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Nonlinear Tomlinson-Harashima precoding for direct-detected double sideband PAM-4 transmission without dispersion compensation

Haiyun Xin,1,2 Kuo Zhang,1 Deming Kong,2 Qunbi Zhuge,1 Yan Fu,1,2 Shi Jia,2 Weisheng Hu,1 and Hao Hu2,*

1Shanghai Institute for Advanced Communication and Data Science, State Key Laboratory of Advanced Optical Communication Systems and Networks, Shanghai Jiao Tong University, Shanghai 200240, China
2DTU Fotonik, Technical University of Denmark, Ørsteds Plads, Building 343, DK-2800 Kgs. Lyngby, Denmark
*huhao@fotonik.dtu.dk

Abstract: In this paper, we propose a nonlinear Tomlinson-Harashima pre-coding (THP) scheme for nonlinear distortion suppression in direct-detected double sideband (DSB) PAM-4 transmission systems. Based on the traditional THP, the feedback term is modified by introducing nonlinear components. In this way, more accurate feedback can be obtained to mitigate the signal distortions, especially the nonlinear distortions including the signal-to-signal beating interference and nonlinear power series caused by chromatic dispersion and square-law detection. Meanwhile, we also propose to only reserve the nonlinear kernels with adjacent tap products in nonlinear THP, for the purpose of computation complexity reduction. To verify the effectiveness, transmissions of double sideband (DSB) PAM-4 signal in 1550nm window are experimentally demonstrated. Volterra FFE is adopted on the receiver side to suppress linear and nonlinear pre-cursors. We optimize various parameters of hardware and apply appropriate simplification to the nonlinear THP kernels. The results indicate that, the proposed nonlinear THP can lead to up to three folds BER reduction, compared to the conventional linear THP. Finally, with the combination of proposed nonlinear THP and conventional Volterra FFE, we successfully transmit 84-Gbps PAM-4 and 107-Gbps PAM-4 respectively over 80 km and 40 km under the hard decision forward error correction (HD-FEC) threshold of 3.8 × 10⁻³.

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1. Introduction

With the emergence of diverse new scenarios, including mobile fronthaul, high-speed optical access, data center interconnects and so forth, advanced optical short-reach transmission systems with both upgraded capacity and extended reach are urgently needed, to bear the continuous traffic growth [1]. With the advantages of low-cost and simple structure, the combination of intensity modulation and direct detection (IM-DD) and pulse amplitude modulation (PAM), is an attractive solution for a short reach transmission system [2–4].

However, the double sideband (DSB) feature makes intensity modulated signal suffer from the power fading caused by chromatic dispersion, which substantially limits the achievable capacity and transmission distance. Recently, various solutions have been proposed to avoid the undesirable spectrum nulls in IM-DD system. One straightforward way is compensating the chromatic dispersion (CD) with a dispersion compensation module (DCM), which however needs additional optical devices and incurs high insertion loss [5,6]. Another promising method is the single sideband (SSB) based transmission with digital signal processing schemes, such as Kramers-Kronig (KK) algorithm or iterative subtraction, for
signal-to-signal beating interference (SSBI) cancellation [7–10]. Using such a method, the
generation of SSB signal lies on additional optical filtering or complex in-phase and
quadrature (IQ) modulation. In addition, electrical CD pre-compensation can be employed at
the transmitter side for DSB signal [11,12], which also relies on the IQ modulation with two
DACs.

In all the above mentioned solutions, the benefit of PAM signal from other formats is not
fully utilized. In fact, as a single-carrier baseband signal, it is possible to remove the power
fading caused inter-symbol interference (ISI) on the decision samples, without eliminating the
spectrum nulls in the frequency domain, because the decision levels are sampled at a rate (one
two sample per symbol, 1sps) lower than Nyquist sampling criterion (>2sps). In [13], the essence
of power fading caused ISI for PAM signal is theoretically analyzed. According to the
analysis, feedback equalization, including receiver-side decision-feedback equalization (DFE)
or transmitter-side Tomlinson-Harashima pre-coding (THP), can provide an effective solution
to address the power fading problem. These methods firstly estimate ISI from previous
symbols and then subtract it, consequently the ISI can be eliminated. In addition, transmitter-
side THP is more desirable than receiver-side DFE, since error propagation can be avoided. In
recent years, THP has been applied to various systems. For example, in [13–15], THP is used
to combat the power fading caused distortion. In [16,17], THP is used to deal with ISI
induced by bandwidth limitation from device filtering or fast-than-Nyquist (FTN) signaling.
In these demonstrations, linear THP pre-coding was adopted, which can only address the
linear power fading issue, since the estimated ISI is calculated by a weighted summation of
previously symbols. However, in this way, the mitigation of nonlinear distortion completely
depends on nonlinear FFE in the receiver side, which will be less effective when nonlinear
distortion gets more severe.

In this work, we propose to implement nonlinear THP pre-coding for improving system
performance. We present theoretical analysis on the linear and nonlinear impairments in IM-
DD systems when CD is interacted. Based on the traditional linear THP, nonlinear
components are introduced to feedback a more accurate ISI. In addition, to reduce the
computation complexity, we propose to only reserve the nonlinear kernels with adjacent tap
products. The proposed scheme is experimentally verified in C-band PAM-4 transmission
system. Volterra FFE is implemented in receiver side to mitigate pre-cursors, which could not
be solved by nonlinear THP. We optimize various parameters in the system, including optical
modulation index (OMI), bias voltage, roll-off factor and equalization configurations. Results
indicate that nonlinear THP can lead to up to three folds BER reduction, compared with the
conventional linear THP pre-coding. Finally, transmissions of 84-Gbps PAM-4 over 80 km
and 107-Gbps PAM-4 over 40 km in the C-band are achieved by a combination of nonlinear
THP and Volterra FFE, below the HD-FEC threshold of 3.8 × 10⁻³.

This paper is organized as follows. In Section 2, we theoretically analyze the linear and
nonlinear impairments of IM-DD system in the C-band transmission, and then present the
principle of nonlinear THP scheme. In Section 3, we describe the experimental setup. In
Section 4, we first present the optimization of system parameters including hardware
parameters setting and nonlinear THP configuration, then evaluate the performance of
nonlinear THP. Section 5 concludes this work.

2. Impairments in IM-DD system and nonlinear Tomlinson–Harashima pre-
coding

2.1 Impairments in IM-DD system

According to the theoretical analysis in [18], the signal after the optoelectronic detection by a
photodiode (PD) detection will experience kinds of impairment sources in an IM-DD system.
In this work, we mainly focus on electro-optical modulation, CD effect and square-law
detection.
Assuming the electrical driving current is $m(t)$, then after intensity modulation, the optical power is

$$P_{o,T} = P_o + \eta m(t),$$

where $P_o$ is the average optical power, and $\eta$ is the efficiency factor. Correspondingly, the optical field is expressed as

$$E_{o,T}(t) = \sqrt{P_{o,T}(t)} = \sqrt{P_o \frac{1 + \eta m(t)}{P_o}}.$$  

(2)

For simplicity, $E_{o,T}(t)$ is normalized as $E_{o,(t)} = \sqrt{1 + m'(t)}$, where average power of $m'(t) = \eta m(t)/\sqrt{P_o}$ exactly determines the carrier-to-signal power ratio (CSPR).

Using a Taylor series expansion, we can rewrite $E_{o,T}(t)$ as

$$E_{o,i}(t) = 1 + \sum_{n=1}^{\infty} c_n m'(t)^n = 1 + \frac{1}{2} m'(t) + \sum_{n=2}^{\infty} c_n m'(t)^n,$$

(3)

where $c_n = (-1)^{n-1} (2n-3)!/(2^{n-2} (n-2)!n!)$ for $n = 2, 3, 4, \ldots$.

If the complex CD transfer function in time-domain is $h_{CD}(t)$, the received optical field is

$$E_{o,i}(t) = E_{o,i}(t) \otimes h_{i,o}(t) = 1 + \frac{1}{2} m'(t) \otimes h_{i,o}(t) + \sum_{n=2}^{\infty} c_n m'(t)^n \otimes h_{i,o}(t).$$

(4)

The received optical power becomes

$$P_{o,i}(t) = |E_{o,i}(t)|^2 = 1 + m'(t) \otimes R[h_{i,o}(t)] + \sum_{n=2}^{\infty} c_n m'(t)^n \otimes R[h_{i,o}(t)]$$

(5)

where $R\{\cdot\}$ represents the real part. In Eq. (5), the first term is DC offset and is usually blocked. The second term contains the signal of interest and also reflects the power fading effect. In previous literatures, this term is often expressed in the frequency domain

$$H_{CD}(f) = f(h_{CD}(t)) = \cos(2\pi \beta_l f),$$

(6)

where power fading occurs when $2\pi \beta / \ell f - \frac{\pi}{2}$ is a multiple of $\pi$. Here, for intuitive observation of impairment, we give time domain analysis of the second term. After sampling and expansion, this term is expressed as

$$m'(t) \otimes R[h_{i,o}(t)] = m[kT] + \sum_{q=0}^{\infty} R[h_{i,o}(qT)] m(kT - qT)$$

(7)

ideal signal

$$= m[k] + \sum_{q=0}^{\infty} R[h_{i,o}(q)] m[k - q],$$

power fading induced ISI
where \( k \) indicates the symbol index, \( T \) is symbol duration and \( q \) is a nonzero integer, Eq. (7) can be divided into the ideal signal and the ISI component. Obviously, the power fading induced ISI only contains the weighted summation of adjacent symbols, and thus is a linear impairment.

Apart from the linear impairment in the second term of Eq. (5), the signal also suffers from nonlinear impairments in the third term and forth term. The third term is the received power series of the applied current, and the forth term is known as the signal-signal beating interferences (SSBI). While SSBI has been intensively studied previously, the power series part is usually neglected. It is easy to realize that, due to the introduction of \( h_{CD}(t) \), the high order interferences are with memory, and larger transmission distance will cause severer nonlinear distortion with a longer memory. To sum up, the total inter symbol interference (ISI) in an IM-DD system can be divided into two parts: (1) first order interference, which is also the power fading caused ISI; (2) Second and higher order interferences which are composed of received power series and SSBI. To better eliminate the distortions, both first order interference and high order ones should be taken into account.

2.2 Nonlinear THP pre-coding

In the conventional THP pre-coding, the ISI is estimated by a linear weighted summation of the previous symbols. Such estimation can well match the power fading caused ISI, since this ISI is exactly a linear combination of adjacent symbols as indicated by Eq. (7) or the second terms of Eq. (5). However, the linear combination is not capable to cover nonlinear impairments, i.e. the power series of the applied current and SSBI in the third and forth terms of Eq. (5). Some methods have been reported to solve this issue. In [13], the CSPR is appropriately improved to suppress the two mentioned nonlinear impairments. In this case, the power fading caused ISI is the dominant impairment. Results indicate that the combination of THP in transmitter side and FFE in receiver side can lead to an improved transmission performance. But a high CSPR also means a lower extinction ratio, and will reduce effective SNR. Afterwards, it has been proposed to replace the linear FFE in receiver side with a Volterra-based FFE, generating a THP-Volterra FFE [15], achieving the transmission of 56Gbps PAM-4 over 80 km under KP4 FEC. In this method, the cancellation of nonlinear impairments depends on the Volterra FFE. However, it has been reported that Volterra FFE is less effective than a joint Volterra FFE and Volterra DFE, because of the noise enhancement in FFE, especially in a severely distorted channel [19]. Inspired by this, and considering that THP can be regarded as transmitter-side DFE, we propose to replace traditional linear THP with a nonlinear one, and the working principle is illustrated in Fig. 1. It is noted that, the proposed THP scheme in this work is quite different from [20], where a memory-less nonlinear THP is used to compensate the nonlinear P-I mapping of EML modulator and demonstrated in 1310nm wavelength region. However, it is not able to address the issue in this work, since the nonlinear distortion due to CD and square-law detection has memory while the nonlinear P-I mapping is memory-less.
The input of the nonlinear THP pre-coding module is the original PAM-4 symbols with four levels, denoted as $m[k]$. Each input symbol subtracts the predicted ISI $e[k]$, according to the previous output symbols $x[k-1], x[k-2]...$. In order to guarantee the stability of this feedback system, modulo operation is inserted to confine the output values within $-M$ and $M$, i.e. $x[k] = \text{Mod}(m[k] - e[k], M)$. As a result, the pre-coded output signal $x[k]$, is uniformly distributed and the value is confined within $-M$ and $M$. In the case of PAM-4, $M$ is 4. The prediction of ISI includes both linear kernels and nonlinear kernels. If nonlinear kernels are not used, it is same as traditional linear THP pre-coding. After the fiber channel and PD detection, partial distortions are compensated by pre-coding. Then, the signal is equalized by FFE to mitigate the residual ISI. Here, the output of the FFE is an extended PAM signal with reference levels derived from the transmitter side. Finally modulo operation is used to recover the original PAM-4 signals.

### 2.3 Nonlinear THP kernels and their simplification

Next, we discuss the kernel design for nonlinear THP. Among various nonlinear distortion modelling methods in optical fiber transmissions, Volterra series is the most commonly used in IM-DD systems [21,22]. The Volterra series is essentially a combination of polynomials which can well match signal distortions in IM-DD system according to Eq. (7) in section 2.1. In this work, we will employ Volterra based series to evolve THP. Accordingly, the predicted ISI can be expressed as:

$$
e[k] = \sum_{l=1}^{n} h(l, l)[x[k - l]] + \sum_{l=1}^{n} \sum_{l_1=1}^{L_1} h(l, l_1)[x[k - l]][x[k - l_1]] + ...$$

$$+ \sum_{l=1}^{n} \sum_{l_1=1}^{L_1} \sum_{l_2=1}^{L_2} h(l, l_1, l_2)[x[k - l]][x[k - l_1]][x[k - l_2]] + ...$$

$h(l, l_1, ... l_r)$ is the coefficient of the $n^{th}$ order kernel of FFE. $L_r$ is the tap number of the $n^{th}$ order kernel.

Since high order kernels occupy a major computation complexity due to large amount of multiplications, it is desired to simplify these kernel in order to sustain the low-cost advantage of the IM-DD systems. Because some kernels in the Volterra series are almost equal to zero, the simplification of the nonlinear THP is to select the most important kernels while truncating others. Two aspects are considered in this work. Firstly, as the second order interference is the dominant nonlinear distortion, only second order kernels are reserved. Secondly, product terms with large intervals have negligible contributions to signal distortions, thus they are regarded as less significant kernels and can be pruned. This scheme is termed as the diagonally-pruned scheme and has been used in the Volterra FFE [23,24].

![Diagram of non-linear THP pre-coding and decoding](image.png)
this work, the pruned scheme is extended to the nonlinear THP. Mathematically, $e(k)$ of this simplified nonlinear THP is expressed as:

$$e(k) = \sum_{l=1}^{p} h(l) x(k-l) + \sum_{l=1}^{p} \sum_{i=1}^{p} h(l, l_i) x(k-l) x(k-l_i),$$

(9)

where $P$ is the product memory length, which means that kernels with product terms larger than $P$ sample intervals in THP are pruned. Obviously, this pruned scheme can considerably reduce the high order kernel number of the nonlinear THP. For example, when the memory length of the second order tap is 21 and $P$ is 4, the number of second order kernels can be reduced from 231 to 95.

3. Experimental setup

The experimental setup is as shown in Fig. 2. Two transmission cases are tested, including 84-Gbps over 80 km and 107-Gbps PAM-4 signal over 40 km. At the transmitter side, pseudorandom binary sequence (PRBS) of $2^{15}$ is selected and mapped onto PAM-4 symbols. When THP is used, the PAM-4 symbols are pre-coded according to the description in Section 2. Then, the original or pre-coded PAM-4 symbols are pulse-shaped by a raised cosine filter at two samples per symbol (2sps), and fractionally up-sampled to match the 65 GSa/s sampling rate of arbitrary waveform generator (AWG, Keysight M8195A). Pre-emphasis is used to compensate the non-ideal frequency responses of electrical components, including AWG, electrical amplifier (EA) and electrical attenuator (Att). After that, the generated data sequence is loaded into AWG with a 3-dB bandwidth of 25GHz. Note that, in order to satisfy the length requirement of data loaded into AWG (multiples of 128), the PAM-4 symbols are appropriately truncated before the following operations.

After the AWG, the electrical signal is amplified by an EA (SHF806A) and attenuated by an electrical attenuator, before driving a Mach-Zehnder modulator (MZM). The bias of MZM is not exactly at the quadrature point, but is optimized together with the peak-to-peak voltage for different transmission cases. Considering that the distributed feedback Bragg (DFB) laser source used in the experiment has an output power of 12 dBm and the MZM has a 10-dB insertion loss, the power launched into the standard single mode fiber (SSMF) is around 2 dBm. After the transmission, the signal is amplified by an erbium-doped optical fiber amplifier (EDFA), and the amplified spontaneous emission (ASE) noise is suppressed by a tunable optical filter (TOF). After adjusting power by the variable optical attenuator (VOA), the optical signal is detected by a PIN (U2t, XPDV2120R) with a 50-GHz bandwidth. In the receiver side, a real-time oscilloscope (Keysight, DSA-X 93304Q), with an 80-GSa/s 8-bit sampling and a steep built-in aliasing filter of 33-GHz bandwidth, is employed to digitize the electrical signal for off-line processing, which mainly includes synchronization, resampling and equalization.

The verification of nonlinear THP includes two procedures. First, to obtain the coefficients of THP, the original PAM-4 signal without THP is sent in the transmitter side, and both feed forward equalizer (FFE) and decision-feedback equalization (DFE) are used in the receiver side. The tap coefficients of FFE and DFE are updated simultaneously with common feedback errors. For the FFE part, we adopt a nonlinear Volterra filter operated at 2sps. For the DFE part, we use either linear or nonlinear variants, depending on whether linear or nonlinear THP will be used in the transmitter side.

Then, because the THP can be regarded as transmitter-side DFE, and the derived kernel coefficients of DFE are directly applied to the kernels of THP. Different from the first procedure, the receiver side equalization in this procedure only contains Volterra FFE (vFFE), which outputs an extended PAM signal. At last, a Modulo operation is applied to recover the original PAM-4 data sequence.
Fig. 2. Experimental setup and DSP flows. When THP coefficients need to be obtained or optimized, the original PAM-4 without pre-coding is transmitted, and joint FFE and DFE is adopted in the receiver side. When THP performance needs to be tested, the THP pre-coded PAM-4 signal is transmitted, and only FFE is adopted in the receiver side.

4. Experimental results and discussions

4.1 Parameters optimization using error-propagation free DFE

Prior to assessing the performance of nonlinear THP, we optimize various parameters to achieve an optimal transmission performance. The optimization includes two steps. First, we optimize hardware parameters including the bias voltage, optimal modulation index (OMI) and roll-off factor for different transmission cases. Then, we investigate the optimal kernel configuration in nonlinear THP. For convenience, the optimization is based on the BER results of error-propagation (EP) free DFE, instead of BER results of nonlinear THP. This is reasonable because the BER results of the EP-free DFE can be regarded as a benchmark of the corresponding THP.

Figure 3 shows the BER results of various combinations of OMI and bias voltage. To obtain the optimal performance of each configuration, we apply Volterra DFE with sufficient kernels, where 1st tap length is set as 81, and 2nd tap length is set as 30. As can be seen, the Volterra DFE always has a lower BER than linear DFE in all (OMI, bias) conditions, which hints the superiority of nonlinear THP over linear THP. In addition, as optical modulation index (OMI) increases, the gap increases, since nonlinear feedback kernels can better deal with the aggravated nonlinear distortion caused by smaller CSPR. According to the results, we select the (OMI, bias) pair for the four cases as: (1) 84-Gbps over 80 km with linear THP: (0.31, 1.02V), (2) 84-Gbps over 80 km with nonlinear THP: (0.31, 1.02V), (3) 107-Gbps over 40 km with linear THP: (0.41, 0.92V), (4) 107-Gbps over 40 km with nonlinear THP: (0.41, 0.92V).
Then, we optimize the roll-off factor for the two data rates. Considering that bandwidth limitation mainly originates from AWG and MZM, a larger roll-off factor directly increases the amount of linear ISI in the optical intensity. When the optical intensity is converted to optical field and interacted with dispersion effect, the increased ISI will further cause more severe nonlinear distortion. The BER results with different roll off factors are plotted in Fig. 4. It is found that the optimal roll-off factor for nonlinear DFE is larger than linear DFE.
Fig. 5. BER results as a function of the 2nd tap number $L_2$ and the product memory length $P$.
(a) and (b): 84 Gbps PAM-4 over 80 km. (c) and (d): 107 Gbps PAM-4 over 40 km.

Last, with the optimal hardware parameters (OMI, bias and roll off factor), we optimize the kernel configuration of THP to reduce the computation complexity. Similar to above mentioned measurements, we use the results of EP-free Volterra DFE as a benchmark. For 1st order tap ($L$ in Eq. (9)), we find that the BER reaches a floor when the 1st order tap number is more than 81. Thus, we use 81 taps for linear THP, and also fix this value for nonlinear THP. Then, we investigate the required 2nd order tap number ($L_2$ in Eq. (9)) and product memory length ($P$ in Eq. (9)) in Fig. 5. It can be seen that lower BER can be obtained when larger tap number and memory length are employed. However, the improvement decreases when they become too large. From Fig. 5, for both 84-Gbps PAM-4 over 80 km, the optimal $L_2$ and $P$ are 20 and 10. Correspondingly, the number of multiplication in nonlinear THP kernels is reduced from $20 \times 20 \times 19 / 2 = 210$ to $20 + 20 \times 19 / 2 - 10 \times 9 / 2 = 165$. And for 107-Gbps PAM-4 over 40 km, the $L_2$ and $P$ are selected as 20 and 6, then the multiplication number in nonlinear THP kernels is reduced from 210 to $20 + 20 \times 19 / 2 - 14 \times 13 / 2 = 109$.

4.2 Transmission results based on nonlinear THP

With the optimal hardware parameters and kernel configuration, we test the transmission results when employing nonlinear THP. The kernel coefficients of Volterra DFE derived in Section 4.1 are directly applied to THP. Figure 6 shows the received THP signal before and after Volterra FFE, including eye diagram, electrical spectrum and occurrence of decision samples. These figures were obtained under the condition of 84-Gbps PAM-4 over 80 km distance. In Fig. 6(b), spectra nulls appear in received THP signal, due to CD and square-law
detection. The corresponding occurrence of decision samples shown in Fig. 6(c) has a Gaussian type profile, where the middle levels occupy largest proportion. In contrast, from Figs. 6(d) and 6(e), it is interesting that although the spectra nulls still exist after Volterra FFE, a clear multi-level PAM signal is obtained. Here, we also attach the occurrence of linear THP signal as a reference to compare the performance of linear and nonlinear THP. It is observed that the nonlinear THP reduces the noise variance at each level and has a better performance than linear THP.

Fig. 6. Received signal with nonlinear THP. (a)–(c): Eye diagram, electrical spectrum and occurrences of decision samples before equalization. (d)–(f): Eye diagram, electrical spectrum and occurrences of decision samples after equalization. For (f), linear THP result is attached as a reference.

At last, we investigate the BER results of linear and nonlinear THP under different PIN-PD input powers in Fig. 7. One can easily find that, the nonlinear THP can lead to a considerable BER reduction in comparison with linear THP. For 84-Gbps PAM-4 over 80 km transmission, the BER can be reduced from $5.3 \times 10^{-3}$ to $1.8 \times 10^{-3}$ at 5.5dBm PIN-PD input power. And for 107-Gbps PAM-4 over 40 km transmission, the BER can be reduced from $5.8 \times 10^{-3}$ to $3 \times 10^{-3}$ at 6.3dBm PIN-PD input power. Therefore, by employing nonlinear THP, we can transmit 84-Gbps PAM-4 signal over 80 km and 107-Gbps PAM-4 signal over 40 km under HD-FEC threshold.
Fig. 7. BER results as a function of PIN-PD input optical power. (a): 84-Gbps PAM-4 over 80 km. (b): 107-Gbps PAM-4 over 40 km.

5. Conclusions

In this paper, we propose nonlinear THP to mitigate nonlinear distortions in the IM-DD system when CD is interacted. By introducing nonlinear kernels in the THP, the estimated ISI can better match the signal distortion, thus improving the BER performance. In the experiment, we optimize various parameters, such as bias voltage of MZM, OMI and roll-off factor. In addition, we discuss the simplification of nonlinear THP kernels to reduce computation complexity. It is demonstrated that the BER can be reduced by 2–3 folds by employing the proposed nonlinear THP, compared with the traditional linear THP. As a result, with the combination of proposed nonlinear THP and conventional Volterra FFE, we successfully transmit 84-Gbps PAM-4 signal over 80 km and 107-Gbps PAM-4 signal over 40 km under the HD-FEC threshold of $3.8 \times 10^{-3}$.

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