Chained Turbo Equalization for Block Transmission without Guard Interval

Khoirul Anwar*, Hui Zhou* and Tad Matsumoto*,‡
*Japan Advanced Institute of Science and Technology (JAIST), 1-1 Asahidai, Nomi, Ishikawa, 923-1292 JAPAN
Email:{anwar-k, hui-z, matumoto}@jaist.ac.jp
‡Center for Wireless Communications, University of Oulu, FI-90014 Finland
Email: tadashi.matsumoto@ee.oulu.fi

***Abstract*—This paper proposes chained turbo equalization (CHATUE) algorithm, a simple detection scheme for block transmission without guard interval (GI), where turbo equalizers for the several consecutive frames exchange information about the interference components. Although the block transmission without GI can significantly increase bandwidth efficiency, the received signal suffers from inter-block interference (IBI) and inter-symbol interference (ISI), which can't be effectively compensated by block-by-block equalization. With the CHATUE algorithm, the IBI components are cancelled by utilizing *a* posteriori feedback from the decoders of the neighboring equalizers (future and past). To best utilize the latest results of the sub-optimal reduced complexity equalization, frequency domain soft cancellation and minimum mean squared error filtering (FD/SC-MMSE), we exploit J matrix to convert the Toeplitz channel matrix structure to a circulant matrix. Results of the computer simulation for single carrier block transmission (SCBT) show that the proposed scheme can achieve good performance.**

I. Introduction

Block transmission with cyclic prefix (CP) as the guard interval (GI), for example, orthogonal frequency division multiplexing (OFDM) [1], digital multi-tone (DMT) [2] and single carrier block transmission with CP (SC-CP), [3], etc., have been drawing much attention due to the robustness against fading frequency selectivity. For the SC-CP the latest version of the reduced complexity equalization technique, frequency domain soft cancellation minimum mean squared error (FD/SC-MMSE) [4] turbo equalization can effectively eliminate the interference components without requiring computationally heavy burden.

However, CP transmission causes loss in the bandwidth efficiency. On the contrary, if no GI is employed in the block transmission, the received signal suffers from inter-block interference (IBI), and the conventional FD/SC-MMSE cannot effectively eliminate ISI because the channel matrix no longer has a circulant structure. Therefore, if the FD/SC-MMSE algorithm can be modified so that the interference components can be cancelled without having to transmit the CP, significant improvement in bandwidth efficiency can be expected.

To overcome the problem of IBI due to insufficient GI length, several techniques have been presented. Ref. [5] uses turbo equalization technique to remove the IBI and ISI using the knowledge only from the past block and [6] makes modification at the special pilot configuration for transmission with insufficient GI length.

Recently in [7], single carrier system without CP is proposed by assuming that IBI from the past block can be completely cancelled so that the problem comes only from the future blocks. By further assuming that the interference plus noise can be approximated by complex colored Gaussian noise, [7] applies the Bayesian linear unbiased estimation (BLUE) to cancel the IBI from the future.

However, the past and future equivalent channel should come from the tail and head of the past and future channel response instead from the current observed channel. Therefore the assumed channel matrices in [7] is true if and only if the channel always stay the same in each block which is unrealistic in multipath fading enviroments. Furthermore, combining energy which is spread in many blocks should be considered, by means of the interference knowledge exchange between blocks where the interferences come from, to obtain better performance.

This paper proposes a new frequency domain turbo equalization technique, CHAined TUrbo Equalization (CHATUE) by exchanging the knowledge of IBI components between the neighboring blocks. The received signal samples of several blocks neighboring in time, i.e., past, present, and future, are stored, and their equalizers and decoders exchange *a* posteriori information of the symbols causing IBI. The *a* posteriori information is obtained from the decoders, and is used for soft cancelation of the IBI components in the neighboring equalizer blocks. We exploit a **J** matrix, to convert the equivalent channel matrix from Toeplitz structure into the circulant matrix, which enables us to use the conventional FD/SC-MMSE to preserve its superiority in terms of low computational complexity.

II. System Model

The transmitter and receiver structures assumed in this paper are shown in Fig. 1. Single-Input Single-Output (SISO) antenna system is assumed in this paper, however, its extension to multiple-input multiple-output (MIMO) structure is rather straightforward. At the transmitter, information bits to be transmitted is encoded by the encoder C. Block of coded bits is interleaved by a random interleaver Π and modulated to

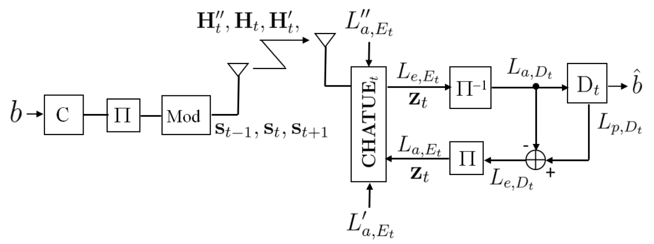

Fig. 1. System model of CHATUE algorithm for the t-th block.

obtain symbol block $\mathbf{s}$. The t-th block is expressed as

$$\mathbf{s}_t = [s_t^{[1]}, s_t^{[2]}, \cdots, s_t^{[K]}]^T \in \mathbb{C}^{K\times 1}, \tag{1}$$

with K being the block length. These blocks are then transmitted over the frequency selective fading channel $\mathbf{H}$ with frequency selectivity due to multipath propagation.

The fading channel gain is assumed to be constant during one block interval, but vary block-by-block. Let $\mathbf{H}'$, $\mathbf{H}$ and $\mathbf{H}''$ denote the equivalent block-wise representation of the channel corresponding to which is for past, current and future blocks, respectively.

When cyclic prefix (CP) is appended at the transmitter side and eliminated at the receiver side, the equivalent channel matrix H becomes circulant. However, in this paper, CP transmission is not assumed, for which the channel matrix for the current block has a Toeplitz structure, as

$$\mathbf{H}_t = \begin{bmatrix} h_t^{[0]} & & & 0 \\ \vdots & h_t^{[0]} & & \\ h_t^{[L-1]} & \vdots & \ddots & \\ & h_t^{[L-1]} & \vdots & h_t^{[0]} \\ & & \ddots & \vdots \\ 0 & & & h_t^{[L-1]} \end{bmatrix} \in \mathbb{C}^{(K+L-1)\times K}. \tag{2}$$

The channel matrix for the interference components from the past block is

$$\mathbf{H}'_{t-1} = \begin{bmatrix} & h_{t-1}^{[L-1]} & \cdots & h_{t-1}^{[1]} \\ & & \ddots & \vdots \\ & & & h_{t-1}^{[L-1]} \\ & & & \\ 0 & & & \end{bmatrix} \in \mathbb{C}^{(K+L-1)\times K}, \tag{3}$$

and from future block

$$\mathbf{H}''_{t+1} = \begin{bmatrix} & & & 0 \\ & & & \\ h_{t+1}^{[0]} & & & \\ \vdots & \ddots & & \\ h_{t+1}^{[L-2]} & \cdots & h_{t+1}^{[0]} & \end{bmatrix} \in \mathbb{C}^{(K+L-1)\times K}. \tag{4}$$

At the receiver, the channel is assumed to be known. The received signal of the current block, $\mathbf{y}$ can be formulated as

$$\mathbf{y}_t = \mathbf{H}_t\mathbf{s}_t + \mathbf{H}'_{t-1}\mathbf{s}'_{t-1} + \mathbf{H}''_{t+1}\mathbf{s}''_{t+1} + \mathbf{n}, \tag{5}$$

where $\mathbf{s}_t$ is the current block, $\mathbf{s}'_t$ and $\mathbf{s}''_t$ is interference components from past and future, respectively, and $\mathbf{n}$ is a zero mean complex additive white Gaussian noise (AWGN) vector with covariance $\mathbf{E}\{\mathbf{n}\mathbf{n}^H\} = \sigma^2\mathbf{I}$. σ^2 denotes the noise variance defined by the specified signal to noise ratio (SNR) per antenna. The equivalent interference component from the past is

$$\mathbf{s}'_{t-1} = [0, \cdots, 0, s_{t-1}^{[L-1]}, \cdots, s_{t-1}^{[-1]}]^T \in \mathbb{C}^{K\times 1}, \tag{6}$$

and from the future is

$$\mathbf{s}''_{t+1} = [s_{t+1}^{[K]}, \cdots, s_{t+1}^{[K+L-1]}, 0, \cdots, 0]^T \in \mathbb{C}^{K\times 1}, \tag{7}$$

where L is channel impulse response length. $\mathbf{s}'_t$, $\mathbf{s}_t$ and $\mathbf{s}''_t$ may be originated from the same user or different users. In the case where the blocks are originated by different users, the system is equivalent to time division multiple access (TDMA).

The CHATUE equalizer for the t-th block receives three log-likelihood ratio (LLRs) inputs:

$$L_{a,E_t} = \frac{P(s_t = +1)}{P(s_t = -1)} \tag{8}$$

for ISI cancelation of the t-th block and

$$L''_{a,E_t} = \frac{P(s_{t-1} = +1)}{P(s_{t-1} = -1)}, \quad L'_{a,E_t} = \frac{P(s_{t+1} = +1)}{P(s_{t+1} = -1)}, \tag{9}$$

as the knowledge of interference components from the past and future blocks.

It is important to note here that L_{a,E_t} is in the form of an extrinsic LLR due to the fact the iteration is performed between equalizer E_t and decoder D_t, while L''_{a,E_t} and L'_{a,E_t} is *a* posteriori LLR from D_{t-1} and D_{t+1}, respectively. The deinterleaved version of L_{e,E_t} by Π^{-1} becomes L_{a,D_t} while the L_{p,D_t} and $L_{e,D_t} = L_{p,D_t} - L_{a,D_t}$ are *a* posteriori and extrinsic LLRs, as output from the decoder, respectively.

III. IBI Cancellation

The current block suffers from IBI components from the past. Furthermore, IBI components from the future is inevitable when sampling continues until the end of the impulse response for the last symbol in the current block in order to preserve the entire channel energy.

Fig. 2(a) describes an equivalent channel of the $\mathbf{y}_t$ block and its relation to the equivalent channels $\mathbf{H}''_t$ and $\mathbf{H}'_t$ which is required to cancell the IBI components in (6) and (7) as expressed by (5).

Fig. 2(b) shows a chain structure to cancel the IBI components by exchanging LLRs from the past and future blocks iteratively. Equalizer for the t-th block, E_t, receives two IBI components L''_{a,E_t} and L'_{a,E_t} from the past and future, respectively. At the same time, the decoder D_t provides two *a* posteriori LLRs, L''_{p,E_t} for the past block and L'_{p,E_t} for the future block.

In order to best utilize the latest version of the FD/SC-MMSE algorithm, we exploit a J matrix [7], given by

$$\mathbf{J} = \left[\begin{array}{c|c} 0_{(K-L+1)\times(L-1)} & \mathbf{I}_{K\times K} \\ \mathbf{I}_{(L-1)\times(L-1)} & \end{array}\right] \in \mathbb{C}^{K\times(K+L-1)}, \tag{10}$$

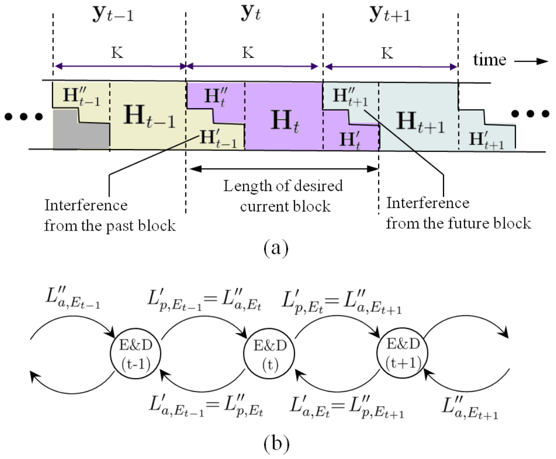

Fig. 2. (a). Equivalent channel of block $\mathbf{y}_t$ that comprises of the equivalent channels $\mathbf{H}'_{t-1}$ of interference components from the past and $\mathbf{H}''_{t+1}$ of interference components from the future blocks and (b). *a* posteriori information exchange between the neighboring blocks.

to convert the Toeplitz current channel matrix $\mathbf{H}_c$ into a circulant matrix, $\mathbf{JH}_c$, and therefore, the received signal block $\mathbf{r}$ becomes

$$\mathbf{r}_t = \mathbf{J}\mathbf{y}_t = \mathbf{J}\mathbf{H}_t\mathbf{s}_t + \mathbf{J}\mathbf{H}'_{t-1}\mathbf{s}'_{t-1} + \mathbf{J}\mathbf{H}''_{t+1}\mathbf{s}''_{t+1} + \mathbf{J}\mathbf{n}. \quad (11)$$

In order to utilize the LLR of the coded symbols from the past and future block decoders, we can calculate the soft symbols $\hat{\mathbf{s}}'_t$, $\hat{\mathbf{s}}_t$, and $\hat{\mathbf{s}}''_t$ in the past, present, and future blocks, respectively. The soft replica of the received signal vector $\hat{\mathbf{r}}$ is given by

$$\hat{\mathbf{r}}_t = \mathbf{J}\mathbf{H}_t\hat{\mathbf{s}}_t + \mathbf{J}\mathbf{H}'_{t-1}\hat{\mathbf{s}}'_{t-1} + \mathbf{J}\mathbf{H}''_{t+1}\hat{\mathbf{s}}''_{t+1}. \quad (12)$$

We, then, perform soft cancellation of the ISI and IBI components, of which residual is given by

$$\tilde{\mathbf{r}}_t = \mathbf{r}_t - \hat{\mathbf{r}}_t. \quad (13)$$

To reduce the computational complexity, we perform all computation in a block-wise format. Therefore, we introduce *restoral term*, $\mathbf{h}_t(k)\hat{s}_t(k)$, to the (13) to obtain $\hat{\mathbf{s}}_t(k)$

$$\begin{aligned} \hat{\mathbf{s}}_t(k) &= \mathbf{r} - \hat{\mathbf{r}} + \mathbf{h}_t(k)\hat{s}_t(k) \\ &= \tilde{\mathbf{r}} + \mathbf{h}_t(k)\hat{s}_t(k), \end{aligned} \quad (14)$$

where $\mathbf{h}_t(k) \in \mathbb{C}^{(K+L-1)\times 1}$ denotes the k-th column vector of the current channel matrix $\mathbf{H}_t$ in (2) and $\hat{s}_t(k)$ for binary phase shift keying (BPSK) is given by

$$\hat{s}_t(k) = \tanh[L_{a,E_t}(s_t^{[k]})/2]. \quad (15)$$

Similarly,

$$\hat{s}'_{t-1}(k) = \tanh[L'_{a,E_t}(s'^{[k]}_{t-1})/2], \quad (16)$$

and

$$\hat{s}''_{t+1}(k) = \tanh[L''_{a,E_t}(s''^{[k]}_{t+1})/2], \quad (17)$$

to form $\hat{\mathbf{s}}'_{t-1}$ and $\hat{\mathbf{s}}''_{t+1}$.

IV. Proposed CHATUE Algorithm

Minimum mean squared error (MMSE) filtering is used to further reduce the residual error in (13), of which criterion is given by

$$\mathbf{w}_t(k) = \arg\min_{\mathbf{w}_t^H(k)} \left|\mathbf{w}_t^H(k)\hat{\mathbf{s}}_t(k) - s_t(k)\right|^2, \quad (18)$$

to obtain the weight of MMSE filtering is

$$\begin{aligned} \mathbf{w}_t(k) &= \left(\mathrm{E}[\tilde{r}_t\tilde{r}_t^H] + \mathbf{h}_t(k)|\mathbf{s}_t(k)|^2\mathbf{h}_t^H(k)\right)^{-1}\mathbf{h}_t(k) \\ &= \left(\mathbf{J}\mathbf{H}_t\mathbf{\Lambda}_t\mathbf{H}_t^H\mathbf{J}^H + \mathbf{J}\mathbf{H}'_{t-1}\mathbf{\Lambda}'_{t-1}\mathbf{H}'^H_{t-1}\mathbf{J}^H \right.\\ &\quad + \mathbf{J}\mathbf{H}''_{t+1}\mathbf{\Lambda}''_{t+1}\mathbf{H}''^H_{t+1}\mathbf{J}^H + \sigma^2\mathbf{J}\mathbf{J}^H \\ &\quad \left. + \mathbf{h}_t(k)|\mathbf{s}_t(k)|^2\mathbf{h}_t^H(k)\right)^{-1}\mathbf{h}_t(k) \\ &= \left(\mathbf{\Sigma} + \mathbf{h}_t(k)|\mathbf{s}_t(k)|^2\mathbf{h}_t^H(k)\right)^{-1}\mathbf{h}_t(k) \end{aligned} \quad (19)$$

A block-wise solution to the MMSE problem can be derived based on the latest version of the FD/SC-MMSE algorithm. Since interleaving is random enough, it is reasonable to assume that the soft symbols are uncorrelated, yielding the symbol-level covariance matrix of the ISI components remaining in the current frame being diagonal, as

$$\mathbf{\Lambda}_t = \mathrm{diag}\{\mathrm{E}[|\hat{s}_t|^2] - |\hat{s}_t|^2\} \in \mathbb{C}^{K\times K}, \quad (20a)$$

$$\mathbf{\Lambda}'_{t-1} = \mathrm{diag}\{\mathrm{E}[|\hat{s}'_{t-1}|^2] - |\hat{s}'_{t-1}|^2\} \in \mathbb{C}^{K\times K}, \quad (20b)$$

$$\mathbf{\Lambda}''_{t+1} = \mathrm{diag}\{\mathrm{E}[|\hat{s}''_{t+1}|^2] - |\hat{s}''_{t+1}|^2\} \in \mathbb{C}^{K\times K}. \quad (20c)$$

When BPSK modulation is assumed, the (20) becomes

$$\mathbf{\Lambda}_t = \mathrm{diag}\{1 - |\hat{s}_t^{[0]}|^2, \cdots, 1 - |\hat{s}_t^{[K]} - 1)|^2\}, \quad (21a)$$

$$\mathbf{\Lambda}'_{t-1} = \mathrm{diag}\{0, \cdots, 0, 1 - |\hat{s}'^{[-L+1]}_{t-1}|^2, \cdots, 1 - |\hat{s}'^{[-1]}_{t-1}|^2\}, \quad (21b)$$

$$\mathbf{\Lambda}''_{t+1} = \mathrm{diag}\{1 - |\hat{s}''^{[K]}_{t+1}|^2, \cdots, 1 - |\hat{s}''^{[K+L-1]}_{t+1}|^2, 0, \cdots, 0\}. \quad (21c)$$

The covariance matrix of the soft cancellation output, including IBI from the past and future block as well as ISI from the current block is given by

$$\begin{aligned} \mathbf{\Sigma} &= \mathbf{J}\mathbf{H}_t\mathbf{\Lambda}_t(\mathbf{J}\mathbf{H}_t)^H + \mathbf{J}\mathbf{H}'_{t-1}\mathbf{\Lambda}'_{t-1}(\mathbf{J}\mathbf{H}'_{t-1})^H \\ &\quad + \mathbf{J}\mathbf{H}''_{t+1}\mathbf{\Lambda}''_{t+1}(\mathbf{J}\mathbf{H}''_{t+1})^H + \sigma^2\mathbf{J}\mathbf{J}^H. \end{aligned} \quad (22)$$

Decomposing the filter output into two parts, the desired component and the residual, and invoking the matrix inversion lemma, the filter output can be expressed as

$$\begin{aligned} z(k) &= \mathbf{w}_t^H(k)\hat{\mathbf{s}}_t(k) = \mathbf{w}_t^H(k)(\tilde{\mathbf{r}} + \mathbf{h}_t(k)\hat{\mathbf{s}}_t(k)) \\ &= (1 + \gamma(k)|\hat{s}_t(k)|^2)^{-1}\mathbf{h}_t^H(k)\mathbf{\Sigma}^{-1} \\ &\quad \cdot(\tilde{\mathbf{r}}_t(k) + \mathbf{h}_t(k)\hat{\mathbf{s}}_t(k)), \end{aligned} \quad (23)$$

where

$$\gamma(k) = \mathbf{h}_t^H(k)\mathbf{\Sigma}^{-1}\mathbf{h}_t(k), \quad (24)$$

and the block wise expression of (23) can be expressed as

$$\mathbf{z} = (\mathbf{I} + \mathbf{\Gamma}\mathbf{S}_t)^{-1}[\mathbf{\Gamma}\hat{\mathbf{s}}_t + \mathbf{H}_t^H\mathbf{J}^H\mathbf{\Sigma}^{-1}\tilde{\mathbf{r}}_t], \quad (25)$$

where

$$\mathbf{\Gamma} = \mathrm{diag}[\mathbf{H}_t^H\mathbf{J}^H\mathbf{\Sigma}^{-1}\mathbf{J}\mathbf{H}_t] \quad (26)$$

is a block-wise expression of $\gamma(k)$ in (24) and

$$\mathbf{S}_t = \mathrm{diag}[|\hat{\mathbf{s}}_t|^2] \in \mathbb{C}^{K\times K}. \tag{27}$$

Now, recall that a circulant channel matrix in the time domain can be converted into a diagonal matrix in the frequency domain by the discrete Fourier transform (DFT), as

$$\mathbf{J}\mathbf{H}_t = \mathbf{F}^H \mathbf{\Phi}\mathbf{F}, \tag{28}$$

to obtain

$$\begin{aligned}\mathbf{\Sigma} \;=\;& \mathbf{F}^H\mathbf{\Phi}\mathbf{F}\mathbf{\Lambda}_t\mathbf{F}^H\mathbf{\Phi}^H\mathbf{F} + \mathbf{J}\mathbf{H}'_{t-1}\mathbf{\Lambda}'_{t-1}(\mathbf{J}\mathbf{H}'_{t-1})^H \\ &+\mathbf{J}\mathbf{H}''_{t+1}\mathbf{\Lambda}''_{t+1}(\mathbf{J}\mathbf{H}''_{t+1})^H + \sigma^2\mathbf{J}\mathbf{J}^H,\end{aligned} \tag{29}$$

of which frequency domain equivalent is

$$\begin{aligned}\mathbf{X} \;=\;& \mathbf{F}\mathbf{\Sigma}\mathbf{F}^H \\ =\;& \mathbf{\Phi}\mathbf{F}\mathbf{\Lambda}_t\mathbf{F}^H\mathbf{\Phi}^H + \mathbf{F}\mathbf{J}\mathbf{H}'^H_{t-1}\mathbf{\Lambda}'_{t+1}\mathbf{H}'^H_{t-1}\mathbf{J}^H\mathbf{F}^H \\ &+\mathbf{F}\mathbf{J}\mathbf{H}''^H_{t+1}\mathbf{\Lambda}''_{t+1}\mathbf{H}''^H_{t+1}\mathbf{J}^H\mathbf{F}^H + \mathbf{F}\sigma^2\mathbf{J}\mathbf{J}^H\mathbf{F}^H\end{aligned} \tag{30}$$

with

$$\mathbf{\Sigma}^{-1} = \mathbf{F}^H\mathbf{X}^{-1}\mathbf{F}. \tag{31}$$

The final output of CHATUE equalizer is given by

$$\begin{aligned}\mathbf{z} \;=\;& (\mathbf{I}_K + \mathbf{\Gamma}\mathbf{S}_t)^{-1}[\mathbf{\Gamma}\hat{\mathbf{s}}_t + \mathbf{H}_t^H\mathbf{J}^H\mathbf{\Sigma}^{-1}\tilde{\mathbf{r}}_r] \\ =\;& (\mathbf{I}_K + \mathbf{\Gamma}\mathbf{S}_t)^{-1}[\mathbf{\Gamma}\hat{\mathbf{s}}_t + \mathbf{F}^H\mathbf{\Phi}^H\mathbf{X}^{-1}\mathbf{F}\tilde{\mathbf{r}}_t].\end{aligned} \tag{32}$$

It is assumed that the output $\mathbf{z}$ can be expressed as

$$\mathbf{z} = \mu\mathbf{s}_t + \nu, \tag{33}$$

where

$$\begin{aligned}\mu \;=\;& \mathrm{E}[\mathbf{z}\cdot\mathbf{s}_t*] \\ =\;& (\mathbf{I}+\mathbf{\Gamma}\mathbf{S}_t)^{-1}\mathbf{\Gamma}\cdot\mathrm{E}[|\mathbf{s}_t|^2] \\ =\;& (\mathbf{I}+\mathbf{\Gamma}\mathbf{S}_t)^{-1}\mathbf{\Gamma},\end{aligned} \tag{34}$$

where $\mathbf{s}_t*$ is the complex conjugate of $\mathbf{s}_t$. Since $\mathrm{E}[|\mathbf{s}_t|^2] = 1$ for BPSK modulation and ν is equivalent noise vector with variance

$$\sigma^2 = \mu(1-\mu). \tag{35}$$

Now, we can convert the CHATUE output into an extrinsic LLR

$$\mathbf{L}_{e,E_t} = \frac{4\Re(\mathbf{z})}{1-\mu}, \tag{36}$$

where $\Re(\mathbf{z})$ denotes the real part of complex $\mathbf{z}$.

V. Computer Simulation

A block diagram of the chain simulations conducted to evaluate performances of the proposed CHATUE algorithm is shown in Fig. 3 for the current t-th block. In fact, each CHATUE equalizer is connected to the equalizers for its corresponding future and past blocks, the mutual information (MI) provided by the decoder for the block $t+2$ to the block $t+1$ has to be taken into account as well as the blocks $t-2$ to the $t-1$. To avoid unacceptable complexity due to the chained structure, we invoke the following assumptions: Since the block length is large and the channel variations are random

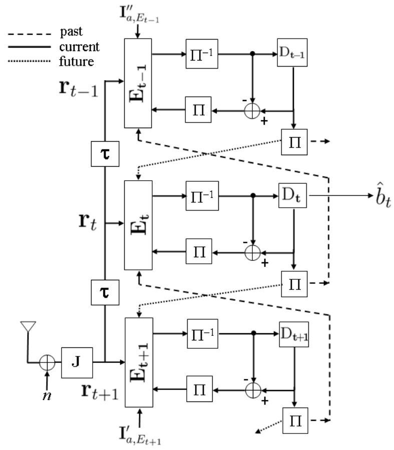

Fig. 3. Simulation setup of a chained turbo equalization structure.

enough, it is reasonable to assume that the *a* posteriori mutual information between $\mathbf{s}$ and $\mathbf{z}$ for the block $t-2$ (causing IBI on the block $t-1$) and is the same as that for the block $t+1$ as

$$I''_{a,E_{t-1}} \approx I'_{a,E_{t+1}}, \tag{37}$$

where I is the mutual information between $\mathbf{s}$ and $\mathbf{z}$ as described in [8]. Note that this assumption is reasonable because mutual information is already in the sense of average. In the future work, we consider the complexity reduction and the delay optimization.

A. EXIT Analysis

To evaluate the convergence property of the proposed CHATUE algorithm, we perform extrinsic information transfer (EXIT) analysis. The sequences of the *a* posteriori LLRs provided by the decoders for the past and the future blocks are approximated as being Gaussian distributed [8].

Fig. 4 shows the EXIT curves of equalizers and decoder at *SNR* = 2dB. Because there are three LLR inputs and one LLR output, the EXIT function of the equalizer will be in four dimensions (4D). To simplify the presentation, in this paper, we plot the lower bound of the EXIT curves by assuming $I'_{a,E_{t-1}} = I''_{a,E_{t+1}} = 0$, as

$$I_{e,E_t} = T_{E_t}(I_{a,E_t}, 0, 0, \mathrm{SNR}), \tag{38}$$

where T_{E_t} is the transfer function of CHATUE equalizer for the t-th block.

The upperbound of the EXIT curve is plotted by assuming $I'_{a,E_{t-1}} = I''_{a,E_{t+1}} = 1$, as

$$I_{e,E_t} = T_{E_t}(I_{a,E}, 1, 1, \mathrm{SNR}). \tag{39}$$

The decoder EXIT curve is calculated by measuring the histogram of the decoder output LLR, obtained by using the bahl-cocke-jelinek-raviv (BCJR) algorithm [9]. A convolutional encoder (CC) with constraint length of 3 and code generator G=[7 5], noted as CC-3(7,5), is assumed. The decoding trajectory is plotted by exchanging the mutual information in the form of LLR.

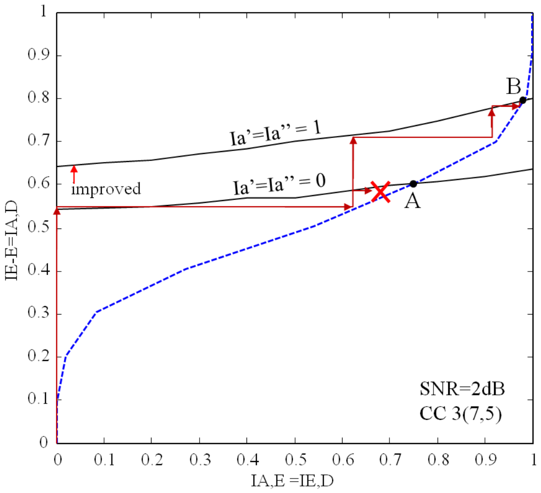

Fig. 4. EXIT analysis of CHATUE algorithm.

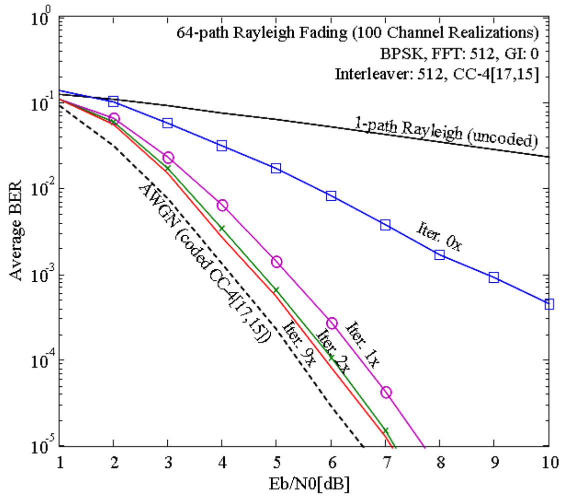

Fig. 5. BER performance of CHATUE algorithm is within about 0.5dB from the lower bound.

The decoder has one input, I_{a,D_t}, one output, I_{e,D_t} and is independent of SNR because of no direct connection to the channel. Transfer function for the decoder EXIT curve is expressed as

$$I_{e,D_t} = T_{D_t}(I_{a,D}), \tag{40}$$

where T_{D_t} is the transfer function of the decoder for the t-th block.

Fig. 4 shows EXIT curves and trajectory for SNR = 2dB. It is found that with the mutual information feedback from the past and future being zero, the EXIT curves intersect before the decoder output mutual information reaches one; with the mutual information feedback from the past and future being one, the EXIT curves do not intersect and the decoder output mutual information reaches one. The trajectory shows that neighboring LLRs help the equalizer to lift up its EXIT curve to avoid the intersection at Point A, and shift the intersection to Point B. Thereby, the performance is improved.

B. BER Performance

Bit error rate (BER) performance in multipath fading channel with 64-path equal average power is shown in Fig. 5 for a CC-4(17,15). The fading gain is assumed constant within one block but fast (randomly) varying block-by-block. The lower-bound is the simulated BER curve of block transmission in AWGN channel using the same parameters, while the upper-bound is the theoretical BER curve of an uncoded system in Rayleigh fading channel. It is found that the proposed CHATUE equalizer can achieve good performance and converges after 2 or 3 iterations (BER with 9 iterations is almost the same as with 2 iterations) which is only about 0.5dB away from the lower bound at BER of 10^{-4}.

VI. Conclusion

In this paper, we have proposed a new equalization technique, which allows us to eliminate the cyclic prefix (or guard interval) transmission. Enhancement in spectrum efficiency can be expected when the channel delay spread is large with respect to the block length. Results of the EXIT chart analysis were also presented to demonstrate the impact of the mutual information feedback from the the past and the future blocks through chain simulations. It has been shown that the mutual information feedback from the past and future can significantly improve the convergence property of the equalization for the current block. Results of the BER simulations were also presented in this paper to demonstrate a good performance which is within 0.5dB away from the lower bound BER of AWGN channel.

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