NON-BINARY LDPC CODED OFDM IN IMPULSIVE POWER LINE CHANNELS

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ABSTRACT

In this paper, we propose Irregular Non-Binary Low Density Parity Check (IR-NB-LDPC) codes with Signed Log Fast Fourier Transform (SL-FFT) decoding algorithm to overcome the harsh environment of power-line communication (PLC) channels. Their performance is compared with Irregular Binary LDPC (IR-B-LDPC) codes using the Sum Product algorithm (SPA). The sparse parity check matrix $\mathbf{H}$ of both codes are constructed using progressive edge growth (PEG) algorithm with a novel initialization of the *apriori* log likelihood ratios (LLR) of each decoder to mitigate the highly impulsive noise in PLC channels. Numerical performance results obtained via simulations show that the proposed system at bit error rate of 10^{-4} achieves a coding gain of more than 21 dB compared to uncoded system and more than 6 dB compared to IR-B-LDPC codes for the same block length in bits and rates, however, this is accomplished with higher decoding complexity.

Index Terms— Non-binary LDPC, binary-LDPC, log likelihood ratios, power-line communication channels, OFDM.

1. INTRODUCTION

Demand is increasing for the use of power-line communication (PLC) channels as effective media for communication such as high speed data transmission and voice signals, since it offers economical communication with cheap installation and reliable connection throughout buildings. However, the characteristics of PLC are not specially designed for communication, and it is a harsh medium for high speed data transmission due to high attenuation, frequency selectivity and impulsive noise [1, 2]. The degradation in bit error rate (BER) can be reduced using LDPC codes with long block length and soft decision decoding algorithms [3], however, this results in higher decoder complexity. In [4], a method is proposed to reduce this complexity by constructing an optimum sparse parity check matrix $\mathbf{H}$ for short and intermediate block lengths using a progressive edge growth (PEG) algorithm. However, in impulsive PLC channels, the noise is no longer Gaussian [3], and a simple estimator is not suitable to calculate the initial values of LLR. Therefore, the BER performance of LDPC codes will be degraded. Various techniques have been suggested to mitigate the impulsive noise. These include: blanking, clipping and combined blanking clipping non-linearity [5]; using irregular quasicyclic low density parity check (QC-LDPC) codes [6]; or concatenation coding schemes between outer Luby transform (LT) codes and inner LDPC codes [7]. Several drawbacks of these techniques such as its effect on the received signal or increase the system complexity. Nakagawa et al. [3] proposed an LLR calculation for binary LDPC codes by substituting the probability density function (PDF) of impulsive noise in the LLR calculation. In practice, this is unrealistic as the substitution ignored the effect of the multipath PLC channel.

In this paper, we derive the log likelihood ratio (LLR) for B-LDPC/NB-LDPC decoders to overcome PLC introduced impairments. Simulation results in highly impulsive PLC channels show that the IR-NB-LDPC codes with a higher Field size $\mathbb{F}_{16}$ improve the system performance compared to the IR-B-LDPC codes and perform better than all previously published techniques [3, 5–7]

The rest of the paper is organized as follows. Section 2 focuses on the system model. The novel application of the derived LLR computations of B-LDPC/NB-LDPC codes over impulsive PLC channels are derived in Section 3. Section 4 presents the LDPC decoding algorithms. Section 5 presents simulation results of the proposed IR-NB-LDPC compared with IR-B-LDPC for coded orthogonal frequency division multiplexing (COFDM) systems in highly impulsive PLC channels. Finally, conclusions are drawn in Section 6.

2. SYSTEM MODEL

A conventional COFDM based PLC system is considered in [5]. After cyclic prefix (CP) removal and Fast Fourier Transform (FFT), the received signal can be expresed as

$$Y_n = H_n D_n + W_n, \ \ \forall n = 0, 1, \ldots, N-1, \qquad (1)$$

where N represents the number of sub-carriers, D_n are the information symbols, and W_n represents the FFT of noise sam-

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ples in time domain, w_k. The noise in PLC channels is a non-Gaussian, i.e. it comprises a mixture of background noise with low power spectral density (PSD) and impulsive noise with high PSD in short bursts. Therefore, it can be modelled as Bernoulli-Gaussian noise mixture that is a special case of the Middleton class-A noise model. At the receiver, the complex PDF of the noise w_k in time domain can be expressed as $p_{w_k}(w_k^R, w_k^I)$ where the w_k^R and w_k^I are the real and imaginary parts of w_k, respectively as given by [8]

$$p_{w_k}(w_k^R, w_k^I) = (1-\alpha)\mathcal{N}(w_k^R, 0, \sigma_w^2)\mathcal{N}(w_k^I, 0, \sigma_w^2) + \alpha\mathcal{N}(w_k^R, 0, \sigma_w^2+\sigma_i^2)\mathcal{N}(w_k^I, 0, \sigma_w^2+\sigma_i^2), \quad (2)$$

where $\mathcal{N}(.)$ denotes the Gaussian PDF, σ_w^2 is the AWGN variance, σ_i^2 is the impulsive noise variance and α is the mixing factor corresponding to the probability of occurrence of impulsive noise. It's worth noting that both PDFs involved in the mixture are zero-mean. Furthermore, H_n represents the frequency response of the multipath PLC model in the frequency band $(0.5-20)$ MHz as given by [2]

$$H(f) = \sum_{l=1}^{L} g_l e^{-(a_0+a_1 f^k)d_l} e^{-j2\pi f \frac{d_l}{v_p}}, \quad (3)$$

where $H(f)$ is the frequency response, L is the total number of multi-paths, l is the path index, g_l is the weighting factor, a_0 and a_1 are attenuation parameters, $k \in [0.5, 1]$ is the exponent of the attenuation factor, d_l is the path length and v_p is the phase velocity of the wave. For the exact parameters of the model, please refer to [2].

3. LLR COMPUTATION

3.1. Binary LDPC (B-LDPC) codes

The optimal soft information is the *a posteriori* log-likelihood ratio (LLR) that is equivalent to the maximum likelihood detector. Therefore, the modified log-likelihood ratios (LLRs) of SPA algorithm in impulsive PLC channels can be derived by including the joint effects of (2) and (3) in the computation of the LLRs in order to achieve improved BER performance. The LLR on the m-th bit can be expressed as

$$\mathrm{LLR}(\mathcal{X}_\mathrm{m}|\mathrm{Y_n}) = \log\left(\frac{\sum_{D_n\in\mathcal{X}_m^{(0)}} e^{-|Y_n-H_nD_n|^2/N_t)}}{\sum_{D_n\in\mathcal{X}_m^{(1)}} e^{-|Y_n-H_nD_n|^2/N_t)}}\right), \quad (4)$$

where $\mathcal{X}_m$ is the m-th bit of the transmitted symbol D_n and N_t is the total noise power which include the Gaussian noise power N_0 and impulsive noise power N_i, i.e.

$$N_t = N_0 + \alpha_i \cdot N_i = N_0(1+\alpha_i/\Gamma), \quad (5)$$

where $\Gamma = \sigma_g^2/\sigma_i^2$ is the Gaussian-to-impulsive noise power ratio and $\sigma_t^2 = N_t/2$. Therefore, the LLRs derivation for $4QAM$ modulation depending on the constellation mapping $C_{4QAM} = [-1-j, -1+j, 1-j, 1+j]$, for $n = 0, 1, \ldots, (L_c-1)/\log_2(M)$, where L_c is the code length of the B-LDPC code and $M = 4$ is the constellation order, are given as

$$\mathrm{LLR}(b_0(n)) = -2(Y_r(n)\,H_r(n) + Y_i(n)\,H_i(n))/\sigma_t^2, \quad (6)$$

$$\mathrm{LLR}(b_1(n)) = 2(Y_r(n)\,H_i(n) - Y_i(n)\,H_r(n))/\sigma_t^2, \quad (7)$$

$$\mathrm{LLR}(s(n)) = [\mathrm{LLR}(b_0(n)), \mathrm{LLR}(b_1(n))]. \quad (8)$$

It is worth mentioning that the n-th index will be dropped for simplicity for the remaining equations in this paper. The LLRs derivation for $16QAM$ modulation depending on the constellation mapping $C_{16QAM} = [-3+3j, -3+1j, -3-3j, -3-1j, -1+3j, -1+1j, -1-3j, -1-1j, 3+3j, 3+1j, 3-3j, 3-1j, 1+3j, 1+1j, 1-3j, 1-1j]$, for $k = 0, 1, \ldots, (L_c-1)/4$, are given as

$$\mathrm{LLR}(b_0) = -2(Y_rH_r + Y_iH_i)/\sigma_t^2, \quad (9)$$

$$\mathrm{LLR}(b_1) = \max\{-2(Y_rH_r + 2H_r^2 + Y_iH_i + 2H_i^2)/\sigma_t^2, 2(Y_rH_r - 2H_r^2 + Y_iH_i - 2H_i^2)/\sigma_t^2\}, \quad (10)$$

$$\mathrm{LLR}(b_2) = 2(-Y_rH_i + Y_iH_r)/\sigma_t^2, \quad (11)$$

$$\mathrm{LLR}(b_3) = \max\{2(Y_rH_i - 2H_r^2 - Y_iH_r - 2H_i^2)/\sigma_t^2, -2(Y_rH_i + 2H_r^2 - Y_iH_r + 2H_i^2)/\sigma_t^2\}, \quad (12)$$

$$\mathrm{LLR}(s) = [\mathrm{LLR}(b_0), \ldots, \mathrm{LLR}(b_3)], \quad (13)$$

where Y_r, Y_i, H_r and H_i are the real and imaginary parts of the received signal and of the frequency response of the PLC channel, respectively, with H_r and H_i computed using (3). Furthermore, $\mathrm{LLR}(b_0), \ldots, \mathrm{LLR}(b_m)$ and $\mathrm{LLR}(s)$ are the soft LLRs corresponding to the bits $b_0, \ldots, b_m$ and the symbol s, respectively, where $m = \log_2(M)$ is the number of bits to represent one symbol in the constellation map.

3.2. Non-binary LDPC (NB-LDPC) codes

The modified LLRs of signed log-FFT decoder can be derived by including the effect of (2) and (3) in (14) as

$$F_n^a = \log\left(\frac{e^{-|Y_n-H_nD_n|^2/N_t}|D_n = a}{e^{-|Y_n-H_nD_n|^2/N_t}|D_n = 0}\right), \; a \in \mathbb{F}_q\backslash\{0\}, \quad (14)$$

Therefore, the modified LLRs of previous $4QAM$ constellation points can be given as

$$F^0 = \mathbf{0}, \quad (15)$$

$$F^1 = 2\left(Y_iH_r - Y_rH_i\right)/\sigma_t^2, \quad (16)$$

$$F^2 = 2\left(Y_rH_r + Y_iH_i\right)/\sigma_t^2, \quad (17)$$

$$F^3 = 2(Y_rH_r + Y_iH_r - Y_rH_i + Y_iH_i)/\sigma_t^2. \quad (18)$$

Furthermore, the modified LLRs of previous $16QAM$ constellation points can written as

$$F^0 = \mathbf{0}, \tag{19}$$
$$F^1 = 2(Y_r H_i - Y_i H_r + 2H_r^2 + 2H_i^2)/\sigma_t^2, \tag{20}$$
$$F^2 = 6\left(Y_r H_i - Y_i H_r\right)/\sigma_t^2, \tag{21}$$
$$F^3 = 4(Y_r H_i - Y_i H_r + H_r^2 + H_i^2)/\sigma_t^2, \tag{22}$$
$$F^4 = 2(Y_r H_r + Y_i H_i + 2H_r^2 + 2H_i^2)/\sigma_t^2, \tag{23}$$
$$F^5 = 2(Y_r H_r + Y_r H_i - Y_i H_r + Y_i H_i + 4H_r^2 + 4H_i^2)/\sigma_t^2, \tag{24}$$
$$F^6 = 2(Y_r H_r + 3Y_r H_i - 3Y_i H_r + Y_i H_i + 2H_r^2 + 2H_i^2)/\sigma_t^2, \tag{25}$$
$$F^7 = 2(y_r H_r + 2y_r H_i - 2y_i H_r + y_i H_i + 4H_r^2 + 4H_i^2)/\sigma_t^2, \tag{26}$$
$$F^8 = 6\left(Y_r H_r + Y_i H_i\right)/\sigma_t^2, \tag{27}$$
$$F^9 = 2(3Y_r H_r + Y_r H_i - Y_i H_r + 3Y_i H_i + 2H_r^2 + 2H_i^2)/\sigma_t^2, \tag{28}$$
$$F^{10} = 6(Y_r H_r + Y_r H_i - Y_i H_r + Y_i H_i)/\sigma_t^2, \tag{29}$$
$$F^{11} = 2(3Y_r H_r + 2Y_r H_i - 2Y_i H_r + 3Y_i H_i + 2H_r^2 + 2H_i^2)/\sigma_t^2, \tag{30}$$
$$F^{12} = 4(Y_r H_r + Y_i H_i + H_r^2 + H_i^2)/\sigma_t^2, \tag{31}$$
$$F^{13} = 2(2Y_r H_r + Y_r H_i - Y_i H_r + 2Y_i H_i + 4H_r^2 + 4H_i^2)/\sigma_t^2, \tag{32}$$
$$F^{14} = 2(2Y_r H_r + 3Y_r H_i - 3Y_i H_r + 2Y_i H_i + 2H_r^2 + 2H_i^2)/\sigma_t^2, \tag{33}$$
$$F^{15} = 4(Y_r H_r + Y_r H_i - Y_i H_r + Y_i H_i + 2H_r^2 + 2H_i^2)/\sigma_t^2. \tag{34}$$

It is worth highlighting that when the number of modulation levels M is equal to $\mathbb{F}_q$ there is no information loss and the channel LLRs are directly passed to the decoder. While, in the case of $q > M$ and M divides q, the symbols over $\mathbb{F}_q$ can be generated by the sum of the LLRs of the demodulated symbols. For example, when $M = 4$ and $q = 16$ the LLR of a received symbol being 11 is $F^2 + F^3$.

4. LDPC DECODING

4.1. B-LDPC Codes

B-LDPC codes are a class of linear block codes that can be constructed using sparse parity check matrix $\mathbf{H}$ of dimensions $m \times n$. Irregular LDPC codes using the sum-product algorithm (SPA) can approach the Shannon capacity on the AWGN channel within 0.0045 dB using a large block length [9, 10]. The latency introduced by long block codes can be solved by using the PEG algorithm proposed in [4]. PEG ensures the extrinsic information in the iterative SPA decoder is unaffected and no girth cycles of length four are generated in the Tanner graph. Therefore, the optimum symbol node degree distribution of ones in $\mathbf{H}$ for length $L_c = 1008$ and code rate $R_c = 1/2$ is given as

$$\mathbb{F}_2 = 0.47532x^2 + 0.279537x^3 + 0.0348672x^4 + 0.108891x^5 + 0.101385x^{15} \tag{35}$$

The iterative SPA is listed in Algorithm 1 [9, 10]. This algorithm can be used to decode the B-LDPC codes using the LLRs derivation in (6-13).

Algorithm 1: Sum-Product Algorithm (SPA)

Initialization:
Iteration $= 1$,
$L(D_n) = \log \frac{P(Y_n|D_n=0)}{P(Y_n|D_n=1)}$,
$\lambda_{n\to m}(D_n) = L(D_n)$,
$\Lambda_{m\to n}(D_n) = 0$,

while $\mathbf{H}\hat{\mathbf{D}} \neq \mathbf{0}$ and iteration $\leq$ max iteration **do**

  Update checks to nodes: for each m**, and** $n \in \mathcal{N}(m)$**, compute** $\Lambda_{m\to n}(D_n) = 2\tanh^{-1} \prod_{n'\in\mathcal{N}(m)\setminus n} \tanh\left[\lambda_{n'\to m}(D_{n'})/2\right]$,

  Update nodes to checks: for each n**, and** $m \in \mathcal{M}(n)$**, compute** $\lambda_{n\to m}(D_n) = L(D_n) + \sum_{m'\in\mathcal{M}(n)\setminus m} \Lambda_{m'\to n}(D_n)$,

  For each n**, compute** $\lambda_n(D_n) = L(D_n) + \sum_{m\in\mathcal{M}(n)} \Lambda_{m\to n}(D_n)$

  Decision: $\hat{D}_n = \begin{cases} 0 & \lambda_n(D_n) \geq 0 \\ 1 & \textbf{otherwise}, \end{cases}$

  Iteration = Iteration+1,

end

4.2. NB-LDPC Codes

NB-LDPC codes are a class of linear block codes first proposed by Davey and Mackay [11], when the Galois field $\mathbb{F}_q$ is a binary extension field with order $q = 2^p$. NB-LDPC codes outperform B-LDPC codes for the same length in binary bits and code rate on channels with noise bursts [10–12], but involve higher computational complexity. In practice, NB-LDPC codes are decoded in either the probability or logarithmic domain. The latter has the advantage of reduced complexity and numerical stability. Therefore, the SL-FFT decoding algorithm that exhibits lower decoding complexity compared to other decoding algorithms is utilized in this paper. The optimum symbol node degree distribution of the non-zero elements in $\mathbf{H}$ over $\mathbb{F}_{16}$ for the code length $L_c = 252$ and rate $R_c = 1/2$ is given as [4]

$$\mathbb{F}_{16} = 0.772739x^2 + 0.102863x^3 + 0.113797x^4 + 0.010601x^5 \tag{36}$$

The SL-FFT is listed in Algorithm 2 and 3 [13, 14]. These algorithms can be used to decode NB-LDPC codes incorporating the LLRs derivations in (15-34).

Algorithm 2: LOG-FFT Decoding Algorithm

Initialization:
Iteration $=1$,
$F_n^a = \log\left(\frac{e^{-|Y_n - H_n D_n|^2/N_t}|D_n=a}{e^{-|Y_n - H_n D_n|^2/N_t}|D_n=0}\right)$, $a \in \mathbb{F}_q\backslash\{0\}$,
$R_{m,n}^a = 0$,
$Q_{m,n}^a = F_n^a$,

while $\mathbf{H}\hat{\mathbf{D}} \neq \mathbf{0}$ and iteration $\leq$ max iteration **do**

Permute $Q_{m,n}$ according to $a = \mathbf{H}_{m,n} \in \mathbb{F}_q\backslash\{0\}$
$\widetilde{Q}_{m,n} = \mathcal{P}_a(Q_{m,n}), \ \forall m,n$

Transform to signed-log domain:
$\widetilde{\varphi}_{m,n} = (\widetilde{\varphi}_{m,n}(s), \widetilde{\varphi}_{m,n}(r))$**, with**
$\widetilde{\varphi}_{m,n}(s) = 1, \ \widetilde{\varphi}_{m,n}(r) = \widetilde{Q}_{m,n}, \ \forall m,n$, where s and r are the sign and magnitude of $\widetilde{Q}_{m,n}$

Transform to Fourier domain using Fast Walsh-Hadamard Transform:
$\widetilde{\Phi}_{m,n} = \texttt{FWHT}(\widetilde{\varphi}_{m,n}), \ \forall m,n$

Update check node messages:
$\widetilde{\Theta}_{m,n}(s) = \prod_{k\in\mathcal{N}_m\backslash n} \widetilde{\Phi}_{m,k}(s), \ \forall m,n$
$\widetilde{\Theta}_{m,n}(r) = \sum_{k\in\mathcal{N}_m\backslash n} \widetilde{\Phi}_{m,k}(r), \ \forall m,n$

Take the Inverse Fourier Transform for check nodes:
$\widetilde{\theta}_{m,n} = \texttt{IFWHT}(\widetilde{\Theta}_{m,n}), \ \forall m,n$

Extracted the magnitude using signed-log domain:
$\widetilde{R}_{m,n} = \widetilde{\theta}_{m,n}(r), \ \forall m,n$

Inverse permutation of $\widetilde{R}_{m,n}$ according to $a = \mathbf{H}_{m,n} \in \mathbb{F}_n\backslash\{0\}$,
$R_{m,n} = \mathcal{P}_a^{-1}(\widetilde{R}_{m,n}), \ \forall m,n$

Update variable nodes:
$Q_{m,n} = F_n^a + \sum_{k\in\mathcal{M}_n\backslash m} R_{k,n} - \alpha_{m,n}$,
$\alpha_{m,n} = \max_a Q_{m,n}$

Tentative decoding:
$\hat{D}_n = \arg\max_a F_n^a + \sum_{k\in\mathcal{M}_n} R_{k,n}, \ \forall n$

Iteration = Iteration+1,

end

Algorithm 3: Signed-Log Domain

$z(s) = \begin{cases} x(s) & x(s) = y(s) \ \textbf{ or } \ x(m) \geq y(m) \\ -x(s) & \textbf{otherwise,} \end{cases}$

$\gamma = \begin{cases} 1 & x(s) = y(s) \\ -1 & \textbf{otherwise,} \end{cases}$

$z(m) = \max{}^*[x(m),\ y(m)] + \log(1 + \gamma e^{-|x(m)-y(m)|})$

5. SIMULATIONS AND RESULTS

For fair BER comparison, the rate, code length and fields are constructed to have the same amount of binary information bits. Therefore, a $(1008, 504)$ IR-B-LDPC code constructed using (35) with the iterative SPA algorithm, is compared with a $(252, 126)$ IR-NB-LDPC code constructed using (36) with the iterative SL-FFT algorithm. For both systems, the maximum number of decoding iterations is set to 50. Additionally, 1024 sub-carriers per OFDM symbol and a cyclic prefix of 256 samples are utilized with $4QAM$ and $16QAM$ as modulation with bit random interleaver. The performance is evaluated over two practical multipath PLC channels with 4 and 15 paths [2]. Fig. 1 and 2 demonstrate the BER performance of the proposed IR-NB-LDPC COFDM system compared with IR-B-LDPC COFDM and uncoded systems, respectively. The proposed LLRs calculated using (6 - 13) and (15 - 34) after substitution of $\alpha = 0$ in (5) work efficiently with the SL-FFT decoder and give approximately 7 dB coding gain compared to SPA decoder and more than 21 dB compared to the uncoded multipath PLC channels at $P_e = 10^{-4}$ for all cases.

Moreover, both LDPC decoders are examined in highly impulsive PLC channels for $\alpha = 0.3$ and $\Gamma = 0.1$ implying that the impulsive noise is 10 times stronger than the Gaussian noise. The proposed LLRs calculated using (6 - 13) and (15 - 34) after substitution of α and Γ in (5) result in supe-

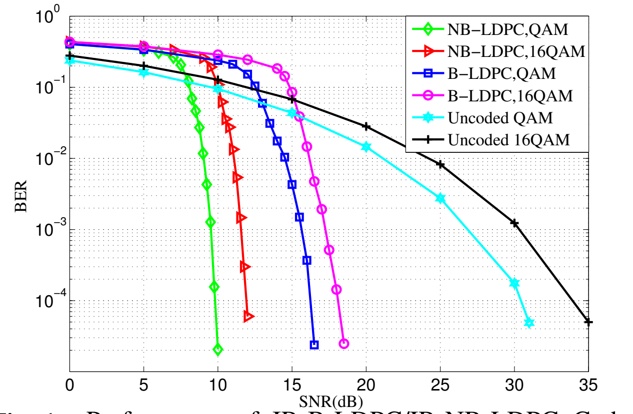

Fig. 1: Performance of IR-B-LDPC/IR-NB-LDPC Coded OFDM in 4 path PLC channel.

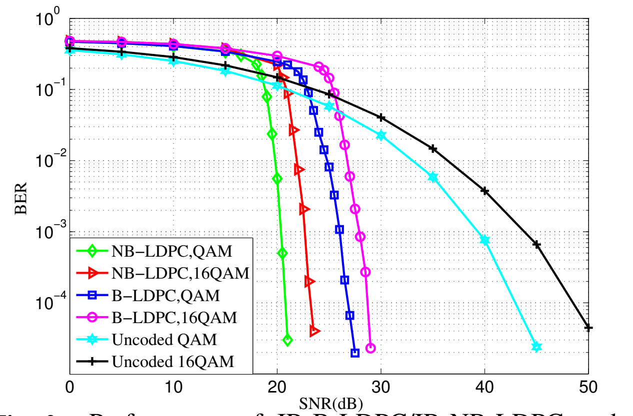

Fig. 2: Performance of IR-B-LDPC/IR-NB-LDPC coded OFDM in 15 path PLC channel.

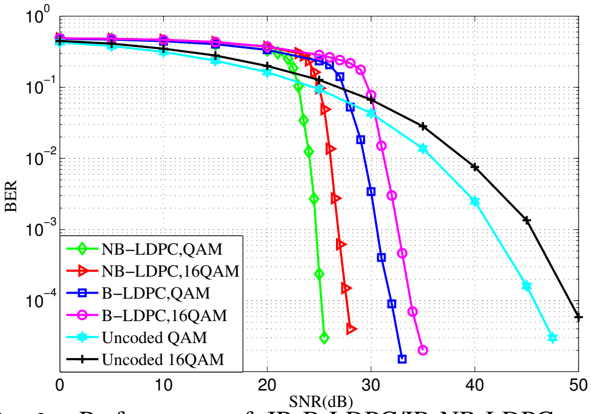

Fig. 3: Performance of IR-B-LDPC/IR-NB-LDPC coded OFDM in 4 path PLC channel with $\alpha = 0.3$ and $\Gamma = 0.1$.

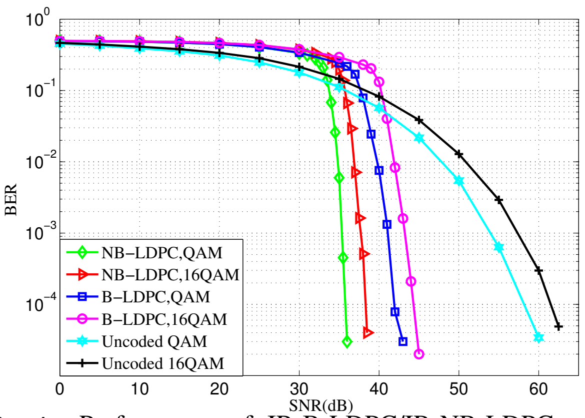

Fig. 4: Performance of IR-B-LDPC/IR-NB-LDPC coded OFDM in 15 path PLC channel with $\alpha = 0.3$ and $\Gamma = 0.1$.

rior BER performance for the SL-FFT decoder over impulsive noise and give approximately 6 dB coding gain compared to the SPA decoder and more than 20 dB compared to uncoded multipath PLC channels at $P_e = 10^{-4}$ for all cases.

6. CONCLUSION

The BER performance of the IR-B-LDPC/IR-NB-LDPC coded OFDM system has been significantly improved using LLR computation as initials of SPA/SL-FFT decoders with the optimum construction of $\mathbf{H}$. The performance is simulated for the same block length in bits and rate in two typical PLC channels. It is evident from the results that the proposed IR-NB-LDPC codes over $\mathbb{F}_{16}$ outperform the IR-B-LDPC codes over $\mathbb{F}_2$ in highly impulsive PLC channels. Therefore, the derivation of LLRs with SL-FFT decoder have superior performance in the presence of burst errors and makes the IR-NB-LDPC codes are very attractive for practical purposes to mitigate the impulsive noise in PLC channels.

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