Journal of Marine Science and Technology

Volume 20 | Issue 5 Article 7

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Li, Chi-Min; Li, Guo-Wei; and Liu, Han-Yi (2012) "PERFORMANCE COMPARISON OF THE STBC-OFDM DECODERS IN A FAST FADING CHANNEL," *Journal of Marine Science and Technology*: Vol. 20: Iss. 5, Article 7.
DOI: 10.6119/JMST-011-0506-3
Available at: https://jmstt.ntou.edu.tw/journal/vol20/iss5/7

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Journal of Marine Science and Technology, Vol. 20, No. 5, pp. 534-540 (2012)
DOI: 10.6119/JMST-011-0506-3

PERFORMANCE COMPARISON OF THE STBC-OFDM DECODERS IN A FAST FADING CHANNEL

Chi-Min Li, Guo-Wei Li, and Han-Yi Liu

Key words: space-time block code, OFDM, fast fading channel.

ABSTRACT

In a fast fading channel, the orthogonality of the two consecutive STBC-OFDM symbols will be destroyed and the received signals can not be perfectly separated. As a result, the bit error rate (BER) performance will be degraded seriously. In this paper, we analyze several co-channel interference (CCI) cancellation decoders for the STBC-OFDM in a fast fading channel including the Alamouti method, SIC method, ML method, DMLD method, and a proposed DZFD method. The proposed DZFD method has the advantage of reducing the computational complexity for the DMLD method. We also conduct the outdoor channel measurements to verify the CCI cancellation ability for different methods.

I. INTRODUCTION

Orthogonal Frequency Division Multiplexing (OFDM) with Space-Time Block Coding (STBC-OFDM) or the Space-Frequency Block Coding (SFBC-OFDM) has been widely investigated in the current wireless communication systems, such as IEEE 802.11 a/g, Digital Video Broadcasting for Handheld (DVB-H), and IEEE 802.16 Wireless MAN, due to its high bandwidth efficiency and the ability of overcoming frequency selective fading channel [1, 6]. A conventional OFDM system transforms the frequency selective fading channel into multiple flat fading channels so that the orthogonal transmit diversity techniques, i.e., STBC or SFBC can be applied. However, in a fast fading channel, the orthogonality of the two consecutive STBC-OFDM symbols will be destroyed. As a result, the bit error rate (BER) performance of the OFDM system will be degraded dramatically and the co-channel interference (CCI) arises [11]. Similarly, the SFBC-OFDM suffers the same problem when the frequency responses of the two adjacent subcarriers are not identical. In this case, the adjacent channel interference (ACI) happens and the received signals can not be perfectly separated.

In [5], a Diagonalized Maximum Likelihood Decoder (DMLD) has been proposed to remove the CCI of the STBC-OFDM under the fast fading environment. The DMLD adopts a specific generated matrix to maintain the orthogonality of the STBC or SFBC scheme and uses the maximum likelihood (ML) criterion to improve the BER performance. In this paper, we will analyze several STBC-OFDM decoders that have the ability to ease the CCI problem. To avoid the intensive computation of the DMLD method, a simple Diagnolized Zero Forcing Decoder (DZFD) for the STBC-OFDM system is also given. Simulation results will show that the DZFD and the DMLD decoders have the similar BER performance. Nevertheless, the DZFD method can greatly reduce the computational complexity as compared with the DMLD method.

This paper is organized as follows: the CCI problem of the conventional STBC-OFDM is described and analyzed in Section II. Several methods that ease the CCI problem are summarized in Section III. The BER performances of these schemes are presented for both the computer simulations and field experiments in Section IV. Finally, some conclusions of this paper are given in Section V.

II. STBC-OFDM IN A TIME-VARYING FADING CHANNEL

In this paper, superscripts $(.)^T$, $(.)^*$, $Q(.)$ and $(.)^H$ represent the transpose, complex conjugate, hard decision and complex conjugate transpose respectively. The system model of a STBC-OFDM system can be depicted in Fig. 1. To simplify the analysis, we consider the simple STBC g_2 encoder with two transmit antennas and one receive antenna. The input symbol vector of the STBC encoder is denoted as $X = [X(0), X(1), \ldots, X(2N-1)]^T$, where N is the number of the subcarriers. Let $X_1 = [X(0), X(1), \ldots, X(N-1)]^T$ and $X_2 = [X(N), X(N+1), \ldots, X(2N-1)]^T$, after the STBC encoder, the gener-

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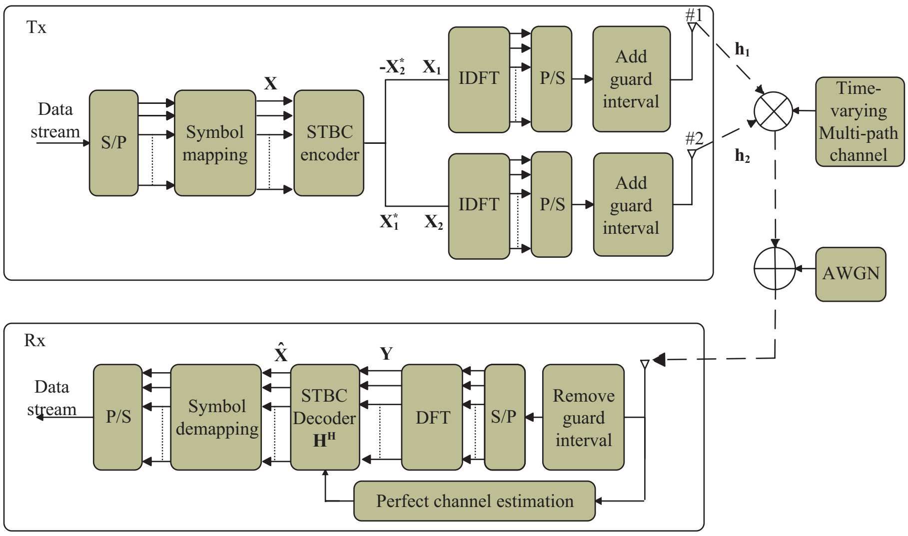

Fig. 1. System Model of the STBC-OFDM.

ated STBC g_2 coded data symbols are

$$g_2^{STBC} = \begin{bmatrix} X_1 & X_2 \\ -X_2^* & X_1^* \end{bmatrix} \quad (1)$$

Through the signal manipulations such as the Inverse Discrete Fourier Transform (IDFT), addition and removal of the cyclic prefix and the DFT, we can obtain the received symbol vector **Y** as

$$\mathbf{Y} = \begin{bmatrix} Y_t(k) \\ Y_{t+1}^*(k) \end{bmatrix} = \begin{bmatrix} H_{1,t}(k) & H_{2,t}(k) \\ H_{2,t+1}^*(k) & -H_{1,t+1}^*(k) \end{bmatrix} \begin{bmatrix} X_1(k) \\ X_2(k) \end{bmatrix} + \begin{bmatrix} Z_t(k) \\ Z_{t+1}^*(k) \end{bmatrix}$$
$$= \mathbf{HX} + \mathbf{Z} \qquad k = 0, 1, 2, ..., N-1 \quad (2)$$

where t denotes the time instant, k is the index for subcarrier, $H_{1,t}(k)$, $H^*_{1,t+1}(k)$, $H_{2,t}(k)$, $H^*_{2,t+1}(k)$ and $Z_t(k)$, $Z^*_{t+1}(k)$ are the DFT of the channel impulse responses for the first and the second transmit antennas and the channel noises respectively.

In Eq. (2), assume that the complex channel gains between the consecutive two OFDM symbols are the same (i.e. $H_{1,t}(k) = H_{1,t+1}(k)$, $H_{2,t}(k) = H_{2,t+1}(k)$). Then, the channel matrix H will be orthogonal, i.e.

$$\mathbf{H^H H}$$
$$= \begin{bmatrix} H_{1,t}(k) & H_{2,t}(k) \\ H_{2,t+1}^*(k) & -H_{1,t+1}^*(k) \end{bmatrix}^H \begin{bmatrix} H_{1,t}(k) & H_{2,t}(k) \\ H_{2,t+1}^*(k) & -H_{1,t+1}^*(k) \end{bmatrix}$$
$$= \begin{bmatrix} |H_{1,t}(k)|^2 + |H_{2,t+1}(k)|^2 & H_{1,t}^*(k)H_{2,t}(k) - H_{1,t+1}^*(k)H_{2,t+1}(k) \\ H_{1,t}(k)H_{2,t}^*(k) - H_{1,t+1}(k)H_{2,t+1}^*(k) & |H_{1,t+1}(k)|^2 + |H_{2,t}(k)|^2 \end{bmatrix}$$
$$= \begin{bmatrix} \varphi_k & 0 \\ 0 & \varphi_k \end{bmatrix} = \varphi_k \mathbf{I_2}, \quad (3)$$

where $\varphi_k = |H_{1,t}(k)|^2 + |H_{2,t+1}(k)|^2 = |H_{1,t+1}(k)|^2 + |H_{2,t}(k)|^2$. Using Eq. (3), the transmitted symbol vector can be detected simply as follows:

$$\hat{\mathbf{X}} = \frac{1}{\varphi_k} Q(\tilde{\mathbf{X}}) = \frac{1}{\varphi_k} Q(\mathbf{H^H Y}) = \frac{1}{\varphi_k} Q(\varphi_k \mathbf{X} + \mathbf{H^H Z}) \quad (4)$$

Note that we have assumed the receiver has the perfect knowledge of the channel responses. However, in the actual fast fading environments, the channel response of two consecutive OFDM symbol are not the same. Actually, the two channel responses for the STBC-OFDM system are time-varying and frequency selective. In such a fast fading environment, the channel matrix **H** will be no longer orthogonal, i.e.

$$\mathbf{H^H H} = \begin{bmatrix} \varphi_1(k) & \varepsilon(k) \\ \varepsilon^*(k) & \varphi_2(k) \end{bmatrix} \neq \varphi_k \mathbf{I_2}, \quad (5)$$

$\varphi_1(k) = |H_{1,t}(k)|^2 + |H_{2,t+1}(k)|^2$, $\varphi_2(k) = |H_{1,t+1}(k)|^2 + |H_{2,t}(k)|^2$, $\varepsilon(k) = H^*_{1,t}(k)H_{2,t}(k) - H^*_{1,t+1}(k)H_{2,t+1}(k)$. Applying Eq. (5), the detected output vector $\tilde{\mathbf{X}}$ can be rewritten as

$$\tilde{\mathbf{X}} = \begin{bmatrix} \tilde{X}_1(k) \\ \tilde{X}_2(k) \end{bmatrix} = \begin{bmatrix} \varphi_1(k)X_1(k) + \varepsilon(k)X_2(k) + Z_t'(k) \\ \varepsilon^*(k)X_1(k) + \varphi_2(k)X_2(k) + Z_{t+1}'(k) \end{bmatrix}, \quad (6)$$

$$Z' = \begin{bmatrix} Z_t'(k) \\ Z_{t+1}'(k) \end{bmatrix} = \mathbf{H^H Z}$$

where $\varepsilon(k)X_2(k)$ and $\varepsilon^*(k)X_1(k)$ are the co-channel interferences (CCI). These interferences degrade the system performance seriously.

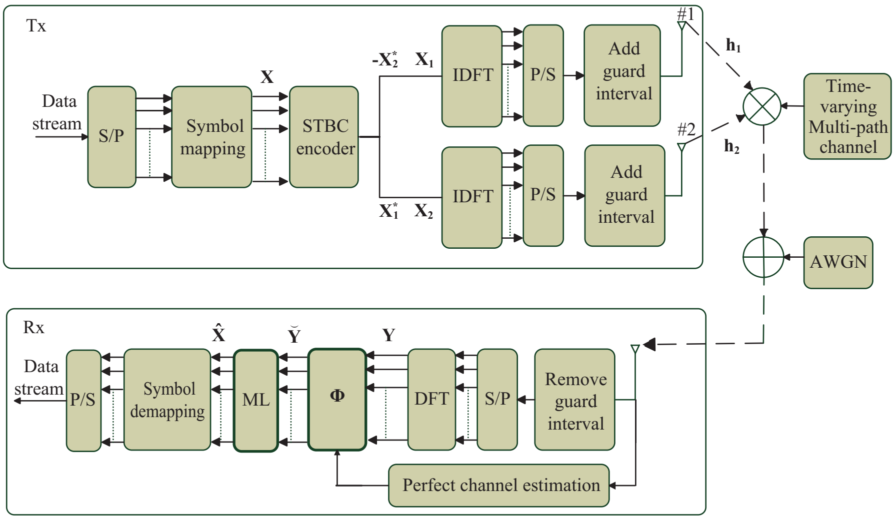

Fig. 2. STBC-OFDM with DMLD Method.

III. CCI SIGNAL DETECTION METHODS

In this section, we describe several signal detection methods to reduce the CCI problem including the SIC method, a conventional ML method, the DMLD method and a proposed DZFD method.

1. SIC Detection Method

Successive Interference Cancellation (SIC) method was proposed in [11]. The detection procedure can be described as follows:

$$<1> \tilde{X}_1(k) = \varphi_1(k)X_1(k) + \varepsilon(k)X_2(k) + Z_t^{'}(k)$$
$$<2> \tilde{X}_2(k) = \varphi_2(k)X_2(k) + \varepsilon^*(k)X_1(k) + Z_{t+1}^{'}(k)$$
$$<3> compare\ \varphi_1(k)\ and\ \varphi_2(k)\ (if\ \varphi_1(k) > \varphi_2(k))$$
$$<4> \hat{X}_1(k) = Q(\tilde{X}_1(k)/\varphi_1(k))$$
$$<5> \breve{X}_2(k) = \tilde{X}_2(k) - \varepsilon^*(k)\hat{X}_1(k)$$
$$<6> \hat{X}_2(k) = Q(\breve{X}_2(k)/\varphi_2(k))$$

SIC method decodes the symbol with the higher SNR firstly. For example, in Step. (3), if the channel gain for the first symbol is greater than the channel gain for the second transmitted symbol, i.e., $\varphi_1(k) > \varphi_2(k)$, SIC receiver decodes the first symbol in Step. (4). Otherwise ($\varphi_1(k) \leq \varphi_2(k)$), SIC receiver decodes the second symbol in Step. (4) and first symbol in Step. (6) respectively. SIC method has the advantage of low computational complexity, yet, it has the error propagation problem that will reduce the BER performance of the receiver [4, 9].

2. Conventional Maximum Likelihood (ML) Detection Method

The ML method estimates the most probable transmitted signal *via* Eq. (7).

$$\hat{\mathbf{X}} = \arg \min_{\hat{X}_1(k), \hat{X}_2(k) \in C_M} \left\| \mathbf{Y} - \mathbf{H} \begin{bmatrix} \hat{X}_1(k) \\ \hat{X}_2(k) \end{bmatrix} \right\|^2, \tag{7}$$

where C_M denotes the constellation points, and the constellation size is denoted as $|C_M|$. ML signal detection needs $|C_M|^2$ times calculation of the ML metric to solve Eq. (7). Instead of using the simple linear calculation as the SIC method, the ML method needs to solve every possible symbol combinations to achieve a better BER performance. However, computational complexity is the main problem for such the ML or Maximum Posterior Probability (MAP) decoders [3].

3. DMLD Detection Method

The DMLD method was proposed in [10]. We explain the DMLD method briefly. The STBC-OFDM system with DMLD can be illustrated in Fig. 2. Let **Φ**, as defined in Eq. (9), be a matrix that diagonizes the channel matrix **H**.

$$\mathbf{\Phi H} = \mathbf{diag}(\chi_1, \chi_2) \tag{8}$$

$$\mathbf{\Phi} = \begin{bmatrix} \Phi_{11} & \Phi_{12} \\ \Phi_{21} & \Phi_{22} \end{bmatrix} \tag{9}$$

where χ_1 and χ_2 are the complex numbers. Substituting Eq. (2) and Eq. (9) into Eq. (8), we have

$$\mathbf{\Phi} = \begin{bmatrix} \dfrac{\Phi_{12} H_{1,t+1}^*(k)}{H_{2,t}(k)} & \Phi_{12} \\ \Phi_{21} & -\dfrac{\Phi_{21} H_{1,t}(k)}{H_{2,t+1}^*(k)} \end{bmatrix}, \tag{10}$$

where Φ_{12} and Φ_{21} are arbitrary determined complex values.

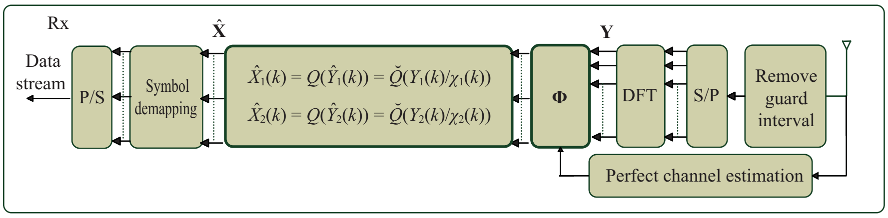

Fig. 3. Rx Structure of the DZFD Method.

Let $\mathbf{\Phi}_{12} = H_{2,t}(k)$ and $\mathbf{\Phi}_{21} = H^*_{2,t+1}(k)$, Then $\mathbf{\Phi}$ becomes

$$\mathbf{\Phi} = \begin{bmatrix} H^*_{1,t+1}(k) & H_{2,t}(k) \\ H^*_{2,t+1}(k) & -H_{1,t}(k) \end{bmatrix} \tag{11}$$

Eq. (11) reveals that $\mathbf{\Phi}$ can be easily determined after the channel estimation of $\mathbf{H}$. Besides, χ_1 and χ_2 have the relation

$$\begin{aligned} \chi &= \chi_1 = \chi_2 \\ &= H^*_{1,t+1}(k)H_{1,t}(k) + H_{2,t}(k)H^*_{2,t+1}(k). \end{aligned} \tag{12}$$

Therefore, the received signal can be separable if the receiver multiplies $\mathbf{\Phi}$ before the symbol detection in Eq. (13).

$$\begin{aligned} \breve{\mathbf{Y}} &= \mathbf{\Phi}\mathbf{Y} \\ &= diag(\chi,\chi)\mathbf{X} + \mathbf{\Phi}\mathbf{Z} \end{aligned} \tag{13}$$

After that, the ML decision for the DMLD method can be determined as

$$\begin{aligned} &\hat{X}_{j,ML}(k) = \arg \min_{\hat{X}_{j,ML}(k)\in C_M} \left\| \breve{Y}_j(k) - \chi(k)\hat{X}_{j,ML}(k) \right\|^2, \\ &j = 1, 2 \end{aligned} \tag{14}$$

where C_M denotes the constellation points [10].

4. A Proposed DZFD Detection Method

The DMLD adopts a specific generated matrix $\mathbf{\Phi}$ to maintain the orthogonality of the STBC scheme and uses the maximum likelihood (ML) criterion to improve the BER performance. However, the operation of ML method involves high computational complexity especially when the symbols are modulated with the high order modulations such as 32QAM, 64QAM, ..etc. To reduce the computational cost and maintain a similar BER performance of the DMLD method, a modified diagonal CCI detection method using the simple zero forcing (ZF) criterion is described in Eq. (15). In Eq. (13), we note that after the operation of $\mathbf{\Phi}$, there is no CCI occurred theoretically.

If we divide the χ directly from the output signal of $\mathbf{\Phi}$, we have

$$\begin{aligned} \hat{Y}_1(k) &= \breve{Y}_1(k)/\chi(k) \\ &= \frac{1}{\chi(k)}(\chi(k)X_1(k) + Z_t'(k)) = X_1(k) + Z_t''(k) \\ \hat{Y}_2(k) &= \breve{Y}_2(k)/\chi(k) \\ &= \frac{1}{\chi(k)}(\chi(k)X_2(k) + Z_{t+1}'(k)) = X_2(k) + Z_{t+1}''(k) \end{aligned} \tag{15}$$

where $Z_t''(k) = Z_t'(k)/\chi(k)$, $Z_{t+1}''(k) = Z_{t+1}''(k)/\chi(k)$. Then, hard decision $Q(\bullet)$ is used to decide the most possible signals of the transmitted symbols as $\hat{X}_1(k) = Q\left(\hat{Y}_1(k)\right)$ and $\hat{X}_2(k) = Q\left(\hat{Y}_2(k)\right)$. The receiver structure of the DZFD is depicted in Fig. 3.

In brief, the DZFD method also adopts the generated matrix $\mathbf{\Phi}$ to separate the received signal and uses the ZF criterion to reduce the computational complexity of the DMLD method.

It is well-known that the conventional ZF criterion has the noise enhancement drawback. The noise will be amplified if the decoder compensates the channel attenuation simply by dividing the received signal with a very small channel gain. In the proposed DZFD method, the noise enhancement happens only when the divisor χ is small. However, this happens only when all the four independent or low-correlated channel gains are small (Eq. 12). Due to low-correlated property under the fast-fading environment, this problem happens rarely and the following simulations will demonstrate this observed behavior.

IV. SIMULATION RESULTS

In this section, the Alamouti detection method [2], SIC detection method, ML detection method, DMLD detection method and the DZFD detection method are compared in the fast fading channel scenarios. Table 1 lists the parameters for this simulation. Table 2 is the COST207 TU Channel Model employed in the simulation. We consider the STBC-OFDM system with two transmit antennas and one receive antenna.

1. BER Performance Analysis

Fig. 4 shows the BER performance of various detection methods in the STBC-OFDM system when the speed of the mobile is at 120 km/hr. The Alamouti method suffers severe

Table 1. Simulation parameters.

Carrier frequency	2.5 GHz	
System bandwidth	3 MHz	
No. of subcarriers	128	
Subcarrier spacing	23.4 kHz	
Sampling duration	0.33 μs	
Symbol duration	42.7 μs	
Modulation	16QAM	
Environment	COST 207 TU	
User velocity	120 km/hr	300 km/hr

Table 2. COST 207 TU channel model.

	Path Delay (us)	Power (dB)
1	0	-3
2	0.2	0
3	0.5	-2
4	1.6	-6
5	2.3	-8
6	5	-10

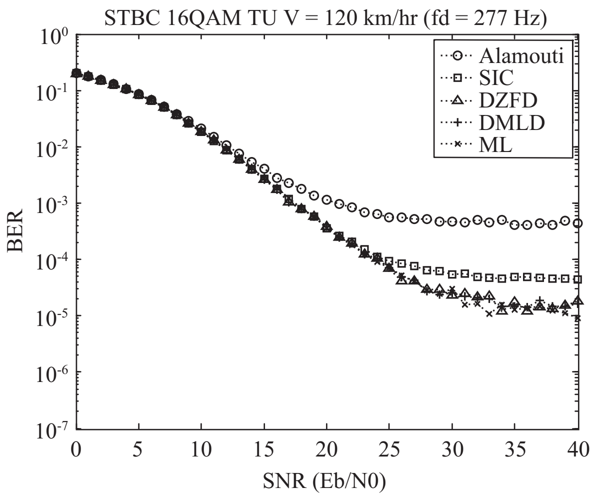

Fig. 4. BER performance (120 km/hr).

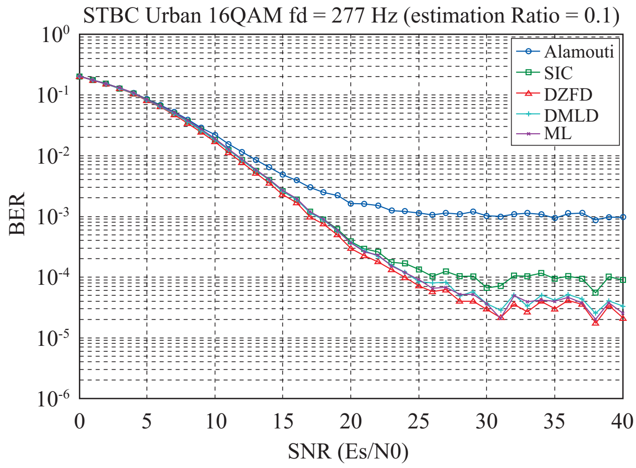

Fig. 5. Robustness simulation.

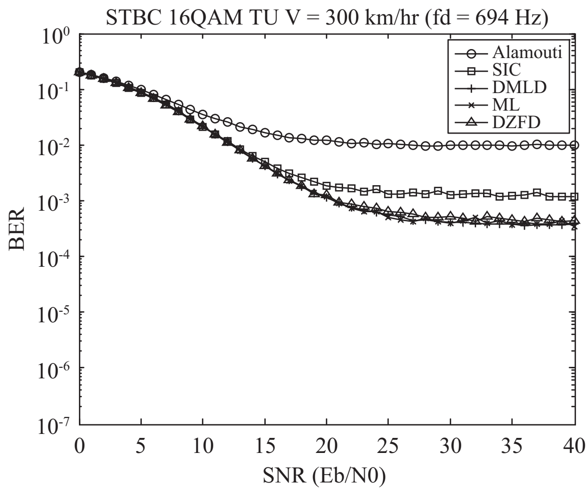

Fig. 6. BER performance (300 km/hr).

performance degradation due to the CCI and Inter Carrier Interference (ICI). Besides, the performance improvement of the SIC method is limited because of the error propagation problem. Meanwhile, the performance of DMLD method, the DZFD method and ML method are almost the same. Fig. 5 is the robustness simulation of different decoders under the 10% channel estimation errors. That is, we replace the channel matrix $\mathbf{H}$ with the $\mathbf{H}' = \mathbf{H} + \Delta\mathbf{H}$ to perform the BER simulation for different decoders, $E\{|\Delta\mathbf{H}|/|\mathbf{H}|\} = 0.1$, $E\{.\}$ denotes the expectation operation. Results show that the performance of DMLD method, the DZFD method and ML method are very similar. If the speed of the mobile increases to 300 km/hr such as a high-speed rail scenario, the BER performances of various detection methods are shown in Fig. 6. The effect of CCI dominants the BER performance and the error floor occurred at the low SNR region for every method.

Furthermore, we analyze these methods by using the measured outdoor channels shown in Fig. 7. The receiver was placed on the top of an 11-floor building and was marked as Rx in Fig. 7. The transmitter was moving along the MS path. Besides, the center frequency of the receiver is 2.44 GHz with 5 MHz signal bandwidth. Fig. 8 is an example of non-line of sight (NLOS) delay profile of the path measured by using the RUSK channel sounder. The BER performance of various detection methods in the measured channel are shown in Fig. 9. Results show that the ML method has the best performance among these methods under the measured scenario. The DMLD, DZFD method outperform the SIC and the Alamouti detection method and have the similar BER performance. However, the DZFD has the advantage of low computational complexity than the DMLD method in the following complexity analysis.

Table 3. Computational complexity of various detection methods.

	Alamouti	SIC	ML	DMLD	DZFD
Complexity	$8 \times N$	$15 \times N$	$(6 \times \lvert C_M \rvert^2) \times N$	$(4 \times \lvert C_M \rvert + 6) \times N$	$10 \times N$
QPSK	$8 \times N$	$15 \times N$	$96 \times N$	$22 \times N$	$10 \times N$
16QAM	$8 \times N$	$15 \times N$	$1536 \times N$	$70 \times N$	$10 \times N$
64QAM	$8 \times N$	$15 \times N$	$24576 \times N$	$262 \times N$	$10 \times N$

* *N* denotes the number of subcarriers.

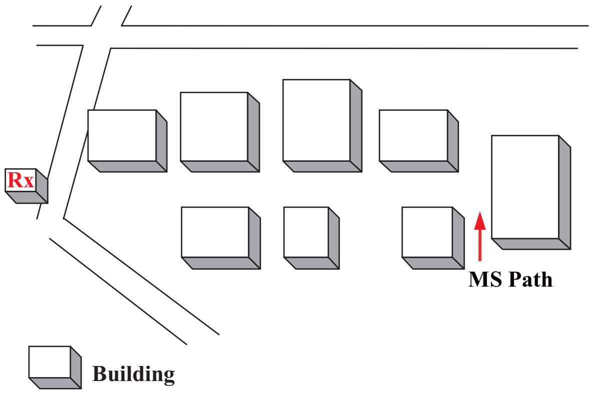

Fig. 7. Layout of the measurement.

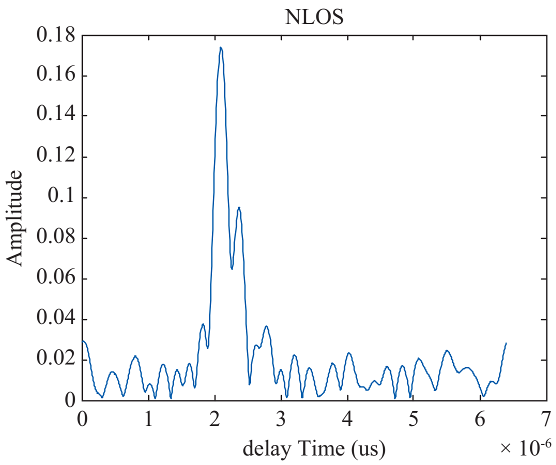

Fig. 8. The NLOS delay profile of the measured outdoor channel.

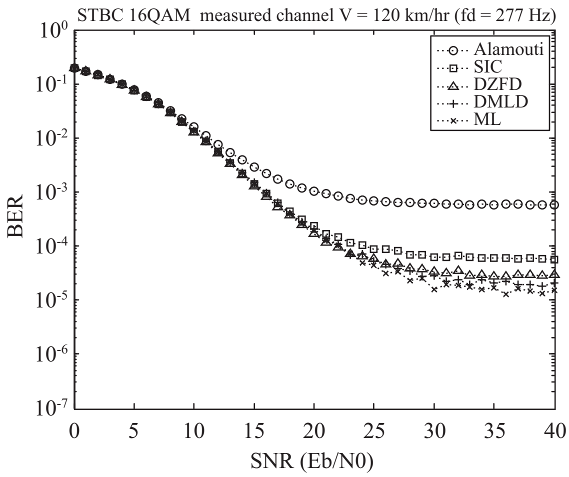

Fig. 9. BER Performance under the measured outdoor channel (120 km/hr).

2. Computational Complexity Analysis

In this study, we consider only the complex multiplication for each subcarrier. The complexity for the Alamouti can be deduced from Eq. (4). The number of the complex multiplication of the $\mathbf{H}^H Y$ is 4 for each subcarrier in the considered 2 by 1 STBC-OFDM scheme. Besides, the number of calculation for φ_k and the complex division is also 4. Therefore, Alamouti method requires 8 complex multiplications for each subcarrier. For the SIC method, it requires 7 additional complex multiplications for the Steps <4>~<6>, as compared with the Alamouti method. Note that one complex division is equivalent to two complex multiplications. Therefore, SIC method requires 15 complex multiplications for each subcarrier.

In Eq. (7), we can note that each constellation combination requires 6 complex multiplications. Therefore, ML method requires $(6 \times |C_M|^2) \times N$ complex multiplications for all the subcarriers where N is the number of subcarrier. The complexity calculations for the DMLD and DZFD can be illustrated in Eq. (14) and Eq. (15) respectively. In Eq. (12) and Eq. (13), the number of the complex multiplication for the $\mathbf{\Phi Y}$ is 4 and the number is 2 for the χ. Besides, $4 \times |C_M|$ multiplications are needed for the calculation of Eq. (14). However, only two complex divisions are required (four complex multiplications) for Eq. (15).

Table 3 summaries various detection methods in terms of the number of the multiplication. It reveals that the DZFD method can greatly reduce the computational complexity compared with the DMLD method and the ML method.

V. CONCLUSIONS

In this paper, we analyze several CCI cancellation decoders for the STBC-OFDM in a fast fading channel including the Alamouti method, SIC method, ML method, DMLD method and the DZFD method. A simple DZFD method is given to reduce the computational complexity of the DMLD method. We also conduct the outdoor channel measurements to verify the CCI cancellation ability of different methods. Results show that the ML method has the best performance under both the simulated and the measured scenarios. The DMLD and the DZFD have almost the same performance and outperform the SIC and the Alamouti detection method. Yet, the DZFD has the advantage of simple calculation than the DMLD method.

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