Comparative Study of Time-Domain and Frequency-Domain Channel Estimation in MIMO-OFDM Systems

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Abstract—**The time-domain channel estimator is compared with the frequency-domain channel estimator when multiple-input-multiple-output (MIMO) orthogonal frequency division multiplexing (OFDM) is employed. The time-domain maximum likelihood channel estimation (TMLE) and its mean square error (MSE) are derived assuming the knowledge of the maximum channel length. If we use only one OFDM symbol for the estimation, the preamble sequence of each transmit antenna uses a different set of subcarriers in MIMO-OFDM systems. As a result, the MSE of the frequency-domain least square channel estimation (FLSE) is derived assuming that linear interpolation is employed. Numerical results show that the MSE of TMLE is smaller than that of FLSE and that the performance difference increases as the number of transmit antenna increases.**

I. Introduction

Orthogonal frequency division multiplexing (OFDM) systems have been applied to wireless communications, such as wireless local area network (LAN) [1], due to their robustness to multipath fading and the high bandwidth efficiency. On the other hand, the use of multiple antennas at both transmitter and receiver enhances the channel capacity of the wireless system significantly [2], [3]. To exploit this capacity, the Bell Lab. layered space time (BLAST) architecture was proposed in [4]. Alternatively, the space time code (STC) was used as a method to achieve the capacity limit by utilizing coding and diversity gains [5]. Multiple transmit and receive antennas are combined with OFDM to improve the capacity and reliability of communications [6]. When MIMO-OFDM systems are implemented in conjunction with either BLAST or STC, the channel state information is required for proper decoding. Thus, the channel estimation becomes more important as the number of antennas increases.

In single-input-single-output (SISO) OFDM systems, the frequency-domain least square channel estimator (FLSE) is generally used for the preamble-based channel estimation. Although the complexity of FLSE is small, the mean square error (MSE) of the estimation is relatively high [7]. In order to reduce the MSE, the time-domain ML channel estimator (TMLE) is considered in [8] for SISO-OFDM systems. The minimum mean square error (MMSE) channel estimation in the frequency domain is considered in [9], which requires the channel statistics such as power and the probability distribution of the channel coefficients.

Provided that only one OFDM symbol is used for estimating the channel as is assumed in 802.16 system [10]; then, the preamble sequence of each transmit antenna uses a different set of subcarriers in MIMO-OFDM systems. For FLSE, the estimated channel coefficients are interpolated to obtain the channel estimates of the subcarriers where the preamble is not located. However, the interpolation error may drastically magnify the MSE of FLSE in frequency selective channels, which OFDM systems generally assume [1], [10]. Thus, we derive the MSE of FLSE in the presence of interpolation, and show that the MSE of FLSE increases significantly in frequency selective channels. In order to avoid the interpolation error, the time-domain channel estimation is considered as an alternative method. The MSE of TMLE is derived assuming the knowledge of the maximum channel length. The comparison between the TMLE and the FLSE shows that the MSE of TMLE is significantly smaller than that of FLSE in frequency selective channels.

II. Signal Model

MIMO-OFDM systems considered in this paper are shown in Fig. 1. At each transmit antenna, the data is modulated

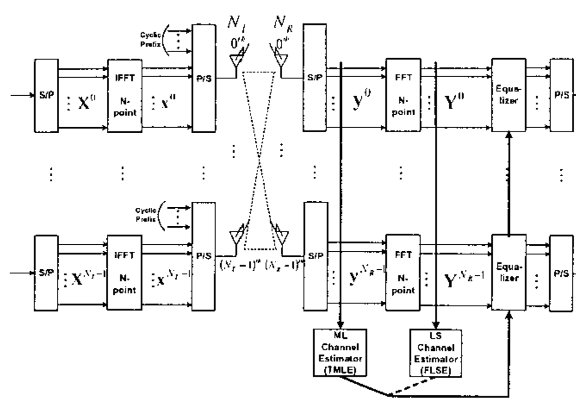


Fig. 1. Block Diagram of MIMO-OFDM Systems

by inverse fast Fourier transform (IFFT) and a cyclic prefix (CP) with length ν is added. To avoid intersymbol interference (ISI), $\nu > L-1$, where L is the maximum length of each channel. The system has N_T transmit antennas and N_R receive antennas. We assume frequency selective channels, which remain unchanged during at least one OFDM symbol.

Provided the preamble sequence $\{X_k^t\}$ for t^{th} transmit antenna in the frequency domain, the preamble sequence is allocated at subcarrier k where the subcarrier index k is the element of the position set $\mathbf{p}^t$. In other words, the position set $\mathbf{p}^t$ of the t^{th} transmit antenna indicates the location of the preamble sequence. Given N_T transmit antennas, the time-domain transmit signal at the t^{th} transmit antenna is expressed as

$$x_n^t = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} X_k^t \exp\left(j\frac{2\pi nk}{N}\right)$$
$$X_k^t = \begin{cases} \text{nonzero} & k \in \mathbf{p}^t \\ 0 & \text{otherwise} \end{cases} \quad (1)$$

The index k denotes the k^{th} subcarrier.

Given N_R receive antennas, the time-domain received signal at the r^{th} receive antenna after removing the CP is given by

$$y_n^r = \sum_{t=0}^{N_T-1} \sum_{l=0}^{L-1} x_{|n-l|_N}^t h_l^{tr} + z_n^r, \quad 0 \le n \le N-1 \quad (2)$$

where h_l^{tr} is the channel parameter between the t^{th} transmit antenna and the r^{th} receive antenna. The indices n and l indicate time and channel, respectively. N is the preamble length and equal to the IFFT size because we assume that the estimation duration is one OFDM symbol. z_n^r is assumed to be additive white Gaussian noise (AWGN). $|n-l|_N$ indicates "$n-l$ modulo N."

Eq. (2) can be written in matrix form as

$$\mathbf{y} = \mathbf{x}\mathbf{h} + \mathbf{z} \quad (3)$$

where $\mathbf{y} = [\mathbf{y}^0\ \mathbf{y}^1 \cdots \mathbf{y}^{N_R-1}]$, $\mathbf{y}^r = [y_0^r\ y_1^r \cdots y_{N-1}^r]^T$ and the superscript $[\cdot]^T$ indicates transpose. $\mathbf{x}$ is a $N \times LN_T$ matrix

$$\mathbf{x} = \left[\mathbf{x}^0\ \mathbf{x}^1 \cdots \mathbf{x}^{N_T-1}\right]$$
$$\mathbf{x}^t = \begin{bmatrix} x_0^t & x_{N-1}^t & \cdots & x_{N-L+1}^t \\ x_1^t & x_0^t & \cdots & x_{N-L+2}^t \\ \vdots & \vdots & \ddots & \vdots \\ x_{N-1}^t & x_{N-2}^t & \cdots & x_{N-L}^t \end{bmatrix}, \quad (4)$$

and $\mathbf{h}$ is the time-domain channel matrix with dimension $LN_T \times N_R$

$$\mathbf{h} = \begin{bmatrix} \mathbf{h}^{00} & \mathbf{h}^{01} & \cdots & \mathbf{h}^{0(N_R-1)} \\ \mathbf{h}^{10} & \mathbf{h}^{11} & \cdots & \mathbf{h}^{1(N_R-1)} \\ \vdots & \vdots & \ddots & \vdots \\ \mathbf{h}^{(N_T-1)0} & \mathbf{h}^{(N_T-1)1} & \cdots & \mathbf{h}^{(N_T-1)(N_R-1)} \end{bmatrix} \quad (5)$$

where $\mathbf{h}^{tr} = \left[h_0^{tr}\ h_1^{tr} \cdots h_{L-1}^{tr}\right]^T$. Finally, $\mathbf{z} = \left[\mathbf{z}^0\ \mathbf{z}^1 \cdots \mathbf{z}^{N_R-1}\right]$ and $\mathbf{z}^r = \left[z_0^r\ z_1^r \cdots z_{N-1}^r\right]^T$ is a zero-mean white Gaussian vector with covariance matrix $\mathbf{C}_{\mathbf{z}^r} = E\left\{\mathbf{z}^r\mathbf{z}^{rH}\right\} = \sigma_z^2\mathbf{I}_N$ where I_N is the $N \times N$ identity matrix and the superscript $[\cdot]^H$ indicates Hermitian transpose.

In the frequency domain, the received signal at the r^{th} receive antenna is represented as follows:

$$\mathbf{Y}^r = \sum_{t=0}^{N_T-1} \mathbf{X}^t\mathbf{H}^{tr} + \mathbf{Z}^r \quad . (6)$$

where $\mathbf{Y}^r = \left[Y_0^r\ Y_1^r \cdots Y_{N-1}^r\right]^T$, $\mathbf{X}^t = diag\left(X_0^t, X_1^t, \cdots, X_{N-1}^t\right)$ and $\mathbf{Z}^r = \left[Z_0^r\ Z_1^r \cdots Z_{N-1}^r\right]^T$. $\mathbf{H}^{tr} = \left[H_0^{tr}\ H_1^{tr} \cdots H_{N-1}^{tr}\right]^T$ is the frequency-domain channel between the t^{th} transmit antenna and the r^{th} receive antenna. The relation between $\mathbf{h}^{tr}$ and $\mathbf{H}^{tr}$ is given by

$$\mathbf{H}^{tr} = \mathbf{W}_L\mathbf{h}^{tr} \quad (7)$$

where $[\mathbf{W}_L]_{p,q} = \exp\left(-j\frac{2\pi pq}{N}\right)$, $(N \times L)$.

III. Time-domain ML Estimator

In this section, the TMLE is derived for MIMO-OFDM systems. The MSE of TMLE is also derived as the performance measure. Since the channel equalization is performed in the frequency domain, the MSE of TMLE is converted into that of the frequency-domain channel. In addition, the preamble condition that minimizes the MSE is obtained.

Given $\mathbf{h}$, the conditional probability density function is given by

$$\Lambda(\mathbf{y}|\mathbf{h}) = \frac{1}{(\pi\sigma_z^2)^{NN_R}} \exp\left\{-\frac{1}{\sigma_z^2}Tr\left\{[\mathbf{y}-\mathbf{x}\mathbf{h}]^H[\mathbf{y}-\mathbf{x}\mathbf{h}]\right\}\right\} \quad (8)$$

where $Tr\{\cdot\}$ denotes the trace of a square matrix. The ML estimation of $\mathbf{h}$ can be found by maximizing Eq. (8). This is equivalent to minimizing the log likelihood function:

$$\Lambda_L(\mathbf{y}|\mathbf{h}) = Tr\left\{[\mathbf{y}-\mathbf{x}\mathbf{h}]^H[\mathbf{y}-\mathbf{x}\mathbf{h}]\right\}. \quad (9)$$

Since $\Lambda_L(\mathbf{y}|\mathbf{h})$ is a convex function over $\mathbf{h}$, the estimation can be obtained by searching $\mathbf{h}$ to satisfy the following condition

$$\frac{\partial\Lambda_L(\mathbf{y}|\mathbf{h})}{\partial\mathbf{h}} = 0. \quad (10)$$

This condition implies that the time-domain ML estimate is given by

$$\hat{\mathbf{h}} = \left(\mathbf{x}^H\mathbf{x}\right)^{-1}\mathbf{x}^H\mathbf{y}, \qquad \text{(TMLE)}. \quad (11)$$

For $\mathbf{x}$ with dimension $N \times LN_T$, rank($\mathbf{x}$) = rank($\mathbf{x}^H\mathbf{x}$) $\le$ min(N, LN_T). If $N < LN_T$, the inverse of $LN_T \times LN_T$ square matrix $\mathbf{x}^H\mathbf{x}$ cannot exist because rank($\mathbf{x}^H\mathbf{x}$) $\le N < LN_T$. The necessary and sufficient condition to have a unique channel estimate (i.e., $\mathbf{x}^H\mathbf{x}$ is nonsingular) is $N \ge LN_T$.

From Eq. (3) and Eq. (11), the MSE of TMLE over the time-domain channel is given by

$$\begin{aligned}\mathrm{MSE}_{h,\mathrm{TMLE}} &= \frac{1}{LN_TN_R}E\left[Tr\left\{\left(\hat{\mathbf{h}}-\mathbf{h}\right)^H\left(\hat{\mathbf{h}}-\mathbf{h}\right)\right\}\right]\\ &= \frac{1}{LN_TN_R}E\left[Tr\left\{\left(\mathbf{x}^H\mathbf{x}\right)^{-1}\mathbf{x}^H\mathbf{z}\mathbf{z}^H\mathbf{x}\left(\mathbf{x}^H\mathbf{x}\right)^{-1}\right\}\right]\\ &= \frac{\sigma_z^2}{LN_T}Tr\left\{\left(\mathbf{x}^H\mathbf{x}\right)^{-1}\right\}\end{aligned} \tag{12}$$

Using Eq. (7) and Eq. (12), the MSE of TMLE over the frequency-domain channel is given by

$$\begin{aligned}\mathrm{MSE}_{H,\mathrm{TMLE}} &= \frac{1}{NN_TN_R}\\ &\times E\left[\sum_{t=0}^{N_T-1}\sum_{r=0}^{N_R-1}Tr\left\{\left(\hat{\mathbf{H}}^{tr}-\mathbf{H}^{tr}\right)^H\left(\hat{\mathbf{H}}^{tr}-\mathbf{H}^{tr}\right)\right\}\right]\\ &= \frac{1}{N_TN_R}E\left[\sum_{t=0}^{N_T-1}\sum_{r=0}^{N_R-1}Tr\left\{\left(\hat{\mathbf{h}}^{tr}-\mathbf{h}^{tr}\right)^H\left(\hat{\mathbf{h}}^{tr}-\mathbf{h}^{tr}\right)\right\}\right]\\ &= \frac{\sigma_z^2}{N_T}\sum_{t=0}^{N_T-1}Tr\left\{\left(\mathbf{x}^{t^H}\mathbf{x}^t\right)^{-1}\right\}\end{aligned} \tag{13}$$

In [8], the optimum preamble condition was derived for the TMLE in SISO systems. Similarly, in MIMO-OFDM systems, the optimum preamble condition to minimize the MSE Eq. (12) or Eq. (13) is

$$\mathbf{x}^H\mathbf{x} = \frac{N}{N_T}\mathbf{I}_{LN_T}, \tag{14}$$

if the sum of power of the preamble sequence in each transmit antenna is constant (e.g., $\sum_{k=0}^{N-1}|X_k^t|^2 = N/N_T,\ t \in \{0,1,\cdots,N_T-1\}$). The preamble condition Eq. (14) is the extension of the preamble condition in [8] for SISO systems. Employing the optimum preamble condition Eq. (14), the MSE of TMLE is given by

$$\mathrm{MSE}_{h,\mathrm{TMLE}}^{opt} = \frac{N_T}{N}\sigma_z^2, \tag{15}$$

$$\mathrm{MSE}_{H,\mathrm{TMLE}}^{opt} = \frac{LN_T}{N}\sigma_z^2. \tag{16}$$

As the number of transmit antenna increases, both $\mathrm{MSE}_{h,\mathrm{TMLE}}^{opt}$ and $\mathrm{MSE}_{H,\mathrm{TMLE}}^{opt}$ increase because total transmit power is assumed to be the same regardless of the number of the transmit antenna. Note that $\mathrm{MSE}_{H,\mathrm{TMLE}}^{opt}$ is L times larger than $\mathrm{MSE}_{h,\mathrm{TMLE}}^{opt}$ because the noise is increased L times through FFT.

IV. Frequency-domain LS Estimator

In this section, the FLSE is investigated in MIMO-OFDM systems. As a performance measure, the MSE is analytically derived assuming that linear interpolation is used without interpolation error. Then, the MSE of FLSE is compared with that of TMLE that is obtained in Eq. (16).

From [7], the frequency-domain LS estimate is given by

$$\hat{H}_k^{tr} = \frac{Y_k^r}{X_k^t}, \quad k \in \mathbf{p}^t \qquad \text{(FLSE)}. \tag{17}$$

For FLSE in MIMO-OFDM systems, the preamble sequence of each transmit antenna should use a different set of subcarriers because only one OFDM symbol is assumed to be used for training signal. In other words, the following condition should be satisfied.

$$\begin{aligned}&\mathbf{p}^i \cap \mathbf{p}^j = \emptyset \qquad (i \neq j,\ \ 0 \leq i,j \leq N_T-1)\\ &\bigcup_{t=0}^{N_T-1}\mathbf{p}^t = \{0,1,\cdots,N-1\}\end{aligned} \tag{18}$$

In order to satisfy the above condition, $\mathbf{p}^t$ may be defined as follows,

$$\begin{aligned}\mathbf{p}^t \equiv \{t, t+N_T, \cdots, t+(N/N_T-1)N_T\}\\ t \in \{0,1,\cdots,N_T-1\}.\end{aligned} \tag{19}$$

For the t^{th} transmit antenna, the interpolation is needed in order to obtain the channel parameters at the subcarriers that are not the element of $\mathbf{p}^t$.

If the coherence bandwidth of the channel is large compared with the subcarrier spacing, i.e., L is small, the coefficients of real channel have the linear relationship as,

$$\begin{aligned}H_k^{tr} = \frac{p}{N_T}H_{|t+(m+1)N_T|_N}^{tr} + \left(1-\frac{p}{N_T}\right)H_{|t+mN_T|_N}^{tr}\\ k \in (\mathbf{p}^t)^c\end{aligned} \tag{20}$$

where $m = \lceil (k-N_T)/N_T \rceil$, and $p = k - |t+mN_T|_N$. The operator $\lceil\cdot\rceil$ indicates the lowest integer larger than $(\cdot)$. In this case, we may use linear interpolation to obtain the channel parameters of the t^{th} transmit antenna at the subcarriers that are not the element of $\mathbf{p}^t$. When linear interpolation is used, the channel estimates are given by

$$\begin{aligned}\hat{H}_k^{tr} = \frac{p}{N_T}\hat{H}_{|t+(m+1)N_T|_N}^{tr} + \left(1-\frac{p}{N_T}\right)\hat{H}_{|t+mN_T|_N}^{tr}\\ k \in (\mathbf{p}^t)^c\end{aligned} \tag{21}$$

where m and p are the same as those of Eq. (20). From Eq. (20) and Eq. (21), we consider the MSE of FLSE assuming that interpolation error disappears under the channel with large coherence bandwidth.

Using Eq. (6), Eq. (17), Eq. (20), and Eq. (21), the MSE without interpolation error is as follows:

$$
\begin{aligned}
\mathrm{MSE}_{H.\mathrm{FLSE}}^{No} &= \frac{1}{N N_T N_R} \\
&\times E\left[\sum_{t=0}^{N_T-1}\sum_{r=0}^{N_R-1} Tr\left\{\left(\hat{\mathbf{H}}^{tr}-\mathbf{H}^{tr}\right)^H\left(\hat{\mathbf{H}}^{tr}-\mathbf{H}^{tr}\right)\right\}\right] \\
&= \frac{\sigma_z^2}{N N_T}\sum_{t=0}^{N_T-1}\left\{\sum_{k\in\mathbf{p}^t}\frac{1}{|X_k^t|^2}+\right. \\
&\left.\sum_{k\in(\mathbf{p}^t)^c}\left(\frac{(p/N_T)^2}{\left|X^t_{|t+(m+1)N_T|_N}\right|^2}+\frac{(1-p/N_T)^2}{\left|X^t_{|t+mN_T|_N}\right|^2}\right)\right\} \\
&\geq \left(\frac{2N_T^2+1}{3N_T^2}\right)\sigma_z^2
\end{aligned}
\tag{22}
$$

where the lower bound is obtained when $|X_k^t|^2 = 1$ for $k \in \mathbf{p}^t$ and $t = \{0, 1, \cdots, N_T - 1\}$. This can be proved using "Lagrange Multiplier Method"[8]. When $N_T = 1$, the lower bound of the MSE becomes σ_z^2, which is the same as the MSE from [7]. As the number of transmit antenna increases, we can obtain the averaging effect because the several independent noise signals are added through interpolation. The averaging effect of noise improves the SNR; accordingly, the MSE decreases as the number of transmit antenna increases.

Although the MSE of FLSE without interpolation error decreases as the number of transmit antenna increases, the MSE of TMLE is always smaller than that of FLSE if $L \leq \left(\frac{2N_T^2+1}{3N_T^3}\right)N$

$$
\mathrm{MSE}_{H.\mathrm{TMLE}}^{opt} \leq \mathrm{MSE}_{H.\mathrm{FLSE}}^{No} \quad \text{subject to } L \leq \left(\frac{2N_T^2+1}{3N_T^3}\right)N. \tag{23}
$$

For the channel with large coherence bandwidth, it is not difficult to satisfy the constraint of the maximum channel length because $L \ll N$. In addition, L is generally smaller than $N/4$ because the CP length is generally at most 25% of the IFFT size [1], [10].

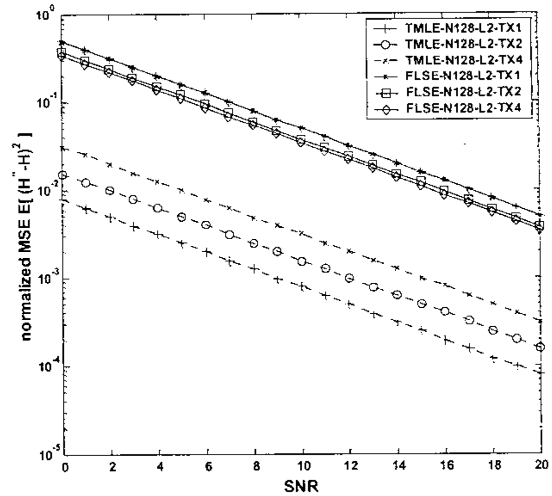
If the coherence bandwidth of the channel is small, the interpolation error cannot be neglected. Considering the effect of interpolation error, the MSE of FLSE can be written as

$$
\begin{aligned}
\mathrm{MSE}_{H.\mathrm{FLSE}} &= \mathrm{MSE}_{H.\mathrm{FLSE}}^{No} + I_{err}\left(\text{type}, N_T, L/N\right) \\
&\geq \left(\frac{2N_T^2+1}{3N_T^2}\right)\sigma_z^2 + I_{err}\left(\text{type}, N_T, L/N\right).
\end{aligned}
\tag{24}
$$

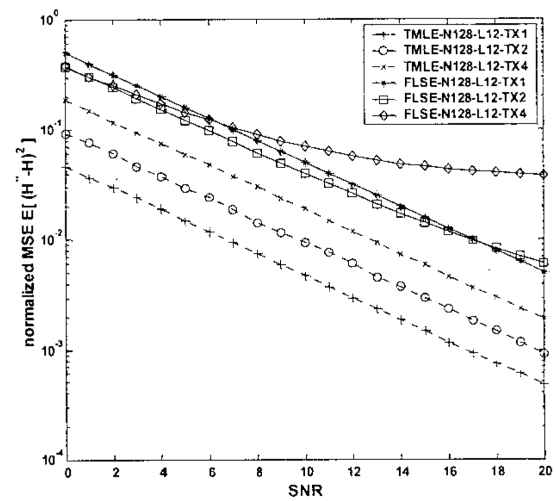
The interpolation error I_{err} depends on interpolation type, N_T and L/N. Generally, as N_T or L/N increases, I_{err} increases significantly, while $\mathrm{MSE}_{H.\mathrm{TMLE}}$ is directly proportional to L or N_T. In the next section, the interpolation error I_{err} is evaluated numerically.

V. Numerical Results

Fig. 2 shows the comparison between the TMLE and the FLSE as the number of transmit antenna increases from one to four. The total number of subcarriers N is 128 where all the subcarriers are used for carrying the preambles. By varying the channel length L from two to twelve, the effect of coherence bandwidth to the MSE of the estimation is analyzed. We assume that the channel consists of L independent taps, in which the power of each tap is the same. In addition, the number of transmit antenna N_T changes from one to four, which illustrates the influence of employing MIMO to the SNR losses and interpolation errors. The preamble is transmitted through Rayleigh fading channel, in which the correlation between antennas does not exist. In addition, we assume that the channel remains unchanged during one OFDM



(a) N=128, L=2



(b) N=128, L=12

Fig. 2. Comparison between the TMLE and the FLSE as the number of transmit antennas increases from one to four

symbol. The dashed lines indicate the MSE of TMLE Eq. (16) and the solid lines denote the MSE of FLSE using linear interpolation. To achieve the lower bound of the $\mathrm{MSE}_{H,\mathrm{FLSE}}$, the optimum preamble sequence is used (e.g., $|X_k^t|^2 = 1$, $k \in \mathrm{p}^t$, $t \in \{0, 1, \cdots, N_T - 1\}$). Fortunately, this preamble sequence satisfies the optimum preamble condition of TMLE Eq. (14).

In Fig. 2(a), the length of channel L is two, i.e., the coherence bandwidth is large compared with the subcarrier spacing. When $N_T = 1$, the SNR difference between the TMLE and the FLSE is about 18 dB which agrees with the analytical results of TMLE Eq. (16) and FLSE Eq. (22). If the number of transmit antenna increases by two times, the MSE of TMLE grows directly proportional to N_T (i.e., the SNR difference is 3 dB) because the signal energy used for estimating channel is reduced by half. On the contrary, the MSE of FLSE decreases as described in Eq. (22). The SNR differences to attain the same MSE between $N_T = 1$ and $N_T = 2$ and between $N_T = 1$ and $N_T = 4$ are respectively for 1.25dB and 1.63dB. However, the performance gap between the TMLE and the FLSE remains large despite the decrease in the MSE of FLSE.

In order to evaluate the MSE of the estimation in frequency selective channels, the length of channel L is increased to 12 in Fig. 2(b). i.e., the coherence bandwidth is $1/6$ of that in Fig. 2(a). When $N_T = 1$ in TMLE, the SNR difference between $L = 2$ and $L = 12$ is about 8 dB which agrees with the MSE of TMLE derived in Eq. (16). In case of FLSE, the MSE remains the same. However, as the number of transmit antenna increases, the MSE of TMLE grows proportional to the number of transmit antenna, while the MSE performance of FLSE degenerates severely. In other words, the performance difference between the TMLE and the FLSE increases as the number of transmit antenna increases. The performance degradation of FLSE is due to the increase of interpolation error that is represented as the floor in the MSE curve of FLSE. When $N_T = 4$ in FLSE, the MSE is saturated at about 5×10^{-2}. In low SNR, the effect of interpolation error is dominated by the noise. However, the interpolation error appears apparently in high SNR. In light of the above, the MSE curves of FLSE intersects one another because the effect of the interpolation error appears at high SNR. Since OFDM systems generally assume the frequency selective channel, i.e., $L/N = 1/32 \sim 1/4$, [1][10], the effect of interpolation error in FLSE should be seriously considered, and the interpolation should be designed carefully. However, even if the interpolation error is set aside, the MSE of TMLE is smaller than that of FLSE.

VI. Conclusions

In this paper, we have compared the time-domain ML channel estimator (TMLE) with the frequency-domain LS channel estimator (FLSE) when MIMO-OFDM is employed under the assumption that the estimation duration is one OFDM symbol. Using the approach in [8], the TMLE and its MSE were derived for MIMO-OFDM systems. The FLSE using linear interpolation was investigated and the MSE of FLSE is analytically derived when there is no interpolation error. The effect of interpolation error was evaluated numerically. Through numerical results, we have shown that the the MSE of TMLE is smaller than that of FLSE and that the performance difference increases as the number of transmit antenna increases.

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