128-Gbaud PAM4 O-Band Transmission Using Advanced MLSE with Simple LLR Calculation for SD-FEC Scheme

Shuto Yamamoto, Hiroki Taniguchi, Akira Masuda, Masanori Nakamura, Yoshiaki Kisaka
NTT Network Innovation Laboratories, NTT Corporation, 1-1 Hikarinooka, Yokosuka, Japan
shuto.yamamoto.at@hco.ntt.co.jp

Abstract: We propose simple methods to calculate LLR for IM-DD system with MLSE and SD-FEC and show that an advanced MLSE with the LLR calculation methods makes NGMI higher in 128-Gbaud PAM4 in 20-GHz bandwidth limitation. © 2023 The Author(s)

1. Introduction

Recent increase of data-center traffic induced by the massive use of cloud services that handle rich content. A huge number of connection ports are supported by Ethernet deployed economically and applied to intra- and inter-data center networks. Ethernet has been already completed standardization up to 400GbE at 100-Gbps per channel with O-band and intensity-modulation and direct-detection (IM-DD) schemes in IEEE 802.3 [1]. The next-generation Ethernet links such as 800GbE or 1.6TbE will require the increase of data rates even more. The higher modulation rate leads to bandwidth limitation (BWL) by transmission systems using narrowband and lower cost devices. BWL distorts the waveform of received signals as the inter symbol interference (ISI). To achieve the higher capacity transmission economically, due to the nonlinear response caused by the drivers, modulators, and photo detectors, conventional linear equalization schemes are no longer able to cope with the severe nonlinear ISI. Against this, several studies about solving severe ISI with device nonlinearity in IM-DD systems are reported [2-6].

We have been focusing on the strong tolerance with maximum likelihood sequence estimation (MLSE) which is applied in a receiver-side digital signal processing (DSP) technique. We have proposed an advanced MLSE based on nonlinear-channel estimation (NL-MLSE) for the higher estimation accuracy of channel response [7]. In IEEE 802.3, an application of soft-decision (SD) forward error correction (FEC) to high-baudrate IM-DD system is discussed [8, 9], and a performance evaluation for MLSE in SD-FEC scheme is required. In general, SD-FEC is performed based on logarithm likelihood ratio (LLR) which is calculated from the received-signal level after digital equalizations. On the other hand, LLR in MLSE schemes is calculated by soft-output Viterbi algorithm (SOVA) and the calculation is very complicated [10].

In this paper, we propose two simple methods of LLR calculation for IM-DD system with MLSE and SD-FEC in which the simple methods require the number of path-metric data in Viterbi algorithm (VA) to calculate LLR less than that in SOVA. We demonstrate 128-Gbaud PAM4 10-km transmission with 3-dB bandwidth of 20 GHz and NL-MLSE scheme with the simple LLR calculation methods achieves the higher performance in normalized generalized mutual information (NGMI) corresponding to the performance index in SD-FEC system.

2. Simple methods to calculate LLR in MLSE schemes

We propose two simple methods of LLR calculation. One is based on the minimum value of path metric for each bit composing PAM4 symbol in VA as shown below.

$$LLR = \left(\min_{i \in U_0} l_i - \min_{i \in U_1} l_i\right) / (2\sigma^2), \qquad (1)$$

where l_i is i-th path metric in VA and $U_{0,1}$ is a set of indexes corresponding to bit 0 or 1. σ is the standard deviation of the noise distribution in a transmission channel. This method is called Method A in this paper. Method A is very simple but a gain from trace back is not obtained in this method. The trace back is corresponding to a backward sequential estimation in VA and enhances the accuracy of bit decision. Therefore, Method A may not always achieve high performance.

The other is including an effect from trace back, in which the method is called Method B. In Method B, a temporal LLR is obtained from Method A. That is, the temporal LLR, LLR', is calculated by Eq. (1). Next, the bit decision based on the trace back updates the temporal LLR. If the bit decision based on the trace back coincides with the temporal bit decision based on Eq. (1), then the LLR is not updated. That is, $LLR = LLR'$, in which negative and positive LLRs correspond to bit 0 and 1, respectively. In this case, the LLR in Method B is the same value as that in

Table 1. Required numbers of path-metric data for LLR calculation. In the bottom row, $M = 4, d = 5,$ and $T = 20$.

Method A	Method B	SOVA
M^d	$2M^d$	$M^{d+1}T$
1024	2048	81920

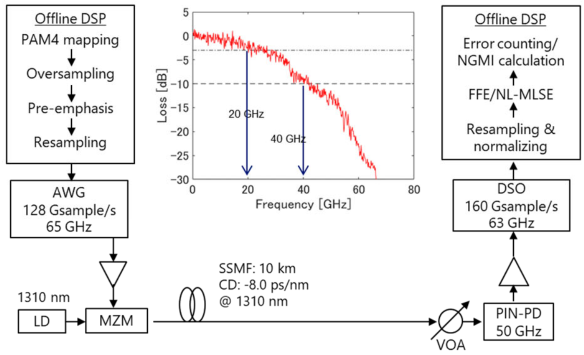

Fig. 1. Experimental configuration and frequency response.

Method A. If the bit decision based on the trace back does not coincide with the temporal bit decision, then the LLR is updated and the update rule is $LLR = -c\,\mathrm{sgn}(LLR')\,/(2\sigma^2)$, where the coefficient c is a constant number. The required numbers of path-metric data to calculate LLR in Methods A, B, and SOVA are shown in Table 1, where M is the number of PAM levels, T is the length of trace back in VA, and d is a memory length of channel response in MLSE. As shown in Table 1, Methods A and B require the number of path-metric data less than that in SOVA. The calculation complexity of Method B is almost same as that of the conventional VA in a hard-decision FEC (HD-FEC) scheme and the complexity is less than that of SOVA.

3. Experimental results

We demonstrate 128-Gbaud PAM4 10-km O-band transmission using NL-MLSE [7] with Method A or B. In this investigation, we evaluate bit error ratio (BER) and NGMI, which correspond to the performance indexes in HD-FEC and SD-FEC schemes, respectively. Figure 1 shows the experimental configuration. The transmission sequence is generated by an off-line DSP and a 65-GHz arbitrary waveform generator (AWG) driven at 1-sample/symbol. A 15-th order pseudo-random binary sequence (PRBS) is utilized as the transmission sequence. The electrical PAM4 signal is modulated to an optical signal at 1310 nm by a Mach-Zehnder modulator (MZM). The optical signal is transmitted to 10-km standard single-mode fiber (SSMF) without any optical amplifiers. The transmitted optical signal is received with a 50-GHz PIN photodiode (PD) after adjusting optical power by a variable optical attenuator (VOA). The amount of chromatic dispersion (CD) is -8.0 ps/nm at 1310 nm. The received signal is then converted into a digital signal sequence by a 63-GHz, 160-GSample/s digital storage oscilloscope (DSO) and demodulated by the conventional feed-forward equalizer (FFE) or NL-MLSE. The finite impulse response (FIR) filter has 45 T/2-spaced taps. The number of T-spaced taps for the adaptive low-pass filter (ALPF) and desired impulse response filter (DIRF) is 5. These filters are updated by the recursive least square (RLS) algorithm. The order of Volterra series expansion to emulate the nonlinear channel response is 3. To ensure the correct adaptation of the filters, the filter taps and kernels are pre-trained by the first 1000 symbols. The length of trace back is 20 in NL-MLSE. Figure 1 also shows the frequency response of the transmission system in which 3-dB and 10-dB bandwidths are 20 GHz and 40 GHz, respectively.

Figure 2 shows the relationship between the coefficient c and NGMI in each received optical power (ROP) for Method B. NGMI is calculated based on bit LLR [11]. As shown in this figure, the larger c makes NGMI lower in the lower ROPs. This means that incorrect bit decisions induced by the trace back increase cross entropy and deteriorate NGMI especially in the lower ROP in which bit error occurs frequently. The lower c realizes the higher NGMI regardless of the value of ROP and the best of c is around 0.02. Figure 3 shows the relationship between ROP and NGMI for the conventional FFE scheme and NL-MLSE scheme with Method A or B, in which c is 0.018 for Method B. As shown in this figure, NL-MLSE with Method A or B achieves the higher NGMI in each ROP than that in the conventional FFE scheme. In the ROP of 3 dBm, Methods A and B realize 18% and 22% increases in NGMI, respectively. Figure 4 show the relationship between BER and NGMI, in which c is 0.018 for Method B. As shown in this figure, the conventional FFE scheme cannot achieve 0.8 of NGMI. On the other hand, Methods A and B achieve 0.931 and 0.962 of NGMI, respectively. In this figure, the dashed line corresponds to the theoretical curve in additive white Gaussian noise (AWGN) channel and SOVA realizes the almost same performance with the theoretical curve. It is assumed that SOVA approximately achieves 0.985 of NGMI in this case. Therefore, Method B with $c = 0.018$ achieves NGMI close to SOVA.

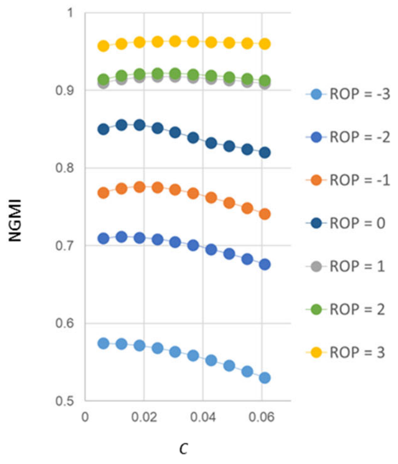

Fig. 2. Relationship between the coefficient *c* and NGMI for each ROP in Method B.

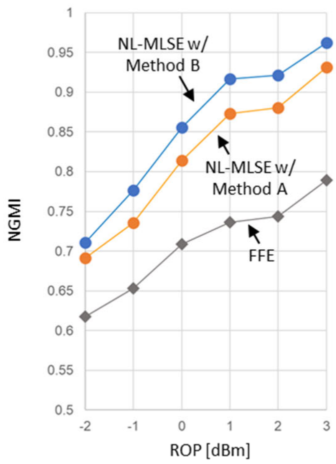

Fig. 3. Relationship between received optical power and NGMI in FFE and NL-MLSE with Method A or B.

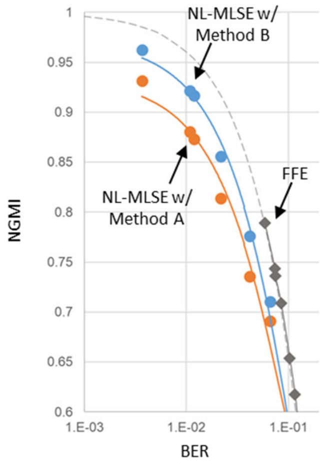

Fig. 4. Relationship between BER and NGMI in FFE and NL-MLSE with Method A or B.

4. Conclusions

We proposed two simple methods to calculate LLR in IM-DD system with MLSE and SD-FEC. The methods required the less complexity than that in SOVA as a conventional calculation scheme. We demonstrated 128-Gbaud PAM4 10-km transmission with 3-dB bandwidth of 20 GHz and showed that NL-MLSE with the simple calculation methods achieved the higher performance not only in BER but also in NGMI than that of a conventional linear equalization scheme. The proposed method including trace back technique realized NGMI close to SOVA. It is indicated that MLSE schemes are applicable to SD-FEC capable high-baudrate IM-DD system utilizing an uncomplicated DSP architecture.

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