Linearity and Noise in Differential Travel Time Acoustic Velocity Measurement

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Abstract - **An improved constant current source for the time interval integrator in the differential travel time velocity sensor, BASS (Benthic Acoustic Stress Sensor [1]), has extended the full scale range of velocity over which the measurement of velocity is linear from 100 cm/s to 500 cm/s[1]. Slow speed response in the former constant current source introduced non-linearities of 5% to 8% for full scale ranges of 240 cm/s. When corrected, linearity is restored to 1.5% over the entire range. Electronic noise (standard deviation) is less than 0.5% of reading with a minimum noise of 0.3 mm/s equivalent velocity for a 15 cm path length.**

I. INTRODUCTION

Measurement of fluid flow by acoustic travel time difference is a technique that has been elaborated upon for almost a century. In the last twenty years, simultaneous pulses directed oppositely along a single path have been used. The difference in arrival times of the two pulses is measured to determine the fluid velocity component along the path. Velocity varies linearly with differential delay of the pulses to an excellent approximation for oceanic flows. Short time differences must be measured accurately and without offset at zero time difference. In BASS (Benthic Acoustic Stress Sensor), this time difference is determined by differencing a pair of current integrators, each of which is turned on by the arrival of the acoustic pulse, but which are both turned off together. A constant current source is steered from a sink to an integrating capacitor and back to a sink to affect this integrating interval. Linearity of the measurement thus depends upon the constancy of this current for various intervals.

It was discovered in BASS that as the full scale range was increased from 120 cm/s to 240 cm/s and higher for wave measurements, deviations from linearity of 5%-8% were observed. Tests were performed with coaxial delay lines to generate pulse delays electronically. Some of the apparent non-linearity resulted from reflections in the coaxial delay lines but when careful impedance matching of the coaxial lines and the switches removed reflections, non-linearity remained. The cause of this non-linearity was a slow response of the constant current source to voltage changes required for steering the current from sink to capacitor and back to sink. A circuit modification corrected the problem. The search for the cause and the solution for this non-linearity revealed important characteristics of the differential time delay measurement.

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II. DIFFERENTIAL TIME INTERVAL MEASUREMENTS

Pulses of sound travel through fluid at the group velocity with respect to the fluid. If the fluid is moving, the pulse traveling downstream arrives before the pulse traveling upstream. The difference in travel time is closely approximated by

$$\delta t=2dv/c^2$$

where δt is the travel time difference, d is the pathlength, v is the component of fluid velocity along the path, and c is the speed of sound in the fluid [2]. The absolute travel time is approximately 100μs which is typically 1000 times as large as the differential travel time for oceanic velocities. (The absolute travel time does give a valuable measure of heat content through the dependence of c on temperature[3].) Measuring travel time differences not only reduces the effect of timing noise on velocity uncertainty by at least 1000, it removes a large noise contribution to the measurement of small velocities where the differential travel time approaches zero but the absolute travel time remains large. The differential travel time however varies over a time range corresponding to a velocity range from 0.3 mm/sec to 300 cm/s typically (the least resolvable velocity change to full scale) or four orders of magnitude. Even though offsets are not affected by non-linearities in this measurement, the accurate determination of mid- to upper-range values of velocity are sensitive to such errors.

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III. DUAL INTEGRATORS

Bursts of fifteen cycles of 1.75 mHz sound are transmitted to permit the amplitude of the received signal to build in the narrow bandwidth transducers. Zero crossings of the received signals are counted and the fourteenth negative going zero crossing is used to define the pulse arrival time. In this way, sensitivity to amplitude change is minimized and voltage noise (Johnson noise) translates to minimum time noise. A flip-flop changes state upon this event, one for each receiver. The transition from high to low of the flip-flop associated with signal 1 causes the first integrator to start. The input label "Start integrate 1" in Fig. 1 is driven by the flip-flop. Constant current source 1, indicated by label "CCS 1" flows through Q3 to a ground sink before the flip-flop changes state but when "Start integrate 1" switches, Q5 is turned off, turning Q3 off, and the emitter coupling of Q5 to Q6 causes Q6 to turn on, which turns on Q4. This steers the constant current from a ground sink to the integrating capacitor C19.

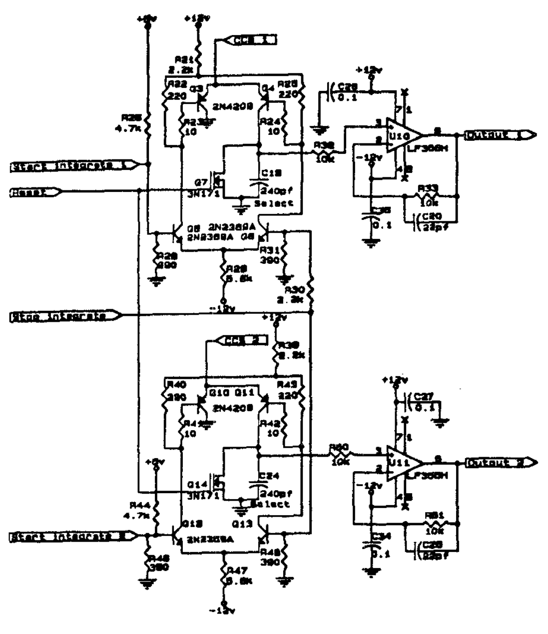

Fig. 1 Differential time to voltage circuit. Dual integrator.

Between measurements, C19 and C24, the integrating capacitor of the second integrator, are discharged by Q7 and Q14 respectively while the input "Reset" is high. "Reset" goes low before the acoustic pulse is expected. During the time that the constant current charges C19, the voltage tracked by FET input voltage follower U10 rises linearly with time. The input capacitance of U10 is in parallel with C19 and the sink capacitance of Q7 is also in parallel with C19. Neither capacitance shows detectable change with voltage as C19 charges from 0v to 5v. Nor does the β of Q4 change enough over the variation in V_{ce} from 8v to 3v as C19 charges to introduce non-linearities as large as 1 part in 1000. Consequently, the output labeled "Output 1" should be linear with time to 0.1% from 0v to 5v.

A similar integrator is turned on by the detection of the second acoustic pulse, which could be before, after, or at the same time as the first acoustic pulse. The change of state of the flip-flop controlled by the second signal starts integrator 2 by turning off Q12 and Q10, turning on Q13 and Q11, and steering a second constant current source, "CCS 2", from a ground sink to the integrating capacitor C24. This voltage is followed by U11 to provide the voltage level of the integrator at "Output 2".

A coincidence circuit detects when both flip-flops have made transitions and turns off both integrators by a low at "Stop integrate". This turns off Q6 and Q4 which turns on Q5 and Q3 by emitter coupling in the first integrator. The constant current is steered from the integrating capacitor, C19, to a ground sink and the voltage on C19 stops rising. The same thing happens at the same time at the second integrator. Q13 and Q11 turn off, Q12 and Q10 turn on, and the voltage on C24 stops rising.

The turn on and turn off transitions of Q4 and Q11 in the two integrators first add and then subtract small charges from the integrating capacitors through their Miller capacitances. The voltage swing of the bases of Q4 and Q11 are the same when the integrators have charged to 5v as when they have charged very little so the amount of charge added or subtracted only varies as the Miller capacitance varies with V_{bc}. This is only a small effect, of order 1 part in 1000. Consequently, the integrators are linear in time once the addition of charge at the turn on point and the subtraction of charge at the turn off point are neglected. Care is taken to ensure both integrators turn on fully before the coincidence circuit turns them both off. This ensures that the switching effects are cancelled to first order by the difference amplifier that follows "Output 1" and "Output 2".

IV. LINEARITY MEASUREMENTS

Linearity and scaling of the entire receiver and differential time to voltage circuit is measured for each BASS whenever it is reconfigured for a new velocity range. Integrating capacitors C19 and C24 are replaced by greater values for higher velocity range. This calibration is done with precision nanosecond delay lines.

A voltage signal that simulates the acoustic burst is generated by the calibrator at the time the BASS receiver expects it, typically 100μs after BASS transmits. This signal is split, one part fed to one input of the receiver while the other part is delayed by a switch selected set of coaxial delay lines before being fed to the other input of the receiver. Care was taken to match switch impedance to coax impedance since reflections in the calibrator introduce phase shifts, an error in the generation of precise delays. The calibrator has a range of 450ns and was determined to have an error of about 0.7ns over this range with a model 4195A HP Network/Spectrum Analyzer.

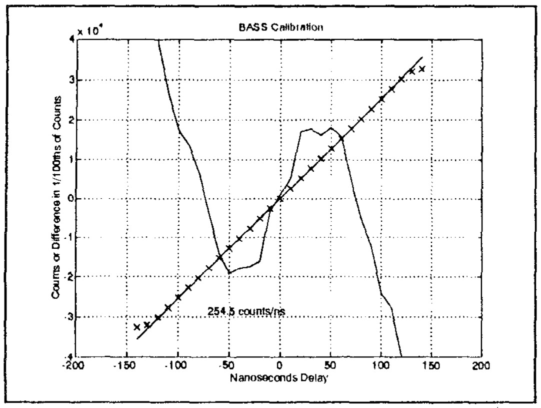

Fig. 2 BASS calibration with the slow constant current source. The straight line is a linear fit, Xs are measured observations, and the curved line is the difference between the observations and the linear fit magnified by 100.

In 1992, BASS moved from deep water onto the shelf, and the velocities it experienced went from 120 cm/s to 240 cm/s and higher. Fig. 2 shows a calibration to 140ns (105 cm/s) in which the deviation from linearity exceeds 3%. Actual measurements are indicated by X, a linear fit over the central 75% of the range is plotted as a straight line, and the difference of the observations from the fit is plotted with a magnification of 100. The error near the ends of the range is large but clipping diodes, that protect the A/D converter from large overvoltages, are on the verge of conducting and significant deviation from linearity at the very end is expected. However, a relatively large error, 3% of reading, is present over the central third of the range and an opposite error is present over the outer third of the range. Careful measurements made when the integrating capacitor was only 120 pf instead of 240 pf showed non-linearities of less than 1%. This range effect on error was puzzling but the cause was found in the constant current source.

V. CONSTANT CURRENT SOURCE

The current switched from sink to integrating capacitor and back to sink, a value about 4.5 ma, is generated by a constant current source. A low temperature coefficient voltage reference diode, U13, is compared to a current sense resistor, R19, as shown in fig. 3A. A portion of this voltage drop from potentiometer R17 is used for this comparison so that adjustment of scale is possible. An op amp difference amplifier, U8, drives a Darlington pair of current regulating transistors, Q1 and Q2. High frequency noise is reduced by filtering the output of the constant current source with a capacitor, C17. Except for the constant current itself, everything is low power, and consequently slow.

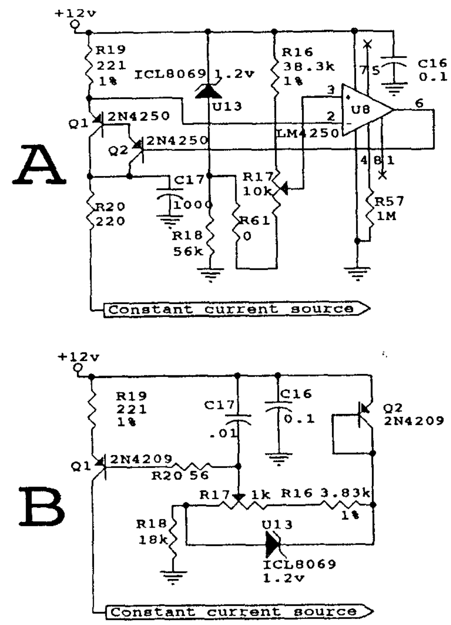

Fig. 3 Constant current sources. 3A, the original constant current source, is precise but slow. 3B, the new constant current source, is fast.

The source is not constant, however, while it is being switched from the sink to the integrating capacitor. The emitters of Q3 and Q4 in fig. 1 rise in voltage by 0.3v and the current decreases instantaneously because of the change in voltage across R20 in fig. 3A. The current change is not immediately sensed at R19 because the change is accomodated by the charge stored in C17. Not a great deal of harm is done by this except that in time, this change is sensed and a corrective action is taken. 50ns later, the current begins to correct itself and is still correcting when the end of the integrating interval is reached. A second error is introduced when the current is switched back to the sink but this does not affect the measurement.

Dynamic response of the original constant current source was sacrificed for stability and precision but this created an unacceptable error in linearity. Because linearity in this circuit is voltage vs. time, poor frequency response from slow slewing of the op amp and limited bandwidth of the Darlington pair causes an error in voltage. The remedy is shown in fig. 3B. A single fast transistor, Q1, is controlled by a portion of the voltage across a low temperature coefficient voltage reference. The temperature coefficient of V_{be} on transistor, Q1, is matched by a similar transistor, Q2, in the reference voltage path. The circuit is simpler as well as fast. It reduces the non-linearity of differential time to voltage measurements below 1% as shown in fig. 4.

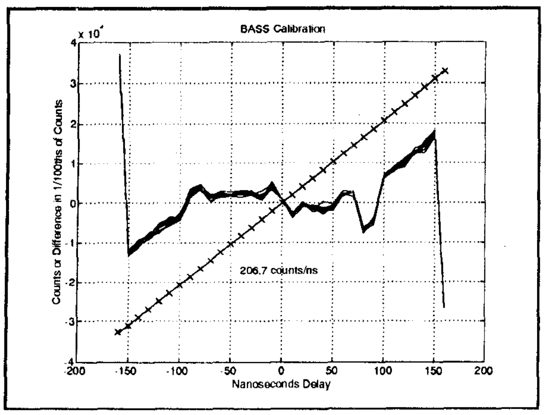

Fig. 4 BASS calibration with fast constant current source. The maximum difference between observations and their linear fit is 1/3%. Ten observations are plotted per delay and this appears to thicken the curve of differences.

The measurements of fig. 4 were taken by setting the nanosecond delay calibrator to 0ns and noting the output of BASS in counts. Increments of delay were added until the BASS output went to zero. Then 10ns delay was added and the output recorded ten times. Another 10ns was added and another ten observations were recorded. When 160ns were switched in and the A/D was saturating and giving full scale readings, the connections of the nanosecond delay to BASS were reversed and BASS counts were recorded for delays from 0ns to 160ns by 10ns intervals. Finally, these data were plotted, a line was fit to the central 70% of the data, and the differences of the observations from the linear fit were plotted magnified by 100. The ten observations per delay make the difference line appear thick. The difference between the observations and the linear fit is at worst 1/3%.

VI. CONCLUSION

Acoustic measurements of current by differential travel time provide high sensitivity and low noise but require care to obtain linearity over high current ranges. The constant current source must respond with the same bandwidth as the current steering circuits. Failure to consider dynamic response of the current source can introduce non-linearity in the differential time measurement. A simple fast transistor constant current source reduces this effect to an acceptable level.

VII. ACKNOWLEDGMENTS

Richard Koehler and Jia Qin Zhang designed and built the nanosecond delay calibrator that made this discovery possible. This is WHOI Contribution No. 8866.

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