TIMING DISCRIMINATOR FOR PULSED TIME-OF-FLIGHT LASER RANGEFINDING MEASUREMENTS

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ABSTRACT

A time-pickoff circuit based on constant fraction discriminator (CFD) timing principle has been developed for pulsed time-of-flight laser rangefinding. It is based on the detection of the crossing point of the trailing edge of the original timing pulse and the leading edge of its delayed replica with a fast ECL comparator. A simplified theory has been presented for the sources of the walk error in constant fraction time discriminators, and three different comparator types have been tested in the CFD developed. It has been noticed that if there are not any crosstalk disturbances, the walk error produced by the limited gain-bandwidth product of the comparator is the dominating walk error source and that walk error can be decreased to +/- 1 mm in a 1:10 dynamic range of input pulses by adding an external offset voltage between the input nodes of the comparator. The bandwidth of the preamplifier used was 100 kHz-100 MHz.

1. INTRODUCTION

The time-pickoff circuit is an essential component in TOF laser radars and more generally in time spectroscopy, which is used for example in nuclear physics measurements 1,2. The purpose of the time-pickoff circuit is to gain a logic output pulse that is precisely related in time to the occurrence of an event. The source of the event may be, for example, the detection of a laser light pulse with varying shape and amplitude.

The performance of the time-pickoff circuit is measured with three parameters: walk error, drift and timing jitter. The walk error, which is often the most important parameter, is the timing error in the time-pickoff circuit as a function of the shape and amplitude of the input pulses. Drift, again, is the long-term timing error as a function of component aging and temperature variations in the timing circuitry. Long-term
drift can often be calibrated so that it is sufficiently small. Timing jitter, which is expressed with the term resolution or precision, is caused by statistical fluctuation of the shape of the measurement pulses and noise in the measurement system. The effect of jitter can be reduced by averaging several successive measurement results.

In both pulsed laser rangefinding and nuclear physics the amplitude and shape of the measurement pulse may change in the measurement process. In nuclear physics, they may vary typically in a relatively large range from one pulse to another due to the detection process, and, for this reason the gain control cannot be used to cancel the amplitude variation. In pulsed time-of-flight laser rangefinding the variations in the received pulse amplitude are induced mostly due to the variation in the measurement distance and object reflection. The effect of these amplitude variations on timing can usually be cancelled by optical gain control with a negligible timing error. However, the optical gain control is not always fast enough to hold a stable optical peak power in the pulses. This happens, for example, when the measurement beam is scanned over a surface. In that situation it is necessary to have a time-pickoff circuit with low walk error, because the time-pickoff circuit can adapt even to two successive pulses, which have varying amplitudes. It is possible that the averaged shapes of the pulses may vary also in the laser rangefinders.

The aim of this work was to develop a timing discriminator for pulsed time-of-flight laser rangefinding measurements with a walk error less than +/-1 mm (+/-7 ps) in an input amplitude range of 1:5. The application environment is a laser radar described in chapter 2 and in more detail in ref. 3. It is a double-channel laser radar which is used in steel industry for measuring the thickness of the lining in a steel converter. Tests were made also with a single-channel laser radar, which is designed for measuring the shapes of ship blocks in shipyards in distances varying from 2 to 30 m 4. In order to decrease the walk error, it was studied how the speed of the timing comparator affects the walk error. Also a compensation method based on the adjustment of the offset voltage of the timing comparator was analyzed. The adjustment procedure should be easy and fast and applicable to all devices of the selected laser type.

2. CONSTRUCTION OF A PULSED TIME-OF-FLIGHT RADAR

The block diagram of a double-channel laser radar used in the application environment in this work is presented in figure 1. The measurement system consists of a transmitter, two receiver channels, two constant fraction discriminators and a time measuring unit. In addition, the system may include also an optical attenuator and a microprocessor unit which measures the power of the received signal and controls the operation of the optical attenuator. The start pulses are taken directly from the outgoing radiation from diffuse reflection. In a single-channel laser radar the whole receiver channel is common to start and stop pulses, which are separated after the time-pickoff circuit to their own channels with logic gates. The laser transmitter sends
optical pulses in 4 kHz frequency. The laser may be either a SH or DH type, the pulse power of which may vary in the range of 5 to 50 W. The FWHM (full-width-at-half-maximum) of the measurement pulses in the channel are in the range of 6 to 8 ns. The lower and upper -3 dB corner frequencies of the whole receiver channel are 100 kHz and 100 MHz, respectively.

The time-pickoff circuit converts the analog measurement pulses to ECL logic pulses which are fed to the TDC (time-to-digital converter). The TDC measures the time difference between the start and stop pulses. It consists of a 100 MHz oscillator and counters which roughly digitize the time interval to be measured. The time fractions between the start and stop pulses and their respective next clock pulse but one are digitized with an interpolation circuit based on analog time-to-amplitude conversion.

3. OPERATION OF A TIME-PICKOFF CIRCUIT

3.1. GENERAL CONSIDERATIONS OF A TIME-PICKOFF CIRCUIT

The time-pickoff circuits can be roughly divided into two groups, constant fraction discriminators and leading edge discriminators. The timing moment may be detected with a comparator, or a tunnel diode. In the leading edge discriminator, the timing moment is the crossing moment of the leading edge of the input pulse and a fixed threshold level. The timing moment is a function of the amplitude and rise time of the input signal. The result of the variation of the timing moment is that the walk error is large and the resolution does not remain at an optimum level.

In constant fraction timing, the timing moment is always located at some fractional point of the leading and trailing edges of the input pulse. Usually the input signal is divided into two parts. The other part of the signal is delayed and inverted. An undelayed, but attenuated signal is summed with it. The timing moment is the point where the rising edge of the non-attenuated pulse has an equal voltage level with the peak level of the attenuated pulse. Then the summed bipolar pulse crosses zero voltage level. The optimum fractional point and the value of the delay are selected so that the resolution value is minimized. The delay is usually implemented with a delay cable and the attenuation with a resistor divider. Pulse inverting is possible to realize with a pulse transformer or with a shorted delay cable. If a comparator is used in the detection of the timing moment, the delayed and undelayed pulses can be fed to the different input nodes, and subtraction of the pulses from each other is not needed.

There are three possible sources for walk error: first, variation of the crossing moment of the two signals in the input of the time-pickoff circuit, second, the limited gain-
bandwidth product of the comparator, and third, the possible crosstalk in the circuit. All these sources may increase walk error, but they may also cancel out the effect of each other.

The crossing moment may vary due to the timing principle itself or because there is an offset voltage between the two input pulses. The offset voltage may introduce walk error, but it may be also used for cancelling the walk error, which is caused by the limited speed of the comparator. The effect of the offset voltage can be seen from the figure 2, where the input pulse of the time-pickoff circuit is divided into two pulses and the other pulse is delayed. The timing moment is the crossing point of the two pulses which are fed to different input nodes of the comparator. From fig. 2 it can be seen that if a positive offset voltage is added to the delayed pulse and if the amplitude of the input pulse is decreased, the timing moment shifts to an earlier position.

The propagation delay of the time-pickoff circuit changes, when the slew rates of the pulses change, because the comparator has a limited gain-bandwidth product. The higher the slew rate of the input pulse, the smaller is the propagation delay of the comparator. If the amplitudes of the input pulses are changed, but the slew rates of the input pulses are kept constant by changing the rise time, the propagation delay remains unchanged, provided that the amplitudes of the input pulses exceed some minimum level.

The third source for the walk error may be some crosstalk between the input pulses and some other pulses in the time-pickoff circuit. In some cases also the disturbance caused by the crosstalk may cancel the effect of some other walk error source, but then the adjustment procedure of the walk error may be very time-consuming and unpredictable.

The timing jitter $\sigma_t$ of the measurement result depends on the slew rate and noise of the input pulse:

$$\sigma_t^2 = \frac{\sigma_n^2}{(dV/dt)_{V=D}^2}$$

where $\sigma_n$ is the r.m.s. noise of the input of the time-pickoff circuit and $(dV/dt)_{V=D}$ is the slew rate of the signal at the timing moment. Eq. 1 is derived for timing, where the input pulse crosses a reference level. From (1) it can be noted that the timing moment should be chosen as a point in which the ratio between noise and slew rate is at minimum. If it is assumed that the distance is measured $n$ times and the noise in the measurement process is randomly distributed, the resolution is improved by factor $\sqrt{n}$ compared to a single measurement result.

There are many sources of noise in TOF laser radars: the signal-induced noise, the excess noise from an avalanche photodiode and amplifiers and the noise from the current caused by background radiation. According to measurements in ref. 3, when the temperature of the target is near room temperature or lower, the effect of the background noise can be neglected. At small signal levels, both the noise from the
electronics and the signal-induced noise must be taken into account, but at high signal levels the signal-induced noise is dominating, if an avalanche photodiode is used as a photodetector. The higher the timing moment is chosen in the leading or trailing edge, the stronger the signal-induced noise is at the timing moment. At the top of the laser output pulse the slew rate is small and, for example, the relaxation oscillations may cause instability in the shape of the pulse. For these reasons the timing moment should not be chosen near the peak of the pulse. The lower the timing moment is at the linear part of the leading or the trailing edges, the better the resolution, because the signal-induced noise is there at its lowest, assuming that the timing moment is chosen from the linear parts of the leading and trailing edges.

3.2. CONSTRUCTION OF THE DEVELOPED CFD

The main principle of the time-pickoff circuit used here is the following: the amplified input pulse is divided into two parts which have equal amplitudes and the other pulse is delayed. The delayed and non-delayed signals are fed to the different input nodes of a fast comparator with ECL logic outputs. The value of the delay is selected so that the two pulses cross each other at the trailing edge of the first pulse and at the leading edge of the latter pulse. The pulse diagram is presented in figure 2. With this method, the advantage is that the resolution is small, because both pulses to the inputs have high slew rates at the timing point. The timing moment is usually selected to be at 30% - 40% from the maximum amplitude, because, according to the measurements, the resolution is smallest at that level.

A simplified schematic diagram of the CFD developed is presented in figure 3. The threshold level of the noise comparator is adjusted to exceed the peak level of the background noise, for example to a level of 100 mV to 150 mV. The output pulse of the noise comparator goes through ECL-flip-flop n:o 1 (type 10H131) to the data input of the flip-flop N2. So the output of the flip-flop N2, which is at the same time the output of the discriminator, can change its state only when the amplitude of the input pulse of the discriminator exceeds the threshold level adjusted to the noise comparator. The output timing pulse from the timing comparator triggers the clock input of flip-flop N2.

Flip-flop N1 has two purposes: it extends the data input pulse of the flip-flop N2 and latches the output nodes of the noise comparator to a stable state until the output of the flip-flop N2 resets the state of flip-flop N1. The latter function is important, because the output nodes of the noise comparator may change at the same time as the input pulses to the timing comparator cross each other. Then the output pulse of the noise comparator can be coupled to the input nodes of the timing comparator through the parasitical capacitances between the bonding wires in a dual comparator circuit inducing timing error.

The presented configuration is a result of evolution. Originally the timing comparator was enabled by the noise comparator through its latch input only for the period of
timing. The idea was that in this case the offset of the timing comparator can be selected freely without oscillations which may be present in the version of fig. 3 if the noise level exceeds the input offset. However, this method needs an extra delay cable in order to adjust the timing for the latch input of the timing comparator. We found that the latching of the timing comparator is not needed, because the possible oscillations in the output of the timing comparator in the construction of fig. 3 are suppressed when the timing pulse exists at its input, and were not found to produce any measurable errors.

4. MEASUREMENTS

Three different comparator types were tested in a CFD circuit described in chapter 3. The types were Analog Devices AD 96687 and Signal Processing Technology HCMP 96870A and SPT 9689. The walk error of the distance meter was measured with several offset voltages with all three comparator types. All the comparators tested are pin-compatible with each other. The main difference between the comparators is in the -3 dB bandwidth of small signal open loop gain. However, not any exact values are given in the datasheets, but it seems that the SPT9689 is the fastest comparator and the AD96687 is the slowest one.

In all walk error measurements a double-channel laser radar was used, except for the measurement presented in figure 8, which was measured with a single-channel laser radar. In the single-channel radar only the comparator SPT9689 was tested. The semiconductor laser used in double-channel radar was a pigtailed DH laser diode CVD-193F (manufactured by Laser Diode Inc.), the pulse power of which was 18 W measured from the end of the fiber. The rise and fall times (between 10 % and 90 %) of the channel pulse (at the input of the CFD) were 4.5 ns and 4.7 ns, respectively, and the FWHM was 7.9 ns measured with an oscilloscope of 1 GHz analog bandwidth. The shape of the channel pulse is presented in figure 4. In a single-channel laser radar the laser type CVD-93F was used. It had a pulse power of 5 W, rise time of 3.3 ns, fall time of 4.3 ns and FWHM of 6.8 ns. The delay cable in the constant fraction discriminator was 190 cm long in the measurements presented in figures 5 to 12. With that cable the crossing point of the undelayed and delayed pulses was about at 41 % of the maximum height of the pulse in a double-channel radar and 28 % in the single-channel radar.

The walk error measurements with zero offset voltage are presented in figure 5. The external offset voltage compensated the internal offset voltage of the comparator. We found that the walk error was clearly largest with the AD 96687 comparator and smallest with the SPT 9689 comparator.
The walk error measurements with different offset voltages are presented in figures 6 to 9. For all comparators the walk error curves have been measured with several values of offset voltage. The smallest walk error, +/- 1 mm in the input amplitude range of 0.2 to 2.0 V, was achieved with SPT 9689 in a double-channel radar. The best measured walk error in the single-channel radar with SPT9689 was about +/- 1.5 mm in the input amplitude range of 0.2 to 2.0 V.

The effect of the timing level on the performance of the CFD was studied by measuring the single-shot resolution values vs. input pulse amplitudes with several delay cable lengths (figure 10). The comparator type was AD96687. The timing levels are indicated as relative levels of the crossing point of the input pulses of the timing comparator compared to the maximum amplitude of the pulses. The timing levels were measured with zero offset voltage. The results show that the resolution depends slightly on the timing level, as expected, and the higher the timing level, the worse the resolution. However, with input amplitudes smaller than 1.3 V the resolution was clearly the best with 41 % timing level and not with the lowest timing level. This is probably due to the fact that with low signal amplitudes, the noise of the amplifiers becomes the dominating noise source and the total noise levels at the timing points at 23 % and 41 % timing levels are somewhat equal. The lower slew rate value at 23 % level gives a worse resolution than at 41 % level at low signal amplitudes.

The effect of crosstalk between the output of the noise comparator and the input of the timing comparator may have an impact on the shape of the walk error curve. It was proved to be true by measuring the walk error curves so that the latch inputs of the noise comparator were not in use. The walk error curves measured with and without using the latch inputs, with different noise comparator threshold values, are presented in figures 11 and 12, respectively. In both measurements the comparator type was AD96687. All curves in the figures are measured with the same delay cable length and the same timing comparator offset voltage. It can be seen that the crosstalk between bonding wires inside the double comparator circuit has a rather strong effect on the shape of the walk error curves and for this reason it is necessary to prevent the change of state of the output nodes of the noise comparator during the timing moment at the inputs of the timing comparator.
5. EVALUATION OF THE WALK ERROR VALUE

In the evaluation of the value of the walk error, two things must be taken into account: the dependence of the propagation delay on the slew rate of the input pulses and the offset voltage in the input nodes of the comparator.

A simplified model of the propagation delay of a comparator has been introduced in Eq. (2). The propagation delay is created by the internal capacitances of the comparator. In the following, we make a simple assumption that the comparator is an amplifier which has a single high-frequency pole and the shape of the input pulses remains the same at all input signal amplitudes. The amplification is linear, but we take into account the effect of operating in a large signal range by including in the calculation the minimum propagation delay \( \tau_D \). It is assumed that the maximum slew rate of the comparator is not exceeded. The open loop gain in the s-domain is:

\[
\frac{U_o(s)}{U_i(s)} = F(s) = \frac{e^{-\tau_D} \cdot A(0)}{1 + sA(0)\tau_0}
\]

where \( A(0) \) is the open loop gain in zero frequency and \( \tau_0 = 1/\omega_0 \) the time constant corresponding to the upper -3 dB corner frequency of the unit gain. The terms \( U_i(s) \) and \( U_o(s) \) are the input and output voltages of the amplifier, respectively. Assuming that the input signal is a linear ramp \( u_i(t) = SR \cdot t \) and taking the inverse Laplace transform from \( U_o(s) \) we can calculate the response time \( t_{delay} \) required to reach some output level \( V_o \):

\[
t_{delay} = \sqrt{\frac{2 \cdot V_o \tau_0}{SR}} + \tau_D
\]

Eq. (3) consists of the minimum propagation delay \( \tau_D \) and another delay, the value of which depends on the slew rate of the input signal. The propagation delay has the simplified form in small signal analysis:

\[
t_{delay} = A + \frac{B}{\sqrt{\text{slew rate}}},
\]

where \( A \) ja \( B \) are constants. The smaller is the constant \( \sqrt{2 \cdot V_o \tau_0} \), the smaller the dispersion of the propagation delay (walk error) of the comparator. In practice it means that the comparator should be as fast as possible.

The effect of the offset voltage summed to the input nodes of the comparator can be evaluated coarsely using input pulses which have linear leading and trailing edges (figure 13). Now we derive the crossing point of the pulses from a pair of equations which define the straight lines of the leading and trailing edges of the pulse:
\[ u - U_{\text{of}} = \frac{U_p}{t_r} \cdot t \]

\[ u - U_p = \frac{-U_p}{t_f} \cdot t \]

where \( U_{\text{of}} \) = offset voltage between the input nodes, \( U_p \) = the amplitude of the input pulses in the comparator nodes, \( t_r \) = rise time (0-100 \%) of the pulse and \( t_f \) = fall time (100\%-0) of the pulse. We can calculate the crossing point \( t_{\text{cross}} \) by solving the pair of equations in the crossing point:

\[ t_{\text{cross}} = \frac{1 - \frac{U_{\text{of}}}{U_p}}{\frac{1}{t_r} + \frac{1}{t_f}} \]  

(6)

If \( U_{\text{of}}, t_r, \text{ja} t_f \) are constants, we come to a simplified solution for the crossing point:

\[ t_{\text{cross}} = C \left( 1 - \frac{D}{\text{slew rate}} \right) \]  

(7)

where \( C \) and \( D \) are constants and the term slew rate equals to the sum of the slew rates of the leading and trailing edges: \( \text{slew rate} = \frac{U_p}{t_r} + \frac{U_p}{t_f} \).

In order to calculate the final walk error curves, we may select the values \( V_0 \) and \( \tau_0 \) in eq. 3 so that the shape of the calculated propagation delay curve is close to the shape of the measured propagation delay curve of AD96687 presented in figure 5. For example, if we choose the value of 0.5 V for \( V_0 \) and assume that the AD96687 has a gain of 170 at the frequency of 100 MHz, the calculated walk error curve comes very close to the measured one. If we sum up equations 3 and 6 and use the measured rise and fall times \( t_r \) and \( t_f \), we can calculate the walk error curves with different offset voltage values. The result of such calculations for AD96687 is presented in figure 14. In equation 6 the rise time \( t_r \) (0-100 \%) used was 4.9 ns and the fall time \( t_f \) (100\%-0) used was 5.6 ns, and in equation 3 the rise and fall times used were 5.6 ns both.

Comparing the calculated walk error curves of figure 14 and the measured walk error curves of figure 6 with each other, it seems that they are reasonably close to each other and it can be concluded that the forming of the walk error curve is based on the factors described above. The value of the walk error can be decreased in a limited amplitude range by adding an external positive voltage between the input nodes of the timing comparator, as the calculations show. The differences between calculated and measured curves are probably due to the facts that the comparator probably has more than one high-frequency poles and that the pulse edges in equation 5 have been approximated to be linear, which in reality is not, of course, quite true.
6. DISCUSSION

To achieve a small walk error, the bandwidth of the timing comparator should be increased. In this way only a small offset voltage is needed to straighten the walk error curve and the result is a decreased walk error value especially at low amplitudes. From eq. 5 it can also be seen that if the rise and fall times of the input pulses are decreased, the walk error decreases.

The new semiconductor technologies make it possible to fabricate faster amplifiers and comparators. There are already commercially available transimpedance and voltage amplifiers with bandwidths ranging from 0.5 to 1 GHz. However, with high bandwidths, the importance of the layout increases and all stray capacitances must be minimized in order to prevent other parts of the circuit from disturbing the input pulses of the timing comparator. The crosstalk could also be minimized by integrating the CFD with the amplifiers on the same semiconductor chip.

ACKNOWLEDGMENTS

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Figure 1. A simplified diagram of a laser radar.

Figure 2. The formation of timing point, when an offset voltage has been added between the input nodes of the timing comparator.
Figure 3. The schematic diagram of a constant fraction time discriminator used in this work.

Figure 4. The delayed and non-delayed pulses measured from the input pins of the comparator AD96687. The offset voltage between the input pins was 36 mV and the timing level was 41%.
Figure 5. The walk error of the CFD for three comparator types with zero external offset voltage and 41 % timing level.

Figure 6. The measured walk error curves of the CFD with AD96687 and different offset voltages. The timing level was 41 %.
Figure 7. The measured walk error curves of the CFD with HCMP96870A and different offset voltages. The timing level was 41%.

Figure 8. The measured walk error curves of the CFD with SPT 9689 and different offset voltages. The timing level was 41%.
Figure 9. The measured walk error curves of SPT 9689 with different offset voltages in single-channel laser radar. The timing level was 41%.

Figure 10. The single-shot resolution of the laser radar as a function of the amplitude of the channel pulse measured with four different timing levels. The comparator type was AD96687 and the offset voltage was 33 mV.
Figure 11. The walk error curves measured with different noise comparator threshold values and using the latch inputs of the noise comparator. The timing level was 41% and the offset voltage of the timing comparator was 33 mV.

Figure 12. The walk error curves measured with different noise comparator threshold values without using the latch inputs of the noise comparator. The timing level was 41% and the offset voltage of the timing comparator was 33 mV.
Figure 13. The simplified input pulses of a CFD.

Figure 14. The walk error curves with different offset voltages for AD96687, which are based on adding the calculated propagation delay and calculated offset correction curves.