Envelope pulsed ultrasonic distance measurement system based upon amplitude modulation and phase modulation

Y. P. Huang and J. S. Wang
Department of Electrical Engineering, National Cheng-Kung University, Tainan, 701 Taiwan, Republic of China

K. N. Huang
Department of Electronic Engineering, I-Shou University, Kaohsiung, 840 Taiwan, Republic of China

C. T. Ho, J. D. Huang, and M. S. Young
Department of Electrical Engineering, National Cheng-Kung University, Tainan, 701 Taiwan, Republic of China

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A novel microcomputer-based ultrasonic distance measurement system is presented. This study proposes an efficient algorithm which combines both the amplitude modulation (AM) and the phase modulation (PM) of the pulse-echo technique. The proposed system can reduce error caused by inertia delay and amplitude attenuation effect when using the AM and PM envelope square wave form (APESW). The APESW ultrasonic driving wave form causes a phase inversion phenomenon in the relative wave form of the receiver. The phase inversion phenomenon sufficiently identifies the “measurement pulse” in the received wave forms, which can be used for accurate time-of-flight (TOF) measurement. In addition, combining a countertechnique to compute the phase shifts of the last cycle for TOF, the presented system can obtain distance resolution of 0.1% of the wavelength corresponding to the 40 kHz frequency of the ultrasonic wave. The standard uncertainty of the proposed distance measurement system is found to be 0.2 mm at a range of 50–500 mm. The APESW signal generator and phase detector of this measuring system are designed on a complex programmable logic device, which is used to govern the TOF measurement and send the data to a personal computer for distance calibration and examination. The main advantages of this APESW system are high resolution, low cost, narrow bandwidth requirement, and ease of implementation.


I. INTRODUCTION

Ultrasonic waves are commonly used to measure the distance from a reflected surface or the depth of liquids. The techniques of ultrasonic distance measurement include the time-of-flight (TOF) method,1–3 the phase-shift analysis of single-frequency continuous wave method,4–6 the combination method of time of flight and phase shift,7 the multifrequency continuous wave and phase-shift method,8 the multifrequency amplitude-modulation (AM)-based ultrasonic system,9 and the binary frequency shift-keyed (BFSK) method.10 In addition, some studies also use digital signal processing (DSP) techniques, i.e., the cross-correlation method.2,7

There are two ultrasonic transducers, one of which is used as a transmitter and the other as a receiver. Generally, a geometric distance measurement can be obtained by measuring the TOF of an ultrasonic wave, and measuring the phase shift between the transmitted and the received waves. The distance \( d \) between the transmitter and receiver is evaluated by

\[
\[
\end{eqnarray}
\]

where \( c \) is the sound velocity and \( T_F \) is the wave flight time. A continuous square wave is commonly used in ultrasonic distance measurement. The resonant signal of the receiver may be written as

\[
V_R(t) = E_R \sin(\omega t + \theta),
\]

where \( E_R \) is the peak value of the received signal, \( \omega \) is the resonant angle of the fundamental frequency of the transducer, and \( \theta \) is the phase-shift angle which is linearly proportional to the measured distance. The phase-shift measurement is limited to \( 2\pi \) radians, or one wavelength (\( \lambda \)). Such a small-range limitation would severely restrict the use of phase detection to measure displacement. One method to overcome this limitation is to add the counting of the integer number of the wavelength, which occurs at the displacement range. The total displacement can be calculated by the formula

\[
d = \left( N + \frac{\theta}{2\pi} \right) \lambda,
\]

where \( N \) is the integer number of the wavelength (\( \lambda \)) counted, and \( \theta \) represents the phase angle of the last pulse which is smaller than \( 2\pi \).
Nevertheless, when using the TOF to measure the distance, the system errors are mostly due to the inertia phenomenon of amplitude delay, the amplitude attenuation of the received signal, and uncertainty in the speed of sound caused by environmental temperature and humidity variations. The effects of temperature and humidity can be satisfactorily compensated for by adding a temperature and a humidity sensor circuit to the ultrasonic transducers. The root cause of the inertia phenomenon of amplitude delay when starting vibration is due to the piezoelectrics of the ultrasonic transducer, which reduces the amplitude of the echo wave form. Figure 1 shows the emission and the echo wave form from a pair of ultrasonic transducers. The inertia delay caused by long rise time will add to the TOF and cause error. That is also the most difficult error to overcome of the TOF method. In addition, the amplitude of the echo wave form will change with distance, which is caused by beam spreading and attenuation. The amplitude decreases very fast in the air, which will directly cause the accuracy of the ultrasonic transducer to decrease. As shown in Fig. 2, for short distance measurements (distance 1), the error caused by inertia delay (delay 1) is smaller because high energy is transferred between emission and reception; but in long distance measurements (distance 2), the error (delay 2) will become larger because of the attenuation of the ultrasonic wave. Additionally, most methods of ultrasonic distance measurement cannot avoid the inertia delay of machine vibration. As a result, the longer the traveling distance of the wave, the more serious the error of TOF caused by the accumulation of the inertia delay and the amplitude attenuation.

Although the error caused by amplitude attenuation can be improved by making the threshold level variable or using the automatic gain controlled (AGC) technique, they still cannot provide very high accuracy. The accuracy of the DSP techniques for phase-shift computation is limited by the resolution of the analog to digital (A/D) converter. In addition, DSP techniques are more expensive due to complex hardware and software. On the other hand, the BFSK method measures the position by signal processing of phase digitized information from the received signal, the TOF, and the phase transition between two different frequencies. However, this method is easily influenced by noise. The BFSK also has an extremely critical problem of lagging, there are approximately ten cycles of uncertainty waiting time (approximately 250 µs for received signal settles to its steady state when using 40 kHz waves) when changing between the two frequencies.

Supplying two pulse trains subsequently to an acoustic transducer will produce an envelope zero phenomenon in the receiver wave. The measurement of TOF can be retrieved by detecting the envelope zero. However, the width of the envelope zero is easily influenced by noise and depends on the signal-to-noise ratio (SNR) of the echo signal. Since controlling the time interval between the two pulse trains will produce a phase inversion which occurs during the envelope zero, it is possible to obtain more accurate TOF measurements by detecting the phase inversion phenomenon. This article proposes a novel driving algorithm for an ultrasonic transmitter, which can overcome the inertia delay problem and achieve accurate distance measurements. We use 40 kHz ultrasonic driving wave forms which combine both amplitude-modulation (AM) and phase-modulation (PM) waves. The phase modulation is used to cause the phase inversion and identify the specific pulses for TOF measurement. The amplitude modulation can enhance and accelerate the phase modulation, and reduce the amplitude attenuation effect. Ten warm up pulses are added to reduce the inertia delay. The distance measurement algorithm is based on both TOF and phase-shift techniques. The proposed system can achieve high accuracy and high resolution in distance measurement.

II. THE MEASUREMENT METHOD

A. Transmitted signals and received signals

The amplitude-modulation and phase-modulation envelope square wave form, which is called APESW in the following, is proposed to drive the ultrasonic transmitter for distance measurement. Figure 3 shows the system block diagram which is used for generating APESW signals and processing received signals from the receiver. The APESW signals are originally generated from a complex programmable
logic device (CPLD) with phase modulation, and then are processed by a level shifter and amplitude-modulation circuit. The modulated signals “transmitted signal $S_T$” are then sent to the driving circuit and form “driving wave form TX” to drive the ultrasonic transmitter. The signals which are received by the ultrasonic receiver are processed by a preamplifier circuit and form “received sinusoid wave form RX.” The “received signal $S_R$” is transferred from a sinusoid form to a square wave form [transistor-transistor logic (TTL) signal] by clamping and a Schmitt-trigger circuit, and is then sampled by a CPLD for analysis.

Figure 4 shows the proposed driving wave form. One APESW wave consists of ten low-amplitude pulses, followed by two high-amplitude waves and then another ten low-amplitude pulses. The first ten 40 kHz low-amplitude square waves are used to warm up both the ultrasonic transmitter and receiver, which can eliminate the inertia delay caused by the piezoelectric effect of the ultrasonic transducer. In order to distinguish from the warm up waves and enhance the real waves for distance measurement, the measurement waves are modulated at both amplitude and phase. The measurement waves have higher amplitudes and start with a double-width pulse and then another ten low-amplitude pulses. The first ten warm up waves (phase I) and the received sinusoid wave form (RX) are described in Fig. 3. The ultrasonic transmitter and receiver are placed face to face, and the wave forms are measured by a mixed signal oscilloscope (54622D, Agilent). After the phase change period (low amplitude in the wave form of RX), the output of the PM detector 2 (PM_DET in the figures) goes from low to high as soon as it detects the phase inversion of the $S_R$ signal.

The measurement wave form is easily identified via the phase-modulation characteristic. Two phase-modulation detectors (PM detectors 1 and 2) are used to detect the phase changing phenomenon of the transmitted and received signals, respectively. The PM detectors detect the phase according to the 40 kHz counter clock at every rising edge of $S_T$ and $S_R$ signals. The outputs of the PM detectors are used for $T_F$ counting. The $T_F$ represents the traveling time between the transmitted and received signals. The measurement of TOF starts from the rising edge of the measurement pulse (the rising edge of PM detector 1) that directly follows the double-width pulse and stops when detecting the phase inversion of the received signal (the rising edge of PM detector 2). The amplitude of the received signal is relatively low during the phase change period, and normally the $S_R$ will not have any output or will only have shorter pulse output during the beginning of this period. As a result, the first divergent phase that is detected by PM detector 2 is the relative measurement pulse. Figures 6(a) and 6(b) show the actual driving wave form (TX) and the received sinusoid wave form (RX) which are described in Fig. 3. The ultrasonic transmitter and receiver are placed face to face, and the wave forms are measured by a mixed signal oscilloscope (54622D, Agilent). After the phase change period (low amplitude in the wave form of RX), the output of the PM detector 2 (PM_DET in the figures) goes from low to high as soon as it detects the phase inversion of the $S_R$ signal.

B. Computation of TOF and distance

As shown in Fig. 7, TOF is the sum of two separated parts. One part estimates the integer number $N$ of full-cycle clock, and the other calculates the phase shift ($\theta_{ST}$) of the last cycle which is smaller than one full cycle. The TOF of the measurement ($T_F$) can be expressed as

$$T_F = \left( N + \frac{\theta_{ST}}{2\pi} \right) T_{\text{period}},$$

where $N$ represents the integer number of the full-cycle clock, $\theta_{ST}$ is the phase-shift quantity between emission and echo waves, and $T_{\text{period}}$ is the time period of the counter-clock.

The system uses the counter-clock ($f_{\text{clk}}=40$ kHz) which is derived from the system clock ($f_{\text{clk}}=40$ MHz) divided by 1000 for calculating values of the integer of $N$. The ultra-
Sonic transmitter driving wave form also aligns with this 40 kHz counterclock. Therefore, the integer number \( N \) can be estimated by \( N=\text{INT}[T_F/T_{\text{period}}]=\text{INT}[T_F(40 \times 10^3)] \), where \( T_F \) is the actual traveling time of the ultrasonic waves, \( T_{\text{period}} \) represents the period of the counterclock, and \( \text{INT} \) represents the integer of the given number. As shown in Fig. 7, \( N \) is counted from the falling edge of the 40 kHz counterclock that corresponds to the rising edge of the measurement wave (the rising edge of PM detector 1), and the counterclock stops at the rising edge of the PM detector 2 which rises as soon as it detects the phase inversion of the received signal.

In order to avoid the limitation caused by the amplitude of the signal and the finite bits of the A/D converter, the phase shifts of the last cycle are computed by a counterclock technique.\(^{14}\) As shown in Fig. 8, the phase shift between transmitted and received signals \( (\theta_{ST}) \) is calculated from the last incomplete cycle. The phase-shift quantity is calculated by the original system clock \( (f_{\text{clk}}=40 \text{ MHz}) \). Therefore, the theoretical maximum distance resolution of the system can reach 0.00865 mm (at 25 °C environment temperature, \( c/f_{\text{clk}}=346 \text{ m/s}/[40 \times 10^6 (1/s)] =0.00865 \text{ mm} \), where \( c \) is velocity and \( f_{\text{clk}} \) is the frequency of the system clock). The 40 MHz counterclock for phase-shift measurement is activated when the system senses the phase inversion of the received signal (the rising edge of PM detector 2). The counterclock is deactivated when the next falling edge of the 40 kHz counterclock is sensed. The counting number \( (m) \) can be translated to phase quantity \( (\varphi) \) by

\[
\varphi = m \times \left( \frac{40 \text{ kHz}}{40 \text{ MHz}} \right) \times 2\pi = \frac{m \times 2\pi}{1000}.
\]

The phase-shift quantity \( (\theta_{ST}) \) between transmitted signals and received signals can then be obtained by

\[
\theta_{ST} = 2\pi - \varphi.
\]

The algorithm derived for computing distance \( d \) is explained as follows. If the velocity of ultrasound \( c \) is constant, then the wavelength \( \lambda \) can be determined as \( \lambda = c/f \) where \( f \) is the frequency of the 40 kHz ultrasonic wave. The distance \( d \) represents the total displacement; Eq. (1) can be rewritten as

\[
d = cT_F = c \left( N + \frac{\theta_{ST}}{2\pi} \right) \frac{1}{f} = \left( N + \frac{\theta_{ST}}{2\pi} \right) \lambda,
\]

where \( N \) is the counted integer number of the given wavelength, \( \theta_{ST} \) is the phase-shift degree of the last cycle, and \( T_F \) is the time of flight. This algorithm for TOF measurement can be easily programmed into a digital microprocessor system providing high accuracy at low cost.
C. Estimation of uncertainty

In order to account for uncertainty, the distance computing equation (1) can be rewritten as

\[ d = kT_Fc, \]

where \( d \) is the distance between the transmitter and target, \( c \) is the sound velocity, \( T_F \) is the wave flight time, and \( k \) is the constant close to 1 which depends on the sensor geometry. Through calibration, the uncertainty contribution due to the constant \( k \) can be made negligible. The sound velocity \( c \) for an ideal gas at a constant pressure is\(^{15} \)

\[ c = \sqrt{\frac{\gamma RT_0}{M}}, \]

where \( \gamma, M, R, \) and \( T_0 \) are the specific heat ratio, molar mass, universal gas constant, and the absolute temperature (in Kelvin), respectively. Disregarding the variable of humidity, the velocity of sound \( c \) which increases as the square root of the absolute temperature can be substituted to centigrade \( t \) and expressed as

\[ c = 331.45 \times \sqrt{1 + \frac{t}{273}}. \]

Equation (10) is suitable for dry air; however, in order to account for the air with moisture (water vapor), the formulas need correction. The approximate equation of the sound velocity ratio over the range for relative humidity \( h \) from 0 to 1.0, and temperature \( t \) from 0 to 30 °C is\(^{6} \)

\[ \frac{c_h}{c_0} = 1 + h(9.66 \times 10^{-4} + 7.2 \times 10^{-5}t + 1.8 \times 10^{-6}t^2 + 7.2 \times 10^{-8}t^3 + 6.5 \times 10^{-14}t^4), \]

where \( c_h \) and \( c_0 \) are the sound velocities at relative humidity \( h \) and dry air, respectively. As a result, the velocity of sound in air varies with temperature \( t \) and air humidity \( h \) and can be expressed as

\[ c = f(t, h). \]

As the measured quantities \( T_F, t, \) and \( h \) can be considered uncorrelated, the standard uncertainty \( u(d) \) of the measured distance can be obtained from Eq. (8) as\(^{12} \)

\[ u(d) = \sqrt{(kT_F)^2 \left[ \left( \frac{df}{dt} \right)^2 u^2(t) + \left( \frac{df}{dh} \right)^2 u^2(h) \right] + (kc)^2u^2(T_F)}, \]

where \( u(t), u(h), \) and \( u(T_F) \) are the standard uncertainties of the temperature, humidity, and time of flight. The sensitivity coefficients for \( t \) and \( h \) can be derived from Eqs. (10) and (11) for standard uncertainty estimation.

III. SYSTEM IMPLEMENTATION

A. Description of the hardware system

The APESW transmitted signals are generated in digital format by a CPLD (FLEX10K10, Altera Corporation). As shown in Fig. 9, the amplitude modulation is achieved by modulating the supplied power (+5 and +12 V) of the output amplifier (CD4069, National Semiconductor). The output amplifier is the last stage circuit for ultrasonic transmitter driving. The voltage levels of the output amplifier CD4069 are +5 and +12 V; however +3.3 V is required for the CPLD signals, thus a level-shift circuit is added to shift the driving signals for the transmitter. After amplitude modulation, the signal levels become +5 and +12 V. The amplitude modulation can reduce the amplitude attenuation effect and enhance the phase modulation. On the other hand, the phase modulation can be simply achieved by controlling the pulse width of the transmitted signal from the CPLD.

The hardware implementation is shown in Fig. 10. The spec of the ultrasonic transmitter/receiver (400ST160/400SR160, Pro-Wave Electronics Corporation, Taiwan) has a 55° beam angle, 40 kHz center frequency, and 2 kHz bandwidth. The receiving sensitivity is ~60 dB at 40 kHz (0 dB = 1 V/μbar). The amplitude-modulation circuit consists of a voltage regulator (7805, Fairchild Semiconductor), an NPN transistor (2SC1815, TOSHIBA, Japan), three resistors (7, 5, and 5 kΩ), and a capacitor (300 pF). The control signal from the CPLD changes the supplied voltage of the output ampli-
The preamplifier (CD4069, National Semiconductor). The supplied voltage is switched between +5 and +12 V, which is used to modulate the amplitude of the driving wave form for the transmitter. When the control signal is at +5 V level, the NPN transistor will be in the saturation region and the regulator will bypass the 7 kΩ resistor and go to ground. As a result, the output voltage of the regulator will be +5 V. On the other hand, when the level of the control signal is 0 V, the NPN transistor will be in the cut-off region and the 7 kΩ resistor will combine with 5 kΩ resistor which will shift the output voltage to +12 V. Since the voltage levels of the output amplifier CD4069 are +5 and +12 V, but +3.3 V is required for the CPLD signals, a level-shift circuit is added to shift the driving signals for the transmitter. The level-shift circuit consists of a buffer (CD4050, National Semiconductor), an open-collector buffer (74HC07, ST Microelectronics), and a 5 kΩ pull-up resistor. The receiver circuit consists of a preamplifier (uA741, Texas Instruments), followed by a clamping circuit and an inverting Schmitt trigger (74HC14, Texas Instruments). The clamping circuit which consists of a diode (1N4148, Fairchild Semiconductor) and a 0.1 μF capacitor clamps the received signals which limits them to a positive level between 0 and 2 V. The clamped signals are then transferred from a sinusoid form to a square wave form by an inverting Schmitt trigger, and the signals are then sampled by the CPLD.

The APESW signal generator and phase detector of this measuring system are designed on a CPLD, which is used to govern the operation of the entire system and compute the measured distance. Figure 11 shows the block diagram of the programmed circuit inside the CPLD. The 40 MHz system clock is divided by 1000 to generate the 40 kHz clock. The 40 kHz signal is used as a reference clock for the amplitude-modulation (AM) block, phase-modulation (PM) block, \(N\) values, and phase-shift quantity (\(\theta_{ST}\)) calculation circuit blocks. The amplitude-modulation block provides the control signal to modulate the power of the output amplifier circuit for the transmitter, and the phase-modulation block modulates the pulse width of the transmitter. The received signals are analyzed through a phase-modulation detector block to detect the measurement pulse. The \(T_F\) calculator block calculates the \(N\) value and phase-shift quantity \(\theta_{ST}\) to measure the TOF. The TOF data are sent to a personal computer (PC) for distance calculation.

![Figure 11](image1.png)

**FIG. 11.** The block diagram of the programmed circuit inside the CPLD.

The TOF data value \(T_F\) from \(N\) and \(\varphi\) and the TOF data value \(T_F\) from \(N\) and \(\varphi\) and sends the data to a PC for distance calculation.

### B. Software block of the system

The software block diagram of the ultrasonic distance measurement system is shown in Fig. 12. The distance measurement signal \(S_T\) which is transmitted from the ultrasonic transducer will receive the reflex signal \(S_R\) from the ultrasonic receiver. The TOF data value \(T_F\) can be obtained by calculating the time difference between \(S_T\) and \(S_R\) using time counter. The time counter uses two different clocks for reference: the 40 kHz clock measures the \(N\) value and the 40 MHz clock is for precise phase-shift \(\theta_{ST}\) calculation. At the beginning, the control algorithm for the transmitters resets the time counter and then sends out ten warm up pulses. Next, the AM and PM waves are transmitted. The \(N\) value time counter will be activated as soon as the phase modulation of \(S_T\) is detected. The \(N\) value counter will be stopped as soon as the phase inversion of \(S_R\) is detected. At the same time, the \(\varphi\) value time counter starts counting and counts till the next falling edge of the 40 kHz clock. The system calculates the time of flight \(T_F\) from \(N\) and \(\varphi\) and sends the data to a PC for distance calculation.

![Figure 12](image2.png)

**FIG. 12.** The software block diagram of the system.

### IV. EXPERIMENTAL RESULTS

The experiment is achieved in a programmable temperature and humidity chamber (GTH-800-00-CP-ST, Giant Force Instrument, Taiwan). Figure 13 shows the block diagram of the experiment system. The experimental embodiment of the distance measurement system consists of two acoustic transducers with a transmitter and a matching receiver, a signal generation and power amplifier system, a voltage regulator dc power supply circuit, and a preamplifier and clamping circuit system. A microprocessor-based controller CPLD governs the operation of the measurement system and sends the time of flight \(T_F\) data to a PC. The
position of the ultrasonic receiver is controlled by a stepping motor and the system is calibrated in a known distance \(d\) by an optical linear scale. The step motor, which is controlled by the PC via the RS-232 interface, has 400 steps per circle and the resolution of the ball screw is 10 \(\mu m/\text{step}\) (i.e., the lateral motion is 4 mm/step resolution). The temperature \(t\) and relative humidity \(h\) of the chamber are sensed and recorded by a thermal/humidity meter (TES-1365, TES Electrical Electronic Corporations, Taiwan). The output of the reading is sent to the PC. As a result, for any given position, the PC can be calculated and displays the distance via the received \(T_F\), \(t\), and \(h\). The error of the measurement can then be used for calibration and examination.

The temperature and relative humidity are controlled at 25 °C and 60%RH, respectively, during the distance measurement experiment. The range of the distance measurement is set as 50–500 mm. The effects of temperature and humidity uncertainties on the distance uncertainty are estimated at a distance of 100 mm. The temperature is regulated over the range of 5–40 °C at a fixed 60%RH, and the relative humidity is controlled over the range of 30%–90%RH at a fixed temperature of 25 °C, respectively. A logged data graph of the actual distance versus measured distance from approximately 50 to 500 mm is shown in Fig. 14(a). There are other factors that might affect the accuracy of the distance measurement: the delay time of the circuit and the actual position of the ultrasonic piezofilm which is hard to predict accurately with a metal cover. In order to eliminate those factors, the actual distance is counted relatively, that is, we set the starting point at 50 mm and increase 10 mm/step. The measured distance and the actual distance are calculated relative to the 50 mm starting point. The plot of deviation versus distance is shown in Fig. 14(b). The experimental standard deviation of the linearity with respect to the distance is found to be 0.0897 mm, while the temperature and humidity uncertainty effects produce standard deviations of 0.18 and 0.017 mm, respectively [see Figs. 14(c) and 14(d)]. The combined stan-
standard uncertainty \( u(d) \) of the measured distance can be obtained by Eqs. (10), (11), and (13). The sensitivity coefficients for \( t \) and \( h \) are about 0.6067 m s\(^{-1}\) °C\(^{-1}\) and 0.0173 m s\(^{-1}\) %RH\(^{-1}\) at 60%RH and 25 °C, respectively. Consequently, the combined standard uncertainty \( u(d) \) is about 0.2 mm in the range of 50–500 mm at temperature of 5–40 °C and a relative humidity of 30%–90%RH with the compensation of the temperature and the humidity.

V. DISCUSSION

The most significant uncertainty factor in the experiment system is the temperature (0.18 mm), which is due to the uncertainty of the employed temperature sensor, the nonideal behavior of the compensation strategy, the noise, wind velocity, and thermal gradients in the measuring path. The accuracy of the employed thermal meter is 0.5 °C. If the error is considered a random variable with uniform distribution, it corresponds to a distance uncertainty about 0.04 mm for a distance measurement of 100 mm. The uncertainty caused by the nonideal behavior of the compensation strategy is also relatively small. The average wind velocity is approximately 0.8 m/s with a standard deviation of 0.05 m/s, which corresponds to a distance uncertainty of about 0.015 mm for a distance measurement of 100 mm. Consequently, the noise and thermal gradients in the measuring path are the significant factors of the temperature uncertainty in the experiment chamber.

The proposed ultrasonic distance measurement algorithm is based on both TOF and phase-shift techniques. When using ultrasonic waves to measure distance in the air, the TOF technique will encounter some system errors. These errors are primarily due to the inertia delay and amplitude attenuation of the received signals. In this article, we propose that the APESW driving wave forms can reduce these two effects. In order to reduce the inertia delay effect, ten warm up pulses are sent to the ultrasonic transducers before distance measurement. The amplitude modulation can reduce the amplitude attenuation effect and enhance the phase modulation. The phase modulation can make the specific pulses for TOF measurement easily identified in the received signals. In addition, in order to avoid the limitation caused by the amplitude of the signal and the finite bits of the A/D converter, the phase shifts of the last cycle are computed by a countertechnique in this system.\(^{14}\) Therefore, the theoretical maximum distance resolution of the system can reach 0.1% of the wavelength. Experimental results indicate that the proposed ultrasonic distance measurement system can accurately measure the distance. The algorithm is simple and can be easily adapted for other microprocessors. Compared to the BFSK phase-shift method,\(^{14}\) this system requires a single 40 kHz signal. This improvement reduces the complex circuit and the error caused by two different frequencies. The main advantages of this APESW wave form system are high resolution, low cost, narrow bandwidth requirement, and ease of implementation.

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