Obtain $\boldsymbol{I}_\mathrm{d}^{(j)}$ from the estimation loop. The taps corresponding to the first $T_-^{(j-1)}$ elements, denoted by $\boldsymbol{I}_\mathrm{d}^{(j)}[1:T_-^{(j-1)}]$, are chosen as real paths. The taps corresponding to the rest of the indexes, denoted by $\boldsymbol{I}_\mathrm{d}^{(j)}[T_-^{(j-1)}+1:T_+^{(j-1)}]$, should be checked using $\tilde{\boldsymbol{I}}_\mathrm{d}^{(j-1)}$. We choose

$$\boldsymbol{I}_\mathrm{r}^{(j)} = \boldsymbol{I}_\mathrm{d}^{(j)}[1:T_-^{(j-1)}] \cup \left(\boldsymbol{I}_\mathrm{d}^{(j)}[T_-^{(j-1)}+1:T_+^{(j-1)}] \cap \tilde{\boldsymbol{I}}_\mathrm{d}^{(j-1)}\right) \tag{15}$$

and update the values of the bound of the channel order as follows

$$T_{-}^{(j)} = L(\boldsymbol{l}_{\mathrm{r}}^{(j)} \cap \tilde{\boldsymbol{l}}_{\mathrm{r}}^{(j-1)}),\quad T_{+}^{(j)} = T_{-}^{(j)} + \Delta, \tag{16}$$

where $L(a)$ counts the number of the elements in set a, and Δ denotes the interval between the lower and the upper bound of the channel order.

Many channel estimators based on the time delay detection (Yang *et al*., 2001; Nicoli *et al*., 2003) involve complicated matrix calculations including eigenvalue decomposition to acquire the channel order and detect the time delays. However, in the proposed estimator, following from Eqs.(9)~(11), (15), (16), the simple iterative algorithm can obtain the time delay information and the channel parameters jointly without complicated matrix calculations.

PERFORMANCE EVALUATION

In this section, we will investigate the improved channel estimation algorithm performance in multipath Rayleigh fading channels. Consider a 4QAM MIMO-OFDM system with two transmit and two receive antennas. The Alamouti scheme is applied to obtain the transmit diversity. The system occupies a bandwidth of 20 MHz (T=50 ns) operating in the 3 GHz frequency band. The entire channel bandwidth is divided into 512 subchannels. Then the symbol duration is 25.6 μs. A 6.4 μs cyclic prefix (CP) is added to each symbol to combat the intersymbol interference caused by multipath delay spread. This results in a total block length Ts 32 μs.

As shown in Table 1, two different channel environments are considered in the evaluation. Channel A is a 6-path fading channel with sample-spaced time delays and an exponentially decaying power-delay profile. Channel B is the “vehicular A” channel defined by ETSI for UMTS, which has non-sample-spaced multipath. The maximum Doppler frequency is set to be 167 Hz in the two channel environments.

In the estimator, the interpolation factor of the shaping pulse β is set to be 4 in Eq.(7), and the interval between the lower and the upper bound of the channel order Δ is set to be 2 in Eq.(16). So the resolution of the time delay detection is $1/\beta$=0.25.

Table 1 Characteristics of two channel environments

Channel A (sample-spaced)		Channel B (non-sample-spaced)	
Time delays (ns)/T	Power profile (dB)	Time delays (ns)/T	Power profile (dB)
0/0	0	0/0	0
750/15	−4	310/6.2	−1
1500/30	−9	710/14.2	−9
2250/45	−13	1090/21.8	−10
3000/60	−17	1730/34.6	−15
3750/75	−22	2510/50.2	−20

Note: T=50 ns

Table 2 shows the probability of correct acquisition of different path time delays in the sample-spaced multipath channel at SNR=12 dB. As can be seen, the probability of correct acquisition of the time delays is high even for the paths with the smaller power, for example, the fifth and the sixth paths. Although we set the initial value of the upper bound of the channel order to be 8, the probability of error acquisition of the two pseudo paths is very small.

Table 2 Probability of correct acquisition of different path time delays in sample-spaced multipath channel at SNR=12 dB

Tap	Probability
1	1.00
2	1.00
3	0.99
4	0.98
5	0.95
6	0.79
7 (pseudo)	0.19
8 (pseudo)	0.00

Hence, the estimator can adaptively estimate the channel order and the time delays.

Fig.3 shows the detected multipath time delays at SNR=12 dB. Figs.3a and 3c show all the detected time delays in Channels A and B respectively. Obviously, the estimator can correctly acquire the time delays and only catch a few pseudo paths sporadically. As shown in Fig.3b, the estimator acquires the exact value of the time delay of a sample-spaced path. For the non-sample-spaced path, as shown in Fig.3c, the estimated values of the time delay are given by the nearest neighborhood of the original time delay. Then the estimates have small departures from the exact value. These small departures increase some mean square error (MSE) of the estimator as can be seen in Fig.5a, but cause only a very slight performance degradation of the bit error rate (BER) of the system as shown in Fig.5b.

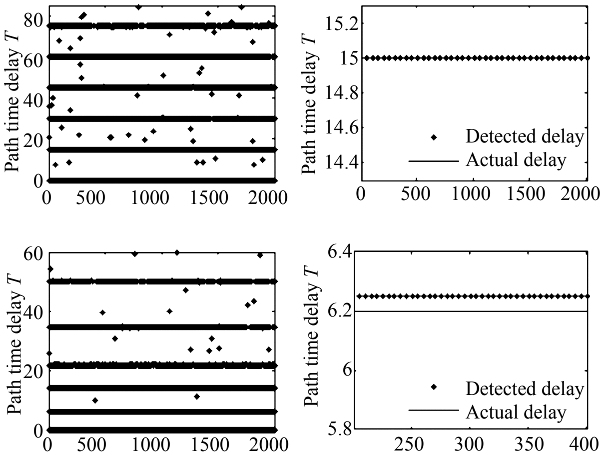


Fig.3 Detecting the channel path time delays at *SNR*=12 dB
(a) Time delays in Channel A; (b) Time delay of a sample-spaced path; (c) Time delays in Channel B; (d) Time delay of a non-sample-spaced path

For the channel with time varying delay spread, we evaluate the tracking ability of the proposed algorithm. The delay of each path increases at a rate of 1 μs/s or 500 ns/s. The simulation results at *SNR*=12 dB are given in Fig.4. We can see that the proposed estimator can track the changes in delay. The tracking curves are scalariform because the resolution of the time delay detection is 0.25.

Fig.5a gives the MSE performances of the estimator in multipath fading channels. In the sample-spaced multipath channel, the MSE of our estimator decreases with increasing signal-to-noise ratio. For example when *SNR*=12 dB, the MSE is about 2×10^{-3}. In the non-sample-spaced multipath channel, as interpreted above, the time delay estimation departures will introduce some performance loss of MSE. The figure also contains the MSE curves of the LS estimator for comparison. As expected, the MSE curves of the proposed estimator are lower than those of the LS estimator.

Accordingly, as seen in Fig.5b, the BER of our estimator is lower than that of LS estimator in either the sample-spaced or the non-sample-spaced multipath channel. The performance of the proposed estimator is quite close to the performance in the case of ideal channel parameters especially in the sample-spaced multipath channel.

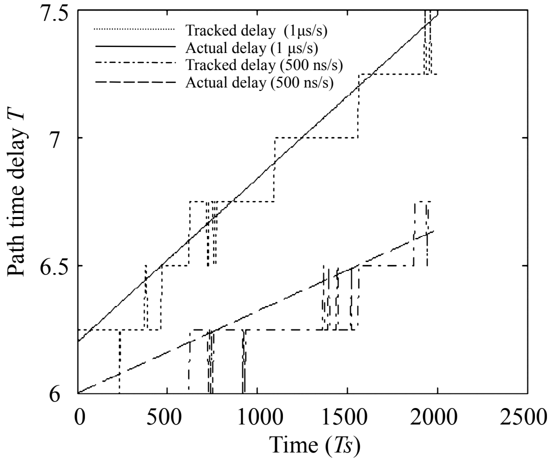
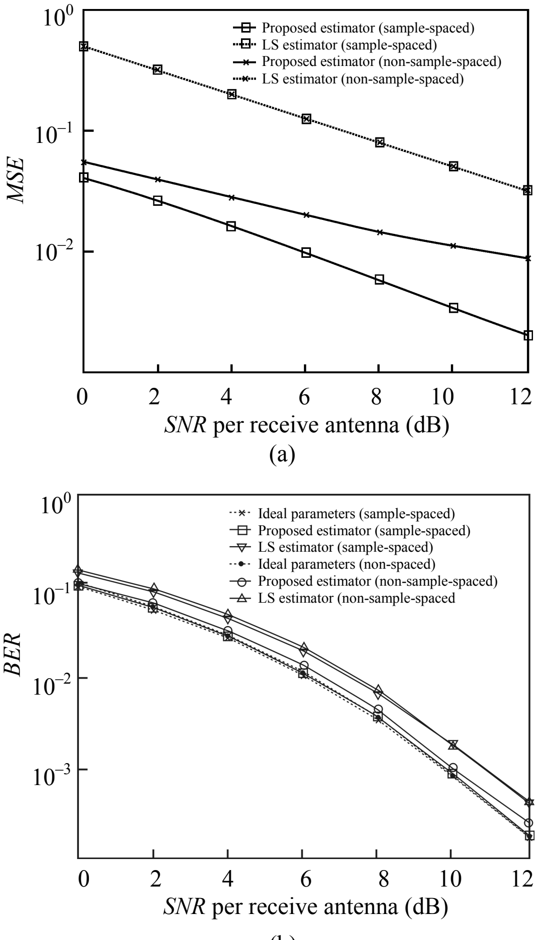


Fig.4 Tracking the channel path time delay changes at *SNR*=12 dB



(a)

(b)

Fig.5 The MSE (a) and BER (b) performance comparison of channel estimation

CONCLUSION

In this paper, we propose an effective channel estimation algorithm for broadband MIMO-OFDM systems by exploiting the sparse nature of the channel. Utilizing the correlation of the channel taps, time delay detection can yield more robustness to the energy leakage. An adjustment scheme is proposed to get the channel order and track the delay spreads. The techniques can increase the accuracy of the time delay detection and, consequently, improve the channel estimation performance. The most attractive advantage is that the complicated matrix calculation is replaced by search steps which can acquire the channel order and estimate the channel parameters without significantly increasing the complexity of the system.

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