A 10 Gb/s PAM-4 Transmitter With Feed-Forward Implementation of Tomlinson-Harashima Precoding in 28 nm CMOS

BYUNGJUN KANG1, (Student Member, IEEE), GYU-SEOB JEONG2, (Member, IEEE), JEONGHO HWANG3, (Member, IEEE), KWANSEO PARK4, (Member, IEEE), HYUNGROK DO1, (Graduate Student Member, IEEE), HYOJUN KIM1, (Graduate Student Member, IEEE), HAN-GON KO5, (Member, IEEE), MOON-CHUL CHO6, (Member, IEEE), AND DEOG-KYOO JONG1, (Fellow, IEEE)

1Department of Electrical and Computer Engineering, Seoul National University, Seoul 08826, South Korea
2Ayar Labs, Santa Clara, CA 95054, USA
3Cadence Design Systems, San Jose, CA 95134, USA
4Department of Electrical Engineering and Computer Sciences, University of California at Berkeley, Berkeley, CA 94720, USA
5ONE Semiconductor, Suwon-si 16226, South Korea
6Samsung Electronics, Hwaseong 18448, South Korea

Corresponding author: Deog-Kyoon Jeong (dkjeong@snu.ac.kr)

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ABSTRACT A 10 Gb/s PAM-4 transmitter (TX) with a modulo-based equalization technique is presented. The proposed feed-forward Tomlinson-Harashima precoding (FF-THP) scheme takes advantage of both Tomlinson-Harashima precoding (THP) and feed-forward equalization (FFE). The vertical eye margin (VEM) is enhanced by removing the precursor inter-symbol interference (ISI) with pretaps while incorporating the modulo operation. The VEMs of equalization methods are derived based on z-domain response (ZDR). The effectiveness of the FF-THP is examined by quantitative analysis and numerical simulation. Especially for a one-pole channel with a precursor, optimized tap coefficients of an FFE and an FF-THP are derived as closed-form concerning the precursor and the first postcursor. Calculations of decision threshold voltage and estimated bathtub curve based on Gaussian noise are featured by using the histogram of an eye diagram. The advantages of the FF-THP over a conventional FFE are measured by a fabricated chip. The proposed TX compensates for a 21 dB channel loss with a level mismatch ratio of 99.1% and with a figure of merit (energy efficiency per sum of channel ISI) of 4.05 pJ/b/ISI. Moreover, the FF-THP achieves 38% and 87.5% improvement on VEM and horizontal eye margin, respectively, compared with an FFE. It is fabricated in 28 nm CMOS technology, occupying an active area of 0.075 mm².

INDEX TERMS Feed-forward equalizer (FFE), inter-symbol interference (ISI), PAM-4, z-domain response (ZDR), Tomlinson-Harashima precoding (THP), transmitter (TX).

I. INTRODUCTION

Due to the rapid growth of demand for the data throughput in wireline interfaces, multi-level signaling such as pulse amplitude modulation (PAM), especially PAM-4, is widely adopted in many standards such as Fibre Channel, InfiniBand, and Ethernet [1]–[3]. Despite the fact that multi-level PAM signaling can significantly increase the data throughput, inter-symbol interference (ISI) of a channel and the reduced signal-to-noise ratio (SNR) substantially degrade signal integrity and bit error rate (BER) performance, thus making it a great challenge to employ PAM-4 signaling on a high-loss channel. While the channel equalization can be done at both the transmitter (TX) and the receiver (RX), there are a few advantages
in equalizing the channel loss on the TX side. Equalization is more straightforward to implement at the TX than at the RX side because TX has the exact information of the input data, whereas the RX may have a sampling error. Therefore asymmetric links, such as DRAM interfaces, may use TX equalization thanks to its simplicity [4].

While a feed-forward equalizer (FFE) in the form of a finite impulse response (FIR) filter is widely employed, because of a scaling factor imposed on by maximum drivable voltage or current, the eye opening on a high-loss channel can be significantly reduced [5]. While the nonlinear decision feedback equalizer (DFE) is widely used for being immune to noise boosting, errors tend to occur in bursts that exacerbate the forward error correction (FEC) performance. Thus, combining the DFE and the FEC in PAM-4 signaling can bring out significant performance degradation [5].

As an alternative, Tomlinson-Harashima precoding (THP) is a viable candidate for TX equalization for a high-loss channel since by nonlinear operation and subtraction of a uniformly distributed transmitted signal [6], [7]. The THP can theoretically equalize a wide range of channels, regardless of the channel loss [8], [9]. Albeit attractive, when it comes to a physical implementation, the use of THP is limited because of the feedback timing constraint and the lack of precursor-handling capability that are the same problem as the DFE.

Various techniques such as pipelining and mapping have been reported to relieve the timing constraint in the THP implementation [10]–[12]. However, even though the timing constraint is alleviated, the precursor ISI has remained a problem equalizing a high-loss channel. Therefore, another approach that has been reported is a model predictive control (MPC) that offers limited controllability of a precursor ISI [13], [14]. This paper proposes a novel feed-forward Tomlinson-Harashima precoding (FF-THP) architecture incorporating the precursor compensation in the modulo-based equalization to build a 10 Gb/s PAM-4 TX. To mitigate the timing issue, the proposed FF-THP utilizes modulo prediction rather than direct calculation. In addition, by implementing 2 pretaps and 10 posttaps, the FF-THP can handle large channel ISI, including a precursor.

The rest of this paper is organized as follows. In Section II, basic concepts and functions of the THP, the FFE, and the proposed FF-THP are introduced. Also, the z-domain analysis is derived in this Section. Especially, the quantitative analysis and numerical simulation are conducted on a simple one pole channel with a precursor to examine the effect of the precursor. Section III describes the overall architecture and the constructed engines of the proposed FF-THP TX. Subsequently, the measurement results of the fabricated chip are presented in Section IV, followed by a summary and a conclusion of this paper in Section V.

II. PROPOSED ARCHITECTURE AND ANALYSIS

A. COMPARISON BETWEEN TOMLINSON-HARASHIMA PRECODING AND FEED-FORWARD EQUALIZER

The architectures of the THP and the FFE are illustrated in Fig. 1(a) and (b). The THP includes the feedback loop with a nonlinear modulo operation, and the tap coefficients of the THP are directly determined by the channel response, $H_{ch}(z)$. On the other hand, the FFE shows the feed-forward structure with the tap coefficients, $w_i$, which perform equalization by the inverse function of the channel. Fig.1(c) presents the swing limitation of the TX and the received data eye. In physical implementations, the output signal range of the TX driver, $M$, is limited. Therefore, the amplitudes of $D_{in}$ of THP and FFE for PAM-4 signaling ($A_{THP}$ and $A_{FFE}$) are represented below.

$$A_{THP} = \frac{3}{4} M.$$  \hspace{1cm} (1)

$$A_{FFE} = \frac{1}{1 + \sum_{i=1}^{M} |w_i|} M.$$  \hspace{1cm} (2)

In the THP, the nonlinear modulo operation adds or subtracts $M$ to the summation of the input and the feedback filter output when the summed signal exceeds the output range, $[−M/2, M/2]$, to guarantee that $Ch_{in,THP}$ remains within the driver output range. As a result, $A_{THP}$ remains the same for a high-loss channel, and the signal amplitude of $Ch_{in,THP}$ shows the uniform distribution. On the contrary, the output of the FFE ($Ch_{in,FFE}$) is summed without a modulo operation. Thus, $A_{FFE}$ is inversely proportional to the sum of $|w_i|$ and the distribution of $Ch_{in,FFE}$ is bell-shaped with signal power concentrated at the center. The amplitudes at the channel outputs ($Ch_{out,THP}$ and $Ch_{out,FFE}$) are derived by multiplication of the amplitude of the main cursor of the
channel (H₀) and the amplitude of Dᵦ, which are H₀AᵀHＰ and H₀AᶠＦＥ, respectively. For a high-loss channel, equalizer tap coefficients are increased, and then both H₀ and AᶠＦＥ are decreased. Consequently, the amplitude of Chₜₒᵤₜ,ＦＦＥ is reduced quadratically by the amount of ISI. However, the amplitude of Chₜₒᵤₜ,ＴＨＰ remains the same by virtue of the nonlinear modulo operation.

**B. PROPOSED FEED-FORWARD TOMLINSON-HARASHIMA PRECODING**

The design process of the proposed FF-THP is illustrated in Fig. 2. {D} and {k} denote sequences of the input data and the quotient resulting from a modulo operation for the present data. M represents the modulus of the modulo operation, and M corresponds to the maximum amplitude of the signal range of Chₜₒᵤₜ, which is [−M/2, M/2] as shown in Fig. 1. The FF-THP inherits the traditional THP operation, which has two main functions: a modulo operation to stabilize the output and a feedback equalization to compensate for a channel loss. These two key features are modified to build the FF-THP. Firstly, the modulo operation is replaced by the addition of a predicted modulo value, {kM}, to the input as shown in Fig. 2(b) and (c), which is essential for the next step of modification. Secondly, the feedback equalizer is reconstructed as the equivalent FFE with pretaps to remove a precursor ISI as shown in Fig. 2(d). Thus, the proposed FF-THP acquires the ability to remove precursors of a channel as well as keeping the modulo operation. The tap coefficients of the FFE are determined to maximize the vertical eye margin (VEM) at the channel output. Because of the increased number of signal levels, FF-THP has some drawbacks requiring a larger input range and more samplers of a receiver than conventional FFE, similar to THP. However, using the structure of FF-THP instead of the feedback equalizer, a feedback time constraint is completely removed in equalization, which enables a high-speed operation. Moreover, a larger eye opening and a larger SNR suitable for multi-level signaling are obtained by predictive modulo operation.

**C. z-DOMAIN ANALYSIS OF VERTICAL EYE MARGIN**

The primary function of an equalizer is providing a response to remove channel ISI. Assuming that a channel has one precursor and N postursors, the z-domain responses (ZDRs) of the channel and the normalized channel (Hᵦ(z) and hᵦ(z)) can be represented as (3) and (4), respectively.

\[
Hᵦ(z) = H₋₁z^1 + H₀ + H₁z⁻¹ + \cdots + H_Nz⁻^N \\
= \sum_{i=-1}^{N} H_i z^{-i},
\]

\[
hᵦ(z) = h₋₁z^1 + h₀ + h₁z⁻¹ + \cdots + h_Nz⁻^N \\
= \sum_{i=-1}^{N} H_i \frac{z}{H₀} z^{-i},
\]

where Hᵢ and hᵢ denote the magnitude of the iᵗʰ tap of a single bit response (SBR) and a normalized SBR, respectively. Since hᵦ(z) is normalized by the main cursor H₀, h₁ is equal to H₁/H₀ with H₀ = 1.

As shown in Fig. 2(a), the feedback filter of the THP is comprised of posttaps concerning only the postcursor of the channel (H₀) and a ZDR of each equalizer. With the amplitude adjusting coefficient is necessary for FFE [5]. Including the amplitude adjustment, the ZDRs of the FFE and the FF-THP using tap coefficients (w₋₂, w₋₁, w₀(= 1), w₁, ..., and wᴺ/₂) equalizing the channel ISI including the precursor are given below.

\[
Hᶠ𝐹𝐸(ｚ) = \frac{1}{1 + h₁z⁻¹ + \cdots + h_Nz⁻^N} = \sum_{i=0}^{N} a_i z⁻^i \times (a₀ = 1, a_n = \sum_{i=1}^{R} -h_i a_{n-i}, \ n = 1, 2, \ldots).
\]

(5)

On the other hand, both FFE and FF-THP have the ability to compensate precursors by using pretaps. Since the output range of the TX is limited between −M/2 and M/2, the amplitude adjusting coefficient is necessary for FFE [5]. Including the amplitude adjustment, the ZDRs of the FFE and the FF-THP using tap coefficients (w₋₂, w₋₁, w₀(= 1), w₁, ..., and wᴺ/₂) equalizing the channel ISI including the precursor are given below.

\[
Hᵗʰᵖ(z) = \frac{1}{1 + h₁z⁻¹ + \cdots + h_Nz⁻^N} = \sum_{i=0}^{N'} w_i z⁻^i.
\]

(6)

\[
Hᶠᶠ₋ᵗʰᵖ(z) = \sum_{i=-2}^{N'} w_i z⁻^i.
\]

(7)

An expression of VEM can be derived by multiplying a ZDR of a channel and a ZDR of each equalizer. With the
combined ZDR, \( R(z) \) representing the received signal, VEM in PAM-L signaling is described below.

\[
R(z) = \sum_i R_i z^{-i}, \quad \text{VEM}_R = \frac{R_0}{L-1} - \sum_{i \neq 0} |R_i|,
\]

(8)

where \( R_i \) denotes the \( i \)th coefficient of \( R(z) \). When a modulo operation is introduced, the amplitude of the data signal becomes \( M/L \), reduced from \( M/(L-1) \) in PAM-L signaling [4]. Therefore, for calculating VEMs of the THP and the FF-THP, (8) must be multiplied by the amplitude ratio of \((L-1)/L\). Calculating \( R(z) \) for three equalizers and using (8), VEMs are represented by (9-11) as follows, assuming that \( N \) and \( N' \) go to infinity.

\[
\text{VEM}_{\text{THP}} = \frac{H_0(1 - h_1 h_{-1})}{L} - \frac{(L-1)H_0}{L} \left(1 + \sum_{n=1}^{N} |\sum_{i=1}^{n} h_i a_{n-i}| \right).
\]

(9)

\[
\text{VEM}_{\text{FFE}} = \frac{H_0}{L - 1} \sum_{i=-2}^{1} |w_i| \left(\sum_{i=-2}^{1} |w_i h_{-i}| \right) - \frac{H_0}{L} \sum_{i=-2}^{1} |w_i| \left(\sum_{j \neq 0} \sum_{i=-2}^{1} |w_i h_{-i+j}| \right).
\]

(10)

\[
\text{VEM}_{\text{FFE-THP}} = \frac{H_0}{L} \sum_{i=-2}^{1} \left(\sum_{j \neq 0} \sum_{i=-2}^{1} \left| w_i h_{-i+j} \right| \right) - \frac{(L-1)H_0}{L} \left(\sum_{j \neq 0} \sum_{i=-2}^{1} \left| w_i h_{-i+j} \right| \right).
\]

(11)

According to the above equations, as the channel has a larger precursor, \( H_{-1} \), the VEM of the THP becomes smaller. Also, as tap coefficients to compensate channel ISI become larger, the VEM of the FFE becomes smaller compared with the VEM of the FF-THP.

**D. ONE POLE CHANNEL WITH ONE PRECURSOR**

To demonstrate the effect of the precursor and channel ISI, a hypothetical wireline channel is taken as an example with exponentially decaying postcursors and one precursor. In this case, the channel response (3) can be simplified to (12). Furthermore, assuming that the channel has unity DC gain, the \( H_0 \) can be represented by \( h_{-1} \) and \( h_1 \).

\[
H_{ch}(z) = H_0 h_{ch}(z) = H_0(h_{-1} z^{-1} + 1 + h_1 z^{-1} + h_1^2 z^{-2} + \cdots) = \frac{1 - h_1}{1 + h_{-1}(1 - h_1)} (h_{-1} z^{-1} + 1 + \sum_{i=-1}^{h_1} z^{-i}).
\]

(12)

Also, the ZDR of the THP, (5), is recalculated as (13).

\[
H_{\text{THP}}(z) = \frac{1}{1 + h_1 z^{-1} + h_1^2 z^{-2} + \cdots} = 1 - h_1 z^{-1}.
\]

(13)

Two pretaps and one posttap coefficient of FFE and FF-THP can be optimized for channel response (12). The tap coefficients are derived based on the partial differentiation of the ISI by each of \( w_{-2}, w_{-1}, \) and \( w_1 \). The optimized tap coefficients are shown below.

\[
w_{-2} = \frac{h_1^2}{1 - h_{-1} h_1}.
\]

(14)

\[
w_{-1} = \frac{h_{-1}}{1 - h_{-1} h_1}.
\]

(15)

\[
w_1 = -h_1 (1 - h_{-1} h_1).
\]

(16)

Applying (13-16) to (9-11), the optimized VEMs of the THP, the FFE, and the FF-THP are featured below with \( h_1 \) and \( h_{-1} \).

\[
\text{VEM}_{\text{THP}} = \frac{(1 - h_1)(1 - h_{-1}(L - 1 + h_1))}{(1 + (1 - h_1) h_{-1} L)}.
\]

(17)

\[
\text{VEM}_{\text{FFE}} = \frac{(1 - h_1)}{(1 + h_1)(1 + (1 - h_1) h_{-1}(L - 1))}
\times \frac{(1 - 3 h_1 h_{-1}(1 - h_{-1}) + ((L - 1) - h_1^3)(h_{-1})^3)}{(1 + (1 - 2 h_1) h_{-1} + (1 - h_1^2)(h_{-1})^2)}.
\]

(18)

\[
\text{VEM}_{\text{FFE-THP}} = \frac{(1 - h_1)}{(1 - h_{-1})}
\times \frac{(1 - 3 h_1 h_{-1}(1 - h_{-1}) + ((L - 1) - h_1^3)(h_{-1})^3)}{(1 + (1 - h_1) h_{-1} L)}.
\]

(19)

From (17-19), the calculated VEMs of THP, FFE, and FF-THP with respect to \( h_{-1} \) and \( h_1 \) for PAM-4 signaling are illustrated in Fig. 3. In Fig. 3(a), when \( h_{-1} \) is 0, the VEM of the THP and the FF-THP are the same. However, since the THP has no control over the precursor, as \( h_{-1} \) increases, the VEM of the THP becomes the smallest. As featured in the plots, the VEM of the FF-THP is the largest among the three.

A numerical simulation is conducted to compare the calculation with the simulation result. The SBR of the simulated channel and the simulated eye diagrams of THP, FFE, and FF-THP are presented in Fig. 4. Considering that natural channel response for \( \sim 20 \text{dB-loss channels features approximately} \) \( h_1 \) of 0.5 [12], to verify the effect of a precursor and
The operation of the TX is switched between FFE mode accurately compensate channel ISI and maximize the VEM. Externally controlled 10-bit coefficients for the two pretaps, implemented to remove the reflection from the channel. The UI pulses. Also, in the DAC, 50\textsuperscript{\textdegree}C. FEED-FORWARD EQUALIZER

B. MODULO PREDICTION ENGINE

The structure of the MPE is presented in Fig. 6. The inputs of the modulo table cell (MTC) are the two last PAM-4 data (D\textsubscript{0} and D\textsubscript{1}), the modulo values for both data (M\textsubscript{0} and M\textsubscript{1}), and the current PAM-4 data (D\textsubscript{2}). Then, it generates the modulo value for the current data (M\textsubscript{2}). It is worth noting that since the MTC depends only on the last two data and the modulo values, it is possible to apply the MTC to another channel if the first and the second posttaps (w\textsubscript{1} and w\textsubscript{2}) are similar to those of a target channel. However, since the MTC considers only w\textsubscript{1} and w\textsubscript{2}, the residual ISI that are not removed by w\textsubscript{1} and w\textsubscript{2} may cause modulo prediction error and induce the additional ISI. Because a wireline channel shows a similar response as a one-pole channel, w\textsubscript{1} and w\textsubscript{2} can sufficiently compensate for the channel response. Therefore, the residual ISI is negligible, and the other tap coefficients are much smaller than w\textsubscript{1} and w\textsubscript{2}. Also, even if a modulo prediction error occurs, when D\textsubscript{1} is −0.375, which corresponds to PAM-4 data 00, whether M\textsubscript{1} is 0 or 1, M\textsubscript{2} depends on D\textsubscript{2} as shown in the simplified table. Consequently, the modulo prediction error can be self-healed, and the burst error can be prevented.

A modulo operation in THP is calculated based on a direct summation of multiplications of data and taps of the feedback equalizer. In MTC, however, a modulo value is predetermined by a channel. Therefore, the burden of digital computation is much reduced. In addition, a modulo look-ahead (MLA) technique is used through 9 modulo prediction units (MPUs), each of which is comprised of two MTCs. They take combination sets of predetermined modulo values of \{-1, 0, 1\}×\{-1, 0, 1\} as previous modulo values (\{M\textsubscript{0}M\textsubscript{1}\}) and generate candidates for M\textsubscript{2} and M\textsubscript{3} \{(M\textsubscript{2}[−1\rightarrow 1] to M\textsubscript{2}[1\rightarrow 1]) and M\textsubscript{3}[−1\rightarrow 1] to M\textsubscript{3}[1\rightarrow 1]\}. The candidates are selected by the last modulo values, M\textsubscript{0} and M\textsubscript{1}. As a result, assisted by the MTC and the MLA technique, the digital computation operates with up to 1.25-GHz clock frequency.

To further enhance the data rate, there are two options: increasing the clock frequency and expanding the parallelism. The MTC is designed considering the first and the second posttaps. Still, since the modulo prediction error can be self-healed, the MTC can be simplified so that it only considers w\textsubscript{1} at the expense of slight degradation of BER. The simplified version of the MTC can enhance the clock frequency. Moreover, expanding the 4-parallel structure to 2\textsuperscript{N}-parallel can nominally increase the data rate by the factor of N-2. Thus, with the simplified MTC and the expansion of parallelism, the data rate can be increased significantly. Also, the MPE is purely a digital structure; immediate improvements in efficiency and data rate are expected for newer technologies.
of the output ($D_0$, $D_{90}$, and $D_{270}$) are generated by the same structure but the time-shifted input data. To generate $D_0$ and $D_{90}$, $D_{-2} + M_{-2}$ and $D_{-1} + M_{-1}$ are required, and they are derived from one clock delayed version of $D_2 + M_2$ and $D_3 + M_3$. Because of the benefit of the feed-forward structure, the FFE is straightforward for pipeline multiplications and summations.

For clarity, the pipelining in the figure is omitted but is implemented in the fabricated chip. As a result, contrary to THP, the digital computation of the FFE does not suffer from the timing issue and operates in high digital clock frequency.

The tap coefficients, $w_i$, corresponding to a specific channel, are determined to maximize a VEM by using the ArgMax function in Mathematica that finds global maximum with given constraints. Optimized for the same SBR, the ratios between the main tap and the other 12-tap coefficients ($w_i/w_0$) remain the same for the FFE and the FF-THP. Instead, the magnitude of the tap coefficients can be greater for the FF-THP because adding the modulo value guarantees that the output remains within the acceptable input range of the DAC driver.

**IV. MEASUREMENT RESULT**

The measurement setup for the 10 Gb/s PAM-4 transmitter is presented in Fig. 8. The vector signal generator generates
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FIGURE 9. Measured 10 Gb/s PAM-4 eye diagram and histogram of TX output of (a) FFE and (b) FF-THP. (c) Distribution of TX output of FFE and FF-THP.

FIGURE 10. (a) Measured insertion loss. (b) Normalized single bit response of the channel.

Decision Threshold \(X\) = \(\text{Mean}(\text{histo.}(X))\) 
\[\pm Q^{-1}(\text{BER})\text{Std.}(\text{histo.}(X)),\] 
where \(d\) and \(\sigma\) denote the magnitude of data and the standard deviation of Gaussian noise, respectively. \(d\) can be substituted by the difference between the mean of \(X\) and the data adjacent to \(X\), and \(\sigma\) can be substituted by the standard deviation of \(X\), respectively. Means and standard deviations of each PAM-4 data level can be obtained from the histogram of the received signal.

Fig. 11 exhibits the measured 10 Gb/s PAM-4 eye diagrams of the fabricated chip compensating the channel. When TX operates in the FF-THP mode, additional two levels appear along with the conventional PAM-4 levels as expected. The proposed FF-THP achieves the level mismatch ratio \(R_{LM}\) of 99.1%, and the VEM is improved by 38.9% compared with the FFE. From the histograms in Fig. 12(c) and (d), means and standard deviations of the data signal are obtained. Estimated based on Gaussian distribution, the decision thresholds and the bathtub curves of the FFE and the FF-THP are presented in Fig. 12(e) and (f). The proposed FF-THP achieves the BER lower than \(10^{-8}\) at the center of the eye and 87.5% increased horizontal eye margin (HEM) compared with the FFE at the BER of \(10^{-5}\).

Fig. 12 features the chip photomicrograph. The proposed TX occupies an active area of 0.075 mm². The power and the area breakdown of the fabricated chip are presented in Fig. 13. The digital area is 0.0322 mm² which takes 53.3% of total power. Without the PRBS generator, the FF-THP solely

a 78.125 MHz reference clock for PLL that generates a 1.25-GHz clock with a 1/16 divider. To measure the performance of the FFE and the FF-THP, display port cable and SMA cables are used. On the other hand, to measure the transmitter output, the output of the test chip is directly connected to the oscilloscope.

Fig. 9 exhibits the measured 10 Gb/s PAM-4 eye diagram and the histogram of the eye diagram. The eye diagram of the TX output features 800 mV of the output range. For this measurement, a lossy channel is not added. The distribution in Fig. 9(b) shows the centralized signal when the TX operates in FFE mode. On the other hand, when the TX operates in FF-THP mode, the signal of the FF-THP is evenly distributed. Because of the widespread distribution, the FF-THP features better SNR than the FFE.

The insertion loss and the normalized SBR of the measured channel are presented in Fig. 10. The channel loss is 21 dB at the Nyquist frequency of 2.5 GHz with the first postcursor of the channel around 0.5, which is the natural response of ~20 dB channel, as mentioned before. Also, the sum of the normalized ISI of the SBR is 1.48 times greater than that of the main cursor.

Before representing the measurements of the channel output of the proposed TX, it is necessary to mention a method that indirectly evaluates the BER performance of TX [15]. Assuming that Gaussian noise is added to the output data, BER for the PAM-4 signal and the decision threshold of the data \(X\) is represented by

\[BER_L = \frac{L - 1}{L} \cdot \text{erfc} \left( \frac{d}{2\sqrt{2}\sigma} \right) \times \log_2 L,\] 
(20)

where \(d\) and \(\sigma\) denote the magnitude of data and the standard deviation of Gaussian noise, respectively. \(d\) can be substituted by the difference between the mean of \(X\) and the data adjacent to \(X\), and \(\sigma\) can be substituted by the standard deviation of \(X\), respectively. Means and standard deviations of each PAM-4 data level can be obtained from the histogram of the received signal.
occupies 0.022 mm². With a 1-V supply, the total power consumptions of digital and analog blocks are 32 mW and 28 mW, respectively.

Table 1 compares the performance of the proposed FF-THP based TX with other PAM-4 TXs that compensate for a high channel loss or large ISI. The sum of channel ISI is an important parameter because VEMs of TX equalizers depend on it. Also, asymmetric link such as memory interface has multi drops, which are indicated by not the channel loss at Nyquist frequency but the sum of channel ISI. In the point of view of a channel ISI, the proposed design, assisted by the pretaps and the modulo-based signaling, can compensate 1.48 of the sum of the normalized ISI that is the largest. As a result, the FF-THP achieves the best FoM₂ of 4.05 pJ/b/ISI with lower than $10^{-8}$ BER.

**V. CONCLUSION**

This paper presents a PAM-4 TX introducing the FF-THP. The proposed TX takes both advantages of the modulo-based
equalization and the controllability over a precursor. Moreover, the quantitative z-domain analysis on channel response and the equalization parts of the THP, the FFE, and the FF-THP is conducted. A simple one pole channel with one precursor is employed to demonstrate the repercussions of the precursor and the effectiveness of the FF-THP. From the analysis, the FF-THP shows the largest VEM among the TX equalization methods when the channel has a precursor or large ISI. Also, considering Gaussian noise, the bathtub curve and the decision threshold voltage are indirectly estimated based on the histogram of the eye diagram.

The fabricated chip employs two pretaps and ten post-taps to compensate for the 21 dB channel loss. The proposed FF-THP presents 87.5% wider HEM at the estimated BER of $10^{-5}$ and 38.9% larger VEM compared with the FFE. In addition, the FF-THP achieves an estimated BER lower than $10^{-8}$ and the FoM2 of 4.05 pJ/b/ISI. Moreover, the digital-based equalization technique can take full advantage of process scaling. The prototype TX can comprehensively equalize any channel having both precursors or large ISI. Therefore, the proposed design can be applied to further high-speed wireline communication and multi-level signaling.

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JEONGHO (JEFFREY) HWANG (Member, IEEE) received the B.S. and Ph.D. degrees in electrical engineering and computer science from Seoul National University, Seoul, South Korea, in 2014 and 2019, respectively.

In 2019, he was a Postdoctoral Researcher with the Inter-University Semiconductor Research Center, Seoul National University. He joined Cadence Design Systems, in 2020, where he is currently a Lead Design Engineer. His research interests include high-speed I/O interfaces, silicon photonics, and on-chip reference oscillators. He was a recipient of the Korea Foundation for Advanced Studies during 2011–2018, the IEEE CASS Student Travel Grant Award in 2019, and the IEEE SSCS Student Travel Grant Award in 2019.

KWANSEO PARK (Member, IEEE) received the B.S. degree in electrical engineering from the Pohang University of Science and Technology, Pohang, South Korea, in 2013, and the Ph.D. degree in electrical and computer engineering from Seoul National University, Seoul, South Korea, in 2019.

In 2019, he was with the Inter-University Semiconductor Research Center, Seoul National University. He is currently a Postdoctoral Researcher at the University of California at Berkeley, Berkeley, CA, USA. His current research interests include the design of high-speed I/O circuits and agile hardware design methodology.

Dr. Park received the Distinguished Ph.D. Dissertation Award from the Department of Electrical and Computer Engineering, Seoul National University, in 2019, the IEEE Solid-State Circuits Society Student Travel Grant Award in ISSCC 2019, and the Best Design Award and the Best Poster Award at the IC Design Education Center Chip Design Contest, International SoC Design Conference, in 2016 and 2014, respectively.

HYOUNJUN KIM (Graduate Student Member, IEEE) received the B.S. degree in electronic engineering from Hanyang University, Seoul, South Korea, in 2017. He is currently pursuing the Ph.D. degree in electrical engineering with Seoul National University, Seoul. His current research interests include clock and data recovery (CDR) and phase-locked loop (PLL) for high-speed I/O applications.

HAN-GON KO (Member, IEEE) received the B.S. and Ph.D. degrees in electrical engineering and computer science from Seoul National University, Seoul, South Korea, in 2015 and 2021, respectively.

In 2021, he joined ONE Semiconductor, Suwon, South Korea, where he is involved in the design of I/O interface and clocking circuits for DDR4 and DDR5 RCD. His research interests include high-speed I/O interfaces, equalizers, PLL, and clocking circuits.

MOON-CHUL CHOI (Member, IEEE) received the B.S. degree in electrical and electronic engineering from Chung Ang University, Seoul, South Korea, in 2015, and the Ph.D. degree in electrical engineering and computer science from Seoul National University, Seoul, in 2021.

In 2021, he joined Samsung Electronics, Hwaseong, South Korea, where he has been involved in the design of DRAM, such as GDDR6 and GDDR7. His research interests include high-speed I/O circuits, transceivers, clock and data recovery (CDR), and development of high-speed DRAM.

DEOG-KYOON JEONG (Fellow, IEEE) received the B.S. and M.S. degrees in electronics engineering from Seoul National University, Seoul, South Korea, in 1981 and 1984, respectively, and the Ph.D. degree in electrical engineering and computer sciences from the University of California at Berkeley, Berkeley, CA, USA, in 1989.

From 1989 to 1991, he was a Member of the Technical Staff with Texas Instruments, Dallas, TX, USA, where he was involved in the modeling and design of BiCMOS gates and the single-chip implementation of the SPARC. He was with the Faculty of the Department of Electronics Engineering, Inter-University Semiconductor Research Center, Seoul National University, where he is currently a Professor. He was one of the co-founders of Silicon Image, now Lattice Semiconductor, Portland, OR, USA, which specialized in digital interface circuits for video displays, such as DVI and HDMI. His current research interests include the design of high-speed I/O circuits, phase-locked loops, and memory system architecture.

Dr. Jeong was a recipient of the ISSCC Takuo Sugano Award in 2005 for the Outstanding Far-East Paper.