Journal of Zhejiang University SCIENCE A
ISSN 1009-3095 (Print); ISSN 1862-1775 (Online)
www.zju.edu.cn/jzus; www.springerlink.com
E-mail: jzus@zju.edu.cn

RLS channel estimation with adaptive forgetting factor in space-time coded MIMO-OFDM systems*

LIANG Yong-ming[1], LUO Han-wen[1], HUANG Jian-guo[2]
([1]*Department of Electronic Engineering, Shanghai Jiao Tong University, Shanghai 200030, China*)
([2]*College of Marine Engineering, Northwestern Polytechnical University, Xi'an 710072, China*)
E-mail: liangyongming@sjtu.edu.cn; luohanwen@hotmail.com; jghuang@nwpu.edu.cn
Received Apr. 16, 2005; revision accepted June 16, 2005

Abstract: Considering that channel estimation plays a crucial role in coherent detection, this paper addresses a method of Recursive-least-squares (RLS) channel estimation with adaptive forgetting factor in wireless space-time coded multiple-input and multiple-output orthogonal frequency division multiplexing (MIMO-OFDM) systems. Because there are three different forgetting factor scenarios including adaptive, two-step and conventional ones applied to RLS channel estimation, this paper describes the principle of RLS channel estimation and analyzes the impact of different forgetting factor scenarios on the performances of RLS channel estimation. Simulation results proved that the RLS algorithm with adaptive forgetting factor (RLS-A) outperformed that with two-step forgetting factor (RLS-T) or with conventional forgetting factor (RLS-C) in both estimation accuracy and robustness over the multiple-input multiple-output (MIMO) channel, i.e., a wide-sense stationary uncorrelated scattering (WSSUS) and frequency-selective slowly fading channel. Hence, we can employ the RLS-A method by adjusting forgetting factor adaptively to track and estimate channel state parameters successfully in space-time coded MIMO-OFDM systems.

Key words: MIMO-OFDM, Channel estimation, RLS algorithm, Adaptive forgetting factor
doi:10.1631/jzus.2006.A0507 **Document code:** A **CLC number:** TN911.23; TN929.5

INTRODUCTION

The very dramatic increase in demand for high-rate data transmission stimulated greater research efforts in developing wideband wireless communication systems which could support high-rate transmission over wireless channels. But we have to face the challenges from the multi-path effects because multi-path propagation of wire wave is usually seen as a harmful factor which leads to one of the most troublesome radio channel problems. However, the multi-path effects can become a beneficial factor for utilizing multiple-input multiple-output (MIMO) systems.

MIMO technique can increase system capacity and spectrum efficiency in rich scattering environments without increasing the bandwidth or transmitted power (Foschini and Gans, 1998). While orthogonal frequency division multiplexing (OFDM) is an effective technique for mitigating the effects of delay spread in frequency selective-fading channels. Moreover, space-time processing techniques could be effectively applied to improve the system performance. Therefore, a system that combines the MIMO system, space-time coding and OFDM can provide spectrally efficient high data rate transmission over a fading channel. As a result, a space-time coded MIMO-OFDM system is an attractive scheme for implementing B3G, 4G wireless cellular systems (Stuber *et al*., 2004), WLAN (Coon *et al*., 2003) and HIPERMAN (Fan *et al*., 2004).

Channel estimation uses training sequence including preamble and pilot for estimating channel state information (CSI) crucial for coherent detection

* Project supported by the National Natural Science Foundation of China (No. 60272079), and the Hi-Tech Research and Development Program (863) of China (No. 2003AA123310)

in wireless communication systems. Training-sequence-aided channel estimation is very common in OFDM or CDMA systems and has been studied by many researchers. Li (2002) introduced a pilot-aided channel estimation method which can attain diversity and multiplexing gain in MIMO-OFDM systems. Schafhuber *et al*.(2003b) demonstrated a channel tracking and estimation method based on adaptive algorithm. This method exploited least-mean-square (LMS) adaptive filter not requiring knowledge of channel statistics (Haykin, 2003). Schafhuber *et al*.(2003a) proposed a channel estimation method based on adaptive Wiener filters for time-varying channel in wireless OFDM systems. For multi-user CDMA systems, a channel estimation method with RLS algorithm for the downlink was presented by Khuabut *et al*.(2004a). Khuabut *et al*.(2004b) introduced improvement of RLS channel estimation in forward link of MC-CDMA systems. Channel estimation is very important for the receiver and is an indirect method for improving the performance of the receiver, i.e., channel estimation supplies the receiver with channel state information for coherent detection. Liu and Xu (1995) introduced a method of multi-user blind channel estimation and spatial channel pre-equalization. Wong and Cheung (2004) proposed a new “fractional self-decorrelation” technique to enhance single-user-type DS-CDMA detectors’ output SINR in blind space-time RAKE receivers. For MIMO systems, Sun *et al*.(2003) proposed RLS based adaptive approach iterative channel estimation combined with MIMO minimum mean square error (MMSE) equalizer by employing soft feedback information. And Benjebbour and Yoshida (2003) presented a simplification of the weights update from a MIMO-RLS to a modified MIMO-LMS algorithm for semi-adaptive ordered successive detection in MIMO wireless systems. Chen and Su (2004) introduced the RLS algorithm for estimating the MIMO channel in spatially correlated environments. As to space-time coded systems, Tarokh *et al*.(1998; 1999) introduced the orthogonal design of space-time block codes and corresponding data detection methods. Chueh *et al*.(2004) proposed an iterative adaptive receiver design using the multi-layer RLS algorithm with applications to layered space-time coded system. Cozzo and Hughes (2003) demonstrated a method of joint channel estimation and data detection in space-time communications. The joint algorithm is realized by an iterative space-time receiver based on the expectation maximization (EM) algorithm. Suthaharan *et al*.(2003) proposed a computationally efficient adaptive channel estimation scheme for space-time block coded (STBC) MIMO-OFDM systems. The paper used the conventional RLS algorithm for channel estimation and MMSE interference cancellation (IC) method for signal detection.

We investigated the RLS channel estimation in space-time coded MIMO-OFDM systems in this paper. Alamouti (1998)’s space-time block code is employed in this paper. Compared with (Li, 2002), our RLS channel estimation scheme does not require the inversion of large matrix and prior knowledge of the second-order statistics (SOS) of channel and noise; compared with (Schafhuber *et al*., 2003a), our scheme does not employ the LMS (Schafhuber *et al*., 2003b) but the RLS adaptive filter instead of Wiener filters without any prior knowledge of the SOS; compared with (Khuabut *et al*., 2004a; 2004b; Liu and Xu, 1995; Wong and Cheung, 2004), our scheme is not applicable for CDMA systems but for MIMO-OFDM systems; compared with (Sun *et al*., 2003; Benjebbour and Yoshida, 2003), considering cyclic prefix (CP) mitigating inter-symbol interference (ISI) effectively, our scheme does not employ equalizers to mitigate ISI in the MIMO-OFDM system; compared with (Chen and Su, 2004), our scheme uses the WSSUS channel in place of the spatially correlated channel; compared with (Chueh *et al*., 2004; Cozzo and Hughes, 2003; Suthaharan *et al*., 2003), our scheme employs the Maximum Likelihood (ML) algorithm for signal detection instead of IC, EM or other algorithms; compared with (Sun *et al*., 2003; Chueh *et al*., 2004; Cozzo and Hughes, 2003), our scheme has low computational complexity because it neither applies iterative principle nor combines channel estimation with signal detection. Our scheme is designed to track and estimate WSSUS slowly time-varying channel in space-time coded MIMO-OFDM systems.

The rest of the paper is organized as follows. Firstly we briefly describe the MIMO-OFDM system and the WSSUS channel model. Secondly we propose a kind of preamble structure for RLS channel estimation. Thirdly the principle of RLS channel estimation is demonstrated in detail. Hereafter, we analyze three RLS algorithms including RLS-A, RLS-T and RLS-C respectively. Computer simulation results are provided to compare and evaluate the performances of different RLS algorithms including RLS-A, RLS-T and RLS-C. Finally, some beneficial conclusions are drawn.

SYSTEM MODEL

As Fig.1 shows, we use N_T transmit antennas, N_R receive antennas, n OFDM symbols and K subcarriers in a MIMO-OFDM system. The transmitted symbol vector is shown below.

$$\boldsymbol{a}[n,k]=[a^{(1)}[n,k],\ldots,a^{(N_T)}[n,k]]^T,\ n\in\mathbb{Z},\ k=0,\ldots,K-1, \quad (1)$$

where $a^{(i)}[n,k]$ denotes the symbol transmitted at the symbol time n, by subcarrier k, and antenna i. The nth OFDM symbol $S_n[m]$ can be obtained by performing an inverse fast discrete Fourier transform (IFFT) to the $\boldsymbol{a}[n,k]$ and inserting a cyclic prefix of length L_{CP},

$$\boldsymbol{S}_n[m]=\begin{cases}\dfrac{1}{\sqrt{KN_T}}\sum_{k=0}^{K-1}\boldsymbol{a}[n,k]e^{j2\pi mk/K}, & m=-L_{CP},\ldots,K-1,\\ 0, & \text{else.}\end{cases} \quad (2)$$

Thus the duration of each OFDM symbol is $N=K+L_{CP}$. The overall baseband transmitted signal is $\boldsymbol{S}[m]=\sum_{n=-\infty}^{+\infty}\boldsymbol{S}_n[m-nN]$.

CHANNEL MODEL

The complex baseband representation of wireless mobile channel impulse response of the N_T transmitter and N_R receiver can be described by Eq.(3).

$$h_{N_TN_R}(t,\tau)=\sum_l \xi^{(l)}_{N_TN_R}(t)c(\tau-\tau^{(l)}_{N_TN_R}), \quad (3)$$

where $\tau^{(l)}_{N_TN_R}$ is the delay of the lth path, and $\xi^{(l)}_{N_TN_R}(t)$ is the complex amplitude of the lth path. $\xi^{(l)}_{N_TN_R}(t)$ is wide-sense stationary (WSS) narrowband complex Gaussian random process, which is statistically independent of different paths and identically distributed (i.i.d.) for different l, N_T and N_R.

The frequency response at time t is

$$\begin{aligned}H_{N_RN_T}(t,f)&=\int_{-\infty}^{+\infty}h_{N_RN_T}(t,\tau)e^{j2\pi f\tau}d\tau\\ &=C(f)\sum_l \xi^{(l)}_{N_RN_T}e^{j2\pi f\tau_k},\end{aligned} \quad (4)$$

where

$$C(f)=\int_{-\infty}^{+\infty}c(\tau)e^{-j2\pi f\tau}d\tau. \quad (5)$$

As Fig.1 shows, the signal from each receiver is formed by the parameter matrix $\boldsymbol{H}[m,l]$ of fading MIMO $N_R\times N_T$ channel, the transmitted signal $\boldsymbol{S}[m]$, and the noise $\boldsymbol{\eta}[m]$. $\boldsymbol{\eta}[m]$ is stationary white Gaussian noise with distribution expressed by $N(0,\sigma_\eta^2\boldsymbol{I})$.

$$\boldsymbol{r}[m]=\sum_{l=0}^{L-1}\boldsymbol{H}[m,l]\boldsymbol{S}[m-l]+\boldsymbol{\eta}[m]. \quad (6)$$

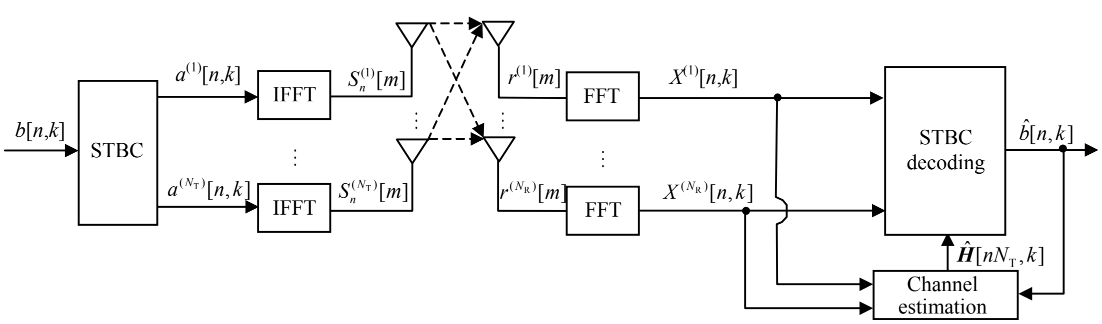

Fig.1 MIMO-OFDM system

We assume the channel of MIMO-OFDM system is uncorrelated scattering (US). And we further assume the length of channel tap L meet the relation $L\leq L_{\mathrm{CP}}+1$ in order to avoid inter-symbol-interference caused by multi-path effects. The receiver signal $\boldsymbol{r}[m]$ is demodulated by removing the cyclic prefix and performing FFT,

$$\boldsymbol{X}[n,k]=\frac{1}{\sqrt{K}}\sum_{m=0}^{K-1}\boldsymbol{r}[nN+m]\mathrm{e}^{-\mathrm{j}2\pi km/K}. \tag{7}$$

If $\boldsymbol{H}[m,l]$ varies negligibly within one OFDM symbol, the input/output relation in this MIMO-OFDM system can be expressed as:

$$\boldsymbol{X}[n,k]=\hat{\boldsymbol{H}}[n,k]\,\boldsymbol{a}[n,k]+\hat{\boldsymbol{\eta}}[n,k], \tag{8}$$

here $\boldsymbol{X}[n,k]$, $\hat{\boldsymbol{H}}[n,k]$ and $\hat{\boldsymbol{\eta}}[n,k]$ are all $N_{\mathrm{R}}\times N_{\mathrm{T}}$ matrixes, and $\boldsymbol{a}[n,k]$ is a $N_{\mathrm{T}}\times N_{\mathrm{T}}$ matrix. The channel matrix $\hat{\boldsymbol{H}}[n,k]$ and noise matrix in the time/frequency domain is obtained through FFT as Eqs.(9) and (10) show.

$$\hat{\boldsymbol{H}}[n,k]=\frac{1}{\sqrt{N_{\mathrm{T}}}}\sum_{l=0}^{L-1}\boldsymbol{H}[nN,l]\mathrm{e}^{-\mathrm{j}2\pi kl/K}, \tag{9}$$

$$\hat{\boldsymbol{\eta}}[n,k]=\frac{1}{\sqrt{K}}\sum_{m=0}^{K-1}\boldsymbol{\eta}[nN+m]\mathrm{e}^{-\mathrm{j}2\pi km/K}. \tag{10}$$

PREAMBLE STRUCTURE

This paper proposes an RLS channel estimation method using preamble symbols. The preamble structure consists of six preamble symbols and sixty data as Fig.2 shows. Barhumi *et al*.(2003) introduced a method which designs pilot structure on the principle of MMSE in MIMO-OFDM systems. This method proposed in (Barhumi *et al*., 2003) requires that training sequences should be orthogonal to each other after phase shift. In order to make the preamble structure in Fig.2 meet the principles of (Barhumi *et al*., 2003), th10e preamble symbols are designed into orthogonal sequences. In order to process signals simply, the preamble sequences in different subcarriers but in the same OFDM symbol are the same in one antenna, and these preamble sequences from different antennas are orthogonal to each other. Fig.2 shows that all preamble sequences are inserted. $\boldsymbol{P}$ is the preamble matrix consisting of the preamble sequences from each antenna. $\boldsymbol{P}=[\boldsymbol{P}_1, \ldots, \boldsymbol{P}_{N_{\mathrm{T}}}]$, here $\boldsymbol{P}_n$ is a preamble sequence of one antenna and a cycle sequence whose period is N_{T}, i.e., $\boldsymbol{P}_{n+N_{\mathrm{T}}}=\boldsymbol{P}_n$.

$$\boldsymbol{P}\boldsymbol{P}^{\mathrm{H}}=\boldsymbol{P}^{\mathrm{H}}\boldsymbol{P}=N_{\mathrm{T}}\boldsymbol{I}. \tag{11}$$

A convenient choice for $\boldsymbol{P}$ is $N_{\mathrm{T}}\times N_{\mathrm{T}}$ FFT matrix. For example, $N_{\mathrm{T}}=4$ gives $\boldsymbol{P}_1=[1,1,1,1]^{\mathrm{T}}$, $\boldsymbol{P}_2=[1,-1,1,-1]^{\mathrm{T}}$, $\boldsymbol{P}_3=[1,-1,-1,1]^{\mathrm{T}}$, $\boldsymbol{P}_4=[-1,-1,1,1]^{\mathrm{T}}$.

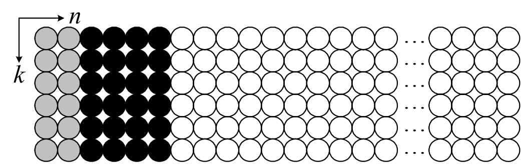

Fig.2 Location of preamble and data symbols for one transmit antenna

RLS ALGORITHM

The RLS channel estimation proposed in this paper consists of three steps as Fig.3 shows. We first determined the relation between input and output to obtain the channel parameters in the frequency domain and performed IFFT to obtain the channel parameters of the time/delay domain. In fact this step is a channel estimation method based on least-square

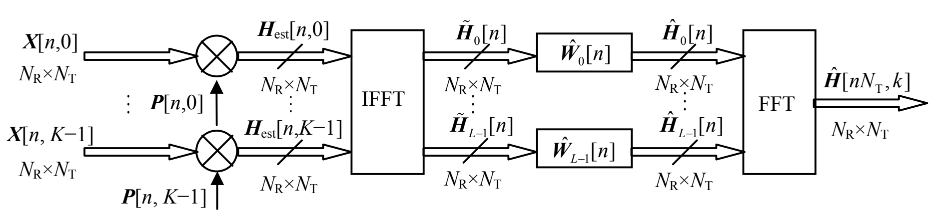

Fig.3 Principle of channel estimation

(LS) algorithm without noise effect (Haykin, 2003). Then, we performed channel estimation with the help of RLS filter. Finally, we performed FFT to estimate the channel parameters in the time/frequency domain.

The RLS channel estimation is shown in Fig.4. Because the MIMO channel is considered as a WSSUS channel, different subchannels will be analyzed respectively in order to simplify the analysis. We can use the channel state parameter $\tilde{\boldsymbol{H}}_l[n]$ to draw the $\hat{\boldsymbol{H}}_l[n]$ in the time/delay domain by means of RLS filtering. If we apply the Wiener filter to channel estimation, we have to presuppose second-order statistics of channel and noise. To reduce the complexity of channel estimation, we propose the channel estimation based on RLS filter. This channel estimation needs no prior knowledge of channel and noise statistics and adjusts weight to estimate channel adaptively.

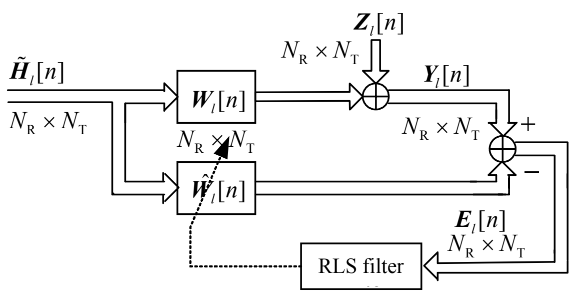

Fig.4 Algorithm of adaptive filter

For RLS filter, the adaptive algorithm is as given below:

$$\hat{\boldsymbol{H}}_l[n] = \hat{\boldsymbol{W}}_l^{\mathrm{H}}[n]\tilde{\boldsymbol{H}}_l[n],\ n\geq 0,\ l=0,1,\ldots,L_{\mathrm{CP}}. \quad (12)$$

Estimation error,

$$\boldsymbol{E}_l[n]=\boldsymbol{Y}_l[n]-\hat{\boldsymbol{W}}_l^{\mathrm{H}}[n]\tilde{\boldsymbol{H}}_l[n]. \quad (13)$$

Weight update,

$$\hat{\boldsymbol{W}}_l[n] = \hat{\boldsymbol{W}}_l[n-1] + \boldsymbol{k}_l[n]\boldsymbol{E}_l^*[n],\ n\geq 1. \quad (14)$$

$\boldsymbol{k}_l[n]$ is time-variant gain vector whose definition is in Eq.(15).

$$\boldsymbol{k}_l[n]=\frac{\boldsymbol{Q}_l[n-1]\tilde{\boldsymbol{H}}_l[n]}{\lambda+\tilde{\boldsymbol{H}}_l[n]\boldsymbol{Q}_l[n-1]\tilde{\boldsymbol{H}}_l[n]},\ n\geq 1, \quad (15)$$

$$\boldsymbol{Q}_l[n]=\frac{1}{\lambda}(\boldsymbol{I}-\boldsymbol{k}_l[n]\tilde{\boldsymbol{H}}_l^{\mathrm{H}}[n])\boldsymbol{P}_l[n-1],\ n\geq 1, \quad (16)$$

here $\tilde{\boldsymbol{H}}_l[n]=\left[\tilde{H}_l[n],\tilde{H}_l[n-1],\ldots,\tilde{H}_l[N-M+1]\right]^{\mathrm{T}}$, M is the length of RLS filter. For each antenna, the relation between the vector $\tilde{\boldsymbol{H}}_l[n]$ and the scalar $\tilde{H}_l[n-i]$ is shown as below:

$$\hat{\boldsymbol{H}}_l[n]=\sum_{m=0}^{M-1}\widehat{W}_l[m]\tilde{H}_l[n-m]=\hat{\boldsymbol{W}}_l^{\mathrm{H}}[m]\ \tilde{\boldsymbol{H}}_l[n]. \quad (17)$$

where $\hat{\boldsymbol{W}}_l[m]=\left[\widehat{W}_l^*[0]\quad \widehat{W}_l^*[1]\quad \ldots\quad \widehat{W}_l^*[M-1]\right]^{\mathrm{T}}$.

Initialization

$$\hat{\boldsymbol{W}}_l[n]=[0,\ 0,\ \ldots,\ 0]^{\mathrm{T}}\ \text{or}\ \hat{\boldsymbol{W}}_l[n]=[1,\ 0,\ \ldots,\ 0]^{\mathrm{T}}. \quad (18)$$

For initialization of RLS recursion, we set

$$\boldsymbol{Q}_l[0]=(\tilde{\boldsymbol{H}}_l[0]\tilde{\boldsymbol{H}}_l^{\mathrm{H}}[0]+\delta\boldsymbol{I})^{-1}=\frac{1}{\delta}\left[\boldsymbol{I}-\frac{\tilde{\boldsymbol{H}}_l[0]\tilde{\boldsymbol{H}}_l^{\mathrm{H}}[0]}{\left\|\tilde{\boldsymbol{H}}_l[0]\right\|^2+\delta}\right], \quad (19)$$

$$\boldsymbol{k}_l[0]=\boldsymbol{Q}_l[0]\tilde{\boldsymbol{H}}_l[0]=\frac{1}{\left\|\tilde{\boldsymbol{H}}_l[0]\right\|^2+\delta}\tilde{\boldsymbol{H}}_l[0], \quad (20)$$

where δ is the regularization parameter.

For $\hat{\boldsymbol{W}}_l[n]$ of RLS filter, if the maximum delay L of actual channel is unknown in advance, $L_{\mathrm{CP}}+1$ taps filter will be applied to channel estimation.

Because we assume that the channel varies negligibly within whole OFDM symbols, we can obtain the channel parameter matrix in the time/frequency domain by performing FFT as Eq.(21) shows.

$$\hat{\boldsymbol{H}}[nN_{\mathrm{T}},k]=\frac{1}{\sqrt{N_{\mathrm{T}}}}\sum_{l=0}^{L-1}\hat{\boldsymbol{H}}_l[n]\mathrm{e}^{-\mathrm{j}2\pi kl/K}. \quad (21)$$

ADAPTIVE SCHEME

The channel estimation scenario is described below. Considering that the channel is slow fading, the channel transfer functions of data symbols, i.e., $\boldsymbol{H}$ can be expressed by the channel transfer functions of the six preamble symbols, i.e., $\boldsymbol{H}_1$, $\boldsymbol{H}_2$, $\boldsymbol{H}_3$, $\boldsymbol{H}_4$, $\boldsymbol{H}_5$ and $\boldsymbol{H}_6$ in the frequency domain. Eq.(22) is the channel transfer functions.

$$\boldsymbol{H}=\alpha_1\boldsymbol{H}_1+\alpha_2\boldsymbol{H}_2+\alpha_3\boldsymbol{H}_3+\alpha_4\boldsymbol{H}_4+\alpha_5\boldsymbol{H}_5+\alpha_6\boldsymbol{H}_6, \quad (22)$$

where

$$\alpha_1+\alpha_2+\alpha_3+\alpha_4+\alpha_5+\alpha_6=1. \quad (23)$$

Here $\alpha_j\geq 0, j=1,\ldots,6$.

There are two schemes for setting the weight values in this paper, i.e., Scheme A and Scheme B.

In Scheme A, the weight values including α_1, α_2, α_3, α_4, α_5 and α_6 can be set equal values.

$$\alpha_j=\alpha_{j+1}=1/6,\ j=1,\ldots,5. \quad (24)$$

Scheme A is not reasonable. Because the channel is varying, equal weight values can not reflect different roles of the different preamble symbols.

In Scheme B, α_1, α_2, α_3, α_4, α_5 and α_6 can be set different weight values that reflect different roles of the different preamble symbols. Scheme B outperforms Scheme A because $\boldsymbol{H}_1$, $\boldsymbol{H}_2$, $\boldsymbol{H}_3$, $\boldsymbol{H}_4$, $\boldsymbol{H}_5$ and $\boldsymbol{H}_6$ play different roles on data channel $\boldsymbol{H}$. It is obvious that $\boldsymbol{H}_1$, $\boldsymbol{H}_2$, $\boldsymbol{H}_3$, $\boldsymbol{H}_4$, $\boldsymbol{H}_5$ and $\boldsymbol{H}_6$ play more and more roles correspondingly in data channel $\boldsymbol{H}$ as the distances between $\boldsymbol{H}_i$ (i=1,2,…,6) and $\boldsymbol{H}$ becomes shorter gradually in the time domain. So it is reasonable to set the weight values as follows:

$$0\leq\alpha_j\leq\alpha_{j+1}\leq 1,\ j=1,\ldots,5. \quad (25)$$

So the optimal selection of weight values is based on the channel characteristics and experiences and needs large computer simulations to determine. Therefore we can present a suboptimal scheme as follows:

$$\begin{cases}\alpha_j=\alpha_{j-1}+q,\\ \alpha_j<1/6,\\ j=2,\ldots,6.\end{cases} \quad (26)$$

Considering α_1 playing 1/6 role on $\boldsymbol{H}$ at most, we can assume α_1=0.10 and obtain the values of α_2, α_3, α_4, α_5, α_6 and q.

As $\boldsymbol{H}$ is decided by not only the weights, i.e., α_1, α_2, α_3, α_4, α_5, α_6 but also the transfer functions, i.e., $\boldsymbol{H}_1$, $\boldsymbol{H}_2$, $\boldsymbol{H}_3$, $\boldsymbol{H}_4$, $\boldsymbol{H}_5$, $\boldsymbol{H}_6$, we can analyze the roles of the transfer functions. Since it is crucial for RLS channel estimation algorithm to select the forgetting factors to estimate the channel transfer functions of pilot symbols, there are three scenarios including RLS-C, RLS-T and RLS-A proposed in this paper.

In RLS-C scenario with conventional forgetting factor, we design the equal forgetting factor value for the whole preamble symbols. For example, RLS-C applies the parameters as follows:

$$\begin{cases}\lambda_{j+1}=\lambda_j=0.9,\ j=1,\ldots,5,\\ \delta=0.1,\end{cases} \quad (27)$$

here δ is the regularization parameter. The forgetting factor λ_j is a constant.

In RLS-T scenario with two-step forgetting factor, as Fig.2 shows, we design the low forgetting factor value for the first two preamble symbols in order to make the communication system depend on the channel information estimated from the recent correct preamble symbol. Further we design the high forgetting factor value for the later four preamble symbols in order to make the communication system depend on the statistical information of channel response, which is drawn by prior estimation. For example, RLS-T employs the parameters as follows:

$$\begin{cases}\lambda_{j+1}=\lambda_j=0.6,\ j=1,\\ \lambda_{j+1}=\lambda_j=0.9,\ j=3,\ldots,5,\\ \delta=0.1,\end{cases} \quad (28)$$

here the low value forgetting factor of the first two preamble symbols is a constant, i.e., λ=0.60 and the high value forgetting factor of the later four preamble symbols is another constant, λ=0.90.

In RLS-A scenario with adaptive forgetting factor, we design variant forgetting factor value for all preamble symbols and apply an initial forgetting factor value to the first preamble symbol, and make all forgetting factors have a fixed positive learning rate, with the remainder preamble symbols all having different forgetting factors. Selection of the initial forgetting factor value and the learning rate is based on the factors such as channel characteristics, algorithm requirements and computer simulation results. For example, RLS-A applies the parameters as follows:

$$\begin{cases} \lambda_{j+1}=\lambda_j+v,\ j=1,\dots,5, \\ \lambda_1=0.9, \\ v=0.01, \\ \delta=0.1, \end{cases} \tag{29}$$

here v is the learning rate.

SIMULATIONS

The parameters of the simulation system are shown in Table 1. And the simulation channel is the Jakes model shown in Table 2. We apply 16 taps to capture channel parameters in the time/delay domain. Our space-time coding scenario is a rate 1/2 space-time orthogonal block code as Alamouti designed. And the receiver uses the ML algorithm to detect the space-time code. After computer simulations, some curves of convergence, MSE, BER and robustness were obtained. Thus we can compare and analyze the RLS algorithms, i.e., RLS-C, RLS-T and RLS-A.

Table 1 Simulation parameters

Simulation parameters	Parameters values
MIMO scheme	4×4 antenna pairs
Carrier frequency	4 GHz
Bandwidth	6 MHz
Subcarrier number	64
Subcarrier separation	100 kHz
Length of CP	15
Modulation scheme	QPSK
Space-time coder	Alamouti STBC, 1/2 rate
Space-time decoder	ML
RLS adaptive scheme	Scheme B
RLS forgetting factor	RLS-C, RLS-T, RLS-A
RLS filter length	M=4
LMS step parameter	u=0.03
LMS filter length	M=4

Table 2 Channel model

Channel type	Path No.	Delay (ns)	Power delay profile (dB)	Mobile velocity (km/h)
ITU Jakes channel	1	0	–1.78	
	2	260	0.00	
	3	520	–7.47	3
	4	780	–10.00	
	5	1040	–12.62	

Computational complexity

Because the LS algorithm is only a part of the LMS or RLS algorithm, its computational complexity is lowest. The computational complexity of the LMS algorithm is lower than the RLS algorithm. If the computational complexity order of the LMS algorithm is $\gamma[M(L_{CP}+1)]$, the computational complexity order of the RLS algorithm will be $\gamma[M^2(L_{CP}+1)]$.

MSE

Fig.5 shows very obviously that MSE performances are in the order RLS-A>RLS-T>RLS-C although RLS-C can very easily select forgetting factor values. Therefore, selection of the adaptive or two-step forgetting factor should be based on channel characteristics, experimentations and experiences. Fig.5 also shows that RLS-A, RLS-T and RLS-C have the same computational complexity and better MSE performance than LMS and LS algorithms.

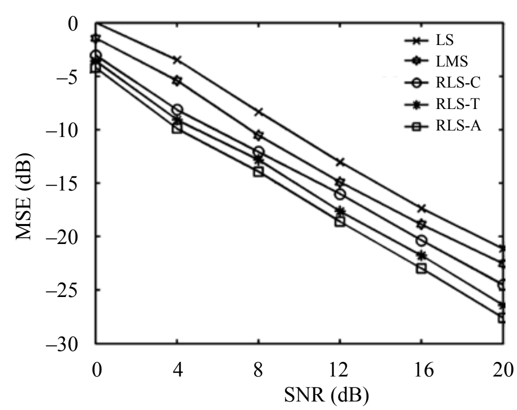

Fig.5 MSE versus SNR

BER

Fig.6 shows that RLS-A outperforms RLS-T and RLS-C in terms of BER performance and that the three RLS algorithms have better BER performance than LMS or LS algorithm. So it is preferable to employ RLS algorithm instead of LMS or LS algorithm to track and estimate channel characteristics. However, the relatively low computational complexity of LMS and LS is probably a trade-off between good estimation accuracy and low computational complexity.

Robustness

Although RLS-A, RLS-T and RLS-C have

nearly no enormous difference in robustness as Fig.7 shows, we found that RLS-A is less sensitive than RLS-T and RLS-C to the increase of the maximum Doppler frequency on condition of SNR=0 dB, and that among the three RLS algorithms RLS-C is most sensitive to maximum Doppler frequency. RLS-A, RLS-T and RLS-C have obviously better performance than LMS and LS algorithm on condition of different maximum Doppler frequencies. As a result, compared to RLS-T and RLS-C, the RLS-A algorithm performs more efficiently in high fading channels.

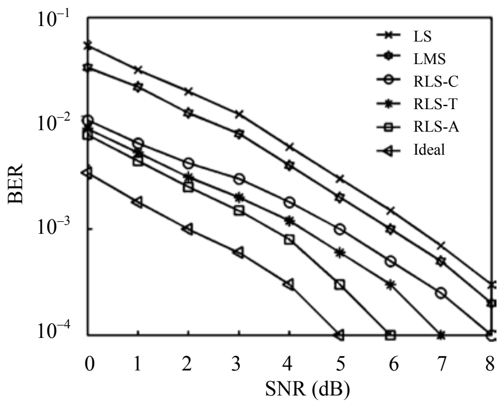

Fig.6 BER versus SNR

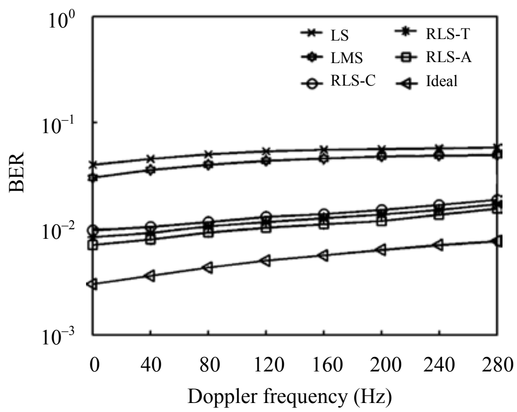

Fig.7 BER versus maximum Doppler frequency

Adaptive scheme comparison

As stated above, Scheme A has equal weight values. RLS-C′, RLS-T′ and RLS-A′ represent Scheme A's three RLS algorithms as Fig.8 shows. Scheme B obviously outperforms Scheme A in terms of BER performance. So Scheme B is designed specially to reflect channel characteristics.

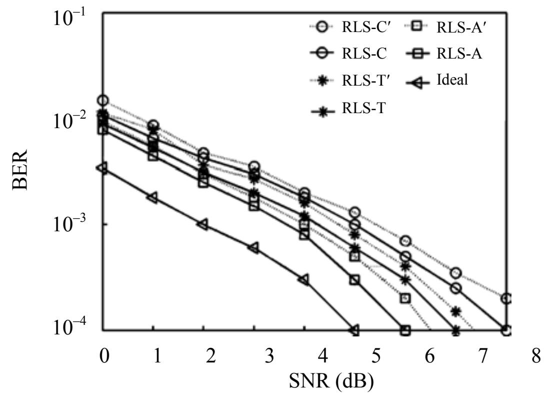

Fig.8 Scheme A versus Scheme B

CONCLUSION

This paper demonstrates a preamble-aided RLS channel estimation method with adaptive forgetting factor in space-time coded MIMO-OFDM systems. The method exploits preamble symbols to obtain channel parameters based on LS algorithm without noise effect, and obtain channel parameters by applying RLS algorithm. Simulation results proved that the RLS algorithm with adaptive forgetting factor outperforms those with two-step and conventional forgetting factors in terms of estimation accuracy and robustness performances in frequency-selective fading channels.

So we can select a reasonable forgetting factor scenario in space-time coded MIMO-OFDM systems according to different channel characteristics. Furthermore, when the channel conditions are too bad or the forgetting factor is not selected reasonably, the RLS algorithm with conventional forgetting factor or two-step one will be unstable or unrobust. Therefore, the RLS-A channel estimation method proposed in this paper can overcome disadvantages caused by bad propagation environments by means of adjusting forgetting factor value adaptively in space-timed coded MIMO-OFDM systems. As a result, preamble-aided RLS channel estimation with adaptive forgetting factor can be applied more efficiently to the MIMO-OFDM systems, such as B3G of cellular systems, 802.11n WLAN, HIPERMAN, etc.

References

Alamouti, S.M., 1998. A simple transmit diversity technique for wireless communications. *IEEE Journal Select. Areas*

Commun., **16**(8):1451-1458. [doi:10.1109/49.730453]

Barhumi, I., Leus, G., Moon, M., 2003. Optimal training design for MIMO-OFDM systems in mobile wireless channels. *IEEE Trans. on Signal Processing*, **51**(6):1615-1624. [doi:10.1109/TSP.2003.811243]

Benjebbour, A., Yoshida, S., 2003. Simplified Weights Update Algorithm for Semi-adaptive Ordered Successive Detection in MIMO Wireless Systems. The 57th IEEE VTC, **2**:989-993.

Chen, Y.C., Su, Y.T., 2004. MIMO Channel Estimation in Spatially Correlated Environments. 15th IEEE International Symposium on Personal, Indoor and Mobile Radio Communications, **1**:498-502.

Chueh, J., Hayoung, Y., Vucetic, B., 2004. An Iterative Adaptive Receiver Design Using the Multi-layer RLS Algorithm with Applications to Layered Space Time Coded Systems. 15th IEEE International Symposium on Personal, Indoor and Mobile Radio Communications, **2**:1342-1346.

Coon, J., Siew, J., Beach, M., Nix, A., Armour, S., McGeehan, J., 2003. A Comparison of MIMO-OFDM and MIMO-SCFDE in WLAN Environments. Global Telecommunications Conference, **6**:3296-3301.

Cozzo, C., Hughes, B.L., 2003. Joint channel estimation and data detection in space-time communications. *IEEE Trans. on Communications*, **51**(8):1266-1270. [doi:10.1109/TCOMM.2003.815062]

Fan, L.H., Hoshyar, R., Fazel, S., Roberts, R., Politis, C., Tafazolli, R., 2004. New MAC Design to Accommodate Joint Detection Techniques in a MIMO-OFDM-based HIPERMAN System. The 13th IEEE Workshop on Local and Metropolitan Area Networks, LANMAN, p.223-228.

Foschini, G.J., Gans, M.J., 1998. On limits of wireless communications in a fading environment when using multiple antennas. *Wireless Pers. Commun.*, **6**(3):311-335. [doi:10.1023/A:1008889222784]

Haykin, S., 2003. Adaptive Filter Theory. 4th Edition. Publishing House of Electronics Industry, Beijing.

Khuabut, P., Kunarutanapruk, S., Transongcharoen, P., Jitapunkul, S., 2004a. RLS Channel Estimation with Forgetting Factor Adaptation for the Downlink of MC-CDMA System. Proceeding of 2004 IEEE International Conference on Networking, Sensing and Control: ICNSC2004, Taipei, Taiwan, p.1160-1164.

Khuabut, P., Kunaruttanapruk, S., Transongcharoen, P., Jitap-unkul, S., 2004b. On the Improvement of RLS Channel Estimation in Forward Link of MC-CDMA System. IEEE CCECE 2004-CCEGI, Niagara Falls.

Li, Y., 2002. Simplified channel estimation for OFDM systems with multiple transmit antennas. *IEEE Trans. Wireless Commun.*, **1**(1):67-75. [doi:10.1109/7693.975446]

Liu, H., Xu, G.H., 1995. Multiuser Blind Channel Estimation and Spatial Channel Pre-equalization. ICASSP-95, **3**:1756-1759.

Schafhuber, D., Matz, G., Hlawatsch, F., 2003a. Adaptive Wiener Filters for Time-varying Channel Estimation in Wireless OFDM Systems. IEEE ICASSP2003, **4**: 688-691.

Schafhuber, D., Rupp, M., Matz, G., Hlawatsch, F., 2003b. Adaptive Identification and Tracking of Doubly Selective Fading Channels for Wireless MIMO-OFDM Systems. IEEE SPAWC-03, Rome, Italy.

Stuber, G.L., Barry, J.R., McLaughlin, S.W., Li, Y., Ingram, M.A., Pratt, T.G., 2004. Broadband MIMO-OFDM wireless communications. *Proceedings of the IEEE*, **92**(2):271-294. [doi:10.1109/JPROC.2003.821912]

Sun, Y., Yee, M., Sandell, M., 2003. Iterative Channel Estimation with MIMO MMSE-turbo Equalization. 2003 IEEE 58th Vehicular Technology Conference, **2**: 1278-1282.

Suthaharan, S., Nallanathan, A., Kannan, B., 2003. A Computationally Efficient Channel Estimation with Signal Detection for MIMO-OFDM Systems. 14th IEEE Proceedings on Personal, Indoor and Mobile Radio Communications, **2**:1245-1249.

Tarokh, V., Seshadri, N., Calderbank, A.R., 1998. Space-time codes for high data rate wireless communication: performance criterion and code construction. *IEEE Trans. Inform. Theory*, **44**(2):744-765. [doi:10.1109/18.661517]

Tarokh, V., Jafarkhani, H., Calderbank, A.R., 1999. Space-time block codes from orthogonal designs. *IEEE Trans. Inform. Theory*, **45**(5):1456-1467. [doi:10.1109/18.771146]

Wong, K.T., Cheung, S.K., 2004. Fractional self-decorrelation to enhance single-user-type DS-CDMA detectors output SINR in blind space-time RAKE receivers. *IEEE Communications Letters*, **8**(6):336-338. [doi:10.1109/LCOMM.2004.827424]