Journal of Network and Computer Applications 34 (2011) 116–120

Contents lists available at ScienceDirect

Journal of Network and Computer Applications

journal homepage: www.elsevier.com/locate/jnca

A frequency domain subspace blind channel estimation method for trailing zero OFDM systems

Song Wang [a,1], Jinli Cao [a,2], Jiankun Hu [b,*]

[a] School of Engineering and Mathematical Sciences, La Trobe University, VIC 3086, Australia
[b] School of Computer Science and Information Technology, RMIT, VIC 3001, Australia

ARTICLE INFO

Article history:
Received 16 April 2010
Received in revised form
20 August 2010
Accepted 20 August 2010

Keywords:
Orthogonal frequency division multiplexing
Blind channel estimation
Trailing zero
Subspace

ABSTRACT

The technique of orthogonal frequency division multiplexing (OFDM) is widely used for high-speed data transmission in indoor and outdoor communication networks. Trailing zero (TZ) OFDM systems possess some favorable features compared to conventional cyclic prefixed OFDM systems. In this paper, we propose a blind channel estimation method for TZ-OFDM systems in the frequency domain. The significance of the proposed method is threefold. First, it gives an insight into the characteristics of TZ-OFDM systems from a frequency domain perspective. The characteristics revealed are not clearly manifest in the time domain. Second, by exploiting the desirable property brought by the trailing zeros and making use of the frequency domain samples available at the TZ-OFDM receiver, the frequency domain subspace based blind channel estimator is straightforward in concept and algorithm derivation as opposed to a time domain formulation. Third, the proposed method explicitly addresses the issue of unknown channel length. The new method is theoretically proven and numerically shown to work effectively for channels of unknown length. Simulation results demonstrate the satisfactory performance of the proposed method as compared to the existing methods and its robustness to channel overestimation.

1. Introduction

Orthogonal frequency division multiplexing (OFDM) is an attractive multicarrier transmission technique for combating frequency-selective distortion and narrowband interference (Sklar, 2001; Prasad, 2004) in indoor and outdoor communication networks (Thompson et al., 2006). Conventional OFDM systems apply a cyclic prefix (CP) to eliminating inter-block interference (IBI). However, CP-OFDM systems fail to recover transmitted symbols if there is a channel spectral null located on a subcarrier (Wang and Giannakis, 2001). To address this issue, the CP is replaced by trailing zeros (TZ), added to each transmitted block. As an appealing alternative to CP-OFDM, not only are TZ-OFDM systems insensitive to channel spectral nulls, but they also enjoy maximum diversity gain (Wang and Giannakis, 2001). Despite the seeming relationship between CP- and TZ-OFDM systems, the results obtained for the former cannot be carried over directly to the latter.

To detect transmitted data correctly, reliable estimation of time-dispersive channels is important for CP- and TZ-OFDM systems. There are two classes of channel estimation methods, training-based and blind. In the former, training sequences (also called pilots) are transmitted to enable channel estimation at the receiver. However, training sequences consume system bandwidth, thus reducing network spectral efficiency. In addition, to track channel variations in wireless applications, pilot symbols have to be transmitted periodically, causing further loss in system throughput. Compared to training-based schemes, blind channel estimation methods do not suffer from these drawbacks. Indeed, blind identification techniques have been recognized as a bandwidth efficient means (Giannakis et al., 2001) for increasing network capacity. Since the celebrated work of Moulines et al. (1995), the subspace (SS) based estimation framework has played a critical role in blind channel identification for single-carrier systems (e.g., Giannakis et al., 2001; Tsai and Tseng, 2008) and for OFDM systems (e.g., Muquet et al., 2002a,b). It is also useful in remotely continuous monitoring of medical patients where fast and accurate estimation of large volume of medical images such as ECG is critical (e.g., Hu and Han, 2009).

In this paper, we propose a frequency domain subspace approach to blind channel estimation for TZ-OFDM systems. Although the frequency domain is particularly suitable for analyzing TZ-OFDM systems (Manton, 2002), due to the fact that

* Corresponding author. Tel.: +61 3 9925 9793; fax: +61 3 9662 1617.
E-mail addresses: song.wang@latrobe.edu.au (S. Wang), j.cao@latrobe.edu.au (J. Cao), jiankun@cs.rmit.edu.au (J. Hu).
[1] Tel.: +61 3 9479 3744; fax: +61 3 9471 0524.
[2] Tel.: +61 3 9479 3035; fax: +61 3 9470 3060.

1084-8045/$ - see front matter © 2010 Elsevier Ltd. All rights reserved.
doi:10.1016/j.jnca.2010.08.009

there is no trivial link in terms of problem characterization between the time and frequency domains, to the best of our knowledge, there has been no frequency domain blind TZ-OFDM channel estimation method previously available. The significance of our proposed frequency domain method is threefold. Firstly, useful features of TZ-OFDM systems can be better revealed in the frequency domain than in the time domain. Not only do we explain from a frequency domain viewpoint why TZ-OFDM systems are not sensitive to channel spectral nulls, but we also show that the spectrum of source symbols and the unknown channel gets spread out in the frequency domain, a feature which we utilise in the estimation of the unknown channel. Secondly, capitalizing on the discrete Fourier transform (DFT) operation performed in OFDM systems facilitates channel estimation in the frequency domain. Indeed, the proposed algorithm directly use the frequency domain samples available at the TZ-OFDM receiver to carry out channel estimation. Thirdly, unlike many time domain blind estimation methods, prior knowledge of channel length is no longer a prerequisite in the proposed frequency domain algorithm, which then leads to robust channel length estimation. That is, the proposed method is robust against channel length overestimation.

In this paper, we provide an insight into the characteristics of TZ-OFDM systems from a frequency domain perspective. The characteristics revealed are not clearly manifest in the time domain. For example, it is an important property of TZ-OFDM systems that symbol detection is always guaranteed irrespective of channel spectral nulls; this characteristic of TZ-OFDM systems unfolds in a straightforward manner in the frequency domain as opposed to the time domain. By exploiting the desirable property brought by the trailing zeros, and making use of the frequency domain samples available from the DFT at the receiver, we present a frequency domain realization of subspace decomposition, the central idea for subspace based blind channel identification. Compared to the SS time domain algorithm in Scaglione et al. (1999) and Muquet et al. (2002b), the proposed method is straightforward in concept and algorithm derivation and explicitly addresses the issue of unknown channel length (e.g., in a mobile scenario). Furthermore, the new method is theoretically proven and numerically shown to work effectively for channels of unknown length. In this case, an overestimated channel vector is obtained, and the proposed method is robust to channel over-estimation. Simulations show that our method performs as well as the SS time domain algorithm in Muquet et al. (2002b) and outperforms the correlation matching algorithm in Backx et al. (2007) for a wide range of SNR.

The rest of the paper is organized as follows. Section 2 presents the TZ-OFDM system model. Section 3 discusses the frequency domain properties of TZ-OFDM systems and derives the frequency domain blind estimation algorithm. Section 4 shows simulation results, where we compare the proposed algorithm with the existing methods. The conclusion is given in Section 5.

Notation: Bold uppercase (lowercase) letters are used for matrices (vectors). Superscripts $(\cdot)^T$ and $(\cdot)^{\mathcal{H}}$ denote the transpose and Hermitian transpose, respectively. $E\{\cdot\}$ represents the mathematical expectation. $\|\cdot\|$ denotes the Euclidean norm.

2. System model

It is known that at the transmitter side, a TZ-OFDM system operates in the same manner as a CP-OFDM system except that trailing zeros are used as guard intervals. Specifically, each source information block of P symbols is first modulated by the IDFT to form the complex baseband time-domain signal block $\mathbf{s}_m = [s_m(0)\ s_m(1)\ \cdots\ s_m(P-1)]^T$, where m is the block index. A sequence of trailing zeros of length T is then added to each $\mathbf{s}_m$ so that the total number of data samples per block for transmission is $N=P+T$. The information symbols padded with zeros are sent over a channel modeled by a finite impulse response (FIR) filter $\mathbf{h} = [h_0\ h_1\ \cdots\ h_L]^T$, of which the channel order L may be unknown but upper bounded by T, the length of trailing zeros.

Assuming perfect timing and carrier synchronization, the discrete-time signal in the mth received block is

$$x_m(n) = s_m(n) * h_n + w_m(n) \tag{1}$$

where $w_m(n)$ is the additive white Gaussian noise (AWGN) with zero mean and variance σ^2 (σ may or may not be known), and uncorrelated with $s_m(n)$. The symbol * in (1) represents linear convolution. Unlike CP-OFDM systems, the receiver of a TZ-OFDM system does not discard the guard interval. That is, all of the N samples in each received block go through the DFT demodulation, yielding the frequency domain observation block

$$\mathbf{y}_m = [Y_m(0)\ Y_m(1)\ \cdots\ Y_m(N-1)]^T \tag{2}$$

where $Y_m(k)$, $k=0,1,\ldots,N-1$, are the DFT output of the mth received block.

The objective of the blind channel estimation problem at hand is to determine the unknown channel vector $\mathbf{h}$ based only on frequency domain observation blocks $\mathbf{y}_m$.

3. Frequency domain subspace based blind channel estimation

Due to the trailing zeros inserted between transmitted blocks, not only is IBI removed, but also the convolutive channel is converted to a multiplicative one with the length of the transmitted block and channel vector increased by zero padding. This makes the linear convolution in (1) equivalent to the N-point circular convolution of $s_m(n)$ with h_n, which leads to the following frequency domain representation of (1) given that the frequency domain block $\mathbf{y}_m$ [see (2)] is available at the receiver:

$$Y_m(k) = S_m(k)H(k) + W_m(k), \quad k = 0,1,\ldots,N-1 \tag{3}$$

where $H(k)$, $S_m(k)$ and $W_m(k)$ are, respectively, the channel's frequency response, the frequency samples of $s_m(n)$ and $w_m(n)$. Note that the point-wise multiplication in (3) for TZ-OFDM systems differs from CP-OFDM systems in the size of the DFT (Wang and Manton, 2009). Alternatively, (3) can be expressed in vector-matrix form

$$\mathbf{y}_m = \mathbf{\Lambda}_H \mathbf{W}_{N\times P} \mathbf{s}_m + \mathbf{w}_m \tag{4}$$

where $\mathbf{\Lambda}_H = \mathrm{diag}\{H(0), H(1), \ldots, H(N-1)\}$, $\mathbf{W}_{N\times P}$ denotes the $N\times P$ matrix formed by the first P columns of the orthonormal DFT matrix $\mathbf{W}_N$ (so that $\mathbf{W}_N^{\mathcal{H}}\mathbf{W}_N = \boldsymbol{I}$), and $\mathbf{w}_m$ is defined analogously to (2).

As (3) and (4) show, the trailing zeros introduce spectral redundancy to the TZ-OFDM systems, and as a result the spectrum of $s_m(n)$, $n=0,1,\ldots,P-1$, in each block and the spectrum of the unknown channel are spread out over the frequency domain to form $S_m(k)$ and $H(k)$, respectively, $k=0,1,\ldots,N-1$. It is the spectral redundancy of the spread spectrum of $s_m(n)$ and h_n that underpins the subsequent channel identification process.

Since the channel is of order L, the channel's z-transform $H(z) = \sum_{i=0}^{L} h_i z^{-i}$ has L roots. Based on the equivalence between the DFT and z-transform, there can be at most L spectral nulls among $H(k)$, $k=0,1,\ldots,N-1$, which means that there can only be a maximum of L zero diagonal elements in $\mathbf{\Lambda}_H$. Hence, the rank of $\mathbf{\Lambda}_H$ is at least P as $L \le T$, leading to $\mathrm{rank}(\mathbf{\Lambda}_H \mathbf{W}_{N\times P}) = P$. Let the $N\times N$ autocorrelation matrix of $\mathbf{y}_m$ be $\mathbf{R}_y = E\{\mathbf{y}_m \mathbf{y}_m^{\mathcal{H}}\}$. Assuming that the autocorrelation matrix of $\mathbf{s}_m$, $\mathbf{R}_s = E\{\mathbf{s}_m \mathbf{s}_m^{\mathcal{H}}\}$, is of full rank, it is readily seen that $\mathbf{R}_y$ has rank P.

Remark. From (4), we can see that with $(\mathbf{\Lambda}_H \mathbf{W}_{N\times P})$ being of full column rank, the transmitted symbols $\mathbf{s}_m$ can always be recovered even when the channel has spectral nulls among the N frequencies. This is an important characteristic of TZ-OFDM systems, which is an advantage over CP-OFDM systems. This characteristic of TZ-OFDM systems is readily revealed through the above frequency domain analysis thanks to the distinctive structure of $\mathbf{\Lambda}_H$, which shows the channel's spectrum, including spectral nulls. By contrast, such a characteristic cannot be revealed in such a straightforward fashion in the time domain.

Performing subspace decomposition (Moulines et al., 1995) on $\mathbf{R}_y$ gives rise to the noise subspace, spanned by the columns of the matrix $\mathbf{U}=[\mathbf{u}^{(1)}\ \mathbf{u}^{(2)}\ \cdots\ \mathbf{u}^{(T)}]$, where $\mathbf{u}^{(j)}$ are known to be the eigenvectors associated with T smallest eigenvalues of $\mathbf{R}_y$. As the orthogonal complement of the noise subspace, known as the signal subspace, is spanned by the columns of $(\mathbf{\Lambda}_H \mathbf{W}_{N\times P})$, we have

$$(\mathbf{u}^{(j)})^{\mathcal{H}} \mathbf{\Lambda}_H \mathbf{W}_{N\times P} = \mathbf{0}, \quad j=1,2,\ldots,T \tag{5}$$

Expanding the left-hand side of (5) yields

$$(\mathbf{u}^{(j)})^{\mathcal{H}} \mathbf{\Lambda}_H \mathbf{W}_{N\times P} = h \mathbf{\Lambda}_U^{(j)} \mathbf{W}_{N\times P}$$

where $h=[H(0)\ H(1)\ \cdots\ H(N-1)]$, and $\mathbf{\Lambda}_U^{(j)}$ denotes the diagonal matrix whose entries are the elements of $(\mathbf{u}^{(j)})^{\mathcal{H}}$.

Now we shall show that h is uniquely determined (up to a complex scale factor) by solving

$$h \mathbf{\Lambda}_U^{(j)} \mathbf{W}_{N\times P} = \mathbf{0}, \quad j=1,2,\ldots,T \tag{6}$$

Since h is made up of the DFT of the channel coefficients, the channel vector $\mathbf{h}$ can be obtained by taking the first $L+1$ elements of the IDFT of h.

Theorem 1. *The nontrivial solution h to (6) uniquely (up to a constant scalar) determines the channel's frequency response $H(k)$, $k=0,1,\ldots,N-1$.*

Proof. Without loss of generality, we assume that $H(z)=\sum_{i=0}^{L} h_i z^{-i}$ has L distinct roots v_i, $i=1,2,\ldots,L$, each of which is used to form an $N\times 1$ Vandermonde vector $\mathbf{v}_i=[1\ v_i^{-1}\ v_i^{-2}\ \cdots\ v_i^{-(N-1)}]^{\mathcal{H}}$. (The case of multiple roots of $H(z)$ can be handled by the generalized Vandermonde vectors, Golub and Van Loan, 1996.) It is a standard result that the set $[\mathbf{v}_i^{\mathcal{H}}: i=1,\ldots,L]$ is a basis for the left null space of $\mathcal{H}$, where $\mathcal{H}$ is the $N\times P$ Toeplitz matrix with the first row equal to $[h_0\ 0\ \cdots 0]$ and the first column equal to $[h_0\ \cdots h_L\ 0\ \cdots\ 0]^T$. It can be shown that

$$\mathbf{W}_N \mathcal{H} = \mathbf{\Lambda}_H \mathbf{W}_{N\times P} \tag{7}$$

It follows from (7) and $\mathbf{v}_i^{\mathcal{H}}\mathcal{H}=\mathbf{0}$ that $\boldsymbol{v}_i^{\mathcal{H}}\mathbf{\Lambda}_H\mathbf{W}_{N\times P}=\mathbf{0}$, where $\boldsymbol{v}_i$ is the vector comprising the N-point DFT of $\mathbf{v}_i$. Further, the set $[\boldsymbol{v}_i^{\mathcal{H}}: i=1,\ldots,L]$ forms a basis for the left null space of $(\mathbf{\Lambda}_H\mathbf{W}_{N\times P})$. Given that $(\mathbf{u}^{(j)})^{\mathcal{H}}$ is in the null space of $(\mathbf{\Lambda}_H\mathbf{W}_{N\times P})$ [see (5)], $(\mathbf{u}^{(j)})^{\mathcal{H}}$ can be expressed as a linear combination of $\boldsymbol{v}_i^{\mathcal{H}}$. That is,

$$(\mathbf{u}^{(j)})^{\mathcal{H}} = b_1^{(j)}\boldsymbol{v}_1^{\mathcal{H}} + b_2^{(j)}\boldsymbol{v}_2^{\mathcal{H}} + \cdots + b_L^{(j)}\boldsymbol{v}_L^{\mathcal{H}} \tag{8}$$

where the coefficients $b_l^{(j)}$ are not all zero.

Suppose there exists another $1\times N$ vector $\overline{h}$ such that $\overline{h}\mathbf{\Lambda}_U^{(j)}\mathbf{W}_{N\times P}=\mathbf{0}$, $j=1,2,\ldots,T$. This leads to

$$(\mathbf{u}^{(j)})^{\mathcal{H}}\mathbf{W}_N\overline{\mathcal{H}} = \mathbf{0}, \quad j=1,2,\ldots,T$$

where $\overline{\mathcal{H}}$ is defined in the same way as $\mathcal{H}$ except to replace $h_0,\ldots,h_L$ with $\overline{h}_0,\ldots,\overline{h}_L$. From this equation and (8), by taking the IDFT we get $(b_1^{(j)}\mathbf{v}_1^{\mathcal{H}}+b_2^{(j)}\mathbf{v}_2^{\mathcal{H}}+\cdots+b_L^{(j)}\mathbf{v}_L^{\mathcal{H}})\overline{\mathcal{H}}=\mathbf{0}$. It follows that $\mathbf{v}_i^{\mathcal{H}}$, $i=1,\ldots,L$, constitute the basis vectors of the left null space of $\overline{\mathcal{H}}$. This means that the roots v_i, $i=1,2,\ldots,L$, of $H(z)$ are also the roots of $\overline{H}(z)=\sum_{i=0}^{L}\overline{h}_i z^{-i}$. Thus, $\overline{h}=\alpha h$, where α is a nonzero constant. □

In practice, when only finite data are available we use the sample covariance matrix $\hat{\mathbf{R}}_y$ of frequency domain blocks $\mathbf{y}_m$:

$$\hat{\mathbf{R}}_y = \frac{1}{N_s}\sum_{m=1}^{N_s}\mathbf{y}_m\mathbf{y}_m^{\mathcal{H}}$$

The blind channel estimator is then obtained in the least squares sense by solving

$$\hat{h} = \arg\min_{\|h\|=1}\left(\sum_{j=1}^{T}\|h\mathbf{\Lambda}_U^{(j)}\mathbf{W}_{N\times P}\|^2\right) \tag{9}$$

whose solution $(\hat{h})^{\mathcal{H}}$ is the eigenvector associated with the smallest eigenvalue of $\sum_{j=1}^{T}(\mathbf{\Lambda}_U^{(j)}\mathbf{W}_{N\times P})(\mathbf{\Lambda}_U^{(j)}\mathbf{W}_{N\times P})^{\mathcal{H}}$. Since the $1\times N$ vector $\hat{h}$ consists of the channel's frequency response, taking the IDFT of $\hat{h}$ yields the estimated channel coefficients.

Remarks. (1) The IDFT of $\hat{h}$ results in an augmented version of the channel estimate. When the channel order is known, channel coefficients are given by the first $L+1$ elements of the IDFT of $\hat{h}$. When the channel order is unknown, we need to take the first $T+1$ entries of the IDFT result, yielding an overestimate of the channel coefficients. As shown by the simulations in Section 4, the proposed frequency domain method is robust to channel over-estimation.

(2) In the absence of noise or using the true autocorrelation matrix, the proposed method achieves exact channel identification. In other words, with the channel order known, exact channel coefficients (up to an inherent scale factor) are obtained; with the channel order unknown, the solution becomes the true channel coefficients padded with $T-L$ zeros.

Although we have derived the algorithm for TZ-OFDM systems without virtual carriers, the proposed method can be readily adapted for the situation where virtual carriers are present. Suppose that there are J nonvirtual carriers among P subcarriers. In this case, the algorithm proceeds along a similar line, where we only include those nonzero entries in $\mathbf{s}_m$ [see (4)]. Accordingly, the matrix $\mathbf{W}_{N\times P}$ in (4) is to be changed to $\mathbf{W}_{N\times J}$, which leads the autocorrelation matrix $\mathbf{R}_y$ to be of rank J. Lastly, $\mathbf{W}_{N\times P}$ should be replaced by $\mathbf{W}_{N\times J}$ in the equations, such as (5), (6) and (9), leading up to the channel estimation.

4. Simulations

In this section, we evaluate the performance of the proposed frequency domain subspace channel estimation method in comparison with the SS time domain algorithm in Muquet et al. (2002b) and the correlation matching algorithm in Backx et al. (2007). To fully test the performance of the proposed method, in the simulation study we conduct investigations in the following three aspects: (a) assessing the estimation performance of our method under a range of SNR values with the number of received blocks fixed, (b) evaluating the estimation performance of our method with a fixed SNR value and a varying number of received blocks, and (c) testing the robustness of our method when the channel length is unknown.

The TZ-OFDM system used in the simulations comprised $P=32$ subcarriers carrying QPSK modulated symbols. A sequence of $T=8$ trailing zeros was appended to each transmitted block. The FIR channel is of order $L=4$. AWGN was applied to the channel output. The performance measure adopted in this section is the mean-square-error (MSE) in dB, defined as

$$\text{MSE (dB)} = 10\log_{10}\left(\frac{1}{N_r}\sum_{i=1}^{N_r}\|\hat{\mathbf{h}}_i - \mathbf{h}\|^2\right) \tag{10}$$

where N_r is the number of Monte Carlo runs, $\mathbf{h}$ is the true (unit-norm) channel vector, and $\hat{\mathbf{h}}_i$ is the estimated channel vector (with unit norm) from the ith run.

In the first simulation, we compared the performance of the proposed method with the SS time domain algorithm in Muquet et al. (2002b) and the correlation matching algorithm in Backx et al. (2007) in terms of the MSE vs SNR. We used a total of $N_s = 100$ received TZ-OFDM blocks for channel estimation. The channel order was assumed to be known. The comparative results are reported in Fig. 1. It can be seen that although the correlation matching algorithm shows robustness against observation noise at low SNR (10 dB or less), our method and the SS time domain method exhibit far better performance in the range of mid and high SNR. It is also observed from Fig. 1 that the proposed method performs consistently with the SS time domain algorithm. This is because the rationale behind both methods is identical, namely subspace decomposition.

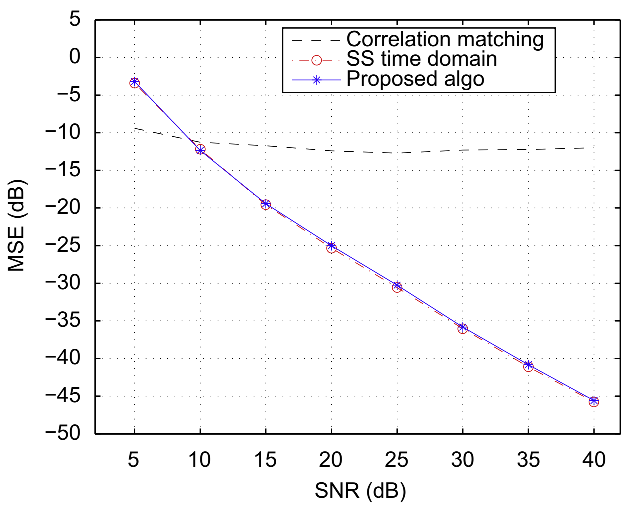

Fig. 1. Performance comparison of the proposed method with the correlation matching and SS time domain algorithms for SNR = 5–40 dB.

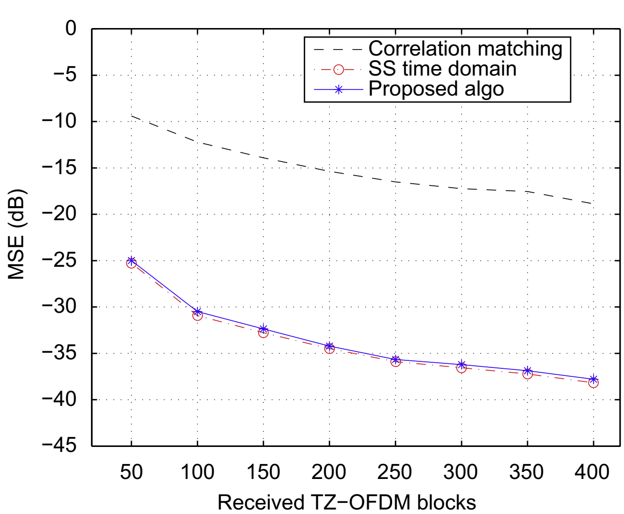

Fig. 2. Performance comparison of the proposed method with the correlation matching and SS time domain algorithms for $N_s = 50$–400 blocks.

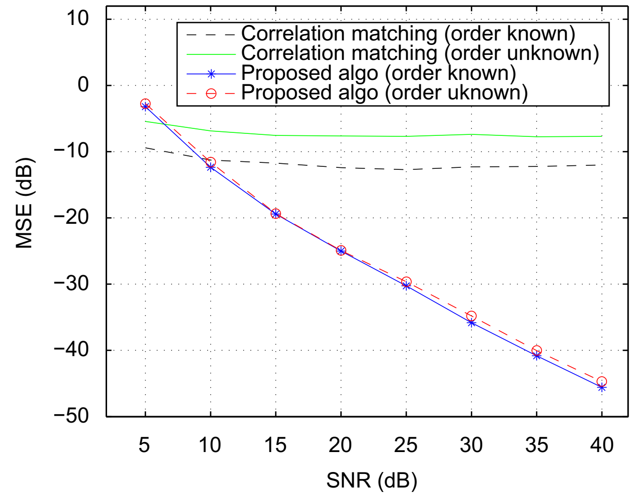

Fig. 3. Channel overestimation with channel order unknown.

In the second simulation, we fixed the SNR at 25 dB and varied the number of received TZ-OFDM blocks from 50 to 400. Again we assumed the knowledge of the channel order. Figure 2 illustrates the MSE of all three methods. As can be seen from Fig. 2, by increasing the number of received data blocks, the MSE for all methods decreases. It is also shown in Fig. 2 that both the proposed and SS time domain methods outperform the correlation matching algorithm. As analyzed earlier, with the channel order known, our method performs as well as the SS time domain algorithm.

In the third simulation, we assessed the performance of the proposed method when the channel order is unknown. In this case, an overestimated channel vector was obtained. A total of $N_s = 100$ received blocks was used in the simulation. As the issue of unknown channel length is not explicitly addressed in the SS time domain algorithm in Muquet et al. (2002b), Fig. 3 only presents the simulated MSE vs SNR for the proposed and correlation matching methods. It is shown in Fig. 3 that our method is strongly robust to channel overestimation.

5. Conclusion

We have developed a frequency domain subspace based blind channel estimator for TZ-OFDM systems. The proposed method reveals some important characteristics of TZ-OFDM systems from a frequency domain perspective. It is noted that the characteristics revealed are not explicit in the time domain. Built upon the favorable features of TZ-OFDM systems, the proposed method presents a straightforward formulation of channel estimation in the frequency domain. Moreover, the new method can deal with channels of unknown length. Simulation results show that our method performs favorably compared to the existing methods and is robust to channel overestimation.

References

Backx FD, Vinhoza TTV, Sampaio-Neto R. Blind channel estimation for zero-padded OFDM systems based on correlation matching. In: Proceedings of the IEEE vehicular technology conference (VTC), 2007. p. 1308–11.

Giannakis GB, Hua Y, Stoica P, Tong L, editors. Signal processing advances in wireless and mobile communications. Trends in channel estimation and equalization, vol. 1, Prentice Hall PTR; 2001.

Golub GH, Van Loan CF. Matrix computations. 3rd ed. Johns Hopkins University Press; 1996.

Hu J, Han F. A pixel-based scrambling scheme for digital medical images protection. Journal of Network and Computer Application 2009;32(4):788–94.

Manton JH. An OFDM interpretation of zero padded block transmission. Syst Control Lett 2002;47:393–9.

Moulines E, Duhamel P, Cardoso JF, Mayrargue S. Subspace methods for the blind identification of multichannel FIR filters. IEEE Trans Signal Process 1995;43(2): 516–25.

Muquet B, de Courville M, Duhamel P. Subspace-based blind and semi-blind channel estimation for OFDM systems. IEEE Trans Signal Process 2002a;5: 1699–712.

Muquet B, Wang Z, Giannakis GB, de Courville M, Duhamel P. Cyclic prefixing or zero padding for wireless multicarrier transmissions?IEEE Trans Commun 2002b;50(12):2136–48.

Prasad R. OFDM for wireless communications systems. Artech House; 2004.

Scaglione A, Giannakis GB, Barbarossa S. Redundant filterbanks precoders and equalizers, part I and Part II. IEEE Trans Signal Process 1999;47(7):1988–2022.

Sklar B. Digital communications: fundamentals and applications. 2nd ed. Prentice Hall PTR; 2001.

Thompson EA, Harmison E, Carper R, Martin R, Isaacs J. Robot teleoperation featuring commercially available wireless network cards. Journal of Network and Computer Applications 2006;29(1):11–24.

Tsai T, Tseng D. Subspace algorithm for blind channel identification and synchronization in single-carrier block transmission systems. Signal Process 2008;88:296–306.

Wang Z, Giannakis GB. Linearly precoded or coded OFDM against wireless channel fades? In: Proceedings of the IEEE workshop on signal processing advances for wireless communications (SPAWC), 2001. p. 267–70.

Wang S, Manton JH. A cross-relation-based frequency-domain method for blind SIMO-OFDM channel estimation. IEEE Signal Process Lett 2009;16(10):865–8.