Turbo Packet Combining for Broadband Space-Time BICM Hybrid-ARQ Systems with Co-Channel Interference

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Abstract—In this paper, efficient turbo packet combining for single carrier (SC) broadband multiple-input–multiple-output (MIMO) hybrid–automatic repeat request (ARQ) transmission with unknown co-channel interference (CCI) is studied. We propose a new frequency domain soft minimum mean square error (MMSE)-based signal level combining technique where received signals and channel frequency responses (CFR)s corresponding to all retransmissions are used to decode the data packet. We provide a recursive implementation algorithm for the introduced scheme, and show that both its computational complexity and memory requirements are quite insensitive to the ARQ delay, i.e., maximum number of ARQ rounds. Furthermore, we analyze the asymptotic performance, and show that under a sum-rank condition on the CCI MIMO ARQ channel, the proposed packet combining scheme is not interference-limited. Simulation results are provided to demonstrate the gains offered by the proposed technique.

Index Terms—Automatic repeat request (ARQ) mechanisms, multiple-input–multiple-output (MIMO), single carrier (SC), unknown co-channel interference (CCI), intersymbol interference (ISI), frequency domain methods.

I. INTRODUCTION

SPACE-TIME-BIT-INTERLEAVED coded modulation (ST–BICM) with iterative decoding is an attractive signaling scheme that offers high spectral efficiencies over multiple-input–multiple-output (MIMO)-intersymbol interference (ISI) channels [1]–[5]. To combat ISI in single carrier (SC) broadcast ST–BICM transmission, frequency domain equalization, initially introduced for single antenna systems [6]–[9], has been proposed using iterative (turbo) processing [10]. It is a receiver scheme that allows high ISI cancellation capability at an affordable complexity cost. In practical systems, unknown co-channel interference (CCI) caused by other transmitters (distant users and/or neighboring cells) who simultaneously use the same radio resource can dramatically degrade the link performance. This limitation can be overcome by using the so-called hybrid–automatic repeat request (ARQ) protocols, where channel coding is combined with ARQ [11], [12]. In hybrid–ARQ, erroneous data packets are kept in the receiver and used to detect/decode the retransmitted frame [13]–[19].

This technique is often referred to as “packet combining”. Practical packet combining schemes have been addressed in [20]. In [21], an elegant information-theoretic framework has been introduced to analyze the throughput and delay of hybrid–ARQ under random user behavior. Interestingly, the authors have shown that hybrid–ARQ systems are not interference limited, i.e., arbitrarily high throughput can be achieved by simply increasing the transmit power of all users even when multi-user detection (MUD) techniques are not used at the receiver. Motivated by the above considerations, we investigate efficient low-complexity turbo frequency domain reception techniques for SC broadband ST–BICM signaling with hybrid–ARQ operating over CCI-limited MIMO channels.

The powerful diversity–multiplexing tradeoff tool, initially introduced by Zheng and Tse for coherent delay-limited, i.e., quasi-static, MIMO channels [22], has been elegantly extended by El Gamal et al. to MIMO ARQ channels with flat fading, and referred to as diversity–multiplexing–delay tradeoff [23]. The authors have proved that the ARQ delay, i.e., maximum number of ARQ protocol rounds, improves the outage probability performance for large classes of MIMO ARQ channels [23]. In particular, they have demonstrated that the diversity order can be increased due to ARQ even when the MIMO ARQ channel is long-term static, i.e., the MIMO channel is random but fixed for all ARQ rounds. The diversity–multiplexing–delay tradeoff has then been characterized in the case of block-fading MIMO ARQ channels, i.e., multiple fading blocks are allowed within the same ARQ round [25]. In [26], the outage probability of MIMO-ISI ARQ channels has been evaluated under the assumptions of short-term static channel dynamic, and Chase-type ARQ [27], i.e., the data

1In non-ergodic, i.e., block fading quasi-static channels, the outage probability is a meaningful measure that provides a lower bound on the block error probability. It is defined as the probability that the mutual information, as a function of the channel realization and the average signal-to-noise ratio (SNR), is below the transmission rate [24].

2In the case of short-term static dynamic, the ARQ channel realizations are independent from round to round. This dynamic applies to slow ARQ protocols where the delay between two rounds is larger than the channel coherence time.
packet is entirely retransmitted. It has been shown that, as in the flat fading case, ARQ presents an important source of diversity, but its influence becomes only minimal when the ARQ delay is increased. This observation suggests that the design of practical packet combining schemes should target a high diversity order for early ARQ rounds. Supplementary retransmissions are then used to correct rare erroneous data packets, when they occur.

More recently, packet combining for MIMO ARQ systems has been investigated (e.g. [28]–[36]). Turbo combining techniques, where decoding is iteratively performed through the exchange of soft information between the soft-input–soft-output (SISO) packet combiner and the SISO decoder, have been proposed for the MIMO- ISI ARQ channel using unconditional minimum mean square error (MMSE)-aided combining [26], [37]. These approaches have then been extended to broadband MIMO code division multiple access (CDMA) [26], [37]. These approaches have then been extended to broadband MIMO code division multiple access (CDMA) systems with ARQ [38]. Time domain turbo packet combining for CCI-limited MIMO- ISI ARQ channels has been introduced in [39].

In this paper, we investigate efficient turbo receiver techniques for SC ST–BICM transmission with Chase-type ARQ over broadband MIMO channel with unknown CCI. We introduce a frequency domain MMSE-based turbo packet combining scheme, where all ARQ rounds are used to decode the data packet. By using an identical cyclic prefix (CP) word for multiple retransmissions of a symbol block, we perform transmission combining at the signal level. The frequency domain soft MMSE packet combiner performs soft ISI cancellation and retransmission combining in the presence of unknown CCI jointly over all received signal blocks. We also provide an efficient recursive implementation algorithm for the proposed scheme, and show that both the computational load and memory requirements are quite insensitive to the ARQ delay. The complexity order is only cubic in terms of the number of transmit antennas. Received signals and channel frequency responses (CFR) corresponding to all ARQ rounds are used without being required to be stored in the receiver. We analyze the asymptotic performance of the proposed combining scheme. Interestingly, we show that under a rank-condition on the MIMO ARQ channel corresponding to unknown CCI, the proposed combining scheme is not interference-limited, i.e., unknown CCI can be completely removed. Finally, we provide numerical simulation results for some scenarios to validate our findings.

The remainder of the paper is organized as follows. In Section II we describe the ARQ system under consideration, along with the communication model in the presence of unknown CCI. In Section III, we present the frequency domain turbo packet combining scheme we propose in this paper, and analyze both its complexity and memory requirements. In Section IV, we carry out the asymptotic performance analysis, and provide representative numerical results that demonstrate the gains achieved by the proposed scheme. Finally, we point out conclusions in Section V.

Notation:
- Superscripts *, †, and $^H$ denote conjugate, transpose, and Hermitian transpose, respectively. $\mathbb{E} [\cdot]$ is the mathematical expectation of the argument ($\cdot$).
- Let $X$ be a square matrix, diag $\{ X \}$ denotes the row vector corresponding to the diagonal of $X$, and $\text{tr} \{ X \}$ denotes the trace of $X$. When $X_1, \ldots, X_M \in \mathbb{C}^{N \times Q}$, diag $\{ X_1, \ldots, X_M \}$ denotes the $MN \times MQ$ matrix whose diagonal blocks are $X_1, \ldots, X_M$, diag $\{ x \}$ is the $N \times N$ diagonal matrix whose diagonal entries are the elements of the complex vector $x \in \mathbb{C}^N$. $(X)_{m,m}$ denotes the $m$th diagonal entry of matrix $X$.
- $I_N$ is the $N \times N$ identity matrix, and $0_{N \times Q}$ denotes an all zero $N \times Q$ matrix. For $i = 0, \ldots, T-1$, $E_{i,N}$ is a $N \times NT$ zero matrix where the $i$th $N \times N$ block is equal to $I_N$.
- Operator $\otimes$ denotes the Kronecker product, and $\delta_{m,n}$ is the Kronecker symbol, i.e., $\delta_{m,n} = 1$ for $m = n$ and $\delta_{m,n} = 0$ for $m \neq n$.
- For each sequence of matrices $X_0, \ldots, X_{T-1}$ (respectively, scalars $x_0, \ldots, x_{T-1}$), $\overline{X} \triangleq \frac{1}{T} \sum_{i=0}^{T-1} X_i$ denotes its time average (respectively, $\overline{x} \triangleq \frac{1}{T} \sum_{i=0}^{T-1} x_i$).
- $U_T$ is a $T \times T$ unitary matrix whose $(m,n)$th element is $(U_T)_{m,n} \triangleq \frac{1}{\sqrt{T}} \exp \left\{ -j \frac{2\pi mn}{T} \right\}$ for $m, n = 0, \ldots, T-1$, where $j = \sqrt{-1}$. $U_{T,N} = T N \times T N$ defined as $U_{T,N} \triangleq U_T \otimes I_N$.
- For each vector $x \in \mathbb{C}^N$, $x_j$ denotes the discrete Fourier Transform (DFT) of $x$, i.e., $x_j = U_Q x$.
- The acronym i.i.d. means “independent and identically distributed”.

II. ARQ System Model

A. SC–MIMO ARQ Transmission Scheme

We consider an SC multi-antenna-aided transmission scheme where the transmitter and the receiver are equipped with $N_T$ transmit (index $t = 1, \ldots, N_T$) and $N_R$ receive (index $r = 1, \ldots, N_R$) antennas, respectively. The MIMO channel is frequency selective and is composed of $L$ symbol-spaced taps (index $l = 0, \ldots, L-1$). The energy of each tap $l$ is denoted $\sigma_l^2$, and the total energy is normalized to one, i.e., $\sum_{l=0}^{L-1} \sigma_l^2 = 1$.

Each information block is initially encoded then interleaved with the aid of a semi-random interleaver II. The resulting frame is serial to parallel converted and mapped over the elements of the constellation set $S$ to produce symbol matrix $S \in S^{N_T \times T}$, where $T$ is the number of channel use (c.u). A CP word, whose length is $T_{CP} \geq L - 1$, is then appended to $S$, thereby yielding matrix $S' \in S^{N_T \times (T+T_{CP})}$. This allows the prevention of inter-block interference (IBI) and the exploitation of the multipath diversity of the MIMO broadband channel. We suppose that no channel state information (CSI) is available at the transmitter and assume infinitely deep interleaving. Therefore, transmitted symbols are independent and have equal transmit power, i.e.,

$$\mathbb{E} [s_{t,i} s_{t,i'}^*] = \delta_{t-t',i-i'}.$$  \hspace{1cm} (1)

At the upper layer, an ARQ protocol is used to help correct erroneous frames. An acknowledgment message is generated after the decoding of each information block. Therefore, when the decoding is successful, the receiver sends back a positive acknowledgment (ACK) to the transmitter, while the feedback
of a negative acknowledgment (NACK) indicates that the decoding outcome is erroneous. Let $K$ denote the ARQ delay, and $k = 1, \ldots, K$ denote the ARQ round index. When the transmitter receives an ACK feedback, it stops the transmission of the current block and moves on to the next information block. Reception of a NACK message incurs supplementary retransmission of the current block and moves on to the next information block. We focus on Chase-type ARQ, i.e., the number of taps $L$, $H_{N'}^{CCI(k)}$, and $\sigma^2_{\nu'}$ completely unknown at the receiver). As the desired user, the interferer employs a CP-aided transmission strategy. Its transmitted symbols $s_{CCI(k)}^{i}$ at each round $k$ verify the independence/energy-normalization condition (1) as useful symbols. Therefore, the signal-to-interference ratio (SIR) at each receive antenna is given as

$$SIR = \frac{N_T}{N_T \sum_{l'=0}^{L-1} \sigma^2_{\nu'}/}. \quad (3)$$

We assume perfect frame synchronization between the interferer and the desired user. They can differ in terms of the CP word length, which depends on the delay of the multipath channel, but are synchronized in terms of the useful symbol frames. Under this assumption, CP deletion yields the following baseband received $N_R \times 1$ signal at round $k$ and channel use $i$,

$$y_i^{(k)} = \sum_{l=0}^{L-1} H_i^{(k)} s_{i-(l)} \mod T + \sum_{l'=0}^{L-1} H_i^{CCI(k)} s_{i-(l')} \mod T + n_i^{(k)},$$

where $n_i^{(k)} \sim \mathcal{CN}(0, \sigma^2_{\nu} I_{N_R})$ denotes the receiver thermal noise. The SC–MIMO ARQ communication scheme at round $k$ is depicted in Fig. 1. In the following, we assume perfect channel estimation at each ARQ round $k$ (i.e., $H_i^{(k)}$ are perfectly known) while CCI channel matrices $H_i^{CCI(k)}$ are completely unknown at the receiver side.

1) Single-Round Communication Model: To derive the block communication model corresponding to ARQ round $k$,
we consider the following block signal vector,
\[ y^{(k)} = \begin{bmatrix} y_0^{(k)} & \cdots & y_{T-1}^{(k)} \end{bmatrix}^\top \in \mathbb{C}^{N_R T}, \]
that groups signals corresponding to the entire symbol frame. Vector \( y^{(k)} \) can be expressed as,
\[ y^{(k)} = H^{(k)} s + w^{(k)}, \]
where
\[ s \triangleq \begin{bmatrix} s_0^\top & \cdots & s_{T-1}^\top \end{bmatrix}^\top \in \mathbb{S}^{T N_T}, \]
\[ w^{(k)} \triangleq \begin{bmatrix} w_0^{(k)}^\top & \cdots & w_{T-1}^{(k)}^\top \end{bmatrix}^\top \in \mathbb{C}^{N_R T}, \]
is a block circulant matrix that can be block diagonalized in a Fourier basis as
\[ H^{(k)} = U_{T,N_R}^H \Lambda^{(k)} U_{T,N_T}, \]
where
\[ \Lambda^{(k)} \triangleq \text{diag} \left\{ \Lambda_0^{(k)}, \cdots, \Lambda_{T-1}^{(k)} \right\} \in \mathbb{C}^{N_R T \times N_T T}. \]
Exploiting (10) and the block circulant structure of \( H^{(k)} \), we get
\[ \Lambda_i^{(k)} = \sum_{l=0}^{L-1} H_l^{(k)} \exp \left\{ -2\pi il T \right\}. \]
Applying the DFT \( U_{T,N_R} \) on signal vector \( y^{(k)} \) yields the single-round frequency domain communication model
\[ y_f^{(k)} = \Lambda^{(k)} s_f + w_f^{(k)}, \]
where \( y_f^{(k)}, s_f, \) and \( w_f^{(k)} \) denote the DFT of \( y^{(k)}, s, \) and \( w^{(k)} \), respectively.

2) Multi-Round Communication Model: Let us suppose that received signals and channel matrices corresponding to ARQ rounds \( 1, \cdots, k \) are available at the receiver. First, we introduce the signal vector notation
\[ \mathbf{y}^{(k)} \triangleq \begin{bmatrix} y_f^{(1)} & \cdots & y_f^{(k)} \end{bmatrix}^\top \in \mathbb{C}^{k N_R}, \]
where received signals corresponding to multiple ARQ rounds are grouped in such a way to construct \( k N_R \) virtual receive antennas. Similarly, we define,
\[ H^{(k)} \triangleq \begin{bmatrix} H_f^{(1)} & \cdots & H_f^{(k)} \end{bmatrix}^\top \in \mathbb{C}^{k N_R T \times N_T}, \]
\[ W^{(k)} \triangleq \begin{bmatrix} W_f^{(1)} & \cdots & W_f^{(k)} \end{bmatrix}^\top \in \mathbb{C}^{k N_R}, \]

The block signal vector that serves for jointly performing, at ARQ round \( k \), packet combining and equalization in the presence of CCI is constructed similarly to (5),
\[ \mathbf{y}^{(k)} \triangleq \begin{bmatrix} y_0^{(k)} & \cdots & y_{T-1}^{(k)} \end{bmatrix}^\top \in \mathbb{C}^{k N_R T}, \]
and can be expressed as,
\[ y^{(k)} = H^{(k)} s + w^{(k)}, \]
where
\[ w^{(k)} \triangleq \begin{bmatrix} w_0^{(k)}^\top & \cdots & w_{T-1}^{(k)}^\top \end{bmatrix}^\top \in \mathbb{C}^{k N_R T}. \]
Matrix \( H^{(k)} \) has the same structure as (9), where its first \( T k N_R \times N_T \) block column is equal to
\[ \begin{bmatrix} H_0^{(k)} & \cdots & H_{T-1}^{(k)} & 0_{N_T \times (T-L)k N_R} \end{bmatrix}^\top. \]
\( H^{(k)} \) can be factorized similarly to (10) as,
\[ H^{(k)} = U_{T,k N_R}^H \Lambda^{(k)} U_{T,N_T}, \]
where
\[ \Lambda^{(k)} \triangleq \text{diag} \left\{ \Lambda_0^{(k)}, \cdots, \Lambda_{T-1}^{(k)} \right\} \in \mathbb{C}^{k N_R T \times N_T T}, \]
and matrices \( \Lambda_i^{(k)}, \ k' = 1, \cdots, k, \) are given by (12). The multi-round frequency domain communication model at ARQ round \( k \) is then expressed as,
\[ \mathbf{y}^{(k)} = \Delta^{(k)} s_f + \mathbf{w}^{(k)}, \]
where \( y_f^{(k)} \) and \( w_f^{(k)} \) denote the DFT of \( y^{(k)} \) and \( w^{(k)} \), respectively.

III. FREQUENCY DOMAIN TURBO PACKET COMBINING IN THE PRESENCE OF UNKNOWN CCI

A. General Description
At each ARQ round, the decoding of a data packet is performed by iteratively exchanging soft information in the form of log-likelihood ratio (LLR) values between the soft packet combiner, i.e., the joint transmission combining and equalization unit, and the soft-input–soft-output (SISO) decoder. Let us suppose that, at ARQ round \( k \), all received signals and channel matrices corresponding to previous rounds \( k \) are available at the receiver. Note that this assumption could not be feasible in practice since the receiver...
will require a huge memory. In Subsection III-D, we show that the proposed turbo packet combining algorithm requires little memory while it uses signals and CSIs corresponding to all ARQ rounds 1, · · · , k. The block diagram of the frequency domain turbo packet combining receiver at ARQ round k is depicted in Fig. 2.

First, the multiple ARQ rounds frequency domain block signal vector \( \mathbf{y}_k^{(k)} \) is constructed. Second, the soft packet combiner estimates the covariance of unknown CCI plus noise, then computes the multi-transmission MMSE filter that takes into account both co-antenna interference (CAI) and ISI while suppressing unknown CCI. These two elements are then used with a priori information to compute extrinsic LLRs corresponding to coded and interleaved bits. Only extrinsic information is fed back to the soft packet combiner to help perform transmission combining and equalization in the next turbo iteration. The iterative soft packet combining and decoding process is stopped after a preset number of turbo iterations and decision about the data packet is performed. The ACK/NACK message is then sent back to the transmitter depending on the decoding outcome. Note that during the first iteration a priori LLR values are the output of the SISO decoder obtained at the last iteration of previous round \( k - 1 \).

### B. Properties of CCI plus Noise Covariance

In this subsection, we focus on covariance properties of CCI plus noise present in both the single-round and multi-round communication models given by (6) and (18), respectively. These properties present an important ingredient in the turbo packet combining algorithm we introduce in Subsection III-C.

Let \( \Theta_k \) denote the covariance of CCI plus noise \( \mathbf{w}_k^{(k)} \) present in received signal (4) at round k,

\[
\Theta_k \triangleq \mathbb{E} \left[ \mathbf{w}_k^{(k)} \mathbf{w}_k^{(k)H} \right] \in \mathbb{C}^{NR \times NR}. \tag{25}
\]

Let us group covariance matrices corresponding to rounds \( 1, \cdots, k \) in the block diagonal matrix

\[
\Xi_k \triangleq \text{diag} \{ \Theta_1, \cdots, \Theta_k \} \in \mathbb{C}^{KNR \times KNR}. \tag{26}
\]

**Proposition 1:** The covariance \( \Xi_k \) of the CCI plus noise block vector \( \mathbf{w}_k^{(k)} \) present in the multi-round communication model (18) after \( k \) rounds is expressed as

\[
\Xi_k = I_T \otimes \Theta_k \in \mathbb{C}^{TKNR \times TKNR}. \tag{27}
\]

**Proof:** The expression in (27) is easily obtained by calculating the mathematical expectation of \( \mathbf{w}_k^{(k)} \mathbf{w}_k^{(k)H} \). In the derivation, we only exploit the independence between the entries of \( \mathbf{H}_i^{(k)} \) and \( \mathbf{H}_j^{(k)} \) \( \forall i, l, k \), and the fact that CCI symbols satisfy (1). No assumption on the structure of the CCI block matrix is used. A detailed proof of (27) in the case of sliding-window aided time-domain detection can be found in [39, Subsection III-C].

**Proposition 2:** The covariance \( \Theta_k \) of the single-round CCI plus noise block vector \( \mathbf{w}_k^{(k)} \) at ARQ round k is

\[
\Theta_k = I_T \otimes \Theta_k. \tag{28}
\]

**Proof:** The proof follows by simply invoking Proposition 1 for one round.

**Proposition 3:** Covariance matrices of frequency domain CCI plus noise vectors \( \mathbf{w}_f^{(k)} \) and \( \mathbf{w}_f^{(k)} \) (corresponding to the DFTs of \( \mathbf{w}_k^{(k)} \) and \( \mathbf{w}_k^{(k)} \), respectively) are \( \Xi_k \) and \( \Theta_k \), respectively.

**Proof:** The proof of Proposition 3 follows from the fact that \( \Xi_k \) and \( \Theta_k \) are block circulant and block diagonal matrices.

Proposition 1 indicates that the covariance of the multi-round CCI plus noise vector can be obtained by separately computing single-round covariances using Proposition 2. This result greatly impacts the computational complexity of the proposed algorithm as it will be shown in Subsection III-D.

### C. Proposed Scheme

In this subsection, we derive the frequency domain MMSE-based soft packet combiner that cancels both CAI and ISI jointly over multiple ARQ rounds in the presence of unknown CCI.

To combine signals corresponding to ARQ rounds 1, · · · , k, we use conventional soft parallel interference cancellation (PIC) (of both multi-round CAI and ISI) and unconditional MMSE filtering techniques [3]. Therefore, at each turbo iteration of ARQ round k, the MMSE-based soft packet combiner produces a complex scalar decision \( \tilde{z}_k^{(k)} \) that serves for computing extrinsic LLR values corresponding to coded and interleaved bits mapped over symbol \( s_{t,i} \). Let \( \varphi_{t,i} \) denote the vector of a priori LLRs of bits corresponding to symbol \( s_{t,i} \), and available at the input of the soft combiner at a particular turbo iteration. \( \mathbf{a}_{t,i} \triangleq \mathbb{E} \left[ s_{t,i} \mid \varphi_{t,i} \right] - \mathbb{E} \left[ s_{t,i} \right] \) denotes the conditional variance of \( s_{t,i} \). By invoking either the orthogonal projection theorem or Lagrangian methods, and using (11) and (27), soft MMSE-based packet combining at ARQ round k, can be performed in the frequency domain as,

\[
\mathbf{z}_f^{(k)} = \mathbf{G}_k \tilde{\mathbf{y}}_f^{(k)} - \Theta_k \tilde{\mathbf{s}}_f^{(k)}, \tag{29}
\]

where \( \mathbf{z}_f^{(k)} \) is the DFT of \( \mathbf{z}^{(k)} \triangleq \left[ z_{1,1}^{(k)}, \cdots, z_{N_T,1}^{(k)} \right]^T \in \mathbb{C}^{N_F T} \), i.e., \( \mathbf{z}_f^{(k)} = \mathbf{U}_f^H \mathbf{z}^{(k)} \). \( \tilde{\mathbf{s}}_f^{(k)} \) denotes the DFT of the soft symbol vector \( \mathbf{s} \triangleq \mathbb{E} \left[ s \mid \varphi_{t,i} \right] \) \( \forall (t,i) \), and

\[
\mathbf{G}_k = \text{diag} \left\{ \Delta_k^{(k)H} \mathbf{B}_1^{(k)H}, \cdots, \Delta_k^{(k)H} \mathbf{B}_{k-1}^{(k)H} \right\}, \quad \mathbf{G}_k = \mathbf{C}_k - I_T \otimes \text{diag} \left\{ (\mathbf{C}_k)_{1,1}, \cdots, (\mathbf{C}_k)_{N_T,N_T} \right\}, \tag{30}
\]

\[
\mathbf{B}_k^{(k)} = \Delta_k^{(k)} \Xi_k \mathbf{B}_k^{(k)H} + \Xi_k; \quad \mathbf{C}_k^{(k)} = \Delta_k^{(k)} \mathbf{B}_k^{(k)H} \Delta_k^{(k)}, \tag{31}
\]

\[
\Xi_k = \text{diag} \left\{ \sigma_{1,i}^{(k)}, \cdots, \sigma_{N_T,i}^{(k)} \right\} \in \mathbb{R}^{N_T \times N_T}. \]

Matrices \( \mathbf{C}_k^{(k)} \) and \( \Xi_k \) denote time averages of \( \mathbf{C}_k^{(k)} \) and \( \Xi_k \), respectively, as defined in Section I. The input for the soft
demapper can be extracted from \( z_f^{(k)} \) as \( z_{t,i}^{(k)} = e_{t,i}^H \Theta_k^{(0)} z_f^{(k)} \) where \( e_{t,i} \) is the \((iN_T + t)\)th vector of the canonical basis. As it can be seen from the forward–backward filtering structure in (29), the frequency domain MMSE filter explicitly cancels soft CAI and ISI while it only requires the covariance of unknown CCI plus noise. Note that both Propositions 1 and 2 are used to derive (29).

To obtain estimates of unknown CCI plus noise covariance matrices \( \Theta_1, \ldots, \Theta_k \), required by (31), let us consider the single-round frequency domain communication model (13). Proposition 3 indicates that the covariance of \( w_f^{(k)} \) is \( \Theta_k = I_T \otimes \Theta_k \). Therefore, with respect to the block diagonal structure of (13), unknown CCI plus noise covariance \( \Theta_k \) can directly be estimated in the frequency domain at each turbo iteration, with the aid of a priori LLRs, according to the following average,

\[
\Theta_k = \frac{1}{T} \sum_{t=0}^{T-1} \left( y_{t,i}^{(k)} - \Lambda_i^{(k)} \hat{s}_f \right) \left( y_{t,i}^{(k)} - \Lambda_i^{(k)} \hat{s}_f \right)^H. \tag{32}
\]

\( y_{t,i}^{(k)} \) and \( \hat{s}_f \) denote the DFTs of \( y_i^{(k)} \) and \( s \) at frequency bin \( i \), respectively, i.e., \( y_{t,i}^{(k)} = E_{i,N_T} y_i^{(k)} \) and \( \hat{s}_f = E_{i,N_f} \hat{s}_f \). Covariance matrices \( \Theta_1, \ldots, \Theta_{k-1} \) are similarly estimated at ARQ rounds 1, \ldots, \( k-1 \), respectively, and correspond to estimates obtained at the last turbo iteration. In other words, when the decoding outcome is erroneous, a NACK message is fed back to the transmitter, and the unknown CCI plus noise covariance estimate obtained at the last iteration is saved in the receiver to help perform packet combining at the next ARQ round.

\[\] D. Implementation Aspects

We first provide an efficient implementation of the proposed scheme since turbo combining requires at each turbo iteration the computation of matrix inverses \( B_0^{(k-1)} \), \( \cdots, B_{k-1}^{(k-1)} \in \mathbb{C}^{kN_R \times kN_R} \) given by (31). Second, we analyze the computational complexity and memory requirements of the proposed implementation algorithm.

1) An Efficient Implementation Algorithm: The special structure of the frequency domain ARQ channel matrix (23) together with the matrix inversion lemma [40] allow us to express the inverse of \( B_i^{(k)} \) as,

\[
B_i^{(k)} = \Xi_k^{(k)} - \Xi_k^{(k)} \Lambda_i^{(k)} \left( \hat{\Sigma} + D_i^{(k)} \right)^{-1} \Lambda_i^{(k)H} \Xi_k^{(k)}, \tag{33}
\]

where \( D_i^{(k)} \) is obtained according to the following recursion,

\[
\begin{align*}
D_i^{(k)} &= D_i^{(k-1)} + \Lambda_i^{(k)H} \Theta_k^{(k-1)1} A_i^{(k-1)} \\
D_i^{(0)} &= 0_{N_T \times N_T}.
\end{align*}
\]

Therefore, matrices \( C_i^{(k)} \), \( \cdots, \) \( C_i^{(k-1)} \) are simply computed as,

\[
C_i^{(k)} = D_i^{(k)} - D_i^{(k)} \left( \hat{\Sigma} + D_i^{(k)} \right)^{-1} D_i^{(k)}, \tag{35}
\]

while the forward filtering part of (29) is calculated at each ARQ round \( k \) as,

\[
\Gamma^{(k)} \hat{\Sigma}_f^{(k)} = \Gamma^{(k)} \hat{\Sigma}_f^{(k)} \tag{36}
\]

where

\[
\begin{align*}
\hat{\Sigma}_f^{(0)} &= \hat{\Sigma}_f^{(k-1)} + \Lambda_i^{(k)H} \left( I_T \otimes \Theta_k^{(k-1)} \right) y_{f}^{(k)} \\
\hat{\Sigma}_f^{(0)} &= 0_{N_T \times 1}.
\end{align*}
\]

(37)

The proposed turbo packet combining algorithm is summarized in Table I. Note that, during the first iteration of round \( k \), the anti-causal parts in recursions (34) and (38), i.e., \( D_i^{(k-1)} \) and \( \hat{\Sigma}_f^{(k-1)} \), respectively, correspond to the output of these recursions at the last iteration of previous round \( k-1 \).
2) Computational Complexity and Memory Requirements: The proposed recursive implementation algorithm avoids storing received signals and CFRs corresponding to multiple ARQ rounds. It also prevents the computation of $kN_R \times kN_R$ matrix inverses. This dramatically reduces the implementation cost since the complexity order of directly computing $B_i^{-1}(k)$ is cubic against $kN_R$, and is greatly increased from round to round. In the following, we analyze both the complexity and memory requirements of the proposed scheme, and compare them with those of the LLR-level combining technique.

First, note that in the case of LLR-level packet combining, frequency domain MMSE equalization is separately performed for each ARQ round. Therefore, $T$ inversions of $N_R \times N_R$ matrices are required to compute the forward and backward filters. Since in general it is required to have more receive than transmit antennas, especially when CCI is present in the system, an implementation similar to that introduced in the previous subsection is beneficial because only $T$ inversions of $N_T \times N_T$ matrices will be required. In this case, the two variables in recursions (34) and (38) are computed at ARQ round $k$ as,

$$R_{NR} = \begin{bmatrix} 1 & \delta_R \end{bmatrix}_{N_R \times N_R}, \quad R_{N_T} = \begin{bmatrix} 1 & \delta_T \end{bmatrix}_{N_T \times N_T},$$

(42)

Signal level using signals and CFRs corresponding to all ARQ rounds, without being required to be explicitly stored in the receiver. This is performed with the aid of the two variables $D_i(k)$ and $\tilde{y}_j(k)$ in recursions (34) and (38), respectively. This translates into a memory size of $2TN_T(N_T + 1)$ real values. Therefore, the computational complexity and storage requirements are less sensitive to the ARQ delay. The technique requires only a few more additions and a bit more memory compared to LLR-level combining. Table II summarizes implementation requirements and reports the relative costs for some modulation schemes.

### IV. PERFORMANCE EVALUATION

#### A. Asymptotic Performance Analysis

In the following, we provide a frame-based analysis where we derive system conditions under which perfect CCI cancellation holds. We suppose that the interferer CSI is perfectly known, and investigate the influence of its channel properties on the interference cancellation capability of the proposed packet combining scheme in the high SNR regime.

**Theorem 1:** We consider a CCI-limited MIMO ARQ system with $N_T$ transmit and $N_R$ receive antennas, and ARQ delay $K$. Let $\Theta_{CCI}$ denote the CCI covariance at ARQ round $k = 1, \cdots, K$, i.e., the covariance of the global noise at the receiver is $\Theta_k = \Theta_{CCI} + \sigma^2 I_{N_R}$, and $\rho_k$ be the rank of $\Theta_{CCI}$. We assume perfect LLR feedback from the SISO decoder. The frequency domain soft MMSE packet combiner provides perfect CCI suppression for asymptotically high SNR if

$$\sum_{u=1}^{k} \rho_u < kN_R - N_T. \quad (39)$$

**Proof:** See the Appendix.

We now proceed to derive an upper bound on $\rho_k$, where we incorporate the rank of the CCI fading channel. Under the assumption that CCI symbols satisfy (1), i.e., infinitely deep interleaving, we get

$$\Theta_{CCI} = \sum_{l=0}^{L-1} H_{U}^{CCI(l)} H_{U}^{CCI(l)^H}.$$  

(40)

Relative costs refer to the relative number of arithmetic additions $\Delta C$ and memory $\Delta M$ required by the proposed scheme compared to LLR-level combining. With respect to storage requirements and number of arithmetic additions in Table II, we have $\Delta C = \Delta M = \frac{2TN_T + 1}{\log_2 |S|}$.  

#### TABLE II

**Summary of Memory and Arithmetic Additions Required by the Proposed and LLR-Level Combing Schemes, and Relative Cost Evaluation**

| Combining scheme | Memory | Arithmetic Additions | Relative Costs |
|------------------|--------|----------------------|---------------|
| LLR-Level        | $TN_T \log_2 |S|$ | $TN_TN_R(k - 1) \log_2 |S|$ | QPSK 8-PSK 16-QAM |
| Proposed         | $2TN_T(N_R + 1)$ | $2TN_TN_R(k - 1)(N_T + 1)$ | $N_T^2 \frac{2N_T - 1}{2} \frac{N_T}{2}$ |
Let us write each CCI channel matrix as

$$
\mathbf{H}_{i}^{CCI(k)} = \mathbf{R}_{NT}^{1/2} \mathbf{A}_{i}^{CCI(k)} \mathbf{R}_{NT}^{1/2}, \quad \forall i',
$$  \hspace{1cm} (41)

where $\mathbf{A}_{i}^{CCI(k)} \in \mathbb{C}^{N_R \times N_T}$ characterizes the scattering environment between the CCI transmitter and receiver [41], and $\mathbf{R}_{NT}$ and $\mathbf{R}_{NT}^r$ are the correlation matrices controlling the receive and transmit antenna arrays, and are in general independent, and using (40) and (41), we get

$$
\rho_k \leq \min \left\{ N_R, \sum_{i'=0}^{L'-1} \text{rank}\left\{ \mathbf{H}_{i}^{CCI(k)} \mathbf{H}_{i'}^{CCI(\Theta)^H} \right\} \right\}
$$

$$
= \min \left\{ N_R, \sum_{i'=0}^{L'-1} \text{rank}\left\{ \mathbf{A}_{i}^{CCI(k)} \mathbf{R}_{NT} \mathbf{A}_{i'}^{CCI(\Theta)^H} \right\} \right\}
$$

$$
\leq \min \left\{ N_R, \sum_{i'=0}^{L'-1} \text{rank}\left\{ \mathbf{A}_{i}^{CCI(k)} \right\} \right\}. \quad (43)
$$

A closer look at Theorem 1 and upper bound (43) provides interesting system interpretations.

- **Impact of CCI Fading Channel:** First, note that the CCI cancellation capability of the frequency domain MMSE packet combiner is related to the CCI channel rank. When the interferer has a rank-deficient channel matrix at a certain ARQ round, interference can completely be removed (at subsequent rounds) if the sum-rank condition in Theorem 1 is satisfied. In practice, the channel rank can dramatically drop in the case of the so-called pinhole channel, where the transmitter and receiver are largely separated and are surrounded by multiple scatterers [41]. In this scenario, the channel can even prevent multipath from building up since the thin air pipe connecting transmitter and receiver scatterers is very long. For instance, in a system with $N_T = 3$ transmit and $N_R = 2$ receive antennas, and an unknown interferer who is experiencing one path ($L' = 1$) channel realizations with rank equal to two, CCI can be removed at the second ARQ round because the sum-rank condition (39) holds for $k \geq 2$.

- **Impact of the Number of Transmit Antennas and ARQ Delay:** Condition (39) suggests how, for a given CCI channel profile, the number of transmit antennas $N_T$ and ARQ rounds $K$ are chosen to achieve perfect CCI cancellation. For instance, if transmission is corrupted by CCI with quasi-static channel rank $\rho$, and if the ARQ delay allowed by the upper layer is $K$, then only $N_T < K (N_R - \rho)$ transmit antennas can be allocated to the user of interest to achieve interference suppression at the latest at ARQ round $K$, where $\rho_0$ is the rank of $\Theta_k^{CCI}$, i.e., $\rho_k = \rho_0 \forall k$. Increasing the ARQ delay will relax the condition on the number of transmit antennas and therefore allow for an increase in the diversity and/or multiplexing gains depending on the diversity-multiplexing-delay trade-off operating point [23]. Note that when $N_T' \ll N_T$, the CCI channel rank dramatically drops, and therefore CCI suppression is achieved even when a short ARQ delay $K$ is required.

- **Interaction with the Scheduling Mechanism:** In the case of opportunistic communications, interference with co-channel users who have high channel ranks can be prevented. For instance, when a retransmission is required on the reverse link, the base station (BS) can choose the timing of the next ARQ round in such a way that transmission simultaneously occurs with that of a user with low channel rank. This is feasible since the BS has complete knowledge about user CSIs in the reverse link. The same scheduling mechanism can be used in the forward link if all users provide the BS with feedback information about their channel ranks. When the system suffers from CCI caused by neighboring cells, the sum-rank condition (39) can be achieved by simply increasing the number of ARQ rounds because the CCI channel rank tends to be constant over time.

**B. Numerical Results**

In this subsection, we provide block error rate (BLER) performance results for the proposed combining technique. Our focus is to demonstrate the superior performance of the introduced scheme compared to LLR-level combining. We also evaluate BLER performance for scenarios where the interferer has rank-deficient channel matrices to corroborate the theoretical analysis in Subsection IV-A.

In all simulations, we consider a BICM scheme where the encoder is a $\frac{1}{2}$-rate convolutional code with polynomial generators $(35, 23)_b$, and the modulation scheme is quadrature...
phase shift keying (QPSK). The length of the code bit frame is 1032 bits including tails. The ARQ delay is $K = 3$, and the $E_b/N_0$ ratio appearing in all figures is the SNR per useful bit per receive antenna. We consider a $L = 2$ path MIMO-ISI channel profile where $\sigma^2_0 = \sigma^2_1 = \frac{1}{2}$. In practical wireless systems, the wireless channel may have more than two paths due to severe frequency selective fading. In this paper, we restrict ourselves to $L = 2$ for the sake of simulation simplicity. Performance in the case of severe frequency selective fading channels can be found in [38]. We use both the matched filter bound (MFB) per ARQ round and the outage probability [26] of the CCI-free MIMO-ISI channel as absolute performance bounds to evaluate the CCI cancellation capability and diversity order achieved by the proposed combining scheme. The number of turbo iterations is set to five and the Max-Log-MAP version of the maximum a posteriori (MAP) algorithm is used for SISO decoding.

We first investigate performance for scenarios where the user of interest and the interferer have the same number of transmit antennas ($N_T = N_R = 2$) and identical channel profiles, i.e., $L = L'$, equal power taps, and CCI fading channel coefficients are i.i.d. In Fig. 3, we compare the BLER performance of the proposed scheme with that of LLR-level combining for a ST–BICM code with rate $R = 2$, i.e., $N_T = 2$. The number of receive antennas is $N_R = 2$, and $\text{SIR} = 3\text{dB}$. We observe that the proposed scheme significantly outperforms LLR-level combining. The performance gap at ARQ round $k = 3$ is about 1dB for $\text{BLER} \leq 10^{-2}$. Note that both combining schemes fail to perfectly cancel CCI since performance curves tend to saturate for high $E_b/N_0$ values. Fig. 4 reports performance of both techniques when SIR is increased to $5\text{dB}$. In this case, the performance gap between the two schemes is reduced. The CCI cancellation capability is also improved as can be seen from the steeper slopes of BLER curves. In Fig. 5, we evaluate the performance for a high rate ST–BICM code where $R = 4$, i.e., $N_T = 4$. Only $N_R = 2$ receive antennas are considered, and $\text{SIR} = 5\text{dB}$. The proposed scheme dramatically outperforms LLR-level combining, i.e., the performance gap at ARQ round $k = 3$ is about 4dB at $7 \times 10^{-3} \text{BLER}$. The proposed scheme also offers higher cancellation capability and diversity order than LLR-level combining.

We now turn to scenarios where the interferer has a rank-deficient uncorrelated MIMO channel, i.e., $\text{rank} \{ \mathbf{A}_U^{(\text{CCI})} \} < \min (N_T', N_R)$, $\forall l'$, $\delta_{Tx} = \delta_{Rx} = 0$, and assume that the rank is constant over all ARQ rounds. In Fig. 6, we report the BLER performance of the proposed scheme for a CCI-limited MIMO system with settings similar to Fig. 3, i.e., $N_T = N_R = 2$, and $\text{SIR} = 3\text{dB}$. The interferer experiences flat fading, i.e., $L' = 1$, and only has $N_T' = 1$ transmit antenna. Therefore, with respect to (43), $\rho_k = 1 \forall k$. Note that in this interference scenario, the perfect CCI cancellation condition (39) holds for $k \geq 2$. We observe that both the CCI cancellation capability
and the diversity order of the proposed scheme are improved. The performance gain with respect to the case of $N_T' = 2$ and $L' = 2$ is about 1.5dB at $3 \times 10^{-3}$ BLER and round $k = 3$, and the slope of the BLER curve at round $k = 3$ is similar to that of the MFB curve. Fig. 7 compares the performance of the proposed scheme for two scenarios with heavy CCI, i.e., SIR = 1dB. The ST-BICM code has rate $R = 4$, i.e., $N_T = 4$, and the number of receive antennas is set to $N_R = 4$. In the first scenario (Scenario 1), the interferer has $N_T' = 4$ transmit antennas, $L' = 2$ equal power taps, and i.i.d. fading coefficients, while in the second scenario (Scenario 2), $N_T' = 2$, $L' = 1$, and the CCI channel rank is equal to two. Therefore, $\rho_k = 2$ for all $k$, and condition (39) holds for $k \geq 2$. It is clear that in the second scenario, better CCI cancellation capability is achieved for $k \geq 2$. For instance, the performance gap for $k = 3$ is more than 2dB at $2 \times 10^{-2}$ BLER. Also, the diversity order of the CCI-free MIMO-ISI channel is almost achieved.

V. CONCLUSION

In this paper, we investigated efficient iterative turbo packet combining for broadband ST-BICM transmission with hybrid ARQ over CCI-limited MIMO-ISI channels. We have introduced a frequency domain turbo combining scheme where signals and CFRs corresponding to all ARQ rounds are combined in a MMSE fashion to decode the data packet at each round. The covariance of the overall (over all ARQ rounds) CCI plus noise required by the frequency domain MMSE soft packet combiner is constructed by separately computing the covariance related to each round. The proposed technique has a complexity order cubic against the product of the number of receive antennas and ARQ delay. This limitation is overcome by an optimized recursive implementation algorithm where complexity is only cubic in term of the number of transmit antennas. We evaluated the computational load and memory requirements, and found that the introduced recursive technique only requires few arithmetic additions and memory compared to conventional LLR-level combining schemes. We analyzed the effect of CCI channel rank on performance. Interestingly, under a sum-rank condition, the frequency domain MMSE soft packet combiner can completely remove CCI for asymptotically high SNR. Finally, we provided simulation results where we showed that the proposed technique achieves BLER performance superior to LLR-level combining, and offers high CCI cancellation capability and diversity order for many interference scenarios.

APPENDIX

PROOF OF THEOREM 1

Under the assumption of perfect LLR feedback from the SISO decoder, the frequency domain soft packet combiner output (29), at ARQ round $k$, can be expressed as,

$$z_{f_{\text{perfect LLR}}}^{(k)} = As_f + x_f^{(k)},$$

where $A$ is the diagonal matrix of frequency domain symbol gains,

$$A = \text{diag} \left\{ (G_0^{(k)})_{1,1}, \cdots, (G_0^{(k)})_{N_T',N_R}, \cdots, (G_{T-1}^{(k)})_{1,1}, \cdots, (G_{T-1}^{(k)})_{N_T',N_R} \right\},$$

with $G_i^{(k)} = \Delta_i^{(k)} \Xi_i^{-1} \Delta_i^{(k)}$ and $x_f^{(k)}$ is the filtered CCI plus thermal noise at the output of the packet combining filter. Its covariance matrix is

$$G_f^{(k)} = \text{diag} \left\{ G_0^{(k)}, \cdots, G_{T-1}^{(k)} \right\}.$$ 

Now, let us examine the structure of matrix $G_i^{(k)}$ for asymptotically high SNR, i.e., $\sigma_f^2 \to 0$. Let $\Pi_1, \Pi_2, \cdots, \Pi_k, \Pi_{k+1}$ be the low-rank decompositions of matrices $\Theta_0^{(k)}_{\text{CCI}}, \cdots, \Theta_k^{(k)}_{\text{CCI}}$, where $\Pi_1 \in \mathbb{C}^{N_T \times \rho_1}, \cdots, \Pi_k \in \mathbb{C}^{N_T \times \rho_k}$. For the sake of notation simplicity, we write $\Sigma_{k=1}^k \rho_k = \rho$. It follows that the rank of $\Pi = \text{diag} \{ \Pi_1, \cdots, \Pi_k \}$ is $\rho$, and $\Xi_k = \Pi \Pi^H + \sigma^2 I_{kN_R}$ is a square invertible matrix. Therefore, it has an eigenvalue decomposition (EVD) that can be expressed as,

$$\Xi_k = [P_p \ P_{kN_R-\rho}] \begin{bmatrix} \Psi + \sigma^2 I_{\rho} & \sigma^2 I_{kN_R-\rho} \\ \sigma^2 I_{kN_R-\rho} & P_{kN_R-\rho}^H \end{bmatrix} \begin{bmatrix} P_p^H \\ P_{kN_R-\rho}^H \end{bmatrix},$$

where $PP^H = P^H P = I_{kN_R}$ since $\Xi_k$ is symmetric. This condition yields the following set of equalities,

$$P_{kN_R-\rho} P_{kN_R-\rho} = I_{kN_R},$$

$$P_{kN_R-\rho}^HP_{kN_R-\rho} = 0,$$

$$P_p P_p^H + P_{kN_R-\rho} P_{kN_R-\rho}^H = I_{kN_R}.$$

Therefore, a Taylor expansion of $\Xi_k^{-1}$ when $\sigma^2 \to 0$, is given as,

$$\Xi_k^{-1} = P_p \Psi^{-1} P_p^H + \sigma^2 I_{kN_R} + O(\sigma^2).$$

Note that $\Psi$ does not have any null diagonal element, i.e., $\Psi$ is invertible. Indeed, multiplying the left and right sides of (47)
by $P^H$ and $P$, respectively, and with respect to (48a), we get,
\[ P^H \Pi P = Y. \]  
By noting that $P^H \Pi$ is $\rho \times \rho$ and has rank equal to $\rho$, it follows that $Y^{-1} = (\Pi^H P)\rho^{-1} (P^H \Pi)^{-1}$. Therefore, when $\sigma^2 \to 0$, we have,
\[ G_i^{(k)} = \Delta^{(k)} P Y^{-1} P^H \Delta^{(k)} + \sigma^2 \Delta^{(k)} + O(\sigma^2). \]

Since the time domain channel coefficients are i.i.d., it follows that the $kN_R \times N_T$ matrix $\Delta^{(k)}$ has full-column rank unless all fading coefficients are equal to zero. If $\rho + N_T < kN_R$, i.e., $\rho < kN_R - N_T$, then all the first $\rho$ columns of $\Xi_k$ (column vectors of $P$) are in the kernel of $\Delta^{(k)^H}$, i.e., $\Delta^{(k)^H} P = 0_{N_T \times \rho}$. It follows that, when $\sigma^2 \to 0$,
\[ G_i^{(k)} = \sigma^2 \Delta^{(k)^H} + O(\sigma^2). \]

Therefore, when SNR$\to\infty$, we get
\[ \text{SNR}_{\text{MF}} = \frac{1}{\sigma^2} \sum_{i=0}^{T-1} \sum_{u=1}^{k} \left( \Delta^{(k)^H} \Delta^{(k)} \right) + O(\sigma^2), \]
\[ = \frac{1}{\sigma^2} \sum_{i=0}^{T-1} \sum_{u=1}^{k} \left( H^{(u)^H} H^{(u)} \right) + O(\sigma^2), \]

where SNR$_{\text{MF}}$ corresponds to the instantaneous matched filter (MF) SNR in the case of $k$ rounds CCI-free MIMO-ISI ARQ channel.

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