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Robust Receiver Algorithms to Mitigate Partial-Band and Partial-Time Interference in LDPC-coded OFDM Systems

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***Abstract*—Orthogonal frequency division multiplexing (OFDM) systems are vulnerable to narrow-band jamming signals. We jointly tackle two problems: channel estimation in the presence of unknown interference, and decoding with imperfect channel knowledge. In this paper, we propose robust, yet simple, receiver algorithms consisting of both channel estimation and information decoding. The receiver conducts threshold tests to detect interference followed by pilot erasure and channel estimation. Then, channel estimation error and unknown interference statistics are dealt with by the *robust log-likelihood ratio (LLR) calculations* for soft iterative decoding. The proposed receiver algorithm does not require any statistical knowledge of interference and its complexity is linear against the length of codewords. Simulation results show that the bit-error-rate (BER) performance of the proposed system is only $2 \sim 3$ dB away from a genie system where channel information and interference parameters are perfectly known. We also demonstrate that soft decision feedback from a decoder to enhance channel estimation achieves additional $0.5 \sim 1$ dB improvement.**

I. Introduction

Orthogonal frequency-division multiplexing (OFDM) techniques are widely used in wireless systems, e.g., 4G LTE and WLAN standards. Increased use of wireless communication techniques inevitably results in coexistence of different systems in the same spectrum. Consequently, interference has become the main factor for system performance degradation. Interference mitigation techniques are crucial to improve the performance for future wireless systems.

In the literature, the interference mitigation problem is separately discussed as two problems. The first problem considers OFDM channel estimation in the presence of interference with unknown parameters. In [1], a Gaussian mixture model is proposed for the interference distribution, where the interference variance is assumed to be unknown. However, this variance is assumed to be drawn from the inverse Gamma distribution. The authors in [2] also consider a Gaussian mixture interference model and propose a redescending estimator based on extreme value theory for OFDM channel estimation. The second problem deals with contaminated OFDM symbol detection assuming perfect channel estimation. To address the issue of unknown interference, robust log-likelihood ratios (LLRs) [3], [4], parametric LLRs [5], piecewise-linear LLR approximation [6], expectation maximization algorithm [7], and structured channel coding [8] were proposed.

We note that decoupling the interference mitigation task from the channel estimation task ignores the effects of channel estimation errors on the detection of contaminated data symbols. Also, this decoupling approach rules out the possibility of joint designs or the design of closed-loop (i.e., adaptive or iterative) receivers. These motivate our study on robust OFDM receiver designs including channel estimation and symbol detection. In [9], the authors apply time-domain prediction-error filter (PEF) to mitigate narrow-band interference signals (NBI) with constant frequency. However, the constant frequency interference models have not captured the time dynamics and frequency correlation properties, e.g., Bluetooth signals are characterized as frequency hopping partial-band and partial-time interference (PBPTI) for OFDM symbols. In this paper, we propose a frequency-domain method to deal with the PBPTI. A robust OFDM receiver that uses low-density parity-check (LDPC) codes and soft iterative decoding is proposed. The contributions of this paper are summarized as:

1) We propose two prior LLR metrics, *robust LLR* and *dynamic LLR*, to resolve interference detection error, channel estimation error, and unknown interference statistics. The prior LLRs can be used for soft iterative LDPC decoder.
2) The proposed receiver algorithm only assumes the knowledge of noise variance, which can be estimated from previous transmissions, and it does not require the statistics of the OFDM channel or the interference. The complexity of receiver processing is linear in the block length of LDPC codes.
3) Closed-loop system designs using decision feedback from a soft decoder to improve channel estimation is discussed.

The rest of this paper is organized as follows. In Section II, we present the channel and interference model. Section III describes the open-loop LDPC-OFDM system and the robust receiver algorithms. Section IV discusses decision feedback methods. Simulation results are provided in Section V. Finally,

This work is conceived and completed during the first author's internship at MERL in the summer of 2011.

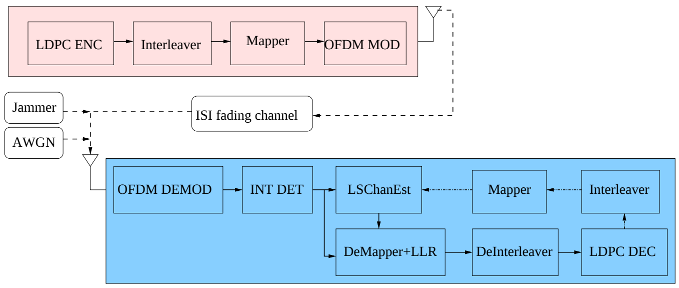

Fig. 1. System Diagram

concluding remarks are given in Section VI.

Notation: For matrix $\mathbf{A}$, let $\mathbf{A}^*$ denote its Hermitian. We define $\mathcal{CN}(0, \sigma^2)$ as a circularly symmetrical complex Gaussian distribution with zero mean and variance σ^2. We also use $\lceil a \rceil$ to denote the maximum integer no smaller than a. For two sets $\mathcal{A}$ and $\mathcal{B}$, we use $|\mathcal{A}|$, $\mathcal{A} \setminus \mathcal{B}$, $\mathcal{A} \bigcap \mathcal{B}$, and $\mathcal{A} \bigcup \mathcal{B}$ for set cardinality, set difference, intersection, and union, respectively.

II. System Model

Consider a baseband inter-symbol-interference (ISI) channel with L taps. Each channel tap is denoted as h_l and is modeled as independent $\mathcal{CN}(0, \sigma_l^2)$ distributed random variable. The normalized delay of h_l is denoted as τ_l where normalization is taken over the sample interval. Denote the $1 \times L$ vectors $\mathbf{\Sigma}^2 = [\sigma_1^2, \cdots, \sigma_L^2]$ and $\mathbf{\Gamma} = [\tau_1, \ldots, \tau_L]$. Then, $\mathbf{\Sigma}^2$ and $\mathbf{\Gamma}$ describe the power delay profile (PDP) for the ISI channel. We assume that the channels remain unchanged during one OFDM symbol, thereby no inter-carrier interference results. A system diagram is shown in Fig. 1. We describe the system setup in II-A and the model of interference in II-B.

A. System Setup

The input, a $1 \times M$ vector, of binary bits is encoded by a systematic irregular LDPC code which has approximately ρ degrees per check node and γ degrees per bit node in the $M \times N$ parity-check matrix [10]. The output block has length N, and the code rate is $R = \frac{M}{N}$. Other LDPC codes, e.g., [11] and references therein, can be used similarly here. A random interleaver with length N is added to protect against burst errors, followed by a discrete constellation mapping. We consider typical constellations such as BPSK, QPSK, and 16QAM. Extension to other constellations is straightforward. Denote the set of constellation symbols as $\mathcal{S}$ and its cardinality as $|S|$. The output $1 \times \frac{N}{|S|}$ symbol vector is divided equally into J vectors of size $1 \times N_\mathrm{D}$, each of which is loaded into data tones of one OFDM symbol. From the system description, we have $N_\mathrm{D} = \lceil \frac{N}{J|S|} \rceil$.

The OFDM modulator converts the ISI channel in the time domain into parallel flat channels in the frequency domain. Each OFDM symbol consists of N_FFT tones, and is converted to time domain by multiplying with an $N_\mathrm{FFT} \times N_\mathrm{FFT}$ inverse discrete Fourier transform (DFT) matrix $\mathbf{D}$ whose (l, k)-th entry is $\frac{1}{\sqrt{N_\mathrm{FFT}}} \exp\left(j \frac{2\pi k l}{N_\mathrm{FFT}}\right)$. A cyclic prefix (CP) with length Δ longer than the maximum channel delay is added to the beginning of the output vector from the inverse DFT, and the resulting $1 \times (N_\mathrm{FFT} + \Delta)$ vector is transmitted over the ISI channel. The receiver removes the CP and multiplies the resulting vector with the DFT matrix $\mathbf{D}^*$ to convert back to the frequency domain. These two operations comprise the ”OFDM DEMOD” block in the receiver. Denote each entry of one OFDM symbol as s_k. The equivalent channel in the frequency domain at the k-th tone can be expressed as

$$y_k = H_k s_k + w_k + I_k, \qquad k \in \{1, \ldots, N_\mathrm{FFT}\}, \tag{1}$$

where w_k denotes the i. i. d. $\mathcal{CN}(0, \sigma_\mathrm{w}^2)$ additive white Gaussian noise (AWGN), H_k denotes the frequency channel, and I_k denotes the interference. The frequency channel, H_k, is relatee to h_l by the Fourier transform $H_k = \sum\limits_{l=1:L} h_l \exp\left(-j \frac{2\pi k \tau_l}{N_\mathrm{FFT}}\right)$.

For one OFDM symbol with N_FFT tones, N_D tones are for data and N_P tones are for pilots. Denote the index set for data tones as $\mathcal{N}_\mathrm{D}$ and that for pilot tones as $\mathcal{N}_\mathrm{P}$. Then, $s_k = 1$ for $k \in \mathcal{N}_\mathrm{P}$. The pilot tones are equally spaced and their positions remain unchanged during the transmission of one LDPC codeword.

B. Interference Modeling

In what follows, we describe a frequency correlated Gaussian mixture model for the PBPTI I_k. Let $I_k = c z_k u_k$, where c is a coefficient representing the power of interference, u_k is modeled as i. i. d. $\mathcal{CN}(0, 1)$ distributed random variable, and z_k denotes a real random variable determining the power spectrum density (PSD) and location of the interference. The vector $\mathbf{z} = [z_1^2, \ldots, z_{N_\mathrm{FFT}}^2]$ defines a profile for the interference. Fig. 2 illustrates an example of the profile vector of $\mathbf{z}$.

The vector $\mathbf{z}$ depends on the following two random variables and $K+1$ parameters. The first random variable χ is modeled as Bernoulli distributed with probability p equal to 1. It describes the status of interference. When $\chi = 0$, the OFDM symbol is not corrupted by interference, i.e., $\mathbf{z} = \mathbf{0}$; when $\chi = 1$, the OFDM symbol is corrupted by interference. We also assume that interference jointly corrupts consecutive K tones starting from tone ω. The random variable ω is modeled as an integer uniformly distributed between 1 and $N_\mathrm{FFT} - K + 1$. And the parameter K describes the coherence bandwidth of the interference. In other words, there are K consecutive nonzero entries in $\mathbf{z}$. Denote the vector of nonzero entries in $\mathbf{z}$ as $\mathbf{\Phi}$, which describes the PSD of the interference. To normalize the PSD when interference is present, we assume $\|\mathbf{z}\| = 1$ such that the power of interference is absorbed in c. Let the average power of interference be σ_I^2. We thus have $c = \sqrt{\sigma_\mathrm{I}^2 N_\mathrm{FFT}/p}$. The random variables $[\chi, \omega]$ change independently for each OFDM symbol, while the parameters $[p, K, \mathbf{\Phi}]$ are constant. In summary, the interference is modeled as *time-independent* and *frequency-correlated Gaussian mixture*. One example for the profile vector $\mathbf{z}$ with $N_\mathrm{FFT} = 8$, $K = 2$, and $\mathbf{\Phi} = [0.5, 0.5]$ is illustrated in Fig. 2, where

$$\mathbf{z}_1 = \left[\, 0\ 0\ 0\ 0.5\ 0.5\ 0.0\ 0\ 0 \,\right], \quad \mathbf{z}_2 = \left[\, 0.5\ 0.5\ 0\ 0\ 0\ 0\ 0\ 0 \,\right],$$
$$\mathbf{z}_3 = \left[\, 0\ 0\ 0\ 0\ 0\ 0\ 0\ 0 \,\right], \quad \mathbf{z}_4 = \left[\, 0\ 0\ 0\ 0\ 0.5\ 0.5\ 0\ 0 \,\right],$$
$$\mathbf{z}_5 = \left[\, 0\ 0\ 0\ 0\ 0\ 0\ 0.5\ 0.5 \,\right], \quad \mathbf{z}_6 = \left[\, 0\ 0\ 0.5\ 0.5\ 0\ 0\ 0\ 0 \,\right],$$

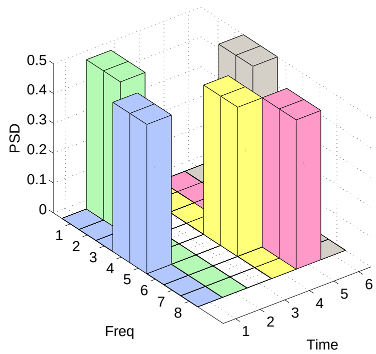

Fig. 2. The profile vector of PBPTI for $N_{\mathrm{FFT}} = 8$, $K = 2$, and $\mathbf{\Phi} = [0.5, 0.5]$.

where the subscript of $\mathbf{z}$ denotes the OFDM symbol index.

We assume that the system only knows the AWGN variance, σ_{w}^2, and it remains constant during transmission. The variance can be estimated from the guard tones where no signals are transmitted or from previous OFDM transmissions. The receiver knows neither the frequency channel H_k nor its PDP. In addition, the receiver does not know the average power of interference, σ_{I}^2, or its profile vector $\mathbf{z}$. in the low interference power regime, the receiver can simply treat interference as AWGN and no interference detection is required. Consequently, we are interested in the regime of high interference power. More specifically, we assume that σ_{I}^2 is higher than the signal power.

III. Robust Receiver Algorithms

In this section, we present the receiver algorithms consisting of channel estimation and symbol detection. Note that the interference and the AWGN form an equivalent Gaussian mixture noise, which depends on the interference statistics $[\chi, \omega, p, K, \mathbf{\Phi}, \sigma_{\mathrm{I}}^2]$. Since these statistics are not accessible at the receiver, it is of high complexity to estimate all of them. For example, estimating ω and K needs two dimensional search over all N_{FFT} tones. The complexity is at least $\mathcal{O}(N_{\mathrm{FFT}}^2)$. Our proposed algorithm detects a subset of all interference during the channel estimation stage. The unknown interference statistics and channel estimation errors are then resolved by robust prior LLR calculation and soft iterative decoding. The complexity of this approach can be shown to be $\mathcal{O}(N_{\mathrm{FFT}})$ or $\mathcal{O}(N)$. We first focus on an open-loop receiver which has three components: interference detection, channel estimation, and soft LLR decoder.

A. Interference Detection

The receiver detects interference using both pilot and data tones. Although pilot-only detection intuitively provides better detection performance, it requires the interference coherence bandwidth be wider than the pilot spacing. Otherwise, a misdetection may occur when interference only corrupts data tones. Note that the corrupted tones have higher amplitude than clean tones when the interference power is much higher than signal power. We use a threshold detection method by searching the received spectrum for $|y_k|^2 > \lambda\sigma_{\mathrm{w}}^2$, where λ is a design parameter. Denote the set of the resulting index as $\mathcal{N}_c$. When $\mathcal{N}_c$ is an empty set, the OFDM symbol is detected to be clean. Otherwise, it is detected as corrupted by jamming signals and the values in $\mathcal{N}_c$ indicate tones where interference is present.

If the profile vector of interference $\mathbf{z}$ and interference power σ_{I}^2 are known, the threshold can be optimally decided [8]. Since the receiver does not have this information, λ is computed by constraining the false alarm probability. The clean received tone signal y_k is $\mathcal{CN}(0, |s_k|^2 + \sigma_{\mathrm{w}}^2)$ distributed when H_k is unknown. Let the constrained false alarm probability be $\mathrm{P_F}$. The threshold can be calculated using the fact that $\beta = |y_k|^2$ is exponentially distributed,

$$\mathrm{P_F} > \int_{\lambda\sigma_{\mathrm{w}}^2}^{\infty} \frac{1}{|s_k|^2 + \sigma_{\mathrm{w}}^2} \exp\left(-\frac{\beta}{|s_k|^2 + \sigma_{\mathrm{w}}^2}\right) \mathrm{d}\beta, \quad s_k \in \mathcal{S},$$

$$\lambda = \left(\frac{\max\limits_k |s_k|^2}{\sigma_{\mathrm{w}}^2} + 1\right) \log\frac{1}{\mathrm{P_F}},$$

where maximization is taken over all constellation points.

It needs to be emphasized that the proposed method only detects a subset of all interference conditions. When the power of interference is not strong on some tones, misdetection can also occur. However, detection error is tolerable and is taken care of by the soft iterative decoding.

B. Pilot Erasure and Channel Estimation

OFDM channel estimation has been widely discussed for systems without interference. When the receiver knows the PDP of channels, optimal Wiener filtering can be used to estimate the frequency channels. When the receiver has no statistical information about the channel, a least-squares (LS) estimation with linear interpolation is typically implemented. Other linear interpolation algorithms, e.g. [12], can be used similarly but require some statistical knowledge of channels. Using the idea of erasure insertion [13], we consider a low-complexity LS algorithm with pilot erasure.

The receiver first erases the pilot tones detected as being corrupted. The index set for the remained pilots can be expressed as $\mathcal{N}_{\mathrm{P}}' = \mathcal{N}_{\mathrm{P}} \setminus \mathcal{N}_c$. The channels for all data tones are recovered by linear interpolation of the receive pilot signals from $\mathcal{N}_{\mathrm{P}}'$. Let k_1 and k_2 be two neighboring pilot indexes in $\mathcal{N}_{\mathrm{P}}'$ such that $k_1 < k < k_2$. A linear interpolation for H_k based on the LS estimates of H_{k_1} and H_{k_2} can be performed as

$$\hat{H}_k = y_{k_1} + \frac{k - k_1}{k_2 - k_1}(y_{k_2} - y_{k_1}),\ k_1, k_2 \in \mathcal{N}_{\mathrm{P}}', k \in \mathcal{N}_{\mathrm{D}}. \quad (2)$$

Here, we assumed every pilot tone is identical to be a constellation of 1.

C. LLR Calculation and Soft Decoder

Let $\mathbf{y}$ and $\mathbf{H}$ denote $1 \times N_{\mathrm{D}}$ vectors of received data tone signals and channels for data tones, respectively. The soft decoder iteratively updates the LLR for each coded bit b_i

$$\mathrm{LLR}(b_i) = \log \frac{\mathrm{P}(b_i = 0|\mathbf{y}, \mathbf{H})}{\mathrm{P}(b_i = 1|\mathbf{y}, \mathbf{H})}, \tag{3}$$

based on the observation of $\mathbf{y}$ and channels. In practice, the estimated channels $\hat{H}_k$ are used to replace H_k for LLR calculations. Let the estimated channel be $\hat{H}_k = H_k + \epsilon_k$, where ϵ_k denotes estimation error modeled as $\mathcal{CN}(0, \sigma_{\mathrm{H}}^2)$ distributed. The equivalent system equation in (1) can be rewritten as

$$y_k = \hat{H}_k s_k \underbrace{- \epsilon_k s_k + w_k + I_k}_{n_k}, \qquad k \in \mathcal{N}_{\mathrm{D}}, \tag{4}$$

where n_k denotes the equivalent noise including channel estimation error, AWGN, and unknown interference. Given s_k and interference statistics, the equivalent noise n_k is $\mathcal{CN}(0, |s_k|^2\sigma_{\mathrm{H}}^2 + \sigma_{\mathrm{w}}^2 + \varsigma_k^2)$ distributed, where ς_k^2 denotes the variance of interference I_k. When the receiver knows ς_k^2, the prior LLR for the bit b_i contained in y_k can be calculated as

$$\mathrm{LLR}(b_i) = \log \frac{\mathrm{f}(y_k|b_i = 0, \hat{H}_k)}{\mathrm{f}(y_k|b_i = 1, \hat{H}_k)} = \log \frac{\sum\limits_{s_k:b_i=0} \mathrm{f}(y_k|s_k, \hat{H}_k)}{\sum\limits_{s_k:b_i=1} \mathrm{f}(y_k|s_k, \hat{H}_k)}$$

$$= \log \frac{\sum\limits_{s_k:b_i=0} \frac{1}{|s_k|^2\sigma_{\mathrm{H}}^2+\sigma_{\mathrm{w}}^2+\varsigma_k^2} \exp\left(-\frac{\left|y_k - \hat{H}_k s_k\right|^2}{|s_k|^2\sigma_{\mathrm{H}}^2+\sigma_{\mathrm{w}}^2+\varsigma_k^2}\right)}{\sum\limits_{s_k:b_i=1} \frac{1}{|s_k|^2\sigma_{\mathrm{H}}^2+\sigma_{\mathrm{w}}^2+\varsigma_k^2} \exp\left(-\frac{\left|y_k - \hat{H}_k s_k\right|^2}{|s_k|^2\sigma_{\mathrm{H}}^2+\sigma_{\mathrm{w}}^2+\varsigma_k^2}\right)}, \tag{5}$$

where $\mathrm{f}(\cdot)$ denotes probability density function, and summation is over the constellation points with a constraint on b_i. For convenience, we call this LLR calculation the *genie LLR* since interference statistics are known.

When the receiver does not know ς_k^2, a robust method was proposed in [4] to obtain the prior LLRs. The maximum-likelihood (ML) estimate of ς_k^2 for $\mathrm{P}(y_k|b_i = 0, \hat{H}_k)$ is used on the numerator in (5), and that for $\mathrm{P}(y_k|b_i = 1, \hat{H}_k)$ is used on the denominator in (5). In other words,

$$\hat{\varsigma}_k^2 = \arg\max_{\varsigma_k^2} \frac{\sum\limits_{s_k:b_i} \exp\left(-\frac{\left|y_k - \hat{H}_k s_k\right|^2}{|s_k|^2\sigma_{\mathrm{H}}^2+\sigma_{\mathrm{w}}^2+\varsigma_k^2}\right)}{|s_k|^2\sigma_{\mathrm{H}}^2 + \sigma_{\mathrm{w}}^2 + \varsigma_k^2}. \tag{6}$$

While this method can be used for a BPSK constellation, for other higher-order constellations, obtaining the ML estimates requires finding the roots for nonlinear polynomials. Here, we propose an alterative method to use the ML estimate of ς_k^2 for $\mathrm{P}(y_k|s_k, \hat{H}_k)$ in each individual term inside the summation,

$$\hat{\varsigma}_k^2 = \arg\max_{\varsigma_k^2} \frac{\exp\left(-\frac{\left|y_k - \hat{H}_k s_k\right|^2}{|s_k|^2\sigma_{\mathrm{H}}^2+\sigma_{\mathrm{w}}^2+\varsigma_k^2}\right)}{|s_k|^2\sigma_{\mathrm{H}}^2 + \sigma_{\mathrm{w}}^2 + \varsigma_k^2}. \tag{7}$$

By setting the derivative of the above equation with respect to ς_k^2 to zero, it follows that $\hat{\varsigma}_k^2 = \frac{1}{\left|y_k - \hat{H}_k s_k\right|^2} - |s_k|^2\sigma_{\mathrm{H}}^2 + \sigma_{\mathrm{w}}^2$. Replacing ς_k^2 in each term of the summation in (5) with $\hat{\varsigma}_k^2$, we obtain a new prior LLR as

$$\mathrm{LLR}(b_i) = \log \frac{\sum\limits_{s_k:b_i=0} \frac{1}{\left|y_k - \hat{H}_k s_k\right|^2}}{\sum\limits_{s_k:b_i=1} \frac{1}{\left|y_k - \hat{H}_k s_k\right|^2}}. \tag{8}$$

The above calculation does not require the knowledge of ς_k^2, hence is called *robust LLR*.

For the OFDM system under consideration, the prior LLRs are calculated based on the results from interference detection. For the tones that are not detected as being corrupted, i.e., $k \in \mathcal{N}_{\mathrm{D}} \setminus \mathcal{N}_{\mathrm{c}}$, the genie LLR in (5) is performed with $\varsigma_k^2 = 0$. For the tones estimated as being corrupted, i.e., $k \in \mathcal{N}_{\mathrm{D}} \bigcap \mathcal{N}_{\mathrm{c}}$, the robust LLR in (8) is used. This LLR calculation method is called *dynamic LLR*. Since the dynamic LLR switches between the robust LLR and the genie LLR based on interference detection for each data tone, it has no extra complexity compared to the robust LLR or the genie LLR. As shown in Fig. 1, the prior LLR vectors from each OFDM symbol are concatenated and deinterleaved. The resulting LLR vector is decoded by the sum-product algorithm in the log domain [14].

The complexity of the proposed receiver algorithm is presented in Table. I. Each component has linear processing complexity in the size of FFT or LDPC block length. As a result, the overall complexity of the proposed receiver architecture is linear in the coded block length as well as the FFT size.

TABLE I
RECEIVER ALGORITHM COMPLEXITY

INT DET	LSChanEst	prior LLR	deinterleaver	LDPC decoding
$\mathcal{O}(N_{\mathrm{FFT}})$	$\mathcal{O}(N_{\mathrm{FFT}})$	$\mathcal{O}(\lvert\mathcal{S}\rvert N)$	$\mathcal{O}(N)$	$\mathcal{O}(N)$

IV. DECISION FEEDBACK

In this section, we describe the decision feedback method to improve channel estimation. The LDPC codes may not be successfully decoded after reaching the maximum number of iterations. However, some bits are decoded with high reliability and can be considered as pilots to improve channel estimation for data tones. We discuss a new linear interpolation method using decoded symbols.

We focus on BPSK modulations because the extension to other modulation formats is straightforward. The decoded bits with posterior LLR higher than θ are reinterleaved and modulated. We denote the resulting set of data tone indices as $\mathcal{N}_{\mathrm{F}}$. A subset of these data tones excluding those being detected as corrupted can be reused as pilot tones as

$$\mathcal{N}_{\mathrm{I}} = \mathcal{N}_{\mathrm{F}} \setminus \mathcal{N}_{\mathrm{c}}. \tag{9}$$

Assume that the feedback symbols and interference are correctly detected. A new estimate of H_k can be obtained as

$$\tilde{H}_k = \frac{y_k \tilde{s}_k^*}{\left|\tilde{s}_k\right|^2} = H_k + \frac{w_k \tilde{s}_k^*}{\left|\tilde{s}_k\right|^2}, \qquad k \in \mathcal{N}_{\mathrm{I}}. \tag{10}$$

A simple idea to improve channel estimation is using new pilot signals $\mathcal{N}_{\mathrm{P}}' \bigcup \mathcal{N}_{\mathrm{I}}$ to interpolate the channels of data tones. However, this method turns out to be ineffective from simulations. We propose a new linear interpolation method to refine channel estimates for tones in $\mathcal{N}_{\mathrm{I}}$.

Note that from (2), channels for data tones are linearly interpolated from pilots in $\mathcal{N}_{\mathrm{P}}'$. Estimation error of $\hat{H}_k$ contains two components: LS estimation error on pilots and interpolation mismatch error. To clarify, Eqn. (2) can be rewritten as

$$\hat{H}_k = H_k + \underbrace{\frac{k_2 - k}{k_2 - k_1} w_{k_1} + \frac{k - k_1}{k_2 - k_1} w_{k_2}}_{e_{\rm E}} + e_{\rm I},\ k \in \mathcal{N}_{\rm D}. \quad (11)$$

The estimation error $e_{\rm E}$ is raised by the LS estimation error w_{k_1} and w_{k_2} on pilot tones, while the interpolation error $e_{\rm I}$ is due to the mismatch between linear interpolation and the real channel value. Since the estimation error w_k in (10) is independent from w_{k_1} and w_{k_2} in (11), $\hat{H}_k$ can be linearly combined with $\tilde{H}_k$ to refine channel estimation as

$$\underline{H}_k = a\tilde{H}_k + b\hat{H}_k, \qquad k \in \mathcal{N}_{\rm I}, \quad (12)$$

where a and b denote the combination coefficients. To obtain an unbiased estimation, it is required that $a + b = 1$. In the case of $a = 1$ and $b = 0$, only the new channel estimate is used. *Two* methods are considered:

1) Method 1: the receiver uses arithmetic mean for channel estimation refinement, i.e., $a = b = 0.5$.
2) Method 2: the variance of the estimation error of $\underline{H}_k$ is minimized[1].

From (10), (11), and (12), an optimization problem can be formulated as

$$\min \left(a^2 + b^2 \underbrace{\left(\left(\frac{k_2 - k}{k_2 - k_1} \right)^2 + \left(\frac{k - k_1}{k_2 - k_1} \right)^2 \right)}_{\vartheta} \right) \sigma_{\rm w}^2, \quad (13)$$
$$\text{s.t. } a + b = 1.$$

The above optimization problem is quadratic and convex. Its global optimal solution exists and can be expressed as $a = 1/(1 + \frac{1}{\vartheta}), b = 1/(1 + \vartheta)$. It should be noted that the linear combination coefficients are independent of $\sigma_{\rm w}^2$. They can be determined independently and stored at the receiver.

Now the new set of pilot index can be obtained as $\mathcal{N}_{\rm P} = \mathcal{N}_{\rm P} \bigcup \mathcal{N}_{\rm I}$, and channels for the remaining data tones $\mathcal{N}_{\rm D} \setminus \mathcal{N}_{\rm I}$ can be estimated using linearly interpolation in (2). New channel values are used to calculate the prior LLR of each bit, and soft decoding is performed using new prior LLRs.

For the n-th iteration feedback, only the data tones not appearing in the previous iterations are added, i.e., $\mathcal{N}_{\rm I}^{(n)} = \left(\mathcal{N}_{\rm F}^{(n)} \setminus \mathcal{N}_{\rm c}\right) \setminus \mathcal{N}_{\rm I}^{(n-1)}$. This ensures independence between the pilot estimation error of $\tilde{H}_k^{(n)}$ and that of $\hat{H}_k^{(n-1)}$. Linear combination of $\tilde{H}_k^{(n)}$ and $\hat{H}_k^{(n-1)}$ as in (12) can be performed by tracking the updated variance of estimation errors.

V. Simulation Results

This section provides simulated BER performance for the proposed robust receiver algorithm in comparison with genie systems. For the first genie system, denote as *NoInt, CSIR*, the received signals are not corrupted by interference and the receiver has perfect channel state information (CSIR). The second genie system, denote as ***CSIR, ISIR***, is corrupted by interference, and the receiver knows CSIR and interference state information (ISIR) $[\chi, \omega, p, K, \mathbf{\Phi}, \sigma_{\rm I}^2]$. For both genie systems, no interference detection and channel estimation are needed. The prior LLRs are calculated using the genie LLR in (5). For our proposed OFDM systems, the power of pilot tones is 3 dB higher than that of the data tones. The system setups are summarized in Table II. Since the power of data tones is normalized, the signal-to-noise ratio (SNR) and interference-to-signal ratio (ISR) can be calculated as $\text{SNR} = \frac{1}{\sigma_{\rm w}^2}$ and $\text{ISR} = \sigma_{\rm I}^2$, respectively. In the following figures, the horizontal and vertical axes represent the bit energy to noise PSD ratio ($E_{\rm b}/N_0$), measured in dB, and BER, respectively.

TABLE II
System Parameters

Parameters	Values
LDPC Soft decoding	$\gamma = 3$, $\rho = 6$, $M = 1540$, $N = 3080$, $R = 0.5$. Max # of Iterations=40
OFDM systems	$N_{\rm FFT} = 1024$, $N_{\rm P} = 71$,$N_{\rm D} = 770$ # of left guards= 91, # of right guards= 91, # of DC=1
ISI Channels	ITU-PED-B
Interference	$p = 0.5$, $\mathbf{\Phi} = [\frac{1}{K}, \ldots, \frac{1}{K}]$
Receiver	$\lambda = 3(\text{SNR} \max \lvert s_i \rvert^2 + 1)$, $\theta = 6$, Max # of feedback iterations=4

Fig. 3 compares the BER performance of our proposed systems, RoLLR and DynLLR, with two genie systems. The RoLLR system uses robust LLR for all data tones, while the DynLLR system calculates dynamic LLR described in Section III-C. For BPSK, QPSK, and 16-QAM constellations, the BER of the DynLLR system is approximately 2 dB away from the genie system with interference, and 3 dB far from the genie system without interference. In addition, the DynLLR system has approximately 2 dB gain compared to the RoLLR system for simulated constellations.

Fig. 4 compares the BER performance of the DynLLR system with genie systems for four groups of $[K, \text{ISR}]$ when BPSK is used. It can be observed that the BER of DynLLR system is degraded when K increases from 50 to 100. This is because more tones are corrupted by interference. The BER of the DynLLR system with $[K = 100, \text{ISR} = 10\text{dB}]$ is the worst among the four groups, because interference detection incurs more errors for low ISR.

Fig. 5 demonstrates the BER comparison among the close-loop, open-loop and genie systems. The two proposed feedback methods in Section IV (labeled as 'FB, M1' and 'FB, M2') have approximately the comparable performance. They also have 0.5 dB gain for $[K = 50, \text{ISR} = 20\text{dB}]$ and 1 dB gain for $[K = 100, \text{ISR} = 10\text{dB}]$ compared to the open-loop system. The extra feedback complexity brings BER improvement.

[1]This minimization does not include the interpolation mismatch error, since its variance, depending on the channel PDP, is not known to the receiver.

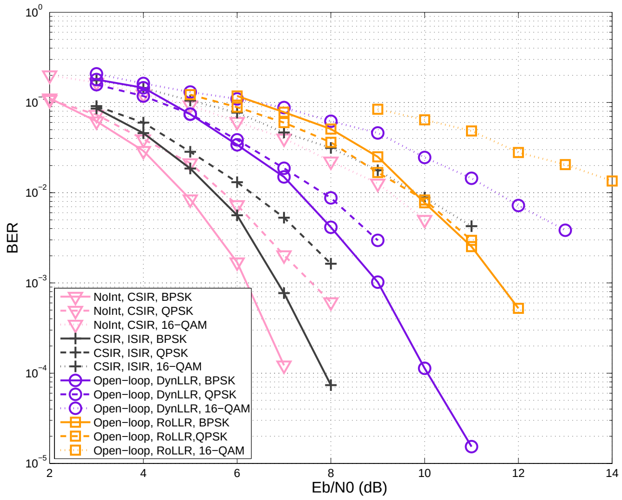

Fig. 3. BER comparison with genie systems for different constellations at $K = 50$ and $\mathrm{ISR} = 20$dB.

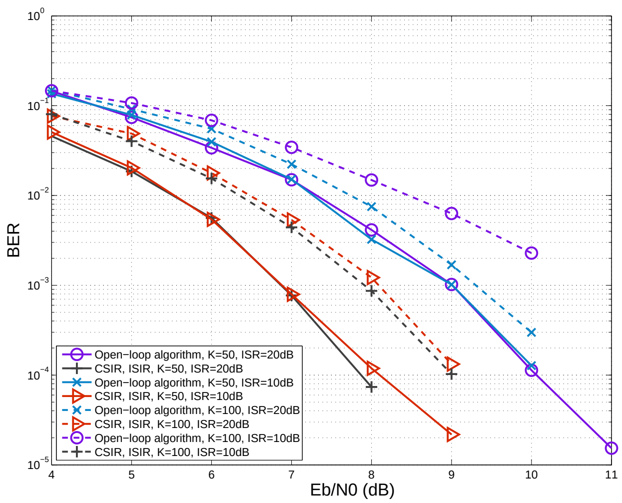

Fig. 4. BER comparison with genie systems for different $[K, \mathrm{ISR}]$ for BPSK.

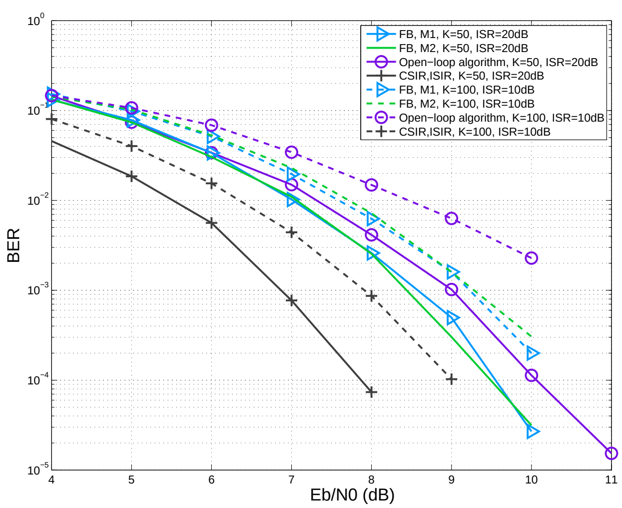

Fig. 5. BER comparison with decision feedback methods for BPSK.

VI. Conclusions

This paper is concerned with interference mitigation, channel estimation, and information decoding for LDPC-coded OFDM systems subject to PBPTI. Interference is detected using threshold methods, and unknown channels are estimated using linear interpolation. To resolve the unknown interference spectrum, we have extended the previously proposed robust LLR calculation for BPSK to QPSK and 16-QAM constellations, and investigated dynamic LLR approaches. Although the presentation and simulations assume the PSD of interference $\mathbf{\Phi}$ to be constant over time, the dynamic LLR approaches can also be used for the case of varying interference PSD, thus simplifying the interference spectrum detection problem. In addition, we have considered decision feedback to improve channel estimation. Two linear combination methods to refine channel estimation for new pilots have been proposed. The BER performance of the proposed systems are approximately $2 \sim 3$dB from the genie systems where CSIR and ISIR are perfectly known. The proposed robust and dynamic LLRs are not restricted to OFDM systems, and can be used by any soft decoding receivers to estimate the channel and to detect the signals in the presence of unknown interference. Future work needs to consider analytical quantification of the performance of the proposed algorithms.

References

[1] M. Morelli and M. Moretti, "Channel estimation in OFDM systems with unknown interference," *IEEE Trans. Wireless. Commun.*, vol. 8, pp. 5338–5347, Oct. 2009.

[2] S. Kalyani and K. Giridhar, "Extreme value theory based OFDM channel estimation in the presence of narrowband interference," in *IEEE GLOBECOM 2006*, Dec. 2006, pp. 1 –5.

[3] A. Ayyar, M. Lentmaier, K. Giridhar, and G. Fettweis, "Robust initial LLRs for iterative decoders in presence of non-Gaussian noise," in *IEEE ISIT 2009*, Jul. 2009, pp. 904 –908.

[4] H. E. Gamal and E. Geraniotis, "Iterative channel estimation and decoding for convolutionally coded anti-jam FH signals," *IEEE Trans. Commun*, vol. 50, pp. 321–331, Feb. 2002.

[5] S. Kalyani and K. Giridhar, "Interference mitigation in turbo-coded OFDM systems using robust LLRs," in *IEEE ICC 2008*, May 2008, pp. 646 –651.

[6] R. Yazdani and M. Ardakani, "Efficient LLR calculation for non-binary modulations over fading channels," *IEEE Tran. Commun.*, vol. 59, pp. 1236 – 1241, May 2011.

[7] L. Sanguinetti, M. Morelli, and H. V. Poor, "BICM decoding of jammed OFDM transmissions using the EM algorithm," *IEEE Trans. Wireless Commun.*, vol. 10, pp. 2800 – 2806, Sep. 2011.

[8] T. Li, W. H. Mow, V. K. N. Lau, M. Siu, R. S. Cheng, and R. D. Murch, "Robust joint interference detection and decoding for OFDM-based cognitive radio systems with unknown interference," *IEEE Journal on Selected Areas in Communications*, pp. 566–575, Apr. 2007.

[9] A. Batra and J. Zeidler, "Narrowband interference mitigation in OFDM systems," in *Proc. of IEEE Military Communications Conference.*, San Diego, CA, Nov. 2008.

[10] D. J. C. MacKay and R. M. Neal, "Near Shannon limit performance of low density parity check codes," *Electronics Letters*, vol. 32, pp. 1645–1646, 1996.

[11] N. Bonello, C. Sheng, and L. Hanzo, "Low-density parity-check codes and their rateless relatives," *IEEE Communications Surveys & Tutorials*, vol. 13, pp. 3–26, 2011.

[12] X. Dong, W. Lu, and A. Soong, "Linear interpolation in pilot symbol assisted channel estimation for OFDM," *IEEE Trans. Wireless. Comm.*, vol. 6, pp. 1910 –1920, 2007.

[13] C. W. Baum and M. B. Pursley, "Bayesian methods for erasure insertion in frequency-hop communication systems with partial-band interference," *IEEE Trans. Commun*, vol. 40, pp. 1231–1238, Jul. 1992.

[14] S. Johnson, *Iterative Error Correction: Turbo, Low-Density Parity-Check and Repeat-Accumulate Codes*. Cambridge University Press, 2010.