Frequency-Domain Estimation of Time-Domain Correlation Matrix for MIMO-OFDM Systems

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***Abstract*—Since the received MIMO-OFDM signal is usually corrupted in the time-domain due to some adverse factors such as frequency offset and large peak-to-average power ratio (PAPR) etc., a high-quality signal may only be obtained in the frequency domain. However, the second-order statistics of the time-domain signal are very often used in a blind or semi-blind channel estimation, which means that an IFFT processor is required in the receiver to achieve the time-domain signal. This additional IFFT incurs a high computational complexity and a long time delay 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 algorithm is proven to be equivalent to the original algorithm of estimating the time-domain correlation matrix, while the former can avoid an IFFT process when high-quality signal is only available in the frequency domain.**

***Index Terms*—MIMO-OFDM system, frequency-domain, correlation matrix estimation, blind.**

I. Introduction

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 multiple-input multiple-output (MIMO) systems. In MIMO orthogonal frequency division multiplexing (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 [1]–[3], a high quality time-domain signal may not be 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 [4] 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 proposed in [5]–[7], 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 to convert the good frequency-domain signal to the desired high-quality time-domain signal, which imposes an additional computational burden and a long time delay in real-time implementation. In this paper, therefore, we will develop a method of estimating the time-domain correlation matrix that is required in the blind and semi-blind time-domain MIMO channel estimation techniques directly from the received frequency-domain signal.

Throughout the paper, we adopt the following notations:
$\otimes$ Kronecker product, $\circledast$ circular convolution,
T Transpose, H Complex conjugate transpose.

II. Preliminary

Consider a MIMO-OFDM system with N_T transmit and N_R receive antennas. For notational simplicity, one OFDM symbol with K subcarriers is considered. In general, 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:

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

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 those proposed in [5]. 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

$$\hat{\mathbf{R}}_{\mathrm{T}} = \frac{1}{K}\sum_{n=0}^{K-1} \mathbf{y}_{P+1}(n)\,\mathbf{y}_{P+1}^H(n). \tag{2}$$

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

$$\hat{\mathbf{R}}(l) = \begin{bmatrix} \hat{R}_{1,1}(l) & \cdots & \hat{R}_{1,N_R}(l) \\ \vdots & \ddots & \vdots \\ \hat{R}_{N_R,1}(l) & \cdots & \hat{R}_{N_R,N_R}(l) \end{bmatrix}, (l = -P, \cdots, P) \tag{3}$$

where
$$\hat{R}_{i_{R1},i_{R2}}(l) \triangleq \frac{1}{K} \sum_{n=0}^{K-1} y_{i_{R1}}(n) y^*_{i_{R2}}(n-l), \tag{4}$$

(2) can be rewritten as

$$\hat{\mathbf{R}}_{\mathrm{T}} = \begin{bmatrix} \hat{\mathbf{R}}(0) & \hat{\mathbf{R}}(1) & \cdots & \hat{\mathbf{R}}(P) \\ \hat{\mathbf{R}}(-1) & \mathbf{R}(0) & \cdots & \hat{\mathbf{R}}(P-1) \\ \vdots & \vdots & \ddots & \vdots \\ \hat{\mathbf{R}}(-P) & \hat{\mathbf{R}}(1-P) & \cdots & \hat{\mathbf{R}}(0) \end{bmatrix}. \tag{5}$$

As the received time-domain signal is often corrupted due to imperfections in MIMO-OFDM systems, a good time-domain signal is not available. In order to achieve the correlation matrix of the time-domain signal for the use of these methods, an IFFT processor is needed in the receiver, which incurs a high computational complexity and a long time delay in real-time implementation. In next section, we propose a new algorithm for the computation of the time-domain correlation matrix directly from the received frequency-domain signal.

III. Frequency-Domain Estimation of the Time-Domain Correlation Matrix

In our previous work [8], we presented briefly an idea of computing the time-domain correlation matrix $\hat{\mathbf{R}}_{\mathrm{T}}$ in the frequency domain without giving theoretical justification. In this section, we develop the detailed estimation algorithm and show its computational advantage.

A. The Frequency-Domain Correlation Matrix $\hat{\mathbf{R}}_{\mathrm{F}}$

Our derivation of the frequency-domain correlation matrix is inspired by the idea of frequency-domain equalization. In the noise-free case, the frequency-domain signal model corresponding to (1) can be written as

$$\begin{bmatrix} \mathbf{Y}_1(m) \\ \vdots \\ \mathbf{Y}_{N_R}(m) \end{bmatrix} = \begin{bmatrix} \mathrm{diag}(\mathbf{F}_1\mathbf{h}_{1,1}) & \cdots & \mathrm{diag}(\mathbf{F}_1\mathbf{h}_{1,N_T}) \\ \vdots & \ddots & \vdots \\ \mathrm{diag}(\mathbf{F}_1\mathbf{h}_{N_R,1}) & \cdots & \mathrm{diag}(\mathbf{F}_1\mathbf{h}_{N_R,N_T}) \end{bmatrix} \begin{bmatrix} \mathbf{X}_1(m) \\ \vdots \\ \mathbf{X}_{N_T}(m) \end{bmatrix} \tag{6}$$

where
$$\mathbf{h}_{i_R,i_T} \triangleq [h_{i_R,i_T}(0), \cdots, h_{i_R,i_T}(L-1)]^T \tag{7}$$

and $X_{i_T}(m)$, $Y_{i_R}(m)$ representing the transmitted and the received frequency-domain signal at the m-th subcarrier and $\mathbf{F}_1$ consists of the first L columns of a $K \times K$ DFT matrix $\mathbf{F}_0$. Let us consider a frequency-domain equalization for (6), namely,

$$\mathbf{Z}(m) = [\mathrm{diag}(\mathbf{F}_2\mathbf{w}_1), \cdots, \mathrm{diag}(\mathbf{F}_2\mathbf{w}_{N_R})] \begin{bmatrix} \mathbf{Y}_1(m) \\ \vdots \\ \mathbf{Y}_{N_R}(m) \end{bmatrix} \tag{8}$$

where $\mathbf{w}_i \triangleq [w_i(0), \cdots, w_i(P)]^T$, $(i = 1, \cdots, N_R)$ is a transversal equalizer of size $P+1$ with respect to the i-th receive antenna and $\mathbf{F}_2$ consists of the first $P+1$ columns of the DFT matrix. Substituting (6) into (8) gives

$$\mathbf{Z}(m) = \left[\sum_{i_R=1}^{N_R} \mathrm{diag}(\mathbf{F}_2\mathbf{w}_{i_R}) \mathrm{diag}(\mathbf{F}_1\mathbf{h}_{i_R,1}), \cdots, \sum_{i_R=1}^{N_R} \mathrm{diag}(\mathbf{F}_2\mathbf{w}_{i_R}) \mathrm{diag}(\mathbf{F}_1\mathbf{h}_{i_R,N_T}) \right] \begin{bmatrix} \mathbf{X}_1(m) \\ \vdots \\ \mathbf{X}_{N_T}(m) \end{bmatrix}. \tag{9}$$

Letting $\mathbf{F}_3$ be the first $L+P$ columns of the DFT matrix and

$$\mathbf{c}_{i_R,i_T} \triangleq [c_{i_R,i_T}(0), \cdots, c_{i_R,i_T}(L+P-1)]^T$$

where
$$c_{i_R,i_T}(n) = w_{i_R}(n) * h_{i_R,i_T}(n),$$

(9) can be rewritten as

$$\mathbf{Z}(m) = [\mathrm{diag}[\mathbf{F}_3(\mathbf{c}_{1,1} + \cdots + \mathbf{c}_{N_R,1})], \cdots \mathrm{diag}[\mathbf{F}_3(\mathbf{c}_{1,N_T} + \cdots + \mathbf{c}_{N_R,N_T})]] \begin{bmatrix} \mathbf{X}_1(m) \\ \vdots \\ \mathbf{X}_{N_T}(m) \end{bmatrix}. \tag{10}$$

If a specific set of equalizers $\mathbf{w}_i$, $(i = 1, \cdots, N_R)$ with respect to the i_T-th transmit antenna is designed such that

$$\mathrm{diag}[\mathbf{F}_3(\mathbf{c}_{1,i_T} + \cdots + \mathbf{c}_{N_R,i_T})] = \begin{cases} \mathbf{I} & \text{if } (i = i_T) \\ \mathbf{0} & \text{if } (i \neq i_T) \end{cases},$$

then the signal sent by this transmit antenna can be recovered, namely,

$$\mathbf{Z}(m) = \mathbf{X}_{i_T}(m).$$

Thus, if such N_T sets of equalizers are determined, all the signals from the N_T transmit antennas can be recovered.

The above shows the idea of frequency-domain equalization for MIMO-OFDM systems. Borrowing this idea, we now derive a frequency-domain correlation matrix. For the k-th subcarrier, (8) gives

$$\mathbf{Z}(m,k) = [\mathrm{diag}(\mathbf{F}_2(k)\mathbf{w}_1), \cdots, \mathrm{diag}(\mathbf{F}_2(k)\mathbf{w}_{N_R})] \begin{bmatrix} Y_1(m,k) \\ \vdots \\ Y_{N_R}(m,k) \end{bmatrix} \tag{11}$$

where $\mathbf{F}_2(k)$ is the k-th row of the matrix $\mathbf{F}_2$. For notational simplicity, the index m of OFDM symbols is dropped from now on without loss of clarity. Thus, (11) can be rewritten as

$$\mathbf{Z}(k) = \left[\mathbf{w}_1^T, \cdots, \mathbf{w}_{N_R}^T\right] \mathbf{Y}'(k) \tag{12}$$

where
$$\mathbf{Y}'(k) \triangleq \left(\mathbf{I}_{N_R} \otimes \mathbf{F}_2^T(k)\right) \begin{bmatrix} Y_1(k) \\ \vdots \\ Y_{N_R}(k) \end{bmatrix}. \tag{13}$$

Note that $\mathbf{Y}'(k)$ can be regarded as an input of the frequency-domain equalizer $\mathbf{w}_i$, $(i = 1, \cdots, N_R)$ at the k-th subcarrier. Let us consider the autocorrelation matrix of $\mathbf{Y}'(k)$,

$$\hat{\mathbf{R}}_{\mathrm{F}} \triangleq \frac{1}{K}\sum_{k=0}^{K-1} \mathbf{Y}'(k)\left(\mathbf{Y}'(k)\right)^H, \tag{14}$$

which is called, for the sake of convenience, the frequency-domain correlation matrix in this paper, even though it is actually an estimate. Using (13) into (14), $\hat{\mathbf{R}}_{\mathrm{F}}$ can be rewritten as

$$\hat{\mathbf{R}}_{\mathrm{F}} = \begin{bmatrix} \boldsymbol{\Delta}_{1,1} & \cdots & \boldsymbol{\Delta}_{1,N_R} \\ \vdots & \ddots & \vdots \\ \boldsymbol{\Delta}_{N_R,1} & \cdots & \boldsymbol{\Delta}_{N_R,N_R} \end{bmatrix} \tag{15}$$

where

$$\boldsymbol{\Delta}_{i_{R1},i_{R2}} \triangleq \frac{1}{K}\sum_{k=0}^{K-1} \mathbf{F}_2^T(k)\,\mathbf{F}_2^*(k)\,Y_{i_{R1}}(k)\,Y_{i_{R2}}^*(k), \quad (i_{R1}, i_{R2} = 1, \cdots, N_R). \tag{16}$$

It will be shown in the next subsection that the frequency-domain correlation matrix $\hat{\mathbf{R}}_{\mathrm{F}}$ contains exactly the same information as the time-domain correlation matrix $\hat{\mathbf{R}}_{\mathrm{T}}$, yet it would significantly facilitate the computation of the second-order statistics in channel estimation by avoiding the IFFT operation converting the frequency-domain signal $Y'_{i_R}(k)$ to the time-domain signal $y'_{i_R}(n)$ in the proposed semi-blind algorithm.

B. Computation of $\hat{\mathbf{R}}_{\mathrm{T}}$ based on $\hat{\mathbf{R}}_{\mathrm{F}}$

Here, we develop a frequency-domain method for the computation of the time-domain correlation matrix $\hat{\mathbf{R}}_{\mathrm{T}}$ in terms of the frequency-domain version $\hat{\mathbf{R}}_{\mathrm{F}}$. Here, we first reveal the relationship between the two matrices. By defining $\phi(k) = \mathrm{e}^{-j2\pi(k/K)}$, (16) can be rewritten as

$$\boldsymbol{\Delta}_{i_{R1},i_{R2}} = \frac{1}{K}\sum_{k=0}^{K-1} \begin{bmatrix} 1 & \phi^{-1}(k) & \phi^{-2}(k) & \cdots & \phi^{-P}(k) \\ \phi^{1}(k) & 1 & \phi^{-1}(k) & \cdots & \phi^{1-P}(k) \\ \phi^{2}(k) & \phi^{1}(k) & 1 & \cdots & \phi^{2-P}(k) \\ \vdots & \vdots & \vdots & \ddots & \vdots \\ \phi^{P}(k) & \phi^{P-1}(k) & \phi^{P-2}(k) & \cdots & 1 \end{bmatrix} Y_{i_{R1}}(k)\,Y_{i_{R2}}^*(k). \tag{17}$$

On the other hand, by letting $y_{i_{R1}}(n) \triangleq \frac{1}{\sqrt{K}}\sum_{k=0}^{K-1} Y_{i_{R1}}(k)\,\mathrm{e}^{j2\pi(kn/K)}$ and $y_{i_{R2}}(n) \triangleq \frac{1}{\sqrt{K}}\sum_{k=0}^{K-1} Y_{i_{R2}}(k)\,\mathrm{e}^{j2\pi(kn/K)}$, $\hat{R}_{i_{R1},i_{R2}}$ as defined in (4) can be expressed as

$$\hat{R}_{i_{R1},i_{R2}}(l) = \frac{1}{K}\sum_{n=0}^{K-1} y_{i_{R1}}(n)\left[\frac{1}{\sqrt{K}}\sum_{k=0}^{K-1} Y_{i_{R2}}^*(k)\,\phi^{n-l}(k)\right] = \frac{1}{K}\sum_{k=0}^{K-1} Y_{i_{R1}}(k)\,Y_{i_{R2}}^*(k)\,\phi^{-l}(k). \tag{18}$$

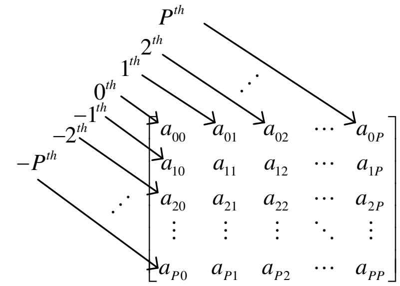

Fig. 1. Position of the diagonal for the partial identity matrix

It is clear that (18) gives the expression for the elements of $\boldsymbol{\Delta}_{i_{R1},i_{R2}}$, i. e.,

$$\boldsymbol{\Delta}_{i_{R1},i_{R2}} = \begin{bmatrix} \hat{R}_{i_{R1},i_{R2}}(0) & \hat{R}_{i_{R1},i_{R2}}(1) & \hat{R}_{i_{R1},i_{R2}}(2) & \cdots & \hat{R}_{i_{R1},i_{R2}}(P) \\ \hat{R}_{i_{R1},i_{R2}}(-1) & \hat{R}_{i_{R1},i_{R2}}(0) & \hat{R}_{i_{R1},i_{R2}}(1) & \cdots & \hat{R}_{i_{R1},i_{R2}}(P-1) \\ \hat{R}_{i_{R1},i_{R2}}(-2) & \hat{R}_{i_{R1},i_{R2}}(-1) & \hat{R}_{i_{R1},i_{R2}}(0) & \cdots & \hat{R}_{i_{R1},i_{R2}}(P-2) \\ \vdots & \vdots & \vdots & \ddots & \vdots \\ \hat{R}_{i_{R1},i_{R2}}(-P) & \hat{R}_{i_{R1},i_{R2}}(1-P) & \hat{R}_{i_{R1},i_{R2}}(2-P) & \cdots & \hat{R}_{i_{R1},i_{R2}}(0) \end{bmatrix}. \tag{19}$$

Thus, $\boldsymbol{\Delta}_{i_{R1},i_{R2}}$ can be expressed in terms of $\hat{R}_{i_{R1},i_{R2}}(l), (l = -P, -P+1 \cdots, P)$ as

$$\boldsymbol{\Delta}_{i_{R1},i_{R2}} = \sum_{l=-P}^{P} \hat{R}_{i_{R1},i_{R2}}(l) \otimes \mathbf{I}_l \tag{20}$$

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. The sequence of the diagonals is demonstrated in Fig. 1. For example, $l = 0$ corresponds to $\mathbf{I}_{(P+1)\times(P+1)}$, while $l = 1$ gives

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

Now we utilize the above result to derive the relationship between the frequency-domain correlation matrix $\hat{\mathbf{R}}_{\mathrm{F}}$ and the time-domain version $\hat{\mathbf{R}}_{\mathrm{T}}$. Using (20) and (15), one can prove

$$\hat{\mathbf{R}}_{\mathrm{F}} = \sum_{l=-P}^{P} \hat{\mathbf{R}}(l) \otimes \mathbf{I}_l. \tag{21}$$

On the other hand, from (5), $\hat{\mathbf{R}}_{\mathrm{T}}$ can be expressed as

$$\hat{\mathbf{R}}_{\mathrm{T}} = \sum_{l=-P}^{P} \mathbf{I}_l \otimes \hat{\mathbf{R}}(l). \tag{22}$$

It is interesting to see from (21) and (22) 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.

TABLE I
COMPLEXITY COMPARISON OF TWO METHODS FOR COMPUTING $\hat{\mathbf{R}}_{\mathrm{T}}$

	Frequency-Domain Algorithm	Time-Domain Algorithm
Step 1	$\mathbf{R}_A(k) = \begin{bmatrix} Y_1(k) \\ \vdots \\ Y_{N_R}(k) \end{bmatrix} \left[Y_1^*(k), \cdots, Y_{N_R}^*(k)\right],$ $(k = 0, \cdots, K-1)$: KN_R^2 multiplications	a K-length IFFT processing: $K\log_2(K)N_R$ multiplications $K\log_2(K)N_R$ additions
Step 2	$\mathbf{R}_B(k) = \left[\mathbf{I}_{N_R} \otimes \mathbf{F}_2^T(k)\right]\mathbf{R}_A(k),$ $(k = 0, \cdots, K-1)$: $KN_R^2(P+1)$ multiplications	$\mathbf{R}_D(n) = \mathbf{y}_{P+1}(n)\mathbf{y}_{P+1}^H(n),$ $(n = 0, \cdots, K-1)$: $KN_R^2(P+1)^2$ multiplications
Step 3	$\mathbf{R}_C(k) = \mathbf{R}_B(k)\left[\mathbf{I}_{N_R} \otimes \mathbf{F}_2^*(k)\right],$ $(k = 0, \cdots, K-1)$: $KN_R^2(P+1)^2$ multiplications	$\hat{\mathbf{R}}_T = \frac{1}{K}\sum_{n=0}^{K-1}\mathbf{R}_D(n)$: $N_R^2(P+1)^2$ multiplications $(K-1)N_R^2(P+1)^2$ additions
Step 4	$\hat{\mathbf{R}}_{\mathrm{F}} = \frac{1}{K}\sum_{k=0}^{K-1}\mathbf{R}_C(k)$: $N_R^2(P+1)^2$ multiplications $(K-1)N_R^2(P+1)^2$ additions	
In Total	Multiplications: $(K+1)N_R^2(P+1)^2 + N_R^2(P+2)$ Additions: $(K-1)N_R^2(P+1)^2$	Multiplications: $K\log_2(K)N_R + (K+1)N_R^2(P+1)^2$ Additions: $K\log_2(K)N_R + (K-1)N_R^2(P+1)^2$

It is obvious from (21) 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 (22). The new frequency-domain correlation matrix estimation algorithm can be described as follows.

Step i) For each subcarrier, compute $\mathbf{Y}'(k)$ using (13) based on the received frequency-domain signal;

Step ii) Estimate the frequency-domain correlation matrix $\hat{\mathbf{R}}_{\mathrm{F}}$ using (14);

Step iii) Obtain $\hat{\mathbf{R}}(l)$, $(l = -P, -P+1, \cdots, P)$ from $\hat{\mathbf{R}}_{\mathrm{F}}$ in (21);

Step iv) Construct $\hat{\mathbf{R}}_{\mathrm{T}}$ from $\hat{\mathbf{R}}(l)$, $(l = 1-P, \cdots, P-1)$ using (22).

Before closing this section, we would like to evaluate the complexity of computing $\hat{\mathbf{R}}_{\mathrm{T}}$ via the proposed frequency-domain method as opposed to the conventional time-domain estimation method. The frequency-domain method requires the calculation of $\mathbf{Y}'(k)\left[\mathbf{Y}'(k)\right]^H$ and $\hat{\mathbf{R}}'_{\mathrm{F}}$, which cost $KN_R^2\left(P^2+3P+3\right)$ and $N_R^2(P+1)^2$ complex multiplications, respectively. Note that the calculation of $\hat{\mathbf{R}}_{\mathrm{F}}$ also requires $(K-1)N_R^2(P+1)^2$ complex additions. On the other hand, the time-domain estimation method mainly consists of the calculation of $\hat{\mathbf{R}}_{\mathrm{T}}$ and a K-point IFFT process. The former requires $(K+1)N_R^2(P+1)^2$ multiplications and $(K-1)N_R^2(P+1)^2$ additions, while the latter involves $N_R K\log_2(K)$ complex multiplications and additions. A detailed comparison of the two methods is given in Table I. Considering that K is in general much larger than N_R and P, $K\log_2(K)N_R \gg N_R^2(P+2)$. Thus, the complexity of the frequency-domain algorithm is approximately the same as that for the calculation of $\hat{\mathbf{R}}_{\mathrm{T}}$ in the time-domain algorithm excluding the IFFT. It means that, in comparison with the conventional time-domain estimation method, the IFFT processing has been avoided in the frequency-domain approach.

IV. CONCLUSIONS

In this paper, a frequency-domain correlation matrix estimation algorithm has been proposed for MIMO-OFDM systems. By computing the time-domain correlation matrix directly from the received frequency-domain signal, the new frequency-domain algorithm is equivalent to the conventional time-domain correlation matrix estimation algorithm. It means that when the good quality received signal is only available in the frequency-domain, an IFFT operation that is required in the conventional time-domain channel estimation can be avoided in the new algorithm.

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