POWER BALANCING IN FBMC-MISO SYSTEMS

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ABSTRACT

This paper addresses the fairness in power allocation optimization for FBMC-MISO systems. Given a power budget we have proposed a sub-optimal strategy to distribute the power among the subcarriers. Firstly, with the aim of mitigating the interference terms as much as possible two precoding techniques have been proposed: WMF and Tx-MDIR. As a benchmark we have considered a transmit filter that is designed according to the ZF criterion. Then, assuming that the noise power dominates over the interferences we have allocated the power according to the solution provided by the max-min SINR problem which balances the SINRs. When the channel can be assumed flat in the subcarrier pass band region both ZF and Tx-MDIR can take advantage of the power control to achieve a BER equal to 10^{-4} by saving 3dB in the transmitted power with respect to the uniform power allocation strategy. Simulation-based results have shown that for very frequency selective channels the BER plot of ZF exhibits an error floor. In this scenario, Tx-MDIR and WMF are able to boost the BER in the Eb/N0 range [15dB-30dB] and [24dB-30dB], respectively.

1. INTRODUCTION

Filter bank based multicarrier (FBMC) systems are a promising solution to overcome the problems that substantially degrade the performance of orthogonal frequency division multiplexing (OFDM) systems. Under perfect synchronization, OFDM is able to smartly remove the inter-symbol interference (ISI) and inter-carrier interference (ICI) by appending a cyclic prefix (CP) at the beginning of each block. However, when synchronization fails the orthogonality conditions are not fulfilled leading to leakage between subcarrier signals. Since FBMC shapes the symbols with waveforms that have a very good time-frequency localization, both the carrier frequency offsets (CFO) and the narrow band interferences (NBI) yield a performance degradation which has less impact in comparison to the OFDM case. Besides, FBMC exhibits a higher spectral efficiency since no redundancy is required to combat ISI and ICI. Nevertheless, under multipath propagation conditions further processing is required to restore the orthogonality properties of the prototype pulse. When the channel can be assumed flat in the subcarrier passband region, single tap equalization suffices. For high frequency selective scenarios broadband processing enhances the link reliability as [1] and [2] shows.

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In this paper we consider a FBMC system that staggers real and imaginary parts of the input symbols according the offset quadrature amplitude modulation (OQAM). This scheme is known as OFDM/OQAM, [3]. Likewise OFDM, the bit error rate (BER) performance crucially depends on the spectral nulls that the channel frequency response may have. One solution to overcome this problem consists of coding the data with a convolutional encoder, [4]. Provided that the channel state information is available at the transmit side the power can be also smartly distributed among the subcarriers with the aim of enhancing the quality of the worst subbands. Furthermore, spatial diversity provides additional degrees of freedom to increase the robustness. However, the research on transmit processing for OFDM systems is much more mature than it is in the FBMC context. In [5] the authors have tailored a joint transmit-receive processing originally devised for OFDM. However, this technique degrades as the channel becomes more frequency selective. Discarding the assumption that the channel coherence bandwidth is higher than the subchannel bandwidth, [6],[7] propose specific designs for FBMC.

Considering an uncoded FBMC system this paper studies how to boost the performance in terms of BER without making any assumption about the channel flatness. With the objective of devising low-complexity architectures we have decoupled the design of the precoders from the power control. This paper tries to give insight into the transmit processing under multipath fading when subcarrier signals overlap both in time and frequency domain. In [6] the authors have proposed optimal and sub-optimal transmit filters aiming at minimizing the transmitted power while quality of service constraints are satisfied. By contrast, given a power budget, the work presented in this paper maximizes the minimum signal to interference plus noise ratio (SINR).

The reminder of the paper is organised as follows. In section II we briefly desrcibe the OFDM/OQAM system model. Given a set of precoders Section III addresses a max-min approach to distribute the power among the subcarriers. With the aim of simplifying the power allocation strategy, two transmit filter designs are devised in Section IV. Simulation results are given in Section V. Finally the conclusion is provided in Section VI.

2. OVERVIEW OF THE OFDM/OQAM SYSTEM MODEL

This section briefly describes the OFDM/OQAM baseband system model in the discrete time domain. Most of the notation has been borrowed from [6].

The transmitted signal is built by multiplexing the symbols in the frequency domain so that each subcarrier signal is shaped with a very spectral efficient waveform. Specifically, we have considered the prototype pulse proposed in [8] with an overlapping factor K=4. The transmitted symbols are precoded on a per-subcarrier basis before being fed to the synthesis filter bank. Taking into account the multi-antenna configuartion of the transmitter, the signal transmitted by the *i-th* ($i = 1,...,N_t$) antenna can be formulated as

$$s_i[k] = \sum_{n=-\infty}^{\infty}\sum_{m=0}^{M-1}\left(b_m^i[n]^* * x_m[n]\right) f_m\left[k - nM/2\right] \quad (1)$$

$$f_m[k] = p[k]exp\left(j\frac{2\pi}{M}m\left(k-(L-1)/2\right)\right) \quad (2)$$

where p[k] accounts for the causal prototype pulse, L for the pulse length and M is the number of bands. Note that the symbols $x_m[n]$ are precoded by linear filters $b_m^i[n]$. The OQAM scheme consists of staggering real and imaginary parts of the complex inputs which are drawn from a QAM constellation. This can be compactly formulated by defining $x_m[n] = d_m[n]\theta_m[n]$. Let $d_m[n]$ be real PAM symbols, the term $\theta_m[n]$ takes real and pure imaginary values depending on the subcarrier index and the time instant.

$$\theta_m[n] = \left\{\begin{array}{ll} 1 & m+n \;\; even \\ j & m+n \;\; odd \end{array}\right\} \quad (3)$$

At the receive side each transmitted signal is distorted by multipath fading and contaminated by additive white Gaussian noise (AWGN). Noise samples are independent and identically distributed, i.e. $E\left\{w[k]w[l]^*\right\} = \delta_{kl}N_0$.

$$r[k] = \sum_{i=1}^{Nt} h_i[k] * s_i[k] + w[k] \quad (4)$$

In order to demodulate the data transmitted on each subband the received signal is fed to the analysis filter bank. At its output, the demodulated signal on the *q-th* band ($q = 0,...,M-1$) is given by $y_q[k] = \left(r[k] * f_q[-k]^*\right)_{\downarrow M/2}$, where $(.)_{\downarrow M/2}$ accounts for the downsampling operation by a factor $M/2$. It must be highlighted that the proposed scheme puts the complexity burden on the transmitter since no equalization is required by the receiver. Thanks to the good spectral confinement of the transmit pulses the demodulated signal are only interfered by the adjacent subcarrier signals. The compact model can be written as follows:

$$y_q[k] = \sum_{m=q-1}^{q+1}\sum_{i=1}^{Nt}\left(b_m^i[k]^* * x_m[k]\right) * g_{qm}^i[k] + w_q[k] \quad (5)$$

$$\begin{array}{l} g_{qm}^i[k] = \left(f_q[-k]^* * h_i[k] * f_m[k]\right)_{\downarrow \frac{M}{2}} \\ k = -L_{g_2},...,L_{g_1} \end{array} \quad (6)$$

The length of $g_{qm}^i[k]$ depends on the channel impulse response and the prototype pulse length. Since not all the taps of $g_{qm}^i[k]$ degrade significantly the signal, only some of which has been considered, thus we have set $L_{g_1} = L_{g_2} = 5$. For the sake of notation equation (5) is reformulated in a matrix way.

$$\begin{aligned} y_q[k] &= \sum_{m=q-1}^{q+1}\sum_{i=1}^{Nt}\left(\mathbf{b}_m^i\right)^H \mathbf{G}_{qm}^i\mathbf{x}_m[k] + w_q[k] \\ &= \sum_{m=q-1}^{q+1}\mathbf{b}_m^H\mathbf{G}_{qm}\mathbf{x}_m[k] + w_q[k] \end{aligned} \quad (7)$$

$$\mathbf{b}_m^i = \left[b_m^i[-L_{b_2}]\cdots b_m^i[L_{b_1}]\right]^T \quad (8)$$

$$\mathbf{x}_m[k] = \left[x_m[k+L_{b_1}+L_{g_1}]\cdots x_m[k-L_{b_2}-L_{g_2}]\right]^T \quad (9)$$

The matrix $\mathbf{G}_{qm}^i$ is a Toeplitz matrix, which first row contains these terms $[g_{qm}^i[-L_{g_2}]...g_{qm}^i[L_{g_1}]0...0]$. Matrix $\mathbf{G}_{qm}$ is constructed by stacking column-wise the convolution matrices $\mathbf{G}_{qm}^i$ ($i = 1,...,N_t$). Vector $\mathbf{b}_m$ is generated likewise.

The real PAM symbols can be detected by de-staggering the outputs of the analysis filter bank.

$$\hat{d}_q[k] = \Re\left(\theta_q[k]^* y_q[k]\right) \quad (10)$$

Taking into account how real and pure imaginary symbols are interleaved we can formulate the sample $z_q[k] = \theta_q[k]^* y_q[k]$ as function of real data streams.

$$\begin{aligned} z_q[k] &= \theta_q[k]^*\left(\sum_{m=q-1}^{q+1}\mathbf{b}_m^H\mathbf{G}_{qm}\mathbf{D}_m[k]\mathbf{d}_m[k] + w_q[k]\right) \\ &= \sum_{m=q-1}^{q+1}\mathbf{b}_m^H\bar{\mathbf{G}}_{qm}[k]\mathbf{d}_m[k] + \bar{w}_q[k] \end{aligned} \quad (11)$$

$$\mathbf{D}_m[k] = diag\left[\theta_m[k+L_{b_1}+L_{g_1}]\cdots\theta_m[k-L_{b_2}-L_{g_2}]\right] \quad (12)$$

$$\mathbf{d}_m[k] = \left[d_m[k+L_{b_1}+L_{g_1}]\cdots d_m[k-L_{b_2}-L_{g_2}]\right]^T \quad (13)$$

By defining the following extended notation $\mathbf{a}_e = [\Re\left(\mathbf{a}^T\right)\Im\left(\mathbf{a}^T\right)]^T$ (10) can be expressed as (14) shows.

$$\hat{d}_q[k] = \sum_{m=q-1}^{q+1}\mathbf{b}_{m,e}^T\bar{\mathbf{G}}_{qm,e}[k]\mathbf{d}_m[k] + \Re\left(\bar{w}_q[k]\right) \quad (14)$$

3. POWER ALLOCATION

Provided that precoders have been already designed, this section addresses the processing to set transmission powers by following a max-min approach. Given a power budget, the strategy consists of maximizing the minimum SINR (max-min SINR). This strategy has been proved to be definitely effective in reducing the BER in the OFDM context, [9]. As Section II features, demodulated streams are degraded by ISI and ICI, which leads to more complex problem in comparison with OFDM. Without loss of generality it has been assumed that precoders have unit norm. Hence the SINR is function of the transmit powers p_q, ($q = 0,...,M-1$).

$$SINR_q = \frac{p_q\left\|\mathbf{b}_{q,e}^T\bar{\mathbf{G}}_{qq,e}[k]\mathbf{e}_l\right\|^2}{i_q + N_0/2} = \frac{p_q\left\|h_q\right\|^2}{i_q + N_0/2} \quad (15)$$

$$\begin{array}{l} i_q = \sum_{m=q-1,m\neq q}^{q+1} p_m\left\|\mathbf{b}_{m,e}^T\bar{\mathbf{G}}_{qm,e}[k]\right\|^2 + \\ p_q\left\|\mathbf{b}_{q,e}^T\bar{\mathbf{G}}_{qq,e}[k]\left(\mathbf{I}-\mathbf{e}_l\mathbf{e}_l^T\right)\right\|^2 \end{array} \quad (16)$$

We have considered that the real PAM symbols have a power equal to 1 as well as they are independent and identically distributed. In notation terms $\mathbf{I}$ accounts for the identity matrix. The vector $\mathbf{e}_l$ is designed to have all the entries equal to 0 except in the *l-th* position where the value is 1. Under the criterion of selecting the row of matrix $\bar{\mathbf{G}}_{qq,e}[k]$ which has the highest coefficients, we have set $l = 1 + L_{b_1} + L_{g_1}$. It must be pointed out that (15) does not depend on the time instant. In this regard, it can be checked that $SINR_q$ remains the same regardless of the value that k may take. For this reason we have dropped the index k when defining the SINRs.

3.1 max-min SINR

The optimization problem can be formulated as follows:

$$\begin{array}{c} \max_{p_0,\ldots,p_{M-1}} \min_{0\leq q\leq M-1} SINR_q \\ s.t \ \sum_{q=0}^{M-1} p_q = P_T, \ \ p_q \geq 0 \end{array} \tag{17}$$

It can be verified that $SINR_q$ is monotonically increasing in p_q and monotonically decreasing in p_{q-1}, p_{q+1}. As a result, the optimal solution adjusts the powers to balance the SINRs, [10]. In case one of the subcarriers is suffering from a deep notch, this subcarrier would get all the power leading to performance degradation. Nevertheless this can be circumvented by resorting to spatial diversity as long as propagation conditions of different links present low-correlation. Then all the bands can provide the same quality of service with no need to allocate most of the power on a single band. Regarding the max-min problem it can be solved by carrying out the singular value decomposition (SVD) of matrix Γ.

$$\Gamma = \left[\begin{array}{cc} \mathbf{D}\psi & (0.5N_0)\,\mathbf{D1} \\ (1/P_T)\,\mathbf{1}^T\mathbf{D}\psi & (0.5N_0/P_T)\,\mathbf{1}^T\mathbf{D1} \end{array} \right] \tag{18}$$

$$\mathbf{D} = diag\left[\frac{1}{\|h_0\|^2}, \ldots, \frac{1}{\|h_{M-1}\|^2}\right] \tag{19}$$

$$[\Psi]_{qi} = \left\{ \begin{array}{cc} \left\|\mathbf{b}_{q,e}^T \bar{\mathbf{G}}_{qq,e}[k]\left(\mathbf{I} - \mathbf{e}_l\mathbf{e}_l^T\right)\right\|^2 & i = q \\ \left\|\mathbf{b}_{i,e}^T \bar{\mathbf{G}}_{qi,e}[k]\right\|^2 & i \in S_q \\ 0 & otherwise \end{array} \right\} \tag{20}$$

$$S_q = \{q-1, q+1\} \cap \{0, \ldots, M-1\} \tag{21}$$

Let $[\mathbf{A}]_{ij}$ be the value that matrix $\mathbf{A}$ takes on *i-th* row and *j-th* column. The elements of vector $\mathbf{1}$ are all identical and equal to 1. The solution of (17) can be computed from the singular vector associated to the maximum singular value of matrix Γ, [10]. The size of matrices $\mathbf{D}$ and Ψ is MxM. Consequently, the complexity to find the optimal power distribution crucially depends on the the number of subbands. It is worth emphasizing that cost functions of (17) can be expressed as a ratio of affine functions, which is quasi-convex. Provided that denominators of the cost functions are restricted to be strictly positive the max-min problem can be transformed to an equivalent linear program, [11]. However, this paper does not try to efficiently solve (17). By contrast we propose a sub-optimal strategy that can substantially reduce the complexity burden. This strategy relies on the assumption that the magnitude of the noise is higher than the magnitude of the interferences, i.e. $N_0/2 >> i_q$. Under the assumption that ISI and ICI terms can be neglected the power can be optimally split as follows.

$$p_q = \frac{P_T}{\|h_q\|^2 \sum_{m=0}^{M-1} \|1/h_m\|^2} \tag{22}$$

This solution was originally formulated in the OFDM context, [9]. It must be mentioned that (22) does not balance the SINRs unless $N_0/2 >> i_q$. Hence the system performance crucially depends on ability of the precoders to remove the interferences.

The authors in [10] have proposed an iterative strategy that jointly addresses the power control and the transmit filters design. However until the stopping criterion is not met it is required to compute several times the SVD of Γ which considerably increases the complexity.

4. PRECODING DESIGN

In Section III it has been shown that a proper transmit processing simplifies the power allocation. To that end, precoders have to mitigate the interferences as much as possible. Keeping in mind this criterion two techniques have been addressed in this section: the transmit matched desired impulse response (Tx-MDIR) and the whitening matched filter (WMF).

4.1 Tx-MDIR

The first strategy sets the coefficients of the transmit filters so that the mean square error (MSE) between the analysis filter bank output and the desired symbol is minimized. As (23) shows apart from the broadband transmit processing the desired impulse response (h_q) has to be optimized too.

$$\begin{array}{l} MSE_q = E\left\{\left|\Re\left(z_q[k]\right) - h_q d_q[k]\right|^2\right\} = \left\|h_q\right\|^2 + \\ \frac{N_0}{2} + \sum_{m=q-1}^{q+1} \left\|\mathbf{b}_{m,e}^T \bar{\mathbf{G}}_{qm,e}[k]\right\|^2 - 2h_q \mathbf{b}_{q,e}^T \bar{\mathbf{G}}_{qq,e}[k]\mathbf{e}_l \end{array} \tag{23}$$

With the objective of jointly designing $\mathbf{b}_{q,e}$ and h_q the sum of the MSEs has been selected to be the cost function. The same approach has been considered in [7] to jointly design transmit and receive filters. In order to comply with the power allocation strategy addressed in Section III, precoders are constrained to have unit norm. Thus, the problem can be formulated as follows.

$$\begin{array}{l} argmin_{\{\mathbf{b}_{0,e},h_0,\ldots,\mathbf{b}_{M-1,e},h_{M-1}\}} \sum_{q=0}^{M-1} MSE_q \\ s.t. \ \ \left\|\mathbf{b}_{q,e}\right\|^2 = 1 \quad q = 0, \ldots, M-1 \end{array} \tag{24}$$

Setting the Lagrangian partial derivatives to 0 we obtain the following results. Likewise the SINR formulation the expressions are not dependent on the time instant.

$$h_q = \mathbf{b}_{q,e}^T \bar{\mathbf{G}}_{qq,e}[k]\mathbf{e}_l \tag{25}$$

$$(\mathbf{I}_q)\,\mathbf{b}_{q,e} = \lambda_q \mathbf{b}_{q,e} \tag{26}$$

$$\begin{array}{l} \mathbf{I}_q = \sum_{m=q-1, m\neq q}^{q+1} \bar{\mathbf{G}}_{mq,e}[k]\bar{\mathbf{G}}_{mq,e}^T[k] + \\ \bar{\mathbf{G}}_{qq,e}[k]\left(\mathbf{I} - \mathbf{e}_l\mathbf{e}_l^T\right)\bar{\mathbf{G}}_{qq,e}^T[k] \end{array} \tag{27}$$

Since receive samples are not equalized the noise power is not included in the solution. This may lead to undesired designs that regardless of substantially mitigating the power that leaks into the surrounding symbols might weight the desired symbol with a magnitude on the same order as the noise power. As a solution, among the possible singular vectors that solve (26) we select the one that maximizes the signal to leakage plus noise ratio (SLNR).

$$SLNR_q = \frac{\left\|\mathbf{b}_{q,e}^T \bar{\mathbf{G}}_{qq,e}[k]\mathbf{e}_l\right\|^2}{\lambda_q + N_0/2} \tag{28}$$

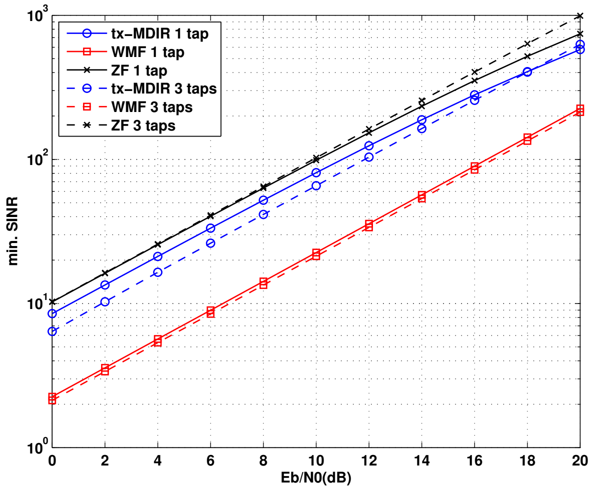

Figure 1: Minimum SINR against Eb/N0. ITU-Veh. A channel model.

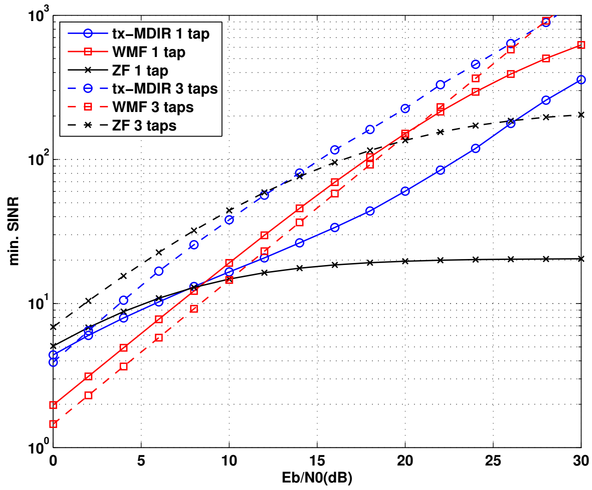

Figure 2: Minimum SINR against Eb/N0. ITU-Veh. B channel model.

4.2 WMF

The whitening matched filter is designed to maximize the signal to leakage ratio (SLR). This criterion aims at providing the optimal signal confinement. Note that this strategy does not take into account the power noise magnitude.

$$SLR_q = \frac{\left\|\mathbf{b}_{q,e}^T \bar{\mathbf{G}}_{qq,e}[k]\mathbf{e}_l\right\|^2}{\left\|\mathbf{b}_{q,e}^T \mathbf{I}_q\right\|^2} \qquad (29)$$

The WMF solution can be formulated as follows:

$$\mathbf{b}_{q,e} = \frac{1}{\alpha_q}\left(\mathbf{I}_q\right)^{-1}\bar{\mathbf{G}}_{qq,e}[k]\mathbf{e}_l \qquad (30)$$

where α_q is selected so that precoders have unit norm, [9].

5. SIMULATION RESULTS

Next, this section evaluates the proposed techniques via simulations. Regarding the parameters of the system we have considered that the transmitter is equipped with 2 antennas and the bandwidth is split into M=512 subbands. Complying with the WIMAX settings only Ma=420 out of the 512 subbands are active. The frequency sampling and the spacing between subcarriers have been set to 5.6MHz and 10.94KHz, respectively. Complex symbols are drawn from a 16QAM constellation. The channel realizations have been generated according to the ITU-Veh.B and ITU-Veh.A models, [12]. In addition to the schemes addressed in Section IV we have also simulated the zero forcing (ZF) described in [6]. The ZF, which was firstly devised in [1], is definetely interesting because it is able to remove the interferences when the channel is low frequency selective.
Figures 1 and 2 depict the minimum SINR under different propagation conditions after having set the transmit power on each band according to the max-min approach. The minimum SINR has been evaluated for different average energy bit to noise ratio (Eb/N0). Taking into account that symbols are generated according to a 16QAM scheme, the average energy bit to noise ratio is defined as $Eb/N_0 = \frac{P_T}{2M_aN_0}$.
In Figure 1, where the power delay profile is generated according to the ITU-VehA channel model, setting one tap per antenna provides virtually the same performance as the multiple tap configuration. In this scenario the original properties of the prototype pulse are nearly restored which means that interferences are practically removed. Therefore, from Figure 1 it can be inferred that the ZF technique presents the highest SNR figure. By contrast, the magnitude of the desired signal drops drastically when the WMF is applied.
Note that simulated techniques behave differently under more challenging scenarios as Figure 2 shows. At low Eb/N0s the ZF with 3 taps exhibits the best performance. Conversely, for moderate and high Eb/N0s the Tx-MDIR with 3 taps per antenna outperforms the rest of techniques.
Figure 2 shows that for ZF and Tx-MDIR schemes the more taps we use the better the performance is. In the WMF case the more taps are set the lower the power that leaks into the adjacent symbols is but at the expenses of degrading the SNR, thus it is not clear which is the best configuration.
Finally in Figures 3 and 4 we have plotted the BER. As a benchmark we have simulated the ZF without implementing any power allocation strategy. In other words, the power has been uniformly distributed among the subcarriers. When the channel can be assumed flat in the subcarrier pass band region both the ZF and the Tx-MDIR are able to take advantage of the solution computed by the max-min SINR problem. A BER equal to 10^{-4} can be achieved by saving 3dB in the transmitted power with respect to the uniform power allocation strategy.
For highly frequency selective channels the BER plots of ZF and Tx-MDIR exhibit an error floor since the interference terms are not completely removed. Since at low Eb/N0s the noise power dominates over the interferences the technique that weights the desired symbols with the highest coefficient achieves the best results. On the contrary, are the interferences that dominate over the noise at high Eb/N0s. For this reason, the poor interference rejection exhibited by the ZF leads to saturation. Although the Tx-MDIR also presents an

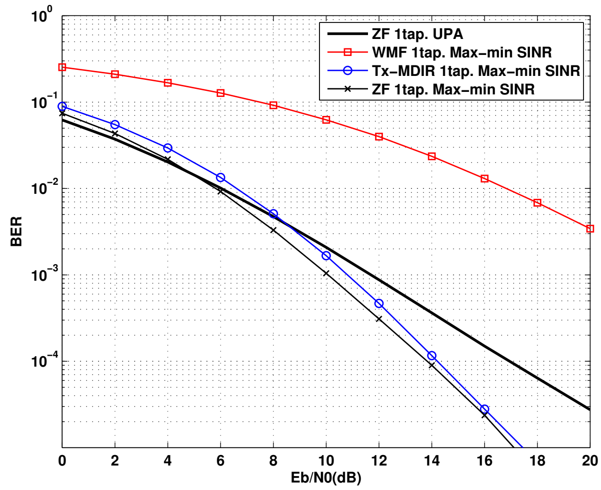

Figure 3: BER against Eb/N0. ITU-Veh. A channel model.

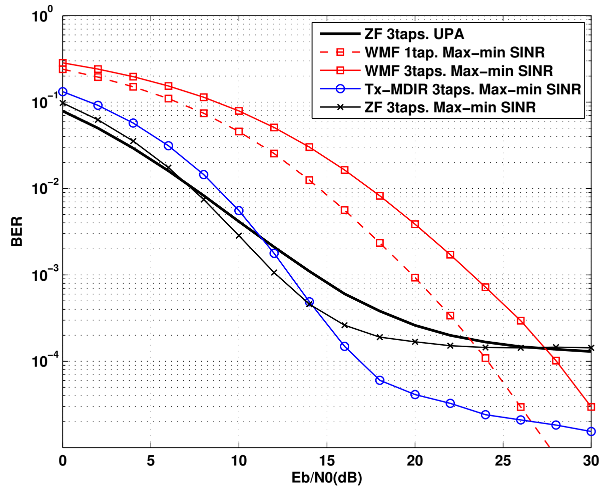

Figure 4: BER against Eb/N0. ITU-Veh. B channel model.

error floor this techniques proves to be more efficient in mitigating the interferences. As a result, the Tx-MDIR is able to boost the BER performance in the Eb/N0 range [15dB-30dB]. As for the WMF no error floor is exhibited. However, as Figure 4 depicts, the WMF technique is only able to outperform the ZF in the Eb/N0 range [24dB-30dB]. At low and moderate Eb/N0s this technique provides poor results since the noise is not included in the cost function to be optimized. It is worth emphasizing that space-time processing should be disregarded in favour of pure space processing when the channel cannot be modelled flat at each subcarrier.

6. CONCLUSION

This paper has tackled the power allocation problem for FBMC-MISO systems. Following a two-step approach precoders and optimal power weights have been computed independently. With the aim of reducing the complexity burden it has been assumed that transmit filters completely remove the interferences. Under this assumption the power distribution among subcarriers has been computed by solving the max-min SINR problem. The simulations have shown that the Tx-MDIR and WMF techniques are able to take advantage of the max-min SINR solution to increase the robustness in a so challenging scenario that discards the channel flatness assumption. This work sets path to gain more knowledge on the behaviour of different spatial precoding techniques for FBMC. It also can be considered a starting point to devise spatial broadcasting strategies for multicarrier systems different from OFDM.

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