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The Design of Clocked-Comparator-Based Time-Interval Measurement Circuit for Pulse ToF Measurement

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Abstract—A clocked-comparator-based time-interval measurement approach is proposed for the measurement of the ToF of wave pulses in a threshold-crossing-detection-based pulse ToF range-finding application. As compared to the conventional TDC-based pulse ToF measurement approach, the proposed approach does not require the use of extra channels in the case where multiple echoes can be expected for every wave pulse launched. As compared to the conventional (flash) ADC-based pulse ToF measurement approach, the proposed approach allows for a simpler backend digital processing and does not require the comparators used in the measurement to be skew-less. Meanwhile, the approach leverages on the strength of clocked comparator circuit which can be designed with a very wide sampling bandwidth. A clocked-comparator-based time-interval measurement circuit with a ~ 7 GHz bandwidth was assembled using low cost COTS components. The measurements done using the measurement circuit agree to within 10 ps with the same measurements done using a 13 GHz, 40 GSa/s oscilloscope and are precise to within ± 15 ps in terms of their 2σ range.

Index Terms—clocked comparator, pulse time-of-flight, time-interval measurement, time to digital conversion.

I. INTRODUCTION

RANGEFINDER measures the distance between itself and an object of interest by measuring the ToF (Time-of-Flight) of a wave, typically sound [1-5], radio [6, 7], or light wave [8-18], beamed from the rangefinder towards the object and back. Knowing the propagation speed v of the wave, the distance between the rangefinder and the object is equal to $v \cdot \Delta t / 2$ where Δt is the time needed by the wave to travel from the rangefinder to the object and back.

Rangefinders that utilize a pulse-shaped wave are typically used for range measurements that require a relatively simple wave generation and echo processing, a high precision result, a high speed acquisition, and the capability to deal with echoes which may vary greatly in strength and which may come from multiple objects [4, 9, 12, 17, 18]. In a pulse ToF range-finding, the narrower the wave pulses, the better the measurement accuracy, precision, and capability to separate multiple objects located at similar distance from the rangefinder [4, 7, 18].

To determine the time when a pulse echo is received, a pulse ToF rangefinder may implement a correlation or threshold crossing detection. Correlation-based detection is the optimum approach in maximum-likelihood-sense to detect pulses in the presence of additive noise. However, the electronics needed to implement a correlation receiver is relatively complex due to the complexity of correlation process and the necessity for the receiver to work linearly. Hence, depending on the needs, threshold crossing detection is still used in a lot of pulse ToF rangefinders including the recent ones [5, 6, 13-18].

When threshold crossing detection is used in a pulse ToF rangefinder design, since the received pulse echo power may vary, a phenomenon called walk (see Fig 1) will occur and consistent pinpointing of a set reference time point cannot be done using a simple threshold crossing detection: the crossing time of a threshold will vary about the reference time point and errors of almost the same magnitude as the rise time of the wave pulse used in the measurement may occur. Typically, a wave pulse with a rise time of < 2 ns is used in a threshold-crossing-detection-based pulse laser and radio range-finding. This corresponds to a potential distance measurement errors of < 30 cm when a simple threshold crossing detection is used. To reduce the effect of walk without increasing the bandwidth of the wave pulse used, either a compensation which is based on a feature of the received pulse echo signal (width, rise time, etc) [14-17] or an interpolation done to determine the onset of the rising edge of the received pulse echo signal [5, 6] is used. In general, the method which relies on more threshold crossings is more resilient to the effect of fading channel and closely spaced multiple echoes which are relevant in some applications [6].

Fig. 1. Walk due to variation in the received pulse echo power in pulse ToF rangefinder utilizing threshold crossing detection for pulse echo detection.

Now, a pulse ToF rangefinder may rely on a TDC (Time-to-Digital Converter) or an ADC (Analog-to-Digital Converter) based approach for the measurements of the ToF of the wave pulses launched. If it relies on a TDC-based approach, the rangefinder will require at least two TDC channels for each pulse echo that is expected to be received for every wave pulse launched in order to allow for the implementation of pulse ToF measurements that are less affected by walk. Combined with the ps resolution required for the precision and accuracy of the measurement, the resulting multi-channel TDC circuit can be very large as demonstrated in [11]. Meanwhile, an ADC-based pulse ToF rangefinder will only require one ADC channel for the implementation of ToF measurements with reduced walk effect. However, in a pulse ToF range-finding that utilizes wave pulse with relatively wide bandwidth, the use of high-sample-rate ADC is required. For example, to precisely reproduce the shape of the edges of a pulse with 1.5 ns rise/fall time, an ADC with multi GSa/s sampling rate will be required.

This paper presents an alternative approach to the measurement of time-interval which is a compromise between the above mentioned approaches. According to the proposed approach, to measure the ToF of a wave pulse, one or more clocked comparator channels (also known as dynamic comparator) are used to repeatedly compare the received pulse echo signal to one or a set of thresholds (Fig. 2). One channel is needed for each threshold required in the measurement and no extra channels are needed should multiple echoes are expected to be received for every wave pulse launched.

This paper is organized as follows: Section II discusses on how clocked comparator can be used to measure a time-interval with ps resolution. Section III discusses on a number of issues that may affect the accuracy of time-interval measurements done according to the proposed approach adversely and on the strategies that can be used to mitigate them. In Section IV, a method to characterize the sampling bandwidth of a clocked comparator is introduced. The sampling bandwidth of the clocked comparator used is an important parameter in the design of a time-interval measurement circuit done according to the proposed approach. Lastly, Section V will present an experiment which demonstrates the design of an accurate time-interval measurement done according to the proposed approach using low cost COTS (Commercial-Of-The-Shelf) components and Section VI will conclude this paper. This paper does not intend to discuss on how to design a clocked comparator circuit. This information can be found from a number of other readily available sources [19-21].

II. CLOCKED-COMPARATOR-BASED TIME-INTERVAL MEASUREMENT

A time-interval measurement (e.g. the measurement of pulse ToF, pulse width, rise/fall time, propagation delay, etc) typically involves the quantification of the interval Δt between the instant when a signal v_i or a pair of signals $v_{i,a}$ and $v_{i,b}$ each, in succession, crosses a set threshold value v_{th} or a pair of set threshold values $v_{th,a}$ and $v_{th,b}$. Mathematically:

$$\Delta t = t_b - t_a \quad (1)$$

where t_a and t_b is the instant when the earlier and the latter threshold crossing occurs, respectively. The value of $v_{th,a}$ and $v_{th,b}$ may or may not be the same and the relevant threshold crossings, each, may be the one that occurs when the value of v_i or $v_{i,a}/v_{i,b}$ is changing in the + or – direction.

A. The Proposed Approach

A clocked-comparator-based Δt measurement is carried out by periodically comparing a given signal v_i or a pair of given signals $v_{i,a}$ and $v_{i,b}$ to one or a pair of set threshold values (v_{th} or $v_{th,a}$ and $v_{th,b}$) as illustrated in Fig. 3. If v_{th} or $v_{th,a}/v_{th,b}$ and v_i or $v_{i,a}/v_{i,b}$ is fed into the – and + input of a comparator, respectively, a logic “1” at the output of the comparator means that the currently sampled value of v_i or $v_{i,a}/v_{i,b}$ is above the respective threshold value and vice versa. In this way, we can make distinctions between whether neither, either one, or all of the thresholds have been crossed by comparing the present and the past output d_o or $d_{o,a}/d_{o,b}$ of the comparators. This is basically the generalization of the propagation delay measurement method we proposed in [22].

Fig. 2. The block diagram of a pulse ToF rangefinder which utilizes clocked comparators for pulse ToF measurement according to the proposed approach.

Fig. 3. (a) A pulse width Δt_w measurement, (b) a rise time Δt_r measurement, (c) A ToF Δt_{ToF} measurement, and (d) a propagation delay Δt_p measurement done according to the proposed approach.

To measure a Δt with high precision, the interval has to be measured with high resolution Δt_{res} so that the quantization noise of the measurement result is small. The quantization error and the quantization noise of a Δt measurement done according to the proposed approach ranges between $-\Delta t_{res}$ to $+\Delta t_{res}$ and is equal to $\Delta t_{res}/\sqrt{3}$, respectively. Meanwhile, the Δt_{res} of a Δt measurement done according to the proposed approach depends on the effective sampling rate $f_{s,eff}$ of the measurement, i.e.:

$$\Delta t_{res} = \frac{1}{f_{s,eff}} \quad (2)$$

A clocked comparator, however, has a limit to how fast it can be repeatedly strobed. To measure a Δt with a resolution of tens of ps or less, either one or a combination of time-interleaving and equivalent-time sampling should be used. A multi-phase output DLL (Delay Locked Loop) can be used to generate a set of clock signals with tens of ps phase difference [23], allowing the implementation of time-interleaved Δt measurement with tens of ps resolution. If coherent equivalent-time sampling is used, Δt_{res} can be calculated to be equal to as follows:

$$\Delta t_{res} = \frac{T_i T_\phi}{LCM(T_i T_\phi)} \quad (3)$$

where T_i and T_ϕ is the period of v_i or $v_{i,a}$ and $v_{i,b}$ and the clock signal ϕ used to strobe the clocked comparators used in the measurement, respectively, and LCM is the least common multiple operator [22]. A Δt measurement with ps resolution can be achieved through the use of this sampling method.

B. Comparison with the Traditional Approaches

When a group of comparators are used to detect the crossing of a number of thresholds by a certain signal, the arrangement of the comparators used will be similar to the way comparators are arranged in a flash ADC. The role of these comparators, however, are different. In the case of the proposed approach, each comparator is a separate Δt measurement unit and the output of the comparators are not combined together to form a digital word as in the case of a flash ADC. Due to this reason, the digital back-end processing required for Δt measurement purpose is simpler in the case of the proposed approach as shown in Fig. 4. In this case, the circuitries required to produce a digital time-stamp every time a threshold crossing occurs are a logic transition detector (can be a XOR gate and a shift register) for each comparator and a shared counter used to count the current sample number. Meanwhile, to do the same, an output aggregator (a bubble-error-corrected thermometer code encoder or a Wallace tree adder), a set of digital comparators, a set of logic transition detectors, and a counter are needed to process the output of the comparators of a flash ADC.

As compared to the TDC-based Δt measurement approach, the proposed approach has a similar level of complexity. As illustrated in Fig. 5, the circuit architecture of each Δt measurement channel of the time-interleaved implementation of the proposed approach is similar to that of a Vernier TDC typically used in a high resolution TDC-based Δt measurement. The main difference is just that there is no string of inverters in the input signal v_i path in the case of the proposed approach which corresponds to the lead signal v_{lead} path of a

Fig. 4. The digital back-end circuitries needed to determine the instants of threshold crossing for (a) a flash ADC (2-bit/3-threshold example) and (b) the proposed approach (3-threshold example).

Fig. 5. The circuit architecture of (a) a Vernier TDC with calibration function and (b) a Δt measurement channel of the time-interleaved implementation of the proposed approach.

Vernier TDC circuit. Meanwhile, the circuit structure of the arbiter used in a Vernier TDC can be very similar to that of a strongARM comparator which is a type of clocked comparator.

III. ACCURACY CONSIDERATIONS

A. DC-Offset

Comparator DC-offset ϵ_{th} is a measure on the difference between the effective threshold value $v_{th,eff}$ and the assigned threshold value v_{th} of a comparator:

$$\epsilon_{th,a} = v_{th,a,eff} - v_{th,a} \quad (4)$$

$$\epsilon_{th,b} = v_{th,b,eff} - v_{th,b} \quad (5)$$

The amount of Δt measurement error in a clocked-comparator-based Δt measurement that is caused by DC-offset is equal to:

$$\delta_{off} = \frac{\epsilon_{th,b}}{\frac{d}{dt}v_i(t_b)} - \frac{\epsilon_{th,a}}{\frac{d}{dt}v_i(t_a)} \quad (6)$$

$$\delta_{off} = \frac{\epsilon_{th,b}}{\frac{d}{dt}v_{i,b}(t_b)} - \frac{\epsilon_{th,a}}{\frac{d}{dt}v_{i,a}(t_a)} \quad (7)$$

when the measurement has one given signal v_i or two given signals $v_{i,a}$ and $v_{i,b}$, respectively.

δ_{off} can be reduced by adjusting $v_{th,a}$ and $v_{th,b}$ such that $v_{th,a,eff}$ and $v_{th,b,eff}$ are closer to their intended values or by implementing a clocked comparator offset reduction circuit technique such as the one proposed in [24].

B. Skew

The skew between the input signal paths $\delta_{i,skew}$ and strobing signal paths $\delta_{\phi,skew}$ to each of the clocked comparator channel used in a clocked-comparator-based Δt measurement affects the accuracy of the measurement adversely. This issue, however, is only relevant when more than one clocked comparator channels are used in the measurement. The amount of Δt measurement error caused by the skews is equal to:

$$\delta_{skew} = \delta_{i,skew} - \delta_{\phi,skew} \quad (8)$$

δ_{skew} can be measured by setting the threshold value of the comparators used in the measurement to the same effective value and shorting the v_i input of the comparators together. Knowing δ_{skew} , δ_{skew} can be offset from the measurement results.

C. Sampling Bandwidth

The role of a clocked comparator is to convert a continuous-time signal into a discrete-time one and to compare the discrete-time version of the continuous-time signal to a certain threshold value. The sampling bandwidth of a clocked comparator can be thought of as the bandwidth of the sample and hold function of the comparator if it can be clocked at ∞ Hz. The sampling bandwidth of the clocked comparator circuit used in a Δt measurement done according to the proposed approach plays an important role to the accuracy and/or the proper implementation of the measurement.

Firstly, the delay across a linear circuit is basically inversely related to its bandwidth. At 50% threshold, the delay across a linear circuit is around the 10% to 90% rise time of the step response of the circuit [25] which is inversely proportional to the bandwidth of the circuit [26]. Hence, a circuit with a wider bandwidth will suffer from a smaller amount of delay variation due to, for example, PVT (Process, Voltage, and Temperature) variation.

Now, suppose that a pulse ToF measurement is done according to the proposed approach and, to reduce walk effect, the pulse-width-based walk error compensation technique proposed in [14] is implemented in the measurement. Here, the width of received pulse echoes are measured at a fixed voltage value which is also the detection threshold for the pulses to allow for a low-complexity but accurate walk-error-compensated pulse ToF measurements. Now, if the sampling bandwidth of the clocked comparator used in the measurement reduces, the walk error of each pulse echo received will appear to be larger as illustrated in Fig. 6. At the same time, since a low detection threshold is typically used for improved detection sensitivity and dynamic range, the pulse echo widths as measured at a fixed voltage level will also appear to be larger. However, in this case, a wider width should actually correspond to a smaller walk error. Hence, if the walk error compensation curve generated under the typical case is used here, the amount of compensation applied will be too small. The opposite will happen if there is an increase in the sampling bandwidth.

Fig. 6. Compensation error in a pulse-width-based walk error compensation due to change in bandwidth (reduction). δ_2 is the amount of compensation that will be applied to the measurement results and is insufficient. The opposite will happen if, instead, bandwidth increases.

Meanwhile, if a pulse ToF measurement is based on the onset of the received pulse echoes which is determined through an interpolation, the previously said delay variation will not be detrimental to the accuracy of the measurement. This is due to the nature of interpolation in which the interpolated points will just get adjusted according to the shape of the received pulse echo. In this case, the clocked comparator circuit used in the measurement does not need to have a very wide sampling bandwidth and can be optimized in different regards. However, the sampling bandwidth of the circuit should still be sufficient so that the received pulse echo signal is not significantly attenuated due to insufficient bandwidth.

IV. CLOCKED COMPARATOR INPUT FREQUENCY RESPONSE CHARACTERIZATION METHOD

The sampling bandwidth of a clocked comparator that is to be used in a Δt measurement done according to the proposed approach need to be characterized prior to the measurement for the reasons explained in Section III C. [27] and [28] each proposes a method to characterize the input frequency response of a clocked comparator. The method proposed in [28] is only applicable for characterizations that are done through simulations since it requires the generation of pulses with a very narrow pulse width. While the method proposed in [27] is applicable for both simulated and experimental characterizations, the method requires the generation of step signals with very sharp edges that need to be placed accurately. ps level accurate placement is needed to accurately characterize a clocked comparator with multi GHz sampling bandwidth.

The proposed characterization method comes from a certain understanding on how clocked comparators work. Basically, a clocked comparator behaves linearly in the early processes of its comparison cycles, i.e. a clocked comparator processes the difference of its input signals linearly before deciding on a logic “1”/“0” output. Fig. 7 shows a model that describes the operation of a clocked comparator. The model is adapted from [20, 28] which consists of a subtractor, an LTV (Linear Time-Varying) filter, an ideal sampler, and a slicer.

The impulse response of an LTV system is time-varying and described by the function $h(t, \tau)$ where τ is the time of arrival of the impulse. The output $v_o(t)$ of the LTV filter block of the model can be related to its input $v_i(t) = v_{i+}(t) - v_{i-}(t)$ through:

$$v_o(t) = \int_{-\infty}^{+\infty} v_i(\tau) \cdot h(t, \tau) \cdot d\tau \quad (9)$$

If the LTV filter always responds the same way every time the clocked comparator is strobed at, say, $t = t_n$ and $t = t_m$ where n and m are integer indexes, assuming that the responses of the LTV filter have always completely died down first before a latter strobe is applied, $h(t+t_n, \tau+t_n)$ will be equal to $h(t+t_m, \tau+t_m)$. Now, based on the model, the overall output $d_o[n]$ of a clocked comparator is either a logic “1”/“0” depending on the output $v_s[n]$ of the ideal sampler block of the model in which $d_o[n]$ will be equal to logic “1” when $v_s[n] > 0$ and vice versa. $v_s[n]$ is an instantaneous sample of $v_o(t)$ t_{obs} seconds after the comparator is strobed the n -th time. Hence, the value of $d_o[n]$ depends on the following integration:

$$v_s[n] = v_o(t_n + t_{obs}) = \int_{-\infty}^{+\infty} v_i(\tau) \cdot \Gamma(\tau, n) \cdot d\tau \quad (10)$$

where $\Gamma(\tau, n) = h(t_n + t_{obs}, \tau)$. Fig. 8-9 illustrate above processes.

$\Gamma(\tau)$ (assuming that $\Gamma(\tau, n) = \Gamma(\tau, m)$) is dubbed as the ISF (Impulse Sensitivity Function) of a clocked comparator. It is a clocked comparator’s weighting factor across time. Every time a clocked comparator is strobed, v_i will be averaged across time according to the comparator’s ISF and the current value of v_s will be equal to the result of the averaging process. Should we be able to slide the strobing position of a clocked comparator continuously and record the corresponding v_s value for each strobe position, the resulting continuous time waveform is the cross correlation between v_i and Γ . In frequency domain:

$$V_s(j\omega) = V_i(j\omega) \cdot \Gamma^*(j\omega) \quad (11)$$

where $*$ is the complex conjugate unary operator. The input frequency response of a clocked comparator is equal to $\Gamma^*(j\omega)$.

Knowing that a clocked comparator averages the difference of its input signals across time starting from the time it is strobed prior to deciding on a certain output, the clocked comparator input frequency response characterization method whose setup and procedure is illustrated in Fig. 10 is proposed. The method requires 3 sweep types: 1) a sweep in the frequency ω of a sinusoidal test signal v_i fed into one of the comparator inputs, 2) several sweeps in the value of a DC threshold signal v_{th} fed into the other comparator input, and 3) several sweeps in the strobing position of the comparator across each period of v_i .

The sweeps are done to determine the apparent amplitude A_s of v_i as v_i is time-averaged across a number of v_i frequency values ω_n where n is an integer index. The goal is to construct a $|\Gamma(j\omega)|$ plot for the comparator. A_s can be determined as:

$$A_s = \frac{v_{th,m} - v_{th,n}}{2} \quad (12)$$

where $v_{th,m}$ and $v_{th,n}$ is the upper and lower v_{th} value in which d_o ceases to toggle as v_{th} is increased and decreased, respectively, as the strobing position of the comparator is swept across the period of v_i . Strobing the comparator with a v_{th} above and below them will always give a logic “0” and “1” output, respectively.

When the method is used to experimentally characterize the input frequency response of a clocked comparator, the ω sweep and v_{th} sweeps can be implemented using an RF signal generator and a DAC (Digital to Analog Converter), respectively. To implement the strobing position sweeps, the

Fig. 7. A model describing the operation of a clocked comparator.

Fig. 8. The time-varying impulse response of the LTV filter block of the clocked comparator model (Fig. 7). Here, the comparator is strobed at $t = t_0$ and t_1 .

Fig. 9. An illustration on how $v_s[n]$ is derived every time a clocked comparator is strobed. $v_s[n]$ is the average of the $v_i(t)$ with $\Gamma(\tau, n)$ as the weighting factor.

Fig. 10. The proposed clocked comparator input frequency response characterization method. Here, v_i and v_{th} is fed into the comparator’s + and – input, respectively, and v_{th} is swept from a low value. Coherent equivalent-time sampling can be used to sweep the comparator’s strobing position very finely.

coherent equivalent-time sampling can be used. A fractional-N PLL (Phase Locked Loop) synchronized to the RF signal generator can be used to generate a strobing signal ϕ with a period T_ϕ that is very close to the period T_i of v_i which results in a coherent sampling with a very fine time step.

V. EXPERIMENT

Here, we will demonstrate the implementation of an accurate clocked-comparator-based Δt measurement. The setup of the experiment is shown in Fig. 11 and the Δt measured is the delay across a 25 cm, 50 cm, and 48 inch RG-316 cable.

A. Experiment Setup

In the experiment, a clocked-comparator-based Δt measurement circuit is built from the demo boards of low-cost COTS ICs (Integrated Circuits) and a logic analyzer (for data acquisition purpose). The demo boards are the demo boards of: 1) Hittite clocked comparator IC HMC874, 2) Hittite fractional-N PLL IC HMC832, 3) Analog DAC IC AD5754. The considerations on the choice are summarized in Table I and are discussed in more details as this section progresses.

A pulse generator is used to launch a repetitive pulse into the cable being measured. Here, the pulse repetition rate is upper bounded since reflection points exist in the path connecting the pulse generator and the comparators, i.e., at the v_i input of the comparators (since $R_{in,a}$ and $R_{in,b} \neq Z_0$) and at the T-junction that splits the path of the pulse to the two comparators. To allow for reflections to die down first before a new pulse is launched into the path, a 50 MHz and 20 MHz repetition rate is used in the 25 & 50 cm and 48 inch cable delay measurement, respectively.

The Δt measurement resolution is set to ~ 1 ps by means of coherent equivalent-time sampling. To do this, given a pulse repetition rate of, e.g., 50 MHz, the frequency of the clock ϕ of the comparators can be set to ~ 49.9975 MHz. This can be done using the chosen PLL IC with the following setup: $\frac{N+n/2^{24}}{D} = \frac{43+16.7 \frac{4080}{9}}{44} \frac{9}{2^{24}}$ given a phase detector frequency of 50 MHz where N , n , and D is the value of the integer and the fractional part of the feedback divider and the output divider of the PLL, respectively. Here, the 24-bit resolution of the fractional part of the feedback divider of the PLL is an overkill since achieving an exactly 1 ps resolution is not needed as long as the resolution of the measurement can be determined (e.g. using (3)).

B. Accuracy Considerations

1) DC-Offset Nullification and Skew Compensation

Before the measurements are done, the value of $v_{th,a}$ and $v_{th,b}$ are adjusted so that their effective values are almost at their intended values. A constant voltage generated by a third DAC (not shown in Fig. 11 (a)) whose value is equal to the 50% level of the pulse launched by the pulse generator is fed into the v_i input of the comparators. $v_{th,a}$ and $v_{th,b}$ are then adjusted so that each of the comparator outputs approximately the same number of “1”s and “0”s after a large number of stroblings. The amplitude and the rise time of the pulse launched by the pulse generator is ~ 1 V and ~ 500 ps, respectively. Hence, the slope of the rising edge of the pulse is ~ 1.6 V/ns. To keep δ_{off} to below 0.5 ps (half the measurement resolution), the voltage resolution of the DAC should be < 0.8 mV. The voltage resolution of the DAC IC is ~ 0.1526 mV and, hence, sufficient. Note that what matters here is the voltage resolution of the DAC, not the resolution of the DAC in terms of its digital input word width.

After the δ_{off} nullification, the δ_{skew} of the circuit is measured. This is done by connecting the v_i input of the comparators to the pulse generator using a pair of cables with matched length.

Fig. 11. (a) The block diagram of and (b) a photograph on the experiment setup.

TABLE I
COMPONENTS CHOICE CONSIDERATIONS

Part	Param.	Consideration	Value	Remark
Comp. HMC 874	Sampling bandwidth	Accuracy <i>or</i> Width of pulse wave	~ 7 GHz	Allows for < 100 ps accurate walk-error compensated <i>or</i> sub-ns-wide-pulse ToF measurement
PLL HMC 832	Divider resolution	Meas. resolution	N: 19-bit n: 24-bit D: 1-62	Lower resolution is actually required since the meas. resolution does not need to be as close to 1 ps
DAC AD 5754	Voltage resolution	Accuracy (DC-offset nullification)	~ 0.1526 mV	< 0.8 mV is required given a pulse edge slope of ~ 1.6 V/ns and a meas. resolution of ~ 1 ps

2) Sampling Bandwidth

Before the measurements, the input frequency response of the clocked comparator IC used is characterized using the method discussed in section IV. The result is shown in Fig. 12. With a sampling bandwidth of ~ 7 GHz, the rise time of the step response of the comparator can be expected to be < 50 ps [26]. If the sampling bandwidth can vary by, e.g., $\pm 40\%$ across the expected PVT range (a clocked comparator's sampling bandwidth is proportionally related to its input transistors transconductance [20, 29]), then the rise time of the step response of the comparator can change by around -14 ps to +33 ps. If the width of a pulse as measured by the comparator can vary due to PVT variation by about the same amount, the error in the amount of compensation applied in a pulse ToF measurement that utilizes the pulse-width-based walk error compensation proposed in [14] can be expected to be < 100 ps. Meanwhile, if interpolation is used to determine the onset of a

pulse for a walk-effect-tolerant pulse ToF measurement, the 7 GHz sampling bandwidth of the comparator can be considered to be very wide. However, the comparator will be able to deal with a pulse that has a very narrow pulse width such as the echoes of the sub-ns laser pulse used in [18]. Clocked comparator with a sampling bandwidth of >10 GHz can be fabricated under a standard CMOS technology [27, 28].

C. Experiment Results

The results of the experiment are shown in Table II. We perform the experiment 3×, redoing the δ_{off} nullification and δ_{skew} measurement procedure each time. The results are shown side by side with the results of the same delay measurement done using a 13 GHz, 40 GSa/s oscilloscope. Similar δ_{off} nullification and δ_{skew} measurement are done on and between the channels of the oscilloscope, respectively. The results of the measurements performed using the measurement circuit and the oscilloscope agree to within 10 ps from one another.

VI. CONCLUSIONS

We have discussed on how a time-interval can be measured using clocked comparator(s). To do this, basically, a clocked comparator is used to determine whether a given signal has crossed a certain threshold through a repeated comparison between the value of the signal and the threshold. Time-interval measurements with ps resolution can be achieved under the proposed approach through the use of time-interleaving and/or equivalent-time-sampling. The effect of DC-offset, skew, and sampling bandwidth to the accuracy of time-interval measurements done according to the proposed approach have been discussed. A method to characterize the sampling bandwidth of a clocked comparator has also been discussed.

The proposed time-interval measurement approach is applicable for the measurement of the ToF of wave pulses in a pulse ToF range-finding application. The approach is a compromise between the conventional TDC-based and ADC-based time-interval measurement approach typically used in the said application. As compared to the TDC-based approach, no extra channels are needed should multiple echoes are expected to be received for every wave pulse launched. As compared to the ADC-based approach, a simpler back-end digital processing can be used and the skews among the comparator channels do not need to be nullified. The skews can instead be measured and compensated from the measurement results. Lastly, the approach can leverage on the strength of clocked comparators which can be designed with a very wide sampling bandwidth.

We demonstrated the implementation of an accurate clocked-comparator-based time-interval measurement. A clocked-comparator-based time-interval measurement circuit with ~7 GHz bandwidth was built using low cost COTS components. The measurements done using the measurement circuit agree to within 10 ps with the same measurements done using a 13 GHz, 40 GSa/s oscilloscope and are precise to within ± 15 ps in terms of their 2σ range.

For our future works, we will be incorporating the proposed Δt measurement technique in the design of a low-cost RF rangefinder for cm-level accurate ranging applications.

Fig. 12. The result of the input frequency response characterization of the clocked comparator IC used in the delay measurement experiment.

TABLE II
THE RESULTS OF THE DELAY MEASUREMENT EXPERIMENT

Cable length	Exp. No.	According to the measurement circuit (300 single shots $\mu \pm 2\sigma$)	According to a 13 GHz, 40 GSa/s oscilloscope (10,000 single shots $\mu \pm 2\sigma$)
25cm	1	1189.99ps $\pm$ 7.74ps	1192.94ps $\pm$ 1.69ps
	2	1190.99ps $\pm$ 4.09ps	
	3	1183.59 $\pm$ 5.85ps	
50cm	1	2349.64ps $\pm$ 13.9ps	2351.83ps $\pm$ 1.69ps
	2	2352.46 $\pm$ 6.80ps	
	3	2343.80ps $\pm$ 3.41ps	
48"	1	5729.31ps $\pm$ 11.1ps	5733.26ps $\pm$ 1.79ps
	2	5729.46ps $\pm$ 11.2ps	
	3	5729.10ps $\pm$ 11.4ps	

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