Adaptive Log-Likelihood-Ratio for Optical Channels with Non-Additive-White-Gaussian-Noise

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Abstract: An adaptive LLR calculation algorithm is proposed for non-additive-white-Gaussian-noise (non-AWGN) optical channels. The noise distribution is continuously updated based on previous decisions. The performance is tested experimentally using 16-QAM and soft-DQPSK systems.

1. Introduction

While equalization and forward-error-correction (FEC) are mostly designed under the additive-white-Gaussian noise (AWGN) assumption, this assumption is often not true for optical channels. For example, phase noises, fiber/device nonlinearities, and residual inter-symbol-interaction (ISI) can cause the noise to be non-additive, signal-dependent, time-correlated (”colored”), I-Q correlated (”non-circular”), and not Gaussian. The distribution of the noise may also be time-varying, due to changes in the channel conditions [1,2], and needs to be continuously or periodically updated. Knowing the noise distribution is important for forward-error-correction (FEC) decoders, as it impacts how the demodulator computes the log-likelihood-ratio (LLR) (for soft-FECs), or the bit decisions (for hard-FECs). The mutual information (MI) between the LLR and the transmitted bits, and the pre-FEC BER can be used to predict the FEC performance and the post-FEC bit-error-rate (BER) when soft-FEC and hard-FEC are employed, respectively. This paper proposes a method to adaptively estimate the noise distribution based on previous decisions. Experimental results show that the proposed method is able to reduce the required optical signal-to-noise ratio (ROSNR) by 0.6 dB for 16-QAM with strong nonlinearity caused by semiconductor optical amplifier (SOA) (at MI between 0.82 to 0.9 bit/bit), and by up to 3.5 dB for strongly-filtered soft-differential-quadrature-shift-keying (DQPSK) [3] (at a pre-FEC BER of 4.8×10^{-3}).

1.1. Conventional LLR Calculation

Let $X \in \mathscr{X} = \{x_1, x_2, ..., x_M\}$ be the transmitted symbol that comes from an M-QAM constellation. Let $Y = X + Z$ be the received symbol, where Z is the band-limited additive noise. Let $\mathscr{X}_{j,0}$ and $\mathscr{X}_{j,1}$ be the subsets of constellation points whose j-th bit in the label is 0 and 1, respectively. The LLR of the j-th bit in a symbol under maximum *a posteriori* detection is defined as

$$LLR(b_j) = \log \frac{\sum_{x\in\mathscr{X}_{j,0}} P_{X|Y}(x|y)}{\sum_{x\in\mathscr{X}_{j,1}} P_{X|Y}(x|y)} = \log \frac{\sum_{x\in\mathscr{X}_{j,0}} P_{Y|X}(y|x)P_X(x)}{\sum_{x\in\mathscr{X}_{j,1}} P_{Y|X}(y|x)P_X(x)}. \tag{1}$$

When the noise is modeled by AWGN with variance σ^2, we have

$$P_{Y|X}(y|x) = \frac{1}{\sqrt{2\pi\sigma^2}} \exp - \frac{||y-x||^2}{2\sigma^2}. \tag{2}$$

1.2. LLR Calculation for Optical Channels

There have been research works to improve the FEC performance via improving the quality of LLR for optical channels, when the noise is not exactly AWGN [4–6]. Bosco et al. showed that for optical intensity-modulation direct-detection (IMDD) with on-off keying (OOK) modulation, the noise distribution is signal dependent, and Gaussian approximation would yield a significant performance penalty [4]. In the works by Zibar et al. [5], and Zhou et al. [6], they showed that for coherent and pulse amplitude modulated (PAM) optical systems, using non-identical Gaussian distributions to model the conditional distribution $P_{Z|X}(z|x)$ can yield a performance gain. However [4–6] all require a training stage to estimate the noise distribution, and once the estimation stage is completed, the noise distribution will be fixed. Borujeny and Kschischang [7] showed that the noise distribution can be time-varying, thus an adaptive technique is desired for noise distribution estimation.

2. Adaptive LLR Calculation

We propose an adaptive LLR calculation algorithm for optical communication channels. In our proposed scheme, each constellation point has its own noise distribution, $P_{Z|X}(\mathbf{z}|x_i)$, where the bold letter denotes the vector representation of a complex number, e.g., $\mathbf{z} = \begin{bmatrix} \Re(z) \\ \Im(z) \end{bmatrix}$. The noise distribution is approximated by an elliptical Gaussian distribution, which is characterized by a mean μ_i, and a covariance matrix Σ_i. Let $\Theta_i = \{\mu_i, \Sigma_i\}$ be the set of parameters that define $P_{Z|X}(\mathbf{z}|x_i)$, and $\Theta = \{\Theta_i : i \in \{1,2,...,M\}\}$. Let the superscript denote the time index, e.g., $y^{(l)}$ and $z^{(l)}$ are the received signal and noise at time l, and $\Theta^{(l)}$ is the l-th estimate of Θ. $\Theta^{(0)}$ is initialized to $\mu_i^{(0)} = 0$ and $\Sigma_i^{(0)} = I_{2\times2}\sigma^2$, where σ^2 is estimated from the average noise power. After receiving $y^{(l)}$, we first let $\Theta^{(l)} = \Theta^{(l-1)}$. The conditional probability $P_{Y|X}$, given the parameter $\Theta^{(l)}$ is

$$P_{Y|X;\Theta^{(l)}}(y^{(l)}|x_i;\Theta^{(l)}) = \det(2\pi\Sigma_i^{(l)})^{-1/2} \exp\{-\frac{1}{2}(\mathbf{y^{(l)}} - \mathbf{x_i} - \mu_\mathbf{i}^{(\mathbf{l})})^T (\Sigma_i^{(l)})^{-1} (\mathbf{y^{(l)}} - \mathbf{x_i} - \mu_\mathbf{i}^{(\mathbf{l})})\}. \tag{3}$$

Let $p_i^{(l)}$ be the probability of $X^{(l)} = x_i$, given $y^{(l)}$ and $\Theta_i^{(l)}$

$$p_i^{(l)} = P_{X|Y;\Theta^{(l)}}(x_i|y^{(l)};\Theta^{(l)}) = \frac{P_{Y|X;\Theta^{(l)}}(y^{(l)}|x_i;\Theta^{(l)})P_X(x_i)}{\sum_j P_{Y|X;\Theta^{(l)}}(y^{(l)}|x_j;\Theta^{(l)})P_X(x_j)}. \tag{4}$$

$\Theta^{(l)}$ can be updated as follows,

$$\begin{aligned} \mu_i^{(l)\prime} &= (1-\lambda p_i^{(l)})\mu_i^{(l)} + \lambda p_i^{(l)}(y^{(l)} - x_i) \\ \Sigma_i^{(l)\prime} &= 1-\lambda p_i^{(l)})\Sigma_i^{(l)} + \lambda p_i^{(l)}(\mathbf{z_i^{(l)}} - \mu_i^{(l)})(\mathbf{z_i^{(l)}} - \mu_i^{(l)})^T \end{aligned} \tag{5}$$

then assign $\mu_i^{(l)} = \mu_i^{(l)\prime}$ and $\Sigma_i^{(l)} = \Sigma_i^{(l)\prime}$, where $z_i^{(l)} = y^{(l)} - x_i$, and λ is a scaling parameter subject to optimization. The LLR for the j-th bit in the l-th symbol can be computed by substituting (4) in (1).

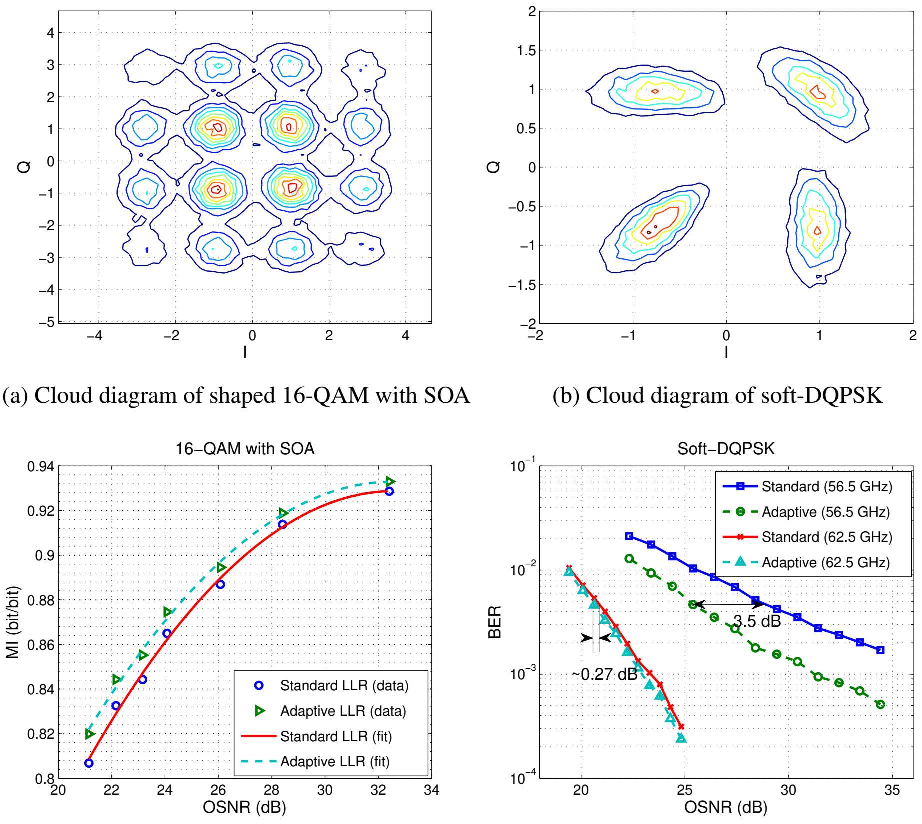
It is also worth noting that the proposed adaptive LLR calculation may also be used for systems with hard-FECs, in which case only the signs of the LLR are needed, and they may be computed by finding $\arg\max_{i\in\{1,2,...,M\}} p_i^{(l)}$.

3. Experimental Results

We experimentally evaluate the performance of the adaptive LLR calculation for both systems with soft-FEC and systems with hard-FEC. Back-to-back 16-QAM transmissions with SOA is used to test its performance for soft-FEC, when there is strong SOA-induced nonlinearity; Back-to-back soft-DQPSK transmission [3] is used to test its performance for hard-FEC, when there is strong distortion caused by wavelength-selective-switch (WSS) filtering. For both cases, a white Gaussian noise source is used to adjust the OSNR, and the signal is processed using offline digital-signal-processing (DSP) before passed to the demodulator.

The dual-polarization (DP) 16-QAM signal has a symbol rate of around 90 Gbaud, and is probabilistically constellation-shaped (PCS) to have an input entropy of 3.79 bit/polarization. The signal is amplified by an SOA before it is filter by a wavelength selective switch (WSS) with a 3-dB bandwidth of 150 GHz. The input power to the SOA is adjusted to be 2.2 dBm, which are in the nonlinear operating region of the SOA. The signal is detected by a coherent receiver and sampled by a real-time oscilloscope at a sampling rate of 200 GSa/s. Fig. 1a shows the cloud diagram of the post-DSP 16-QAM symbols. Fig. 1c shows the MI between the LLR and the transmitted bit, for both the standard LLR calculation and the adaptive LLR calculation. Assuming the soft-FEC has a MI threshold between 0.82 and 0.9, the adaptive LLR is able to reduce the ROSNR by 0.6 dB.

The back-to-back DP-QPSK signal has a symbol rate around 68 GBaud, and is filtered by a WSS, that is configured to have a 3-dB bandwidth of 56.5 GHz and 62.5 GHz, to mimic severe and moderate filtering effects caused by concatenated WSS's and transceiver band-limitations. The post-DSP symbols are used to perform soft differential decoding, and adaptive LLR and BER calculations are conducted on the soft-differentially decoded symbols. Note that a conventional coherent transceiver is used for illustration purposes, but in practice, some DSP blocks (such as carrier phase recovery) can potentially be simplified/removed with the use of soft differential decoding. Fig. 1b shows the cloud diagram of the post-DSP soft-DQPSK symbols, where the strong distortion might be attributed to the filtering and residual inter-symbol-interference (ISI). Fig. 1d shows the pre-FEC BER calculated by the standard slicer and the proposed adaptive slicer. With a decoding threshold of 4.8×10^{-3} [8], the adaptive slicer is able to reduce the ROSNR by 3.5 dB when the WSS bandwidth is 56.5 GHz, and by 0.27 dB when the WSS bandwidth is 62.5 GHz.



(a) Cloud diagram of shaped 16-QAM with SOA

(b) Cloud diagram of soft-DQPSK

(c) Shaped 16-QAM with SOA

(d) Soft-DQPSK

Fig. 1: a) and c) 16-QAM with PCS and input entropy = 3.79 bit/symbol. 90 GBaud back-to-back transmission with SOA was conducted. a) is measured when OSNR is 26.08 dB. b) Mutual information between the LLR and the transmitted bits is computed, and the soft-FEC is assumed to have a decoding threshold of between 0.82 and 0.9. b) and d) Soft-DQPSk 68 GBaud back-to-back transmission. b) WSS bandwidth is 56.5 GHz, and OSNR is 28.4 dB. d) WSS's with bandwidths of 56.5 GHz and 62.5 GHz are used. Pre-FEC BER is computed, and the hard-FEC is assumed to have a decoding threshold around 4.8×10^{-3} [8].

4. Conclusion

An adaptive LLR calculation algorithm is proposed for non-AWGN optical channels. This algorithm can be applied to both soft-FEC proetected and hard-FEC protected systems. Experiments show a ROSNR reduction of 0.6 dB and up to 3.5 dB for 16-QAM with SOA and soft-DQPSK systems, respectively.

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