A Cross-Relation-Based Frequency-Domain Method for Blind SIMO-OFDM Channel Estimation

Song Wang and Jonathan H. Manton

***Abstract*—Single-input multiple-output (SIMO) orthogonal frequency division multiplexing (OFDM) is an appealing multi-carrier transmission technique for combating frequency-selective fading and increasing signal-to-noise ratio. To retrieve transmitted data correctly, reliable estimation of time-dispersive channels is important. This letter develops an improved cross-relation (CR) based blind SIMO-OFDM channel estimation method in the frequency domain. The significance of the proposed algorithm is twofold. First, it is highly data-efficient in that the SIMO-OFDM channels can be blindly identified using a single received data block. Second, the proposed method only requires an upper bound rather than the exact knowledge of the channel length. Simulation results show that the new method performs favorably compared to the existing CR- and subspace-based methods.**

***Index Terms*—Blind channel estimation, orthogonal frequency division multiplexing, single-input multiple-output systems.**

I. Introduction

ORTHOGONAL frequency division multiplexing (OFDM) transmission equipped with multiple receive antennas constitutes a single-input multiple-output (SIMO) OFDM system. SIMO-OFDM transmission is effective in combating frequency-selective fading and increasing signal-to-noise ratio (SNR) due to the spatial diversity accomplished by multiple antennas [1].

Being bandwidth efficient, blind identification of single-carrier systems has been well studied [2]. Among various blind system/channel identification schemes, deterministic methods have the advantages of assuming no statistical structure about the input signal and exhibiting satisfactory performance with a small sample size. As an important deterministic blind channel estimator, the cross-relation (CR) approach was originated in [4] and has played a useful role in subsequent algorithm development, e.g., [5], [7]. More recently, we have presented CR-based blind estimation methods for single-carrier systems in [6] and OFDM systems without cyclic prefix (CP) in [11]. Capitalizing on the inverse discrete Fourier transform (IDFT) and DFT operations performed in OFDM as well as the property brought by the CP, in this letter we deliver a frequency-domain realization of the CR principle for CP OFDM systems. A distinct feature of this work is that it deals with channels of unknown length while the knowledge of channel length is prerequisite for methods in [6] and [11].

A CR-based blind estimation method for single-input two-output CP OFDM channels was proposed in [7], which can identify the channels with just one data block. By contrast, subspace-based blind SIMO-OFDM channel estimators (e.g., [8]) need many more data records. Although the method in [7] is data efficient, channel length must be known a priori. To maintain data-efficiency with equivalent or better performance and remove the restrictive condition on the channel length, we propose an improved CR-based blind SIMO-OFDM channel estimation method. The new method can blindly identify the channels using a single received data block and shows an improved performance over the method in [7]. More importantly, the proposed algorithm only requires an upper bound rather than the exact knowledge of the channel length. Furthermore, our method renders an L $(L \geq 2)$ generalization of the two antenna case in [7].

By formulating the CR in the frequency domain, we first estimate the frequency response of the SIMO-OFDM channels up to a subcarrier frequency-dependent scaling ambiguity. To uniquely identify the channels' impulse response (up to a scale factor across all subcarriers), the frequency-dependent scaling ambiguity must be removed. This is achieved by exploiting the zero padding structure in the IDFT of the channel frequency response. We also show that the proposed method can be readily adapted for OFDM systems with virtual carriers.

II. SIMO-OFDM System Model

Consider an N-subcarrier, single-input L-output OFDM system. Specifically, a data stream of digitally encoded source symbols is divided into blocks of length N. Each block of source data is modulated by the IDFT, prepended with the CP, digital-analog converted and then transmitted. Denote the frequency-domain source block by

$$\mathbf{D}_m = [D_m(0), D_m(1), \ldots, D_m(N-1)] \tag{1}$$

where m is the block index on the OFDM signal and $D_m(k) \neq 0, k = 0, 1, \ldots, N-1$. The timing and carrier synchronization are assumed to be perfect. Due to the CP, the effect of the multipath radio channel is limited to a point-wise multiplication of the frequency-domain source data by the frequency response of the channel [3]. Therefore, at the receiver, after removing the CP and performing the DFT demodulation, the received signal in the mth block at the lth channel is given by

$$X_m^{(l)}(k) = D_m(k)H^{(l)}(k) + W_m^{(l)}(k) \tag{2}$$

where $H^{(l)}(k)$, $X_m^{(l)}(k)$ and $W_m^{(l)}(k)$ are respectively the channel's frequency response, the frequency response of the received signal and the additive white Gaussian noise (AWGN), $l = 1, 2, \ldots, L$, and $k = 0, 1, \ldots, N-1$.

Manuscript received May 05, 2009; revised June 12, 2009. First published June 23, 2009; current version published July 29, 2009. The associate editor coordinating the review of this manuscript and approving it for publication was Prof. Chong-Yung Chi.

S. Wang is with the Department of Electronic Engineering, La Trobe University, Melbourne, VIC 3086, Australia (e-mail: song.wang@latrobe.edu.au).

J. H. Manton is with the Department of Electrical & Electronic Engineering, University of Melbourne, Melbourne, VIC 3010, Australia (e-mail: j.manton@ieee.org).

Color versions of one or more of the figures in this paper are available online at http://ieeexplore.ieee.org.

Digital Object Identifier 10.1109/LSP.2009.2025926

With the SIMO-OFDM model (2), we aim to estimate the channel's finite impulse response (FIR) $h^{(l)}(0), h^{(l)}(1), \ldots, h^{(l)}(M-1)$, $l = 1, 2, \ldots, L$, using just one received OFDM block $[X_m^{(l)}(0), X_m^{(l)}(1), \ldots, X_m^{(l)}(N-1)]$. Suppose that M is the maximum length of the L channels[1], where M is not necessarily known a priori, but upper bounded by a known integer $\bar{M}$, i.e., $M \leq \bar{M}$.

III. Cross-Relation-Based Frequency-Domain Blind SIMO-OFDM Channel Estimation

In the absence of noise, it follows from (2) that any channel pair (i, j) satisfies

$$X_m^{(i)}(k) H^{(j)}(k) = X_m^{(j)}(k) H^{(i)}(k). \quad (3)$$

The above relationship can be viewed as the frequency-domain counterpart of the time-domain CR [4] for multicarrier systems. For each $k = 0, 1, \ldots, N-1$, $1 \leq i, j \leq L$ and $i \neq j$, we define $\boldsymbol{\mathcal{X}}_m^{i,j}(k) \in \mathcal{C}^{1\times L}$ as

$$\boldsymbol{\mathcal{X}}_m^{i,j}(k) = [0 \ \cdots \ 0 - X_m^{(j)}(k) \ 0 \ \cdots \ 0 \ X_m^{(i)}(k) \ 0 \ \cdots \ 0] \quad (4)$$

where the ith entry is $-X_m^{(j)}(k)$ and the jth entry is $X_m^{(i)}(k)$. For each $k = 0, 1, \ldots, N-1$, from (3) and (4) we can form

$$\mathbf{X}_m(k)\mathbf{H}(k) = \mathbf{0} \quad (5)$$

where $\mathbf{H}(k) = [H^{(1)}(k) \ H^{(2)}(k) \ \cdots \ H^{(L)}(k)]^T$ and

$$\mathbf{X}_m(k) = [\boldsymbol{\mathcal{X}}_m^{1,2}(k)^T \ \boldsymbol{\mathcal{X}}_m^{1,3}(k)^T \ \cdots \ \boldsymbol{\mathcal{X}}_m^{1,L}(k)^T \ \boldsymbol{\mathcal{X}}_m^{2,3}(k)^T \cdots \boldsymbol{\mathcal{X}}_m^{2,L}(k)^T \cdots \ \boldsymbol{\mathcal{X}}_m^{L-1,L}(k)^T]^T \in \mathcal{C}^{L(L-1)/2\times L}$$

Theorem 1: If all channels share no common zero, any nontrivial solution $\tilde{\mathbf{H}}(k)$ *to* $\mathbf{X}_m(k)\tilde{\mathbf{H}}(k) = \mathbf{0}$, $k = 0, 1, \ldots, N-1$, is given by $\tilde{\mathbf{H}}(k) = \alpha_k \mathbf{H}(k)$, where α_k is a nonzero scalar dependent on k.

Proof: It follows from $\mathbf{X}_m(k)\tilde{\mathbf{H}}(k) = \mathbf{0}$ that for any channel pair (i, j), $X_m^{(i)}(k)\tilde{H}^{(j)}(k) = X_m^{(j)}(k)\tilde{H}^{(i)}(k)$, or equivalently, $D_m(k)H^{(i)}(k)\tilde{H}^{(j)}(k) = D_m(k)H^{(j)}(k)\tilde{H}^{(i)}(k)$. As $D_m(k) \neq 0$ [see (1)], we have

$$H^{(i)}(k)\tilde{H}^{(j)}(k) = H^{(j)}(k)\tilde{H}^{(i)}(k). \quad (6)$$

For $k = 0, 1, \ldots, N-1$, there exists at least one channel q such that $H^{(q)}(k) \neq 0$. This is because if $H^{(l)}(k) = 0$ for each $l = 1, \ldots, L$, then $H^{(l)}(z)|_{z=e^{j2\pi k/N}} = 0$, which contradicts no common zero among all channels. When $H^{(q)}(k) \neq 0$, we need to show that $\tilde{H}^{(q)}(k) \neq 0$. We proceed with the proof by contradiction. Assume $\tilde{H}^{(q)}(k) = 0$. According to (6),

$$H^{(i)}(k)\tilde{H}^{(q)}(k) = H^{(q)}(k)\tilde{H}^{(i)}(k). \quad (7)$$

If $H^{(q)}(k) \neq 0$ and $\tilde{H}^{(q)}(k) = 0$, then $\tilde{H}^{(i)}(k) = 0$. This means $\tilde{H}^{(l)}(k) = 0$ for each $l = 1, \ldots, L$, which contradicts that $\tilde{\mathbf{H}}(k)$ is a nontrivial solution to $\mathbf{X}(k)\tilde{\mathbf{H}}(k) = \mathbf{0}$. Thus, $\tilde{H}^{(q)}(k) \neq 0$ when $H^{(q)}(k) \neq 0$, or equivalently, $H^{(q)}(k)/\tilde{H}^{(q)}(k) = \alpha_k$, where α_k is a nonzero scalar dependent on k. It follows from (7) that $H^{(i)}(k) = \alpha_k \tilde{H}^{(i)}(k)$. That is, $H^{(l)}(k) = \alpha_k \tilde{H}^{(l)}(k)$, $l = 1, \ldots, L$. □

[1]This implies $h^{(p)}(0),\ h^{(p)}(M-1) \neq 0$ for some channel $p, 1 \leq p \leq L$.

Theorem 1 shows that based on (5), the channel's frequency response $H^{(l)}(k)$, $l = 1, \ldots, L$, can be identified up to a subcarrier frequency-dependent scaling ambiguity, i.e., $H^{(l)}(k) = \alpha_k \tilde{H}^{(l)}(k)$. The frequency-dependent scaling ambiguity α_k must be resolved so that taking the IDFT on $H^{(l)}(k)$ yields the channel's impulse response $h^{(l)}(n)$. Expanding the IDFT of $H^{(l)}(k)$ gives

$$h^{(l)}(n) = \sum_{k=0}^{N-1} H^{(l)}(k) W^{-kn} = \sum_{k=0}^{N-1} \alpha_k \tilde{H}^{(l)}(k) W^{-kn} \quad (8)$$

where $W = e^{-j2\pi/N}$ and the scaling factor $1/N$ is omitted. Incorporating all channels, we rewrite (8) as

$$\begin{bmatrix} \tilde{\mathbf{\Gamma}}_1^{(1)} \\ \tilde{\mathbf{\Gamma}}_2^{(1)} \\ \tilde{\mathbf{\Gamma}}_1^{(2)} \\ \tilde{\mathbf{\Gamma}}_2^{(2)} \\ \cdots \\ \tilde{\mathbf{\Gamma}}_1^{(L)} \\ \tilde{\mathbf{\Gamma}}_2^{(L)} \end{bmatrix} \begin{bmatrix} \alpha_0 \\ \alpha_1 \\ \vdots \\ \alpha_{N-1} \end{bmatrix} = \begin{bmatrix} \mathbf{h}^{(1)} \\ \mathbf{0} \\ \mathbf{h}^{(2)} \\ \mathbf{0} \\ \cdots \\ \mathbf{h}^{(L)} \\ \mathbf{0} \end{bmatrix} \quad (9)$$

where $\mathbf{h}^{(l)} = [h^{(l)}(0) \ h^{(l)}(1) \ \cdots \ h^{(l)}(M-1)]^T$, and $\tilde{\mathbf{\Gamma}}_1^{(l)}$ [$\tilde{\mathbf{\Gamma}}_2^{(l)}$] is the top M rows [bottom $N-M$ rows] of the matrix

$$\mathbf{W} \cdot \mathrm{diag}\{\tilde{H}^{(l)}(0), \tilde{H}^{(l)}(1), \ldots, \tilde{H}^{(l)}(N-1)\} \quad (10)$$

with $\mathbf{W}$ denoting the $N \times N$ IDFT matrix whose elements are $\{\mathbf{W}\}_{p,q} = W^{-pq}$, for $p, q = 0, 1, \ldots, N-1$. It is seen from (9) that the vector $[\alpha_0 \ \alpha_1 \ \cdots \ \alpha_{N-1}]^T$ is in the null space of

$$\tilde{\mathbf{\Gamma}}_M = \left[(\tilde{\mathbf{\Gamma}}_2^{(1)})^T \ (\tilde{\mathbf{\Gamma}}_2^{(2)})^T \ \cdots \ (\tilde{\mathbf{\Gamma}}_2^{(L)})^T\right]^T \in \mathcal{C}^{L(N-M)\times N}. \quad (11)$$

In order to uniquely (up to a constant scalar) determine the channel vector $\mathbf{h} = \left[(\mathbf{h}^{(1)})^T \ (\mathbf{h}^{(2)})^T \ \cdots \ (\mathbf{h}^{(L)})^T\right]^T$, the nullity of $\tilde{\mathbf{\Gamma}}_M$ has to be one.

Theorem 2: The nullity of $\tilde{\mathbf{\Gamma}}_M$ is one.

Proof: First we prove that $\mathrm{nullity}(\mathbf{\Gamma}_M) = 1$, where $\mathbf{\Gamma}_M$ is written in the same format as $\tilde{\mathbf{\Gamma}}_M$ except to replace $\tilde{H}^{(l)}(k)$ with $H^{(l)}(k)$. It is known that the set $\{\mathbf{v}_i = [1 \ W^i \ W^{2i} \ \cdots \ W^{(N-1)i}]^T : i = 0, 1, \ldots, N-1\}$ is a basis for $\mathcal{C}^N$ and that the first vector ($[1 \ 1 \ \cdots \ 1]^T$ for $i = 0$) of this set is in the null space of $\mathbf{\Gamma}_M$. Now we need to show that $\mathbf{v}_i$ for $i = 1, 2, \ldots, N-1$, and their linear combination are not in the null space of $\mathbf{\Gamma}_M$. As

$$\mathbf{\Gamma}_M[\mathbf{v}_1 \ \mathbf{v}_2 \cdots \mathbf{v}_{N-1}] = \mathbf{T} = [(\boldsymbol{\mathcal{T}}^{(1)})^T \ (\boldsymbol{\mathcal{T}}^{(2)})^T \cdots (\boldsymbol{\mathcal{T}}^{(L)})^T]^T$$

where $\boldsymbol{\mathcal{T}}^{(l)} \in \mathcal{C}^{(N-M)\times(N-1)}$ is a Toeplitz matrix whose first row is $[h^{(l)}(M-1) \ h^{(l)}(M-2) \cdots h^{(l)}(0) \ 0 \cdots 0]$ and first column is $[h^{(l)}(M-1) \ 0 \cdots 0]^T$, it follows that each column of $\mathbf{T}$ always contains the nonzero element $h_p(0)$ or $h_p(M-1)$ for some channel p, $1 \leq p \leq L$. So $\mathbf{v}_1, \mathbf{v}_2, \ldots, \mathbf{v}_{N-1}$ are not in the null space of $\mathbf{\Gamma}_M$.

Next we shall prove that any linear combination of $\mathbf{v}_1, \mathbf{v}_2, \ldots, \mathbf{v}_{N-1}$ is not in the null space of $\mathbf{\Gamma}_M$. We proceed by contradiction. Suppose that there exist coefficients

$a_1, a_2, \ldots, a_{N-1}$, not all 0, such that $\mathbf{\Gamma}_M(a_1\mathbf{v}_1 + a_2\mathbf{v}_2 + \cdots + a_{N-1}\mathbf{v}_{N-1}) = \mathbf{0}$, i.e., $\mathbf{T}\,[a_1\ a_2 \cdots a_{N-1}]^T = \mathbf{0}$. Expanding this equation yields $h^{(l)}(n) * a_n = 0$, where $*$ denotes linear convolution and $l = 1, \ldots, L$. This gives $H^{(l)}(z)A(z) = 0$, where $H^{(l)}(z)$ and $A(z)$ are the z-transform of $h^{(l)}(n)$ and a_n, respectively. As all channels share no common zero, we get $A(z) = 0$. Based on the equivalence between the z-transform and DFT, $A(k) = 0$, where $A(k)$ is the DFT of a_n. This leads to $a_n = 0$ for $n = 1, 2, \ldots, N-1$, which is in contradiction with the assumption that $a_1, a_2, \ldots, a_{N-1}$ are not all 0. Thus, $\mathrm{nullity}(\mathbf{\Gamma}_M) = 1$. Since $\tilde{\mathbf{\Gamma}}_M = \mathbf{\Gamma}_M \mathrm{diag}\{1/\alpha_0, 1/\alpha_1, \ldots, 1/\alpha_{N-1}\}$, $\mathrm{rank}(\tilde{\mathbf{\Gamma}}_M) = \mathrm{rank}(\mathbf{\Gamma}_M)$. Therefore, $\mathrm{nullity}(\tilde{\mathbf{\Gamma}}_M) = 1$. □

Theorem 2 establishes $\mathrm{nullity}(\tilde{\mathbf{\Gamma}}_M) = 1$, and the construction of $\tilde{\mathbf{\Gamma}}_M$ is directly associated with the channel length M. Hence, it is necessary to inspect the following two cases:

1) Case A: $\hat{M} > M$: In this case, $\mathrm{nullity}(\tilde{\mathbf{\Gamma}}_{\hat{M}}) > 1$ because there exists more than one $\mathbf{v}_i$, $0 \le i \le N-1$, such that $\tilde{\mathbf{\Gamma}}_{\hat{M}}\mathbf{v}_i = 0$.

2) Case B: $\hat{M} < M$: In this case, $\tilde{\mathbf{\Gamma}}_{\hat{M}}$ has full (column) rank, i.e., $\mathrm{nullity}(\tilde{\mathbf{\Gamma}}_{\hat{M}}) = 0$.

Based on the above analysis, $\mathrm{nullity}(\tilde{\mathbf{\Gamma}}_M) = 1$ holds only for the true M, which offers an approach to determining the channel length if it is not known a priori.

In the presence of noise, the SIMO-OFDM channels are estimated in the least squares sense as follows:

1) Find $\tilde{\mathbf{H}}(k) = \arg\min_{\|\mathbf{H}(k)\|=1} \|\mathbf{X}_m(k)\mathbf{H}(k)\|$, for $k = 0, 1, \ldots, N-1$.
2) Go to the next step if the channel length is known; otherwise, for $M = 1, \ldots, \bar{M}$, compute the ratio of the second smallest singular value to the smallest singular value of $\tilde{\mathbf{\Gamma}}_M$. The value of M corresponding to the largest ratio is the channel length.
3) Determine the vector $\boldsymbol{\alpha} = [\alpha_0\ \alpha_1\ \cdots\ \alpha_{N-1}]^T$ that minimizes $\|\tilde{\mathbf{\Gamma}}_M\boldsymbol{\alpha}\|$ subject to $\|\boldsymbol{\alpha}\| = 1$.
4) With $\tilde{\mathbf{H}}(k) = [\tilde{H}^{(1)}(k)\ \tilde{H}^{(2)}(k)\ \cdots\ \tilde{H}^{(L)}(k)]^T$ obtained in Step 1), compute $\alpha_k\tilde{H}^{(l)}(k)$ for $k = 0, 1, \ldots, N-1$ and $l = 1, \ldots, L$. Finally, take the N-point IDFT on $\alpha_k\tilde{H}^{(l)}(k)$, and the first M elements of the IDFT are the FIR estimate for channel l.

To end this section, we make the following remarks.

i) Both the proposed algorithm and the method in [7] can estimate the SIMO-OFDM channels using a single received block. Besides the better performance of the proposed algorithm shown by the simulations in Section IV, our method does not need to know the channel length a priori whereas the method in [7] only works for the case of known channel length. It is noted that with the channel length known, the method in [7] is computationally more efficient than our method as only one singular value decomposition (SVD) is needed in the method in [7] and the size of the SVD is smaller than that in Step 3) of our algorithm.

ii) For relatively slow time-varying channels where channel parameters can be considered constant over multiple blocks, it is possible to enhance estimation performance by including more received blocks. The inclusion of more blocks only affects the estimation of $\tilde{\mathbf{H}}(k)$, i.e., $\tilde{\mathbf{H}}(k) = \arg\min_{\|\mathbf{H}(k)\|=1} \left\| \left[\mathbf{X}_1^T(k)\ \mathbf{X}_2^T(k) \cdots \mathbf{X}_K^T(k)\right]^T \mathbf{H}(k) \right\|$, by using K received blocks. The rest of the algorithm remains the same. The associated computational cost in estimating $\tilde{\mathbf{H}}(k)$ using K blocks is $\mathcal{O}(KNL^3(L-1)/2)$[9].

iii) Although we have derived the algorithm for OFDM systems without virtual carriers, the proposed method can be adapted for the situation where virtual carriers are present. Suppose that there are J nonvirtual carriers among N subcarriers, i.e., $D_m(k) = 0$ in (1) for $k = J, J+1, \ldots, N-1$. It follows that $J \ge 2M-1$ must hold for the SIMO-OFDM channels to be identifiable [7]. The presence of $(N-J)$ virtual carriers changes the range of subcarrier index k from $k = 0, 1, \ldots, N-1$ to $k = 0, 1, \ldots, J-1$, and accordingly the vector $\boldsymbol{\alpha} = [\alpha_0\ \alpha_1\ \cdots\ \alpha_{J-1}]^T$ and $\mathbf{W}$ in (10) is the $J \times J$ IDFT matrix.

IV. Simulations

Simulations were run to test the proposed method for channel length estimation and parameter identification. In the simulations we chose the same single-input two-output OFDM channel example as in Section V, [7]. The source symbols were QPSK modulated and AWGN was applied to each channel output. Unless stated otherwise, the OFDM system consisted of 16 subcarriers with no virtual carriers. The SNR(dB) of the SIMO-OFDM system is defined by

$$\mathrm{SNR(dB)} = 10\log_{10}\left(\frac{E\left\{\sum_{l=1}^{L}\sum_{k=0}^{N-1}|X_m^{(l)}(k)|^2\right\}}{E\left\{\sum_{l=1}^{L}\sum_{k=0}^{N-1}|W_m^{(l)}(k)|^2\right\}}\right).$$

The performance measure is the mean-square-error (MSE) in dB, as defined in Section IV, [11]. The inherent scalar ambiguity in the channel estimate was removed using [10].

In the first simulation, we evaluated the performance of the proposed method against that of the methods in [7] and [8], referred to as WLC and AMZ algorithms, respectively. As both WLC and AMZ algorithms cannot handle unknown channel length, for the sake of performance comparison, we assume that the channel length is known. The purpose of this simulation study is to demonstrate the high data-efficiency of the proposed method. Fig. 1 presents the MSE of all three methods. It is observed from Fig. 1 that both the proposed and WLC algorithms can blindly estimate the channels using just one received OFDM block with our method outperforming the WLC algorithm. To achieve a comparable performance, the subspace-based AMZ algorithm has to use 40 blocks of data.

In the second simulation, we incorporated virtual carriers in the OFDM system, where 12 out of 64 carriers were virtual carriers. Channel length was assumed to be known. Fig. 2 compares the performance of our method with the WLC algorithm (one received block used in both methods). As can be seen from Fig. 2, our method performs slightly better than the WLC algorithm in the presence of virtual carriers.

In the third simulation, we chose a different number of data blocks for the proposed algorithm (channel length assumed to be known). It is shown in Fig. 3 that more data blocks lead to better estimation accuracy. As was analyzed in Section III, the performance improvement comes with an increased computational cost in Step 1) of the proposed algorithm.

In the last simulation, we used the proposed algorithm to estimate the channel length, assuming only the knowledge of an

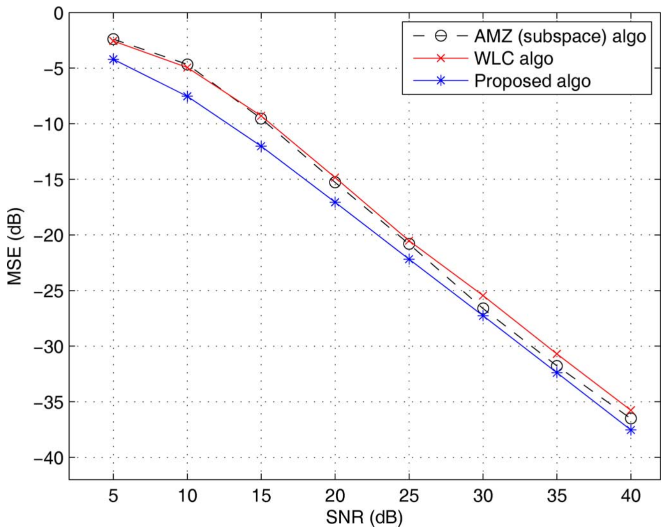

Fig. 1. Performance comparison of the proposed algorithm with WLC and (subspace-based) AMZ algorithms. One received data block used for the proposed and WLC algorithms. Forty data blocks used for the AMZ algorithm.

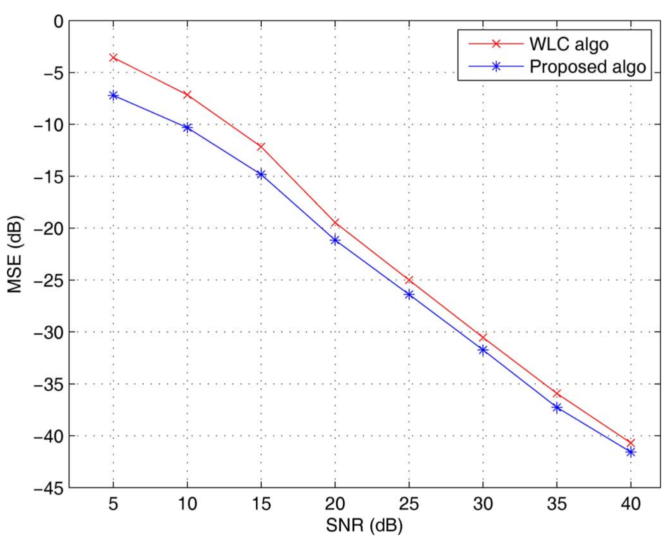

Fig. 2. Comparison of the proposed algorithm with the WLC algorithm in the presence of virtual carriers. One block of data used in both methods.

upper bound ($\bar{M} = 10$). The channel length was estimated over a range of SNR using one received OFDM block. It can be seen from Table I that the estimated channel length coincides with the true length ($M = 5$) at relatively high SNR. At these higher SNR values, the MSE matches the corresponding curve in each of Figs. 1 and 3. At low SNR (12 dB or less), channel length estimation is inaccurate. In the presence of length mismatch, subsequent channel parameter identification turns out to be ineffective, as analyzed in Cases A and B in Section III.

V. Conclusion

We have presented a CR-based blind SIMO-OFDM channel estimation method in the frequency domain. The new method can blindly identify the SIMO-OFDM channels using a single received data block, thus it is well suited to situations where fast channel variations occur. Moreover, the proposed method requires only the upper bound of the channel length, which is a clear advantage over many existing blind channel estimation methods. Numerical simulations show that our method is much more data-efficient than subspace-based blind SIMO-OFDM channel estimators (e.g., [8]) and has an improved performance over the CR-based method [7]. In the case of unknown channel length, the proposed method is most effective at high SNR.

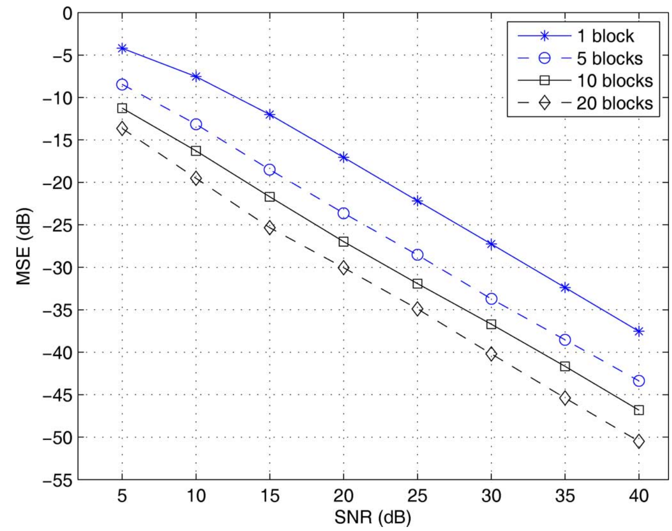

Fig. 3. Performance of the proposed method using one, five, ten, or 20 data blocks.

TABLE I
Channel Length Estimation With $\bar{M} = 10$.

SNR (dB)	2	5	10	15	20	25	30	35	40	45
Channel length estimate	7	7	6	5	5	5	5	5	5	5

References

[1] B. Sklar, *Digital Communications: Fundamentals and Applications*, 2nd ed. Upper Saddle River, NJ: Prentice-Hall PTR, 2001.

[2] *Signal Processing Advances in Wireless and Mobile Communications*, G. B. Giannakis, Y. Hua, P. Stoica, and L. Tong, Eds. Upper Saddle River, NJ: Prentice-Hall PTR, 2001, vol. 1, Trends in Channel Estimation and Equalization.

[3] R. Prasad, *OFDM for Wireless Communications Systems*. Norwell, MA: Artech House, 2004.

[4] G. Xu, H. Liu, L. Tong, and T. Kailath, “A least-squares approach to blind channel identification,” *IEEE Trans. Signal Process.*, vol. 43, pp. 2982–2993, 1995.

[5] Y. Huang and J. B. Benesty, “A class of frequency-domain adaptive approaches to blind multichannel identification,” *IEEE Trans. Signal Process.*, vol. 50, pp. 11–24, 2003.

[6] S. Wang, J. H. Manton, D. B. H. Tay, C. Zhang, and J. C. Devlin, “An FFT based method for blind identification of FIR SIMO channels,” *IEEE Signal Process. Lett.*, vol. 14, pp. 437–440, 2007.

[7] H. Wang, Y. Lin, and B. Chen, “Data efficient blind OFDM channel estimation using receiver diversity,” *IEEE Trans. Signal Process.*, vol. 51, pp. 2613–2623, 2003.

[8] F. O. Alayyan, K. Abed-Meraim, and A. M. Zoubir, “Blind MMSE channel identification and equalization algorithms for OFDM systems,” in *Proc. Int. Symp. Signal Processing and Its Applications*, Feb. 2007, pp. 1–4.

[9] G. H. Golub and C. F. Van Loan, *Matrix Computations*, 3rd ed. Baltimore, MD: Johns Hopkins Univ. Press, 1996.

[10] D. R. Morgan, J. Benesty, and M. M. Sondhi, “On the evaluation of estimated impulse responses,” *IEEE Signal Process. Lett.*, vol. 5, pp. 174–176, 1998.

[11] S. Wang and J. H. Manton, “Blind channel estimation for non-CP OFDM systems using multiple receive antennas,” *IEEE Signal Process. Lett.*, vol. 16, pp. 299–302, 2009.