ITERATIVE JOINT TONE-INTERFERENCE CANCELLATION AND DECODING FOR MIMO-OFDM

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ABSTRACT

In this paper, multi-input multi-output orthogonal frequency-division multiplexing (MIMO-OFDM) is considered under the effects of both multipath fading and multi-tone interference (MTI). The MIMO-OFDM system makes joint use of channel and orthogonal space-frequency block coding (OSFBC) on the transmit side and iterative processing on the receiver side for robustness and improved performance against the fading and MTI effects of the channel. The new iterative receiver is implemented by either an optimal a posteriori probability (APP) space-frequency detector (SFD) or a soft information-aided minimum mean-squared error (MMSE) combiner at its front-end and a soft input-soft output (SISO) channel decoder at its back-end. Both receivers are compared in terms of their computational complexities and bit-error rate (BER) performances. Both iterative receivers provide a improvement in performance after only a few detection/decoding iterations. Also, despite its suboptimality, the MMSE receiver achieves a BER performance close to that of the APP detector at a significantly lower cost.

1. INTRODUCTION

Orthogonal frequency-division multiplexing (OFDM) is an efficient technology emerging in growing number of applications within terrestrial digital video broadcasting (DVB), wireless local area networks (LANs), 4G cellular systems, high bit-rate digital subscriber lines (HDSLs), asymmetric digital subscriber loops (ADSLs) and next-generation tactical communication systems. In many cases, OFDM-based communication systems are subject to both the fading effects of the channel and narrowband interference (NBI), which may arise in various forms and is often modelled as multi-tone interference (MTI) [1]. For this reason, as the range of OFDM applications grows so does the importance of designing interference and fading resistant OFDM transceivers.

One way to provide interference mitigation capability is to employ multiple transmit and/or receive antennas and to incorporate orthogonal space-frequency coding (OSFBC) to the OFDM transceiver [2]. This approach relies on the premise that not all subcarriers in OSFBC are hit by the interference so that the post-decoding signal-to-interference plus noise ratio (SINR) is still high enough to achieve good bit-error rates (BERs). Notice that, while the primary reason for having SFC is fading, the robustness against MTI comes as a side benefit which cannot necessarily be achieved through space-time block coding as pointed out by [3]. The interference rejection performance is analyzed and presented for an Alamouti coded OFDM system in [4] under the effects of frequency-selective fading and MTI.

Despite its advantages, both because OSFBC is essentially a precaution taken at the transmitter and because conventionally it uses a simple ML space-frequency decoder that is interference unaware, the full potential of OSFBC-OFDM is never fully utilized. To alleviate this problem, in this paper we consider a MIMO-OFDM system which combines the advantages of concatenated channel and SFC at the transmitter side and iterative processing on the receiver side. The receiver of the proposed system consists of a front-end soft-input soft-output (SISO) detector used for joint interference suppression and space-frequency decoding (SFD) that is coupled with a back-end channel decoder for iterative processing. The front-end module is implemented first by the optimal *a posteriori* probability (APP) detection rule taking MTI structure into account. Then as a lower complexity alternative, a minimum mean-squared error (MMSE) detector is also proposed for the same task. However, because the MMSE detector alone is not able to process or generate soft information, it is combined with an *a priori* soft interference canceller and an *a posteriori* probability mapper so as to make it suitable for turbo processing. The two iterative receivers are compared in terms of the computational complexity and BER performance with respect to various signal-to-noise (SNR), signal-to-interference (SIR) ratios and the fractions of interfered bandwidth. The proposed system is also compared with a coded MIMO-OFDM system employing just spatial multiplexing and not OSFBC so as to evaluate the interference mitigation ability of OSFBC. The simulations show that both receivers provide a significant improvement in the interference mitigation and diversity performance after a few decoding iterations in all transmission scenarios. Moreover the BER performance of the MMSE receiver comes within 1 dB (in both SIR and SNR) to that of the APP receiver at a significantly lower cost.

Notice that several other turbo receiver architectures have been proposed in the context of space-time/space-frequency coded OFDM systems. In [5], a MIMO system employing spatial multiplexing and subject to MUI is considered and an iterative receiver with joint *a posteriori* probability (APP) detection and MUI cancellation is proposed. In [6], an low-density parity-check (LDPC) based space-time coded OFDM system is investigated and an iterative APP-based receiver is developed for joint detection and decoding. A similar APP receiver is presented for iterative decoding of turbo space-frequency coded OFDM in [7]. Then, in [8] an iterative receiver based on linear filtering is proposed for a bit-interleaved coded MIMO-OFDM system as a lower complexity alternative to the APP decoding based receivers above. Among these works, [6, 7, 8] disregard the impact of interference on the coded MIMO-OFDM system whereas [5] only considers an uncoded MIMO system with only spatial multiplexing. On the other hand, in this paper, the performance issues for OSFBC-OFDM is addressed under multipath fading and MTI and iterative receivers are proposed with interference aware SISO space-frequency detectors.

2. SYSTEM MODEL

We consider an OSFBC-OFDM system with N_t transmit and N_r receive antennas, all assumed to be uncorrelated. The elements of a bit sequence, $u_i \in \{0,1\}$, for $i = 1,\ldots,L_u$, are encoded by a convolutional encoder to form the coded bit sequence with elements $c_j \in \{1,-1\}$ for $j = 1,\ldots,L_c$. This sequence is then bitwise interleaved to form the sequence with elements d_j for $j = 1,\ldots,L_c$ that is partitioned into groups of m_o bits which are mapped onto the complex symbols of an M-ary PSK or QAM symbol alphabet $\mathscr{A} = \{A_1, A_2, \ldots, A_M\}$ where $M = 2^{m_o}$. The resulting com-

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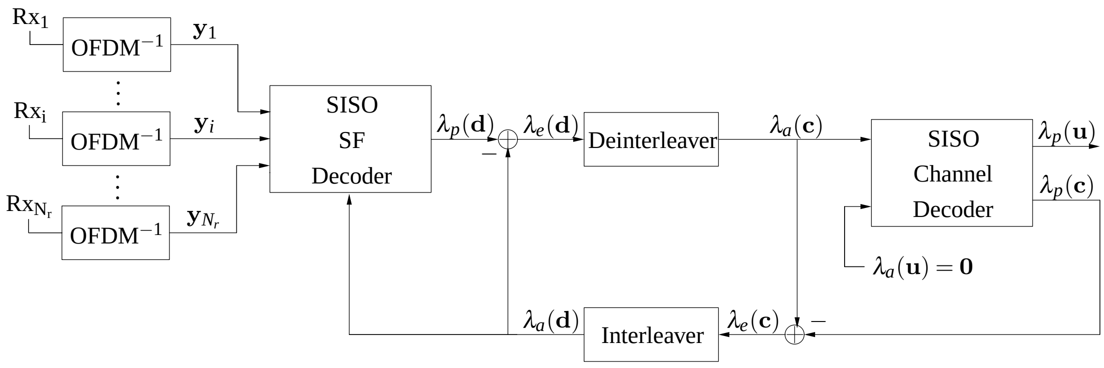

Figure 1: General block diagram of the iterative space-frequency receiver.

plex symbol sequence with elements a_k for $k = 1,\ldots,L_a$ is parsed into P blocks of length k_o where $L_a = k_o P$ and in vector form $\mathbf{a}_p = \left[a_{k_o p+1}\ a_{k_o p+2}\ \cdots\ a_{k_o(p+1)}\right]^T$ for $p = 0,\ldots,P-1$. Each block $\mathbf{a}_p$ is then encoded by a rate-k_o/n_o OSFB encoder employing a size $n_o \times N_t$ block matrix $\mathscr{G}$ to form the length-n_o code symbol blocks $\mathbf{x}_{i,p} = \left[x_{i,n_o p+1}\ x_{i,n_o p+2}\ \cdots\ x_{i,n_o(p+1)}\right]$ at each transmit branch where $i = 1,\ldots,N_t$. Considering that there are P encoded blocks altogether, the OSFB code block at the ith antenna is represented as $\mathbf{x}_i = [\mathbf{x}_{i,0}\ \ldots\ \mathbf{x}_{i,p}\ \ldots\ \mathbf{x}_{i,P-1}]$. Finally each of the N_t transmit antenna sequences is OFDM modulated with $N = n_o P$ subcarriers and transmitted over independent channels.

Assuming a single interference source, the received discrete signal model after OFDM demodulation at the jth receive antenna and the nth subband is

$$y_{j,n} = \sum_{i=1}^{N_t} h_{j,i,n} x_{i,n} + \gamma_{j,n} g_{j,n} b_{j,n} + \nu_{j,n}, \qquad n = 1,\ldots,N, \qquad (1)$$

where $h_{j,i,n}$ is the channel fading coefficient between the ith transmit and the jth receive antenna, $x_{i,n}$ is nth symbol transmitted from the ith transmit antenna, $g_{j,n}$ is the complex channel gain between the interference source and the jth receive antenna, $b_{j,n}$ is the interference signal observed by the jth receive antenna, $\gamma_{j,n}$ is an interference indicator function (i.e. $\gamma_{j,n}$ equals to 1 when the n-th subband is interfered and 0 otherwise.), and $\nu_{j,n}$ represents the zero-mean, complex, additive white Gaussian noise (AWGN) with two-sided power spectral density of $N_0/2$.

The signal model in (1) can be expressed in block form and in terms of the OSFB encoder input symbols, a_k, rather than the coded symbols, $x_{i,n}$. This can be done by first parsing the received OFDM block at each antenna into P length-n_o subblocks and by defining

$$\begin{aligned}
\mathbf{y}_{j,p} &= \left[y_{j,n_o p+1}\quad y_{j,n_o p+2}\quad \cdots\quad y_{j,n_o(p+1)}\right],\\
\mathbf{b}_{j,p} &= \left[b_{j,n_o p+1}\quad b_{j,n_o p+2}\quad \cdots\quad b_{j,n_o(p+1)}\right],\\
\boldsymbol{\nu}_{j,p} &= \left[\nu_{j,n_o p+1}\quad \nu_{j,n_o p+2}\quad \cdots\quad \nu_{j,n_o(p+1)}\right],
\end{aligned}$$

as the vectors collecting the noisy channel observations, interference tones and noise symbols, respectively, for the pth block ($p = 0,\ldots,P-1$) and jth receive antenna. Then, collecting all receive antenna signals in

$$\begin{aligned}
\mathbf{y}_p &= \left[\mathbf{y}_{1,p}\quad \ldots\quad \mathbf{y}_{j,p}\quad \ldots\quad \mathbf{y}_{N_r,p}\right]^T,\\
\mathbf{b}_p &= \left[\mathbf{b}_{1,p}\quad \ldots\quad \mathbf{b}_{i,p}\quad \ldots\quad \mathbf{b}_{N_r,p}\right]^T,\\
\boldsymbol{\nu}_p &= \left[\boldsymbol{\nu}_{1,p}\quad \ldots\quad \boldsymbol{\nu}_{i,p}\quad \ldots\quad \boldsymbol{\nu}_{N_r,p}\right]^T,
\end{aligned}$$

respectively, and using the OSFBC relation between $\mathbf{a}_p$ and $\mathbf{x}_{i,p}$, the pth receive block can be expressed as

$$\mathbf{y}_p = \mathbf{H}_p \mathbf{a}_p + \boldsymbol{\Gamma}_p \mathbf{G}_p \mathbf{b}_p + \boldsymbol{\nu}_p. \qquad (2)$$

Here $\mathbf{H}_p$ is the channel submatrix whose structure is dependent on the chosen block code $\mathscr{G}$; specifically, e.g.,

$$\mathbf{H}_p = \begin{bmatrix}
h_{1,1,8p+1} & h_{1,2,8p+1} & h_{1,3,8p+1} & 0\\
h_{1,2,8p+2} & -h_{1,1,8p+2} & 0 & -h_{1,3,8p+2}\\
h_{1,3,8p+3} & 0 & -h_{1,1,8p+3} & h_{1,2,8p+3}\\
0 & h_{1,3,8p+4} & -h_{1,2,8p+4} & -h_{1,1,8p+4}\\
h^*_{1,1,8p+5} & h^*_{1,2,8p+5} & h^*_{1,3,8p+5} & 0\\
h^*_{1,2,8p+6} & -h^*_{1,1,8p+6} & 0 & -h^*_{1,3,8p+6}\\
h^*_{1,3,8p+7} & 0 & -h^*_{1,1,8p+7} & h^*_{1,2,8p+7}\\
0 & h^*_{1,3,8(p+1)} & -h^*_{1,2,8(p+1)} & -h^*_{1,1,8(p+1)}\\
\vdots & \vdots & \vdots & \vdots\\
h_{N_r,1,8p+1} & h_{N_r,2,8p+1} & h_{N_r,3,8p+1} & 0\\
\vdots & \vdots & \vdots & \vdots\\
0 & h^*_{N_r,3,8(p+1)} & -h^*_{N_r,2,8(p+1)} & -h^*_{N_r,1,8(p+1)}
\end{bmatrix},$$

for $\mathscr{G}_3$ in [9], and $\mathbf{G_p}$ and $\boldsymbol{\Gamma}_\mathbf{p}$ are the diagonal matrices holding the channel and interference indicator coefficients of the interferer, respectively. Notice that each of the P subblocks are detected and decoded separately. Therefore, in the sequel, we will limit ourselves to the decoding of one subblock and drop the subscript p for simplicity.

3. ITERATIVE SPACE-FREQUENCY RECEIVER

The block diagram of the proposed iterative receiver is shown in Fig. 1 where, following the OFDM demodulation, joint interference cancellation/SFD and decoding iteration is performed by a soft information exchange between the SISO space-frequency decoder and the channel decoder. For front-end operation we propose two alternative approaches, one based on the optimal symbol-by-symbol APP detection rule, and the other on the joint use of soft interference cancellation (SIC), MMSE combining and probability mapping. The back-end SISO channel decoder is implemented by the BCJR algorithm, which is well-known in the turbo processing literature. For this reason, we skip the details of this algorithm and focus only on the operations of the APP- and SIC/MMSE-based SISO blocks and the comparison of their computational complexity.

3.1 APP-Based Receiver

Note from the signal model in Section II, every complex symbol a_k corresponds to an m_o-bit codeword, i.e., $a_k \leftarrow \{d_{k,1}\ \ldots\ d_{k,m}\ \ldots\ d_{k,m_o}\}$ where $d_{k,m} = d_{(k-1)m_o+m}$ for $k = 1,\ldots,L_a$ and $m = 1,\ldots,m_o$. Then, considering only the first decoding block (indexed by $p = 0$), a symbol-by-symbol APP detector processing the noisy observation block $\mathbf{y}$ in (2) computes the *a posteriori* bit

LLR values on each $d_{k,m}$ for $k = 1,\ldots,k_o$ and $m = 1,\ldots,m_o$ by

$$\lambda_p(d_{k,m}) = \log \frac{p(d_{k,m}=1 \mid \mathbf{y})}{p(d_{k,m}=-1 \mid \mathbf{y})} = \log \frac{\sum_{\forall \mathbf{a}:(d_{k,m}=1)} \exp\left(\mathscr{K}(\mathbf{a}) + \sum_{k=1}^{k_o} \log P(a_k)\right)}{\sum_{\forall \mathbf{a}:(d_{k,m}=-1)} \exp\left(\mathscr{K}(\mathbf{a}) + \sum_{k=1}^{k_o} \log P(a_k)\right)} \quad (3)$$

where the summations in the numerator and the denominator are over all possible $\mathbf{a}$ blocks for which $d_{k,m}$ is 1 or -1, respectively, $P(a_k)$'s are the *a priori* symbol probabilities for symbols a_k and $\mathscr{K}(\mathbf{a})$ is the maximum likelihood distance metric formed by the channel observations as

$$\mathscr{K}(\mathbf{a}) = -\frac{1}{2}(\mathbf{y}-\mathbf{Ha})^H \left(\sigma^2\mathbf{I}+\mathbf{G\Gamma G}^H\right)^{-1}(\mathbf{y}-\mathbf{Ha}) \quad (4)$$

where H denotes the Hermitian transpose. Assuming that the successive bits are independent, the expression in (3) can be rewritten as

$$\lambda_p(d_{k,m}) = \lambda_a(d_{k,m}) + \lambda_e(d_{k,m}) \quad (5)$$

where $\lambda_a(d_{k,m}) = \log \frac{P(d_{k,m}=1)}{P(d_{k,m}=-1)}$ is the *a priori* bit LLR from the back-end channel decoder and

$$\lambda_e(d_{k,m}) = \log \frac{\sum_{\forall \mathbf{a}:(d_{k,m}=1)} \exp\left(\mathscr{K}(\mathbf{a}) + \sum_{(l,q)\neq(k,m)} \frac{1}{2} d_{l,q}\lambda_a(d_{l,q})\right)}{\sum_{\forall \mathbf{a}:(d_{k,m}=-1)} \exp\left(\mathscr{K}(\mathbf{a}) + \sum_{(l,q)\neq(k,m)} \frac{1}{2} d_{l,q}\lambda_a(d_{l,q})\right)} \quad (6)$$

is the extrinsic bit LLR to be passed onto the back-end channel decoder.

Note that in order to reduce the computational complexity and to avoid numerical instability, the extrinsic LLR in (6) can be computed by

$$\lambda_e(d_{k,m}) = \max{}^*_{\forall \mathbf{a}:(d_{k,m}=1)} \left[\mathscr{K}(\mathbf{a}) + \sum_{(l,q)\neq(k,m)} \frac{1}{2} d_{l,q}\lambda_a(d_{l,q})\right] - \max{}^*_{\forall \mathbf{a}:(d_{k,m}=-1)} \left[\mathscr{K}(\mathbf{a}) + \sum_{(l,q)\neq(k,m)} \frac{1}{2} d_{l,q}\lambda_a(a_{l,q})\right] \quad (7)$$

where $\max^*[\cdot,\cdot]$ operation is defined as $\max^*[\mathrm{x},\mathrm{y}] = \max[\mathrm{x},\mathrm{y}] + \log\left(1+\mathrm{e}^{-|\mathrm{x}-\mathrm{y}|}\right)$.

3.2 SIC/MMSE-Based Receiver

A lower complexity alternative to APP-based space-frequency decoding is to employ MMSE combining at the front-end. However, because the MMSE combiner alone is not capable of using or generating soft information, it has to be combined with an *a priori* soft interference canceller (SIC) and an *a posteriori* probability mapper so as to make it suitable for soft information exchange and thus for turbo processing. This approach is similar to the linear filtering techniques used for the turbo equalization in [10] and for the space-frequency equalization for bit-interleaved coded modulation (BICM) on MIMO-OFDM in [8], with the distinction of taking into account the OSFBC and the MTI.

The proposed receiver first uses the *a priori* symbol expectations collected in the vector $\bar{\mathbf{a}} = \left[\bar{a}_1 \ldots \bar{a}_k \ldots \bar{a}_{k_o}\right]^T$ where each symbol expectation is computed using the *a priori* symbol probabilities from the channel decoder, i.e., $\bar{a}_k = \sum_{q=1}^{M} A_q P(a_k = A_q)$. Then, the expected interference for the kth symbol is expressed as

$$\bar{\mathbf{y}}_k = \mathbf{H}(\bar{\mathbf{a}} - \bar{a}_k\mathbf{e}_k), \quad (8)$$

which is subtracted from the channel observation vector $\mathbf{y}$ to form the vector with reduced interference

$$\tilde{\mathbf{y}}_k = \mathbf{y} - \bar{\mathbf{y}}_k = \mathbf{H}(\mathbf{a} - \bar{\mathbf{a}} + \bar{a}_k\mathbf{e}_k) + \mathbf{\Gamma G b} + \boldsymbol{\nu} \quad (9)$$

where $\mathbf{e}_k$ is a length-k_o all-zero column vector except for its kth entry which is equal to 1.

The vector in (9) is used as input to the combiner whose coefficient vector at time k, $\mathbf{w}_k = \left[w_{k,1}\ w_{k,2}\ \ldots\ w_{k,N_r n_o}\right]^T$, is chosen to minimize the MSE between the combiner output and the data symbol a_k:

$$J_{\mathrm{MSE}}(\mathbf{w}_k) = E\left[|a_k - \mathbf{w}_k^H \tilde{\mathbf{y}}_k|^2\right]. \quad (10)$$

The optimum solution that minimizes the cost function in (10) is

$$\mathbf{w}_{k,\mathrm{opt}} = \left(\sigma^2\mathbf{I} + r_b\mathbf{G\Gamma G}^H + \mathbf{HR}_a\mathbf{H}^H + |\bar{a}_k|^2\mathbf{hh}^H\right)^{-1}\mathbf{h}_k r_a$$

where $\mathbf{h}_k = \mathbf{He}_k$, $\mathbf{R}_a$ is the diagonal covariance matrix

$$\mathbf{R}_a = \mathrm{E}\left[(\mathbf{a}_k - \bar{\mathbf{a}}_k)(\mathbf{a}_k - \bar{\mathbf{a}}_k)^H\right] \quad (11)$$

and r_a and r_b are the average powers of the transmitted signal and the interferer, respectively. Once the coefficient vector $\mathbf{w}_k$ is evaluated, the combiner output is obtained by

$$z_k = \mathbf{w}_k^H(\mathbf{y}_k - \mathbf{H}\bar{\mathbf{a}}_k + \bar{a}_k\mathbf{h}_k). \quad (12)$$

In this respect, note that the information contained in z_k is extrinsic, since the a priori information about the desired symbol a_k at time k is left out of the soft cancellation process.

The purpose of the SIC/MMSE combiner is to serve as part of a turbo-type detector/decoder structure. This means that the extrinsic information in the form of either M-ary *a posteriori* symbol probabilities or binary LLRs needs to be extracted from the combiner output sequence. As in [8] and [10], this information is produced by viewing the combiner output sequence as produced by an AWGN channel with input a_k, i.e.,

$$z_k = \mu_k a_k + n_k \quad (13)$$

where μ_k is the channel gain and n_k is complex white Gaussian noise with zero mean and variance σ_k^2. This is equivalent to saying that z_k admits a complex Gaussian distribution, i.e., $z_k \sim \mathscr{N}(\mu_k a_k, \sigma_k^2)$. The parameters μ_k and σ_k^2 are calculated at each time instant using the combiner structure as $\mu_k = \mathbf{w}_k^H\mathbf{h}_k r_a$, $\sigma_k^2 = \mathbf{w}_k^H\mathbf{h}_k\left(1-\mathbf{h}_k^H\mathbf{w}_k\right)r_a$. Once μ_k and σ_k^2 are computed, the APP bit LLRs $\lambda_p(d_{k,m})$ for $k = 1,\ldots k_o$ and $m = 1,\ldots,m_o$ (again considering the first decoding block only) can be expressed as in (5) with $\lambda_a(d_{k,m})$ is as defined before and the extrinsic bit LLR

$$\lambda_e(d_{k,m}) = \log \frac{\sum_{\forall a_k:(d_{k,m}=1)} \exp\left(\tilde{\mathscr{K}}(\mathbf{a}) + \sum_{(l,q)\neq(k,m)} \frac{1}{2} d_{l,q}\lambda_a(d_{l,q})\right)}{\sum_{\forall a_k:(d_{k,m}=-1)} \exp\left(\tilde{\mathscr{K}}(\mathbf{a}) + \sum_{(l,q)\neq(k,m)} \frac{1}{2} d_{l,q}\lambda_a(d_{l,q})\right)}$$

where $\tilde{\mathscr{K}}(a_k) = -\frac{|z_k-\mu_k a_k|^2}{2\sigma^2}$. Notice that similar to its APP counterpart, $\lambda_e(a_{k,m})$ can be computed with the $\max^*[\cdot,\cdot]$ operation.

3.3 Complexity Analysis

We also present a comparison of the computational load of both proposed receivers over the decoding of a single block $\mathbf{a}$. For both receivers, the first iteration needs less effort because of the lack of *a priori* information. It is sufficient to compute $\mathscr{K}(\cdot)$ for the APP-based SISO receiver. Likewise, additional load due to the computation of the expected symbols and the covariance matrix appears in the second and further iterations for the MMSE-based receiver. Tables 1 and 2 summarize the computational complexity analysis.

Table 1: Computational Complexity of the APP Receiver

APP	1^{st} iteration	Other iterations
ADD	$sk_oW_1^{\dagger}$	$sk_oW_1'^{\dagger}$
MUL	$sk_oM_1^{\ddagger}$	$sk_oM_1'^{\ddagger}$
$\max^*[\cdot,\cdot]$	$(2^{k_os-1}-1)2sk_o$	$(2^{k_os-1}-1)2sk_o$

$^{\dagger}W_1 = 2^{k_om_o}n_oN_r(k_o+1)$, $W_1' = 2^{k_om_o}(n_oN_r(k_o+1)+2(m_ok_o-1))$
$^{\ddagger}M_1 = 2^{k_om_o}n_oN_r(k_o+3)$, $M_1' = 2^{k_om_o}(n_oN_r(k_o+3)+2m_ok_o)$

Table 2: Computational Complexity of the SIC/MMSE Receiver

MMSE	1^{st} iteration	Other iterations
ADD	$\phi = k_o(W_2+2n_oN_r+s2^{s+1})^{\diamond}$	$\phi + k_o(n_oN_r+2^s+2s)$
MUL	$\psi = k_o((M_2+2n_oN_r+2)^{\diamond}$ $+s(2^{s+2}+2s+1))$	$\psi + k_o(2^{s+1}+2+s)$
EXP	sk_o2^s	$sk_o(2^s+1)$
LOG	sk_o	sk_o
INV	k_o	k_o

$^{\diamond}W_2 = n_oN_r(k_o(k_o-1)+n_oN_r(k_o+2)-1), M_2 = n_oN_r(n_oN_r(k_o+1)+k_o^2)$

4. SIMULATION RESULTS

The proposed iterative receivers for the OSFBC-OFDM transmitter are simulated in the presence of multipath fading and MTI. The transmitted bits are encoded by a rate-$1/2$ convolutional code with memory 4 and the generator $(23,\,35)_8$ in octal representation. The coded bits are mapped onto complex QPSK symbols and a block code matrix $\mathscr{G}_3$ in [9] is used for OSFB encoding to distribute the QPSK symbols onto 3 transmit antennas. The simulation parameters are as follows: $M = 4$, $m_o = 2$, $N_t = 3$, $k_o = 4$, $n_o = 8$. On all transmit branches OFDM modulation is performed with $N = 128$ subcarriers. The frequency-selective multipath channels are modeled by 6-tap finite impulse response (FIR) filters with complex-Gaussian distributed coefficients and exponentially decaying power delay profile. The cyclic prefix length is set to 7 symbols so that the intersymbol interference effects are avoided. The estimation of the channel coefficients or interferer power which can be incorporated into the iterative receiver is omitted and all channel state information is assumed to be perfectly known to the receiver. The SNR and SIR are defined as SNR $= E_s/N_0$ and SIR $= E_s/N_j$, respectively where N_j is the average interferer power. The fraction of interfered bandwidth is denoted with ρ. In all simulations, the iterative receivers are stopped after the 5 detection/decoding iterations beyond which only a marginal gain is observed.

The performances of both the APP- and SIC/MMSE-based iterative receivers at an SIR of -8 dB are shown in Fig. 2 as a function of the SNR. As noted from the figure both sytems offer an improvement of two orders of magnitude between their first and last iterations at SNR = 10 dB. Furthermore after the last iterations, the performance of the SIC/MMSE-based iterative receiver is only 0.9 dB away from that of the APP-based iterative receiver at 1×10^{-4} BER level, where the former requires approximately 7.2% of the operational cost of the latter (i.e. computed for the considered system parameters through Tables 1 and 2).

A similar comparison is shown in Fig. 3, this time with respect to SIR at a fixed SNR of 10 dB. As observed at 10^{-4} BER level, iterative processing for OSFBC-OFDM provides a gain of 7 dB SIR improvement in 5 iterations for both receivers. Besides its reduced computational complexity, the MMSE-based receiver has only a 0.7-dB SIR loss with respect to the APP-based ones. Note that even at low SIRs the first iteration offers a significant improvement with both receivers in contrast to the behavior in Fig. 2.

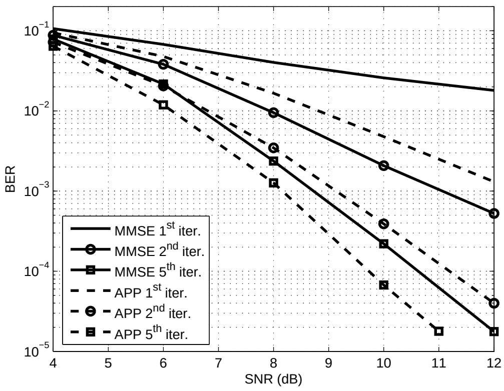

Figure 2: BER vs. SNR performance of the proposed receivers. ($N_r = 1$, $SIR = -8$ dB, $\rho = 0.75$).

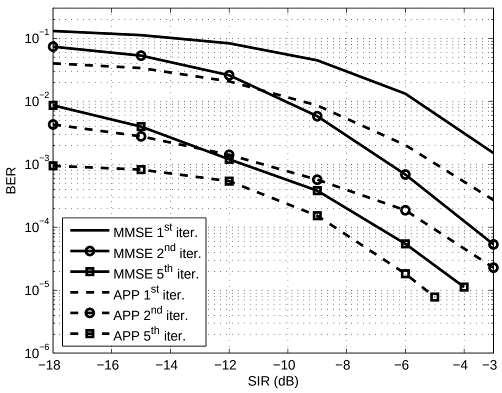

Figure 3: BER vs. SIR performance of the proposed receivers. ($N_r = 1$, $SNR = 10$ dB, $\rho = 0.75$).

The effect of the fraction of the interfered bandwidth (ρ) on the iterative receiver performance is shown in Fig. 4 for SNR = 10 dB and SIR = - 6 dB. Although not depicted in the figure, as the SIR increases, the BER improvement obtained by each iteration grows as well for large ρ values such as 0.9 or 1. Another important observation is the continuous performance improvement of the iterations through small ρ values. Since the total interference power is fixed, as ρ increases, the effective interference power per subcarrier is reduced proportionally, and this leads to a trade-off between the number of interfered subcarriers versus the tone power per subcarrier in [4], where an Alamouti-based non-iterative SFC-OFDMA system is considered. In contrast, we do not observe any trade-off with the proposed MMSE-based system. As ρ increases, the randomness and the rareness of the tones decrease, which leads to more contiguously-hit subcarriers. Because the outer SISO decoder's working principle relies on sequence-based correction, when long series of contiguously corrupted bit information arrives at the SISO module, it cannot succeed in correcting them, especially in the low SIR regime.

To evaluate the OSFBC gain in a MIMO-OFDM system, the MMSE-based iterative receiver is also simulated for a MIMO transmitter employing only spatial multiplexing, and comparisons with respect to the SNR and SIR are presented in Fig. 5 and Fig. 6, respectively. In Fig. 5, the BER curves for both OSFBC-OFDM and MIMO-OFDM are shown for the first and fifth iterations, and

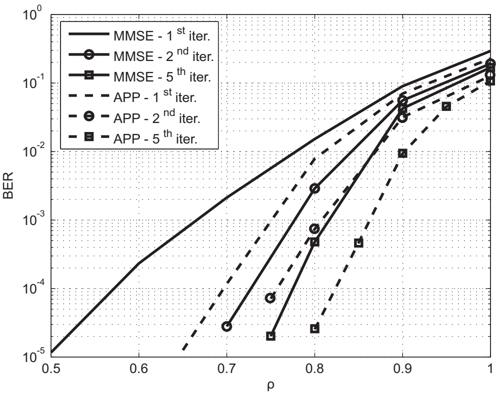

Figure 4: BER vs. ρ performance of the proposed receivers ($N_r = 1$, $SNR = 10$ dB, $SIR = -6$ dB).

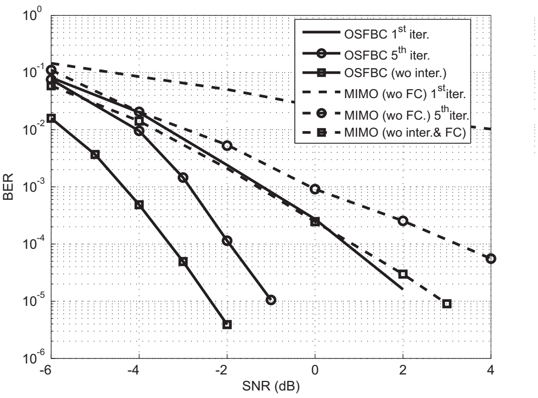

Figure 5: BER vs. SNR performance of the proposed receivers with and without OSFBC ($N_r = 3$, $SIR = -5$ dB, $\rho = 0.8$).

interference-free reference curves (the first iterations) are drawn for comparison. For SIR$= -5$ dB, the OSFBC-OFDM system is subject to a performance degradation of 4 dB in SNR at a BER of 10^{-4}. The MIMO-OFDM system employing spatial multiplexing has a loss of more than 8 dB to the interference although this is not explicitly depicted in the figure. This result agrees with the interference mitigation capability of space-frequency coding as demonstrated in [4]. Moreover, the space-frequency coded system approaches within 1 dB of the interference-free performance in the fifth iteration, whereas the same quantity is almost 2 dB for the uncoded system. Fig. 6 shows the BER vs. SIR performances of both systems. It is observed that frequency coding enables the system to resist 4 dB more SIR for a BER of 10^{-4} at the fifth iteration. Applying an iterative process makes the uncoded system also more robust to interference, albeit its performance in the fifth iteration can only reach the performance of the OSFBC system's first iteration.

5. CONCLUSION

In this paper, we presented two iterative SF detection/decoding architectures for a MIMO OSFBC-OFDM system subject to frequency-selective fading and MTI. The two iterative receivers differ from each other in their front-end detector modules, where one is implemented by an APP detection algorithm and the other by a combination of the SIC/MMSE combining and probability mapping operations. It is demonstrated with simulations that combining the advantages of OSFBC and iterative processing provides not only increased robustness against MTI but also a significant gain in terms

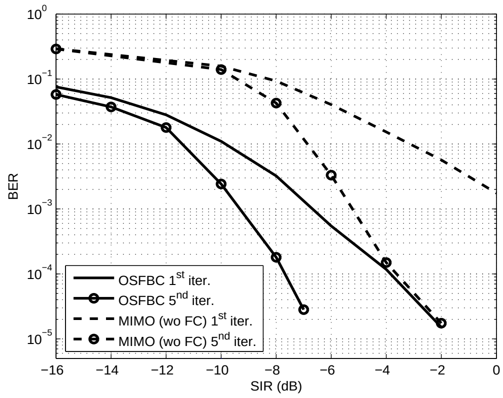

Figure 6: BER vs. SIR performance of the proposed receivers with and without OSFBC. ($N_r = 3$, $SIR = -5$dB, $\rho = 0.8$).

of both the SNR and SIR. It is further noted that the MMSE-based receiver achieves performance comparable to that of an APP-based receiver, only with at a significantly lower cost.

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