Digital Time of Flight Measurement for Ultrasonic Sensors

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Abstract. Ultrasonic sensor measurements are mostly based on the determination of the time of flight (T.o.F.). The paper presents the development of a digital algorithm for pulse-echo measurement applications, based on the use of a cross-correlation function to determine the T.o.F.. Some experimental results are presented, and the possibility to realize a low cost real-time measurement system is considered.

I. Introduction

Ultrasonic sensors are widely used in industrial applications to measure both distance in air and flow velocity. The operating principle is based on the measurement of the time of flight (T.o.F.), that is the time necessary for an ultrasonic wave to travel from the transmitter to the receiver through the target, over which it is reflected back. The length of this path is directly proportional to the T.o.F..

Different techniques can be used to generate ultrasonic waves. Among them, continuous wave [1] and pulse-echo techniques [2], [3] are widely known. More complex methods involving the modulation of either continuous or pulse waves have been reported, for instance in [4], [5] and [6].

In continuous wave methods, the transmitter generates a continuous output, whose echo is detected by a separate receiver. Accuracy depends on the measurement of phase shift between the transmitted and reflected wave. Although better performances than with pulse-echo measurements can be obtained, complex hardware is required to measure phase and, in most cases, different frequencies need to be used in order to determine the number of integer wavelengths in the phase shift.

Pulse-echo techniques are widely used in commercial systems for industrial applications. With this method a short train of waves is generated, enabling the same transducer to be used both as a transmitter and as a receiver. The simplest and most common way to detect the echo signal is a threshold method, where detection occurs when the signal crosses a given amplitude level [7]. Although it has proved to be simple and cheap, the technique may suffer from poor resolution, particularly when the echo pulse has been much attenuated. An improvement consists in the adoption of an adjustable amplitude threshold, in which case detection does not depend on the magnitude of the signal but only upon its shape [8]. In the presence of noise due to air turbulences, and/or mechanical vibrations close to the transducer operating frequency, both methods are known to achieve a repeatability of some tenths of the ultrasonic wavelength.

In this paper the use of digital signal processing methods in a pulse-echo measurement application is presented. The transmitted and echo pulses are transduced into electrical signals and digitized. Their cross-correlation is computed in digital form, and the delay between the two is estimated as the time index corresponding to the peak of the correlation function. From this the T.o.F. is obtained. The reasons for this approach are mainly two; i) from a conceptual point of view, signal processing algorithms can provide improved resolution and at least comparable accuracy, and ii) a measurement system based on dedicated digital hardware is at present not only feasible, but also competitive in terms of cost and performances.

An analysis of the proposed method will be carried out with particular emphasis on the algorithms that have been selected. Consideration will also be given to the problem of real-time measurement, that motivates some of the choices in the implementation. Finally, experimental results will be presented.

II. Analysis of the Problem

Measurement of the time of flight can be considered in the general framework of time delay estimation. The transducer generates a signal, $s(t)$, that propagates to the target and is reflected back, being detected after a delay D. The T.o.F. measurement system has to determine the length of this time interval.

Ideally, the echo is undistorted and it is assumed to have been attenuated by a coefficient α, so that one can express it in the form $\alpha s(t)$. In practice, these hypotheses can be only approximately satisfied; besides this, external disturbances such as turbulences and vibrations, as mentioned in the previous section, will also affect the signal. Finally, quantization noise is introduced when signals are converted and digitized.

The measurement system acquires the two digital sequences $x_T(nT)$ and $x_E(nT)$ representing the transmitted and echo signals, respectively, that can be written in the form:

$$x_T(nT) = s(nT) + v(nT); \qquad x_E(nT) = \alpha\, s(nT-D) + n(nT) \tag{1}$$

where T is the sampling interval, while $v(nT)$ and $n(nT)$ take into account the discrepancies from the ideal model, and can be considered as zero mean uncorrelated random processes.

The correlation of the two sequences is given by:

$$C(kT) = \sum_{n=-\infty}^{+\infty} x_T(nT)\, x_E(nT+kT) \tag{2}$$

and, according to the hypotheses given above on $v(nT)$ and $n(nT)$, the statistical expectation of this sequence is:

$$E[C(kT)] = \alpha\, C_{ss}(kT-D) \tag{3}$$

where $C_{ss}(kT)$ is the sampled auto-correlation function of the signal $s(t)$. For a finite energy signal, it can be shown that $|C_{ss}(kT)| \le C_{ss}(0)$, therefore $E[C(kT)]$ will have a peak for $k = k_D$, with $D = k_D T$ if the delay is an integer multiple of the sampling interval, otherwise $D = (k_D + \delta)T$ with $|\delta| \le 0.5$. In practice, when a single measurement is taken, the delay can be estimated by finding the peak of the correlation (2).

The signal generated by ultrasonic sensors is a short

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train of ultrasonic waves. The electrical equivalent of the transducer is a high quality resonant circuit, therefore the signal that is generated in response to the transmit command can be represented in the form:

$$s(t) = a(t) \cdot \sin(2\pi f_0 t + \phi_0) \tag{4}$$

where f_0 is the resonant frequency of the transducer, a(t) represents the envelope of its response and has a finite duration.

For this class of signals it can be shown that the following relationship holds:

$$C_{ss}(\tau) \cong C_{aa}(\tau) \cdot \frac{1}{2} \cos 2\pi f_0 \tau \tag{5}$$

where $C_{aa}(\tau)$ is the auto-correlation of the envelope function a(t). The behavior of $C_{ss}(\tau)$ is depicted in fig. 1, that refers to the transmission pulse from an ultrasonic transducer. Unfortunately, in this case the resolution obtained in estimating the delay D by determining the peak of the function C(kT) depends on the period of the sinusoidal term in (5), rather than on the sampling interval T. This yields a resolution of $1/f_0$, that, in terms of distance, corresponds to one half of the ultrasonic wavelength λ. Although this may be comparable with some echo threshold methods, the approach can be further refined to improve the result.

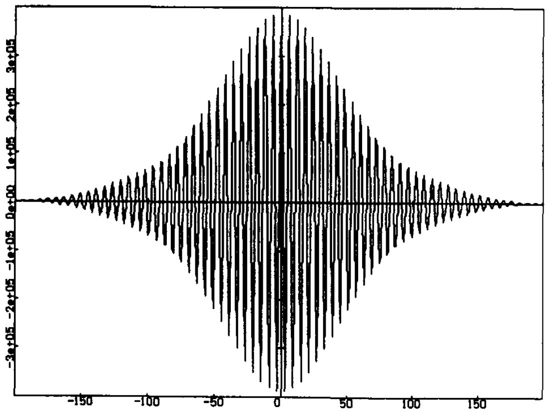

Fig.1. Auto-correlation of a typical transmitted pulse.

It should be noticed that the form of equation (5) is quite similar to (4), except for the replacement of a(t) with $C_{aa}(\tau)$ as the envelope function. However, estimation of the delay D could also be based on $C_{aa}(\tau)$ itself, which has the same properties as $C_{ss}(\tau)$. The advantage is that resolution can be expected to improve in this way, since the sinusoidal term is no longer present.

III. The Measurement Algorithm

The algorithm presented in this paper is based on the idea of obtaining from the sampled sequences $x_T(nT)$ and $x_E(nT)$ a sequence that is related to the auto-correlation function $C_{aa}(\tau)$, from which the delay D can be estimated with a resolution of the same order as the sampling interval T.

In some applications, such as the measurement of distances about one meter or more, the T.o.F. can be quite large, and very long data sequences would be acquired if the receiver signal is sampled continuously. However, since s(t) in (4) has a finite duration, a different arrangement can be considered. In fact, samples taken before the echo is received need not be acquired, since the only relevant information is their count. This allows the acquisition of a much shorter sequence of echo samples, $x_E(nT)$. Its length N in fact can be equal to that of the sampled transmission sequence $x_T(nT)$, the only requirements being that the time interval NT is longer than the duration of s(t) and that the echo is entirely contained in the sequence. In this way, to obtain the T.o.F. the delay estimate provided by the algorithm must be algebraically added to sample count, the latter representing the number of samples that were discarded before the acquisition of the echo sequence was started.

It is apparent from equation (4) that the transducer output can be interpreted as an amplitude modulated signal with carrier frequency f_0. The first step of the algorithm is then to recover from the sequences $x_T(nT)$ and $x_E(nT)$ the sampled envelope a(nT); one way to proceed is by demodulation. Different methods were analyzed and tried on experimental data: the final choice was to realize demodulation by a non-linear squaring operation, followed by low-pass digital filtering. The sequences that are obtained will be indicated as $y_T(nT)$ and $y_E(nT)$, respectively for the transmitted and echo signal. Besides its simplicity, the main advantage of this kind of demodulation is that it does not require to accurately determine the carrier frequency f_0. This avoids the need for a specific set up of the system if different ultrasonic transducers have to be used.

Low-pass filtering eliminates the spectral components that are generated around the frequency $2f_0$ as a consequence of squaring, and also reduces the noise content in the signal. The measured amplitude spectrum of the trasmitted signal is shown in fig. 2 for one of the sensors that has been used in the experiments. The signal can be seen to have quite a narrow bandwidth, therefore the low-pass digital filter can also have a low cutoff frequency. Since the same filter must be used both for the transmitted and the echo sequence, the signal to noise ratio of the latter signal dictates such features as transition bandwidth and stopband attenuation, which ultimately determine the order of the filter.

The baseband signals that are obtained by demodulation are in general much oversampled. In fact, if the bandwidth of a(t) is B_a, the highest frequency in the signal s(t) is $f_0 + B_a$, therefore the sampling interval T must be such that: $T < 1 / 2(f_0 + B_a)$. This suggests that the sequences $y_T(nT)$ and $y_E(nT)$ can be decimated without any loss in information: decimation by a factor K (with K integer) would

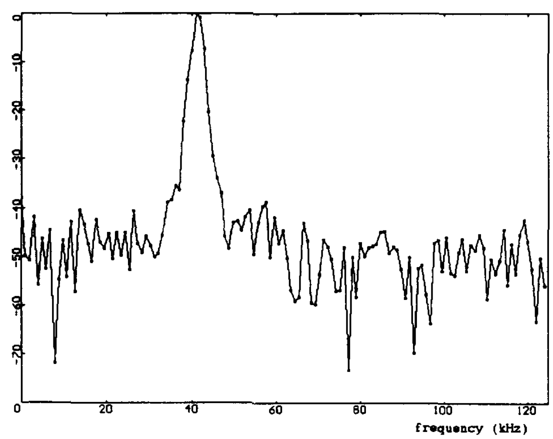

Fig.2. Amplitude spectrum of the transmitted signal.

yield the sequences $y_T(nKT)$ and $y_E(nKT)$. From a computational point of view decimation has the advantage that a lower number of data samples is processed in the operations that follow, since a sequence with N samples has its length reduced to N/K. However, it should be reminded that the sampling interval is increased as a consequence. Therefore, the question has to be considered carefully, to avoid any loss in resolution for the estimation of the delay D.

A possible solution of the problem lies in the interpolation of the samples of the obtained correlation sequence. The correlation of the demodulated and decimated sequences $y_T(nKT)$ and $y_E(nKT)$ can be written as:

$$C\,(kKT) = \sum_{n=0}^{(N/K)-1} y_T(nKT)\; y_E(nKT+kKT), \qquad k = -\frac{N}{K}+1,\ \dots,\ \frac{N}{K}-1 \qquad (6)$$

Indicating with k_D the index of the peak sample in the sequence, $D \approx (k_D + \delta)KT$, with $|\delta| \le 0.5$. This means that if the delay D has a fractional component (i.e., $\delta \ne 0$), the peak of the corresponding continuous correlation function does not correspond to one of the samples.

The value of $C(k_D KT)$ would correspond to that of the actual correlation peak only if the sequence C(kKT) is shifted by δKT, which could be obtained by the convolution of C(kKT) with a suitable function of δ. Considering the matter in the opposite direction, if the analytical expression of this function is known, the fractional index δ can be determined from the samples of C(kKT) through an interpolation formula.

Even when such analytical expression is found, it can be assumed that in most practical cases the relationship will not be satisfied exactly. Therefore, it was decided to avoid this complication and rely instead on an approximate interpolation formula. Given the samples with index k_D, $k_D - 1$ and $k_D + 1$, a simple second order polynomial approximation of $C(\tau)$ yields a parabola, and the fractional index δ is obtained as the abscyssa of its vertex. Deleting the term KT from the argument of C(kKT) for simplicity:

$$\delta = \frac{C(k_D + 1) - C(k_D - 1)}{2\cdot[2\;C(k_D) - C(k_D - 1) - C(k_D + 1)]} \qquad (7)$$

Simulation analysis of this approximate relationship showed that it introduces an error which in most cases is only marginally larger than with exact interpolation formulas.

IV. Experimental Results

The proposed method was verified experimentally by means of the experimental set up shown in figure 3. T_1 is a 40 kHz ultrasonic transducer that acts both as a transmitter and as a receiver, converting an electrical signal into acoustic and viceversa. After being filtered by a band pass amplifier, whose center frequency is synchronous with the transducer operating frequency, signals from T_1 are acquired by a digitizer. Recalling from fig. 2 that the typical bandwidth of the signals is lower than 10 kHz, a 250 kHz sampling rate was adopted. Both the transducer and the digitizer are controlled by the processing unit, where the data sequences are stored and processed.

This system arrangement would enable to acquire both $x_T(nT)$ and $x_E(nT)$. However, it should be noted that the signal corresponding to the echo is the result of a double conversion process. In fact, the electrical signal applied to the transmitter is converted into an acoustic wave, and the acoustic echo is converted back into an electrical

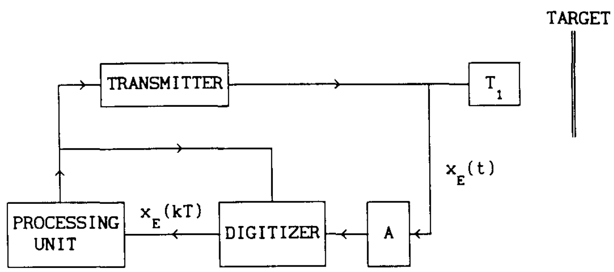

Fig.3. Block diagram of the experimental set-up.

signal. Differences can occur, due to the transducer bandwidth and to the electric impedances of the transmitter and receiver circuits. In order to avoid differences between $x_T(nT)$ and $x_E(nT)$ due to mismatches of the above parameters, it has been chosen to digitize $x_T(nT)$ through a separate receiver, according to the arrangement of fig. 4.

In this scheme T_1 and T_2 are two matched 40 kHz ultrasonic transducers: T_1 acts as the transmitter, while the other is the receiver. Transmission parameters (amplitude and length of the pulse train) have been rigorously standardized. The processing unit simultaneously drives the transmitter and the digitizer, whose input is connected to the receiver, and accepts the acquired data from the digitizer for storage. Strictly speaking, the resulting sequence $x_T(nT)$ is delayed by the time of flight between the transducers T_1 and T_2, however the difference can be corrected as a result of calibration procedures.

With the described system a set of experimental results have been obtained for three different transducer-to-target distances. The length of the data sequences is 256 points, and the T.o.F. has been determined according to the algorithm described in the previous section. Different decimation factors have been tried; eventually, decimation by 4 was settled for, but factors up to 8 still enable to achieve similar results.

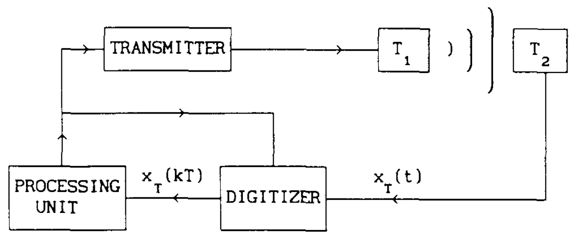

Fig.4. Acquisition of the reference transmitted pulse.

The linear regression of experimental data, presented in fig. 5, provides a very good fit. From the calculated slope a sound velocity of 345 m/s can be determined, while the intercept corresponds to an offset of 10 points. This may be due to time delays between the digitizer trigger and the actual start of the transmit pulse, to the above mentioned T.o.F. difference affecting $x_T(nT)$, or to a constant bias in the evaluation of the origin for measuring distance. Once the offset has been corrected for, performances of the proposed method are good, and a sensitivity of approximately 0.7 mm can be obtained.

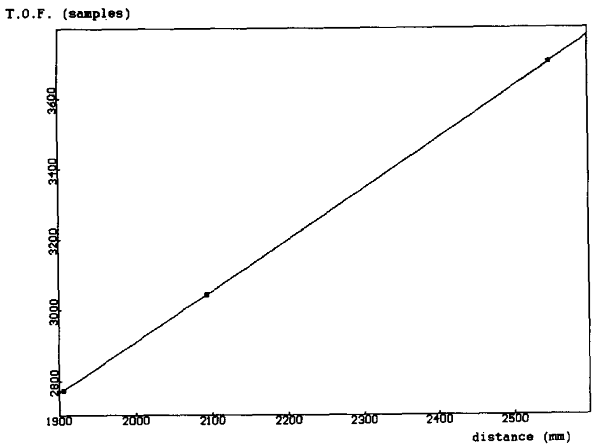

Fig.5. Linear fit of the experimental data.

V. Final Remarks

The approach presented in the paper provides interesting performances in the measurement of the time of flight also with cheap ultrasonic transducers. In the experiments carried out so far a good agreement with reference data has been obtained; without external noise the attainable resolution is of the order of a few tenths of the ultrasonic wavelength. A preliminary analysis suggests that the proposed method is sufficiently insensitive to the effects of superimposed wideband noise if care is taken to process the sampled data sequences by appropriate digital filters.

The measurement algorithm can be implemented on state-of-the-art digital signal processing systems at moderate cost. A preliminary assessment of the computational effort required by the algorithm to process 256-point data sequences, with decimation by 4 after demodulation, shows that a measurement can be obtained in less than 1 ms on a modern DSP chip. This would enable the realization of a low cost real-time measuring instrument, that can be easily integrated, for instance, into a more complex automation system.

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