The Integrator Technique of Time Delay Measurement for Ultra Wideband Communications Systems

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Abstract

The integrator technique of time delay measurement is revisited and its suitability to many aspects of ultra wideband (UWB) communications systems is indicated. In the paper basic properties of the integrator technique, measurement practicalities and sensitivity issues are reviewed.

1. Introduction

Ultra wideband (UWB) communications systems currently are a major area of research, leading to many emerging applications. Recent comprehensive reviews can be found in [1-3]. In UWB systems, narrow pulses on nanosecond time scales are employed. These have to be detected prior to any signal processing that may be employed for information extraction. Pulse detection requires techniques capable of working on these time scales.

In this paper we revisit and review integrator techniques of time delay measurement, which have been developed over many years [4-19], which can achieve picosecond time resolution, are simple to implement and are ideally suited to many areas of UWB applications. Areas of special significance would be position location and modulation techniques such as pulse position modulation and pulse width modulation. The integrator technique of time delay measurement has already been applied to position location [14, 17], jitter measurement in digital repeaters [8], microwave and optical propagating path characterization [10, 11, 12, 15, 19], high speed transistor measurement [5, 7], fast photo detector characterisation [16].

For large signal-to-noise ratio (SNR) the resolution is set by component limitations, but picosecond accuracy can readily be achieved with available operational amplifiers on a single pulse basis [7] and much more accurately with averaging. On the other hand, for low SNR the integrator as a time delay estimator is near optimum [18].

2. Basic properties

The integrator technique of time delay measurement basically consists of taking the difference between normalised waveforms of the “unit-step” type and integrating (Fig 1). A “unit-step” type of waveform is one where the initial value is zero and the steady-state value is finite and non-zero (Fig. 1(b)). Otherwise, the actual shape of the waveform is immaterial. The measurement technique can also be applied to waveforms of the “delta-function” type (Fig. 1(a)), by converting them to “unit-step” pulses through an additional stage of integration, but thereafter the same signal processing would follow. Since the actual wave shape is immaterial, very narrow pulses can be employed.

The time delay t_d that is measured in this manner is formally given by [4, 6]

$$t_d = \int_{-\infty}^{\infty} \left(\frac{v_1(t)}{v_1(\infty)} - \frac{v_2(t)}{v_2(\infty)} \right) dt \qquad \ldots(1)$$

If the two waveforms in Fig. 1 were the input and output of a two port network, the time delay t_d so measured would in fact be the first moment of the impulse response of the network and can be readily evaluated from the two-port transfer function. The time delay t_d can be simply measured with an operational amplifier differencing integrator [5, 7].

Extension of the delay time integrator technique to measurements of the second time moments has given results for pulse width widening of “delta-function” type of waveforms, or equivalently rise time changes in the corresponding “unit-step” type waveforms after integrating the “delta-function” type waveforms [9]. Extension of the technique to time moments higher than the second has been achieved [13].

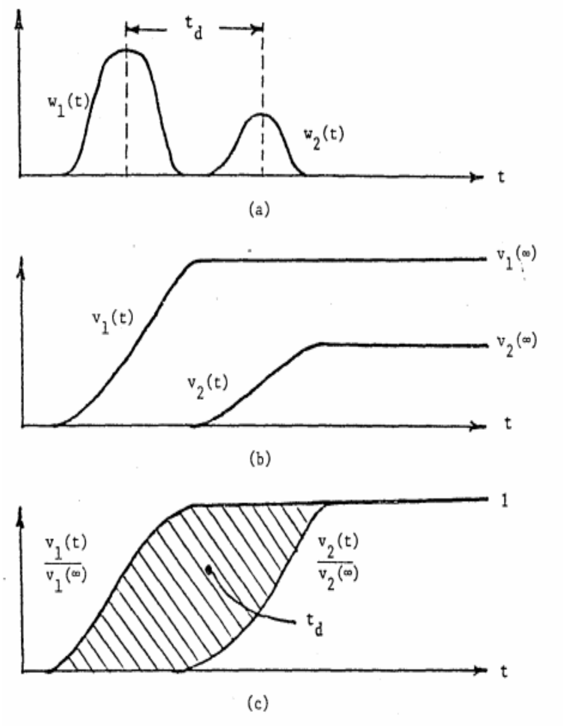


Fig. 1 Illustrating the integrator as a time delay estimator
(a) Waveforms of the "delta-function" type
(b) Waveforms of the "unit-step" type
(c) Time delay t_d as an area

The technique has also been extended to bandpass waveforms with a carrier [10-12]. Here we have an additional variant in that there are now two waveform envelopes, the in-phase and quadrature envelopes, p(t) and q(t) respectively, which can be detected with a synchronous demodulator. After detection both p(t) and q(t) are functions of the phase of the local oscillator (l.o.) with respect to the phase of the r.f. carrier. It is advantageous to adjust the l.o. phase so that $p(\infty)$ is maximum (and simultaneously $q(\infty) = 0$). The p(t), q(t) so obtained will be denoted by $\hat{p}(t)$ and $\hat{q}(t)$. The formula for the time delay t_d now becomes the group delay t_g of the propagating path, given by

$$t_g = \int_{-\infty}^{\infty} \left(\frac{\hat{p}_1(t)}{\hat{p}_1(\infty)} - \frac{\hat{p}_2(t)}{\hat{p}_2(\infty)} \right) dt \qquad \ldots(2)$$

where $\hat{p}_1(\infty)$, $\hat{p}_2(\infty)$ are the steady state values of $\hat{p}_1(t)$, $\hat{p}_2(t)$. Other formulae involving q(t) give results for attenuation dispersion and group delay dispersion.

There are advantages in using bandpass signals in the practical implementation of the integration measurement, as it solves the problems that arise in practice when normalisation required in eqn. 1 is carried out, as explained in the next Section.

3. Measurement details

A simple delay measuring circuit is shown Fig. 2. Basically it consists of a summing integrator with $v_1(t)$ and $v_2(t)$ as its inputs and one of $v_1(t)$, $v_2(t)$ phase inverted. In practice $v_1(\infty)$ and $v_2(\infty)$ are not equal and to achieve the normalisation in eqn. 1 some means of adjusting the amplitude of either $v_1(\infty)$ or $v_2(\infty)$ is required. If the steady state values of the two waveforms are not equalised, the output of the integrator does not reach a steady state value. One way of equalising the two amplitudes is to include a variable gain element in one of the inputs to the differentiating integrator. In Fig. 2 it is $v_2(t)$ that is adjusted by means of a variable gain stage.

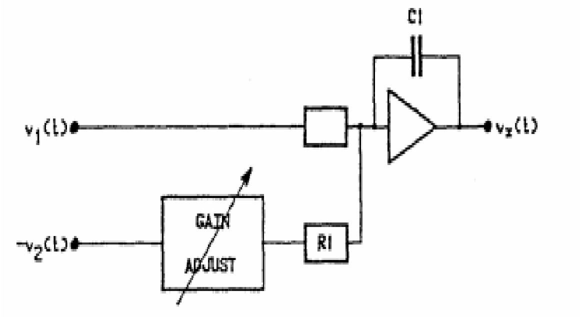
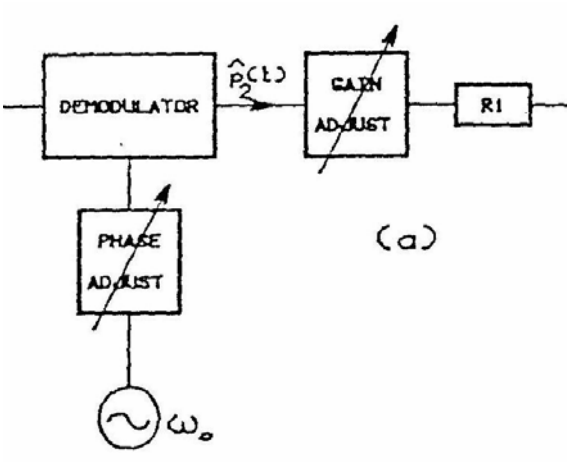


Fig. 2 Basic circuit for measuring time delay

The variable gain stage of Fig. 2 unfortunately introduces a time delay of its own, which is not easy to compensate for when picosecond time scale path differences have to be considered and the spurious time delay is a function of the gain setting. A much better way is to use carrier phase adjustment to vary the envelope amplitude of one of the waveforms [15, 17]

Figure 3(a) shows both phase and gain adjustment when synchronous demodulation is used, where it is assumed that the maximum value of $\hat{p}_2(t)$, i.e. $\hat{p}_2(\infty)$, is too large and must be reduced. Since the l.o. phase adjustment will also vary the steady-state value $p_2(t)$, it will by itself equalise the two inputs to the differentiating integrator for integration purposes (Fig. 3(b)), but now the r.h.s. of eqn. 2 will not be correct. This will introduce some error in the measurement, but this error has been shown to be negligible in many circumstances and in any case the error can be compensated [15, 17]. This phase-for-gain technique completely solves the gain adjustment problem.



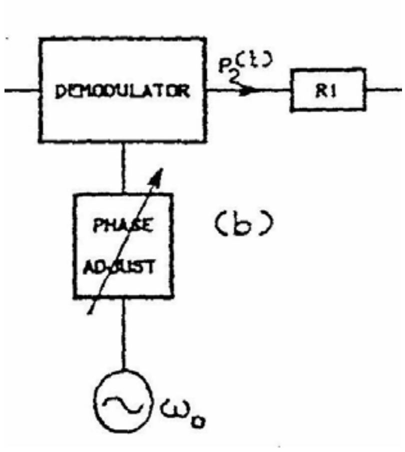


Fig. 3 Illustrating the phase-for-gain adjustment
(a) Full phase and gain adjustment for bandpass signals
(b) Phase adjustment substituting for gain adjustment

4. Sensitivity

The operational amplifier integrator is able to achieve picosecond sensitivities [5, 7, 8], because we are only interested in the steady-state value of the integrator output and not in the manner in which this output is reached. It turns out that to first order the operational amplifier frequency response does not affect the steady-state output. Second order effects set a limit to the maximum gain that the integrator can achieve and this sets a limit to the sensitivity.

For high SNR's, as a "rule of thumb", the sensitivity limit is given by 1/1000 of the reciprocal of operational amplifier gain x band width product in Hz. For a gain x bandwidth product of 1 GHz, the sensitivity is a few picoseconds [7]. This is not the ultimate that can be obtained. By careful design even higher sensitivities can be achieved and even still better results with averaging.

When SNR is low, the sensitivity limit is set by the noise present in the signal itself. Input noise contributes an output which is interpreted by the measurement process as a fluctuation in the quantity that is being measured. Errors introduced by noise after one stage of integration have been calculated, showing that δt_d, the standard deviation in the t_d estimate, is given by

$$\delta t_d = \sqrt{\frac{T}{2(SNR)B}} \,, \qquad \ldots(3)$$

where T is the integration time, SNR is the signal-to-noise ratio and B is the noise bandwidth [17].

Further work [18] has shown that integrator is a near optimal time delay estimator, approaching the Cramer-Rao limit at least within a factor of 2.

5. Conclusion

The integrator technique of time delay measurement has been reviewed and its suitability to UWB systems pointed out. Of particular significance is its applicability to position location systems.

6. References

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