Subspace Decomposition Approach to Multi-User MIMO Channel Estimation in SC-FDE Systems

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***Abstract*—Multiuser multiple-input multiple-output (MU-MIMO) wireless systems that employ single-carrier frequency-domain equalization (SC-FDE) for uplink transmissions can provide high data rates with increased spectral efficiency under severe channel conditions. However, the availability of accurate channel estimates at the base station (BS) receiver is crucial for achieving peak performance. In this paper, we investigate the use of subspace decomposition and derive a novel algorithm for the blind estimation of MU-MIMO channels in SC-FDE systems, as specified by the 3GPP LTE. By exploiting the long data blocks available in LTE standards, our proposed blind algorithm can obtain accurate estimates of the MU-MIMO channels over every block of transmitted data. This provides for a bandwidth-efficient solution in SC-FDE systems by eliminating (reducing) the need to allocate an entire block of pilot sequences for each active transmitter. Furthermore, since the channel estimation is deployed at the BS, the computational complexity is not considered to be a significant burden for future systems in exchange for the increased spectral efficiency. The results of simulations over fading channels, using realistic system parameters representative of LTE-Advanced, demonstrate the advantages of our proposed blind subspace-based channel estimation algorithm and support the feasibility of the resulting MU-MIMO SC-FDE scheme with reduced training.**

***Keywords*—blind channel estimation; subspace decomposition; MIMO, LTE-A; SC-FDE; SC-FDMA**

I. INTRODUCTION

Single-carrier frequency division multiple-access (SC-FDMA) has attracted considerable attention recently, following its adoption for uplink transmission in 3GPP Long Term Evolution (LTE) mobile wireless communication systems [1]. This choice has been motivated in part by its low peak-to-average power ratio (PAPR), which benefits the mobile user equipment (UE) when compared to orthogonal FDMA (OFDMA) techniques [2]. In SC-FDMA, the base station (BS) allocates resources to all users within a cell and decides which UE can share the same frequency-time resources, in a way that will minimize interference among them [2]. In the special case when mobile users occupies all available subcarriers for transmission, rather than sharing with other users, the single-carrier frequency-domain equalization (SC-FDE) terminology becomes more appropriate since in effect, the multiple-access is realized in the spatial, as opposed to the frequency domain.

In its Release 10, LTE Advanced (LTE-A) aims to meet the IMT-Advanced data rate and spectral efficiency requirements by incorporating multi-antenna transceiver systems [3]. Multiple single antenna UEs can transmit simultaneously on the same frequency band in order to increase spectral efficiency while multi-antenna receivers are used at the base station BS, thereby forming a virtual MIMO sub-systems. In the receiver, a multi-user equalizer, such as a MMSE detector [4]-[6], can separate the signals of the different transmitting UE, for which the operation is transparent since all the computations are performed at the BS [2]. The ability of the BS to separate signals that are simultaneously received on the same frequencies is dependent upon accurate channel state information (CSI). LTE-A specifically provides orthogonal demodulation reference (pilot) signals using different cyclic time shifts on a complete frame [7] to enable the BS to obtain channel estimates for each one of the supported physical layer UEs in uplink MU-MIMO [8].

As an alternative to the use of pilot signals, which consume precious time-bandwidth resources, one may resort to blind (or semi-blind) estimation approaches that take advantage of known structural properties of the received signals. In recent years, blind (and semi-blind) channel estimation techniques have been proposed that exploit different types of precoding structures in single-carrier MIMO block transmissions with cyclic prefix, such as [9]. However, while these methods lead to improved performance and robustness in the channel estimation, such an advantage comes at the price of a capacity loss due to precoding [10]. As an alternative to precoding, subspace decomposition techniques have been employed successfully for the blind estimation of MIMO channels in Cyclic Prefix (CP) based OFDM systems [11],[12]. Subspace methods have a clear structure (i.e. minimization of a quadratic cost function) and can achieve good performance provided certain identifiability conditions are satisfied. They are known to identify the unknown MIMO channel up to a matrix ambiguity of reduced size, which can be determined using a small number of pilots (hence semi-blind) [13], [14]. The channel estimate is considered to be as accurate as the received signal correlation matrix can be evaluated, if the ambiguity is presumed to have been removed correctly [10], [13]. Recently, this type of approach was successfully demonstrated for the single-user case in SC-FDE systems [15].

In this paper, we propose to extend the blind subspace decomposition approach and apply it to the estimation of MU-MIMO channels in SC-FDE systems, where each simultaneous active UE transmits on the same time and frequency resources utilizing all of the available subcarriers. To this end, we recast the MU-MIMO channel estimation problem as the minimization of a quadratic cost function involving the eigenvalue decomposition of the array signal correlation matrix. In LTE, the number of transmitted data symbols in one block (FFT size) can be as large as 2048. This allows for high accuracy in estimating the correlation matrix over one block of received data, which in-turn is reflected into more accurate channel estimates. When used in conjunction with the MMSE multiuser equalizer, This accurate CSI leads to the desired high performance in the MU-MIMO SC-FDE transmission while simultaneously increasing the spectral efficiency due to reduction in the required number of pilot symbols. Monte-Carlo simulations over fading channels, using a realistic LTE-Advanced system model, demonstrate the advantages of our proposed blind subspace-based channel estimation algorithm and support the feasibility of the resulting MU-MIMO SC-FDE scheme.

The rest of the paper is organized as follows: Section II describes the structure of the considered MU-MIMO SC-FDE system. Subspace decomposition for MIMO channels estimation is described in Section III. In Section IV, we present our simulation study in support of the proposed approach via Monte-Carlo style runs to evaluate channel estimation accuracy and the corresponding BER. Finally, we conclude in Section V.

II. SYSTEM MODEL AND RECEIVER STRUCTURE

We consider an uplink multiuser MIMO SC-FDE block transmission system with cyclic prefix (CP). Each of the N_T active UE transmitters has a single antenna element to transmit to the N_R antenna (demodulators) at the base station (eNodeB) using the same time and frequency resources in a synchronous fashion.

A. Transmitter and Receiver Structure

Denote the *M*-ary quadrature amplitude modulated (M-QAM) data transmitted by the i^{th} mobile user at the l^{th} block (frame) as :

$$\mathbf{x}_i^l = \left[\, x_i^l(0),\ x_i^l(1), \dots, x_i^l(N-1)\right]^T \tag{1}$$

where $x_i^l(n)$ is the n^{th} complex-valued symbol in user *i*'s data frame, and *N* is the block size. Before transmission through the channel, a cyclic prefix (CP) of length, N_{CP}, is added. Each of the N_T users' transmitters performs these operations on the l^{th} data block (frame) in an identical fashion and the UEs are scheduled to transmit on the same time and frequency resources.

At the MU-MIMO SC-FDMA receiver, each of the N_R antenna elements receives the sum of the N_T active users' signals after passing through the multipath channels between each of the UE transmitters and each of the receiver antenna elements (sensors). The $N_R \times 1$ received signal vector follows the standard equivalent baseband finite impulse response MIMO (FIR-MIMO) channel model [13], [14] with N_T transmitters and N_R receivers. Therefore, we can express the received signal on the N_R antenna elements at each sampled-time instance as:

$$\boldsymbol{y}^l(n) = \sum_{m=0}^{M} \mathbf{h}^l(m)\boldsymbol{x}^l(n-m) + \boldsymbol{z}^l(n) \tag{2}$$

with $\boldsymbol{y}^l(n) = \left[\, y_1^l(n)\ , y_2^l(n), \dots, y_{N_R}^l(n)\right]^T$, $\boldsymbol{x}^l(n) = \left[\, x_1^l(n)\ , x_2^l(n), \dots, x_{N_T}^l(n)\right]^T$ is N_T-dimensional signal vector, *M* is the channel order, $\{\mathbf{h}^l(m)\}_{m=0,\dots,M}$ are the unknown $N_R \times N_T$ matrix-value impulse response channel coefficients at time delay, *m*, given by:

$$\mathbf{h}^l(m) = \begin{bmatrix} h_{1,1}^l(m) & \cdots & h_{N_T,1}^l(m) \\ h_{1,2}^l(m) & \cdots & h_{N_T,2}^l(m) \\ \vdots & \ddots & \vdots \\ h_{1,N_R}^l(m) & \cdots & h_{N_T,N_R}^l(m) \end{bmatrix} \tag{3}$$

The noise, $\mathbf{z}^l(n)$, is modeled as zero mean white Gaussian with variance σ_z^2 ; hence $E[\mathbf{z}_j^l \mathbf{z}_j^{l^H}] = \sigma_z^2 \mathbf{I}$, where $\mathbf{I}$ is the $N_R \times N_R$ identity matrix.

Also, after removing the cyclic prefix at the receiver, the received signal on antenna element *j* can be expressed as :

$$\boldsymbol{y}_j^l = \sum_{i=0}^{N_T-1} \mathbf{h}_{i,j}^l \otimes \boldsymbol{x}_i^l + \mathbf{z}_j^l \tag{4}$$

where $\mathbf{y}_j^l = \left[\, y_j^l(0), y_j^l(1), \dots, y_j^l(N-1)\right]^T$, $\mathbf{h}_{i,j}^l = \left[\, h_{i,j}^l(0)\ , h_{i,j}^l(1), \dots, h_{i,j}^l(M)\ \right]^T$ is a $(M+1)\times 1$ impulse response vector of the wireless channel between the i^{th} user's transmit antenna and the receiver's j^{th} antenna during transmission of the l^{th} data block, the operator $\otimes$ denotes circular convolution, and *M* is the channel order that is assumed not to exceed the CP length, N_{CP}. Also, $\mathbf{z}_j^l = \left[\, z_j^l(0)\ , z_j^l(1), \dots, z_j^l(N-1)\right]^T$ is the noise vector on the j^{th} antenna during transmission of the l^{th} data block.

B. Multiuser MIMO Equalizer Structure

To separate the N_T mobile users' data signals at the (BS) receiver side, we take the *N*-point DFT of (2) and stack the frequency-domain symbols at each subcarrier *k* in a $N_R \times 1$ vector as [6]:

$$\mathbf{Y}^l(k) = \sum_{i=0}^{N_T-1} \mathbf{H}_i^l(k)\, X_i(k)\ + \mathbf{Z}^l(k), \qquad k = 0,1,\dots,N-1 \tag{5}$$

where $\mathbf{H}_i^l(k) = \left[\, H_{i,1}^l(k), \dots, H_{i,N_R}^l(k)\right]$ contains the i^{th} user estimated channel frequency response at subcarrier, *k*, on each of the receiving antenna elements, and $\mathbf{Z}^l(k)$ is the frequency-domain noise expressed as: $\mathbf{Z}_i^l(k) = \left[\, Z_1^l(k), \dots, Z_{N_R}^l(k)\right]$. By expressing the $N_R \times N_T$ channel transfer function matrix at subcarrier *k* as:

$$\mathbf{H}^l(k) = \begin{bmatrix} H_{1,1}^l(k) & \cdots & H_{N_T,1}^l(k) \\ H_{1,2}^l(k) & \cdots & H_{N_T,2}^l(k) \\ \vdots & \ddots & \vdots \\ H_{1,N_R}^l(k) & \cdots & H_{N_T,N_R}^l(k) \end{bmatrix} \tag{6}$$

and defining vector $\mathbf{X}^l(k) = \left[X_1^l(k), \dots, X_{N_T}^l(k)\right]^T$, then we can rewrite (5) as:

$$\mathbf{Y}^l(k) = \mathbf{H}^l(k)\,\mathbf{X}^l(k) + \mathbf{Z}^l(k), \qquad k = 0,1,\dots,N-1 \qquad (7)$$

The matrix-valued filter, $\mathbf{W}^l(k)$, that minimizes the error, $\mathbf{e}^l(k) = \mathbf{W}^l(k)^H\,\mathbf{Y}^l(k) - \mathbf{X}^l(k)$ in the mean square sense (MSE) (i.e., MMSE) is given by [4]- [6]:

$$\mathbf{W}^l(k) = \left[\,\mathbf{H}^l(k)\mathbf{H}^{l^H}(k) + N_T\sigma_Z^2\,\mathbf{I}\right]^{-1}\ \mathbf{H}^l(k) \qquad (8)$$

Then we let $\widehat{\mathbf{X}}^l(k) = \mathbf{W}^l(k)^H\mathbf{Y}^l(k)$ denote the MMSE filtered signal in the frequency-domain. Now, taking the N-point IDFT of each of the individual users filter signals, we get the sampled-time domain pre-detection symbols.

In order for this MU-MIMO system to function properly as explained, it is imperative that we obtain accurate channels estimates to be able to filter and separate the N_T uplink data streams relying on (8). It is clear from (8) that the accuracy of the channels estimates and the actual channels properties will determine ultimately the performance of this system.

III. SUBSPACE METHOD FRAMEWORK

Subspace (SS) based blind channel estimation algorithms, which were first introduced in [16] for the SIMO case, and later expanded to the MIMO scenario [14], [17], derive their properties from the second-order statistics (SOS) of the received signals. We formulate the application of subspace decomposition for the (semi) blind estimation of the sampled time-domain MIMO channels of the N_T simultaneously active UE [14] assuming that the channels variations are sufficiently slow to be considered static over the duration of one data block. Then we take their individual DFT to obtain the corresponding frequency-domain estimates for the application of (8).

If we take $L < N$ successive observations of (2) and stack them into an $N_R L\times 1$ vector as:

$$\underline{\mathbf{y}}^l(n) = [\,\boldsymbol{y}^l(n)^T, \boldsymbol{y}^l(n-1)^T, \dots, \boldsymbol{y}^l(n-L+1)^T]^T \quad (9),$$

we can now express (2) as:

$$\underline{\mathbf{y}}^l(n) = \ \mathcal{H}_L(\mathbf{h}^l)\,\underline{\mathbf{x}}^l(n) + \underline{\mathbf{z}}^l(n) \qquad (10)$$

where $\underline{\mathbf{x}}^l(n) = [\,\mathbf{x}^l(n)^T, \mathbf{x}^l(n-1)^T, \dots, \mathbf{x}^l(n-L-M+1)^T]^T$, $\underline{\mathbf{z}}^l(n) = [\,\mathbf{z}^l(n)^T, \mathbf{z}^l(n-1)^T, \dots, \mathbf{z}^l(n-L+1)^T]^T$ and the channel matrix $\mathcal{H}_L(\mathbf{h}^l)$ is the $N_R L \times N_T\,(L+M)$ generalized Sylvester matrix of order L associated to the matrix $\mathbf{h}^l = [\,\mathbf{h}^l(0)^T, \mathbf{h}^l(1)^T, \dots, \mathbf{h}^l(M)^T]^T$ expressed as:

$$\mathcal{H}_L(\mathbf{h}^l) = \begin{bmatrix} \mathbf{h}^l(0) & \mathbf{h}^l(1) & \cdots & \mathbf{h}^l(M) & 0 & \cdots & 0 \\ 0 & \mathbf{h}^l(0) & \mathbf{h}^l(1) & \ddots & \mathbf{h}^l(M) & \ddots & \vdots \\ \vdots & \ddots & \ddots & \ddots & \ddots & \ddots & 0 \\ 0 & \cdots & 0 & \mathbf{h}^l(0) & \mathbf{h}^l(1) & \cdots & \mathbf{h}^l(M) \end{bmatrix} \quad (11)$$

We assume that $\underline{\mathbf{y}}^l(n)$ in (10) is a wide-sense stationary process and the signals $\underline{\mathbf{x}}^l(n)$ and noise $\underline{\mathbf{z}}^l(n)$ are mutually independent and zero mean. Let $\boldsymbol{R}_y$ be the $N_R L \times N_R L$ correlation matrix of $\underline{\mathbf{y}}^l(n)$ expressed as:

$$\boldsymbol{R}_y^l = E[\,\underline{\mathbf{y}}^l(n)\underline{\mathbf{y}}^l(n)^H\,] = \ \mathcal{H}_L(\mathbf{h}^l)\boldsymbol{R}_x^l\mathcal{H}_L(\mathbf{h}^l)^H + \sigma_z^2\mathbf{I}_{N_R L} \qquad (12)$$

where the matrix $\boldsymbol{R}_x^l = E[\,\underline{\mathbf{x}}^l(n)\,\underline{\mathbf{x}}^l(n)^H]$ is positive-definite as assumed for $L > N_T M$ [17], and $\mathcal{H}_L(\mathbf{h}^l)$ has full column rank $N_T\,(L+M)$. The maximum channel order, M, is assumed to be known [18] . Therefore, $\boldsymbol{R}_y$ can be written as :

$$\boldsymbol{R}_y^l = \mathbf{U}_x\boldsymbol{\Lambda}_x\mathbf{U}_x^H + \sigma_z^2\mathbf{U}_z\mathbf{U}_z^H \qquad (13)$$

where $\mathbf{U}_x = [\mathbf{u}_1, \dots, \mathbf{u}_{N_T(L+M)}]$ denotes the signal eigenvectors, and $\mathbf{U}_z = [\mathbf{u}_{N_T(L+M)+1}, \dots, \mathbf{u}_{N_R L}]$ denotes the noise eigenvectors. $\boldsymbol{\Lambda}_x = diag(\lambda_1, \dots, \lambda_{N_T(L+M)})$, with $\lambda_1 \geq \lambda_2 \geq \cdots \geq \lambda_{N_T(L+M)} \geq \sigma_z^2$ are the signal eigenvalues. It is shown in [19], [16] that the signal subspace spanned by the eigenvectors corresponding to the largest $N_T\,(L+M)$ eigenvalues spans the same space as the columns of the channel matrix (i.e., range($\mathbf{U}_x$) = range ($\mathcal{H}_L(\mathbf{h}^l)$), while the rest of the eigenvectors, $r = N_R L - N_T\,(L+M)$ span the noise subspace (i.e., range ($\mathbf{U}_z$) = range ($\mathcal{H}_L(\mathbf{h}^l)^\perp$). Using orthogonality relation between noise and signal subspace is the cornerstone of the subspace method that leads to the standard result of:

$$\mathbf{U}_z^H\mathcal{H}_L(\mathbf{h}^l) = \mathcal{H}_L(\mathbf{h}^l)^H\mathbf{U}_z = 0 \qquad (14)$$

This orthogonality relationship permits identification of the channel matrix up to a right multiplication of an invertible $N_T \times N_T$ ambiguity matrix.

To demonstrate how the MIMO channels identification process is carried-out [14], we first denote the noise subspace eigenvectors by ($\mathbf{g}_m$, m = 1, ..., r), and partition each eigenvector as:

$$\mathbf{g}_m = \left[\,\mathbf{g}_0^{(m)^T},\ \mathbf{g}_1^{(m)^T},\ \dots, \mathbf{g}_L^{(m)^T}\right]^T \qquad (15)$$

where $\mathbf{g}_k^{(m)}$, k = 0, 1, ..., L are of size $N_R \times 1$. Then, define the $N_R\,(M+1) \times (L+M)$ matrix as:

$$\mathcal{G}_m = \begin{bmatrix} \mathbf{g}_0^{(m)} & \mathbf{g}_1^{(m)} & \cdots & \mathbf{g}_L^{(m)} & 0 & \cdots & 0 \\ 0 & \mathbf{g}_0^{(m)} & \mathbf{g}_1^{(m)} & \ddots & \mathbf{g}_L^{(m)} & \ddots & \vdots \\ \vdots & \ddots & \ddots & \ddots & \ddots & \ddots & 0 \\ 0 & \cdots & 0 & \mathbf{g}_0^{(m)} & \mathbf{g}_1^{(m)} & \cdots & \mathbf{g}_L^{(m)} \end{bmatrix} \qquad (16)$$

It is can be shown [13], [16], that for each of the columns of size $N_R\,(M+1) \times 1$ of the matrix $\mathbf{h}^l$ (i.e., $\mathbf{h}_i^l$, i = 1, ..., N_T) we have:

$$\mathbf{h}_i^H\left(\sum_{m=1}^{r}\mathcal{G}_m\mathcal{G}_m^H\right)\mathbf{h}_i = 0 \qquad (17)$$

and the dimension of the null-space of the matrix:

$$\mathcal{Q} = \sum_{m=1}^{r}\mathcal{G}_m\mathcal{G}_m^H \qquad (18)$$

is N_T . Therefore, this leads to:

$$\mathbf{h}^l = \mathbf{B}\mathbf{R}^{-1} \qquad (19)$$

where $\mathbf{B}$ is the $N_R\,(M+1) \times N_T$ matrix of the eigenvectors of $\mathcal{Q}$ corresponding to the eigenvalues that are equal to zero. $\mathbf{R}^{-1}$ is an invertible $N_T \times N_T$ ambiguity matrix. Determination of the ambiguity matrix requires the use of a small number of pilot symbols interleaved in each UE data block as has been discussed in the literature [14]. We only need ($N_T\,/M\,N_R$) of the required number of transmitted pilot symbols when compared with traditional training-based channel estimation techniques.

We estimate the correlation matrix, $\boldsymbol{R}_y^l$, and its corresponding eigenvectors in a similar fashion as in [15] by first creating an $N_R L \times (N - L + 1)$ matrix, $\mathbf{A}^l$ as :

$$\mathbf{A}^l = \left[\ \underline{\mathbf{y}}^l(L-1)\ \ \underline{\mathbf{y}}^l(L)\ \ldots\ \underline{\mathbf{y}}^l(N-1)\ \right] \qquad (20)$$

where $\underline{\mathbf{y}}^l(n)$ is given by (9). Then, we can obtain the sample correlation matrix of data block, l, as:

$$\boldsymbol{\Phi}_y^l = \frac{\mathbf{A}^l \mathbf{A}^{l^H}}{N - L + 1} \qquad (21)$$

The expected value of (21) is $\mathrm{E}[\boldsymbol{\Phi}_y^l] = \boldsymbol{R}_y^l$, hence the eigenvalues and eigenvectors of $\boldsymbol{\Phi}_y^l$ should approximate those of $\boldsymbol{R}_y^l$, as long as the block of data symbols is long enough so that the time average approximates the stochastic average.

IV. Simulation Results

We evaluate the performance of our proposed MU-MIMO SC-FDE system employing channel estimates obtained by subspace (SS) decomposition methods through the use of Monte-Carlo simulations. Therefore, we constructed a Matlab simulation with parameters that resemble an LTE-A like system to evaluate the effectiveness and performance of the formulated subspace decomposition based MIMO channel estimation algorithm in such practical application. The main simulation parameters are listed in Table I. The performance measures we consider here are normalized mean square error (NMSE), and bit error rate (BER). We define the energy per bit to noise single sided spectral density ratio as E_b/N_o, and we define the NMSE as:

$$\mathrm{NMSE} = \frac{\mathrm{E}\left[\,\left|\hat{H}_{i,j}^l(k) - H_{i,j}^l(k)\right|^2\right]}{\mathrm{E}\left[\,\left|\,H_{i,j}^l(k)\right|^2\right]}$$

Each simulation point in the results was produced as the average of 100 Monte-Carlo realizations. Each path of the fading MIMO channel from each of the UE antenna to each of the N_R receive antenna elements was modeled as frequency-selective Rayleigh faded of 10 taps with uniform power delay profile. The Doppler spread follows Jake's model with a maximum fading rate, f_d. We also assume very low spatial correlation between the users' channels on the N_R receive antenna elements. For the purpose of applying our SS decomposition channel estimation, the channel was assumed to be constant over at least one data block (quasi-static), when it is gradually varying with the specified f_d . Fig. 1 shows the NMSE of the channels estimates with SS decomposition channel estimation done over every data block with f_d = 5 Hz, when we vary the data block size N. It is evident that estimation performance degrades as the size of the data block decreases as was stated previously in the introduction. The frequency- domain equalizer of Section II-B was implemented to separate the UE symbols in the frequency-domain, and then recover the transmitted data in the sampled time-domain of each UE using an individual IFFT operation. We show BER curves in Fig. 2 when the four active UE are transmitting using 4-QAM, 16-QAM, and 64-QAM formats. For each modulation format, a reference BER curve was generated where the system was assumed to have perfect knowledge of the channels (actual values) as a bench-mark. By inspecting Fig. 2, we can deduce the sensitivity of the SS decomposition estimation algorithm to noise level, which is manifested in deviation from the *perfect estimate* BER curve when 4-QAM is deployed. This deviation is less significant for the 16-QAM

Table I. Simulation Parameters

Parameter	Value
Number of symbols in a block (N)	2048
Data block duration	66.67 µs
Cyclic prefix lenght	4.69 µs
Modulation type	4-QAM/16-QAM/64-QAM
Number of active UE (N_T)	4
Number of of receive antenas (N_R)	8
Channel order (M)	9
Stacking Parameter (L)	42

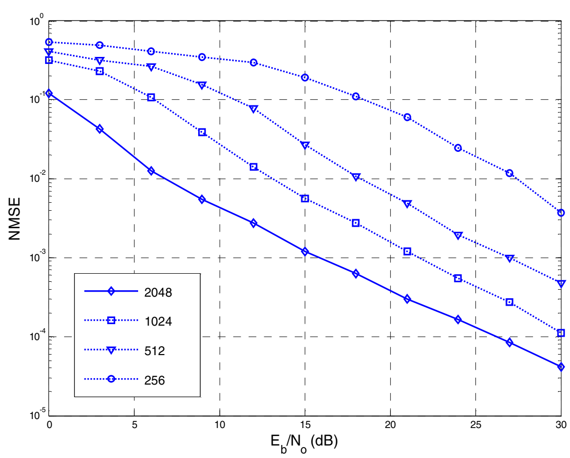

Fig. 1. NMSE of SS-based MIMO channel estimates as a function of SNR with N_T =4, N_R = 8, and diffrent values of block size N at f_d = 5Hz.

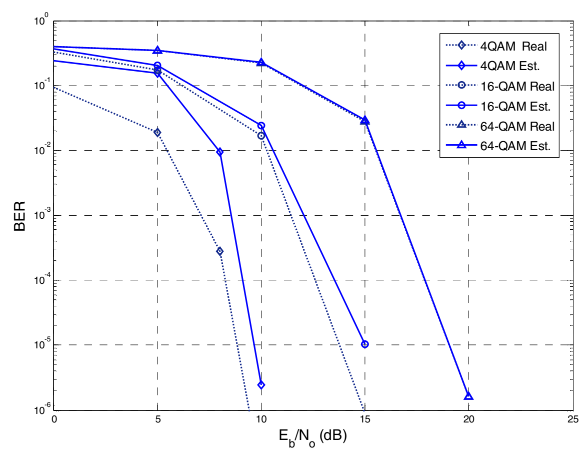

Fig. 2. BER of MIMO SC-FDE system employing SS-based MIMO channel estimatimation with N_T =4, N_R = 8, and N =2048 at f_d = 5Hz.

format, and virtually non-existent in the 64-QAM case. This is because the latter two modulation formats require relatively higher SNR than 4-QAM, which will also be reflected in better SS decomposition estimation for the MIMO channel. In the case of 16-QAM modulation, the bit-error-rate (BER) of the proposed semi-blind scheme is almost identical to that of a system employing the true (but unknown) channels. Finally, Fig. 3 compares BER performance curves of the three considered modulation formats with Doppler rates of 5 Hz, 70 Hz, and 140 Hz, when SS decomposition MIMO channel estimation was performed once every 4 data blocks, and the MMSE weights are used over the entire 4 data blocks. Inspection of Fig. 3 leads us to conclude that reducing computational cost is possible while maintaining desired data rates in a low (5Hz) to moderate (70Hz) Doppler rate environments at a specific SNR. Also, we are able to trade-off data rate for improved performance at the available SNR as is commonly know. Also, using the specified parameters of our simulated system, we only need 5.6% of the required number of transmitted pilot symbols when compared with traditional training-based channel estimation techniques.

V. CONCLUSION

In this paper, we have investigated the use of subspace decomposition for the purpose of blind channel estimation in MU-MIMO uplink SC-FDE systems that utilize a MMSE frequency-domain equalizer. The channel estimation was reformulated as the minimization of a quadratic cost function, computed from the noise subspace eigenvectors of an array output correlation matrix. The large number of symbols in each LTE data block allows for high accuracy in estimating the correlation matrix, which in-turn produces very accurate channel estimates. This accurate CSI, when used in conjunction with the MMSE multiuser equalizer, leads to the desired high performance in the MU-MIMO SC-FDE transmission while simultaneously increasing the spectral efficiency due to reduction in the number of required pilot symbols. The advantages of our proposed blind subspace-based channel estimation algorithm and the feasibility of the resulting MU-MIMO SC-FDE schemes were demonstrated through simulations of multi-user uplink transmissions in an LTE-Advanced type of environment. In particular, the proposed blind subspace decomposition channel estimation is a strong contender to complement the operation of MU-MIMO SC-FDE systems and support improved spectral efficiency in high data rate set-ups.

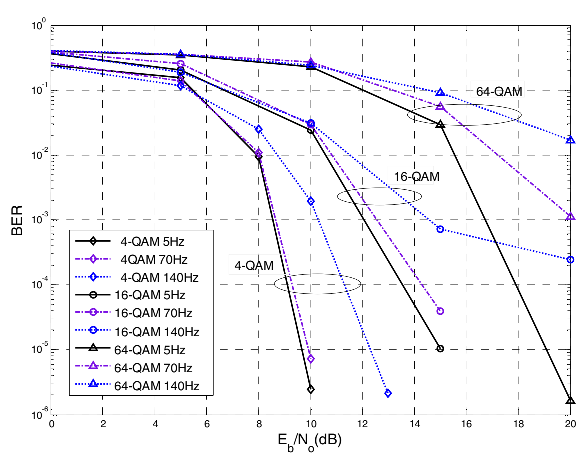

Fig. 3. BER with Doppler rates of 5Hz, 70Hz, and 140Hz when channel estimation is performed every 4 data blocks for the N_T=4 MIMO users.

REFERENCES

[1] 3GPP TS 36.201, "Evolved Universal Terrestrial Radio Access (E-UTRA); LTE Physical Layer General Description," Release 10, 2010.

[2] S. Sesia, I. Toufik, M. Baker, *LTE–the UMTS Long Term Evolution : from theory to practice*, 2nd ed., Chichester, UK: John Wiley & Sons, 2011.

[3] G. Bauch, G. Dietl, "Multi-user MIMO for achieving IMT-Advanced requirements," *Proc. Telecommunications, 2008. ICT 2008. International Conference on*, June 2008.

[4] C. Park, Y. Wang, G. Jöngren, D. Hammarwall, "Evolution of uplink MIMO for LTE-advanced," IEEE Comm. Mag., vol. 49, pp. 112-121, February 2011.

[5] J. Milleth, K. Kuchi, V. Ramaswamy, Dhivagar, K. Giridhar, B. Ramamurthi, "Localized and distributed SC-FDMA and OFDMA for uplink multiple access," *IEEE 802.16 Broadband Wireless Acess Working Group (IEEE C802.16m-08/169r1)* , pp. 1-10, March 3, 2008.

[6] G. Berardinelli, C. Manchón, L. Deneire,T. Sorensen, B. Troels P. Mogensen, K. Pajukoski, "Turbo receivers for single user MIMO LTE-A uplink," *Proc. IEEE 69th Vehicular Technology Conference Spring (VTC 2009-Spring)*, April 2009.

[7] 3GPP TS 36.211, "Evolved Universal Terrestrial Radio Access (E-UTRA); Physical Channels and Modulation," Release 10, 2010.

[8] A. Tahat, "A matrix scheme to extrapolation and interpolation for a 4G MIMO OFDM system," *Proc. IEEE 72nd Vehicular Technology Conference Fall (VTC 2010-Fall)*, Sept. 2010.

[9] C. Shin, R. Heath, E. Powers, "Non-redundant precoding-based blind and semi-blind channel estimation for MIMO block transmission with a cyclic prefix," IEEE Trans. Sig. Proc., vol. 46, pp. 2509-2523, June 2008.

[10] J. Wu, "How much coherent interval should be dedicated to non-redundant diagonal precoding for blind channel estiamtion in single-carrier block trasmission?," IEEE Trans. Wireless Com., vol. 9, pp. 2560-2574, August 2010.

[11] C. Tu, B. Champagne, "Subspace-based blind channel estimation for MIMO-OFDM systems with reduced time averaging," IEEE Trans. Vehicular Tech., vol. 59, pp. 1539-1544, March 2010.

[12] F. Gao , Y. Zeng , A. Nallanathan, T.-S. Ng "Robust subspace blind channel estimation for cyclic prefixed MIMO OFDM systems: algorithm, identifiability and performance analysis", *IEEE J. Sel. Areas Commun.*, vol. 26, pp. 378 -388, Feb. 2008.

[13] P. Loubaton, E. Moulines, and P. Regalia, "Subspace method for blind identification and deconvolution," in *Signal Processing Advances in Wireless & Mobile Communications, vol. I: Trends in Channel Estimation*, G. Giannakis, Y. Hua, P. Stoica, and L. Tong, Eds. Englewood Cliffs, NJ: Prentice-Hall PTR, 2001, pp. 115–117.

[14] S. Visuri, V. Koivunen, "Resolving ambiguities in subspace-based blind receiver in MIMO channels," *Proc. 36th Asilomar Conf. on Signals, Systems, and Computers*, pp. 589-593, Nov. 2002.

[15] C. D'Amours, B. Champagne, A. Dahmane, "Subspace decomposition for channel estimation in SC-FDE systems," *Proc. IEEE 77th Vehicular Tehcnology Conference Spring (VTC 2013-Spring)*, pp. 1-5, June 2013.

[16] E. Moulines, P. Duhamel, J.-F. Cardoso, S. Mayrarue, "Subspace methods for the the blind identification of multichannel FIR filters," IEEE Trans. Sig. Proc., vol. 43, pp. 516-525, February. 1995.

[17] K. Abed-Meraim, Y. Hu, "Blind identification of multi-input multiple-output system using minimum noise subspace," IEEE Trans. Sig. Proc., vol. 45, pp. 254-258, Jan. 1997.

[18] D. Kotoulas, P. Koukoulas, N. Kalouptsidis, "Subspace projection based blind channel order estimation of MIMO systems," IEEE Trans. Sig. Proc., vol. 54, pp. 1351-1363, April 2006.