Alamouti Space-Time Coded OFDM Systems in Time- and Frequency-Selective Channels

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***Abstract*— We propose low-complexity equalizers for Alamouti space-time coded orthogonal frequency-division multiplexing (OFDM) systems in time- and frequency-selective channels, by extending the approach formerly proposed for single-antenna OFDM systems. The complexity of the proposed algorithm is linear in the number of subcarriers by exploiting the band structure of the frequency-domain channel matrix and a band LDL^H factorization. We design minimum mean squared error (MMSE) block linear equalizers (BLE) and block decision-feedback equalizers (BDFE) with and without windowing. We also develop a low-complexity algorithm that adaptively selects the useful bandwidth of the channel matrix. Simulation results show that the proposed algorithm produces a correct estimate of the bandwidth parameter.**

I. Introduction

In recent years, the spatial dimension in a wireless communication system has been explored by employing multiple transmit and/or receive antennas. This offers many benefits over the traditional single antenna system, including multiplexing gain which leads to higher capacity [2], and diversity gain which leads to more reliability [3]. Particularly the Alamouti scheme [1] for a system with two transmit antennas and one receive antenna is optimum in both the capacity and the diversity. Also, the Alamouti scheme does not require channel state information (CSI) at the transmitter and yields a low complexity maximum-likelihood decoding algorithm.

The increasing demand for higher date rates requires transmission over a broadband channel which is frequency-selective. As a result, intersymbol interference (ISI) is introduced, which severely degrades the system performance. Using OFDM can turn a frequency-selective channel into a set of parallel frequency-flat channels and renders simple one-tap equalization for each subcarrier. This allows one to apply the Alamouti scheme on each subcarrier separately to enable the spatial diversity. Note that next to this spatial diversity, it is also possible to enable frequency diversity.

However, to decouple the signals transmitted from different antennas and different subcarriers, the Alamouti space-time coded OFDM system requires the channel between individual transmit and receive antenna pairs to remain constant during two consecutive OFDM symbol periods. Doppler shifts due to high mobility however cause a time-selective or time-varying channel which destroys the orthogonality among antennas and subcarriers. The introduced interference severely degrades the performance of the one-tap equalizer [12]. To combat these time-varying distortions, nontrivial equalization techniques are required.

Contributions: We design low-complexity equalizers for Alamouti space-time coded OFDM systems in time- and frequency-selective channels based on the previous approach for single-antenna systems [5] [6]. Note that other low-complexity approaches such as [8] can be extended in a simpler fashion. The MMSE-BLE and MMSE-BDFE have a complexity that is linear in the number of subcarriers by exploiting the band structure of the frequency-domain channel matrix with band LDL^H factorization. Additionally, the minimum band approximation error (MBAE) sum-of-exponentials (SOE) window can be used to make the channel more banded and lower the error floor caused by the channel approximation error. We also develop an adaptive algorithm to further reduce the complexity which selects the bandwidth parameter adaptively with small performance loss. Note that based on ML detection, similar low-complexity equalizers have been developed in [16].

Notation: We use upper (lower) bold face letters to denote matrices (column vectors). $(\cdot)^*$, $(\cdot)^T$ and $(\cdot)^H$ represent conjugate, transpose, and complex conjugate transpose (Hermitian), respectively. $[\mathbf{A}]_{m,n}$ indicates the entry in the mth row and nth column of $\mathbf{A}$. We use the symbol $\circ$ to denote the Hadamard (element-wise) product and $\otimes$ to denote the Kronecker product. $E\{\cdot\}$ stands for the statistical expectation. $(a)_{\text{div}N}$ and $(a)_{\text{mod}N}$ are defined as the quotient and remainder after division of a by N. $\text{diag}(\mathbf{a})$ is a diagonal matrix with the vector $\mathbf{a}$ on the diagonal. $\mathbf{0}_{m\times n}$ represents the $m \times n$ all-zero matrix and $\mathbf{1}_{m\times n}$ the $m \times n$ all-one matrix. Finally, $\mathbf{I}_N$ denotes the $N \times N$ identity matrix and $\mathbf{F}$ denotes the unitary DFT matrix.

II. System Model

We consider a single-user OFDM system with two transmit antennas and one receive antenna as illustrated in Fig. 1. We assume that the two SISO channels from the two transmit antennas to the receive antenna are both time- and frequency-selective. They both have a maximum channel delay spread that is smaller than the OFDM cyclic prefix (CP) length L. Assume the OFDM system has N subcarriers, N_A of which are active. The remaining $N_V = N - N_A$ virtual subcarriers

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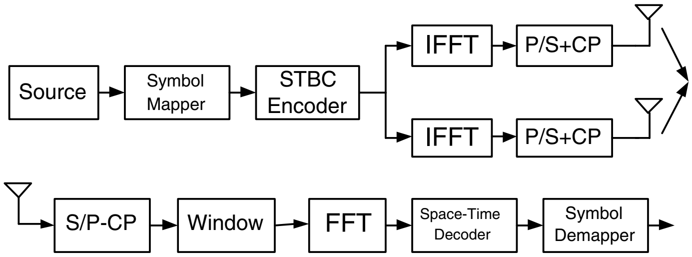

Fig. 1. Alamouti coded OFDM system

are used as frequency guard bands, with $N_V/2$ virtual carriers on both ends of the spectral band.

The bit streams at the transmitter are grouped and mapped into complex symbols. Since we assume the channel delay spread is smaller than the CP length L, after removing the CP at the receiver, it is enough to consider only the two consecutive OFDM symbols which constitute an Alamouti codeword. Assume $\mathbf{s}_i, i = 1, 2$ are the two consecutive OFDM symbols which can be written as

$$\mathbf{s}_i = [\mathbf{0}_{N_V/2\times 1}^T \;\; \tilde{\mathbf{s}}_i^T \;\; \mathbf{0}_{N_V/2\times 1}^T]^T \tag{1}$$

where the $\mathbf{0}$'s indicate the guard bands and $\tilde{\mathbf{s}}_i$ is the data vector of length $N_A = N - N_V$, which yields a set of data symbols with power σ_s^2.

During the first OFDM symbol period, $\mathbf{s}_1$ and $\mathbf{s}_2$ are sent from transmit antenna 1 and 2 respectively. Then, $-\mathbf{s}_2^*$ and $\mathbf{s}_1^*$ are sent from transmit antenna 1 and 2 respectively during the second OFDM symbol period. The IFFT operation converts the frequency-domain signal to a time-domain signal. After the parellel/serial conversion, the CP is added and the overall length-$(N+L)$ vectors are sent from the two transmit antennas simultaneously. At the receiver, after removing the CP, the received signals in two consecutive OFDM symbol periods can be written as

$$\mathbf{y}_1' = \mathbf{H}_{1,1}'\mathbf{F}^H\mathbf{s}_1 + \mathbf{H}_{2,1}'\mathbf{F}^H\mathbf{s}_2 + \mathbf{n}_1' \tag{2}$$

$$\mathbf{y}_2' = -\mathbf{H}_{1,2}'\mathbf{F}^H\mathbf{s}_2^* + \mathbf{H}_{2,2}'\mathbf{F}^H\mathbf{s}_1^* + \mathbf{n}_2' \tag{3}$$

where $\mathbf{y}_i'$ is the received $N \times 1$ vector in the ith symbol period, $\mathbf{H}_{i,j}'$ is the $N\times N$ time-domain channel matrix between transmit antenna i and the receive antenna in the jth symbol period, and $\mathbf{n}_i'$ is the $N \times 1$ circularly symmetric zero-mean white complex Gaussian random noise vector with covariance $E\{\mathbf{n}_i'\mathbf{n}_i'^H\} = \sigma_n^2\mathbf{I}_N$ and $E\{\mathbf{n}_i'\mathbf{n}_j'^H\} = \mathbf{0}_{N\times 1}$.

After the serial/parallel conversion, the FFT operation converts the received time-domain signal back to the frequency domain. Before the FFT, a time-domain receiver window is often used to make the frequency-domain channel matrix more banded [4]. In that case, we obtain

$$\mathbf{y}_1 = \mathbf{F}\mathbf{W}\mathbf{y}_1' + \mathbf{F}\mathbf{W}\mathbf{n}_1' \tag{4}$$

$$\mathbf{y}_2 = \mathbf{F}\mathbf{W}\mathbf{y}_2' + \mathbf{F}\mathbf{W}\mathbf{n}_2' \tag{5}$$

where $\mathbf{W} = \mathrm{diag}(\mathbf{w})$ with $\mathbf{w}$ the time-domain receiver window. Note that for classical OFDM (i.e., unwindowed), we have $\mathbf{W} = \mathbf{I}_N$.

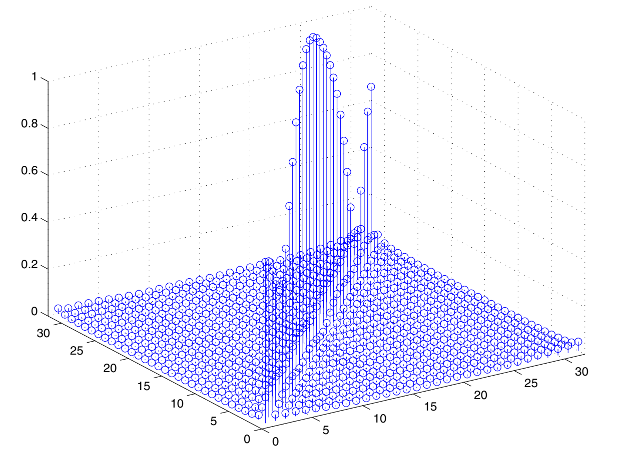

Fig. 2. An example of a frequency-domain channel matrix ($N = 32$)

Stacking $\mathbf{y}_1$ and $\mathbf{y}_2^*$ in one vector, we obtain

$$\begin{bmatrix} \mathbf{y}_1 \\ \mathbf{y}_2^* \end{bmatrix} = \begin{bmatrix} \mathbf{H}_{1,1} & \mathbf{H}_{2,1} \\ \mathbf{H}_{2,2}^* & -\mathbf{H}_{1,2}^* \end{bmatrix} \begin{bmatrix} \mathbf{s}_1 \\ \mathbf{s}_2 \end{bmatrix} + \begin{bmatrix} \mathbf{n}_1 \\ \mathbf{n}_2^* \end{bmatrix} \tag{6}$$

where $\mathbf{H}_{i,j} = \mathbf{F}\mathbf{W}\mathbf{H}_{i,j}'\mathbf{F}^H$ and $\mathbf{n}_i = \mathbf{F}\mathbf{W}\mathbf{n}_i'$.

The time-domain $N \times N$ channel matrix $\mathbf{H}_{i,j}'$ is defined as

$$[\mathbf{H}_{i,j}']_{m,n} = h_{i,j}[m-1, (m-n)_{\mathrm{mod}N}] \tag{7}$$

where $h_{i,j}[n, l]$ is the discrete-time equivalent impulse response of the continuous-time multipath channel $h_{i,j}(t, \tau)$

$$h_{i,j}[n, l] = h_{i,j}(nT_s, lT_s) \tag{8}$$

where $T_s = T/N$ is the sampling period and T is the duration of one OFDM symbol (without CP duration). When the channel is time-invariant, $\mathbf{H}_{i,j}'$ is a circulant matrix and $\mathbf{F}\mathbf{H}_{i,j}'\mathbf{F}^H$ is a diagonal matrix which makes the traditional simple OFDM one-tap equalizer possible. However, when the channel is time-varying, this is no longer true. The non-diagonal matrix $\mathbf{F}\mathbf{H}_{i,j}'\mathbf{F}^H$ gives rise to intercarrier interference (ICI). Fortunately, as shown in [9], [10], [8], the frequency-domain channel matrix $\mathbf{F}\mathbf{H}_{i,j}'\mathbf{F}^H$ is almost banded with the most significant elements around the main diagonal as shown in Fig. 2. This allows for low-complexity equalization architectures as proposed in [5], [4], [8]. Receiver windowing can be introduced before the FFT to make the frequency-domain channel matrix even more banded [4], [8], thereby improving the equalization performance.

In order to allow for low-complexity equalization, we approximate the frequency-domain channel matrix $\mathbf{H}_{i,j}$ by its banded version

$$\mathbf{B}_{i,j} = \mathbf{H}_{i,j} \circ \mathbf{\Theta}_Q \tag{9}$$

where $\mathbf{\Theta}_Q$ is the $N\times N$ Toeplitz matrix defined as $[\mathbf{\Theta}_Q]_{m,n} = 1$ for $|m-n| \le Q$ and $[\mathbf{\Theta}_Q]_{m,n} = 0$ for $|m-n| > Q$ [1].

[1]Due to the use of the $N_V/2$ guard carriers at both edges of the spectrum, only the middle N_A columns of $\mathbf{H}_{i,j}$ are useful. Hence, only the middle N_A columns of $\mathbf{\Theta}_Q$ are of any importance. So designing $\mathbf{\Theta}_Q$ to be banded or circularly banded makes no difference. For simplicity reasons, we design it to be banded.

The bandwidth parameter Q is used to control how many off-diagonal elements should be included to give a good approximation of the banded frequency-domain channel matrix. As shown later, tuning Q allows for a trade-off between equalizer complexity and performance. Q can be chosen according to some rules of thumb in [8]. Usually we take $1 \leq Q \leq 5$, which is much smaller than the number of subcarriers N.

Rewrite (6) as

$$\mathbf{y} = \mathbf{H}\mathbf{s} + \mathbf{n} \tag{10}$$

where $\mathbf{H}$ is a 2×2 block matrix of $N \times N$ approximately banded matrices with bandwidth parameter Q. Using a sepecific permutation matrix, we can now turn $\mathbf{H}$ into a $N \times N$ approximately banded block matrix of 2×2 matrices with block bandwidth parameter Q. Let us therefore define the permutation matrix $\mathbf{P}_{M,N}$ as an $MN \times MN$ matrix with 1's at the positions $\{(i+1, (i)_{\mathrm{div}M} + 1 + N(i)_{\mathrm{mod}M})\}_{i=0}^{MN-1}$ and 0's elsewhere (it is easy to verify that $\mathbf{P}_{M,N}^T \mathbf{P}_{M,N} = \mathbf{I}$). Left multiplying $\mathbf{y}$ in (10) with the permutation matrix $\mathbf{P}_{2,N}$, we obtain

$$\begin{aligned} \mathbf{y}_P &= \mathbf{P}_{2,N}\mathbf{y} = \mathbf{P}_{2,N}\mathbf{H}\mathbf{P}_{2,N}^T\mathbf{P}_{2,N}\mathbf{s} + \mathbf{P}_{2,N}\mathbf{n} \\ &= \mathbf{H}_P\mathbf{s}_P + \mathbf{n}_P \end{aligned} \tag{11}$$

where $\mathbf{H}_P = \mathbf{P}_{2,N}\mathbf{H}\mathbf{P}_{2,N}^T$, and $\mathbf{y}_P = \mathbf{P}_{2,N}\mathbf{y}$ and $\mathbf{s}_P = \mathbf{P}_{2,N}\mathbf{s}$ are the permuted received and transmitted signal, in which the data from the same subcarriers of different transmit antennas are grouped together in $\mathbf{s}_P$, and the received data from the same subcarriers in two consecutive OFDM symbol periods are grouped together in $\mathbf{y_P}$, similar to (10) in [13]. The matrix $\mathbf{H}_P$ is an $N \times N$ approximately banded block matrix of 2×2 matrices with block bandwidth parameter Q, and thus could be approximated by $\mathbf{H}_P \circ (\mathbf{\Theta}_Q \otimes \mathbf{1}_{2\times 2})$. However, for simplicity reasons, we view $\mathbf{H}_P$ as a $2N \times 2N$ approximately banded matrix with bandwidth parameter $2Q+1$, and approximate it by $\mathbf{H}_P \circ \mathbf{\Theta}_{2Q+1}$. Hence, we obtain the following input-output relation:

$$\mathbf{y}_P = (\mathbf{H}_P \circ \mathbf{\Theta}_{2Q+1})\mathbf{s}_P + \mathbf{n}_P = \mathbf{B}_P\mathbf{s}_P + \mathbf{n}_P \tag{12}$$

where the noise covariance matrix becomes

$$\begin{aligned} \mathbf{C}_{nn} &= E\{\mathbf{n}_P\mathbf{n}_P^H\} \\ &= \sigma_n^2 \mathbf{P}_{2,N} \begin{pmatrix} \mathbf{F}\mathbf{W}\mathbf{W^H}\mathbf{F^H} & \mathbf{0} \\ \mathbf{0} & (\mathbf{F}\mathbf{W}\mathbf{W^H}\mathbf{F^H})^* \end{pmatrix} \mathbf{P}_{2,N}^T. \end{aligned} \tag{13}$$

If no windowing is applied before the FFT, $\mathbf{C}_{nn} = \sigma_n^2\mathbf{I}_{2N}$.

III. Space-Time Decoding

In this section, we extend the low-complexity equalizers designed for SISO-OFDM in [4] to the MISO system using the Alamouti coding scheme. For simplicity, we assume that the receiver has perfect CSI. In practice, the techniques developed in [6] can be used to estimate the channel. First we focus on the MMSE-BLE to estimate the transmitted symbols which outperforms other linear approaches [11]. The non-banded MMSE-BLE requires a complexity of $O(N^3)$, which makes it impractical for real systems with a large number of subcarriers (in the DVB-T and DVB-H standard N can be up to 6816). In order to reduce the complexity, the nearly banded structure of the frequency-domain channel matrix is exploited, by using the band LDL^H factorization [5]. By designing a good window, the channel matrix can be made even more banded. We use the MBOE-SOE windowing developed in [4] which can be written as $[\mathbf{w}]_n = \sum_{q=-Q}^{Q} b_q \exp(j2\pi qn/N)$, where b_q is designed to reduce the band approximation error. As shown in [4], in that case the covariance matrix of the windowed noise $\mathbf{C}_{nn}$ is also banded, which still enables the low-complexity MMSE-BLE.

Next, we derive the MMSE-BDFE with and without windowing which can achieve a better performance than the BLE. Finally, we propose an adaptive equalizer which selects the bandwidth parameter Q adaptively according to the instantaneous channel condition in order to further reduce the complexity with small performance loss.

We assume $\underline{\mathbf{B}}_P$ is the frequency-domain channel matrix when $\mathbf{W} = \mathbf{I}_N$ (no windowing is applied), and $\mathbf{B}_P$ is the frequency-domain channel matrix when $\mathbf{W} = \mathrm{diag}(\mathbf{w})$ (MBAE-SOE windowing is used before the FFT).

A. Banded Linear Equalizers

The MMSE-BLE without windowing ($\mathbf{W} = \mathbf{I}_N$) can be written as

$$\hat{\mathbf{s}}_P = \mathbf{G}_{\mathrm{MMSE-BLE}}\mathbf{y}_P \tag{14}$$

$$\mathbf{G}_{\mathrm{MMSE-BLE}} = \underline{\mathbf{L}}^{-H}\ \underline{\mathbf{D}}^{-1}\ \underline{\mathbf{L}}^{-1}\underline{\mathbf{B}}_P^H \tag{15}$$

$$\underline{\mathbf{B}}_P^H\underline{\mathbf{B}}_P + \sigma_n^2/\sigma_s^2\mathbf{I}_{2N} = \underline{\mathbf{L}}\ \underline{\mathbf{D}}\ \underline{\mathbf{L}}^H \tag{16}$$

where the banded lower triangular matrix $\underline{\mathbf{L}}$ and the diagonal matrix $\underline{\mathbf{D}}$ can be computed by the low-complexity band LDL^H factorization.

When the MBAE-SOE window is used ($\mathbf{W} = \mathrm{diag}(\mathbf{w})$), the windowed MMSE-BLE becomes

$$\hat{\mathbf{s}}_P = \mathbf{G}_{\mathrm{W-MMSE-BLE}}\mathbf{y}_P \tag{17}$$

$$\mathbf{G}_{\mathrm{W-MMSE-BLE}} = \mathbf{L}^{-H}\mathbf{D}^{-1}\mathbf{L}^{-1}\mathbf{B}_P^H \tag{18}$$

$$\mathbf{B}_P^H\mathbf{B}_P + 1/\sigma_s^2\mathbf{C}_{nn} = \mathbf{L}\mathbf{D}\mathbf{L}^H \tag{19}$$

where $\mathbf{C}_{nn}$ is the windowed noise covariance matrix expressed in (13). As before, the banded lower triangular matrix $\mathbf{L}$ and the diagonal matrix $\mathbf{D}$ can be computed by the low-complexity band LDL^H factorization.

Since $\hat{\mathbf{s}}_P$ is only the permuted version of $\hat{\mathbf{s}}$, $\hat{\mathbf{s}}$ can be recovered by $\hat{\mathbf{s}} = \mathbf{P}_{2,N}^T\hat{\mathbf{s}}_P$.

Complexity: Similar to the analysis in [4], the MMSE-BLE without windowing in (14) requires approximately $(32Q^2 + 76Q + 34)N$ complex operations per OFDM symbol, and the windowed MMSE-BLE in (17) requires approximately $(32Q^2+80Q+37)N$ complex operations per OFDM symbol[2]. Since usually the bandwidth parameter Q is chosen to be very small, the computational complexity for the banded MMSE-BLE is linear in N, much smaller than the previously proposed

[2]There are still a few very small entries in $\mathbf{B}_P$ due to the fact that we approximate a banded block matrix with block bandwidth parameter Q by a banded matrix with bandwidth parameter $2Q+1$. These could possibly be exploited to reduce the complexity even further.

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equalizers which are quadratic [10] or even cubic [11] in the number of subcarriers N.

B. Banded Decision Feedback Equalizers

The BDFE also exploits the LDL^H factorization algorithm to achieve a low-complexity equalizer [4]. The feedforward filter $\mathbf{F}_F$ and feedback filter $\mathbf{F}_B$ are designed according to the MMSE approach [14]. $\mathbf{F}_B$ is designed to be strictly upper triangular, such that successive cancellation can be used during the feedback process [4] [15]. The MMSE-BDFE without windowing ($\mathbf{W} = \mathbf{I}_N$) can be written as

$$\mathbf{F}_B = \underline{\mathbf{L}}^H - \mathbf{I}_{2N} \quad (20)$$

$$\mathbf{F}_F = \underline{\mathbf{L}}^H \mathbf{G}_{\mathrm{MMSE-BLE}} = \underline{\mathbf{D}}^{-1} \underline{\mathbf{L}}^{-1} \underline{\mathbf{B}}_P^H \quad (21)$$

The MMSE-BDFE with windowing ($\mathbf{W} = \mathrm{diag}(\mathbf{w})$) can be written as

$$\mathbf{F}_B = \underline{\mathbf{L}}^H - \mathbf{I}_{2N} \quad (22)$$

$$\mathbf{F}_F = \underline{\mathbf{L}}^H \mathbf{G}_{\mathrm{W-MMSE-BLE}} = \underline{\mathbf{L}}^H \mathbf{L}^{-H} \mathbf{D}^{-1} \mathbf{L}^{-1} \mathbf{B}_P^H \quad (23)$$

Complexity: The MMSE-BDFE without windowing has the same complexity as the MMSE-BLE without windowing which requires $(32Q^2 + 76Q + 34)N$ complex operations per OFDM symbol, and the windowed MMSE-BDFE has a complexity of $(64Q^2 + 128Q + 65)N$ complex operations per OFDM symbol.

C. Banded Adaptive Equalizer

Even for a high Doppler spread channel, the channel is not always changing significantly within certain OFDM intervals. The channel bandwidth parameter Q can therefore be chosen adaptively according to the instantaneous channel variation. A similar idea has been proposed in [16].

The algorithm can be summarized as follows:

1) Define a threshold α and a maximum bandwidth parameter Q_{max}.
2) Compute the energy in each pair diagonal of every frequency-domain channel matrix $\mathbf{H}_{i,j}$ without windowing,
   $P_Q = \sum_{i,j,q} |[\mathbf{H}_{i,j}]_{q,q\pm Q}|^2, \forall 0 \leq Q \leq Q_{max}$
3) $Q = 0, P = P_0$;
   while $P / \sum_{i=0}^{Q_{max}} P_i < \alpha$ and $Q \leq Q_{max}$ repeat
      $Q = Q + 1$;
      $P = P + P_Q$;
   end
4) Use the obtained Q as bandwidth parameter for windowing and equalization.

IV. Simulation Results

In this section, the proposed decoding algorithms are examined and compared by simulation. We consider an Alamouti space-time coded OFDM system with $N = 128$ and $N_A = 96$. The maximum channel delay spread and the CP length are the same and equal to $L = 32$ (we use the WLAN scenario which is analogous to that of [16]). The two SISO channels from the transmit antennas to the receive antenna are assumed to be independent and Rayleigh distributed, with an exponential power delay profile, and Jakes' Doppler spectrum. The channel magnitudes $h_{i,j}[n,l]$ are assumed to be circularly symmetric zero-mean uncorrelated complex Gaussian random variables. The complex symbols are assumed to be quaternary phase-shift keying (QPSK) only. We consider a high mobility case where the normalized Doppler frequency is $f_d/\Delta f = 0.12$ with f_d the maximum Doppler frequency shift and $\Delta f = 1/T$ the subcarrier spacing. As argued in [4], this value generally represents a high Doppler spread condition.

Fig. 3 compares the BER performance of the MMSE-BLE and the MMSE-BDFE for different $Q^{'}$s (the non-banded case corresponds to $Q = N - 1$). The classical Alamouti decoding fails completely due to the high mobility which destroys the orthogonal structure of the channel matrix. The performance is getting better as Q increases. When only the MMSE-BLE is considered, the non-banded BLE has the best performance. However, the computational complexity is cubic in the number of subcarriers [11], compared to the linear behavior for the proposed banded equalizers. BDFE outperforms BLE when they have the same bandwidth parameter Q. All the banded equalizers have an error floor due to the band approximation error of the channel. The error floor can be reduced by increasing Q.

Fig. 4 shows the BER performance of the equalizers with MBAE-SOE windowing compared to the non-windowed case. It is shown that windowing can greatly improve the system performance by reducing the error floor, since windowing makes the channel matrix more banded. Especially the BDFE with windowing outperforms the non-banded MMSE-BLE. This is in contrast to the methods proposed in [16], which can not reach a better performance than the non-banded MMSE-BLE equalizer. It is worth noting that the windowed MMSE-BLE/BDFE with $Q = 1$ has a better performance than the non-windowed MMSE-BLE/BDFE with $Q = 2$, but the complexity of the windowed MMSE-BLE/BDFE with $Q = 1$ is approximately $54\%/82\%$ of the one for the non-windowed MMSE-BLE/BDFE with $Q = 2$. Meanwhile, the windowed MMSE-BLE outperforms the non-windowed MMSE-BDFE when having the same bandwidth parameter Q. Thus, appropriate window design realizes a good trade-off between system performance and computational complexity.

Fig. 5 shows the BER performance of the adaptive algorithm for $\alpha = 0.991$ and $Q_{max} = 2$ using MMSE BDFE with windowing. The simulation shows that $0.7\%/74.5\%/24.8\%$ of the total number of OFDM symbols is detected with $Q = 0/Q = 1/Q = 2$ respectively. Since the computational complexity for $Q = 0$ / $Q = 1$ is 11% / 45% compared to the computational complexity for $Q = 2$ and step 2 of the algorithm only requires $2(2Q_{max} + 1)N$ additional operations per OFDM symbol, the overall adaptive approach saves 40.7% computational complexity with a small performance loss compared to BDFE with $Q = 2$.

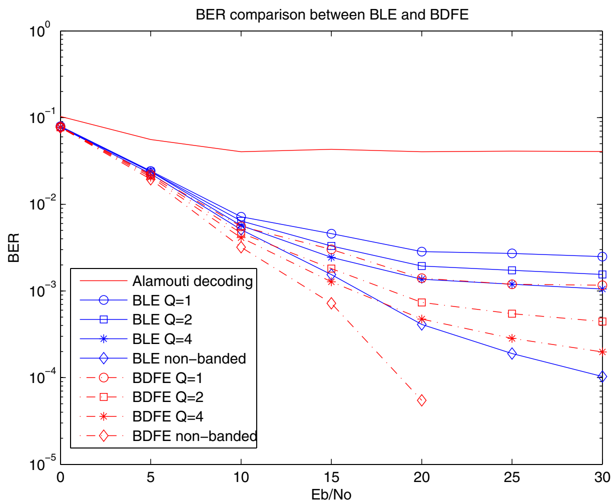

Fig. 3. BER comparison between BLE and BDFE

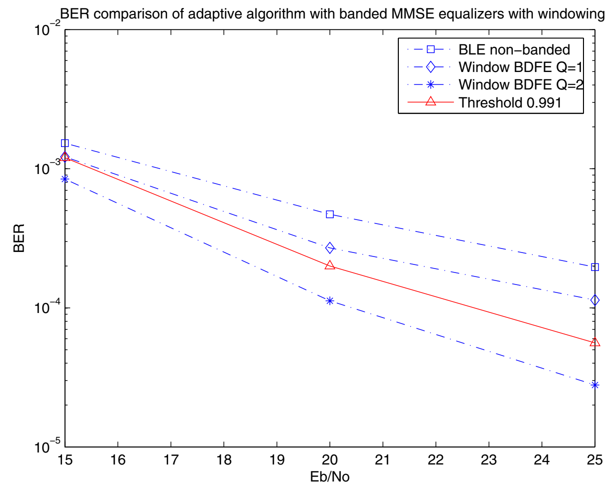

Fig. 5. BER comparison of adaptive algorithm with windowed BDFE

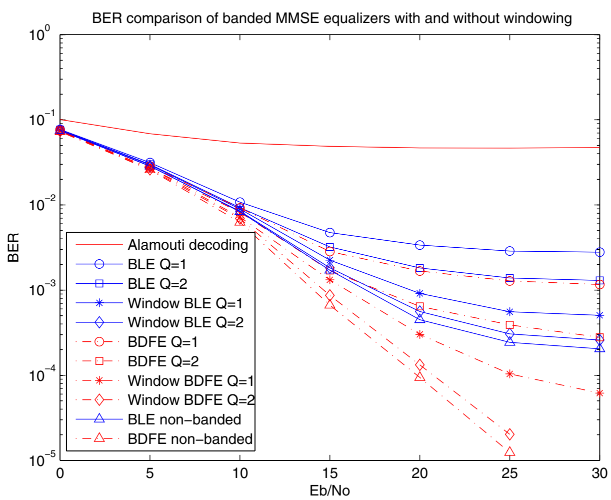

Fig. 4. BER comparison of banded MMSE equalizers with and without windowing

V. Conclusion

We have designed banded MMSE equalizers for Alamouti space-time coded OFDM systems in time- and frequency-selective channels. By using the band LDL^H factorization on the banded frequency-domain channel matrix, the equalizers have a low complexity which is linear in the number of subcarriers. The MBAE-SOE windowing can be applied before the FFT at the receiver to make the channel more banded, which reduces the error floor caused by the channel approximation error. An adaptive algorithm has also been developed to further reduce the complexity, which selects the bandwidth parameter Q adaptively with small performance loss.

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