Performance Evaluation of Low-Complexity Decision-Feedback Detector for SM-MIMO-OFDM Systems

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***Abstract*—Spatially multiplexed multiple–input multiple-output orthogonal frequency division multiplexing (SM-MIMO-OFDM) technology is envisaged to raise transmission capacity over the wireless channels up to 10bps/Hz, allowing for ubiquitous access to multimedia services in the 4G cellular networks. Future transceivers will incorporate MIMO antenna arrays of larger dimensions than those specified in the presently existing standards (with maximum 4 antennas). Thus, a crucial part of the cost-effective system design will deal with the low-complexity detector architecture, providing for optimal or suboptimal performance. In this paper, we propose such a detection technique and examine its error rate performance in the conditions of the dynamic channel model.**

***Index Terms*—decision feedback, minimum mean square error, QR decomposition, spatial multiplexing**

I. Introduction

Scantiness of the radio bandwidth is a major challenge for development of the broadband wireless communication systems capable to meet throughput requirements of the next generation access network. It is well-known that the use of multiple antenna technologies can enable high data rates suited for Internet and multimedia services, and increase transmission range and reliability at no loss of spectral efficiency. By employing multiple antennas, comprising the multiple-input multiple-output (MIMO) architecture, a set of spatial channels is created and can be used to facilitate additional data traffic subchannels, or to introduce coding redundancy in the transmissions without decreasing the throughput.

Spatial multiplexing (SM) MIMO technology can significantly increase bandwidth efficiency (up to the future 10bps/Hz). It multiplexes a number of independent data streams across spatial dimensions, i.e. transmit (Tx) antennas separated by location or space. At the receiver, multiple antennas and signal processing algorithms are used to separate SM streams.

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Recently, combinations of MIMO and multicarrier transmission techniques have attracted considerable research attention, due to the capability of multicarrier technology to resolve the frequency selectivity problem of the wireless channel. The emphasis has been placed on MIMO-OFDM [1] as a main candidate for the next generation wireless broadband interfaces, operating in the non-line-of-sight (NLOS), doubly-selective propagation environments, due to its relatively low implementation complexity. Benefits of SM-MIMO-OFDM solutions are essential, since the theoretical spectral efficiency is increased in proportion to the number of Tx antennas, each of which sends a unique stream of data symbols [2].

SM-MIMO detection algorithms are generally divided into two classes: linear and non-linear. The non-linear approaches typically offer better performance than their linear counterparts. The maximum likelihood detector (MLD) is known to be the optimal non-linear detector of the lattice-type signals [3]. However, its implementation, having computational effort of order $M^{N_{\text{tx}}}$, where M is the size of the modulation constellation and N_{tx} is the number of Tx antennas, is prohibitively complex for systems with many antennas and high-rate modulation schemes. Although the localised search algorithms, which perform suboptimal decoding on a selected sphere within the lattice, were proposed to reduce MLD complexity [4][5], their computational load still depends on the noise variance and is excessive for the lower signal-to-noise power ratios (SNRs). In this connection simpler linear and non-linear (typically, decision-feedback) suboptimal detection schemes have received attention in most system prototypes of higher spatial diversity orders.

Depending on the availability of the SNR information at the receiver, a zero-forcing (ZF) or a minimum mean square error (MMSE) criterion [6] can be adopted in the linear detector design. In the presence of the precise SNR information, the MMSE scheme is known to be the optimal linear detector.

The decision-feedback (DF) algorithms aim to improve the linear scheme output by capitalising on the finite-alphabet property of the modulated data symbols. Vertical Bell Labs layered architecture for space-time communication (V-BLAST) [7] is known to be the optimal DF scheme.

This paper addresses symbol error rate (SER) performance evaluation of the SM-MIMO-OFDM system with DF detection. To reduce computing complexity of the receiver, we present a suboptimal detector based on the sorted QR decomposition (SQRD). Computer simulations show that its performance is almost the same as that of the optimal V-BLAST detector, whereas the complexity is lower by an order of magnitude.

The rest of the paper is organised as follows. Section II describes a general concept of MIMO-OFDM. Section III is dedicated to the optimal linear and DF detector design, followed by the reduced-complexity algorithm description. Computational complexity of the algorithms is compared in Section IV. Section V includes several performance evaluation examples and is followed by conclusion in Section VI.

II. System and Channel Model

We consider a discrete-time baseband SM-MIMO-OFDM model that includes transmitter with N_{tx} antennas, receiver with N_{rx} ($N_{\mathrm{rx}} \geq N_{\mathrm{tx}}$) antennas and an equivalent discrete-time bandlimited channel model specified for each spatial layer (SL).

In the transmitter, a serial binary stream is divided into N_{tx} parallel streams, each of which passes through a PSK or QAM modulator creating a block of N complex-valued signal symbols $x_{n,m}$, $n \in [0, N-1]$ at its output, where *m* denotes the serial index of the block ($m \geq 0$). This block is then inverse Discrete Fourier transformed (IDFT), prepended with a cyclic prefix (CP) of a length N_{cp} to eliminate inter-block interference, converted to a serial sequence of samples and transmitted by the corresponding Tx antenna.

It is assumed that the channel remains constant during one processing block (a block without CP), so there is no loss of orthogonality between subcarriers. Reasonable system design conditions for this assumption are discussed in the work [8], pointing out that the maximum Doppler shift, characterising time variation of the channel impulse response (CIR), must not exceed 0.01 of the processing block rate.

To reflect the CIR process for the (i,j)th SL, where $i \in [0, N_{\mathrm{tx}}-1]$ and $j \in [0, N_{\mathrm{rx}}-1]$ denote Tx and Rx antenna indices respectively, we adopt the bandlimited filter model with a strict response energy enclosure within the first L taps ($L \leq N_{\mathrm{cp}}+1$) with the tap-gains described by the vector $\underline{\mathbf{h}}_m(j,i) = [\underline{h}_{l,m}(j,i)]_{L\times 1}$, $l \in [0, L-1]$. $\underline{\mathbf{h}}_m(j,i)$ is linked to the commonly used block-wise quasi-static approximation of the wide-sense stationary uncorrelated scattering (WSSUS) *K*-path response model [9]

$$g(t,j,i) = \sum_{m=0}^{\infty}\sum_{k=0}^{K-1} a_{k,m}(j,i)\,\delta[t - mT - \tau_k(j,i)] \tag{1}$$

according to the formula

$$\underline{\mathbf{h}}_m(j,i) = \mathbf{\Sigma}(j,i)\,\mathbf{a}_m(j,i), \tag{2}$$

where *k* is the path index, $\tau_k(j,i)$ is the path delay, T is the block duration (including CP), path gains $\mathbf{a}_m(j,i) = [a_{k,m}(j,i)]_{K\times 1}$, $k \in [0, K-1]$ represent zero-mean complex Gaussian random variables (CGRVs) produced by lowpass-filtered independent stochastic processes, and the elements of the band-limiting matrix $\mathbf{\Sigma}(j,i)$ are

$$[\mathbf{\Sigma}(j,i)]_{l,k} = \frac{\sin \pi [l - B\tau_k(j,i)]}{\pi [l - B\tau_k(j,i)]}, \tag{3}$$

where $B = (N + N_{\mathrm{cp}})/T$ is the effective bandwidth.

With regard to the MIMO model features, it is assumed that channels at different SLs are homogeneous [9], implying non-correlatedness of their responses, i.e. $\mathrm{E}[\underline{\mathbf{h}}_m(j_1,i_1)\,\underline{\mathbf{h}}_m(j_2,i_2)^{\mathrm{H}}] = \mathbf{0}$ for $\forall i_1, i_2 \in [0, N_{\mathrm{tx}}-1]$, $\forall j_1, j_2 \in [0, N_{\mathrm{rx}}-1]$, where $i_1 \neq i_2$, $j_1 \neq j_2$. This property can be satisfied by setting proper spacing of the antenna elements in the transmitter and receiver arrays, and array elevation ensuring NLOS operation. Furthermore, we let the intrablock CIR correlation matrix $\mathbf{R}_{\underline{\mathbf{hh}}}(j,i) = \mathrm{E}\left[\underline{\mathbf{h}}_m(j,i)\,\underline{\mathbf{h}}_m(j,i)^{\mathrm{H}}\right]$ be identical for all SLs (denoted simply as $\mathbf{R}_{\underline{\mathbf{hh}}}$). The latter assumption holds reasonably well if antenna spacing inside the array is small in comparison with propagation distance, so that the multipath environment can be considered specular, with fixed scattering delays [9], i.e. τ_k, $\mathbf{\Sigma}$ in (1) and (2) are SL-independent.

At the receiver side, the block at the output of each Rx antenna is separated from the CP and DFT-transformed after that. The resultant N_{rx} sets of symbols are parallel-forwarded for processing in the channel estimator and MIMO detector. Since the OFDM system can be interpreted as N parallel narrowband channels, which do not interfere with each other in frequency, but are subject to SM, the received symbol on the *n*th subcarrier at the *j*th antenna within the *m*th block is described as [10]

$$y_{n,m}(j) = \sum_{i=0}^{N_{\mathrm{tx}}-1} x_{n,m}(i)\,h_{n,m}(j,i) + w_{n,m}(j), \tag{4}$$

where $h_{n,m}(j,i)$ is the channel frequency response (CFR) gain, corresponding to the (i,j)th SL, and $w_{n,m}(j)$ is the white Gaussian noise (WGN) sample, which represents a realisation of the CGRV with $\mathrm{E}[w_{n,m}(j)] = 0$ and $\mathrm{E}[|w_{n,m}(j)|^2] = \sigma_{\mathrm{w}}^2$ over all j, n, m.

III. SM-MIMO Detectors

At the receiver, the detector's objective is to restore sent data symbols based on the received symbol observations (4). To accomplish this mission, the channel response information has to be available. Channel response acquisition is performed by the channel estimator. In this work, perfect CSI knowledge is assumed at the receiver, though channel estimation errors generally should be taken into account as they lead to remarkable performance degradation [11].

A. Linear MMSE Detector

Let the linear transformation (4) be written in the matrix notation as

$$\mathbf{y}_{n,m} = \mathbf{H}_{n,m}\mathbf{x}_{n,m} + \mathbf{w}_{n,m}, \tag{5}$$

where $\mathbf{x}_{n,m} = [x_{n,m}(i)]_{N_{\mathrm{tx}}\times 1}$, $i \in [0, N_{\mathrm{tx}}-1]$, $n \in [0, N-1]$, $m \geq 0$ is the vector of data symbols at the input of N_{tx} Tx antennas on the *n*th subcarrier inside the *m*th block; $\mathbf{y}_{n,m} = [y_{n,m}(j)]_{N_{\mathrm{rx}}\times 1}$, $j \in [0, N_{\mathrm{rx}}-1]$ is the vector of symbols at the output of N_{rx} Rx antennas; $\mathbf{w}_{n,m} = [w_{n,m}(j)]_{N_{\mathrm{rx}}\times 1}$ is the vector of WGN samples affecting received signal; and $\mathbf{H}_{n,m} = [h_{n,m}(j,i)]_{N_{\mathrm{rx}}\times N_{\mathrm{tx}}}$, $i \in [0, N_{\mathrm{tx}}-1]$, $j \in [0, N_{\mathrm{rx}}-1]$ is the MIMO channel transform matrix.

Herein the data symbols $x_{n,m}(i)$, transmitted by different antennas, are assumed to be random, statistically independent, and to have the average power σ_{d}^2, i.e. $\mathrm{E}[\mathbf{x}_{n,m}\mathbf{x}_{n,m}^{\mathrm{H}}] = \sigma_{\mathrm{d}}^2\mathbf{I}_{N_{\mathrm{tx}}\times N_{\mathrm{tx}}}$. The noise samples are characterised as independent identically distributed (IID) CGRVs with $\mathrm{E}[\mathbf{w}_{n,m}\mathbf{w}_{n,m}^{\mathrm{H}}] = \sigma_{\mathrm{w}}^2\mathbf{I}_{N_{\mathrm{rx}}\times N_{\mathrm{rx}}}$.

If the noise variance σ_{w}^2 is known at the receiver (for example, it could be a function of the front-end amplifier gain), the optimal linear detector is obtained using the MMSE criterion to produce the transmitted symbol estimates [6]:

$$\hat{\mathbf{x}}_{n,m} = (\mathbf{H}_{n,m}^{\mathrm{H}}\mathbf{H}_{n,m} + \sigma_{\mathrm{w}}^2/\sigma_{\mathrm{d}}^2\,\mathbf{I})^{-1}\mathbf{H}_{n,m}^{\mathrm{H}}\mathbf{y}_{n,m}, \tag{6}$$

which are subsequently demodulated (sliced on the constellation), restoring sent binary data.

Note that the inverse in (6) always exists for $\sigma_{\mathrm{w}}^2 \neq 0$, due to the Hermitean matrix $\mathbf{H}_{n,m}^{\mathrm{H}}\mathbf{H}_{n,m} + \sigma_{\mathrm{w}}^2/\sigma_{\mathrm{d}}^2\,\mathbf{I}$ being positive definite. Hence the computing algorithm is always stable. It is also evident from (6) that in the absence of noise ($\sigma_{\mathrm{w}}^2 = 0$), the transmitted symbols can be detected only if $\mathbf{H}_{n,m}$ has full column rank, i.e. $N_{\mathrm{rx}} \geq N_{\mathrm{tx}}$. It is however an interesting question how extra Rx diversity ($N_{\mathrm{rx}} > N_{\mathrm{tx}}$) can improve detection performance.

B. V-BLAST MMSE Detector

In the linear detector, the symbol estimates stacked in the vector $\hat{\mathbf{x}}_{n,m}$ (6) undergo demodulation independently from each other. It is possible to couple the linear solution with the decision feedback from the demodulator to improve detection accuracy.

The V-BLAST DF scheme [7] formulates detection as a procedure of successive cancellation of the spatial interference, starting from the most reliable decision, i.e. the symbol with the largest average SNR. Detection of this first symbol is the same as in the linear method by slicing the corresponding entry of $\hat{\mathbf{x}}_{n,m}$ (6) on the constellation. After that the demodulated symbol's contribution is cancelled from the received signal $\mathbf{y}_{n,m}$ that is accompanied by zeroing the corresponding column of the transform matrix $\left[\mathbf{H}_{n,m}^{\mathrm{T}} \quad \sigma_{\mathrm{w}}/\sigma_{\mathrm{d}}\,\mathbf{I}_{N_{\mathrm{tx}}\times N_{\mathrm{tx}}}\right]^{\mathrm{T}}$ (refer to [11] for details of the transform matrix derivation). The following step determines which symbol is to undergo the linear processing next, using the updated transform matrix and received vector, i.e. ordering is established. For optimal performance this ordering has to be updated after each recursion. The described process is repeated until the last symbol is demodulated (Table 1).

TABLE 1. V-BLAST MMSE DETECTION ALGORITHM

Set $\tilde{\mathbf{y}} = \left[\mathbf{y}_{n,m}^{\mathrm{T}} \quad \mathbf{0}_{N_{\mathrm{tx}}\times 1}^{\mathrm{T}}\right]^{\mathrm{T}}$, $\tilde{\mathbf{H}} = \left[\mathbf{H}_{n,m}^{\mathrm{T}} \quad \sigma_{\mathrm{w}}/\sigma_{\mathrm{d}}\,\mathbf{I}_{N_{\mathrm{tx}}\times N_{\mathrm{tx}}}\right]^{\mathrm{T}}$	
For $i = 0$ to $N_{\mathrm{tx}}-1$ do	
(a)	$\mathbf{G} := \tilde{\mathbf{H}}^{+\mathrm{T}}$
(b)	$k_i := \underset{j\in[0,\,N_{\mathrm{tx}}-1],\ j\neq k_l,\ l\in[0,\,i)}{\arg\min} \left\lVert (\mathbf{G})_j \right\rVert^2$
(c)	$\tilde{x}(k_i) := \mathrm{demod}[(\mathbf{G})_{k_i}^{\mathrm{T}}\,\tilde{\mathbf{y}}]$
(d)	$\tilde{\mathbf{y}} := \tilde{\mathbf{y}} - (\tilde{\mathbf{H}})_{k_i}\,\tilde{x}(k_i)$
(e)	$(\tilde{\mathbf{H}})_{k_i} := \mathbf{0}$
Return $\tilde{\mathbf{x}}_{n,m} = [\tilde{x}(0) \quad \cdots \quad \tilde{x}(N_{\mathrm{tx}}-1)]^{\mathrm{T}}$	

It should be noted that although the ordering update following each decision represents an optimal procedure, the original criterion to establish priorities of the detected symbols, namely average post-detection noise power [7], is suboptimal and can be replaced by a more accurate metric as proposed by Kim [12]. It has been shown that the log-likelihood ratio (LLR), exploiting both average and instantaneous noise power, yields better ordering that significantly improves performance for the case of BPSK modulation. However, for the more common QPSK and QAM schemes performance gain is minor.

C. SQRD MMSE Detector

The recursive computation of the pseudoinverse in the V-BLAST algorithm (Table 1) represents the main complexity contribution. If the decoding order is known a priori, then V-BLAST implementation can be considerably simplified by performing the row permutation in $\tilde{\mathbf{y}}$ and the identical column permutation in $\tilde{\mathbf{H}}$ and applying the QR decomposition (QRD) to the resultant matrix [13]. The new form enables low-complexity successive detection without the pseudoinverse computation, starting from the symbol with the highest SNR (to be the last in the permuted vector $\tilde{\mathbf{y}}$) up to the symbol with the smallest SNR (to be the first in the permuted $\tilde{\mathbf{y}}$). Gram-Schmidt orthogonalisation procedure is known as an efficient means to calculate the QR-decomposed form of a matrix. The reduced-complexity version of the DF detector is presented in Table 2.

TABLE 2. REDUCED-COMPLEXITY DECISION-FEEDBACK MMSE DETECTION THROUGH SORTED QR DECOMPOSITION (SQRD)

Set $\tilde{\mathbf{H}} = \left[\mathbf{H}_{n,m}^{\mathrm{T}} \quad \sigma_{\mathrm{w}}/\sigma_{\mathrm{d}}\, \mathbf{I}_{N_{\mathrm{tx}}\times N_{\mathrm{tx}}}\right]^{\mathrm{T}}$, $\mathbf{Q} = \mathbf{0}_{(N_{\mathrm{rx}}+N_{\mathrm{tx}})\times N_{\mathrm{tx}}}$ Set k_i, $i \in [0, N_{\mathrm{tx}}-1]$ For $i = N_{\mathrm{tx}}-1$ down to 0 do (a) $(\mathbf{P})_{k_i} := \mathbf{Q}^{\mathrm{H}}(\tilde{\mathbf{H}})_{k_i}$ (b) $\mathbf{q} := (\tilde{\mathbf{H}})_{k_i} - \mathbf{Q}(\mathbf{P})_{k_i}$ (c) $[\mathbf{P}]_{k_i,k_i} := \lVert \mathbf{q} \rVert$ (d) $(\mathbf{Q})_{k_i} := [\mathbf{P}]_{k_i,k_i}^{-1}\, \mathbf{q}$ Return
Set $\tilde{\mathbf{y}} = \mathbf{Q}^{\mathrm{H}}\mathbf{y}_{n,m}$ ($\tilde{\mathbf{y}} = \mathbf{Q}^{\mathrm{H}}\overline{\mathbf{y}}_{n,m}$) For $i = 0$ to $N_{\mathrm{tx}}-1$ do (e) $\tilde{x}(k_i) := \mathrm{demod}\left[[\mathbf{P}]_{k_i,k_i}^{-1}\,[\tilde{\mathbf{y}}]_{k_i}\right]$ (f) $\tilde{\mathbf{y}} := \tilde{\mathbf{y}} - (\mathbf{P})_{k_i}\,\tilde{x}(k_i)$ Return $\tilde{\mathbf{x}}_{n,m} = [\tilde{x}(0) \quad \cdots \quad \tilde{x}(N_{\mathrm{tx}}-1)]^{\mathrm{T}}$

Note that QRD in Table 2 needs ordering information. Due to a slow change of the channel response between two successive blocks, the symbol ordering for the current block can be assumed to be the same as for the previous block. Thus, QRD (Table 2) is followed by calculation of the ordering information (Table 3), which is buffered to be used in the next block.

TABLE 3. SUBOPTIMAL ORDERING BASED ON SINGLE INVERSION OF QR-DECOMPOSED TRANSFORM MATRIX $\tilde{\mathbf{H}} = \mathbf{Q}\mathbf{P}$

Set $\mathbf{S} = \mathbf{P}^{-\mathrm{T}}$ For $i = 0$ to $N_{\mathrm{tx}}-1$ do $k_i := \underset{\substack{j\in[0,\, N_{\mathrm{tx}}-1] \\ j\neq k_l,\ l\in[0,\, i)}}{\arg\min} \left\lVert (\mathbf{S})_j \right\rVert^2$ Return

IV. COMPUTATIONAL COMPLEXITY

In this section, we assess the computational load required by each of the presented detection algorithms. The detection complexity is defined as the total amount of complex multiplications and additions per subcarrier per block. The corresponding expressions have been derived in [11] and are summarised in Table 4.

TABLE 4. COMPUTATIONAL COMPLEXITY OF VARIOUS DETECTORS BASED ON THE MMSE WEIGHTING

Det.	**Multiplicative complexity**	**Additive complexity**
Linear	$(N_{\mathrm{rx}}+N_{\mathrm{tx}})N_{\mathrm{tx}}^2 + N_{\mathrm{rx}}N_{\mathrm{tx}}$	$(N_{\mathrm{rx}}+N_{\mathrm{tx}}-2)N_{\mathrm{tx}}^2$
V-BLAST	$\left(\frac{1}{3}N_{\mathrm{rx}}+\frac{7}{12}N_{\mathrm{tx}}+\frac{13}{6}\right)N_{\mathrm{tx}}^3$ $+\frac{1}{4}\left(6N_{\mathrm{rx}}+\frac{29}{3}\right)N_{\mathrm{tx}}^2$ $+\frac{1}{6}(19N_{\mathrm{rx}}-7)N_{\mathrm{tx}}$ $-(2N_{\mathrm{rx}}+1)$	$\left(\frac{1}{3}N_{\mathrm{rx}}+\frac{7}{12}N_{\mathrm{tx}}+\frac{5}{6}\right)N_{\mathrm{tx}}^3$ $+\left(N_{\mathrm{rx}}+\frac{17}{12}\right)N_{\mathrm{tx}}^2$ $+\frac{1}{6}(16N_{\mathrm{rx}}-23)N_{\mathrm{tx}}$ $-(2N_{\mathrm{rx}}-1)$
SQRD	$\left(N_{\mathrm{rx}}+\frac{7}{6}N_{\mathrm{tx}}+4\right)N_{\mathrm{tx}}^2$ $+\left(2N_{\mathrm{rx}}+s+\frac{5}{6}\right)N_{\mathrm{tx}}$	$\left(N_{\mathrm{rx}}+\frac{7}{6}N_{\mathrm{tx}}+2\right)N_{\mathrm{tx}}^2$ $+\left(N_{\mathrm{rx}}-\frac{8}{3}\right)N_{\mathrm{tx}}$
Remarks: * Pseudoinverse is computed by means of the Gauss-Jordan elimination procedure [15] ** s denotes number of multiplications in the square root operation		

One can see that the SQRD implementation has the same order of complexity, $O(N_{\mathrm{rx}}N_{\mathrm{tx}}^2 + N_{\mathrm{tx}}^3)$, as the linear detection scheme. This is due to the matrix inversion complexity reduction by means of QRD. At the same time the optimal V-BLAST detector needs an order of magnitude higher number of mathematical operations ($O(N_{\mathrm{rx}}N_{\mathrm{tx}}^3 + N_{\mathrm{tx}}^4)$).

It should be noted that further complexity reduction is possible for SM-MIMO-OFDM systems relying on the strong correlation between the adjacent subcarriers [14], so that detection ordering could be assumed the same for a subcarrier group. However, the reduced number of mathematical operations is still within the same order of magnitude, and performance degradation is observed in the channels with large root-mean-square delay spread.

V. NUMERICAL ANALYSIS

In order to find out which detector is the most efficient, it is necessary to evaluate SER performance of the presented SM-

MIMO-OFDM under the condition of the doubly-selective channel model. This is achieved by a range of Monte Carlo simulations.

A. System Configuration

Consider a discrete-time fully synchronised baseband SM-MIMO-OFDM system with the processing block length $N = 64$ (i.e. 64 subcarriers in the effective bandwidth B). CP length is set to $N_{cp} = 7$ to accommodate CIR with a modelled length of $L = 8$ samples.

The modelled multipath channel is block-wise time-variant (refer to Section II), with the interblock response variation according to the Jakes model [16] and the maximum Doppler shift $0.01B/N$ (i.e. the worst-case scenario from the time selectivity standpoint). The multipath model adopted for simulation has the following parameters: $K = 3$ equipowered IID components with the excess delays $\tau_0 = 2B^{-1}$, $\tau_1 = 3.7B^{-1}$ and $\tau_2 = 5.4B^{-1}$. The modelled MIMO channel properties are in full accordance with the assumptions made in Section II. In all experiments perfectly estimated CSI is assumed to be available at the receiver.

B. SER Performance

In the conventional SER performance evaluation scenario (Fig.1), systems with the linear and DF MMSE detectors and two modulation schemes (QPSK and 16QAM) are compared. Optimality of the V-BLAST DF detector is clearly visible. SER improvement by V-BLAST with respect to the linear detection scheme reaches 12.5dB for the QPSK (lower-order) modulation in the higher SNR regime. The suboptimal SQRD DF detector demonstrates performance, which closely approaches that of V-BLAST with only a marginal SER loss. It should be noted that the worst-case time selectivity of the channel has been adopted in the experiment; hence slower channel response variation in the practical systems would lead to V-BLAST and SQRD detectors being almost identical in the performance sense, while the SQRD complexity is an order of magnitude smaller (refer to Section IV).

Fig.1 also illustrates importance of the DF detection ordering. The optimal ordering (in the V-BLAST and SQRD algorithms), based on detected symbol SNR, guarantees much more significant SER gain than DF detection in an arbitrary symbol order (see the non-sorted QRD SER graph for comparison).

The second simulation experiment aims to assess the impact of the receive diversity order (i.e. the number of Rx antennas) on the system performance. Fig.2 shows a persistent SER improvement with N_{rx} increase for both the linear and DF detector types. However, the DF scheme benefits from extra Rx diversity in a slightly better way than its linear counterpart. One should also note that the extra Rx diversity effect on SER improvement is more pronounced in the higher SNR operational modes.

The square MIMO system configurations, for which $N_{tx} = N_{rx}$, are characterised by the minimum implementation complexity for the spectral efficiency level achieved. However, such systems are prone to received symbol recovery failures when two or more spatial channels appear linearly dependent, constraining the detector's degrees of freedom. Therefore the number of receive antennas is recommended to be bigger than minimum, to maximise likelihood of correct detection.

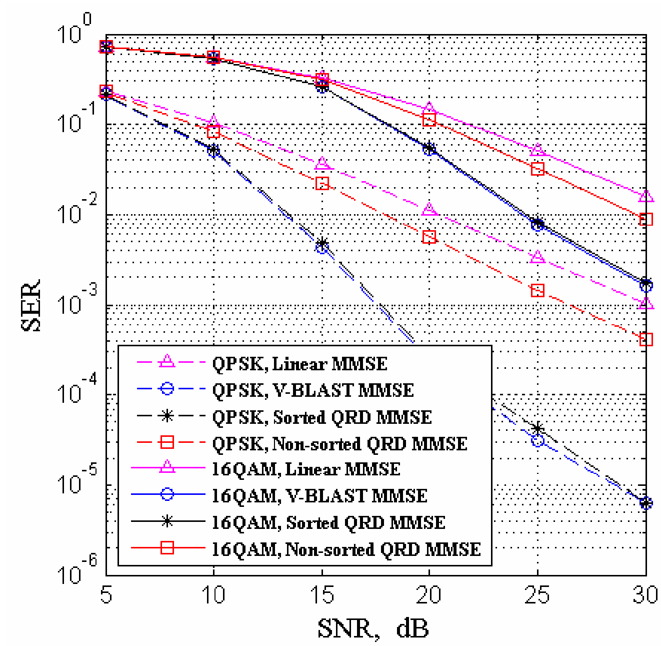

Fig.1. SER of various detectors in the 4x4 system

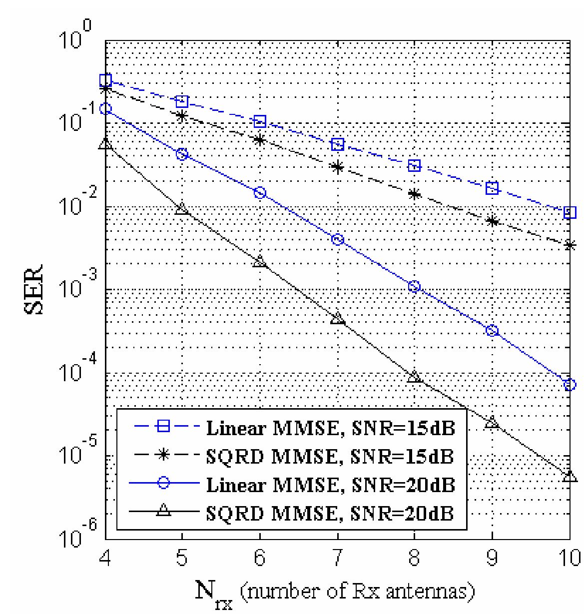

Fig.2. SER dependence on the receive diversity (16QAM-modulated system with 4 Tx antennas)

Fig.3 shows that the square SM-MIMO systems with larger antenna array dimensions exhibit increasingly worse SER than the single-input single-output configuration ($N_{tx} = N_{rx} = 1$) if the linear detector is employed. The proposed SQRD DF detector completely changes the trend. In the lower SNR modes (not shown in Fig.3), it maintains the same SER as the single-input single-output system. When the operational mode is characterised by a considerably higher SNR value, the SM-MIMO system with the SQRD detector surprisingly exhibits a better performance than the single-antenna configuration. Furthermore, SER gain tends to grow with the array dimensions until some asymptotic bound. For example, SER of the system with $N_{tx} = N_{rx} = 7$ is an order of magnitude smaller than SER of the system with $N_{tx} = N_{rx} = 1$ at $SNR = 30\,\text{dB}$.

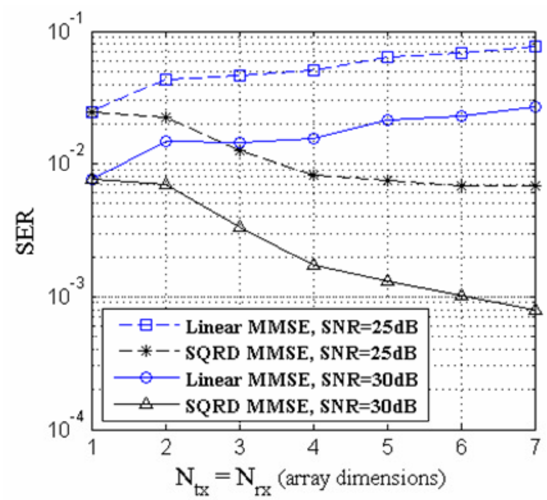

Fig.3. SER dependence on the MIMO array dimensions in the 16QAM-modulated square system

VI. Conclusions and Future Work

In this work, design and performance analysis of the MIMO-OFDM system with spatial multiplexing has been investigated. The emphasis in the discussion has been placed on the MIMO system configurations with higher antenna array dimensions, which will be employed in the next-generation broadband wireless communications systems, and for which the low complexity of detection algorithm is a crucial requirement. A good detector candidate, as proposed in this paper, is the SQRD DF scheme. It guarantees SER close to the optimal DF detector (V-BLAST) in the channels with the time variation constraint while its complexity is an order of magnitude smaller than that of V-BLAST.

With regard to the general detector design trends, a comparative analysis of the linear and DF solutions has revealed an increasing performance gain of the DF detector class observed for the higher SNR operational modes and higher Rx diversity orders. Another important remark is the linear detector performance deterioration with the increase of Tx diversity. This is in contrast to the DF detectors, which are capable of improving performance of the MIMO configurations with many antennas in the higher SNR regimes.

Being entirely dedicated to the best detector choice and Tx and Rx diversity order effects, this paper however leaves a number of important receiver design aspects beyond its scope. The first remark concerns the absence of the realistic CSI acquisition mechanism in the presented system model. The reader is recommended to refer to our more comprehensive work [11] for applicable methods.

Second, the optimal design of any MMSE detector requires exact SNR information at the receiver. In practice, this has to be estimated in the system operation process. An interesting question for further research is the impact of inaccurate SNR setting on the detector performance. To solve this and other related problems, it is possible to use analytical procedures of MSE characterisation of the system output [11].

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