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Subspace Decomposition Channel Estimation for Multiple Virtual MIMO SC-FDMA Systems

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Abstract—**In this paper, we investigate the use of a subspace decomposition MIMO channel estimation algorithm for the blind estimation of the multiple virtual MIMO (V-MIMO) channels in Single Carrier Frequency Division Multiple Access (SC-FDMA) systems in order to increase its spectral efficiency. Groups of multiplexed users are formed, where each user is equipped with a single antenna, to transmit on the same frequency and time resources. When deployed in a V-MIMO system that resembles the LTE standard parameters, the proposed blind algorithm can obtain accurate estimates of the V-MIMO channels for each active user over every block of transmitted data. In addition, it is necessary to have knowledge of the channels state information to perform the users grouping and frequency allocation (scheduling). According to the LTE standard, sounding reference signals are employed in conjugation with Interleaved- frequency division multiple-access (I-FDMA). By eliminating the need to transmit an entire block of pilot symbols for each active transmitter, this permits for additional bandwidth efficiency in V-MIMO SC-FDMA systems in which multiple users occupy the same set of subcarriers as indicated by the LTE standard. The advantages of employing our proposed blind subspace-based channel estimation algorithm are verified by simulations results over fading channels, which demonstrate its practicality for use in the resulting V-MIMO SC-FDMA scheme with reduced training.**

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

I. Introduction

Following its employment for uplink transmission in 3GPP Long Term Evolution (LTE) mobile wireless communication systems [1], single-carrier frequency division multiple-access (SC-FDMA) has received significant attention in recent years. Its low peak-to-average power ratio (PAPR) is the key feature that benefits the mobile user equipment (UE) when compared to orthogonal FDMA (OFDMA) techniques [2] that are employed in the downlink. To increase the spectral efficiency in various wireless communication technologies, multiple-input multiple-output (MIMO schemes have been deployed to increase their respective spectral efficiencies. Nevertheless, particularly in small devices of the user equipment (UE), practical applications and implementation of MIMO for the uplink is limited due to size and cost [3], [4]. In order to alleviate these issues, the virtual MIMO (V-MIMO) approach has emerged for the uplink for the scenario of the simultaneous transmission of multi-user signals, with single transmit antenna each, on the same frequency and time resources [3-7].

In [8], in order to increase spectral efficiency and mitigate interference, the authors had considered the special case when mobile users occupied all available subcarriers for transmission, rather than using a subset of all available subcarriers in the allocated bandwidth as frequently performed in LTE systems. In this scenario, the single-carrier frequency-domain equalization (SC-FDE) terminology becomes more appropriate since in effect, the multiple-access is realized in the spatial only, as opposed to the frequency domain. In [9], using *perfect* knowledge of the corresponding spatial channels, the authors have demonstrated in an SC-FDMA system that multiple single antenna UEs can effectively transmit simultaneously on the same frequency band while multi-antenna receivers are used at the base station BS, thereby forming virtual MIMO sub-systems.

The base station (BS) manages all resource allocations including users' grouping for multiplexing and spectrum allocation for all simultaneously active UE within a cell. It decides when and which UE can share the same frequency-time resources, in a way that will minimize interference among them [2-7]. Since scheduling and accessing the available spectrum are critical issues that directly affect SC-FDMA performance [5-6], considerable amount of research works have been conducted to develop procedures and algorithms for the optimal selection of partners to form V-MIMO groups in SC-FDMA systems. Channel state information (CSI) is an essential component of the functions of these procedures and algorithms. Moreover, as the ability of the BS to separate signals that are simultaneously received on the same frequencies is greatly dependent upon accurately estimating the channels. LTE-A specifically provides orthogonal demodulation reference (pilot) signals using different cyclic time shifts on a complete frame to enable the BS to obtain channel estimates for each one of the supported physical layer UEs in uplink MU-MIMO [10]. Therefore, eliminating or reducing the need for transmission

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of such pilot symbols will further improve the efficiency in the utilization of the time-frequency resource blocks. In this paper, we propose to extend the blind subspace decomposition approach for application in estimating the V-MIMO channels in SC-FDMA systems, where each simultaneous active UE in a group transmits on the same time and frequency resources utilizing only a subset of the available subcarriers in the allocated frequency band as depicted in Fig. 1. This variation in subcarrier allocations as opposed to the one deployed in the SC-FDE V-MIMO scenario in [8], where each active UE occupied all of the available subcarriers in the frequency band, will require independent parallel processing of each of V-MIMO system users groups. As a result, special processing has to be performed to properly separate the V-MIMO groups to conduct the blind subspace channel estimation. Since the subspace channel estimation approach depends greatly on the high accuracy in estimating the correlation matrix over one block of received data [8], this variation will also reflect on the accuracy of produced channel estimates. We investigate in this paper the performance of the proposed blind V-MIMO subspace channel estimation in SC-FDMA systems when employed in conjunction with the MMSE multiuser equalizer to increase the spectral efficiency due to reduction in the required number of pilot symbols.

The rest of the paper is organized as follows: Section II describes the structure of the considered V-MIMO SC-FDMA 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

Each of the N_T active UE transmitters in the V-MIMO SC-FDMA system has a single antenna element to transmit to the N_R antennas (demodulators) at the base station (eNodeB) using the same time and frequency resources in a synchronous fashion [3], [8]. The considered uplink multiuser MIMO block transmission system uses interleaved carrier distribution (i.e., IFDMA) [2] as illustrated in Fig. 1.

A. Transmitter and Receiver Structure

As was depicted in [9], the input and output of the transmitter and receiver are complex modulation symbols that can reach up to 64-level quadrature amplitude modulation in the case of very strong channels [9], [10]. For each user, the data block that is being sent consists of N complex modulation symbols. The M-ary quadrature amplitude modulated (M-QAM) data transmitted by the i^{th} mobile user at the l^{th} block (frame) as :

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

where $x_i^l(n)$ is the n^{th} complex-valued symbol in user i's data frame, and N is the block size. The N point discrete

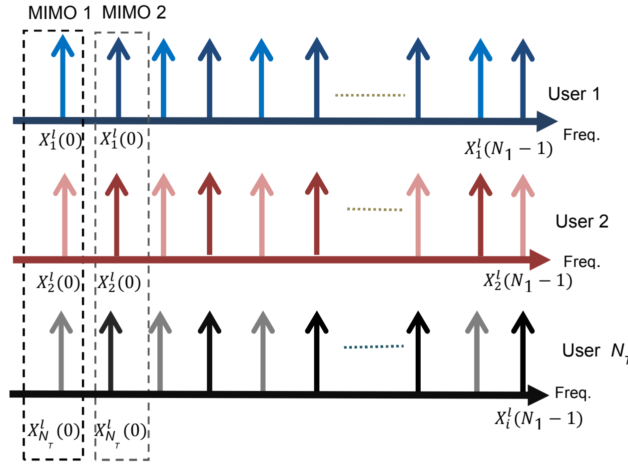

Fig. 1. Subcarriers distribution in a V-MIMO SC-FDMA system where each simultaneous active UE in a MIMO group transmits on the same time and frequency resources utilizing only a subset of the available subcarriers.

Fourier Transform (DFT) takes N modulation symbols at a time and maps it to N out of N_2 orthogonal sub-carriers spread over the available frequency band [7]. The bandwidth spreading factor is denoted by Q and given by $Q = N_2/N$. An SC-FDMA system can handle up to Q source signals with each signal occupying a set of N orthogonal subcarriers [9]. After subcarrier mapping, an N_2 point inverse discrete Fourier Transform (IDFT) generates the time domain representation of the N sub-carriers symbols and a parallel-to-serial operation places it into a time sequence suitable for transmission [2]. Before transmission through the channel, a cyclic prefix (CP) of length, N_{CP}, is added. Each of the N_T users' transmitters in the Q groups of V-MIMO transmitters performs these operations on the l^{th} data block (frame) in an identical fashion, where it is understood that the UEs are scheduled to transmit on the same time and frequency resources within the same group as depicted in Fig. 1.

After passing through the multipath channels between each of the UE transmitters and each of the receiver antenna elements (sensors), each of the N_R antenna elements receives the sum of the N_T active users' signals of the Q V-MIMO groups at the V-MIMO SC-FDMA receiver [7-8]. The inverse of the operations that were performed at the transmitters will be performed on the received (sum) signal at each of the elements [7]. The result is that the sum of each group of N_T transmitted signals are separated and placed in a time-domain vector at each of the N_R antennas.

Due to utilization of IFDMA in the V-MIMO SC-FDMA transmission, we process each of the Q V-MIMO groups independently in parallel arrangements because of the orthogonality among the individual IFDMA subcarriers in the SC-FDMA transmitters [2].

For a given V-MIMO group, the $N_R \times 1$ received signal vector, at each sampled-time domain instance, has the form of the standard equivalent baseband finite impulse response MIMO (FIR-MIMO) channel model described in [13-14] and used in [8] that has N_T transmitters and N_R receivers.

As was shown in [8], we can express the received signal on the N_R antenna elements at each sampled-time instance as:

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

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:

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

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

The received signal on antenna element j, after the removal of the cyclic prefix, can be written as [8]:

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

where $\mathbf{y}_j^l = \left[\, y_j^l(0), y_j^l(1), \dots, y_j^l(N-1)\right]^T$, $\mathbf{h}_{i,j}^l = [\, h_{i,j}^l(0)\ , h_{i,j}^l(1), \dots, h_{i,j}^l(M)\ \ ]^T$ 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 ⊗ denotes circular convolution, and M represents the channel order, with the assumption that it does not 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

We take the N-point DFT of (4) and stack the frequency-domain symbols at each subcarrier k in a $N_R \times 1$ vector as [11]:

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

at the BS V-MIMO receiver side to separate the N_T mobile users' data signals. In (5), $\mathbf{H}_i^l(k) = \left[\, H_{i,1}^l(k), \dots, H_{i,N_R}^l(k)\right]$ contains the i^{th} user estimated channel frequency response at subcarrier, k, on each of the receiving antenna elements. The frequency-domain noise, $\mathbf{Z}^l(k)$, is expressed as: $\mathbf{Z}_i^l(k) = \left[\, Z_1^l(k), \dots, Z_{N_R}^l(k)\right]$. We express channel transfer function matrix of size the $N_R \times N_T$ 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 define vector $\mathbf{X}^l(k) = \left[\, X_1^l(k), \dots, X_{N_T}^l(k)\right]^T$ in order that we can rewrite (5) as:

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

The error minimizing, in the mean square sense (MSE), matrix-valued filter, $\mathbf{W}^l(k)$, of the error $\mathbf{e}^l(k) = \mathbf{W}^l(k)^H\, \mathbf{Y}^l(k) - \mathbf{X}^l(k)$ (i.e., MMSE) is given by [11], [8]:

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

Then, the MMSE filtered signal is expressed in the frequency-domain as, $\widehat{\mathbf{X}}^l(k) = \mathbf{W}^l(k)^H \mathbf{Y}^l(k)$. Now, to obtain the sampled-time domain pre-detection symbols, an N-point IDFT is performed on of each of the individual users filter signals.

In order for this V-MIMO system to function properly as was mentioned previously, 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

The authors of [8],[12] had demonstrated the advantages and effective use of the well-known Subspace (SS) based blind channel estimation for SC-FDE and SC-FDMA systems and formulated novel channel estimation algorithms for each particular scenario. SS based blind estimation methods were first introduced in [13] for the SIMO case, and later expanded to the MIMO scenario [14], derive their properties from the second-order statistics (SOS) of the received signals. In [8] the authors had re-casted the SS method and formulated an algorithm for application in SC-FDE V-MIMO systems. Here we treat each of the Q V-MIMO groups of the SC-IFDMA transmission in the SC-FDMA system as an independent SC-FDMA subsystem. Employing the SS MIMO channel estimation from [14-15] to the currently proposed V-MIMO SC-FDMA system in a similar manner to [8] is a delicate operation. We investigate the applicability and feasibility of this scenario where the total available subcarriers in the frequency-band are sub-divided among the V-MIMO groups rather than a single group of the SC-FDE system as in [8]. Therefore, we re-describe the subspace decomposition for the (semi) blind estimation of the sampled time-domain V-MIMO channels of the N_T simultaneously active UEs [8], where it is understood that this algorithm will be applied to the parallel processing of the Q V-MIMO groups for the blind estimation of the simultaneously active UEs channels in the group. It is assumed that the variations in the channels are sufficiently slow to be considered static over a time interval of one data block. Then individual DFT operations are carried-out to obtain the corresponding frequency-domain estimates for the application of (8).

A number of L successive observations ($L < N$) are taken of (2) and are stacked into an $N_R L\times1$ vector as:

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

Then (2) can be expressed as:

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

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 generalized Sylvester channel matrix $\mathcal{H}_L(\mathbf{h}^l)$ is the $N_R L \times N_T (L + M)$ matrix of order L associated to the matrix $\mathbf{h}^l = [\,\mathbf{h}^l(0)^T, \mathbf{h}^l(1)^T, \dots, \mathbf{h}^l(M)^T]^T$ is expressed as [15], [8]:

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

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 [14], [15]. Let $\boldsymbol{R}_y$ be the $N_R L \times N_R L$ correlation matrix of $\underline{\mathbf{y}}^l(n)$ expressed as:

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

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 $L > N_T M$ [15], and $\mathcal{H}_L(\mathbf{h}^l)$ has full column rank N_T ($L + M$). It is assumed that, M, the maximum channel order, is known [14-15]. Hence, $\boldsymbol{R}_y$ can be expressed as :

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

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. $\boldsymbol{\Lambda}_x = diag(\lambda_1, \dots, \lambda_{N_T(L+M)})$, are the signal eigenvalues with $\lambda_1 \ge \lambda_2 \ge \cdots \ge \lambda_{N_T(L+M)} \ge \sigma_z^2$. As was stated in [8] and the references therein, using orthogonality relation between noise and signal subspace is the cornerstone of the subspace method that leads to the standard result of:

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

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 [8], [15], we first denote the noise subspace eigenvectors by ($\mathbf{g}_m$, $m = 1, \dots, r$), and partition each eigenvector as:

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

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

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

It can be shown [13-15], that for each of the columns of size N_R ($M+1$) × 1 of the matrix $\mathbf{h}^l$ (i.e., $\mathbf{h}_i^l$, $i = 1, \dots, N_T$) we have:

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

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

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

is N_T. Therefore, this leads to:

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

where $\mathbf{B}$ is the N_R ($M+1$) × N_T matrix of the eigenvectors of Q corresponding to the eigenvalues that are equal to zero [12], [8]. $\mathbf{R}^{-1}$ is an invertible $N_T \times N_T$ ambiguity matrix. To determine the ambiguity matrix, we need to use a small number of pilot symbols interleaved in each UE data block as has been discussed in the literature [15]. We only need (N_T /M N_R) of the required number of transmitted pilot symbols for each V-MIMO group when compared with traditional training-based channel estimation techniques for use in V-MIMO SC-FDMA systems.

IV. Simulation Results

We constructed a Matlab simulation with parameters that resemble the proposed V-MIMO SC-FDMA LTE-A like system to evaluate the effectiveness and performance of the formulated subspace decomposition based MIMO channel estimation algorithm. The performance was evaluated through the use of Monte-Carlo simulations runs. Table I. lists the main simulation parameters. The performance measures we consider here are normalized mean square error (NMSE), defined as:

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

and bit error rate (BER). Both measures were plotted against the energy per bit to noise single sided spectral density ratio, (E_b/N_o). Each simulation point in the results was produced as the average of 200 Monte-Carlo realizations. The frequency-selective MIMO channel model from each of the UE antenna to each of the N_R receive antenna elements had 10 Rayleigh faded paths where the power delay profile was assumed to be uniform. The spatial correlation between the users' channels on the N_R receive antenna elements was assumed to be very low. Also, when performing computations for the SS decomposition channel estimation, the channel was considered to be constant over at least one data block.

TABLE I. SIMULATION PARAMETERS

Parameter	Value
SC-FDMA FFT Size (N_2)	2048
Number of symbols in a block (N)	1024
SC-FDMA Data block duration	66.67 µs
Cyclic prefix length	4.69 µs
Modulation type	16-QAM/64-QAM
Number of active UE (N_T) per Carrier	1 and 4
Number of of receive antenas (N_R)	8
Channel order (M)	9
Doppler frequency (f_d)	5 Hz
Stacking Parameter (L)	42

Fig. 2 shows the NMSE of the channels estimates with SS decomposition channel estimation done over every data block with two different scenarios: 1-user per carrier (i.e.. no V-MIMO) and with 4-users per carrier (i.e., $N_T = 4$). It is evident that estimation performance degrades as the number of users in the V-MIMO group that are sharing the same subcarriers increase, as a result of multiple access interference. However, this reduction in performance is approximately 2dB when the number of active users is equal to 4, and is perfectly justified in return for the gained improvement in spectral efficiency using V-MIMO.

As in [8], 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. 3 when the four active SC-FDMA UE are transmitting using 16-QAM, and 64-QAM formats. A reference BER curve was generated for each modulation format and plotted as a bench-mark, where the system was assumed to have perfect knowledge of the channels (actual values).

By inspecting Fig. 3, we can confirm the sensitivity of the SS decomposition estimation algorithm to noise level, which is manifested in deviation from the *perfect estimate* BER curve in both of the 16-QAM and 64-QAM implementations due less than perfect channel estimation in the SS algorithm at these SNRs. This deviation is less pronounced for the 64-QAM format. This is because the latter requires relatively higher SNR than 16-QAM, which will also be reflected in better SS decomposition estimation for the V-MIMO channels. In addition, we are able to trade-off data rate for improved performance at the available SNR as is commonly known.

V. CONCLUSION

In this paper, we have investigated the use of subspace decomposition for the purpose of blind channel estimation in multiple V-MIMO uplinks of SC-FDMA systems that utilize a MMSE frequency-domain equalizer.

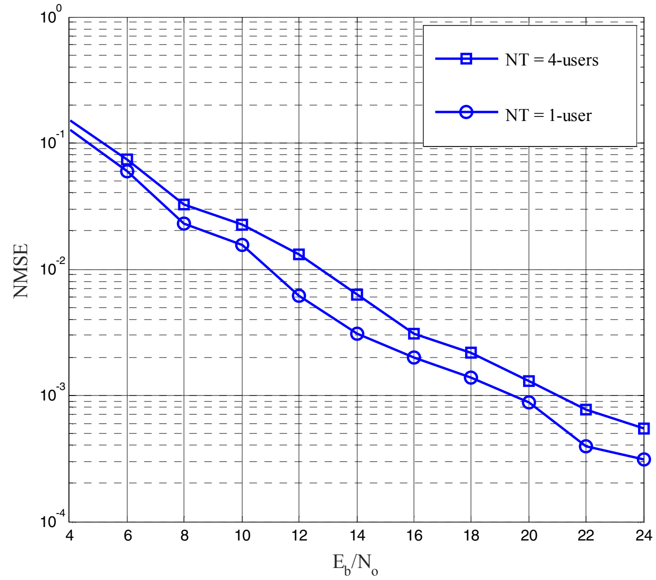

Fig. 2. NMSE of SS-based MIMO channel estimates as a function of SNR with N_T=1 or N_T=4, N_R = 8, and a block size N=1024.

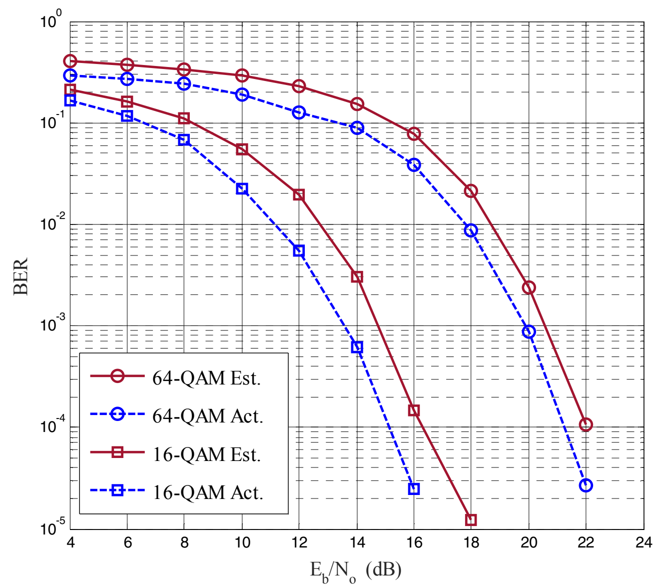

Fig. 3. BER of MIMO SC-FDMA system employing SS-based MIMO channel estimatimation with N_T=4, N_R = 8, and N=1024 and N_2=2048 .

As was similarly performed for the V-MIMO SC-FDE scenario in which the simultaneously active UE occupied all available subcarriers, the channel estimation was computed from the noise subspace eigenvectors of an array output correlation matrix. Desired increase in the spectral efficiency was attained due to reduction in the number of required pilot symbols in the multiple V-MIMO SC-FDMA transmissions. The advantages of our proposed blind subspace-based channel estimation scheme and the feasibility of the resulting V-MIMO SC-FMDA schemes were verified through simulations of multi-user uplink transmissions with parameters that resemble a practical environment setting. The proposed blind subspace decomposition channel estimation can be a strong candidate to support improved spectral efficiency in high data rate set-ups and to enhance the operation of future V-MIMO SC-FDMA systems.

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