Successive Interference Cancellation for MIMO UW-OFDM

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***Abstract*—Unique word - orthogonal frequency division multiplexing (UW-OFDM) is a new signaling concept which has better spectral properties and also a superior bit error ratio (BER) performance over conventional cyclic prefix (CP) based OFDM. In UW-OFDM, the CPs are replaced by deterministic sequences, the so-called unique words (UWs). So far, the performance of UW-OFDM has been well investigated for single-input single-output (SISO) systems. Recently, we expanded our investigation of UW-OFDM to MIMO systems and proposed two linear detection approaches. In this work, we extend our work and propose successive interference cancellation (SIC) based non-linear detection schemes for MIMO UW-OFDM systems. We also compare the results to CP-OFDM with linear and SIC based detection schemes. The results show that UW-OFDM with SIC not only outperforms UW-OFDM with linear detection schemes but also CP-OFDM significantly.**

***Index Terms*—Orthogonal Frequency Division Multiplexing (OFDM), Cyclic Prefix (CP), Unique Word (UW), Multiple-Input Multiple-Output (MIMO)**

I. Introduction

Multicarrier transmission techniques are important transmission techniques for frequency selective environments. The most widely spread variant is the orthogonal frequency division multiplexing (OFDM) technique which is part of many wireless communication standards. In conventional OFDM, a cyclic prefix (CP) is inserted between successive OFDM symbols, so that within the receivers correlation interval, the linear convolution appears as a cyclic convolution. The CP is thrown away at the receiver in order to detect the transmitted data symbols. The randomness of the cyclic prefix implies that the cyclic prefix cannot be used efficiently for particular needs such as synchronization and/or system parameter estimation.

Many variants of OFDM have been introduced in the literature where the conventional CP in OFDM is replaced by a known data independent sequence while still maintaining the cyclicity. Data independent, deterministic guard interval sequences can be optimized for tasks like synchronization or channel estimation. A well discussed such scheme is known symbol padding (KSP) OFDM [1]. A new variant of OFDM has been introduced recently in [2]–[4] which is called as unique word (UW) OFDM. The UW-OFDM signaling concept is different from all other related concepts since the UW is part of the inverse discrete Fourier transform (IDFT) interval. This is achieved by adding redundancy in the frequency domain with the help of a code generator matrix. The design of the code generator matrix is not unique and different code generator matrices yield different performances. Optimum code generator matrices based on the non-systematic design always outperform the systematic design [5].

If the correlation introduced by the generator matrix is advantageously exploited at the receiver, this can result in superior bit error ratio (BER) performance of UW-OFDM over CP-OFDM [3]. This is achieved by using sophisticated detection procedures at the receivers but at the cost of higher complexity. These detection procedures have been well investigated for single-input single-output (SISO) systems. In [6], we extended our investigations of UW-OFDM to multiple-input multiple-output (MIMO) systems where we proposed two linear detection schemes, namely joint detection and a two step approach. Both approaches yield a similar BER performance but differ in the computational complexity. In [7], we introduced a two step linear detection procedure with a much lower computational complexity for orthogonal channels.

So far, we have focused only on linear detection schemes for MIMO UW-OFDM. But it is well known in the literature that linear estimators do not fully eliminate the interference which is caused by the multiple-antenna systems or by the multiple-users. Therefore, interference cancellation schemes are recommended to be used in rich scattering environments to exploit the capacity advantage of multiple antenna systems but at the cost of higher complexity [8]. The successive interference cancellation (SIC) method represents the most effective interference cancellation based reception technique in terms of BER performance and provides some system robustness. SIC detections schemes based on zero forcing (ZF) or minimum mean square error (MMSE) criteria are widely discussed schemes. These detection schemes require multiple calculations of pseudo-inverses which is a computationally expensive task. This complexity is reduced by using the QR decomposition of the channel and successive decoding of the symbols [9]. These SIC detection schemes have been well investigated for CP-OFDM [8]. In this work, we expand our investigations to SIC detection schemes for MIMO UW-OFDM systems. Specifically, we propose a hybrid detection scheme with a low complexity where a combination of linear detection and non-linear demodulation is applied.

The remainder of the paper is organized as follows. In Section II, we provide a brief overview of the UW-OFDM concept. In the subsequent section, we discuss the system model. In Section IV, we first give a brief overview of

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the linear detection schemes, then we discuss the different SIC schemes for UW-OFDM. We perform a computational complexity analysis of the linear and non-linear detection schemes in Section V. The simulation results are presented in Section VI. We conclude our work in Section VII.

Notation: We use lower-case bold face letters $(\boldsymbol{a}, \boldsymbol{b}, ...)$ to indicate vectors and upper-case bold face letters $(\boldsymbol{A}, \boldsymbol{B}, ...)$ to indicate matrices. The superscripts $(\cdot)^*$, $(\cdot)^{\mathrm{T}}$, and $(\cdot)^{\mathrm{H}}$ represent complex conjugate, matrix transpose, and complex conjugate transpose (Hermitian), respectively. The vec$\{\cdot\}$ operator stacks the columns of a matrix into a vector while the unvec$\{\cdot\}$ operator stands for the inverse operation of vec$\{\cdot\}$. The Kronecker product is represented by $\otimes$ and the tr$\{\cdot\}$ operator defines the trace of a matrix.

II. UW-OFDM Basics

Let $\boldsymbol{x}_u \in \mathbb{C}^{N_u \times 1}$ be a unique word of length N_u which is added at the tail of the time domain OFDM symbol such that

$$\boldsymbol{x}' = \begin{bmatrix} \boldsymbol{x}_d \\ \mathbf{0} \end{bmatrix} + \begin{bmatrix} \mathbf{0} \\ \boldsymbol{x}_u \end{bmatrix} = \begin{bmatrix} \boldsymbol{x}_d \\ \boldsymbol{x}_u \end{bmatrix} \in \mathbb{C}^{N \times 1} \tag{1}$$

where, in the first step, a UW-OFDM symbol $\boldsymbol{x} = \begin{bmatrix} \boldsymbol{x}_d^T & \mathbf{0} \end{bmatrix}^T$ is generated with a zero UW. Then in the second step the desired UW is added in the time domain [3]. A zero UW in the first step is achieved by multiplying the input data with a complex valued code generator matrix $\boldsymbol{G}$ before applying the inverse Fourier transform (IFFT). The design of the code generator matrix is not unique, but all of the $\boldsymbol{G}$ matrices shall fulfill the following constraint

$$\boldsymbol{F}_N^H \boldsymbol{B} \boldsymbol{G} = \begin{bmatrix} * \\ \mathbf{0} \end{bmatrix}, \tag{2}$$

where $\boldsymbol{F}_N$ is the N-point DFT matrix with its elements $[\boldsymbol{F}_N]_{k,l} = \frac{1}{\sqrt{N}} e^{-j\frac{2\pi}{N}kl}$ for $k, l = 0, 1, 2, \ldots, N-1$, $\boldsymbol{B} \in \{0,1\}^{N \times N_m}$ inserts the zero subcarriers, and $\boldsymbol{G} \in \mathbb{C}^{N_m \times N_d}$ is the complex valued code generator matrix which correlates N_d data symbols with $N_m = N_d + N_r$. Here N_r, which is typically chosen by $N_r = N_u$, accounts for the redundancy.

The design of the $\boldsymbol{G}$ plays an important role regarding the performance of UW-OFDM. Various approaches for the design of the $\boldsymbol{G}$ matrix have been suggested in [3], [5]. All of these $\boldsymbol{G}$ matrices yield different performances for UW-OFDM. But optimum $\boldsymbol{G}$ matrices based on the non-systematic approach always outperform the $\boldsymbol{G}$ matrices based on the systematic approach. In this work, we use a non-systematic code generator matrix that has been obtained by solving a minimization of the trace of the error covariance matrix of a linear MMSE (LMMSE) estimator for a fixed signal-to-noise ratio using steepest descent algorithms and by a random initialization [5].

III. System Model

Consider a MIMO system with M_T transmit antennas and M_R receive antennas where $\boldsymbol{S} \in \mathbb{C}^{N_d \times M_T}$ contains the $N_d \times M_T$ data symbols to be transmitted from M_T transmit antennas. The time domain signal before adding UW is given by

$$\mathbf{X} = \boldsymbol{F}_N^H \boldsymbol{B} \boldsymbol{G} \boldsymbol{S} \in \mathbb{C}^{N \times M_T}. \tag{3}$$

The data received by the ith receive antenna in the frequency domain after subtracting the UW and removing the zero carriers is given by

$$\begin{aligned} \boldsymbol{y}_i &= \sum_{j=1}^{M_T} \boldsymbol{B}^T \boldsymbol{F}_N \boldsymbol{H}_{i,j} \boldsymbol{F}_N^H \boldsymbol{B} \boldsymbol{G} \boldsymbol{s}_j + \boldsymbol{n}_i \\ &= \sum_{j=1}^{M_T} \hat{\boldsymbol{H}}_{i,j} \boldsymbol{G} \boldsymbol{s}_j + \boldsymbol{n}_i \quad \forall \quad i = 1, \ldots, M_R \end{aligned} \tag{4}$$

where $\boldsymbol{H}_{i,j}$ is the circulant channel convolutional matrix containing the channel impulse response and $\hat{\boldsymbol{H}}_{i,j} = \boldsymbol{B}^T \boldsymbol{F}_N \boldsymbol{H}_{i,j} \boldsymbol{F}_N^H \boldsymbol{B} \in \mathbb{C}^{N_m \times N_m}$ is a diagonal matrix containing the channel frequency response from the jth transmit antenna to the ith receive antenna on its main diagonal. The data received on all antennas can be written as

$$\begin{bmatrix} \boldsymbol{y}_1 \\ \vdots \\ \boldsymbol{y}_{M_R} \end{bmatrix} = \begin{bmatrix} \hat{\boldsymbol{H}}_{1,1} & \cdots & \hat{\boldsymbol{H}}_{1,M_T} \\ \vdots & \ddots & \vdots \\ \hat{\boldsymbol{H}}_{M_R,1} & \cdots & \hat{\boldsymbol{H}}_{M_R,M_T} \end{bmatrix} (\boldsymbol{I}_{M_T} \otimes \boldsymbol{G}) \begin{bmatrix} \boldsymbol{s}_1 \\ \vdots \\ \boldsymbol{s}_{M_T} \end{bmatrix} + \begin{bmatrix} \boldsymbol{n}_1 \\ \vdots \\ \boldsymbol{n}_{M_R} \end{bmatrix}. \tag{5}$$

We can rewrite the above equation as

$$\boldsymbol{y} = \boldsymbol{H}_{\text{eff}} \boldsymbol{G}_{\text{eff}} \tilde{\boldsymbol{s}} + \boldsymbol{n} \in \mathbb{C}^{(N_m \cdot M_R) \times 1}, \tag{6}$$

where $\boldsymbol{H}_{\text{eff}} \in \mathbb{C}^{(N_m \cdot M_R) \times (N_m \cdot M_T)}$ contains the overall channel frequency response as shown in Eq. (5), $\boldsymbol{G}_{\text{eff}} = \boldsymbol{I}_{M_T} \otimes \boldsymbol{G} \in \mathbb{C}^{(N_m \cdot M_T) \times (N_d \cdot M_T)}$ is the modified code generator matrix with a block diagonal structure, and $\tilde{\boldsymbol{s}} \in \mathbb{C}^{(N_d \cdot M_T) \times 1}$ contains the transmitted data symbols and is related to $\boldsymbol{S}$ in Eq. (3) via $\tilde{\boldsymbol{s}} = \text{vec}(\boldsymbol{S})$.

IV. Detection Schemes

A. Linear Detection Schemes

In [6], we applied the best linear unbiased estimator (BLUE) and the LMMSE estimator to estimate the data symbols. The equalizer weights for these estimators are calculated as

$$\boldsymbol{E}_{\text{BLUE}} = \left(\boldsymbol{G}_{\text{eff}}^H \boldsymbol{H}_{\text{eff}}^H \boldsymbol{H}_{\text{eff}} \boldsymbol{G}_{\text{eff}} \right)^{-1} \boldsymbol{G}_{\text{eff}}^H \boldsymbol{H}_{\text{eff}}^H, \tag{7}$$

$$\boldsymbol{E}_{\text{MMSE}} = \left(\boldsymbol{G}_{\text{eff}}^H \boldsymbol{H}_{\text{eff}}^H \boldsymbol{H}_{\text{eff}} \boldsymbol{G}_{\text{eff}} + \frac{\sigma_n^2}{\sigma_d^2} \boldsymbol{I}_{(N_d \cdot M_T)} \right)^{-1} \boldsymbol{G}_{\text{eff}}^H \boldsymbol{H}_{\text{eff}}^H, \tag{8}$$

with $\boldsymbol{E} \in \mathbb{C}^{(N_d \cdot M_T) \times (N_m \cdot M_R)}$, $\boldsymbol{I}_{(N_d \cdot M_T)}$ is the identity matrix of size $(N_d \cdot M_T)$, σ_n^2 is the variance of the noise samples, and σ_d^2 is the variance of the data symbols. Note that the computational complexity for these detection schemes increases significantly with an increased number of transmit antennas since a square matrix of size $N_d \cdot M_T$ has to be inverted. The estimated data symbols are then obtained from

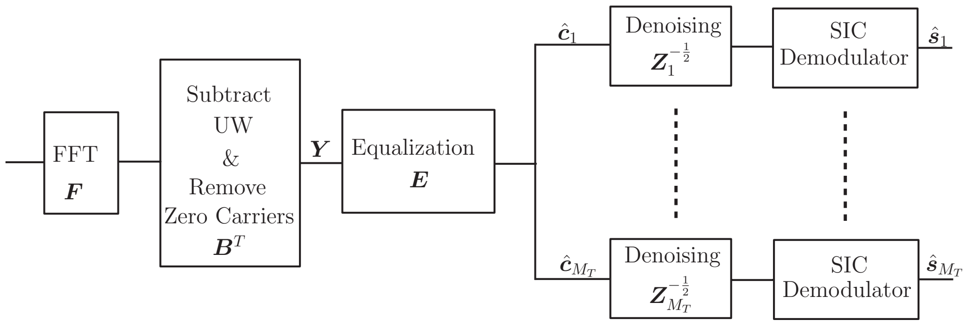

Fig. 1: Receiver architecture using hybrid detection

$$\hat{\boldsymbol{S}} = \mathrm{unvec}(\boldsymbol{E}\boldsymbol{y}) \in \mathbb{C}^{N_d \times M_T}. \tag{9}$$

B. ZF based successive interference cancellation (ZF-SIC) Detection

In [10], a ZF-SIC was presented for SISO UW-OFDM systems. We employ the same procedure for MIMO where we use the linear model in Eq. (6). According to this model,

$$\begin{aligned}\boldsymbol{y} &= \boldsymbol{M}\tilde{\boldsymbol{s}} + \boldsymbol{n}\\ &= \boldsymbol{m}_0 s_0 + \boldsymbol{m}_1 s_1 + \cdots + \boldsymbol{m}_{(N_d\cdot M_t)-1} s_{(N_d\cdot M_t)-1} + \boldsymbol{n},\end{aligned}$$

where $\boldsymbol{M} = \boldsymbol{H}_{\text{eff}}\boldsymbol{G}_{\text{eff}} \in \mathbb{C}^{(N_m\cdot M_R)\times(N_d\cdot M_T)}$ and $\boldsymbol{m}_k$ is the kth column of $\boldsymbol{M}$. The impact of the symbol decision is subtracted from the received signal iteratively. Such an iterative equalizer outperforms the linear equalization schemes but it has a very high computational complexity. Specifically for UW-OFDM, this complexity is much higher than for CP-OFDM since we have to consider the $\boldsymbol{M}$ matrix which has large dimensions, whereas in CP-OFDM the matrix dimensions are significantly smaller since a subcarrier wise detection is carried out.

C. QR decomposition based successive interference cancellation (QR-SIC) Detection

It has been shown in the literature [11], [12] that the ZF-SIC can be restated with the help of a QR decomposition of the effective channel matrix $\boldsymbol{M}$, given as

$$\boldsymbol{M} = \boldsymbol{Q}\bar{\boldsymbol{R}} = \boldsymbol{Q}\begin{bmatrix}\boldsymbol{R}\\ \boldsymbol{0}\end{bmatrix}, \tag{10}$$

where $\boldsymbol{Q} \in \mathbb{C}^{(N_m\cdot M_R)\times(N_m\cdot M_R)}$ is a unitary matrix and $\bar{\boldsymbol{R}}$ contains an upper triangular matrix $\boldsymbol{R} \in \mathbb{C}^{(N_d\cdot M_T)\times(N_d\cdot M_T)}$ and a zero matrix $\boldsymbol{0} \in \mathbb{R}^{(N_r\cdot M_T)\times(N_d\cdot M_T)}$. Multiplying the received signal $\boldsymbol{y}$ with $\boldsymbol{Q}^H$, the resulting signal is

$$\bar{\boldsymbol{s}} = \boldsymbol{Q}^H\boldsymbol{y} = \bar{\boldsymbol{R}}\tilde{\boldsymbol{s}} + \bar{\boldsymbol{n}}, \tag{11}$$

where $\bar{\boldsymbol{n}} = \boldsymbol{Q}^H\boldsymbol{n}$. As $\boldsymbol{Q}$ is a unitary matrix, the statistical properties of $\boldsymbol{n}$ and $\bar{\boldsymbol{n}}$ are the same. In the next step, a symbol wise detection is carried out. Since $\boldsymbol{R}$ has an upper triangular structure, the interference from other transmit layers and also from data symbols can be removed in an iterative manner. Various methods for the QR decomposition have been proposed in the literature. In this work, we employ a Gram-Schmidt based QR decomposition.

D. Hybrid Detection

In [6], we proposed a two step linear detection procedure which has significantly less computational complexity than joint detection. In this procedure, a subcarrier-wise detection is performed in the first step while in the second step an LMMSE based code generator demodulator is used to achieve a similar performance as achieved by joint detection using Eq. (7) and Eq. (8). In this work, we propose a new strategy where we employ a QR decomposition based code generator demodulation, as shown in Fig. 1.

After taking the DFT, subtracting the UW, and removing the zero carriers, the received signal on the lth subcarrier is given as

$$\boldsymbol{y}_l = \boldsymbol{H}_l\boldsymbol{c}_l + \boldsymbol{n}_l \in \mathbb{C}^{M_R\times 1}, \qquad l = 1, ..., N_m \tag{12}$$

where $\boldsymbol{H}_l \in \mathbb{C}^{M_R\times M_T}$ is the channel matrix of the lth subcarrier in the frequency domain and $\boldsymbol{c}_l = \boldsymbol{S}^T[\boldsymbol{G}^T]_l \in \mathbb{C}^{M_T\times 1}$ represents the transmit signal on the lth subcarrier. Here $[\boldsymbol{G}^T]_l$ is the lth column of the matrix $\boldsymbol{G}^T$.

In the first step, we apply linear equalization on each subcarrier. The output of the equalizer is given by

$$\hat{\boldsymbol{c}}_l = \boldsymbol{E}_l\boldsymbol{y}_l \in \mathbb{C}^{M_T\times 1}, \tag{13}$$

where $\boldsymbol{E}_l \in \mathbb{C}^{M_T\times M_R}$ contains the equalizer filter weights and can be calculated using the zero forcing (ZF) or the MMSE criterion. These equalizer weights are calculated as

$$\boldsymbol{E}_{l,\text{ZF}} = (\boldsymbol{H}_l^H\boldsymbol{H}_l)^{-1}\boldsymbol{H}_l^H \tag{14}$$

$$\boldsymbol{E}_{l,\text{MMSE}} = (\boldsymbol{H}_l^H\boldsymbol{H}_l + \sigma_n^2/(\sigma_d^2\beta_l)\boldsymbol{I}_{M_T})^{-1}\boldsymbol{H}_l^H, \tag{15}$$

where β_l is the norm of the lth row of the $\boldsymbol{G}$ matrix and can be precalculated.

In our previous work, we have shown that when a linear code generator demodulation matrix based on the LMMSE criterion is applied in the second step, this provides a similar performance as joint detection. In this work, we apply a QR-SIC based code generator demodulation in the second step on the equalized symbols for each transmit antenna layer. Assuming that the propagation channel is perfectly equalized ($\boldsymbol{E}_l\boldsymbol{H}_l = \boldsymbol{I}_{M_T}$), which is exactly true if Eq. (14) is used and approximately true if Eq. (15) is used, the equalized symbols for the jth transmit antenna layer are given as

$$\hat{\boldsymbol{c}}_j = \boldsymbol{G}\boldsymbol{s}_j + \hat{\boldsymbol{n}}_j, \tag{16}$$

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		Number of flops during coherence time of the channel	Number of flops for each symbol operation
UW-OFDM	Linear Joint Detectors	$N_m^2 N_d M_T^2 M_R + 2N_d^2 N_m M_T^2 M_R + N_d^3 M_T^3$	$N_m^2 N_d M_T^2 M_R + 2N_d^2 N_m M_T^2 M_R + N_d^3 M_T^3 + N_m N_d M_T M_R$
	Two Step Approach	$(2M_T^2 M_R + M_T^3)N_m + 2N_m^2 N_d + N_m^3$	$(2M_T^2 M_R + M_T^3)N_m + 2N_m^2 N_d + N_m^3 + N_m N_d M_T$
	QR-SIC	$N_d^2 N_m M_T^2 M_R$	$N_d^2 N_m M_T^2 M_R + N_m^2 M_R^2 + \frac{N_d M_T (N_d M_T - 1)}{2}$
	Hybrid Detection	$(2M_T^2 M_R + M_T^3)N_m + N_m^2 N_d$	$(2M_T^2 M_R^2 + M_T^3 M_R)N_m + N_m^2 N_d + N_m^2 + \frac{N_d(N_d-1)}{2}$
CP-OFDM	Linear Detectors	$(2M_T^2 M_R + M_T^3)N_m$	$(2M_T^2 M_R^2 + M_T^3 M_R)N_m$
	QR-SIC	$M_T^2 M_R N_m$	$(M_T^2 M_R + M_R^2 + \frac{M_T(M_T-1)}{2})N_m$

TABLE I: Computational complexity comparison of different estimation techniques in flops

where $\hat{\boldsymbol{C}} = [\hat{\boldsymbol{c}}_1, ..., \hat{\boldsymbol{c}}_{N_m}]^T \in \mathbb{C}^{N_m \times M_T}$, $\hat{\boldsymbol{c}}_j$ is the jth column of $\hat{\boldsymbol{C}}$, and $\hat{\boldsymbol{n}}_j = [\boldsymbol{E}_1(j,:)\boldsymbol{n}_1, ..., \boldsymbol{E}_{N_m}(j,:)\boldsymbol{n}_{N_m}]^T \in \mathbb{C}^{N_m \times 1}$ is the noise at the output of the detector. Here $\boldsymbol{E}_l(j,:)$ denotes the jth row of the matrix $\boldsymbol{E}_l$. Since the output noise is not white, we need to carry out a denoising step. To this end, the noise covariance matrix for each transmit layer is

$$\mathbb{E}\left\{\hat{\boldsymbol{n}}\hat{\boldsymbol{n}}_j^H\right\} = \sigma_n^2 \boldsymbol{Z}_j,$$

where $\boldsymbol{Z}_j \in \mathbb{R}^{N_m \times N_m}$ is a diagonal matrix and contains $\boldsymbol{E}_l(j,:)[\boldsymbol{E}_l(j,:)]^H \quad \forall \quad l = 1, \ldots, N_m$ on its diagonal. The equalized symbols $\hat{\boldsymbol{c}}_j$ are first pre-multiplied by $\boldsymbol{Z}_j^{-\frac{1}{2}}$ which results in

$$\boldsymbol{Z}_j^{-\frac{1}{2}}\hat{\boldsymbol{c}}_j = \boldsymbol{Z}_j^{-\frac{1}{2}}\boldsymbol{G}\boldsymbol{s}_j + \boldsymbol{Z}_j^{-\frac{1}{2}}\hat{\boldsymbol{n}}_j. \tag{17}$$

Then we compute the QR-decomposition of $\boldsymbol{Z}_j^{-\frac{1}{2}}\boldsymbol{G}$ as

$$\boldsymbol{Z}_j^{-\frac{1}{2}}\boldsymbol{G} = \boldsymbol{Q}\bar{\boldsymbol{R}}, \tag{18}$$

where $\boldsymbol{Q} \in \mathbb{C}^{N_m \times N_m}$ is a unitary matrix and $\bar{\boldsymbol{R}} = \begin{bmatrix} \boldsymbol{R} \\ \boldsymbol{0} \end{bmatrix} \in \mathbb{C}^{N_m \times N_d}$ with $\boldsymbol{R} \in \mathbb{C}^{N_d \times N_d}$ being an upper triangular matrix. The denoised signal $\bar{\boldsymbol{c}} = \boldsymbol{Z}_j^{-\frac{1}{2}}\hat{\boldsymbol{c}}_j$ is filtered with $\boldsymbol{Q}^H$, as given by

$$\boldsymbol{Q}^H\bar{\boldsymbol{c}} = \bar{\boldsymbol{R}}\boldsymbol{S}_j + \bar{\boldsymbol{n}}, \tag{19}$$

where $\bar{\boldsymbol{n}} = \boldsymbol{Q}^H\boldsymbol{Z}_j^{-\frac{1}{2}}\hat{\boldsymbol{n}}_j$. Since $\boldsymbol{Q}$ is a unitary matrix, the variance of the noise term remains unaffected. Then we start the iterative decoding and remove the interference from other symbols iteratively. Due to the upper triangular structure of $\boldsymbol{R}$, the estimated signals on the M_T layers are free from interference.

V. Complexity Analysis

In this section we perform a computational complexity analysis of the proposed linear and non-linear detection schemes for UW-OFDM. Let $\boldsymbol{A} \in \mathbb{C}^{m \times n}$ be a complex matrix, then a QR decomposition of $\boldsymbol{A}$ has an approximate complexity of mn^2 flops [13]. The inverse of a square matrix $\boldsymbol{A}$ $(m = n)$ has an approximate complexity of n^3 flops [13]. In this work, we use these definitions to calculate the approximate number of complex multiplications required by each detection scheme as summarized in Table I.

As seen from Table I, the linear and non-linear detection schemes, which are based on joint detection (utilize $\boldsymbol{H}_{\text{eff}}$ and $\boldsymbol{G}_{\text{eff}}$ in the detection procedure), have the highest computational complexity. The two step approach using linear estimators and a linear code generator demodulation matrix reduces the complexity significantly. The proposed hybrid detection scheme has a slightly lower complexity than the linear two step detection approach. This is true if channel is time varying. But in practice, the channel is assumed to be constant during the coherence time (for a burst of data in WLAN or for one subframe in LTE-A). Therefore, the coefficients of linear filters are calculated once for this time while the hybrid detection scheme requires an additional $N_m^2 + \frac{N_d(N_d-1)}{2}$ flops for each symbol during the coherence time of the channel.

VI. Simulation Results

In this section, we investigate the performance of UW-OFDM in terms of the BER with the proposed detection schemes. We also compare it with CP-OFDM. We have used LTE-A parameters for a 3 MHz bandwidth in our simulations, as described in Table II. The 3GPP channel models Extended Pedestrian A (EPA) and Extended Vehicular A (EVA) are used for the simulations. The EPA channel model has a much lower frequency selectivity than the EVA channel model. All the simulation results are averaged over 10,000 channel realizations.

	CP-OFDM	**UW-OFDM**
Modulation Scheme	QPSK	
Bandwidth	3 MHz	
Data Carriers (N_d)	180	162
Redundant Carriers (N_r)		18
FFT length (N)	256	
CP/UW duration	4.69 μs	
Channel	EPA and EVA	
MIMO Configuration $M_R \times M_T$	2×2, 4×4	

TABLE II: Simulation parameters

We have chosen two scenarios to show the performance of UW-OFDM with linear and non-linear detection schemes. In the first scenario, we have a 2×2 MIMO system and low frequency selectivity (EPA), we refer to this scenario as "mild scenario". In the second scenario, we have more interference from the antennas (4×4 MIMO) and very high frequency selectivity (EVA). We refer to this scenario as "harsh scenario".

The results show that UW-OFDM with QR-SIC outperforms the linear equalization schemes significantly in both scenarios, as shown in Fig. 2. It is also interesting to observe that the proposed hybrid scheme outperforms the linear equalization significantly. The performance of the hybrid scheme is slightly worse than QR-SIC since we still employ linear equalization in the first step. The performance of UW-OFDM differs for both scenarios and we observe huge gains for UW-OFDM with

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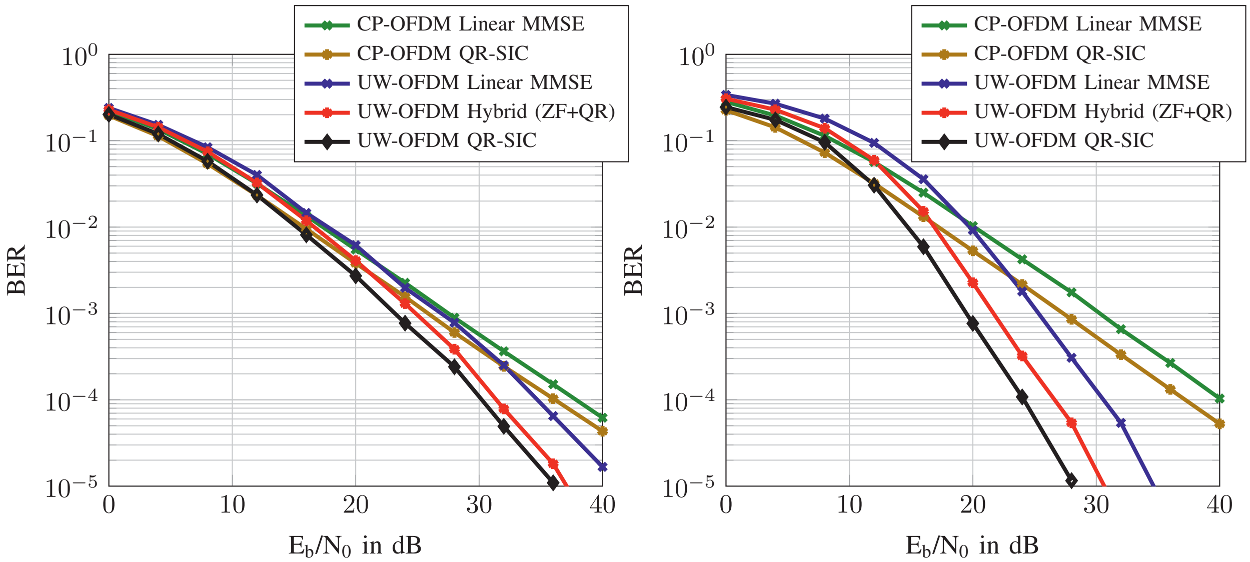

(a) 2×2 MIMO and EPA channel model

(b) 4×4 MIMO and EVA channel model

Fig. 2: Comparison of different detection schemes

linear and non-linear detection schemes over CP-OFDM in the harsh scenario. However, in the mild scenario the performance difference between UW-OFDM and CP-OFDM is reduced and we observe only a slight gain for UW-OFDM with linear equalization in the high SNRs regime. But still, UW-OFDM with hybrid detection has a considerable gain. Moreover, even UW-OFDM with linear equalization outperforms CP-OFDM with QR-SIC at high SNRs in both scenarios. Furthermore, the gain achieved by successive interference cancellation is not significant for CP-OFDM as compared to UW-OFDM.

VII. Conclusion

In this work, we have investigated the performance of UW-OFDM with SIC receivers. Specifically we have proposed a new hybrid detection scheme where we perform linear equalization in the first step and a SIC code generator demodulation in the second step. Moreover, we have performed the complexity analysis of the detection schemes where we have shown that the proposed hybrid scheme has a significantly lower complexity than the SIC receiver. UW-OFDM with SIC receivers shows high performance gains in terms of BER but at the cost of a much higher computational complexity. Note that UW-OFDM with the proposed hybrid equalization provides a trade off between performance and computational complexity. It shows a better BER performance than the linear detection schemes but the BER is slightly worse than for SIC receivers.

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