This full-text paper was peer-reviewed and accepted to be presented at IEEE VTC 2013 Fall

A Reduced-Complexity Detector for OFDMA/SC-FDMA-Aided Space-Time Shift Keying

Mohammad Ismat Kadir, Sheng Chen and Lajos Hanzo
School of ECS, University of Southampton, SO17 1BJ, UK
Email: mik1g09@ecs.soton.ac.uk, sqc@ecs.soton.ac.uk, lh@ecs.soton.ac.uk

Abstract—**We propose a novel reduced-complexity detector for the orthogonal frequency division multiple access (OFDMA)/single-carrier frequency division multiple access (SC-FDMA)-aided space-time shift keying (STSK) architecture. STSK employing OFDMA/SC-FDMA has recently been shown to be beneficial in dispersive multiuser downlink/uplink scenarios. These schemes exhibit excellent performance at a considerably reduced decoding complexity. We can employ a single-stream maximum-likelihood (ML) detector, since a single dispersion matrix (DM) is activated at any signalling interval. In this paper, we propose a new detector, which is capable of further reducing the decoding complexity. The proposed detector is particularly suitable for STSK-based transmission over frequency-selective multiple-input multiple-output (MIMO) channels employing frequency-domain equalization (FDE). The complexity of the proposed scheme is quantified and it is observed that the scheme maintains its superior multiple-antenna performance at a significantly reduced complexity.**

I. Introduction

Space-time shift keying (STSK) [1]–[3] has recently been proposed for striking a flexible diversity versus multiplexing gain tradeoff for multiple-input multiple-output (MIMO) systems, while benefitting from its low-complexity single-stream detector. The key idea behind designing STSK is that it spreads the source information in both the spatial- and time-domain, like the family of linear dispersion codes (LDCs) [4], [5] in order to attain both diversity as well as multiplexing gain. Motivated by the low-complexity design of spatial modulation (SM) [6] as well as space-shift keying (SSK) [7], STSK was conceived for activating only one from a set of LDC-type dispersion matrices (DMs) in addition to the transmission of classic modulation symbols, thereby providing the same diversity as well as multiplexing gain as LDCs, whilst imposing a substantially reduced decoding complexity.

Since the SM and the SSK schemes activate only one antenna element (AE) during each symbol interval, the matched-filter (MF) [6], the maximal-ratio combining (MRC) [8] and maximum-likelihood (ML) [8] detectors using the signal stream of a single AE can be employed, since no inter-element-interference (IEI) is experienced at the receiver. Hence, the STSK scheme may also employ a single-stream ML [1] detector. In order to mitigate the impairments due to the channel-induced dispersion imposed by wideband channels, OFDM-aided STSK [9] and OFDMA/SC-FDMA aided STSK [10] have been proposed for the single-user downlink (DL) and the multiuser downlink/uplink (DL/UL), respectively. All these schemes benefit from the employment of the single-stream ML detector proposed in [1]. To further reduce the complexity of STSK, especially when bandwidth-efficient quadrature amplitude modulation (QAM) is used, two further detectors modifying the MF detectors of [6] were proposed in [11].

Against this background, we propose a new detector, which further reduces the decoding complexity, especially in OFDMA/SC-FDMA-aided STSK arrangements. The proposed detector employs MIMO minimum mean-squared error (MMSE) equalization, followed by the estimation of the index of the DM activated and of the constellation symbol utilized for constructing the STSK codewords. The main contributions of this paper are as follows.

1) We propose a novel detector for the STSK scheme, which is particularly suitable for the OFDMA/SC-FDMA-aided STSK DL/UL proposed in [10] without sacrificing the performance achieved.
2) The complexity of the scheme is quantified in terms of the number of real-valued multiplications and the scheme is found to have a low complexity.

The rest of the paper is organized as follows. Section II gives an overview of the STSK system and the conventional detectors. The proposed detector is illustrated in Section III. In Section IV, the complexity associated with the scheme is evaluated. The complexity as well as performance of the scheme advocated is compared to those of the existing schemes in Section V. Finally, we conclude in Section VI.

Notations: In general, we use capital boldface letter, for example, $\boldsymbol{A}$ to denote a matrix, whereas $\boldsymbol{A}^T$, $\boldsymbol{A}^H$, $\text{tr}(\boldsymbol{A})$ and $\parallel \boldsymbol{A} \parallel$ denote the matrix transpose, the hermitian transpose, the trace and the Euclidean norm of $\boldsymbol{A}$ respectively. The operator $\mid a \mid$ is used for the complex norm of the complex number a, $\otimes$ for the Kronecker product, $\boldsymbol{I}_T$ for the $(T \times T)$-identity matrix and $\Re(\cdot)$ and $\Im(\cdot)$ for the real and the imaginary parts of a complex number, respectively. Furthermore, $\mathcal{E}\left\{\bullet\right\}$ represents the expectation of '$\bullet$' and comp$\left[\bullet\right]$ refers to the complexity associated with the computation of the expression '$\bullet$'.

II. System Overview

In this section, we provide a brief overview of the STSK system, the OFDMA/SC-FDMA-aided STSK and of the existing detectors. The motivation for conceiving a new detector is also outlined.

A. STSK System Architecture

The STSK transmitter maps some of the source bits to the conventional modulation symbols and the rest of the bits to the index of the DM activated from the set of Q matrices, $\boldsymbol{A}_q\,(q = 1,...,Q)$. To be specific, the transmitter spreads the input bits across the M transmit antenna elements (AEs) during T signalling intervals to form the space-time (ST) codeword $\boldsymbol{X}(i) \in \mathbb{C}^{M\times T}$ according to

$$\boldsymbol{X}\,[i] = s\,[i]\,\boldsymbol{A}\,[i]\,, \tag{1}$$

where $s\,[i]$ is an $\mathcal{L}$- phase-shift keying/quadrature amplitude modulation (PSK/QAM) symbol, while $\boldsymbol{A}\,[i]$ is one of the pre-assigned DMs, $\boldsymbol{A}_q\,(q = 1,...,Q)$, which are generated by the optimization method [1], [12] under the power constraint of:

$$\text{tr}(\boldsymbol{A}_q^H \boldsymbol{A}_q) = T \qquad \forall q. \tag{2}$$

The corresponding system is unambiguously referred to as the STSK(M,N,T,Q) scheme operating in conjunction with $\mathcal{L}$-PSK/QAM, where N is the number of the receive AEs. We note that the STSK system and the existing detectors were proposed mainly for a single-user system and hence the user index has been omitted here and in Subsection II-C. The block-based model of a STSK system may be expressed as

$$\boldsymbol{Y}\,[i] = \boldsymbol{H}\,[i]\,\boldsymbol{X}\,[i] + \boldsymbol{V}\,[i]\,, \tag{3}$$

where $\boldsymbol{Y}\,[i] \in \mathbb{C}^{N\times T}$ represents the received signal, $\boldsymbol{X}\,[i] \in \mathbb{C}^{M\times T}$ is the space-time (ST) transmit block and i is the index of the STSK block. Moreover, $\boldsymbol{H}\,[i] \in \mathbb{C}^{N\times M}$ and $\boldsymbol{V}\,[i] \in \mathbb{C}^{N\times T}$ denote the channel impulse response (CIR) and noise elements respectively, obeying the complex-valued zero-mean Gaussian distribution of $\mathcal{CN}(0,1)$ and $\mathcal{CN}(0,N_0)$ respectively, where N_0 indicates the noise variance.

Re-arranging (3) by the linearized system model of [5] employing the vectorial stacking operation vec$\,(\cdot)$, we have

$$\bar{\boldsymbol{Y}}\,[i] = \bar{\boldsymbol{H}}\,[i]\,\boldsymbol{\chi}\boldsymbol{K}\,[i] + \bar{\boldsymbol{V}}\,[i]\,, \tag{4}$$

where

$$\bar{\boldsymbol{Y}}\,[i] = \text{vec}\,(\boldsymbol{Y}\,[i]) \in \mathbb{C}^{NT\times 1}, \tag{5}$$
$$\bar{\boldsymbol{H}}\,[i] \triangleq \boldsymbol{I}\otimes\boldsymbol{H}\,[i] \in \mathbb{C}^{NT\times MT}, \tag{6}$$
$$\boldsymbol{\chi} \triangleq [\text{vec}\,(\boldsymbol{A}_1)\,,\cdots,\text{vec}\,(\boldsymbol{A}_Q)] \in \mathbb{C}^{MT\times Q}, \tag{7}$$
$$\boldsymbol{K}\,[i] \triangleq [\underbrace{0,\cdots,0}_{q-1},\ s\,[i]\,,\underbrace{0,\cdots,0}_{Q-q}]^T \in \mathbb{C}^{Q\times 1}, \tag{8}$$
$$\bar{\boldsymbol{V}}\,[i] = \text{vec}\,(\boldsymbol{V}\,[i]) \in \mathbb{C}^{NT\times 1}. \tag{9}$$

B. OFDMA/SC-FDMA-aided STSK

Recently OFDMA-aided STSK DL and SC-FDMA-aided STSK UL have been proposed in [10], which are capable of mitigating the impairments of realistic dispersive channels while facilitating multi-user transmissions. The SC-FDMA based STSK scheme relying on the interleaved subcarrier allocation has been advocated in the UL scenario, as a benefit of its low peak-to-average power ratio (PAPR).

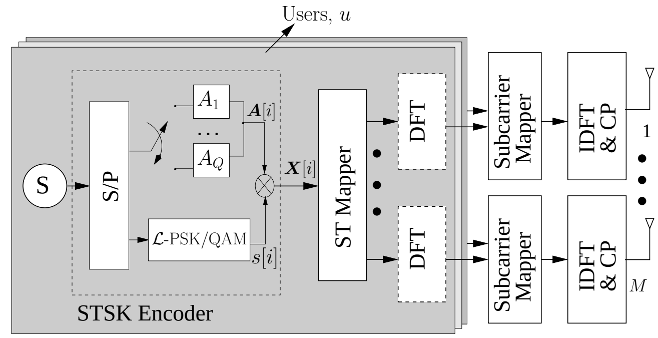

Fig. 1. Simplified architecture of the OFDMA/SC-FDMA-aided STSK transmitter. The dotted 'DFT' block does not exist in the OFDMA-aided scheme.

As shown in Fig. 1, the original frequency-domain (FD) symbols of the OFDMA-aided scheme or the discrete Fourier transformed (DFT) symbols of the SC-FDMA-aided scheme are mapped to a number of subcarriers. The resulting signal is then transmitted after the inverse discrete Fourier transform (IDFT) operation from each of the transmit antenna elements (AEs) of our STSK system.

The signals received by the receive AEs are first equalized in the frequency-domain (FD) using MMSE or zero-forcing (ZF)-based single-tap MIMO FD equalizer. The source information is then estimated by a single-stream based detector in the FD for OFDMA or in the time-domain (TD) for SC-FDMA. We observe that after the employment of the MIMO FDE, the data estimation does not involve the channel matrix. As a result, the single-stream ML detection involves less multiplications and additions and thereby imposes a lower complexity.

C. Existing Detectors

In this section, we briefly describe the optimal ML detector [1], [8] and the two detectors proposed in [11]. In the following, we omit the block index i for the sake of notational simplicity.

The detectors estimate both the index of the activated dispersion matrix (DM) q and the index of the transmitted constellation symbol l corresponding to space-time block index i and thereby estimates the source information.

1) The Optimal ML Detector: The single-stream based ML find the estimates $(\hat{q},\hat{l})$ from:

$$\left(\hat{q},\hat{l}\right) = \arg\min_{q,l}\left\{\left\|\bar{\boldsymbol{Y}} - \bar{\boldsymbol{H}}\boldsymbol{\chi}\boldsymbol{K}_{l,q}\right\|^2\right\} \tag{10}$$

where $\boldsymbol{K}_{l,q}$ represents the equivalent transmit symbol vector $\boldsymbol{K}$ defined in (8) when the $\mathcal{L}$-PSK/QAM symbol s_l is placed in the q -th position of the vector. An exhaustive search over the entire space of $(Q\cdot\mathcal{L})$ is essential for this optimal detector.

2) The Detectors of [11]: Two detectors were proposed in [11] for reduced-complexity detection of STSK schemes, which we refer to as Detector I and Detector II in this paper.

a) Detector I: Detector I normalizes each column $\bar{\boldsymbol{h}}_q$ of $\bar{\boldsymbol{H}}$ to generate the modified equivalent channels, $\boldsymbol{H}' = \left[\frac{\bar{\boldsymbol{h}}_1}{\| \bar{\boldsymbol{h}}_1 \|}, \ldots, \frac{\bar{\boldsymbol{h}}_Q}{\| \bar{\boldsymbol{h}}_Q \|}\right]$ and exhaustively searches through $\boldsymbol{Z} = [z_1, \ldots z_Q]^T = \boldsymbol{H}'\bar{\boldsymbol{Y}}$ to yield:

$$\hat{q} = \arg\max_{q,\forall l'} \Bigg[2 \| \bar{\boldsymbol{h}}_q \| \Big\{ | \Re(z_q) || \Re(s_{l'}) | + | \Im(z_q) || \Im(s_{l'}) | \Big\} - \| \bar{\boldsymbol{h}}_q \|^2 | s_{l'} |^2 \Bigg], \quad (11)$$

where $s_{l'}$ $(l' = 1, 2, \ldots, \mathcal{L}')$ represents the constellation points in the first quadrant only. Having estimated the DM index $\hat{q}$ using (11), the constellation symbol index is estimated from:

$$\hat{l} = \arg\min_{l} | z_{\hat{q}} - \| \bar{\boldsymbol{h}}_{\hat{q}} \| s_l | . \quad (12)$$

Thus the search space for (11) of this two-stage detector is $Q\mathcal{L}'$, which is one-fourth of the search space $Q\mathcal{L}$ for optimal ML detector. This detector, however, has an additional search space of $\mathcal{L}$ to calculate $\hat{l}$ from (12).

b) Detector II: Detector I was further modified in [11] for simplifying (11) as:

$$\hat{q} = \arg\max_{q,\forall v} \left[| \Re(z_q) | x_{v,I} + | \Im(z_q) | x_{v,Q} \right], \quad (13)$$

while (12) invoked for estimating $\hat{l}$ remains unchanged. Note that $x_{v,I}$ and $x_{v,Q}$ indicate the real and imaginary parts of the unit vectors having the same phase angles as those of the first-quadrant constellation points respectively.

The detectors mentioned above are designed for non-dispersive channels, where FDE is not essential. They impose a considerably reduced complexity, especially in case of a high channel coherence interval and higher order QAM constellation. However, in an OFDMA or SC-FDMA aided STSK, MIMO FDE has to be employed and the resulting data estimation does not explicitly involve the channel transfer matrix. Hence detection complexity quantified in terms of the number of multiplications required is considerably reduced [10]. Against this background, we conceived a modified reduced-complexity detector, which estimates the DM index and contellation index separately after MIMO FDE and is particularly useful for STSK transmissions in dispersive channels.

III. The Proposed Detector

The schematic diagram of the OFDMA/SC-FDMA-aided STSK receiver employing the proposed detector is shown in Fig. 2. After the discrete Fourier transform (DFT) operation and subcarrier demapping at the receive AEs, the detector employs a MIMO MMSE FDE and then separately estimates the indices of the activated DM and of the constellation symbol. To be specific, the detector first invokes the MIMO MMSE FDE in an attempt to minimize the average squared error. The MIMO ZF detector suffers from a performance degradation due to the noise enhancement when a subcarrier is in deep fade and hence is not used in the proposed scheme.

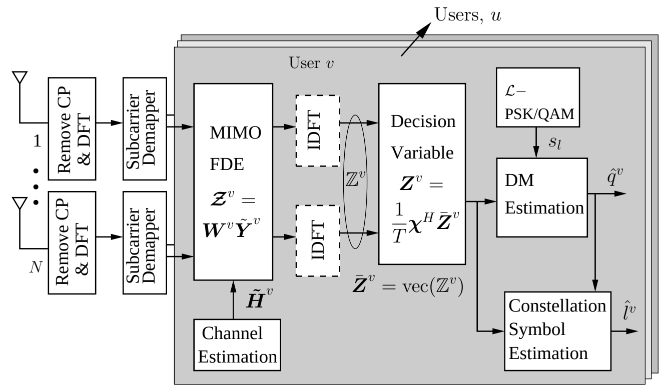

Fig. 2. The receiver architecture of the OFDMA/SC-FDMA-aided STSK employing the proposed detector. The dotted 'IDFT' block does not exist in the OFDMA-aided STSK receiver.

Let the ST codewords corresponding to user u $[u = 0, 1, \ldots, (U-1)]$ of the SC-FDMA-aided STSK in Fig. 1 be denoted by $\boldsymbol{X}^u$, and the transmit blocks obtained by DFT operation of the codeword symbols be represented by $\tilde{\boldsymbol{X}}^u$. For the OFDMA-aided STSK, the DFT-precoding block in Fig. 1 does not exist and we have, $\tilde{\boldsymbol{X}}^u = \boldsymbol{X}^u$.

The block-based FD output $\tilde{\boldsymbol{Y}}$ in Fig. 2 obtained after DFT operation at the receive antennas may be expressed as [10]:

$$\tilde{\boldsymbol{Y}} = \sum_{u=0}^{(U-1)} \tilde{\boldsymbol{H}}^u \tilde{\boldsymbol{X}}^u + \tilde{\boldsymbol{V}} \quad (14)$$

where $\tilde{\boldsymbol{H}}^u$ represents the FD MIMO channel transfer matrix corresponding to user u and $\tilde{\boldsymbol{V}}$ denotes the additive white Gaussian noise (AWGN) matrix. In order to estimate $\tilde{\boldsymbol{X}}^v$ corresponding to a particular user[1] v, the subcarriers are first demapped and multiply the signal $\tilde{\boldsymbol{Y}}^v$ intended for user v by the weight matrix $\boldsymbol{W}^v$, which minimizes $\mathcal{E}\left\{ \left[\boldsymbol{W}^v\tilde{\boldsymbol{Y}}^v - \tilde{\boldsymbol{X}}^v\right] \left[\boldsymbol{W}^v\boldsymbol{Y}^v - \tilde{\boldsymbol{X}}^v\right]^H \right\}$, yielding [10], [13]:

$$\boldsymbol{W}^v = \left[\left(\tilde{\boldsymbol{H}}^v\right)^H \tilde{\boldsymbol{H}}^v + \sigma_N^2 \boldsymbol{I}_M \right]^{-1} \left(\tilde{\boldsymbol{H}}^v\right)^H, \quad (15)$$

where σ_N^2 indicates the noise variance. The MMSE codeword estimate $\boldsymbol{\mathcal{Z}}^v$ of user v can then be expressed as [14, Ch. 5]

$$\boldsymbol{\mathcal{Z}}^v = \boldsymbol{W}^v\tilde{\boldsymbol{Y}}^v. \quad (16)$$

After MIMO FDE, the received signal is passed through the 'IDFT' block, which results in the space-time blocks $\mathbb{Z}^v$. In case of the OFDMA-aided STSK, however, the 'IDFT' block does not exist and we have, $\mathbb{Z}^v = \boldsymbol{\mathcal{Z}}^v$.

Under the idealized assumption of perfect orthogonality of the subcarriers and of perfect synchronization, the scheme is decontaminated from multiuser interference (MUI), albeit the MMSE-based scheme imposes self-interference (SI). Assuming the SI to be low, the decision variable corresponding to

[1]The generalized user has been represented by u, whereas v refers to the intended user.

user v can be expressed as

$$\mathbb{Z}^v = \boldsymbol{X}^v + \tilde{\boldsymbol{V}}^v. \tag{17}$$

Upon applying the vectorial stacking operation vec$(\cdot)$ to both sides of (16), we have

$$\bar{\boldsymbol{Z}}^v = \boldsymbol{\chi}\boldsymbol{K}^v + \bar{\boldsymbol{V}}^v, \tag{18}$$

where

$$\bar{\boldsymbol{Z}}^v = \text{vec}(\mathbb{Z}^v) \in \mathbb{C}^{MT\times 1}, \tag{19}$$

$$\bar{\boldsymbol{V}}^v = vec(\tilde{\boldsymbol{V}}^v) \in \mathbb{C}^{MT\times 1}, \tag{20}$$

and $\boldsymbol{\chi}$ and $\boldsymbol{K}^v$ are defined in (7) and (8), respectively.

The decision variable $\boldsymbol{Z}^v$ is now obtained by multiplying (18) by $\frac{1}{T}\boldsymbol{\chi}^H$ as:

$$\boldsymbol{Z}^v = \frac{1}{T}\boldsymbol{\chi}^H\bar{\boldsymbol{Z}}^v = \left[z_1^v, \dots z_Q^v\right]^T \in \mathbb{C}^{Q\times 1}. \tag{21}$$

We now perform an exhaustive search for the estimates $(\hat{q}^v, \hat{l}^v)$:

$$\begin{aligned}(\hat{q}^v, \hat{l}^v) &= \arg\min_{q,l} \parallel \boldsymbol{Z}^v - \boldsymbol{K}_{l,q} \parallel^2 \qquad (22)\\ &= \arg\min_{q,l}\left(\mid z_q^v - s_l \mid^2 + \sum_{q'\neq q} \mid z_{q'}^v \mid^2\right)\\ &= \arg\min_{q,l}\left(\mid z_q^v - s_l \mid^2 + \parallel \boldsymbol{Z}^v \parallel^2 - \mid z_q^v \mid^2\right)\\ &= \arg\max_{q,l}\left(\mid z_q^v \mid^2 - \mid z_q^v - s_l \mid^2\right)\\ &= \arg\max_{q,l}\Bigg[2\Big\{\mid \Re(z_q^v) \mid\mid \Re(s_l) \mid\\ &\qquad + \mid \Im(z_q^v) \mid\mid \Im(s_l) \mid\Big\}\\ &\qquad - \mid s_l \mid^2\Bigg]. \qquad (23)\end{aligned}$$

Assuming the $\mathcal{L}$-PSK/QAM constellation to be symmetric about the I- and Q-axis, we can rewrite (23) to estimate the index to activated dispersion matrix q as:

$$\begin{aligned}\hat{q}^v &= \arg\max_{q\ \forall l'}\Bigg[2\Big\{\pm \mid \Re(z_q^v) \mid\mid \Re(s_{l'}) \mid\\ &\qquad \pm \mid \Im(z_q^v) \mid\mid \Im(s_{l'}) \mid\Big\} - \mid s_{l'} \mid^2\Bigg]\\ &= \arg\max_{q\ \forall l'}\Bigg[2\Big\{\mid \Re(z_q^v) \mid\mid \Re(s_{l'}) \mid\\ &\qquad + \mid \Im(z_q^v) \mid\mid \Im(s_{l'}) \mid\Big\}\\ &\qquad - \mid s_{l'} \mid^2\Bigg], \qquad (24)\end{aligned}$$

where l' denotes the index of the constellation symbol $s_{l'}$ that lies in the first quadrant only [11].

The index of the constellation symbol used in forming the STSK codeword can be estimated from:

$$\hat{l}^v = \arg\min_{l} \mid z_{\hat{q}}^v - s_l \mid. \tag{25}$$

Since Eq. (24) and Eq. (25) do not explicitly involve the channel transfer function for calculating $\hat{q}$ and $\hat{l}$, the number of multiplications and additions involved is significantly lower than that imposed by the existing detectors.

The rationale of multiplying $\bar{\boldsymbol{Z}}^v$ by the term $\frac{1}{T}\boldsymbol{\chi}^H$ in (21) is outlined as follows. Substituting the value of $\bar{\boldsymbol{Z}}$ from (18) into (21), we have

$$\begin{aligned}\boldsymbol{Z}^v &= \frac{1}{T}\boldsymbol{\chi}^H\bar{\boldsymbol{Z}}^v \qquad (26)\\ &= \frac{1}{T}\boldsymbol{\chi}^H\left[\boldsymbol{\chi}\boldsymbol{K}^v + \bar{\boldsymbol{V}}^v\right]\\ &= \frac{1}{T}\boldsymbol{\chi}^H(\boldsymbol{\chi})_{q^v}s_{l^v} + \bar{\boldsymbol{N}}_0, \qquad (27)\end{aligned}$$

where $(\boldsymbol{\chi})_{q^v}$ denotes the q^v-th column of the matrix '$\boldsymbol{\chi}$' and we have, $\bar{\boldsymbol{N}}_0 \in \mathbb{C}^{Q\times 1}$.

Now we have $\boldsymbol{\chi}^H = [\text{vec}(\boldsymbol{A}_1), \cdots, \text{vec}(\boldsymbol{A}_Q)]^H \in \mathbb{C}^{Q\times MT}$ and $(\boldsymbol{\chi})_{q^v} = [\text{vec}(\boldsymbol{A}_{q^v})] \in \mathbb{C}^{MT\times 1}$, and hence,

$$\begin{aligned}\boldsymbol{\chi}^H(\boldsymbol{\chi})_{q^v} &= [z_1', z_2', \cdots, \underset{q-th\ position}{\underset{\uparrow}{\text{tr}(\boldsymbol{A}_{q^v}^H\boldsymbol{A}_{q^v})}}, \cdots, z_Q']^T\\ &= [z_1', z_2', \cdots, \underset{q-th\ position}{\underset{\uparrow}{T}}, \cdots, z_Q']^T, \qquad (28)\end{aligned}$$

where $z_i' = \text{tr}(\boldsymbol{A}_i^H\boldsymbol{A}_{q^v})$, $i = 1, 2, \dots, Q$ with $i \neq q^v$ and $\text{tr}(\boldsymbol{A}_{q^v}^H\boldsymbol{A}_{q^v}) = T$ according to (2). Thus we obtain,

$$\boldsymbol{Z} = [z_1', z_2', \cdots, \underset{q^v-th\ position}{\underset{\uparrow}{s_{l^v}}}, \cdots, z_Q']^T + \bar{\boldsymbol{N}}_0. \tag{29}$$

Eq. (29) demonstrates that the ML search of (22) can be simplified to obtain the detection rules of (24) and (25).

We note that the optimal ML detector [1], Detector I and Detector II [11] were conceived mainly for non-dispersive MIMO channels. By contrast, the proposed detector is designed for realistic dispersive scenarios. Detector I and Detector II are capable of reducing the complexity imposed by the optimal ML detector. However, they cannot be directly employed in the SC-FDMA-aided STSK scheme, because the TD data estimation has to be preceded by equalization in FD. In case of OFDMA-aided STSK, however, the optimal ML detector, Detector I and Detector II can be employed using the FD channel transfer matrix. In this context, the OFDMA/SC-FDMA-aided STSK scheme and the corresponding ML detector following MIMO FDE were proposed under multiuser scenarios in [10].

We observe that all the STSK-related detectors impose low decoding complexity compared to LDCs, because a single DM is activated in any signalling interval. The coherent scheme requires channel state information (CSI) for detection of source information, but after MIMO equalization in FD, the search for the most probable codeword in our proposed scheme does not involve the channel matrix and hence the computational

Table I
COMPLEXITY OF THE DETECTION SCHEMES

Detection Scheme	Complexity
ML	$\frac{4MNTQ+4NTQ\mathcal{L}+2NTQ\mathcal{L}}{log_2(Q\cdot\mathcal{L})}$
Detector I	$\frac{4MNTQ+8NTQ+4Q\mathcal{L}'+4\mathcal{L}}{log_2(Q\cdot\mathcal{L})}$
Detector II	$\frac{4MNTQ+8NTQ+2VQ+4\mathcal{L}}{log_2(Q\cdot\mathcal{L})}$
Proposed detector	$\frac{4M^2N+8MN+4MTQ+2Q\mathcal{L}'+Q+2\mathcal{L}}{log_2(Q\cdot\mathcal{L})}$

complexity of the proposed scheme quantified in terms of the number of algebraic operations required becomes significantly reduced compared to the existing detectors. Furthermore, the separate detection of the index of the DM activated and of the constellation symbol considerably reduces the search space, thereby imposing lower complexity.

IV. COMPLEXITY ANALYSIS

We quantify the complexity of the proposed scheme in terms of the number of real-valued multiplications and summarize in Table I along with the complexity of the existing detectors.

The complexity associated with our multiple-antenna-based MMSE FDE is given by[2],

$$\text{comp}\left[\left[\left(\tilde{\boldsymbol{H}}^v\right)^H\tilde{\boldsymbol{H}}^v+\sigma_N^2\boldsymbol{I}_M\right]^{-1}\left(\tilde{\boldsymbol{H}}^v\right)^H\right] = 4MN+4M^2N. \quad (30)$$

The complexity imposed by computing $\boldsymbol{\mathcal{Z}}^v = \boldsymbol{W}^v\tilde{\boldsymbol{Y}}^v$ is written as

$$\text{comp}\left[\boldsymbol{\mathcal{Z}}^v = \boldsymbol{W}^v\tilde{\boldsymbol{Y}}^v\right] = 4MN. \quad (31)$$

Similarly, we have:

$$\text{comp}\left[\boldsymbol{Z}^v = \frac{1}{T}\boldsymbol{\chi}^H\bar{\boldsymbol{Z}}^v\right] = 4MTQ+Q. \quad (32)$$

$$\text{comp}\left[\arg\max_{q\ \forall l'}\left\{2\Big(\mid\Re(z_q^v)\mid\mid\Re(s_{l'})\mid+\mid\Im(z_q^v)\mid\mid\Im(s_{l'})\mid\Big) - \mid s'_{l'}\mid^2\right\}\right] = 2Q\mathcal{L}', \quad (33)$$

and

$$\text{comp}\left[\arg\min_{l}\mid z_{\hat{q}}^v - s_l\mid\right] = 2\mathcal{L}. \quad (34)$$

From the numerical expressions in Table I, we can see that our scheme exhibits a substantially reduced complexity.

[2]The complexity associated with the inversion of the $(M\times M)$ matrix at this stage is $\mathcal{O}\left(M^{2.807}\right)$, if Strassen algorithm is used. This has not been included in our complexity calculation, because we quantify complexity in terms of the number of real-valued multiplications. The proposed scheme, however, has a significantly reduced complexity, even if the complexity associated with this matrix inversion is also taken into account.

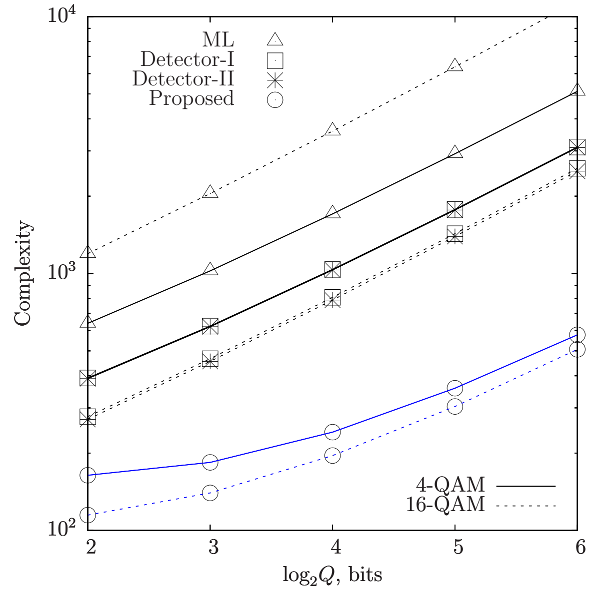

Fig. 3. Complexity of the detection algorithms considered as a function of the number of DMs. The parameters are: $M = 4, N = 4$, with $T = 4$.

V. PERFORMANCE COMPARISON

We have investigated the performance versus complexity of the proposed detector, which was compared to those of the existing detectors in Figs. 3, 4 and 5, upon varying the number of dispersion matrices Q and the number of symbol intervals T in conjunction with $M = 4,\ N = 4$ and employing both the 4-QAM and 16-QAM. We observe that Detector-I and Detector-II impose an identical complexity and both exhibit a lower complexity than the optimal ML detector on OFDMA STSK, albeit their complexity is not significantly reduced for lower-order constellations. On the other hand, our proposed detector exhibits a more significantly reduced complexity for all the scenarios, where we have included the complexity of the MIMO FDE stage in the overall complexity calculation of the proposed detector.

The bit-error ratio (BER) performance of our scheme was also investigated for OFDMA/SC-FDMA-aided single-user STSK (2,2,2,2) employing BPSK modulation for transmission over the COST 207-TU12 dispersive channel model employing 64 subcarriers. We note that Detector I and Detector II cannot be directly employed for the SC-FDMA-aided STSK scheme, because the scheme has to perform MIMO FDE and the source information has to be detected in the TD. In case of OFDMA, however, both the ML detector and the proposed detector can be employed, as mentioned in [10]. We observe that our proposed scheme exhibits a performance comparable to the detector proposed in [10], but at a substantially reduced complexity. While the ML detector for the OFDMA STSK scheme provides better performance, the proposed scheme is sufficiently reduced in complexity.

The performance of Detector II for transmission over narrowband channels is also shown in Fig. 5 for a fair comparison with other schemes. Observe that the proposed detector as

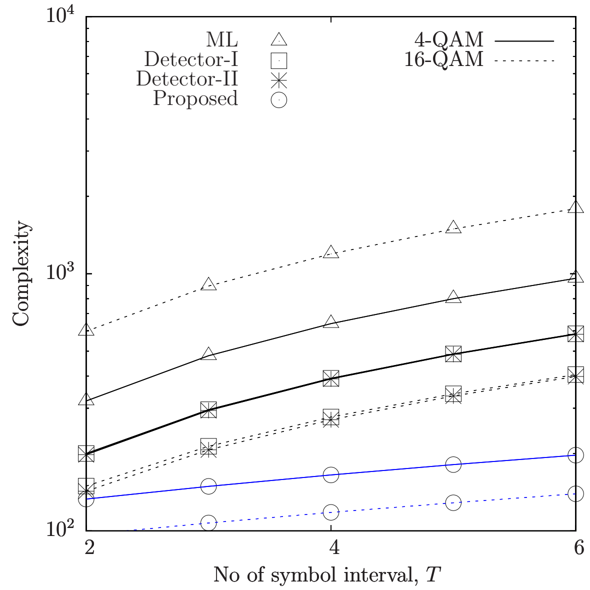

Fig. 4. Complexity of the detection algorithms considered as a function of the number of symbol intervals. The parameters are: $M = 4, N = 4$, with $Q = 4$.

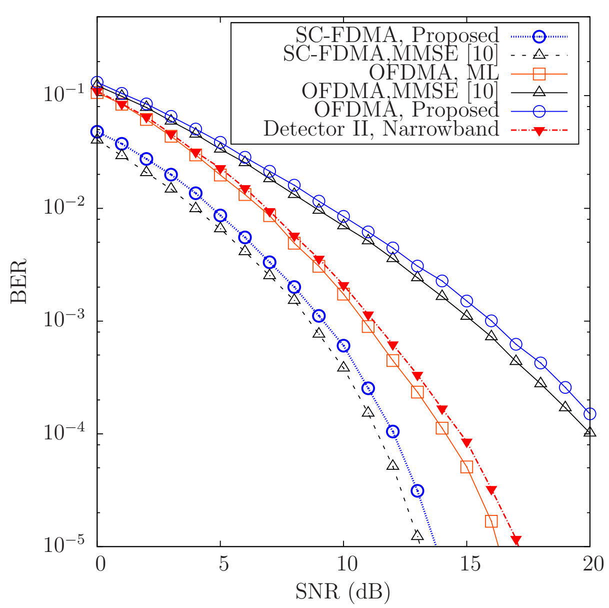

Fig. 5. BER performance of OFDMA/SC-FDMA aided single-user STSK (2,2,2,2) scheme employing BPSK modulation in COST 207-TU12 channel model and the different detectors. The performance of Detector II in narrowband STSK (2,2,2,2), BPSK is also shown as a fair comparioson.

well as the detector advocated in [10] transmitting even in dispersive channel outperforms Detector II in narrowband channel. This is because a beneficial frequency-domain diversity is provided by the DFT-based precoding in our SC-FDMA aided STSK scheme. Thus it can be inferred that the proposed detector reduces the complexity of the existing detector without sacrificing its performance, especially in the SC-FDMA aided STSK uplink.

VI. Conclusions

We proposed a new reduced-complexity detector for the OFDMA-aided STSK DL and the SC-FDMA-aided STSK UL. The proposed detector is inherently immune to inter-element interference (IEI) as a benefit of activating a single DM in the STSK schme and it has a reduced complexity compared to the existing detectors. It is a promising candidate for employment in OFDMA/SC-FDMA aided STSK transmissions over frequency-selective channels even under multiuser scenarios in terms of its performance versus complexity tradeoff.

The effectiveness of our STSK-based system depends mainly on the specific objective function (OF) used for the optimization of the DMs [2], [12]. We have employed an exhaustive search over 10^6 candidate DM sets for minimizing the maximum pairwise symbol error probability (PSEP) under the power constraint given by (2) for the optimization of the DMs. Further detailed discussion on the design of efficient DMs can be found in [15], [16].

References

[1] S. Sugiura, S. Chen, and L. Hanzo, "Coherent and differential space-time shift keying: A dispersion matrix approach," *IEEE Trans. Commun.*, vol. 58, no. 11, pp. 3219–3230, Nov. 2010.

[2] ——, "A universal space-time architecture for multiple-antenna aided systems," *IEEE Commun. Surveys Tuts.*, vol. 14, no. 2, pp. 401–420, 2012.

[3] ——, "Generalized space-time shift keying designed for flexible diversity-, multiplexing- and complexity-tradeoffs," *IEEE Trans. Wireless Commun.*, vol. 10, no. 4, pp. 1144 –1153, Apr. 2011.

[4] B. Hassibi and B. M. Hochwald, "High-rate codes that are linear in space and time," *IEEE Trans. Inf. Theory*, vol. 48, no. 7, pp. 1804–1824, Jul. 2002.

[5] J. Heath, R.W. and A. Paulraj, "Linear dispersion codes for MIMO systems based on frame theory," *IEEE Trans. Signal Process.*, vol. 50, no. 10, pp. 2429–2441, Oct. 2002.

[6] R. Mesleh, H. Haas, S. Sinanovic, C. W. Ahn, and S. Yun, "Spatial modulation," *IEEE Trans. Veh. Technol.*, vol. 57, no. 4, pp. 2228–2241, Jul. 2008.

[7] J. Jeganathan, A. Ghrayeb, L. Szczecinski, and A. Ceron, "Space shift keying modulation for MIMO channels," *IEEE Trans. Wireless Commun.*, vol. 8, no. 7, pp. 3692–3703, Jul. 2009.

[8] J. Jeganathan, A. Ghrayeb, and L. Szczecinski, "Spatial modulation: optimal detection and performance analysis," *IEEE Commun. Lett.*, vol. 12, no. 8, pp. 545–547, Aug. 2008.

[9] M. Driusso, F. Babich, M. I. Kadir, and L. Hanzo, "OFDM aided space-time shift keying for dispersive downlink channels," in *Proc. IEEE VTC Fall, 2012*, Sep. 2012, pp. 1–5.

[10] M. I. Kadir, S. Sugiura, J. Zhang, S. Chen, and L. Hanzo, "OFDMA/SC-FDMA aided space-time shift keying for dispersive multiuser scenarios," *IEEE Trans. Veh. Technol.*, vol. 62, no. 1, pp. 408–414, Jan. 2013.

[11] S. Sugiura, C. Xu, S. X. Ng, and L. Hanzo, "Reduced-complexity coherent versus non-coherent QAM-aided space-time shift keying," *IEEE Trans. Commun.*, vol. 59, no. 11, pp. 3090–3101, Nov. 2011.

[12] L. Hanzo, O. Alamri, M. El-Hajjar, and N. Wu, *Near-Capacity Multi-Functional MIMO Systems (Sphere-Packing, Iterative Detection and Cooperation)*. New York, NY: John Wiley & Sons, 2009.

[13] J. R. Barry, E. A. Lee, and D. G. Messerschmitt, *Digital communication (3rd ed.)*. Berlin, Germany: Springer-Verlag, 2003.

[14] L. L. Yang, *Multicarrier Communications*. Chichester, U.K.: John Wiley & Sons, 2009.

[15] M. Jiang and L. Hanzo, "Unitary linear dispersion code design and optimization for MIMO communication systems," *IEEE Signal Process. Lett.*, vol. 17, no. 5, pp. 497–500, May 2010.

[16] F. Babich, A. Crismani, M. Driusso, and L. Hanzo, "Design criteria and genetic algorithm aided optimization of three-stage-concatenated space-time shift keying systems," *IEEE Signal Process. Lett.*, vol. 19, no. 8, pp. 543–546, Aug. 2012.