Widely Linear Processing in MIMO FBMC/OQAM Systems

Yao Cheng and Martin Haardt
Communications Research Laboratory, Ilmenau University of Technology
P. O. Box 100565, D-98694 Ilmenau, Germany
{y.cheng, martin.haardt}@tu-ilmenau.de
http://tu-ilmenau.de/crl

***Abstract*—In this paper, we investigate the use of widely linear processing in multiple-input multiple-out (MIMO) systems employing filter bank based multi-carrier with offset quadrature amplitude modulation (FBMC/OQAM). The concept of a two-step receiver combining linear processing and widely linear processing is presented. In the first step we cancel the intrinsic interference that prevents us from exploiting the benefits of widely linear processing. Two ways of estimating the interference components are developed and compared. In the second step, a widely linear MMSE receiver is employed taking into account the residual interference. A significant gain over the linear processing based receiver is demonstrated via numerical results.**

I. Introduction

Filter bank based multi-carrier modulation (FB-MC) is regarded as a promising alternative to orthogonal frequency division multiplexing with the cyclic prefix insertion (CP-OFDM). In contrast to CP-OFDM based systems, the insertion of a CP is not needed in systems employing filter bank based multi-carrier with offset quadrature amplitude modulation (FBMC/OQAM), leading to a higher spectral efficiency. Moreover, FB-MC reduces the sidelobes by using spectrally well-contained synthesis and analysis filter banks in the transmultiplexer configuration [1], [2]. Consequently it is able to avoid a high level of out-of-band radiation which CP-OFDM suffers from. These advantages of FB-MC give rise to great research attention on its use in different contexts, such as cognitive radio and professional mobile radio (PMR) networks, where an effective utilization of the available fragmented spectrum is required. In FBMC/OQAM systems, the real and imaginary parts of each complex-valued data symbol are staggered by half of the symbol period [1]. Therefore, the desired signal and the intrinsic interference are separated in the real domain and the pure imaginary domain, respectively.

However, the intrinsic interference in multiple-input multiple-output (MIMO) FBMC/OQAM systems is known as an obstacle of applying a variety of MIMO processing techniques often used when CP-OFDM is employed as the multi-carrier scheme. Widely linear processing is among one of them, while the use of OQAM gives rise to the potential of exploiting its benefits [3]. Widely linear filtering achieves a gain compared to conventional linear filtering in the case where either the correlation of the observation with its complex-conjugated version or the correlation of the observation with the complex-conjugated desired variable is non-zero [4]. The transmission of real-valued data symbols over a complex-valued channel is among such scenarios. Previous works, such as [5], have brought attention to the potential of employing widely linear processing in FBMC/OQAM systems. Nevertheless, it has not been explicitly established how to deal with the presence of the intrinsic interference such that the benefits of widely linear processing can be fully exploited. On the other hand, [7] points out that mitigating the intrinsic interference enables the application of maximum likelihood (ML) detection in MIMO transmissions with FBMC/OQAM.

Our emphasis in this contribution is on the use of widely linear processing in MIMO FBMC/OQAM systems. A two-step receiver is proposed where linear processing and widely linear processing are combined. In the first step, a linear MMSE receiver is applied. An estimate of the intrinsic interference term is obtained either by using the output of the linear MMSE receiver directly or via a reconstruction process by employing the already detected symbols. After the intrinsic interference term is canceled, a widely linear MMSE receiver is employed in the second step. It is derived taking into consideration the residual interference. Extensive simulations are performed to evaluate the performance of the proposed two-step receiver.

The remainder of the paper is organized as follows. In Section II, we first briefly introduce non-circular signals and the concept of widely linear processing followed by the data model of a point-to-point MIMO FBMC/OQAM system. Then the proposed two-step receiver is detailed in Section III. Numerical results are presented in Section IV, before conclusions are drawn in Section V.

II. Preliminaries

A. Non-circular signals and widely linear processing

Let us denote a complex-valued random vector by $\boldsymbol{v} = \boldsymbol{v}_{\mathrm{I}} + \jmath\boldsymbol{v}_{\mathrm{Q}} \in \mathbb{C}^M$, where $\boldsymbol{v}_{\mathrm{I}}, \boldsymbol{v}_{\mathrm{Q}} \in \mathbb{R}^M$ are zero-mean random vectors. The autocorrelation matrix and the pseudo-autocorrelation matrix of $\boldsymbol{v}$ are written as

$$\boldsymbol{\Phi}_{vv} = \mathbb{E}\{\boldsymbol{v}\boldsymbol{v}^{\mathrm{H}}\}$$

and

$$\boldsymbol{\Phi}_{vv^*} = \mathbb{E}\{\boldsymbol{v}\boldsymbol{v}^{\mathrm{T}}\},$$

respectively. The complex-valued random vector $\boldsymbol{v}$ is proper or second-order circular if $\boldsymbol{\Phi}_{vv^*}$ is an all-zero matrix. Otherwise,

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TABLE I
COEFFICIENTS DETERMINED BY THE SYSTEM IMPULSE OF THE SYNTHESIS AND ANALYSIS FILTERS

	$n-3$	$n-2$	$n-1$	n	$n+1$	$n+2$	$n+3$
$k-1$	$0.043\jmath$	$-0.125\jmath$	$-0.206\jmath$	$0.239\jmath$	$0.206\jmath$	$-0.125\jmath$	$-0.043\jmath$
k	$-0.067\jmath$	0	$0.564\jmath$	0	$0.564\jmath$	0	$-0.067\jmath$
$k+1$	$-0.043\jmath$	$-0.125\jmath$	$0.206\jmath$	$0.239\jmath$	$-0.206\jmath$	$-0.125\jmath$	$0.043\jmath$

it is called improper or non-circular [8]. By employing some modulation formats, such as binary phase shift keying (BPSK), amplitude shift keying (ASK), or OQAM, the resulting data signals exhibit non-circularity which may be exploited by widely linear processing at the receiver. It is important to note that for an improper vector $\boldsymbol{v}$, its second-order statistics are described by both $\boldsymbol{\Phi}_{vv}$ and $\boldsymbol{\Phi}_{vv^*}$. By stacking $\boldsymbol{v}$ itself and its complex conjugate $\boldsymbol{v}^*$, a complex-valued augmented vector $\widetilde{\boldsymbol{v}}$ is obtained as

$$\widetilde{\boldsymbol{v}} = \left[\begin{array}{cc} \boldsymbol{v}^{\mathrm{T}}, & \boldsymbol{v}^{\mathrm{H}} \end{array}\right]^{\mathrm{T}} \in \mathbb{C}^{2M}. \tag{1}$$

Moreover, the autocorrelation matrix of $\widetilde{\boldsymbol{v}}$, also as an augmented version of the autocorrelation matrix of $\boldsymbol{v}$, is identified as

$$\boldsymbol{\Phi}_{\widetilde{v}\widetilde{v}} = \left[\begin{array}{cc} \boldsymbol{\Phi}_{vv} & \boldsymbol{\Phi}_{vv^*} \\ \boldsymbol{\Phi}^*_{vv^*} & \boldsymbol{\Phi}^*_{vv} \end{array}\right] \in \mathbb{C}^{2M \times 2M}. \tag{2}$$

When $\boldsymbol{v}$ is non-circular, $\boldsymbol{\Phi}_{\widetilde{v}\widetilde{v}}$ fully characterizes its second-order statistics. This also corresponds to the principle of widely linear filtering [4] where the filter outputs of the input signal and its complex conjugate are combined, leading to an improved performance.

B. System model

Consider a point-to-point MIMO FBMC/OQAM system with M_{T} transmit antennas and M_{R} receive antennas. The channel is assumed to be mildly frequency-selective, where each subchannel can be treated as flat fading. The received vector on the k-th subcarrier and at the n-th time instant is then written as[1] [6], [7]

$$\boldsymbol{y} = \boldsymbol{H}\left(\boldsymbol{d} + \jmath \boldsymbol{u}\right) + \boldsymbol{n} \in \mathbb{C}^{M_{\mathrm{R}}}, \tag{3}$$

where $\boldsymbol{d} \in \mathbb{R}^{M_{\mathrm{T}}}$ is the desired signal on the k-th subcarrier and at the n-th time instant (alternatively written as $\boldsymbol{d}_k[n]$), $\boldsymbol{u} \in \mathbb{R}^{M_{\mathrm{T}}}$ and $\jmath\boldsymbol{u}$ is the pure imaginary intrinsic interference. Here $\boldsymbol{H} \in \mathbb{C}^{M_{\mathrm{R}} \times M_{\mathrm{T}}}$ contains the channel gains between each transmit antenna and each receive antenna, and $\boldsymbol{n}$ denotes the additive white Gaussian noise vector with variance σ_n^2. Note that when the PHYDYAS prototype filter [9] is used and the overlapping factor is chosen to be $K = 4$, the pure imaginary interference on the k-th subcarrier and at the n-th time instant is represented as [7]

$$\jmath \boldsymbol{u} = \sum_{i=n-3}^{n+3} \sum_{j=k-1}^{k+1} c_{ij} \cdot \boldsymbol{d}_j[i], \quad j \neq k \text{ and } i \neq n, \tag{4}$$

where the coefficients c_{ij} represent the system impulse response determined by the synthesis and analysis filters. Equivalently the imaginary units $\jmath$ in the OQAM modulated data symbols are shifted to their corresponding real-valued coefficients such that the transmit symbols can be regarded as all real-valued. The resulting coefficients are presented in Table I. When a linear MMSE receiver is employed, the desired symbols can be obtained as [7]

$$\hat{\boldsymbol{s}} = \boldsymbol{W}^{\mathrm{H}}_{\mathrm{MMSE}} \cdot \boldsymbol{y}, \tag{5}$$

$$\hat{\boldsymbol{d}} = \mathrm{Re}\{\hat{\boldsymbol{s}}\}. \tag{6}$$

To apply widely linear processing, the obstacle is the pure imaginary intrinsic interference. In the sequel, we devise a two-step receiver to tackle this problem such that the benefits of widely linear processing can be exploited.

III. PROPOSED TWO-STEP RECEIVER

We propose a two-step receiving procedure where linear processing and widely linear processing are combined. First, a linear MMSE receiver is applied to the received signal. The goal of the first step is to obtain an estimate of the intrinsic interference $\jmath\hat{\boldsymbol{u}}$. Assuming perfect channel state information at the receiver, the estimated interference component can be subtracted from the received signal as

$$\widetilde{\boldsymbol{y}} = \boldsymbol{H}\left(\boldsymbol{d} + \jmath\boldsymbol{u}\right) - \boldsymbol{H} \cdot \jmath\hat{\boldsymbol{u}} + \boldsymbol{n} = \boldsymbol{H}\left(\boldsymbol{d} + \jmath\boldsymbol{\epsilon}\right) + \boldsymbol{n}, \tag{7}$$

where $\jmath\boldsymbol{\epsilon}$ is the pure imaginary residual interference. We present two ways of obtaining an estimate of the interference. The first method is to directly take the imaginary part of the output of the linear MMSE receiver on the k-th subcarrier and at the n-th time instant as

$$\jmath\hat{\boldsymbol{u}} = \jmath \cdot \mathrm{Im}\{\boldsymbol{W}^{\mathrm{H}}_{\mathrm{MMSE}} \cdot \boldsymbol{y}\}. \tag{8}$$

On the other hand, the idea of the second scheme is to construct an estimate of the interference by using the already detected data symbols shown as follows

$$\jmath\hat{\boldsymbol{u}} = \sum_{i=n-3}^{n+3} \sum_{j=k-1}^{k+1} c_{ij} \cdot \check{\boldsymbol{d}}_j[i], \quad j \neq k \text{ and } i \neq n, \tag{9}$$

where the $\check{\boldsymbol{d}}_j[i]$ are obtained by modulating the information bits detected from $\hat{\boldsymbol{d}}$ using OQAM.

[1] In this expression the index of the subcarrier and the index of the time instant are ignored for simplicity of notations. These indices only appear when explaining the intrinsic interference caused by adjacent subcarriers and time instants (see (4))

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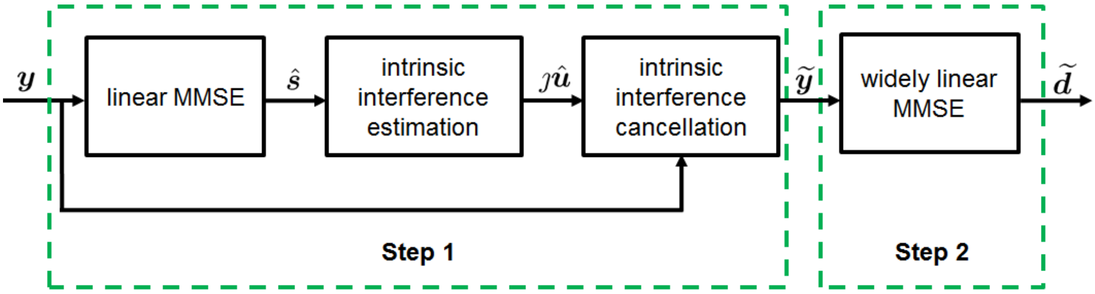

Fig. 1. Block diagram of the proposed two-step receiver

In the second step, a widely linear MMSE receiver is employed on the resulting equivalent received signal $\widetilde{\boldsymbol{y}}$ and its complex conjugate. The detected desired signal is accordingly expressed as

$$\widetilde{\boldsymbol{d}} = \mathrm{Re}\left\{\left[\begin{array}{cc} \boldsymbol{W}_1^{\mathrm{H}} & \boldsymbol{W}_2^{\mathrm{H}} \end{array}\right] \cdot \left[\begin{array}{c} \widetilde{\boldsymbol{y}} \\ \widetilde{\boldsymbol{y}}^* \end{array}\right]\right\}, \tag{10}$$

where

$$\boldsymbol{W}_1 = \left(\boldsymbol{R}_{\widetilde{y}\widetilde{y}} - \boldsymbol{R}_{\widetilde{y}\widetilde{y}^*} \cdot \boldsymbol{R}_{\widetilde{y}\widetilde{y}}^{*\,-1} \cdot \boldsymbol{R}_{\widetilde{y}\widetilde{y}^*}^*\right)^{-1} \cdot \left(\boldsymbol{H} - \boldsymbol{R}_{\widetilde{y}\widetilde{y}^*} \cdot \boldsymbol{R}_{\widetilde{y}\widetilde{y}}^{*\,-1} \cdot \boldsymbol{H}^*\right)$$

and

$$\boldsymbol{W}_2 = \left(\boldsymbol{R}_{\widetilde{y}\widetilde{y}}^* - \boldsymbol{R}_{\widetilde{y}\widetilde{y}^*}^* \cdot \boldsymbol{R}_{\widetilde{y}\widetilde{y}}^{-1} \cdot \boldsymbol{R}_{\widetilde{y}\widetilde{y}^*}\right)^{-1} \cdot \left(\boldsymbol{H}^* - \boldsymbol{R}_{\widetilde{y}\widetilde{y}^*}^* \cdot \boldsymbol{R}_{\widetilde{y}\widetilde{y}}^{-1} \cdot \boldsymbol{H}\right).$$

Here $\boldsymbol{R}_{\widetilde{y}\widetilde{y}}$ and $\boldsymbol{R}_{\widetilde{y}\widetilde{y}^*}$ are the autocorrelation matrix of the equivalent received signal $\widetilde{y}$ and its pseudo-autocorrelation matrix (the correlation matrix of $\widetilde{y}$ and $\widetilde{y}^*$), respectively.

Let P_{s} denote the power per symbol at each transmit antenna. The covariance matrix and the pseudo-covariance matrix of the desired signal $\boldsymbol{d} \in \mathbb{R}^{M_{\mathrm{T}}}$ are

$$\mathbb{E}\{\boldsymbol{d}\boldsymbol{d}^{\mathrm{H}}\} = \mathbb{E}\{\boldsymbol{d}\boldsymbol{d}^{\mathrm{T}}\} = \frac{P_{\mathrm{s}}}{2}\boldsymbol{I}_{M_{\mathrm{T}}}. \tag{11}$$

Assuming that the residual interference is uncorrelated with the desired signal and the noise as well, i.e.,

$$\mathbb{E}\{\boldsymbol{d} \cdot \boldsymbol{\epsilon}^{\mathrm{H}}\} = \mathbb{E}\{\boldsymbol{d} \cdot \boldsymbol{\epsilon}^{\mathrm{T}}\} = \boldsymbol{0}, \tag{12}$$

$$\mathbb{E}\{\boldsymbol{\epsilon} \cdot \boldsymbol{n}^{\mathrm{H}}\} = \boldsymbol{0}, \tag{13}$$

we obtain

$$\boldsymbol{R}_{\widetilde{y}\widetilde{y}} = \boldsymbol{H} \cdot \left(\frac{P_{\mathrm{s}}}{2}\boldsymbol{I}_{M_{\mathrm{T}}} + \boldsymbol{R}_{\epsilon\epsilon}\right) \cdot \boldsymbol{H}^{\mathrm{H}} + \sigma_n^2 \boldsymbol{I}_{M_{\mathrm{R}}}, \tag{14}$$

and

$$\boldsymbol{R}_{\widetilde{y}\widetilde{y}^*} = \boldsymbol{H} \cdot \left(\frac{P_{\mathrm{s}}}{2}\boldsymbol{I}_{M_{\mathrm{T}}} - \boldsymbol{R}_{\epsilon\epsilon}\right) \cdot \boldsymbol{H}^{\mathrm{T}}. \tag{15}$$

Here the covariance matrix of the residual interference related term $\boldsymbol{\epsilon}$ denoted by $\boldsymbol{R}_{\epsilon\epsilon}$ is calculated as

$$\boldsymbol{R}_{\epsilon\epsilon} \approx \frac{1}{N}\sum_{l=1}^{N}(\boldsymbol{u}_l - \hat{\boldsymbol{u}}_l) \cdot (\boldsymbol{u}_l - \hat{\boldsymbol{u}}_l)^{\mathrm{H}}. \tag{16}$$

A block diagram that illustrates the two-step receiver introduced above is shown in Fig. 1.

Notice that to obtain an estimate of the intrinsic interference using (9) there is a delay since three future multi-carrier symbols have to be detected and used (cf. also Table I). By making use of only two detected future symbols or even one, the delay is consequently smaller at the price of a performance degradation due to a larger residual interference term. This issue is further analyzed and discussed in Section IV where the simulation results are shown.

With a focus on point-to-point MIMO FBMC/OQAM systems, we have devised a way of dealing with the intrinsic interference and thus enabling the use of widely linear processing. In fact, it also inspires the design of widely linear processing-based techniques in more complicated communication scenarios. For instance, in uplink transmissions where cooperative MIMO is used, FBMC/OQAM is regarded as a promising alternative to CP-OFDM as the latter suffers from performance degradation due to the lack of synchronization in both the time and the frequency domain. The robustness of FBMC/OQAM in uplink transmissions with timing and frequency misalignments between the nodes has been numerically shown in [10], [11]. It is thus interesting to investigate how widely linear processing can be incorporated into a cooperative MIMO FBMC/OQAM system. In such a case, relay nodes assist the transmissions from the source or sources to the destination such that the source signals arrive at the destination through a number of independent paths and are constructively combined. Unlike the point-to-point MIMO scenario where multiple antennas at the transmit and receive nodes provide spatial diversity, the relay nodes form a virtual antenna array which enhances the reliability of the transmissions without requiring multiple antennas at the nodes. To exploit the cooperative diversity of such a system, distributed beamforming is an effective technique. Compared to linear distributed beamforming algorithms, it is reported in [12] that by applying widely linear processing a significant gain can be obtained. However, it should be noted that when FBMC/OQAM is used and amplify-and-forward is considered as the relaying scheme, the overall interference term observed at the destination that results from the intrinsic interference induced in all phases of the transmissions is a challenge. By employing decode-and-forward instead, the intrinsic interference can then be mitigated at the relay nodes based on the idea introduced in the previous text such that the benefits due to the second-order non-circular property of the equivalently real-valued desired signals can be exploited.

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IV. Simulation Results

In what follows, we evaluate the bit error rate (BER) performance of the proposed two-step receiver for MIMO FBMC/OQAM systems. The number of subcarriers is 512. Each subchannel is considered as Rayleigh flat fading. Perfect channel state information at the receiver is assumed. In addition, the PHYDYAS prototype filter with the overlapping factor $K = 4$ [9] is employed. First, a scenario where $M_{\mathrm{T}} = M_{\mathrm{R}} = 2$ is considered. Fig. 2 presents a comparison between the BER performances of the linear MMSE receiver (as represented by (5) and (6)) and the proposed two-step receiver combining both linear processing and widely linear processing. Note that in the legend of the figure "MMSE + WL-MMSE 1" and "MMSE + WL-MMSE 2" correspond to the two versions of the two-step receiver where the estimated interference is obtained by using (8) and (9), respectively. We also illustrate in Fig. 2 an ideal case where the interference component is completely canceled, i.e., $\boldsymbol{\epsilon} = \mathbf{0}$ in (7), corresponding to the legend "WL-MMSE ideal". A performance improvement is

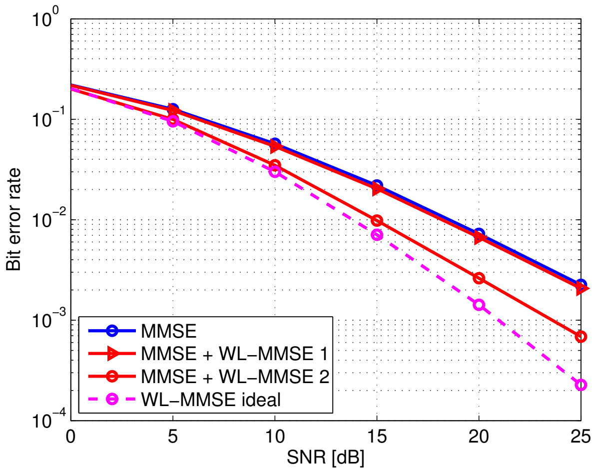

Fig. 2. BER vs. SNR for the case where $M_{\mathrm{T}} = M_{\mathrm{R}} = 2$ ("MMSE + WL-MMSE 1" and "MMSE + WL-MMSE 2" correspond to the two versions of the two-step receiver where the estimated interference is obtained by using (8) and (9), respectively; "WL-MMSE ideal" corresponds to an ideal case where the interference component is completely canceled)

observed by employing our proposed two-step receiver based on (7), (9) and (10) compared to the case where the linear MMSE receiver is used. It can also be seen that when the interference is estimated by using (8), the achieved gain over the linear MMSE receiver is negligible due to a relatively high level of the residual interference. In addition to the non-zero residual interference, the violation of the assumptions (12) and (13) as well as the approximation of $\boldsymbol{R}_{\epsilon\epsilon}$ also contribute to the gap between the performance of the two-step receiver and the ideal case plotted for the purpose of comparison.

In the second example, we consider a $M_{\mathrm{T}} = M_{\mathrm{R}} = 4$ scenario and illustrate the corresponding results in Fig. 3. Similar observations as in the first experiment are obtained. Moreover, it is shown that by increasing the number of transmit and receive antennas, the gain achieved by the proposed two-step receiver over the linear MMSE receiver is even more significant when the estimated interference is obtained based on (9).

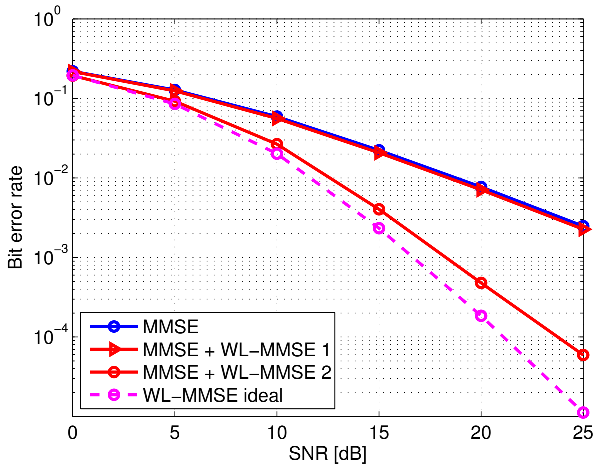

Fig. 3. BER vs. SNR for the case where $M_{\mathrm{T}} = M_{\mathrm{R}} = 4$ ("MMSE + WL-MMSE 1" and "MMSE + WL-MMSE 2" correspond to the two versions of the two-step receiver where the estimated interference is obtained by using (8) and (9), respectively; "WL-MMSE ideal" corresponds to an ideal case where the interference component is completely canceled)

Now we concentrate on the two-step receiver where the intrinsic interference is estimated according to the second scheme as in (9). Its performance is evaluated when a smaller number of detected future multi-carrier symbols are used, and the intrinsic interference is thus only partially canceled. The simulation parameters are the same as in the first experiment. It can be observed in Fig. 4 that when two detected future

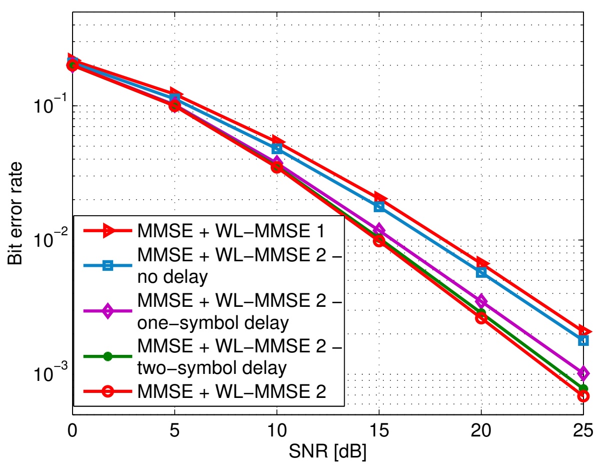

Fig. 4. BER vs. SNR for the case where $M_{\mathrm{T}} = M_{\mathrm{R}} = 2$ ("MMSE + WL-MMSE 1" and "MMSE + WL-MMSE 2" correspond to the two versions of the two-step receiver where the estimated interference is obtained by using (8) and (9), respectively)

symbols are used, the performance degradation compared to the case where all three future symbols contributing to the intrinsic interference are considered is quite small. As the

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delay is further reduced to one symbol, i.e., only one detected future symbol is used, the gap in the performance, which is around 2 dB, is still on an acceptable level. However, when no detected future symbols are utilized at all, the amount of the residual interference is large, and the performance degrades heavily. Still, it outperforms the version of the two-step receiver based on the first scheme of estimating the intrinsic interference as in (8).

Finally, we present a comparison between the proposed two-step receiver and the MMSE-ML scheme in [7]. In this technique, the intrinsic interference is also first mitigated by using the output of a linear MMSE receiver, and the ML detection is applied afterwards. In this example, the scenario is the same as for Fig. 3. For the purpose of comparison, the versions of the two-step receiver with less delay are also considered. The results are shown in Fig. 5. We can observe that the proposed

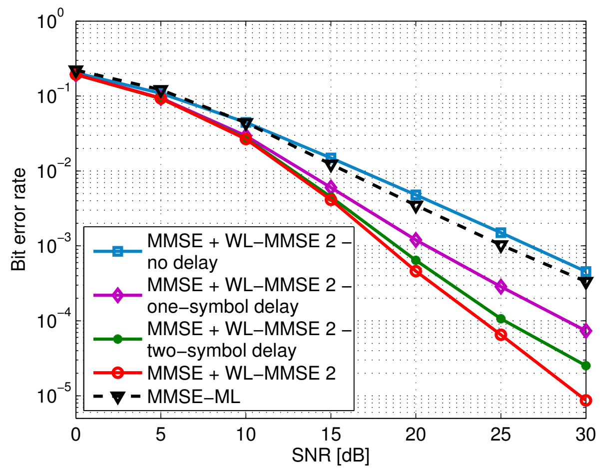

Fig. 5. BER vs. SNR for the case where $M_{\mathrm{T}} = M_{\mathrm{R}} = 4$ ("MMSE + WL-MMSE 2" corresponds to the version of the two-step receiver where the estimated interference is obtained by using (9); "MMSE-ML" corresponds to the scheme combining MMSE and ML in [7])

two-step receiver significantly outperforms the MMSE-ML scheme even when the intrinsic interference is only partially canceled and a smaller delay is incurred. Although the MMSE-ML technique provides a better performance compared to the version of the two-step receiver without taking advantage of any detected future symbols and thus causing no delay, the ML detection in MMSE-ML leads to a higher computational complexity in contrast to the widely linear processing part of the two-step receiver.

V. Conclusion

In this contribution, we develop a two-step receiver for a point-to-point MIMO FBMC/OQAM system. The intrinsic interference term is first estimated by using the output of a linear MMSE receiver and then subtracted from the received signal. A widely linear MMSE receiver is further applied on the resulting signal. Two schemes of estimating the intrinsic interference are presented and analyzed via simulations. In the first scheme, an estimate of the intrinsic interference is obtained by directly taking the imaginary part of the output of the linear MMSE receiver. It causes no delay in the processing but fails to provide a performance improvement compared to the case of a linear MMSE receiver. On the other hand, the second scheme uses the detected adjacent multi-carrier symbols. The two-step receiver with this scheme achieves a very promising performance. Since the knowledge of detected future symbols is required, some delay is incurred. We could cope with this issue by using a smaller number of detected future symbols with a slightly degraded performance.

In the future, the statistics of the residual interference will be analyzed. Some analytical performance evaluations will also be carried out. Moreover, the channel model will not be restricted to the flat fading case. Frequency selective channels will be considered, and the proposed scheme will be adapted to deal with the resulting inter-carrier interference and inter-symbol interference. In addition, the extension of the proposed scheme to cooperative MIMO FBMC/OQAM systems based on the remarks at the end of Section III is also interesting.

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