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LDPC Coded Pilot-Tone Assisted MPSK CO-OFDM System

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Abstract
Pilot-tone assisted log-likelihood ratio (PT-LLR) is derived for LDPC-coded, coherent optical OFDM system in the presence of linear phase noise (LPN). The knowledge of common phase error (CPE) obtained from the pilot-tone is incorporated into the new LLR metric, which eliminates the need for prior CPE estimation and compensation. Compare our metric with the conventional LLR (C-LLR) through extensive simulation using their approximate versions (APT-LLR, AC-LLR). APT-LLR has the same order of complexity as AC-LLR while it outperforms AC-LLR for higher-order modulation formats (16-QAM, 64-QAM) at smaller pilot-tone-to-signal power ratios (PSR). With the help of time-domain blind intercarrier interference (ICI) mitigation, both metrics perform better in the presence of larger laser linewidth.

Keywords—Pilot-tone assisted, log-likelihood ratio, linear phase noise, common phase error, LDPC coded OFDM.

I. INTRODUCTION

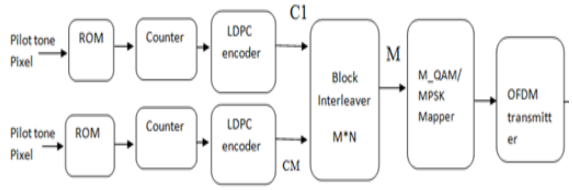
Low-density parity-check (LDPC) coded orthogonal frequency division multiplexing (OFDM) is a suitable coded modulation technique for long-haul optical communication Compared to single carrier, OFDM is more robust to chromatic dispersion (CD) and polarization mode dispersion (PMD). Recently, there have been quite a few experimental demonstrations using LDPC-coded, coherent optical OFDM (CO-OFDM) for high speed long-haul transmission. However, CO-OFDM is prone to linear phase noise caused by both transmitter laser and receiver local oscillator. LPN will cause both common phase error (CPE) and intercarrier interference (ICI) in the frequency domain. ICI can be approximated as additive white Gaussian noise (AWGN) and is negligible when the laser linewidth is small. Pilot subcarriers or pilot-tone have been proposed for CPE estimation.

The performance of the LDPC decoder depends on the calculation of the decoding metric, the log-likelihood ratio (LLR). In conventional LDPC coded CO-OFDM system, CPE compensation is usually performed first, followed by LDPC decoding using the conventional LLR .LLR for OFDM is calculated in the frequency domain, where the signal is corrupted by both CPE and ICI.

Pilot-aided LLR was first derived for LDPC coded pilot-symbol assisted single carrier system with BPSK format. Derive a new LLR metric based on pilot-tone for LDPC coded CO-OFDM in the presence of linear phase noise. A low-power RF-pilot-tone is inserted into the OFDM spectrum. Propose to incorporate the knowledge of CPE obtained from the pilot-tone into the calculation of the bit LLR. The derived bit LLR, defined as pilot-tone assisted LLR (PT-LLR), is evaluated from the likelihood function given the received signal that carries that bit and the pilot-tone that carries the knowledge of unknown CPE. The phase noise term is included into the decoding metric and thus the need of prior phase compensation is eliminated. We compare APT-LLR and AC-LLR in simulation, both of which have the same performance as their non-approximate versions. Moreover, APT-LLR outperforms AC-LLR for higher-order modulation formats (16-QAM, 64-QAM) at smaller pilot-tone to signal power ratios (PSR) with the same order of complexity.

When the laser linewidth becomes too large, tackling only CPE is no longer sufficient and ICI needs to be dealt with. we have proposed a time-domain blind ICI (BL-ICI) mitigation algorithm for non-constant amplitude format. Propose to employ the BL-ICI mitigation algorithm prior to the LLR calculation when ICI is no longer negligible. The performance of both metrics under prior ICI compensation is investigated through simulation. With the help of BL-ICI mitigation, the performance of both metrics has improved in the presence of larger laser linewidth. In this case, APT-LLR still outperforms C-LLR at smaller PSR for 16-QAM and 64-QAM.

II. SYSTEM MODEL



The system model of LDPC-coded, pilot-tone assisted M-QAM/MPSK CO-OFDM is shown in Fig. On the transmitter side, the data streams are encoded into code words using identical LDPC codes, with a code rate equal to , where and are the vector lengths of and , respectively. The outputs of these LDPC encoders are parsed into groups of bits by a block interleaver, with and being the number of bits per constellation point and number of modulated subcarriers per symbol, respectively. The bits are then mapped onto a complex-valued M-QAM or MPSK signal, where *M* = 2*M'*. The modulated subcarriers are zero-padded and transformed to the time-domain with length-*N*FFT IFFT by OFDM transmitter. A pilot-tone is inserted into the OFDM spectrum at zero frequency, which could be achieved by adjusting the IQ-modulator bias. For calculating the new LLR, we only consider CPE and treat the ICI term as AWGN. Assuming channel distortion (CD, PMD) is already removed, the received signal model becomes

$$r(k) = s(k)e^{j\phi} + w(k)$$

Where r(k) s(k), and ǿ are the received signal, transmitted signal, AWGN with E[|w(k)|2]=N0 , and CPE, respectively. k is the frequency subscript, while the time subscript is assumed to be fixed and thus omitted. Assuming the transmitted symbols take on equally likely values from either an M-QAM or an MPSK constellation set $\{A_l\}_{l=0}^{M-1}$ of with average symbol energy equal to $E_s = (1/M)\sum_{l=0}^{M-1}|A_l|^2$. We denote the transmitted and received pilot-tone by $s_p = \sqrt{E_p}$ and r_p, respectively. $\text{PSR[dB]} = 10\log_{10}(E_p/E_{sum})$, where $E_{sum} = NE_s$ is the total energy of data subcarriers in one OFDM symbol, excluding the pilot-tone. At the receiver, the LLR metric is calculated based on the received signal as well as the pilot-tone and passed to the iterative LDPC decoder, where the estimated message is $\overline{\mathrm{m}}$ obtained.

III. DERIVATION OF LLR METRIC

Derive our PT-LLR with consideration of the common phase error, which is contained in the received pilot-tone within each OFDM symbol. Thus, the bit LLR of c_k is calculated from the information contained both in the received signal $r(k)$ and the pilot-tone $r_p(k)$:

$$\lambda(k) = \ln\frac{\sum_{l\in\mathrm{P}_+} p(s(k) = A_l|r(k), r_p(k))}{\sum_{l\in\mathrm{P}_-} p(s(k) = A_l|r(k), r_p(k))}$$

where P_+ and P_- are the sets of all possible *l*-values corresponding to $c_k = 0$ and $c_k = 1$, respectively. Similar to [9, 10], the bit LLR can be re-written as:

$$\lambda(k) = \ln\frac{\sum_{l\in\mathrm{P}_+} p(r(k)|s(k) = A_l, r_p(k))}{\sum_{l\in\mathrm{P}_-} p(r(k)|s(k) = A_l, r_p(k))}$$

Each likelihood function is evaluated by averaging over all possible values of CPE:

$$p(r(k)|s(k) = A_l, r_p(k)) = \int_{-\pi}^{\pi}\left\{\begin{matrix} p(r(k)|s(k) = A_l, r_p(k), \phi) \\ \times\, p\left(\phi|s(k) = A_l, r_p(k)\right) \end{matrix}\right\} d\phi$$

Conditioned on $s(k) = A_l$ and ϕ, the only randomness in $r(k)$ is due to the Gaussian noise $w(k)$. Hence, the first probability term under the integration becomes:

$$p(r(k)\|s(k) = A_l, \phi) = C_l(k)\exp\left[\frac{2}{N_0}|r(k)A_l^*|\cos(\angle r(k) - \angle A_l - \phi)\right]$$

where $C_l(k) = \frac{1}{\pi N_0}\exp\left[-\frac{1}{N_0}(|r(k)|^2 + |A_l|^2)\right]$

the second probability term under the integration is:

$$p\left(\phi|r_p(k)\right) = \frac{\exp\left[\frac{2}{N_0}|r_p(k)s_p^*(k)|\cos(\angle r_p(k) - \phi)\right]}{2\pi I_0\left[\frac{2}{N_0}|r_p(k)s_p^*(k)|\right]}$$

where $I_0[|x|] = \frac{1}{2\pi}\int_{-\pi}^{\pi}\exp[|x|\cos\phi]\,d\phi$ is the zeroth-order modified Bessel function:

$$\lambda(k) = \ln\frac{\sum_{l\in\mathrm{P}_+} e^{-\frac{|A_l|^2}{N_0}} I_0\left[\frac{2}{N_0}|r_p(k)s_p^*(k) + r(k)A_l^*|\right]}{\sum_{l\in\mathrm{P}_-} e^{-\frac{|A_l|^2}{N_0}} I_0\left[\frac{2}{N_0}|r_p(k)s_p^*(k) + r(k)A_l^*|\right]}$$

For high SNR such that $E_s/N_0 \gg 1$, $I_0[|x|]$ may be approximated as: $e^{|x|}/\sqrt{2\pi|x|}$. After the approximation and removal of the ln (.) term, we obtain an approximate PT-LLR (APT-LLR):

$$\lambda(k) \approx \frac{1}{N_0}\left[\begin{array}{l}\max\limits_{l\in P_+}\left(2\left|r_p(k)s_p^*(k)+r(k)A_l^*\right|-\left|A_l\right|^2\right)\\ -\max\limits_{l\in P_-}\left(2\left|r_p(k)s_p^*(k)+r(k)A_l^*\right|-\left|A_l\right|^2\right)\end{array}\right]$$

For MPSK, the constant term $|A_l|^2$ under the max operator could be removed, which leads to:

$$\lambda(k) \approx \frac{2}{N_0}\left[\begin{array}{l}\max\limits_{l\in P_+}\left|\sqrt{E_p}r_p(k)+\sqrt{E_s}r(k)e^{-j2\pi l/M}\right|\\ -\max\limits_{l\in P_-}\left|\sqrt{E_p}r_p(k)+\sqrt{E_s}r(k)e^{-j2\pi l/M}\right|\end{array}\right]$$

where $\theta(k) = 2\pi l/M\ (l = 0,1,\dots M-1)$ is the modulated phase of k-th MPSK signal. In the literature [1, 2, 3], CPE compensation is usually employed first $(r'(k) = r(k)e^{-j\phi})$, followed by LDPC decoding using the conventional LLR (C-LLR): $\lambda_c(k) = \ln \sum_{l\in P_+} A(k,l)/\sum_{l\in P_-} A(k,l)$ with $A(k,l) = \exp\left[-\frac{1}{N_0}\left|r'(k)-A_l\right|^2\right]$. Similarly, we can derive approximate C-LLR (AC-LLR):

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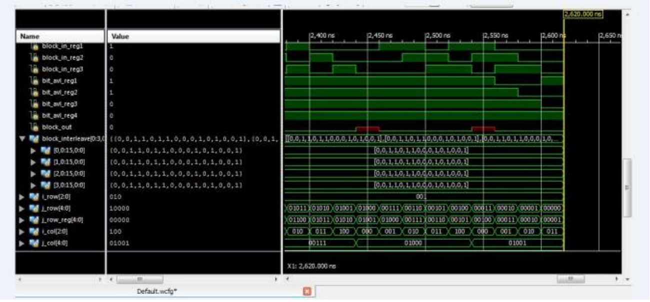
APT-LLR and AC-LLR have the same order of computational complexity. Define the total and effective SNR$_b$ as $\{E_b/N_0\}_{tot}^{dB} = 10\log\{(E_p+E_{sum})/(NN_0)\}$ and $\{E_b/N_0\}_{eff}^{dB} = 10\log\{(E_{sum})/(NN_0)\}$, respectively.

$$\{E_b/N_0\}_{tot} = \{E_b/N_0\}_{eff} + 10\log\left(1+10^{PSR/10}\right).$$

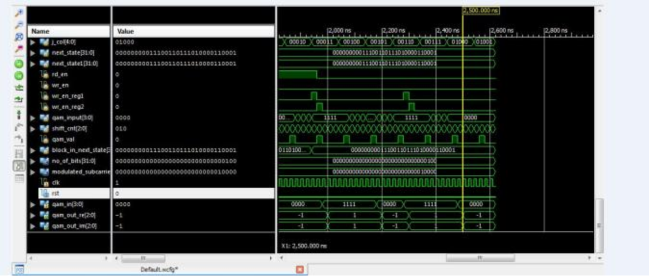
IV. PERFORMANCE RESULT

In encoder block parity check matrix is generated and multiplied with message bits to generate a code word. Code word is ex-ored with the variable nodes to generate column matrix. Column matrix is the input to the block interleaver. Block interleaver converts column matrix into row matrix.Row matrix is input to the M_QAM/MPSK mapper.16-QAM technique is performed in M_QAM mapper.In OFDM transmitter IFFT is performed to convert frequency domain into time domain.

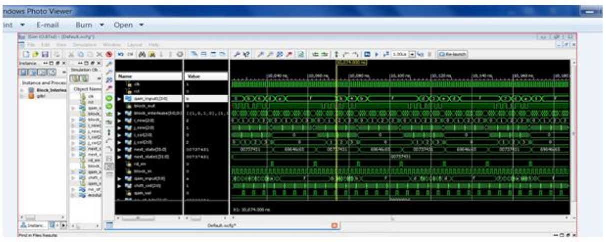
LDPC ENCODER



BLOCK INTERLEAVER



QAM MAPPER



OFDM TRANSMITTER

V. CONCLUSION

The LDPC Coded CO-OFDM system is based on the received signal with the consideration of linear phase noise.Compare the performance of our new metric with the conventional one through simulation. Investigate the performance of both metric under time domain blind ICI mitigation for larger line width.

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