Low Complexity Successive Interference Cancellation for Per-Antenna-Coded MIMO-OFDM Schemes by Applying Parallel-SQRD

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***Abstract*— By using orthogonal frequency division multiplexing (OFDM) in multiple-input multiple-output (MIMO) systems the equalization of frequency selective channels simplifies to a number of parallel MIMO detections. In order to gain from frequency diversity and of the correction capability of the code within a successive interference cancellation (SIC), forward error correction should be implemented for each antenna separately. In this paper we present a novel, computational efficient implementation for SIC in such a per-antenna-coded MIMO-OFDM system. It utilizes a parallelized version of the SQRD algorithm in order to achieve the same detection order for all subcarriers. In comparison to the most applied schemes from literature our approach requires only a fraction of computational complexity with almost the same performance.**

***Index Terms*— MIMO-OFDM, V-BLAST, Successive Interference Cancellation, SQRD, Wireless Communication.**

I. Introduction

In order to exploit the enormous capacity advantage of multiple antenna systems the well-known V-BLAST architecture is a very popular practical implementation [1]. In the past years different receiver implementations for this spatial multiplexing scheme have been proposed, where especially the successive interference cancellation (SIC) with optimized detection ordering achieves a good tradeoff with respect to complexity and performance [2]–[4].

In frequency selective environments the present intersymbol interference (ISI) leads not only to a spatial but also to a temporal superposition, resulting in a two-dimensional equalization problem. In order to implement efficient receivers the application of orthogonal frequency division multiplexing (OFDM) is a promising approach, as the two-dimensional problem is parallelized into N_C common spatial equalizations, with N_C denoting the number of used subcarriers. For the scheme investigated within this paper, forward error correction (FEC) codding is applied to each antenna separately to exploit frequency diversity and to make use of the error correction capability within the SIC. As this per-antenna-coding (PAC) requires the same detection order on each subcarrier an adopted version of the V-BLAST detection algorithm was proposed by van Zelst and Schenk, which requires the repeated calculation of the filter matrices for each carrier [5], [6]. Another approach defining the detection order based on capacity terms was given by Kadous [7]. Within this contribution we present a novel detection scheme with comparable performance but clearly less computational complexity. Therefore, the basic idea of our Sorted QR Decomposition (SQRD) is extended to the N_C parallel MIMO channels resulting in the same detection order on each subcarrier.

Outline of the Paper: The system model is introduced in Section II and in order to simplify the derivation of the SIC algorithms we recall linear equalization in Section III. The three different approaches for ordered SIC are presented in Section IV, where also the pseudo-code of the new Parallel-SQRD (P-SQRD) algorithm is given. The computational complexity and the performance results are investigated in Section V and VI, respectively. The paper is finished by concluding marks in Section VII.

Notation: Matrices are represented by bold capital letters, where the element in row α and column β of a matrix $\mathbf{A}$ is indicated by $[\mathbf{A}]_{\alpha,\beta} = a_{\alpha,\beta}$. Accordingly, vectors are denoted by small capital letters. The matrix transpose, hermitian transpose and Moore-Penrose pseudo-inverse are denoted by $(\cdot)^T$, $(\cdot)^H$ and $(\cdot)^+$, respectively. Furthermore, $\mathbf{I}_\alpha$ represents the $\alpha \times \alpha$ identity matrix and $\mathbf{0}_{\alpha,\beta}$ denotes the $\alpha \times \beta$ all zero matrix. In order to distinguish between variables in time and in frequency domain, we indicate variables in time domain (TD) by an index TD whereas a labeling for variables in frequency domain (FD) is generally omitted.

II. System description

We consider a multiple antenna system with N_T transmit and $N_\mathrm{R} \geq N_\mathrm{T}$ receive antennas applying spatial multiplexing in a frequency selective environment. The channel is assumed to be constant over each frame but changes independently between frames (*block fading channel*) and is perfectly known by the receiver. The transmitter of the per-antenna-coded MIMO-OFDM system is shown in Fig.1.

According to the block diagram the information data is demultiplexed in N_T parallel data streams (*layers*), encoded by a convolutional encoder and after bitwise interleaving mapped to M-QAM or M-PSK symbols $d_i(n)$, $1 \leq i \leq N_\mathrm{T}$, $1 \leq n \leq N_\mathrm{C}$. After transforming the symbols to time domain by using the inverse fast fourier transformation (IFFT) a guard interval (GI) of length N_G is added in form of a cyclic prefix before the sequence of $N_\mathrm{C}+N_\mathrm{G}$ signals $s_{\mathrm{TD},i}(k)$ is transmitted from

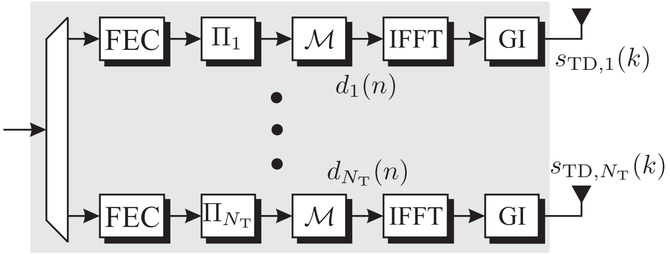

Fig. 1. MIMO-OFDM transmitter with per-antenna-coding

each antenna i. With $\mathbf{s}_{\mathrm{TD},i}(k) = [s_{\mathrm{TD},1}(k), \ldots, s_{\mathrm{TD},N_\mathrm{T}}(k)]^T$ denoting the N_T transmit signals at time instant k and $\mathbf{H}_\mathrm{TD}(\kappa)$, $0 \le \kappa \le N_\mathrm{H}$, representing the $N_\mathrm{R} \times N_\mathrm{T}$ channel matrix taps of the frequency selective channel of order N_H the $N_\mathrm{R} \times 1$ received vector is given by

$$\mathbf{x}_\mathrm{TD}(k) = \sum_{\kappa=0}^{N_\mathrm{H}} \mathbf{H}_\mathrm{TD}(\kappa)\, \mathbf{s}_\mathrm{TD}(k-\kappa) + \mathbf{n}_\mathrm{TD}(k)\,. \tag{1}$$

Here $\mathbf{n}_\mathrm{TD}(k)$ denotes the vector of additive white Gaussian noise at each receive antenna with covariance matrix $\mathrm{E}\{\mathbf{n}_\mathrm{TD}(k)\, \mathbf{n}_\mathrm{TD}^H(k)\} = \sigma_n^2 \mathbf{I}_{N_\mathrm{R}}$. The relation (1) expresses the superposition of transmitted symbols not only in space but also in time direction and thereby points out the two dimensional equalization problem.

At the receiver the cyclic prefix is removed and the fast fourier transformation (FFT) is used to perform the transformation back into frequency domain. As long as $N_\mathrm{G} \ge N_\mathrm{H}$ holds, the application of the cyclic prefix and discrete fourier transformation results in N_C orthogonal MIMO systems. With $\mathbf{d}(n)$ denoting the $N_\mathrm{T} \times 1$ vector of modulated symbols on carrier $1 \le n \le N_\mathrm{C}$ the corresponding received vector in frequency domain is given by [8]

$$\mathbf{y}(n) = \mathbf{H}(n)\, \mathbf{d}(n) + \mathbf{n}(n) \quad \text{for} \quad 1 \le n \le N_\mathrm{C} \tag{2}$$

with the flat MIMO channel for carrier n

$$\mathbf{H}(n) = \sum_{\kappa=0}^{N_\mathrm{H}} \mathbf{H}_\mathrm{TD}(\kappa)\, e^{-j\frac{2\pi}{N_\mathrm{C}}(n-1)\kappa}\,. \tag{3}$$

Due to this separation in N_C non-frequency selective parallel MIMO systems common detection algorithms like linear equalization or successive interference cancelation can be used for each carrier without any modification in case of an uncoded MIMO-OFDM scheme. For per-antenna-coded schemes this is only true for linear equalization, as described next.

III. Linear Equalization

For linear Zero-Forcing (ZF) equalization the received vector $\mathbf{y}(n)$ of each carrier is multiplied by the ZF filter matrix $\mathbf{G}_\mathrm{ZF}(n) = \mathbf{H}^+(n)$ yielding the output vector $\tilde{\mathbf{d}}_\mathrm{ZF}(n) = \mathbf{d}(n) + \mathbf{H}^+(n)\, \mathbf{n}(n)$ with error-covariance matrix

$$\mathbf{\Phi}_{\mathbf{ee},\mathrm{ZF}}(n) = \sigma_n^2 \left(\mathbf{H}^H(n)\, \mathbf{H}(n)\right)^{-1}\,. \tag{4}$$

Consequently, the signal-to-noise-ratio of layer i on carrier n is given by $\mathrm{SNR}_i(n) = 1/[\mathbf{\Phi}_{\mathbf{ee},\mathrm{ZF}}(n)]_{i,i}$. After calculating the Log-Likelihood-Ratios (*L-values*) for each layer by an adequate demodulation $\mathcal{D}$, these L-values are fed to the corresponding decoder and the per layer decoding takes place as shown in Fig. 2.

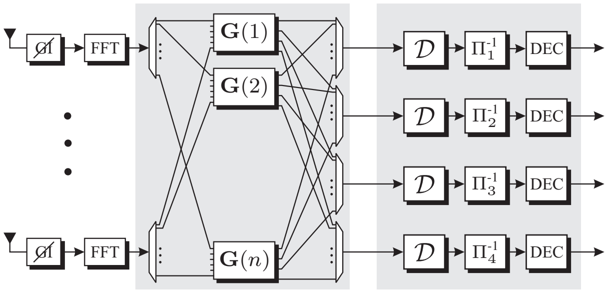

Fig. 2. Linear equalization for a MIMO-OFDM scheme with $N_\mathrm{T} = 4$ transmit antennas and per-antenna-coding

The corresponding MMSE filter matrix is given by $\mathbf{G}_\mathrm{MMSE}(n) = \left(\mathbf{H}^H(n)\, \mathbf{H}(n) + \sigma_n^2 \mathbf{I}_{N_\mathrm{T}}\right)^{-1} \mathbf{H}^H(n)$, which leads to the error-covariance matrix

$$\mathbf{\Phi}_{\mathbf{ee},\mathrm{MMSE}}(n) = \sigma_n^2 \left(\mathbf{H}^H(n)\, \mathbf{H}(n) + \sigma_n^2 \mathbf{I}_{N_\mathrm{T}}\right)^{-1} \tag{5}$$

and to the signal-to-interference-and-noise-ratio $\mathrm{SINR}_i(n) = 1/[\mathbf{\Phi}_{\mathbf{ee},\mathrm{MMSE}}(n)]_{i,i} - 1$. The occuring biase can be considered within the L-value calculation. By introducing the extended channel matrix $\underline{\mathbf{H}}(n)$ and the extended receive vector $\underline{\mathbf{y}}(n)$ for carrier n through [2]

$$\underline{\mathbf{H}}(n) = \begin{bmatrix} \mathbf{H}(n) \\ \sigma_n \mathbf{I}_{N_\mathrm{T}} \end{bmatrix} \quad \text{and} \quad \underline{\mathbf{y}}(n) = \begin{bmatrix} \mathbf{y}(n) \\ \mathbf{0}_{N_\mathrm{T},1} \end{bmatrix} \tag{6}$$

the filter output signal is also given by $\tilde{\mathbf{d}}_\mathrm{MMSE}(n) = \underline{\mathbf{H}}^+(n)\, \underline{\mathbf{y}}(n)$ and the error covariance matrix (5) becomes [4]

$$\mathbf{\Phi}_{\mathbf{ee},\mathrm{MMSE}}(n) = \sigma_n^2 \left(\underline{\mathbf{H}}^H(n)\, \underline{\mathbf{H}}(n)\right)^{-1}\,. \tag{7}$$

Thus, equalization with respect to the MMSE criterion equals ZF filtering with respect to the extended channel model (6). This correspondence is very helpful for implementing the SIC detection with respect to the MMSE criterion [4].

IV. Successive Interference Cancellation

Due to (2), the MIMO-OFDM scheme is separated into N_C orthogonal non-frequency selective MIMO systems and the well-known SIC approach can generally be applied on a per-tone basis. In order to reduce the problem of error propagation within this SIC the order of detection should be optimized [1]–[4] and furthermore we should make use of the error correction capability of FEC before removing the estimated interference. Therefore, on each carrier the layers have to be detected in the same order and consequently no separat optimization of this sequence is possible for each non-frequency selective MIMO system, but an optimization over all carriers has to be performed. In the sequel, we shortly introduce the corresponding approaches proposed by van Zelst and Schenk in [5], [6] and by Kadous in [7] to achieve such an optimized detection sequence for all carriers. Afterwards our new approach called P-SQRD is presented.

A. SINR-Optimization

In order to optimize the detection sequence van Zelst and Schenk proposed to run N_C parallel V-BLAST algorithms with an adopted ordering criterion. In the first detection step, the ZF or MMSE filter matrices $\mathbf{G}(n)$ and the corresponding error covariance matrices $\mathbf{\Phi}_{\mathbf{ee}}(n)$ are calculated for each carrier, whereby the i-th diagonal element $[\mathbf{\Phi}_{\mathbf{ee}}(n)]_{i,i}$ denotes the estimation error on the i-th layer of carrier n. Thus, in case of ZF filtering N_C pseudo-inverses have to be calculated. Afterwards, the diagonal elements of the error covariance matrices are summed up

$$\overline{\mathbf{\Phi}}_{\mathbf{ee},i} = \frac{1}{N_\mathrm{C}} \sum_{n=1}^{N_\mathrm{C}} [\mathbf{\Phi}_{\mathbf{ee}}(n)]_{i,i} \tag{8}$$

and the layer with the smallest overall error is selected as the target layer for each carrier. Subsequently, filtering and demodulation are performed on each carrier according to this layer and after parallel to seriell conversion the channel decoding is performed by Viterbi or BCJR including the calculation of the corresponding code bits [5], [6]. Than these estimated code bits are mapped to QAM/PSK symbols, the estimated interference is canceled out on each carrier and the columns of the target layer are set to zero in the N_C channel matrices $\mathbf{H}(n)$. The detection of the remaining layers takes place in the same way following the V-BLAST philosophy. Overall, this approach requires in case of ZF-filtering the calculation of $N_\mathrm{C}(N_\mathrm{T}-1)$ pseudo-inverses, requiring a considerable complexity.

B. CMOS-Optimization

Another approach to optimize the detection order has been proposed by Kadous in [7]. Within his CMOS (Capacity Mapping Ordering Scheme) algorithm the averaged capacity of layer $1 \leq i \leq N_\mathrm{T}$ after ZF filtering is calculated

$$C(i) = \frac{1}{N_\mathrm{C}} \sum_{n=1}^{N_\mathrm{C}} \log_2\left(1 + \mathrm{SNR}_i(n)\right) \tag{9}$$

and the layer with the maximum $C(i)$ is selected as the current layer of interest. The actual detection process corresponds to the SINR-optimization and for MMSE detection the corresponding SINR is used in the expression for the capacity (9).

C. P-SQRD Approach

As shown in several publications, successive interference cancelation for non-frequency selective multilayer systems can be restarted in terms of the QR decomposition of the channel matrix, where the order of detection is achieved by permuting its columns [2]–[4].

Adopting this idea to MIMO-OFDM, a QR decomposition $\mathbf{H}(n)\mathbf{P}(n) = \mathbf{Q}(n)\mathbf{R}(n)$ of each permuted channel matrix $\mathbf{H}(n)\mathbf{P}(n)$ with permutation matrix $\mathbf{P}(n)$ has to be calculated. For an uncoded system a factorization with optimized detection sequence is efficiently found by the Sorted QR Decomposition (SQRD) and the extension to the MMSE criterion is achieved by the factorization of the extended MMSE channel matrices $\underline{\mathbf{H}}(n)$. However, as the order of detection has to be same on each carrier for MIMO-OFDM systems with per-antenna-coding, we propose an extended version of our SQRD algorithm to find a global permutation matrix $\mathbf{P}$.

In the first step of this Parallel Sorted QR Decomposition (P-SQRD) the squared column norm of each layer over all carriers is calculated

$$\vartheta(i) = \sum_{j=1}^{N_\mathrm{R}} \sum_{n=1}^{N_\mathrm{C}} |h_{j,i}(n)|^2 = \sum_{j=1}^{N_\mathrm{R}} \sum_{\kappa=1}^{N_\mathrm{H}} |h_{\mathrm{TD},j,i}(\kappa)|^2 \tag{10}$$

and equals the squared norm of all fading coefficients $h_{\mathrm{TD},j,i}(\kappa)$ in time domain belonging to transmit antenna i. Following the philosophy of SQRD, the layer with *minimum norm* is determined and permuted to the first position on each carrier. Subsequently, each $\mathbf{H}(n)$ is orthogonalized with respect to the according column vector and the norm (10) is updated in order to denote only that part of each column vector orthogonal to the spanned orthonormal basis. In the second step again the layer with minimum norm is selected, the other columns are orthogonalized with respect to this layer and the norm is again updated. The decomposition of the remaining layer takes place in the same manner and consequently, we basically extended the N_C parallel QR decompositions by a global permutation $\mathbf{P}$ of all channel matrices $\mathbf{H}(n)$. As shown in the sequel, this leads to small overall computational complexity. As already mentioned, the MMSE solution is found by decomposition of the extended channel matrices $\underline{\mathbf{H}}(n)$.

Init: $\mathbf{R}_n = \mathbf{0}$, $\mathbf{Q}_n = [\mathbf{H}_n^T \ \sigma_n \mathbf{I}_{N_\mathrm{T}}]^T$ for $n = 1, \ldots, N_\mathrm{C}$
    $\mathbf{p} = [1\ 2\ \ldots\ N_\mathrm{T}]$
for $i = 1, \ldots, N_\mathrm{T}$
    $\vartheta(i) = \sum_{n=1}^{N_\mathrm{C}} \|\mathbf{Q}_n(1{:}N_\mathrm{R}, i)\|^2 + \sigma_n^2 N_\mathrm{C}$
end
for $i = 1, \ldots, N_\mathrm{T}$
    $\mu = \mathrm{argmin}_{\nu=i,\ldots,N_\mathrm{T}} \vartheta(\nu)$
    Exchange columns i and $i+\mu-1$ in $\mathbf{p}$ and $\boldsymbol{\vartheta}$
    for $n = 1, \ldots, N_\mathrm{C}$
        Exchange columns i and $i+\mu-1$ in $\mathbf{R}_n$ and in the first $N_\mathrm{R}+i-1$ rows of $\mathbf{Q}_n$
        $\mathbf{R}_n(i,i) := \|\mathbf{Q}_n(1{:}N_\mathrm{R}+i, i)\|$
        $\mathbf{Q}_n(1{:}N_\mathrm{R}+i, i) := \mathbf{Q}_n(1{:}N_\mathrm{R}+i, i) / \mathbf{R}_n(i,i)$
        for $\nu = i+1, \ldots, N_\mathrm{T}$
            $\mathbf{R}_n(i,\nu) := \mathbf{Q}_n^H(1{:}N_\mathrm{R}+i-1, i) \cdot \mathbf{Q}_n(1{:}N_\mathrm{R}+i-1, \nu)$
            $\mathbf{Q}_n(1{:}N_\mathrm{R}+i, \nu) := \mathbf{Q}_n(1{:}N_\mathrm{R}+i, \nu) - \mathbf{R}_n(i,\nu)\,\mathbf{Q}_n(1{:}N_\mathrm{R}+i, i)$
            $\vartheta(\nu) := \vartheta(\nu) - |\mathbf{R}_n(i,\nu)|^2$
        end
    end
end

Algorithm 1: P-SQRD-Algorithm for a system with N_T transmit and N_R receive antennas and N_C carrier (gray labeled entries are only required for the MMSE solution)

The pseudo-code[1] of the P-SQRD is given in Algorithm 1, where the gray labeled entries are only required for the MMSE implementation. In order to simplify the description we make us of the Matlab notation for indicating matrix elements, i.e. $\mathbf{A}(\alpha, \beta) = [\mathbf{A}]_{\alpha,\beta}$. To further avoid three-dimensional matrices, the n-th channel matrix $\mathbf{H}(n)$ is denoted by $\mathbf{H}_n$ and corresponding definitions are also used for other matrices.

When the P-SQRD calculation is done, each received vector $\mathbf{y}(n)$ is filtered by $\mathbf{Q}^H(n)$ and due to the upper triangular form of $\mathbf{R}(n)$ the N_T-th layer of each filter output signal

$$\tilde{\mathbf{d}}(n) = \mathbf{Q}^H(n)\,\mathbf{y}(n) = \mathbf{R}(n)\,\mathbf{d}(n) + \mathbf{n}(n) \tag{11}$$

is free of interference. After demodulation the L-values are deinterleaved and fed to the channel decoder. Using the interleaved code bits for remodulation the estimated interference is canceled out and the successive interference cancelation of the remaining layers is performed. The block diagram of the receiver is depicted in Fig. 3.

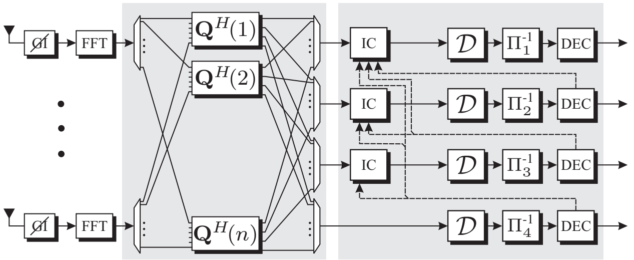

Fig. 3. Successive Interference Cancellation for a MIMO-OFDM scheme with $N_\mathrm{T} = 4$ transmit antennas

V. Computational Complexity

In this section we investigate the computational complexity of the SINR and the P-SQRD approach with respect to complex floating point operations $\mathcal{F}$. In order to achieve simple terms depending only on the system configuration, we count one complex addition as one flop (floating point operation) and a complex multiplication as three flops. Furthermore, the CMOS-approach is omitted as it requires an additional complexity in comparison to the SINR-ordering due to the calculation of the capacity term (9).

In case of ZF-filtering the SINR-approach requires the calculation of $N_\mathrm{T} - 1$ pseudo-inverses for each carrier n. Considering not only these filter calculations but also the detection process (without considering channel decoding) this approach requires approximately $\mathcal{O}\left((\frac{1}{6}N_\mathrm{T}^4 + 2N_\mathrm{R}N_\mathrm{T}^3 + \frac{1}{6}N_\mathrm{T}^3)N_\mathrm{C}\right)$ floating point operations. In contrast, the P-SQRD approach mainly consists of N_C QR decompositions. Using a detailed complexity consideration the overall complexity of this detection scheme is given by $\mathcal{O}\left((4N_\mathrm{R}N_\mathrm{T}^2 + \frac{1}{4}N_\mathrm{T}^2)N_\mathrm{C}\right)$ for the ZF-implementation. Thus, a strong reduction with respect to computational cost is achieved.

[1]The algorithm is given as an extension of the Modified Gram-Schmidt orthogonalization. However, similar expressions can also be achieved using Householder reflexion or Givens rotation for QR decomposition [9]. Furthermore, $\mathbf{p}$ denotes a permutation vector with $\mathbf{P} = \mathbf{I}_{N_\mathrm{T}}(:, \mathbf{p})$.

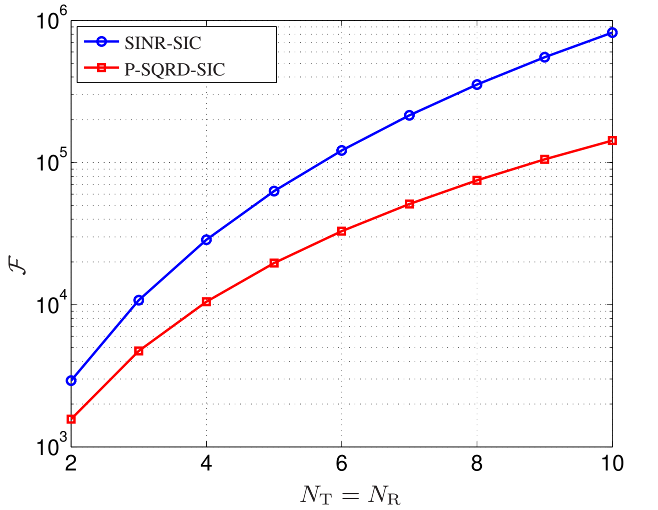

Fig. 4. Number of floating point operations $\mathcal{F}$ for SINR-SIC and P-SQRD-SIC detection with respect to the ZF criterion of a MIMO-OFDM system with $N_\mathrm{T} = N_\mathrm{R}$ antennas and $N_\mathrm{C} = 32$ subcarriers

For a varying, but equal number of transmit and receive antennas $N_\mathrm{T} = N_\mathrm{R}$ Fig. 4 shows the required number of Flops for SINR-SIC and P-SQRD-SIC in case of ZF implementation with $N_\mathrm{C} = 32$ carriers. This figure visualizes the strong decrease in computational complexity achieved by our new approach.

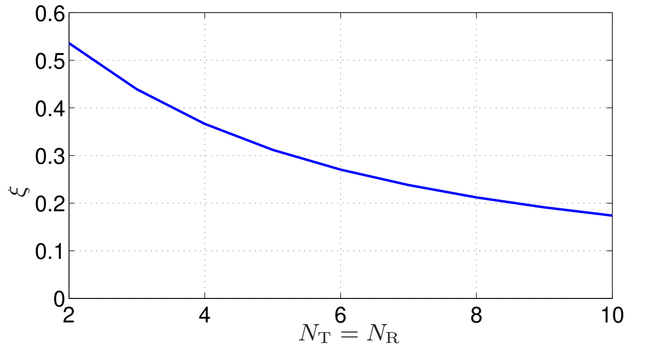

Fig. 5. Fraction of Flops $\xi = \mathcal{F}_\text{P-SQRD-SIC}/\mathcal{F}_\text{SINR-SIC}$ for a MIMO-OFDM system with $N_\mathrm{T} = N_\mathrm{R}$ antennas and $N_\mathrm{C} = 32$ subcarriers and Zero-Forcing criterion

In order to show the fraction of saved complexity, the quotient

$$\xi = \frac{\mathcal{F}_\text{P-SQRD-SIC}}{\mathcal{F}_\text{SINR-SIC}} \tag{12}$$

is depicted for a varying number of antennas $N_\mathrm{T} = N_\mathrm{R}$ in Fig. 5. It indicates an increasing computational advantage of the new scheme for increasing number of antennas. As an example, for a system with $N_\mathrm{T} = N_\mathrm{R} = 4$ antennas the P-SQRD-SIC requires approximately $0.36 \cdot \mathcal{F}_\text{SINR-SIC}$ flops and consequently leads to a strong reduction in computational cost.

VI. Performance Analysis

In this section we investigate the bit error rates (BER) for a per-antenna-coded MIMO-OFDM system with $N_\mathrm{T} =$

$N_{\mathrm{R}} = 4$ antennas. We assume uncorrelated SISO channels of order $N_{\mathrm{H}} = 5$ with a constant power delay profile, i.e. the variance of all fading coefficients is equal to $1/(N_{\mathrm{H}} + 1)$. Furthermore, each OFDM symbol contains a cyclic prefix of length $N_{\mathrm{G}} = 5$ and all $N_{\mathrm{C}} = 32$ subcarriers are used for signal transmission. For the simulations perfect estimation of the channel coefficients and of the noise variance is assumed.

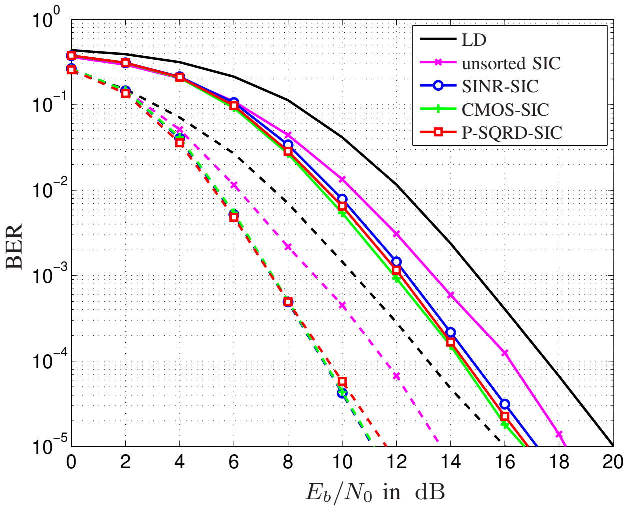

Fig. 6. BER of LD and SIC with respect to ZF (-) or MMSE (- -) criterion for a MIMO-OFDM system with $N_{\mathrm{T}} = N_{\mathrm{R}} = 4$ antennas, channel order $N_{\mathrm{H}} = 5$, $N_{\mathrm{C}} = 32$ subcarriers, guard interval of length $N_{\mathrm{G}} = 5$, 4-QAM symbols and $[7, 5]_8$ convolutional code

In Fig. 6 the bit error rates for linear and successive detection with respect to the ZF- and the MMSE-criterion are shown when the $[7, 5]_8$ convolutional code and 4-QAM modulation is applied on each substream. Obviously, the sorted SIC schemes achieve substantial performance improvements in comparison to linear and successive detection without ordering. The results for P-SQRD-SIC, SINR-SIC and CMOS-SIC are comparable, with minor advantages for the later ones in case of MMSE detection. However, this small performance impairment comes with a strong reduction in computational complexity.

In Fig. 7 the BERs for a system with 64-QAM modulation and the $[133, 171]_8$ convolutional code of constraint length 7 are shown. Again, only a small difference in performance can be observed between the different ordering criterions. This demonstrates the potential of the proposed P-SQRD approach for schemes with high spectral efficiencies.

VII. Summary and Conclusions

In this paper we proposed a new detection scheme for coded MIMO-OFDM systems by introducing an extended version of the SQRD algorithm. This new scheme achieves comparable results to the schemes from literature, however requiring only a fraction of computational complexity. Therefore, the pseudo-code of P-SQRD and an analysis of the computational complexity for both detection schemes was presented. In order to further simplify the detection for systems with a large number of carriers, the authors extended the basic philosophy

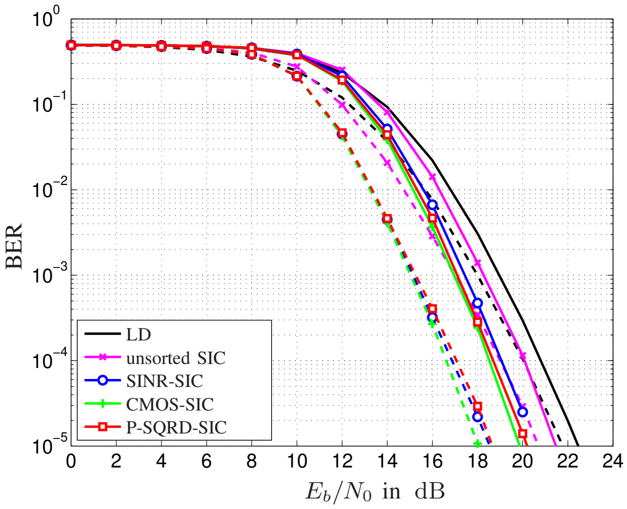

Fig. 7. BER of LD and SIC with respect ZF (-) or MMSE (- -) criterion of a MIMO-OFDM system with $N_{\mathrm{T}} = N_{\mathrm{R}} = 4$ antennas, channel order $N_{\mathrm{H}} = 5$, $N_{\mathrm{C}} = 32$ subcarriers, guard interval of length $N_{\mathrm{G}} = 5$, 64-QAM and $[133, 171]_8$ convolutional code

of the presented P-SQRD algorithm with tools of interpolation theory in [10]. Furthermore, it is also possible to apply the P-SQRD for system with frequency domain equalization, as described in [9]. Thus, a very efficient approach for future WLAN systems has been presented.

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