A 1.62/2.7-Gb/s Adaptive Transmitter With Two-Tap Preemphasis Using a Propagation-Time Detector

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***Abstract*—A 1.62/2.7-Gb/s adaptive transmitter with two-tap preemphasis is presented. The tap coefficients are adjusted by detecting the propagation time through the channel with different lengths. This adaptive transmitter is fabricated in 0.13-μm CMOS technology, and the core area occupies 0.25×0.15 mm^2. The maximum power consumption from a 1.2-V supply is 32.4 mW at 2.7 Gb/s. For a 40-in FR4 PCB trace with a 12.3-dB loss, the measured RMS and peak-to-peak jitters of the recovered data are 30 and 231 ps, respectively, for a 1.62-Gb/s pseudorandom binary sequence (PRBS) of $2^{31}-1$. For a 20-in FR4 PCB trace with a 10.6-dB loss, the RMS and peak-to-peak jitters are 21 and 144 ps, respectively, for a 2.7-Gb/s PRBS of $2^{31}-1$. For both cases, the measured bit error rates are less than 10^{-12}.**

***Index Terms*—Coefficient adaptation, intersymbol interference (ISI) effect, preemphasis, transmitter, Video Electronics Standards Association (VESA) DisplayPort.**

I. Introduction

As THE data rate rises up to multigigabits per second in broadband communications, such as serial-link applications, the loss of the transmission channel becomes severe due to the limited bandwidth. The intersymbol interference (ISI) caused by skin effect and dielectric loss of the channel may degrade the timing jitter of received data and worsen the bit error rate (BER). To overcome this issue, the feedforward equalizer (FFE) at the transmitter is adopted to preemphasize the transmitted data [1]–[3]. Alternatively, the decision-feedback equalizer (DFE) and/or the analog equalizer at the receiver is adopted [4]–[7]. To reduce the ISI effect induced by the channel with different lengths such as copper cable and PCB, adaptive capability is needed for either the FFE or the DFE.

Several adaptation methods [8]–[12] are widely used to adjust the tap coefficients to compensate the unknown channel loss. In [8]–[10], the back channel or the bidirectional transceiver is needed to process the eye-opening information. In [11], the transmitter with an autoconfiguration technique is adopted, but the collaborating receiver is needed. In [12], an adaptation algorithm is used, but the external software and the connection via a serial interface are needed.

This brief presents a 1.62/2.7-Gb/s [13] adaptive transmitter with two-tap preemphasis by using a propagation-time detector, as shown in Fig. 1. By transmitting a step signal through the channel in Fig. 1, the propagation time is estimated to update the digitally controlled tap coefficients. Compared with prior literatures, this technique does not need an additional back channel or a collaborating receiver.

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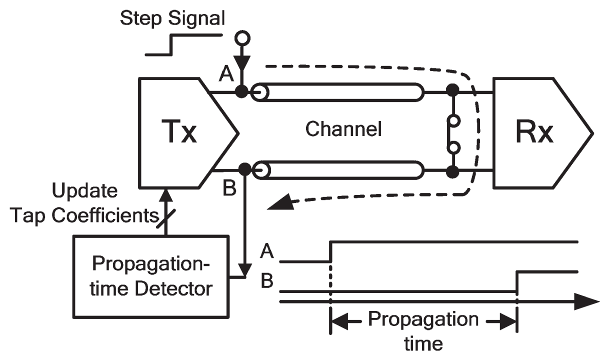


Fig. 1. Adaptive transmitter.

This brief is organized as follows. Section II introduces propagation-time detection, including the analysis of step response. The circuit description is presented in Section III. The experimental results are given in Section IV. Finally, the conclusions are given in Section V.

II. Propagation-Time Detection

In this brief, the channel is realized by the transmission line on an FR4 PCB. To reduce the ISI caused by the transmission line, the tap coefficients are adjusted for different channel lengths. Based on the relation between the loss and the propagation time of the transmission line, the tap coefficients are digitally calibrated in this brief. Propagation-time detection is described as follows.

In Fig. 1, when a step signal is transmitted through the transmission line, the propagation velocity v is given as

$$v = \frac{1}{\sqrt{\mu\varepsilon}} = \frac{2\ell}{t} \tag{1}$$

where μ, ε, ℓ, and t denote the permeability, permittivity, length, and propagation time of the transmission line, respectively. The loss characteristic $H(\omega)$ of the transmission line, which is dependent on the length and frequency, is expressed as

$$H(\omega) = \exp\left[-(k_s\sqrt{\omega} + k_d\omega)\ell\right] \tag{2}$$

where k_s, k_d, and ω denote the parameters of skin effect, dielectric loss, and frequency, respectively. The parameters of k_s and k_d are independent of frequency. By substituting (1) into (2), the loss characteristic is rewritten as

$$H(\omega) = \exp\left[-\frac{(k_s\sqrt{\omega} + k_d\omega)}{2\sqrt{\mu\varepsilon}}t\right] = \exp(-\eta t) \tag{3}$$

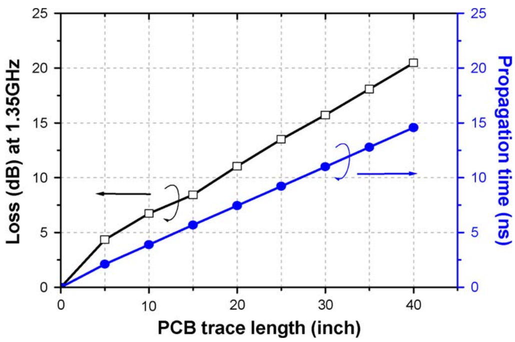


Fig. 2. Simulated loss at 1.35 GHz and propagation time versus length of the transmission line.

where

$$\eta = \frac{k_s\sqrt{\omega} + k_d\omega}{2\sqrt{\mu\varepsilon}}. \tag{4}$$

When the data rate is fixed, the parameter η is assumed constant. The loss of the transmission line is represented in decibels, and (3) is rewritten as

$$\text{loss (in decibels)} = -20\,\log\,H(\omega) = 8.69\eta t. \tag{5}$$

From (5), the propagation time is proportional to the loss (in decibels) of the transmission line. Thus, the loss of the transmission line can be estimated if the propagation time is known. To verify the above derivations, Fig. 2 shows the simulated loss at 1.35 GHz and the propagation time for the transmission line with different lengths.

Once the propagation time is detected, it is quantized into a 5-bit code, i.e., $D[4:0]$, by a 5-bit counter with a 100-MHz reference clock. Thus, the least-significant-bit time t_{LSB} is 10 ns. The estimated channel loss is given as

$$\text{loss}_{\text{est}} = 8.69\eta \cdot t_{\text{LSB}} \cdot D[4:0] = 8.69\eta \cdot t_{\text{LSB}} \cdot m \cdot a_i \tag{6}$$

where m is the ratio between this 5-bit code and tap coefficients a_i. When η, t_{LSB}, and m are properly determined, the tap coefficients are adaptively updated by this 5-bit code to compensate the loss of the transmission line with different lengths.

III. Circuit Description

A. Adaptive Transmitter Architecture

To verify the proposed technique, a 1.62/2.7-Gb/s dual-mode transmitter is implemented for the Video Electronics Standards Association (VESA) DisplayPort application. The adaptive transmitter with two-tap preemphasis is shown in Fig. 3. It is composed of two-tap digitally controlled preemphasis drivers, current-mode-logic D flip-flops (CML DFFs), and a propagation-time detector. This propagation-time detector is composed of a comparator, an exclusive-OR (XOR) gate, and a 5-bit counter with a 100-MHz reference clock.

Initially, the transmitter is operated in propagation-time detection mode, and the terminations of the receiver are isolated, as shown in Fig. 4(a). The input buffers would make the slopes of departure and returning step both larger than 1.2 V/ns. When the step signal is transmitted from one side of the transmission line, the signal Ctrl connected to a 5-bit counter would be high.

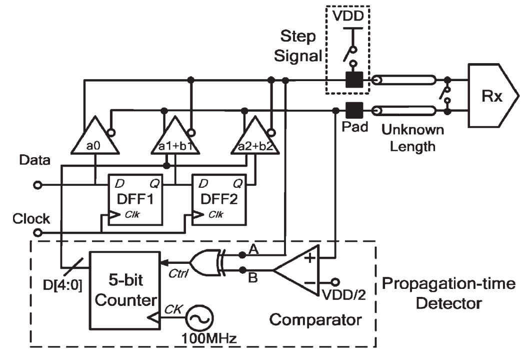


Fig. 3. Adaptive transmitter with two-tap preemphasis.

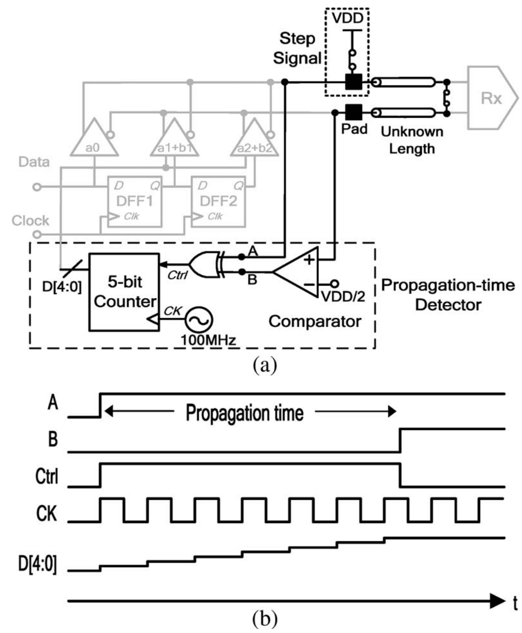


Fig. 4. (a) Transmitter operating in propagation-time detection mode. (b) Timing diagram.

The timing diagram of this mode is shown in Fig. 4(b). The 5-bit counter starts to count up until the step signal arrives in the other side of the transmission line. When the output of the comparator goes high, the signal Ctrl goes low. The 5-bit counter holds the 5-bit code to adjust the tap coefficients.

Then, the transmitter is switched to normal mode to transmit data with preemphasis. Since the propagation-time detector is disabled once the propagation time is detected, the additional power consumption and digital switching noise are eliminated. The building blocks are described as follows.

B. Circuits in This Transmitter

The transmitter with two-tap 5-bit digitally controlled variable tap coefficients is shown in Fig. 5. The main driver is controlled with the fixed tap coefficient a_0. The tap drivers are controlled with the fixed tap coefficients a_1 and a_2 and the variable tap coefficients b_1 and b_2, respectively. The fixed tap coefficients are used to cancel the offset loss of connectors. Thus, the transmitted output is written as

$$Y[n] = a_0 \cdot X[n] - (a_1 + b_1) \cdot X[n-1] - (a_2 + b_2) \cdot X[n-2]. \tag{7}$$

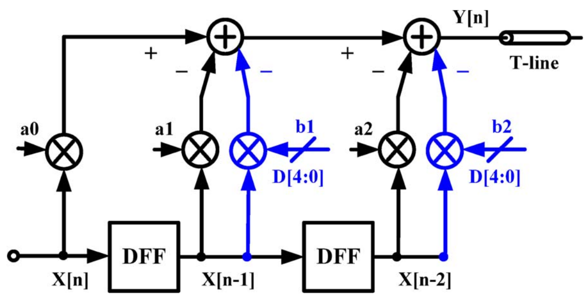


Fig. 5. Block diagram of the transmitter with 5-bit digitally controlled variable tap coefficients.

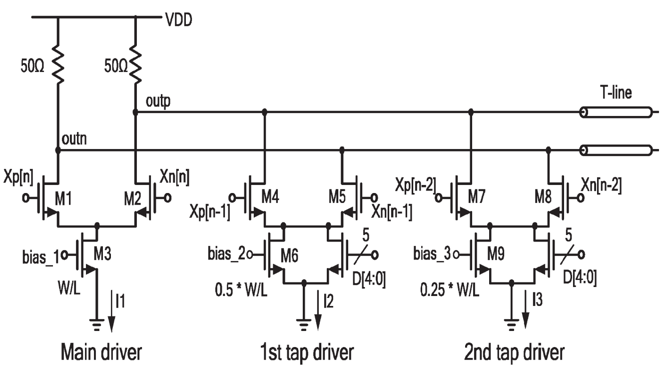


Fig. 6. Transmitter with two-tap 5-bit digitally controlled current sources.

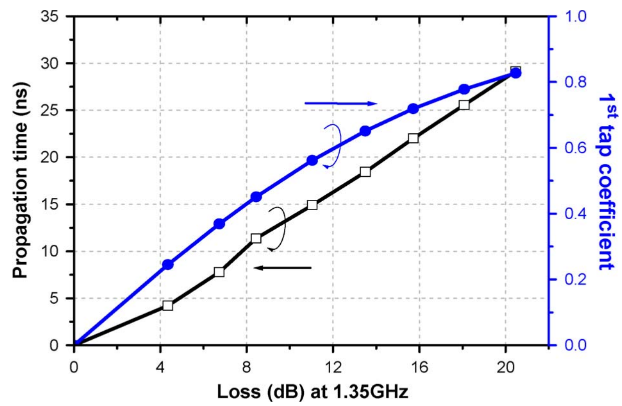


Fig. 7. Simulated propagation time and first tap coefficient versus loss at 1.35 GHz.

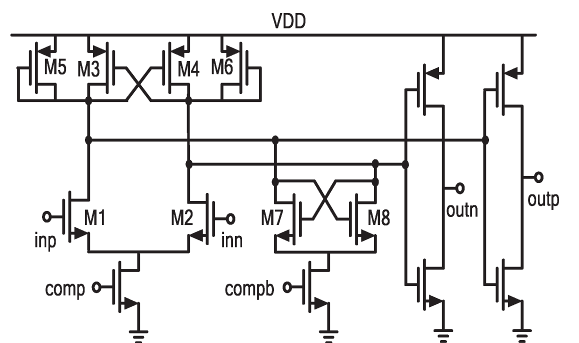


Fig. 8. Comparator.

The combined tap coefficients reduce the additional hardware of two DFFs and two tap drivers, and the output parasitic capacitance. The DFFs in Fig. 5 are realized by the rising-edge-triggered CML DFF. To relax the voltage headroom, the tail current of the CML DFF is removed. The clocking pairs are triggered by the ac-coupled clocks to realize current switching.

From (7), the transmitter with two-tap 5-bit digitally controlled current sources is realized, as shown in Fig. 6. The main driver is composed of the transistors M1–M3. The first and second tap drivers are composed of the transistors M4–M6 and M7–M9, respectively. The ratios of input differential pairs and current sources for the main, first tap, and second tap drivers are 1 : 0.5 : 0.25. In addition, the binary-weighted and digitally controlled current sources are used to adjust the variable tap coefficients.

Neglecting the second tap driver, which is used to further slightly cancel the second postcursor ISI, the preemphasis gain for the main and first tap drivers at a Nyquist frequency of 1.35 GHz is given as

$$\mathrm{Gain}_{\text{pre-E}} = 20 \log \left(\frac{I_1 + I_2}{I_1 - I_2} \right) \tag{8}$$

where I_1 and I_2 are the current sources of the main and first tap drivers, respectively. To achieve a preemphasis gain of 7.5 dB [13], the ratio of I_2/I_1 must be larger than 0.41. The range of the first tap coefficient is designed within 0 and 0.88. The simulated propagation time and corresponding first tap coefficient versus loss at 1.35 GHz are shown in Fig. 7. The parameter η in (5) is evaluated to be 0.081 dB/ns. It indicates that the loss of the channel with different lengths can be estimated from propagation-time detection. The corresponding tap coefficients are updated by digitally controlled current sources to compensate the channel loss.

C. Comparator and 5-Bit Counter

The comparator in the propagation-time detector is shown in Fig. 8. Diode-connected transistors, i.e., M5 and M6, and a cross-coupled pair, i.e., M3 and M4, are used to limit its swing and enhance its speed. Two inverters are used to make a rail-to-rail swing to drive the XOR gate. The simulated delay time of this comparator is 0.25 ns to properly work under a 100-MHz reference clock. The 5-bit up counter is realized by static CMOS logic. If the channel is performed in a non-FR4 channel or at different data rates, the frequency of the reference clock of the 5-bit counter is adjusted to control the compensation ratio.

IV. Experimental Results

This adaptive transmitter is fabricated in a 0.13-μm CMOS process. The die photo is shown in Fig. 9, and the core area is 0.0375 mm^2. For 10-, 20-, and 40-in FR4 PCB traces and a data rate of 1.62 Gb/s, the measured losses at a Nyquist frequency of 0.81 GHz are 2.6, 6.2, and 12.3 dB, respectively, as shown in Fig. 10. For 10- and 20-in PCB traces and a data rate of 2.7 Gb/s, the measured losses at a Nyquist frequency of 1.35 GHz are 3.5 and 10.6 dB, respectively.

To measure the propagation-time detection transient, a step signal is transmitted from one side of the channel. The Ctrl signal is set to high, and the 5-bit counter starts to count simultaneously. Until the step signal arrives to make the Ctrl signal go low, the 5-bit counter holds the 5-bit code. The measured 5-bit code for the 10-, 20-, and 40-in PCB traces are 2, 3, and 4, respectively. Fig. 11 shows the measured 5-bit code to be 4 for a 40-in PCB trace.

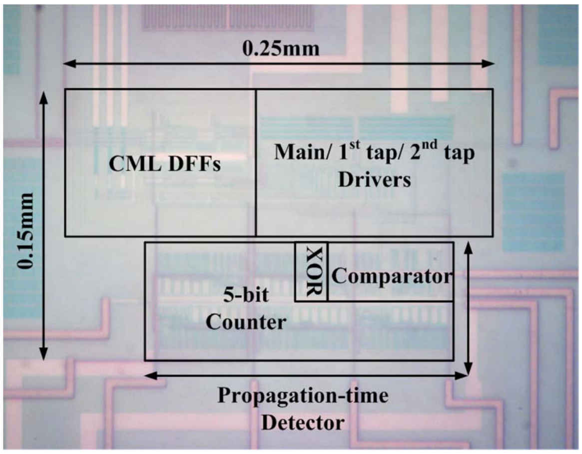


Fig. 9. Die photo.

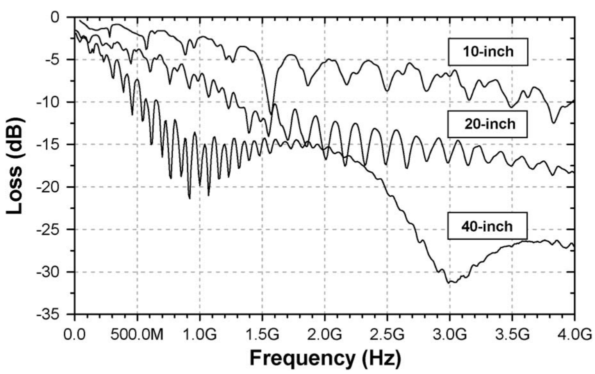


Fig. 10. Measured loss of 10-, 20-, and 40-in FR4 PCB traces.

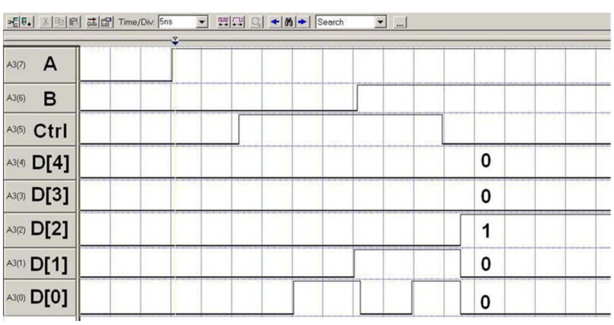


Fig. 11. Measured propagation-time detection transient.

For a 1.62-Gb/s PRBS of $2^{31}-1$, Fig. 12(a) shows the measured eye diagram at the transmitter output without preemphasis. The output swing is 230 mVpp. After passing through a 40-in FR4 PCB trace with a 12.3-dB loss, the eye diagram is shown in Fig. 12(b). Without preemphasis, the measured eye is almost closed, and the jitter cannot be measured. However, by calculating the width of the eye diagram, the estimated RMS and peak-to-peak jitters for the received data are 46 ps (0.075 UI) and 309 ps (0.5 UI), respectively. The transmitter output with preemphasis before transmitted to the PCB trace is shown in Fig. 12(c). The received data at the PCB trace output are shown in Fig. 12(d). With preemphasis, the measured RMS and peak-to-peak jitters for the received data are 30 and 231 ps, respectively. The power consumption values without and with preemphasis from a 1.2-V supply are 21.6 and 30 mW, respectively.

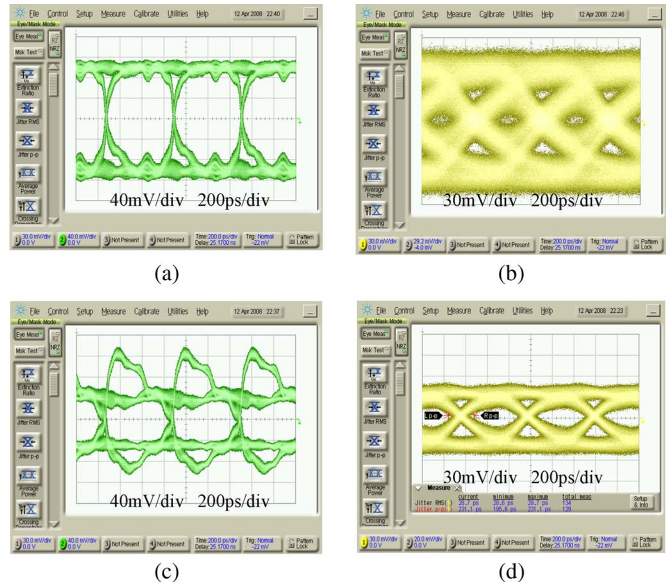


Fig. 12. Measured eye diagrams for a 1.62-Gb/s PRBS of $2^{31}-1$. (a) Transmitter output without preemphasis. (b) Output over a 40-in PCB trace without preemphasis. (c) Transmitter output with preemphasis. (d) Output over a 40-in PCB trace with preemphasis.

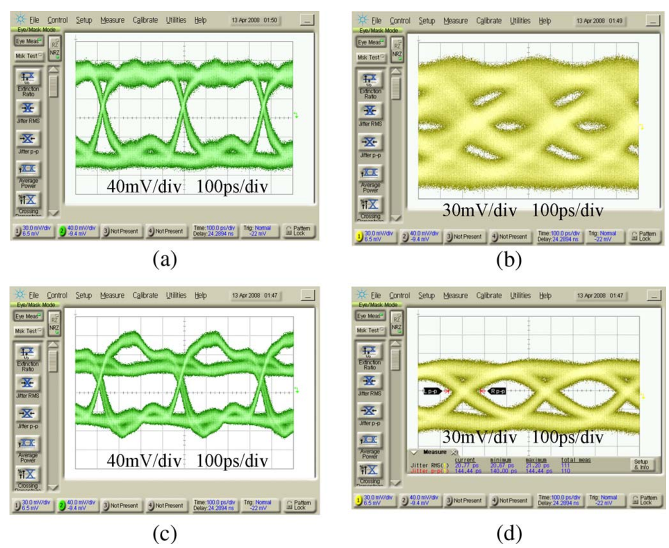
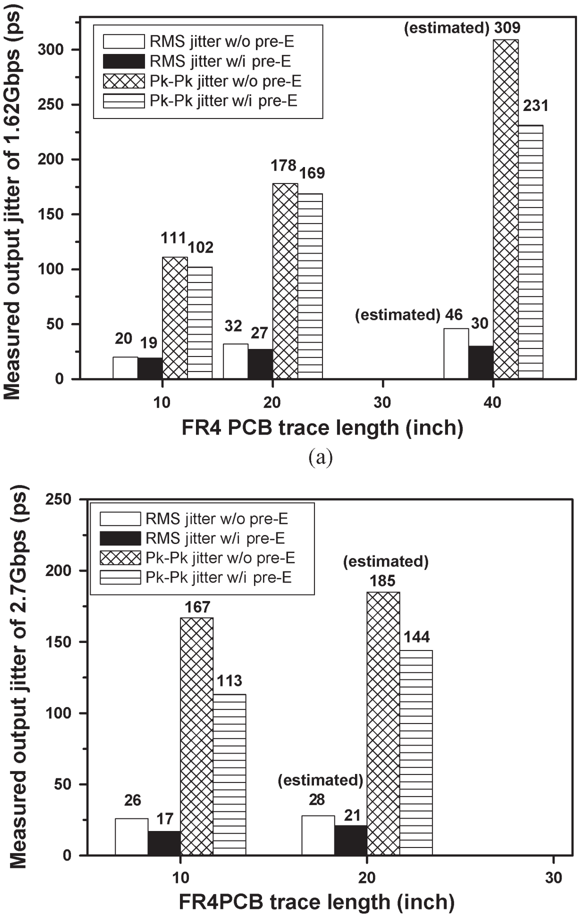


Fig. 13. Measured eye diagrams for a 2.7-Gb/s PRBS of $2^{31}-1$. (a) Transmitter output without preemphasis. (b) Output over a 20-in PCB trace without preemphasis. (c) Transmitter output with preemphasis. (d) Output over a 20-in PCB trace with preemphasis.

For a 2.7-Gb/s PRBS of $2^{31}-1$, Fig. 13(a) shows the measured eye diagram at the transmitter output without preemphasis. The output swing is 230 mVpp. After passing through a 20-in FR4 PCB trace with a 10.6-dB loss, the eye diagram is shown in Fig. 13(b). Without preemphasis, the estimated RMS and peak-to-peak jitters for the received data are 28 and 185 ps, respectively. The transmitter output with preemphasis before transmitted to the PCB trace is shown in Fig. 13(c). The received data at the PCB trace output are shown in Fig. 13(d). With preemphasis, the measured RMS and peak-to-peak jitters for the received data are 21 and 144 ps, respectively. The power consumption values without and with preemphasis from a 1.2-V supply are 25.2 and 32.4 mW, respectively. For VESA DisplayPort application [13], the width of the received eye diagram must be wider than 0.509 UI with a maximum loss of 7.5 dB at



(a)

(b)

Fig. 14. Measured jitters versus FR4 PCB trace length for a PRBS of $2^{31}-1$ at (a) 1.62 Gb/s and (b) 2.7 Gb/s.

TABLE I
SUMMARY OF ADAPTIVE PREEMPHASIS

Parameter		Value		
Length of PCB trace		10-inch	20-inch	40-inch
Channel loss (dB) at	0.81GHz	2.6dB	6.2dB	12.3dB
	1.35GHz	3.5dB	10.6dB	
Output code of the counter		2	3	4
Measured 1st tap coefficient (a1+b1)		0.32	0.38	0.44
Optimal 1st tap coefficient	at 1.62Gbps	0.15	0.34	0.61
	at 2.7Gbps	0.2	0.54	

a Nyquist frequency of 1.35 GHz. In Fig. 13(d), the calculated width of the received eye diagram is equal to 0.611 UI.

The measured jitters of the received data with and without preemphasis versus FR4 PCB trace length for the 1.62- and 2.7-Gb/s PRBSs of $2^{31}-1$ are shown in Fig. 14(a) and (b), respectively. The maximum power consumption from a 1.2-V supply is 32.4 mW with preemphasis at 2.7 Gb/s. The BER is less than 10^{-12}. Table I gives the summary of adaptive preemphasis. The maximum difference between the measured and calculated optimal tap coefficients is 0.17. Finally, Table II gives the performance summary of the proposed transmitter and comparison with other literatures. Compared with previous works, this work has a low power since the propagation-time detector can be turned off once the calibration is done.

TABLE II
SUMMARY AND COMPARISON

	[2]	[3]	[12]	This Work
CMOS	0.13μm	90nm	65nm	0.13μm
Supply	1.2V	1V	1.2V	1.2V
Power	110mW	24mW	45mW	32.4mW
Core area	1 mm^2 (whole chip)	0.036 mm^2	0.2 mm^2	0.0375 mm^2
Data rate	5Gbps	6Gbps	7.5Gbps	1.62/2.7Gbps
Channel loss	33dB	10dB	17dB	12.3/10.6dB
PRBS	2^7-1	2^{31}-1	2^{31}-1	2^{31}-1
BER	< 10^{-12}	<10^{-9}	<10^{-15}	<10^{-12}
Adaptation	No	No	Yes (Software)	Yes (Digital)

V. CONCLUSION

An adaptive transmitter with two-tap preemphasis has been presented. The relation between the loss and the propagation time of a transmission line has been analyzed. The propagation-time detector has been used to measure the propagation time and adjust the tap coefficients. This adaptive transmitter has been fabricated in a 0.13-μm CMOS process. The experimental results have also demonstrated that this adaptive transmitter can compensate the channel loss and reduce the ISI effect.

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