A Frequency-Domain Correlation Matrix Estimation Algorithm for MIMO-OFDM Channel Estimation

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***Abstract*— The second-order statistics of a time-domain signal are very often used in blind and semi-blind channel estimation. Considering that the received signal in MIMO-OFDM systems might be corrupted in the time-domain due to some adverse factors such as frequency offset and large peak-to-average power ratio (PAPR), an IFFT processor is required in the receiver to achieve a good-quality time-domain signal. This additional IFFT incurs extra computational complexity and probably a long time delay as well in real-time communication systems. In this paper, we propose a new algorithm for the computation of the time-domain correlation matrix directly from the received frequency-domain signal. The proposed frequency-domain correlation matrix estimation method is then used to develop a new semi-blind MIMO-OFDM channel estimation approach. A number of computer simulation based experiments are conducted, confirming the effectiveness of the proposed method.**

I. Introduction

Mobile wireless communication is coming to a new era with promises of higher data rate, integrated multimedia services and wide internet accessibility. Due to the distinct advantages of both multiple-input multiple-output (MIMO) and orthogonal frequency division multiplexing (OFDM), MIMO-OFDM as a combination of both technologies has been considered as a strong candidate for future wireless communication systems. When perfect knowledge of the wireless channel is available at the receiver, the capacity of MIMO-OFDM systems grows linearly with the number of transmit or receive antennas, whichever is less. In real wireless environments, however, the channel condition is not known. Therefore, channel estimation is required in MIMO-OFDM systems.

Broadly speaking, MIMO-OFDM channel estimation approaches can be categorized into three classes, training-based method, blind method and semi-blind approach. First, training-based methods, such as the least-square (LS) and the minimum mean square error (MMSE) methods, employ known training signals to render an accurate channel estimation. Blind MIMO-OFDM channel estimation algorithms which exploit the second-order stationary statistics or other properties, normally have a better spectral efficiency. Combining the advantages of both the training-based and the blind algorithms, a semi-blind MIMO channel estimation technique has been proposed in [1] and then extended to MIMO-OFDM systems [2], [3]. With a small number of training pilots, problems such as ambiguities and mis-convergence of the blind methods can be solved. On the other hand, the use of statistical information contained in the available data can improve the accuracy of channel estimation.

An accurate estimation of the 2nd-order statistics of the received signal in the time-domain is essential to blind and semi-blind channel estimation of MIMO systems. In MIMO-OFDM systems, however, the received time-domain signal is often corrupted due to imperfections caused by some factors such as the frequency offset and the larger peak-to-average power ratio (PAPR) etc. Considering the fact that many compensation techniques for these imperfections can only be implemented in the frequency-domain [4], [5], a good time-domain signal is not available. Thus, the existing time-domain MIMO channel estimation techniques cannot be directly applied to MIMO-OFDM systems. For them, to achieve the correlation matrix of the time-domain signal, an IFFT processor is needed in the receiver, which incurs a high computational complexity and a long time delay in real-time implementation. In this paper, we propose a new algorithm for the computation of the time-domain correlation matrix directly from the received frequency-domain signal.

Moreover, we will give an application of the new correlation matrix estimation algorithm. It is well known that the pulse-shaping filter as well as the matched filter are very commonly used in digital communication systems. Perhaps for the sake of simplicity, however, many existing channel estimation methods did not take into consideration either the effect of the pulse-shaping filter in the transmitter or the matched filter in the receiver. As such, these methods have actually been developed for the estimation of the composite channel including the pulse-shaping and matched filters. Considering that both filters are known to the receiver and the only unknown part is the pure multipath channel, ignoring their existence would lead to less accurate estimation results. By utilizing the information of both filters, some improved channel estimation algorithms have been obtained for OFDM systems [6], [7] and CDMA systems [8], [9]. In this paper, based on the proposed correlation matrix algorithm, we propose for MIMO-OFDM systems a new semi-blind channel estimation algorithm that can eliminate the effect of pulse-shaping and matched filtering, and thus give a better channel estimation performance.

Throughout the paper, we adopt the following notations:
$\dagger$ Pseudo-inverse, $\otimes$ Kronecker product,

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T Transpose, H Complex conjugate transpose,
$*$ linear convolution, $\circledast$ circular convolution,
$\| \|_F$ Frobenius norm, and
$\mathrm{vec}()$ a stacking of the columns of the involved matrix into a vector.

II. Data Model and Problem Statement

Consider a MIMO-OFDM system with N_T transmit and N_R receive antennas. The frequency-selective channel can be considered as a combination of L_c multi-paths, namely,

$$\mathbf{H}_c(t) = \sum_{l=0}^{L_c-1} \mathbf{\Gamma}_l \delta(t - t_l)$$

where t_l is the delay of the l-th path and $\mathbf{\Gamma}_l$ is an $N_R \times N_T$ attenuation matrix. Assuming the transmit pulse-shaping filter $g_t(t)$ and the receive matched filter $g_r(t)$, the entire or composite channel can be represented by an $N_R \times N_T$ matrix $\mathbf{H}(t)$, with its (i_R, i_T)-th element as

$$h_{i_R,i_T}(t) = h_{i_R,i_T,c}(t) * g_t(t) * g_r(t) \quad (1)$$

where $h_{i_R,i_T,c}(t)$ is the (i_R, i_T)-th element of $\mathbf{H}_c(t)$. Most of the existing channel estimation literatures focus on the entire discrete-time channel, the sampled version of the continuous-time channel response. Thus, the channel can be regarded as an array of L-tap FIR filters characterized by a number of $N_R \times N_T$ matrices $\mathbf{H}(n)\,(n = 0, 1, \cdots, L-1)$ with $h_{i_T,i_R}(n)$ being its (i_R, i_T)-th element. After removing cyclic prefix whose length is not less than the channel length L, the receive time-domain signal at the i_R-th antenna can be written as:

$$y_{i_R}(n) = \sum_{i_T=1}^{N_T} h_{i_R,i_T}(n) \circledast x_{i_T}(n) + v_{i_R}(n) \quad (2)$$

where $x_{i_T}(n)$ is the transmit time-domain signal at the i_T-th antenna and the noise $v_{i_R}(n)$ is a spatio-temporally uncorrelated noise with zero mean and variance δ_v^2.

An accurate estimation of the 2nd-order statistics of the received signal in the time-domain, $y_{i_R}(n)$, is essential to blind and semi-blind MIMO channel estimation algorithms as that proposed in [1]. By letting

$$\mathbf{y}(n) \triangleq [y_1(n), \cdots, y_{N_R}(n)]^T,$$
$$\mathbf{y}_{P+1}(n) \triangleq \left[\mathbf{y}^T(n), \mathbf{y}^T(n-1), \cdots, \mathbf{y}^T(n-P)\right]^T,$$

the 2nd-order statistics in terms of the correlation matrix $\mathbf{R}_T$ of $\mathbf{y}_{P+1}$ can be estimated by

$$\hat{\mathbf{R}}_\mathrm{T} = \frac{1}{K}\sum_{n=0}^{K-1} \mathbf{y}_{P+1}(n)\,\mathbf{y}_{P+1}^H(n). \quad (3)$$

Note that $\mathbf{y}_{P+1}(n)$ for $n = 0, 1, \cdots, P-1$, can be obtained using $\mathbf{y}(n-j) \triangleq \mathbf{y}(K+n-j)$ for $n < j$ due to the circular convolution. By defining

$$\hat{\mathbf{R}}(l) = \begin{bmatrix} \hat{R}_{1,1}(l) & \cdots & \hat{R}_{1,N_R}(l) \\ \vdots & \ddots & \vdots \\ \hat{R}_{N_R,1}(l) & \cdots & \hat{R}_{N_R,N_R}(l) \end{bmatrix}, \ (l = -P, \cdots, P) \quad (4)$$

where

$$\hat{R}_{i_{R1},i_{R2}}(l) \triangleq \frac{1}{K}\sum_{n=0}^{K-1} y_{i_{R1}}(n) y_{i_{R2}}^*(n-l), \quad (5)$$

(3) can be rewritten as

$$\hat{\mathbf{R}}_\mathrm{T} = \begin{bmatrix} \hat{\mathbf{R}}(0) & \hat{\mathbf{R}}(1) & \cdots & \hat{\mathbf{R}}(P) \\ \hat{\mathbf{R}}(-1) & \mathbf{R}(0) & \cdots & \hat{\mathbf{R}}(P-1) \\ \vdots & \vdots & \ddots & \vdots \\ \hat{\mathbf{R}}(-P) & \hat{\mathbf{R}}(1-P) & \cdots & \hat{\mathbf{R}}(0) \end{bmatrix} \quad (6)$$

It should be mentioned that, in MIMO-OFDM systems, the received time-domain signal is often corrupted due to imperfections caused by some factors such as the frequency offset and the larger peak-to-average power ratio (PAPR) etc. Considering the fact that many compensation techniques for these imperfections can only be implemented in the frequency-domain [4], [5], a good time-domain signal is not available. Thus, the existing time-domain MIMO channel estimation techniques cannot be directly applied to MIMO-OFDM systems. In this case, some algorithms such as that given in [10] resort to the second-order statistics of the frequency-domain signal for channel estimation. However, these algorithms require the estimation of the second-order statistics of a long vector whose size is equal to or larger than the number of subcarriers. Thus, to estimate the correlation matrix reliably, they need a large number of OFDM symbols. In addition, since the matrices involved in these algorithms are of huge size, their computational complexity is extremely high. In contrast, the time-domain methods as that proposed in [1], which are based on second-order statistics of a short vector with a size that is only slightly larger than the channel length, are much more efficient. In these methods, however, to achieve the correlation matrix of the time-domain signal, an IFFT processor is needed in the receiver, which incurs a high computational complexity and a long time delay in real-time implementation. In this paper, we propose a new algorithm for the computation of the time-domain correlation matrix directly from the received frequency-domain signal.

III. Proposed Frequency-Domain Estimation of the Correlation Matrix

The frequency-domain correlation matrix estimation algorithm origins from a frequency-domain equalization approach for the noise-free case, which can be represented by

$$\mathbf{Z}(k) = \left[\mathbf{w}_1^T, \cdots, \mathbf{w}_{N_R}^T\right]\mathbf{Y}'(k) \quad (7)$$

where $\mathbf{w}_i$, $(i = 1, \cdots, N_T)$ is a transversal equalizer of size $(P+1) \times 1$ and

$$\mathbf{Y}'(k) = \left(\mathbf{I}_{N_R} \otimes \mathbf{F}_1^T(k)\right)\begin{bmatrix} Y_1(k) \\ \vdots \\ Y_{N_R}(k) \end{bmatrix}, \quad (8)$$

with $Y_{i_R}(k)$ representing the received frequency-domain signal from the i_R-th receive antenna at the k-th subcarrier and

$\mathbf{F}_1(k)$ is the k-th row of the matrix $\mathbf{F}_1$, which is the first $P+1$ columns of the $K \times K$ DFT matrix $\mathbf{F}_0$. We now calculate the autocorrelation matrix of $\mathbf{Y}'(k)$ using

$$\hat{\mathbf{R}}_\mathrm{F} \triangleq \frac{1}{K}\sum_{k=0}^{K-1} \mathbf{Y}'(k)\left(\mathbf{Y}'(k)\right)^H, \tag{9}$$

which is called, for the sake of convenience, the frequency-domain correlation matrix in this paper, even though it is actually an estimate. Using (8) into (9), $\hat{\mathbf{R}}_\mathrm{F}$ can be rewritten as

$$\hat{\mathbf{R}}_\mathrm{F} = \begin{bmatrix} \mathbf{\Delta}_{1,1} & \cdots & \mathbf{\Delta}_{1,N_R} \\ \vdots & \ddots & \vdots \\ \mathbf{\Delta}_{N_R,1} & \cdots & \mathbf{\Delta}_{N_R,N_R} \end{bmatrix}, \tag{10}$$

where

$$\mathbf{\Delta}_{i_{R1},i_{R2}} \triangleq \frac{1}{K}\sum_{k=0}^{K-1} \mathbf{F}_1^T(k)\,\mathbf{F}_1^*(k)\,Y_{i_{R1}}(k)\, Y_{i_{R2}}^*(k),\ (i_{R1}, i_{R2} = 1, \cdots, N_R). \tag{11}$$

By letting

$$y_{i_R}(n) \triangleq \frac{1}{\sqrt{K}}\sum_{k=0}^{K-1} Y_{i_R}(k)\, \mathrm{e}^{j2\pi(kn/K)},$$

and using (5), (10) and (11), one can verify that

$$\hat{\mathbf{R}}_\mathrm{F} = \sum_{l=-P}^{P} \hat{\mathbf{R}}(l) \otimes \mathbf{I}_l \tag{12}$$

where $\mathbf{I}_l$ is a partial identity matrix, which has its nonzero elements 1 only in the l-th diagonal and zero elements otherwise. For example, $l=0$ corresponds to $\mathbf{I}_{(P+1)\times(P+1)}$, while $l=1$ gives

$$\begin{bmatrix} \mathbf{0}_{P\times 1} & \mathbf{I}_{P\times P} \\ 0 & \mathbf{0}_{1\times P} \end{bmatrix}.$$

On the other hand, from (6), $\hat{\mathbf{R}}_\mathrm{T}$ can be expressed as

$$\hat{\mathbf{R}}_\mathrm{T} = \sum_{l=-P}^{P} \mathbf{I}_l \otimes \hat{\mathbf{R}}(l). \tag{13}$$

It is interesting to see from (12) and (13) that the frequency-domain correlation matrix $\hat{\mathbf{R}}_\mathrm{F}$ and the time-domain correlation matrix $\hat{\mathbf{R}}_\mathrm{T}$ contain exactly the same information, which is dictated by $\hat{\mathbf{R}}(l), (l = -P, -P+1, \cdots, P)$, and the only difference between the two versions is the order of the Kronecker product.

It is obvious from (12) that once $\hat{\mathbf{R}}_\mathrm{F}$ has been calculated, $\hat{\mathbf{R}}(l), (l=-P,-P+1,\cdots,P)$ can be obtained, and then $\hat{\mathbf{R}}_\mathrm{T}$ can be determined directly from (13). The new frequency-domain correlation matrix estimation algorithm can be described as follows.

Step i) For each subcarrier, compute $\mathbf{Y}'(k)$ based on the received frequency-domain signal using (8);

Step ii) Estimate the frequency-domain correlation matrix $\hat{\mathbf{R}}_\mathrm{F}$ using (9);

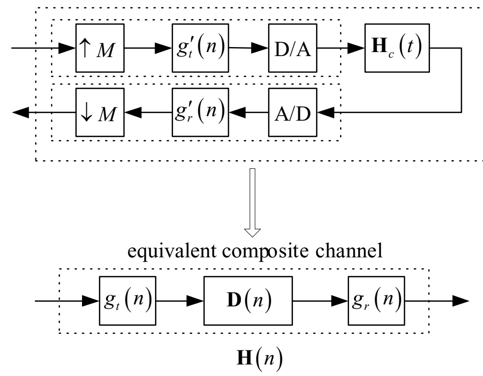

Fig. 1. Discrete-time channel model with pulse shaping

Step iii) Obtain $\hat{\mathbf{R}}(l), (l=-P,-P+1,\cdots,P)$ from $\hat{\mathbf{R}}_\mathrm{F}$ using (12);

Step iv) Construct $\hat{\mathbf{R}}_\mathrm{T}$ from $\hat{\mathbf{R}}(l), (l = 1-P, \cdots, P-1)$ using (13).

IV. Application to Semi-Blind Channel Estimation of MIMO-OFDM Systems

In this section, we will apply the proposed frequency-domain correlation matrix estimation algorithm to revise our semi-blind MIMO-OFDM channel estimation approach previously developed in [2], [3] by utilizing the knowledge of the pulse-shaping filter and matched filter.

A. Brief Overview of Linear Prediction based Semi-Blind MIMO-OFDM Channel Estimation

Based on the second-order statistics, the MIMO linear prediction method can be employed to estimate a blind constraint $\hat{\mathbf{B}}$ for the channel vector $\mathbf{h} \triangleq \mathrm{vec}\left[\mathbf{h}_1, \cdots, \mathbf{h}_{N_R}\right]$ [2], [3], where

$$\mathbf{h}_{i_R} \triangleq \left[h_{i_R,1}(0), \cdots, h_{i_R,1}(L-1), \cdots, h_{i_R,N_T}(L-1)\right]^T.$$

By combining the blind constraint with a training-based LS criterion, a semi-blind cost function for channel estimation can be formulated as

$$\min_{\hat{\mathbf{h}}} \Delta = \left\|\mathbf{Y}_\mathrm{pilot} - \tilde{\mathbf{A}}\hat{\mathbf{h}}\right\|_F^2 + \alpha\left\|\hat{\mathbf{B}}\hat{\mathbf{h}}\right\|_F^2 \tag{14}$$

where $\tilde{\mathbf{A}}$ is a pilot signal matrix, $\mathbf{Y}_\mathrm{pilot}$ is the corresponding signal vector and $\alpha > 0$ is a weighting factor. The solution to this optimization problem is given by

$$\hat{\mathbf{h}} = \left(\tilde{\mathbf{A}}^H\tilde{\mathbf{A}} + \alpha\hat{\mathbf{B}}^H\hat{\mathbf{B}}\right)^\dagger \tilde{\mathbf{A}}^H\mathbf{Y}_\mathrm{pilot}. \tag{15}$$

B. Frequency-Domain Semi-Blind Algorithm with Pulse-Shaping

As the pulse-shaping filter and the matched filter are normally pre-determined in a communication system, their knowledge can be exploited to improve the channel estimation accuracy. The most commonly used pulse-shaping filter in communication systems has the following raised-cosine impulse response [11]

$$g(t) = \operatorname{sinc}\left(\frac{\pi t}{T}\right)\frac{\cos\left(\frac{\beta\pi t}{T}\right)}{1-\left(\frac{2\beta t}{T}\right)^2}$$

where β is the roll-off factor and T the symbol period. In digital communications, the pulse-shaping filter is often realized by an up-sampled raised-cosine FIR filter. Thus, the composite channel model should include the pulse-shaping filter, the analog multi-path channel $\mathbf{H}_c(t)$ and the matched filter as shown in Fig. 1. In this model, an upsampling is usually implemented by inserting $M-1$ zeros between the consecutive input samples prior to pulse-shaping. The transmit filter $g_t(t)$ and the receive filter $g_r(t)$ in (1) are replaced by two root raised-cosine FIR filters $g_t'(n)$ and $g_r'(n)$, whose sampling period is $\frac{T}{M}$. A discrete-time model equivalent to the composite MIMO-OFDM channel can be formulated as an $N_R \times N_T$ matrix $\mathbf{H}(n)$, whose (i_R, i_T)-th element is given by

$$h_{i_R,i_T}(n) = g_t(n) * d_{i_R,i_T}(n) * g_r(n) \tag{16}$$

where $g_t(n) = g_t'(Mn)$, $g_r(n) = g_r'(Mn)$ and $d_{i_R,i_T}(n)$ is the (i_R, i_T)-th element of the equivalent multi-path channel matrix $\mathbf{D}(n)$. It should be mentioned that a common assumption used in many existing algorithms is $g_t(n) * g_r(n) = \delta(n)$, which implies $\mathbf{H}(n) = \mathbf{D}(n)$. In this sense, therefore, the pulse-shaping effect has been neglected. However, this assumption is not true in practical systems. Letting $g(n) \triangleq g_t(n) * g_r(n)$, (16) can be written as

$$h_{i_R,i_T}(n) = g(n) * d_{i_R,i_T}(n) \tag{17}$$

with $g(n) = 0,\ n \notin [0, L_g - 1]$, $d_{i_R,i_T}(n) = 0, n \notin [0, L_d - 1]$, and $L = L_g + L_d - 1$. In what follows, we will improve the semi-blind algorithm proposed in [2], [3] by using (17), namely, we will estimate $\mathbf{D}(n)$, instead of the large-dimensional matrix $\mathbf{H}(n)$, with the information of $g(n)$. Since the number of channel parameters has been considerably decreased, the estimation performance of the new approach is expected to be much better than that of those focusing only on the estimation of the composite channel $\mathbf{H}(n)$. The idea of the new semi-blind algorithm is presented below.

The frequency-domain signal model between the i_R-th receive antenna and the i_T-th transmit antenna can be represented by

$$Y_{i_R}(k) = H_{i_R,i_T}(k)\, X_{i_T}(k),\ \text{for}\ k \in [0, K-1] \tag{18}$$

where $H_{i_R,i_T}(k) \triangleq \mathbf{f}_{0k}\left[\mathbf{h}_{i_R,i_T}^T, \mathbf{0}_{1\times(K-L)}\right]^T$ and $\mathbf{f}_{0k}$ is the k-th row of $\mathbf{F}_0$. From (17), we can derive

$$H_{i_R,i_T}(k) = G(k)\, D_{i_R,i_T}(k) \tag{19}$$

where $G(k) \triangleq \mathbf{f}_{0k}\left[g(0), \cdots, g(L_g-1), \mathbf{0}_{1\times(L-L_g)}\right]^T$ and $D_{i_R,i_T}(k) \triangleq \mathbf{f}_{0k}\left[\mathbf{d}_{i_R,i_T}^T, \mathbf{0}_{1\times(K-L_b)}\right]^T$. Using (19) into (18), we can obtain the received signal after removing the effect of the pulse-shaping and matched filters,

$$\begin{aligned} Y_{i_R}'(k) &\triangleq G^{-1}(k)\, Y_{i_R}(k) \\ &= D_{i_R,i_T}(k)\, X_{i_T}(k). \end{aligned} \tag{20}$$

Now the previous semi-blind algorithm as stated in the previous subsection can be applied to (20) to obtain a frequency-domain channel estimate.

It should also be mentioned that the received signal, after removing the effect of pulse-shaping and matched filters, is only available in the frequency-domain. In the linear prediction-based semi-blind channel estimation method, to derive the correlation matrix $\hat{\mathbf{R}}_\mathrm{T}$ by using (3), an IFFT process is needed to convert the frequency-domain signal to the time-domain version, which incurs an extra computational burden and a long time delay for the implementation. Here, the frequency domain correlation estimation algorithm developed in Section III can be used for obtaining $\hat{\mathbf{R}}_\mathrm{T}$. Since the complexity of the frequency-domain method is approximate to that of the calculation of (3), the IFFT processing has been avoided in the frequency-domain approach. Moreover, since the proposed frequency-domain channel estimation is equivalent to the time-domain method in [2], [3]. Thus, the practical scheme of determining the weighting factor α suggested in [2] can be used directly in the frequency-domain approach.

V. Simulation Results

Here we consider a MIMO-OFDM system with 2 transmit and 4 receive antennas. The number of subcarriers is set to 512, the length of cyclic prefix is 10, and the length of the linear predictor is $P = 4$. In our simulation, the QPSK modulation is used and a Rayleigh channel modelled by a 3-tap MIMO-FIR filter is assumed, in which each tap corresponds to a 2×4 random matrix whose elements are i.i.d. complex Gaussian variables with zero mean and unit variance. A square root raised cosine filter with order 16, oversampling rate 4 and rolloff factor 0.15 is used for pulse-shaping. As shown in [2], [3], the channel estimation performance is associated with $\mathbf{H}(0)$. Accordingly, we define the metric

$$\eta \triangleq \frac{\|\mathbf{H}(0)\|_F^2}{\sum_{n=0}^{2} \|\mathbf{H}(n)\|_F^2}$$

and conduct a simulation study with respect to different ranges of η.

In the experiments, for the purpose of comparison, the composite channel vector $\mathbf{h}$ is first estimated by the LS method. As for the estimation of the pure multi-path channel vector $\mathbf{d}$, we consider the frequency-domain LS and the proposed frequency-domain semi-blind method, all with pulse shaping. For simplicity, we call these three methods as the basic LS, enhanced LS, enhanced semi-blind methods. Note that the enhanced LS method can be easily obtained by setting α to zero in the proposed semi-blind method with pulse-shaping.

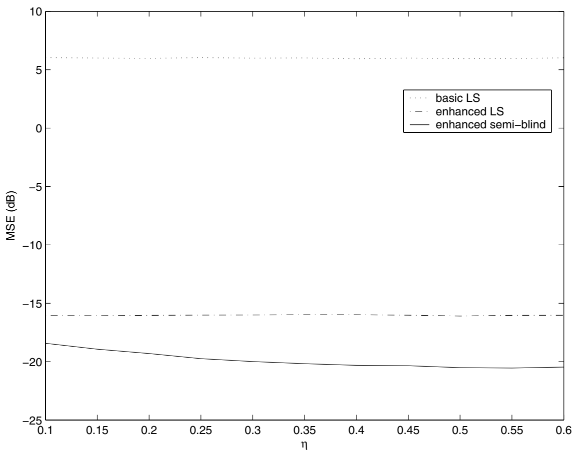

Fig. 2. MSE versus η.

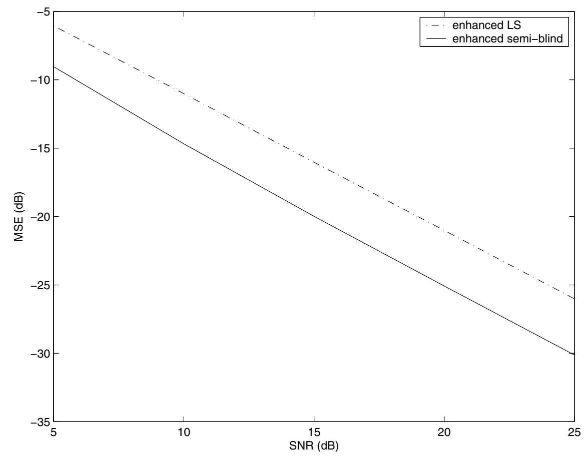

Fig. 3. MSE versus SNR

Experiment 1:*MSE versus* η

In the first experiment, the channel estimation performance in terms of the MSE versus η is investigated. The simulation is undertaken at an SNR of 15dB by 1000 Monte Carlo runs of the transmission of one OFDM symbol at 512 subcarriers, of which 32 are used as pilot for training purpose. Fig. 2 shows the MSE plots resulting from the proposed enhanced LS and semi-blind methods as well as the basic LS estimation. Obviously, the MSE performance of the proposed two enhanced methods are at least 20dB-30dB better than that of the basic LS, indicating that the new approach focusing on the pure multipath channel vector significantly outperforms that for composite channel vector irrespective of pulse-shaping. One can find that the enhanced semi-blind method significantly outperforms the enhanced LS method. In addition, the performance of the semi-blind method improves with the increasing value of η when $\eta < 0.3$, and it remains almost the same when η is in the range of 0.3 to 0.6, which represents typical mobile communication scenarios where the first arrived path is comparable to or stronger than other paths.

Experiment 2: *MSE versus SNR*

Now we investigate the channel estimation performance versus the SNR. The simulation involves 5000 Monte Carlo runs of the transmission of one OFDM symbol, of which 32 are used as pilot for training purpose. Fig. 3 shows the channel estimation results of the two enhanced methods when $\eta > 0.2$. It is seen that the enhanced semi-blind method can achieve nearly 3$\sim$4 dB gains over the enhanced LS method, when the SNR varies from 5 to 25 dB, respectively.

VI. Conclusions

A new algorithm for the estimation of the correlation matrix that is required in blind and semi-blind MIMO-OFDM channel estimation has been presented. By computing the time-domain correlation matrix directly from the received frequency-domain signal, an IFFT operation that is required in the conventional time-domain channel estimation method can be avoided. Furthermore, a new semi-blind approach has been developed for MIMO-OFDM channel estimation by utilizing the new frequency-domain correlation matrix estimation algorithm as well as exploiting the pulse-shaping filter available in the transmitter and the matched filter in the receiver. The effectiveness of the proposed technique has been validated by computer simulations.

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