

# MDL-based root-mean-squared delay spread estimation for MIMO OFDM systems

Xu Xiaodong    Jing Ya    You Xiaohu

(National Mobile Communications Research Laboratory, Southeast University, Nanjing 210096, China)

**Abstract:** This paper studies the application of minimum description length (MDL) criterion for estimating root-mean-squared (RMS) delay spread (RDS) for MIMO OFDM systems. The analytic relationship between the powers and the correlation matrix of multipath components establishes the feasibility of the application of the MDL criterion to RDS estimation. The estimator presents both the estimate of instantaneous RDS and the estimates of noise variance, channel power and SNR of current channel with low computational complexity. Given the powers of the estimated multipath components, the MDL criterion is adopted to acquire the number of paths and the time delays of each path of current channel without making eigendecomposition of the correlation matrix normally required by MDL criterion, following which the noise variance and the power of each path can be estimated. The power delay profile (PDP) and RDS of the current channel are achieved. Simulation results show that the proposed estimator is insensitive to variance of SNR and robust against frequency-selectivity.

**Key words:** multi-input multi-output orthogonal frequency division multiplexing (MIMO OFDM); root-mean-squared (RMS) delay spread (RDS); frequency-selectivity; noise variance; minimum description length criterion

The combination of multi-input multi-output (MIMO) signal processing with orthogonal frequency division multiplexing (OFDM) is regarded as a promising solution for enhancing the data rates of next-generation wireless communication systems operating in frequency selective fading environments. Though OFDM is effective for combating frequency selectivity, the information about frequency-selectivity fading across OFDM subcarriers, and the corresponding time domain root-mean-squared (RMS) delay spread (RDS) can be very useful for improving the performance of OFDM-based wireless communication systems. In the pilot-symbol-aided channel estimation<sup>[1]</sup>, according to the sampling theorem<sup>[2]</sup>, frequency domain interpolation requires the knowledge of frequency-selectivity of the channel, which is normally *a priori* unknown. In general cases, the channel interpolator depends on a likely value (fixed) thus resulting in suboptimal channel interpolation. By detecting the RDS of the channel and hence the frequency-selectivity of the channel, adaptive pilot interpolation techniques can be adopted to improve the system performance. In adaptive modulation techniques, the subband<sup>[3]</sup> is introduced to simplify the task of modem mode signaling and render

feasible the employment of alternative blind detection mechanisms, which are also dependent on such dispersion information. In general, in OFDM-based wireless communication systems, knowledge about RDS of the current channel is helpful for providing better overall system performance, improved radio coverage, and higher data rates.

Frequency domain channel models for the frequency selective radio channel and level crossing statistics in the frequency domain have been studied in Ref. [4], based on which an RMS delay spread estimation technique is proposed<sup>[5]</sup>; however, Ref. [6] shows that LCR measurement may be severely deteriorated due to its sensitivity to noise. Both global (long term) and local (instantaneous) estimation of RDS are investigated in Ref. [7]. According to the analytic relationship between RDS of exponentially decaying power delay profile (PDP) and channel frequency correlation, averaged RDS of the channel is obtained directly from the frequency correlation estimation. Depending on the relationship between signal-to-estimation-error ratio (SER) and SNR given in Ref. [8], the instantaneous channel impulse response (CIR) and instantaneous RDS are estimated through the known SER.

This paper develops an RDS estimator for MIMO OFDM systems through well-known minimum descriptive length (MDL) criterion presented in Ref. [9] based on the analytic relationship between the powers and the correlation matrix of multipath

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**Biographies:** Xu Xiaodong (1975—), male, graduate; You Xiaohu (corresponding author), male, doctor, professor, xhyu@seu.edu.cn.

components. The estimator presents both the estimate of instantaneous RDS and the estimates of noise variance, channel power and SNR of current channel. The noise power can be correctly estimated in our proposed estimator and the influence of noise on RDS estimation can be effectively eliminated from the estimator, therefore, the estimator is insensitive to the variance of SNR and it can work well over a wide range of SNR values. Eigendecomposition of the correlation matrix of the observation vector, normally required by MDL criterion, is avoided in our proposed estimator by noting the analytic relationship between the powers and the correlation matrix of estimated multipath components, which greatly reduces the **computational complexity in the estimator**.

## 1 System Model

The basic baseband-equivalent MIMO OFDM system is shown in Fig.1. Consider an MIMO OFDM system with  $N_t$  transmit antennas and  $N_r$  receive antennas. It is assumed that the channels between different transmit-receive antenna couples are independent and identically distributed multipath Rayleigh fading channels. The multipath Rayleigh fading channel between the  $i$ -th transmit antenna and the  $j$ -th receive antenna can be characterized by

$$h_{j,i}(t, \tau) = \sum_{l=0}^{v-1} h_{j,i}(t, l) \delta(\tau - \tau_l) \quad (1)$$

where  $\{h_{j,i}(t, l)\}$  are complex gains of each path between the  $i$ -th transmit antenna and the  $j$ -th receive antenna at time  $t$ ,  $\tau_l$  is the  $l$ -th path time delay,  $v$  is the number of paths.  $\{h_{j,i}(t, l)\}$  are wide-sense stationary (WSS) narrow-band complex Gaussian processes and the different path gains are mutually uncorrelated, where the sum of average energy of the total channel is normalized to one.

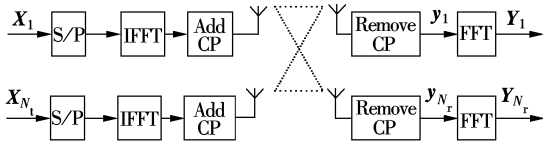


Fig.1 MIMO-OFDM system

We denote  $K$  as the total number of subcarriers, or the FFT size,  $V$  as the relative maximum time delay spread ( $\tau_{v-1}$ ) normalized by system sampling interval. At the receiver side, we add the assumptions that the guard interval duration  $L$  is longer than the channel maximum excess delay. As no *a priori* knowledge is available about true values of  $V$  and  $v$ , the number of multipath components is normally assumed to be equal to  $L$ , we also follow such assumption in this paper.

The OFDM symbol index will be omitted for convenience. The received vector after removing the cyclic prefix can be expressed as

$$\mathbf{y} = \begin{bmatrix} \mathbf{y}_1 \\ \vdots \\ \mathbf{y}_{N_r} \end{bmatrix} = \begin{bmatrix} \mathbf{h}_{1,1}^{\text{circ}} & \cdots & \mathbf{h}_{1,N_t}^{\text{circ}} \\ \vdots & & \vdots \\ \mathbf{h}_{N_r,1}^{\text{circ}} & \cdots & \mathbf{h}_{N_r,N_t}^{\text{circ}} \end{bmatrix} \cdot \left( \mathbf{I}_{N_t} \otimes \left( \frac{1}{\sqrt{K}} \mathbf{F} \right)^H \right) \begin{bmatrix} \mathbf{X}_1 \\ \vdots \\ \mathbf{X}_{N_t} \end{bmatrix} + \mathbf{e} \quad (2)$$

where  $\otimes$  denotes the Kronecker product;  $\mathbf{X}_i$  is the  $K \times 1$  transmitted vector at the  $i$ -th transmit antenna;  $\mathbf{y}_j$  is the  $K \times 1$  received vector at the  $j$ -th receive antenna after removing the cyclic prefix;  $\mathbf{h}_{j,i}^{\text{circ}}$  is a circular matrix with its first column given by  $[\bar{\mathbf{h}}_{j,i}^T, \mathbf{0}_{1 \times (K-L)}^T]^T$ ,  $\bar{\mathbf{h}}_{j,i} = [\bar{h}_{j,i}(0), \dots, \bar{h}_{j,i}(L-1)]^T$ ,  $\{\bar{h}_{j,i}(k)\}$  ( $k = 0, 1, \dots, L-1$ ) are mutually independent multipath components between the  $i$ -th transmit antenna and the  $j$ -th receive antenna, and  $\{\bar{\mathbf{h}}_{j,i}\}$  ( $j = 1, 2, \dots, N_r; i = 1, 2, \dots, N_t$ ) are independent and identically distributed among different antenna couples according to former assumption;  $\mathbf{e}$  is a  $(KN_r) \times 1$  additive (complex) Gaussian noise vector assumed to be independent and identically distributed with zero-mean and variation  $\sigma_N^2$ ;  $\mathbf{F}$  is the Fourier transform matrix and can be written as

$$\mathbf{F} \triangleq \begin{bmatrix} 1 & 1 & \cdots & 1 \\ 1 & e^{-j2\pi/K} & \cdots & e^{-j2\pi(K-1)/K} \\ \vdots & \vdots & & \vdots \\ 1 & e^{-j2\pi(K-1)/K} & \cdots & e^{-j2\pi(K-1)(K-1)/K} \end{bmatrix} \quad (3)$$

At the receiver side, with the assumption that the guard interval duration is longer than the channel maximum excess delay, we obtain the following equation by taking the FFT of  $\mathbf{y}$ :

$$\mathbf{Y} = \begin{bmatrix} \text{diag}\{\mathbf{H}_{1,1}\} & \cdots & \text{diag}\{\mathbf{H}_{1,N_t}\} \\ \vdots & & \vdots \\ \text{diag}\{\mathbf{H}_{N_r,1}\} & \cdots & \text{diag}\{\mathbf{H}_{N_r,N_t}\} \end{bmatrix} \begin{bmatrix} \mathbf{X}_1 \\ \vdots \\ \mathbf{X}_{N_t} \end{bmatrix} + \left( \mathbf{I}_{N_r} \otimes \left( \frac{1}{\sqrt{K}} \mathbf{F} \right) \right) \mathbf{e} \quad (4)$$

where  $\mathbf{H}_{j,i}$  is the channel frequency response between the  $i$ -th transmit antenna and the  $j$ -th receive antenna, and  $\text{diag}\{\mathbf{H}_{ji}\}$  represents the diagonal matrix transformed from the column vector  $\mathbf{H}_{ji}$ . For the convenience of description in channel estimation, Eq. (4) can be rewritten as

$$\begin{aligned} \mathbf{Y} &= \mathbf{I}_{N_r} \otimes [\text{diag}\{\mathbf{X}_1\}, \text{diag}\{\mathbf{X}_2\}, \dots, \text{diag}\{\mathbf{X}_{N_t}\}] \cdot \\ &\quad [\mathbf{H}_{1,1}, \mathbf{H}_{1,2}, \dots, \mathbf{H}_{N_r,N_t}]^T + \left( \mathbf{I}_{N_r} \otimes \left( \frac{1}{\sqrt{K}} \mathbf{F} \right) \right) \mathbf{e} = \\ &\quad \mathbf{I}_{N_r} \otimes [\text{diag}\{\mathbf{X}_1\} \mathbf{F}_\Omega, \text{diag}\{\mathbf{X}_2\} \mathbf{F}_\Omega, \dots, \\ &\quad \text{diag}\{\mathbf{X}_{N_t}\} \mathbf{F}_\Omega] \cdot [\mathbf{h}_{1,1}, \mathbf{h}_{1,2}, \dots, \mathbf{h}_{N_r,N_t}]^T + \end{aligned}$$

$$\left( \mathbf{I}_{N_r} \otimes \left( \frac{1}{\sqrt{K}} \mathbf{F} \right) \right) \mathbf{e} \quad (5)$$

where  $\Omega$  is  $\{0, 1, \dots, L-1\}$  and  $\mathbf{F}_\Omega$  is a transformation matrix consisting of the first  $L$  columns of  $\mathbf{F}$  according to set  $\Omega$ .

## 2 LS Channel Estimation

In order to minimize the mean square error (MSE), we adopt the optimal training sequences proposed in Ref. [10] as pilots for channel estimation in MIMO OFDM systems. Let  $\mathbf{Y} = \{k_1, k_2, \dots, k_M\}$  be the set of pilot tones, which have unity amplitude and should be equally spaced in each OFDM block. Let  $\mathbf{P}_i$  denote the training sequence transmitted by the  $i$ -th antenna. Then, from Eq. (5) we can obtain the received vector of pilot tones as

$$\mathbf{Y}_p = \mathbf{\Psi} \mathbf{h} + \mathbf{\Xi}_p \quad (6)$$

where

$$\mathbf{h} = [\mathbf{h}_{1,1}^T, \mathbf{h}_{1,2}^T, \dots, \mathbf{h}_{N_r, N_t}^T]^T \quad (7)$$

$$\mathbf{\Xi}_p = \left( \mathbf{I}_{N_r} \otimes \frac{1}{\sqrt{K}} \mathbf{F}_Y \right) \mathbf{e} \quad (8)$$

$$\mathbf{\Psi} = \mathbf{I}_{N_r} \otimes [\text{diag}\{\mathbf{P}_1\} \mathbf{F}_{Y,\Omega}, \dots, \text{diag}\{\mathbf{P}_{N_t}\} \mathbf{F}_{Y,\Omega}] \quad (9)$$

where  $\mathbf{F}_Y$  in Eq.(8) is the corresponding rows of  $\mathbf{F}$  according to set  $Y$  and  $\mathbf{F}_{Y,\Omega}$  in Eq. (9) is the corresponding columns of  $\mathbf{F}_Y$  according to  $\Omega$ , respectively. From Eq.(6) we can get the least square (LS) estimate of  $\mathbf{h}$  as

$$\hat{\mathbf{h}} = \mathbf{\Psi}^+ \mathbf{Y}_p = (\mathbf{\Psi}^H \mathbf{\Psi})^{-1} \mathbf{\Psi}^H \mathbf{Y}_p \quad (10)$$

where the pseudo-inverse matrix  $\mathbf{\Psi}^+ = (\mathbf{\Psi}^H \mathbf{\Psi})^{-1} \mathbf{\Psi}^H$ , and  $(\cdot)^+$  denotes the pseudo-inverse operator. Satisfying the condition of  $M \geq N_t L$ , Eq.(10) has a unique solution. The optimal training sequences proposed in Ref. [10] makes  $\mathbf{\Psi}^H \mathbf{\Psi}$  a diagonal matrix, which allows us to rewrite Eq.(10) in a simplified form:

$$\hat{\mathbf{h}} = \frac{1}{M} \mathbf{\Psi}^H \mathbf{Y}_p = \mathbf{h} + \frac{1}{M} \mathbf{\Psi}^H \mathbf{\Xi}_p \quad (11)$$

which can also be denoted as

$$\begin{aligned} & [\hat{\mathbf{h}}_{1,1}^T, \hat{\mathbf{h}}_{1,2}^T, \dots, \hat{\mathbf{h}}_{N_r, N_t}^T]^T = \\ & [\mathbf{h}_{1,1}^T, \mathbf{h}_{1,2}^T, \dots, \mathbf{h}_{N_r, N_t}^T]^T + \frac{1}{M} [(\mathbf{\Psi}^H \mathbf{\Xi}_p)_{1,1}^T, \\ & (\mathbf{\Psi}^H \mathbf{\Xi}_p)_{1,2}^T, \dots, (\mathbf{\Psi}^H \mathbf{\Xi}_p)_{N_r, N_t}^T]^T \end{aligned} \quad (12)$$

The RDS between any receive-transmit antenna couple can be obtained with the aid of the estimated multipath components between corresponding antenna couple. Without loss of generality, the estimated multipath components  $\{\hat{h}_{j,i}(k)\} (k=0, 1, \dots, L-1)$  will be used for describing the estimation

procedure of RDS.

## 3 MDL-Based RDS Estimation

From Eq.(11), the correlation matrix  $\mathbf{E}\{\hat{\mathbf{h}}\hat{\mathbf{h}}^H\}$  can be denoted as

$$\mathbf{E}\{\hat{\mathbf{h}}\hat{\mathbf{h}}^H\} = \mathbf{E}\{\mathbf{h}\mathbf{h}^H\} + \frac{1}{M} \sigma_N^2 \mathbf{I}_{L N_r N_t \times L N_r N_t} \quad (13)$$

Corresponding to Eq.(13), the correlation matrix of  $\hat{\mathbf{h}}_{j,i}$  can be denoted as

$$\mathbf{R} = \mathbf{\Phi} + \frac{1}{M} \sigma_N^2 \mathbf{I}_{L \times L} \quad (14)$$

where  $\mathbf{R} = \mathbf{E}\{\hat{\mathbf{h}}_{j,i} \hat{\mathbf{h}}_{j,i}^H\}$ ,  $\mathbf{\Phi} = \mathbf{E}\{\mathbf{h}_{j,i} \mathbf{h}_{j,i}^H\}$ . Since only  $v$  of the  $L$  components of  $\mathbf{h}_{j,i}$  are nonzero,  $\mathbf{\Phi}$  is a singular matrix and the rank of matrix  $\mathbf{\Phi}$  is  $v$ , or equivalently, the smallest  $L-v$  eigenvalues of  $\mathbf{\Phi}$  are equal to zero. Denoting the eigenvalues of  $\mathbf{R}$  by  $\lambda_1 \geq \lambda_2 \geq \dots \geq \lambda_L$ , it follows that the smallest  $L-v$

eigenvalues of  $\mathbf{R}$  are all equal to  $\frac{\sigma_N^2}{M}$ , i.e.,

$$\lambda_{v+1} = \lambda_{v+2} = \dots = \lambda_L = \frac{1}{M} \sigma_N^2 \quad (15)$$

and we also have

$$\text{tr}(\mathbf{R}) = \sigma_s^2 + \frac{L}{M} \sigma_N^2 = \sum_{i=1}^L \lambda_i \quad (16)$$

where  $\sigma_s^2 = \mathbf{E}\{\mathbf{h}_{j,i}^H \mathbf{h}_{j,i}\}$  denotes the channel power. This implies that the observation space can be partitioned into a signal subspace and a noise subspace. Now if we get the estimate of the channel correlation matrix  $\mathbf{R}$  and the multipath number  $v$  (also the dimension of the signal subspace), the noise variance can be derived.

Whereas, according to the above independence assumption, the  $L$  components of  $\mathbf{h}_{j,i}$  are mutually independent,  $\mathbf{\Phi}$  will be a diagonal matrix. The diagonal elements of  $\mathbf{\Phi}$  are just equal to  $\mathbf{E}\{\mathbf{h}_{j,i} \odot \mathbf{h}_{j,i}^*\}$  ( $\odot$  denotes array multiplication), which are the multipath component powers. From Eq.(14),  $\mathbf{R}$  is also a diagonal matrix. In addition,  $\frac{1}{M^2} \mathbf{E}\{(\mathbf{\Psi}^H \mathbf{\Xi}_p)_{j,i}\}$

$\odot (\mathbf{\Psi}^T \mathbf{\Xi}_p^*)_{j,i}\} = \frac{1}{M} \sigma_N^2 \mathbf{I}_{L \times 1}$ , where  $\mathbf{I}_{L \times 1}$  denotes an  $L \times 1$  vector of all ones, therefore,  $\mathbf{E}\{\hat{\mathbf{h}}_{j,i} \odot \hat{\mathbf{h}}_{j,i}^*\}$ , the powers of the estimated multipath components equal the diagonal elements of correlation matrix  $\mathbf{R}$ , which are also the eigenvalues of  $\mathbf{R}$ . Given  $\mathbf{E}\{\hat{\mathbf{h}}_{j,i} \odot \hat{\mathbf{h}}_{j,i}^*\}$ , the number of paths can be detected through the MDL criterion, based on which the noise variance as well as the power and the time delay of each path can be estimated. Eigendecomposition of the correlation matrix  $\mathbf{R}$  normally required by MDL criterion is

avoided and a novel RDS estimator is proposed. The estimator will present the estimates of the noise variance, channel power, SNR, PDP and RDS.

We estimate  $E\{\hat{\mathbf{h}}_{j,i} \odot \hat{\mathbf{h}}_{j,i}^*\}$  by moving averaging the powers of the most recent  $N$  observation vectors. Let  $m$  denote the  $m$ -th OFDM symbol, we have

$$\hat{\mathbf{R}}(m) = \frac{1}{N} \sum_{i=m-N+1}^m \hat{\mathbf{h}}_{j,i} \odot \hat{\mathbf{h}}_{j,i}^* \quad (17)$$

$\hat{\mathbf{R}}(m)$  are sorted in descending order, and the original positions of each multipath component are recorded as  $\{i_1, i_2, \dots, i_L\}^T$ . Let  $\{\hat{\lambda}_1, \hat{\lambda}_2, \dots, \hat{\lambda}_L\}^T$  denote the sorted vector  $\hat{\mathbf{R}}(m)$ , so there exists  $\hat{\lambda}_1 \geq \hat{\lambda}_2 \geq \dots \geq \hat{\lambda}_L$ , and  $\{i_k\}$  ( $k = 1, 2, \dots, L$ ) reflect the actual positions of  $\{\hat{\lambda}_k\}$  ( $k = 1, 2, \dots, L$ ) in  $\hat{\mathbf{R}}(m)$ . The number of multipath  $v$  can be detected through the well-known MDL criterion presented in Ref. [9]. The objective function of the MDL criterion is defined as

$$\text{MDL}(k) = -N(L-k) \log \left( \frac{\prod_{i=k+1}^L \hat{\lambda}_i^{\frac{1}{L-k}}}{\frac{1}{L-k} \sum_{i=k+1}^L \hat{\lambda}_i} \right) + \frac{1}{2} k(2L-k) \log(N) \quad (18)$$

The number of multipath  $v$  is estimated as

$$\hat{v} = \arg \min_k \text{MDL}(k) \quad k \in \{0, 1, \dots, L-1\} \quad (19)$$

From Eqs.(15) to (19), we now obtain our noise variance and SNR estimator, following which we obtain the RDS estimate:

1) Estimate the number of paths of current channel  $\hat{v}$  by using Eqs.(18) and (19).

2) According to Eqs.(15) and (16), estimate the noise power as

$$\hat{\sigma}_N^2 = \frac{M}{L - \hat{v}} \sum_{i=\hat{v}+1}^L \hat{\lambda}_i \quad (20)$$

and channel power as

$$\hat{\sigma}_S^2 = \sum_{i=1}^{\hat{v}} \hat{\lambda}_i - \frac{\hat{v}}{M} \hat{\sigma}_N^2 \quad (21)$$

3) The estimate of the SNR is then obtained as

$$\hat{\rho} = \frac{\hat{\sigma}_S^2}{\hat{\sigma}_N^2} \quad (22)$$

4) Recalling  $\{\hat{\lambda}_1, \hat{\lambda}_2, \dots, \hat{\lambda}_L\}^T$ ,  $\{i_1, i_2, \dots, i_L\}^T$ , the estimated  $\hat{v}$  and  $\hat{\sigma}_N^2$ , the root-mean-squared delay spread is denoted as

$$\hat{\tau}_{\text{RMS}} = \sqrt{\left( \frac{\sum_{i=1}^{\hat{v}} \left( \hat{\lambda}_i - \frac{\hat{\sigma}_N^2}{M} \right) i_i^2}{\sum_{i=1}^{\hat{v}} \left( \hat{\lambda}_i - \frac{\hat{\sigma}_N^2}{M} \right)} \right) - \left( \frac{\sum_{i=1}^{\hat{v}} \left( \hat{\lambda}_i - \frac{\hat{\sigma}_N^2}{M} \right) i_i}{\sum_{i=1}^{\hat{v}} \left( \hat{\lambda}_i - \frac{\hat{\sigma}_N^2}{M} \right)} \right)^2} \quad (23)$$

## 4 Simulation Results

In the simulation, we assume two different three-tap time-variant Rayleigh fading channels, whose root mean square delay spreads are two samples and ten samples, respectively, as shown in Tab. 1, and the vehicle speed is set to 120 km/h.

**Tab.1** Parameters for the multipath channels A and B

Channel	A	B
The 1st path delay/samples	0	0
The 1st path power/dB	0	0
The 2nd path delay/samples	3	17
The 2nd path power/dB	-3	-4
The 3rd path delay/samples	7	24
The 3rd path power/dB	-10	-6
RMS delay/samples	2	10

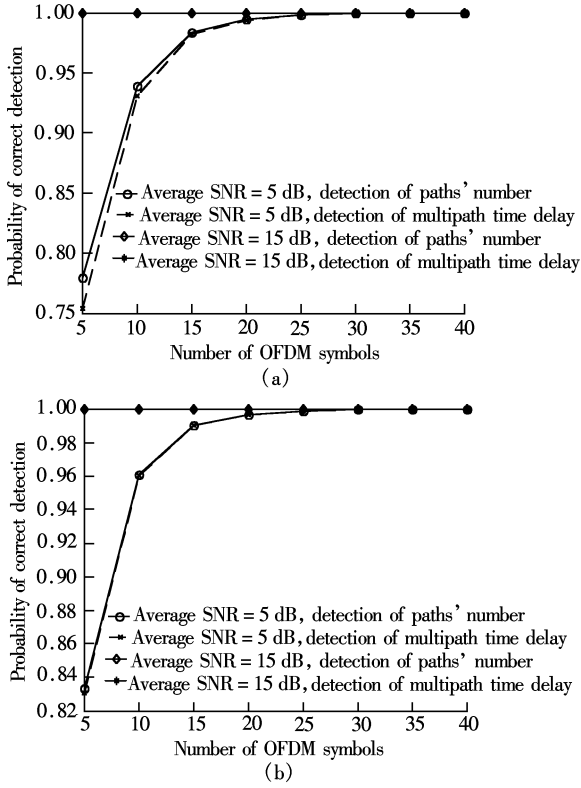
The simulation system we adopted is a  $2 \times 2$  MIMO OFDM system operating at the carrier frequency of 2.4 GHz. System bandwidth of 2 MHz, 256 subcarriers (all modulated, 64 pilot subcarriers) and 32 samples long guard interval are assumed. The root normalized MSE (RNMSE) of the estimated values following

$$e_{\text{RNMSE}} = \sqrt{\frac{1}{J} \sum_{j=1}^J \left( \frac{\hat{v}_j - v_j}{v_j} \right)^2} \quad (24)$$

is used for performance evaluation, where  $J$  is the number of independent trials,  $\hat{v}_j$  is the estimated value of a certain parameter that may be the noise variance, channel power, SNR or RDS, and  $v_j$  is the true value of the parameter for the  $j$ -th trial, where the parameter values are all linear values.

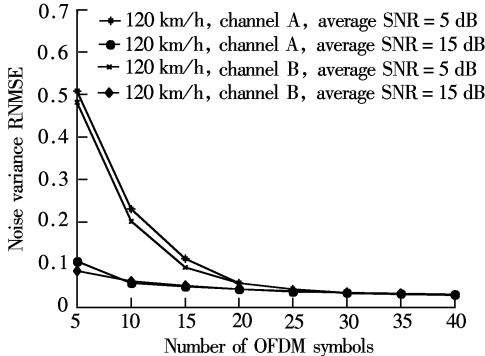
In Fig. 2, we plot the correct detection probabilities of the number of paths and the delays of all valid paths as a function of the number of the observation interval under two different frequency selective fading channels, respectively, through the MDL criterion. Fig.2(a) depicts the detection performance at average SNR of 5 and 15 dB under channel model A that represents the low frequency-selectivity fading channel, and Fig.2(b) depicts those under channel model B that represents the high frequency-selectivity fading channel. It is shown in Figs.2(a) and (b) that the detector converges fast and only a 30 OFDM symbols interval is required for convergence whether at SNR of 5 or 15 dB, for all that at high SNR values, the number of paths and the delay of each path can be detected with probability 1 during short observation intervals; however, at low SNR values, a little longer observation interval is required to achieve the approximates performance. Comparing Fig.2 (b) with Fig.2 (a), it can be found that the detection performance in a high frequency-selectivity

fading channel closely approximate that in a low frequency-selectivity fading channel, demonstrating its strong robustness against frequency selectivity.

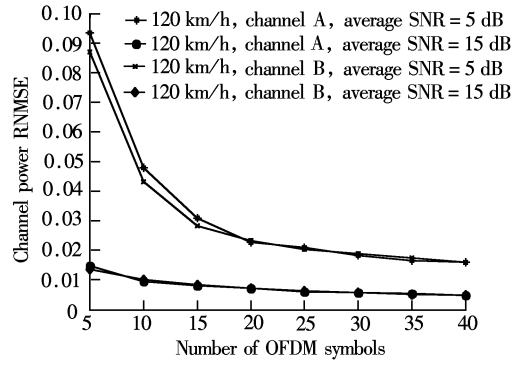


**Fig.2** Correct detection probabilities of the number of paths and delays of all paths at average SNR of 5 and 15 dB at vehicle speed of 120 km/h. (a) Channel model A; (b) Channel model B

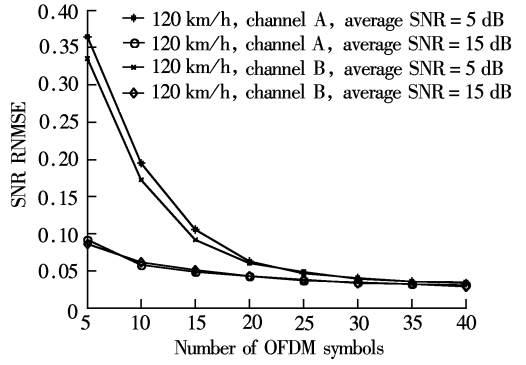
Fig.3 to Fig.6 show the estimator's RNMSE of the noise variance, channel power, SNR and RDS as a function of the number of observation OFDM symbols when average SNR equals 5 and 15 dB under channel model A and channel model B, respectively. As expected, similar characterizations of Fig. 2 will be found from Fig.3 to Fig.6. At the same SNR values, the estimator exhibits approximately identical performance under different frequency selective fading channels; however, under the same channel, the



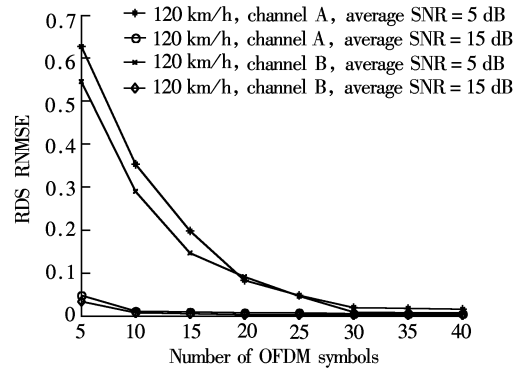
**Fig.3** Noise variance RNMSE at average SNR of 5 and 15 dB at vehicle speed of 120 km/h under channel model A and channel model B



**Fig.4** Channel power RNMSE at average SNR of 5 and 15 dB at vehicle speed of 120 km/h under channel model A and channel model B



**Fig.5** SNR RNMSE at average SNR of 5 and 15 dB at vehicle speed of 120 km/h under channel model A and channel model B



**Fig.6** RDS RNMSE at average SNR of 5 and 15 dB at vehicle speed of 120 km/h under channel model A and channel model B

estimator performs better at high SNR values than at low ones during short observation intervals. Nevertheless the performance difference between both diminishes with the increase of observation interval, which demonstrates its good performance over a wide range of SNR values and its strong robustness against frequency selectivity.

## 5 Conclusion

In this paper a novel RDS estimator for MIMO-OFDM through MDL criterion is proposed based on the analytic relationship between the eigenvalues of

the channel correlation matrix and the multipath component powers. The estimator presents not only the estimate of RDS but also the estimates of the noise variance, channel power and SNR, which will be very useful for adaptive communication systems. Eigendecomposition of correlation matrix of the observation vector, normally required by MDL criterion, is avoided in our proposed estimator by noting the analytic relationship between the estimated multipath component powers and the correlation matrix of estimated multipath components, which greatly reduces the computational complexity in the estimator. At the same time, the estimator shows strong robustness to the change of frequency selectivity and insensitivity to the variance of SNR at high mobile speed environments and it adds no limitations to the modulation type, which is desirable for environment-adaptive communication systems.

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MIMO OFDM 系统中基于 MDL 准则的  
均方根时延扩展估计

徐晓东 井 雅 尤肖虎

(东南大学移动通信国家重点实验室, 南京 210096)

摘要: 研究了 MIMO OFDM 系统中最小描述长度(MDL)准则在均方根时延扩展估计中的应用. 分析了多径分量功率和多径分量的自相关矩阵之间的关系, 使 MDL 可有效地应用于 RDS 估计. 该估计器能以较低的计算复杂度给出 RDS、噪声方差、信道功率和信噪比等多个实时信道参数的准确估计. 根据估计的多径分量功率, 利用 MDL 准则获得了当前信道环境下的路径数及它们各自时延的估计; 根据路径数得到了噪声方差和各径功率的估计, 从而得到当前信道的功率时延谱和均方根时延扩展的估计. 仿真结果表明该估计器可有效对抗信道中信噪比变化和频率选择性的影响.

关键词: MIMO OFDM; 均方根时延扩展; 频率选择性; 噪声方差; 最小描述长度准则

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