

QUADRATURE PHASE DETECTION IN AN ACOUSTIC POSITIONING SYSTEM

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A novel acoustic navigation method is described. The method is based on a continuous waveform transmitted from a tethered object being tracked and received by a bottom node. Accurate range and bearing measurements are made by differential phase detection and phase unwrapping of the received signals. Phase detection is performed using a quadrature phase sensitive detector (QPSD). The paper investigates, using computer simulation, the performance of the system in the presence of noise. The RMS error in the phase estimate was found to be 14° for range estimation and 17° for angle estimation even with a poor SNR of 0 dB. These estimates represent a theoretical equivalent range error of 5.8 mm and angle error of 0.5° , respectively, for an operating frequency of 10 kHz and 5λ hydrophone separation. These results indicate potentially a much better accuracy than that offered by the conventional ultra-short baseline system.

INTRODUCTION

An ability to navigate an underwater platform such as a remotely operated vehicle (ROV) or bottom crawler to precise bottom locations or along a precise path is essential to many scientific experiments and ocean bottom surveys. Similarly, knowledge of the precise position of an underwater moving platform is essential in many operational scenarios including underwater archeology, security surveillance and others.

Three standard techniques permit tracking of underwater objects: Long Baseline (LB), Short Baseline (SB), and Ultra-Short Baseline (USBL) acoustic navigation [1]. USBL systems are frequently used to track tethered and un-tethered underwater vehicles from a surface vessel. The surface vessel has a multi-element acoustic receiver array that measures the angle

as well as the range to an acoustic beacon placed on a tracked object. The system measures an object's range from time-of-flight of an acoustic harmonic pulse (a ping) and an object's bearing from the arrival time (phase) differences among three or more hydrophones. Range and bearing obtained through use of a USBL system are with respect to the transceiver mounted on the vessel. In order to correct for ship angular motions (roll, pitch and yaw), suitable sensors must be installed to monitor these motions.

A novel system was proposed [2] with navigation based on phase monitoring of the continuous acoustic harmonic signal (CW) transmitted from an underwater moving tethered platform and received by the receiving acoustic array positioned on the bottom. This approach potentially offers superior accuracy for both target range and bearing measurements. It has been tried for distance measurements in air with reported accuracy of 0.295 mm at the range of 0.5842 m at frequency of 40 kHz [3]. We shall briefly describe the CW approach and then focus on phase measurement used in our system.

1. NOVEL NAVIGATION SYSTEM OVERVIEW

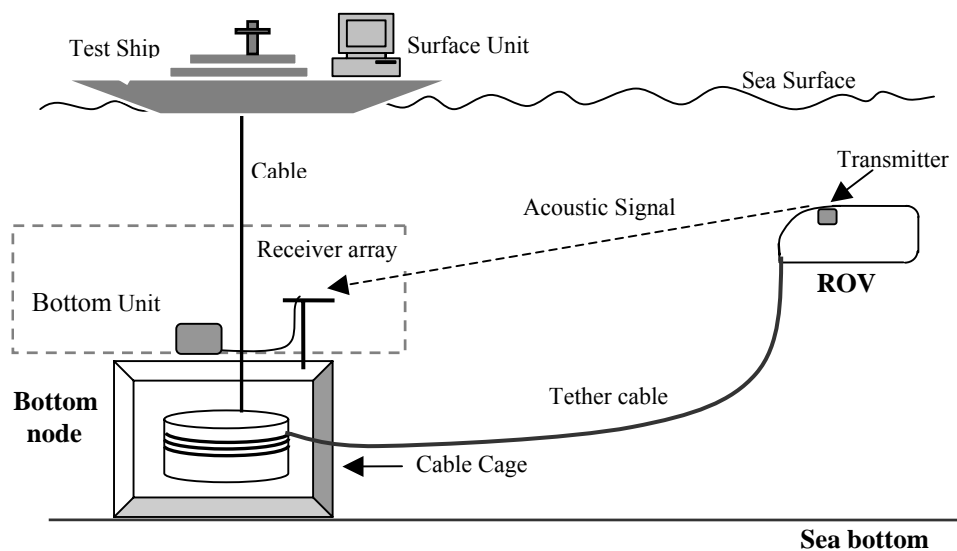


Fig.1 Novel Navigation System

In the proposed scenario, shown in Fig. 1, a cage with tethered ROV or bottom crawler is lowered from the surface vessel to the bottom. On the cage, a Bottom Unit is mounted, consisting of the acoustic receiving array and associated electronics. Upon command from the surface vessel, the ROV leaves the cage for its intended mission. The objective of the navigation system is to monitor the position of the ROV (i.e., to estimate its range and bearing) with respect to the cage (called the bottom node). To accomplish this objective, an acoustic transmitter installed on the ROV is driven by a continuous harmonic electrical signal generated at the node and transmitted over the connecting tether. The acoustic signal transmitted from the ROV is received by a receiving acoustic array located at the node. In this arrangement, the transmitted signal is known precisely at the receiver. The ROV mission

starts from a fixed and known initial position, at a certain known range r_0 from the node (typically the distance from the ROV transmitter while in the cage to the acoustic receiving array). The ROV current position is determined by continuously monitored phase changes between the transmitted and the received signals. The phase changes are caused not only by the time-varying position of the ROV, but also by ambient noise, the presence of multipath and other factors. In this paper, we concentrate only on ambient noise.

Initially, we discuss the range measurement. To commence its mission, the ROV moves from its initial position at range r_0 to a new position at range r with a range increment Δr , that is:

$$r = r_0 + \Delta r. \quad (1)$$

Let the ROV transmit a constant harmonic acoustic signal $x(t) = A \exp(j2\pi f t)$ with wavelength $\lambda = c / f$, where c is the velocity of sound in water.

The received signal $y(t)$ is:

$$y(t) = B \exp(2\pi f (t - \frac{r}{c})). \quad (2)$$

Therefore, the phase difference increment between the transmitted signal $x(t)$ and the received signal $y(t)$ is

$$\Delta\phi = 2\pi f \Delta r / c = \frac{2\pi}{\lambda} \Delta r. \quad (3)$$

Equation (3) means that each range increment between the ROV and the node equal to the wavelength λ will result in a 2π -phase difference between the transmitted and the received signals. By continuous measurement of the phase $\Delta\phi$, we obtain the range increment of the ROV as

$$\Delta r = \frac{\Delta\phi\lambda}{2\pi} \quad (4)$$

and, therefore, the position of the ROV is

$$r = r_0 + \Delta r. \quad (5)$$

The phase measured value is ambiguous due to its periodic nature. With a phase detector, we can measure only a phase ϕ ranging from $-\pi$ to π radians. If phase exceeds π radians, it “flips” back to $-\pi$ radians. The phase change $\Delta\phi$ in Eq. (4) can however go beyond π radians as the distance increment Δr increases. To make a correct phase measurement, we monitor each phase flip and add them to the current measured phase ϕ . This procedure is called phase unwrapping (see Appendix for details).

The total phase is therefore calculated as:

$$\Phi = \text{Unwrap}(\Delta\phi) = 2\pi k + \Delta\phi + \phi_0 \quad (6)$$

where k is the number of phase flips and ϕ_0 is the initial phase at $r = r_0$. Fig.2 depicts the proposed system flowchart for range measurement.

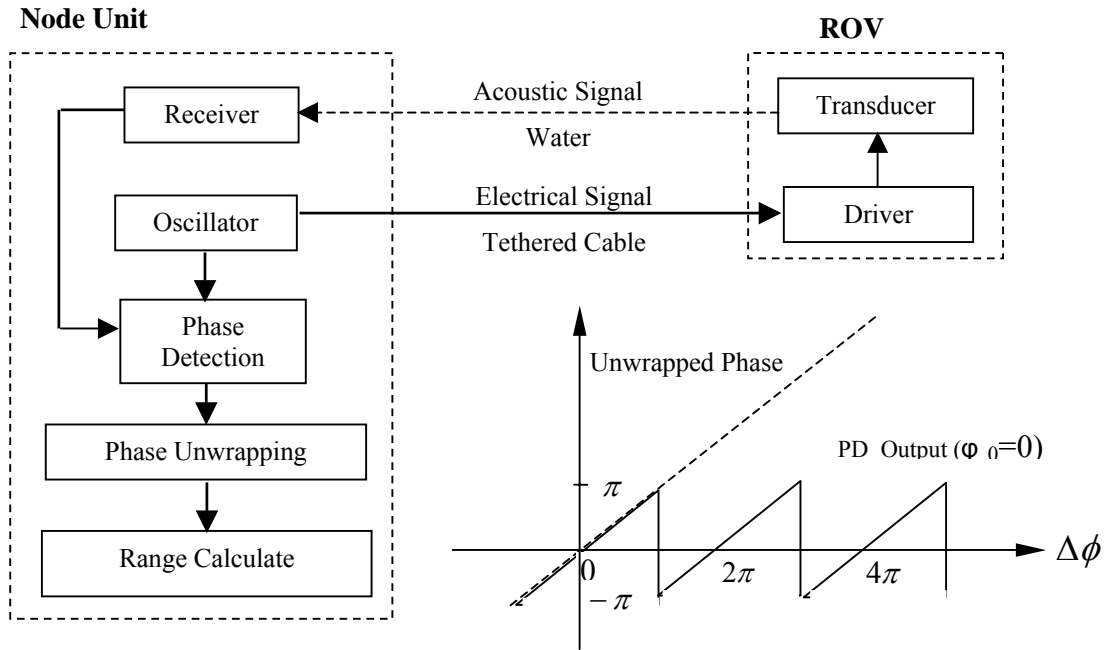


Fig.2 System Flowchart for Range Measurement

Differential phase measurement and unwrapping can also be applied to azimuth and elevation measurements. Let us consider at this point azimuth only as determined by a simple array of two receivers (hydrophones), as shown in Fig. 3.

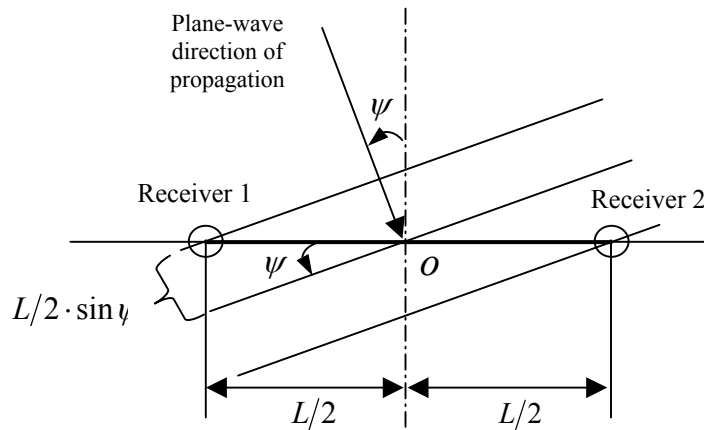


Fig.3 Geometry Relating DOA and Phase of Arrival at Two Receivers

The difference in arrival time of a plane-wave at two receivers is related to the Direction of Arrival (DOA) of the wave-front relative to the receivers or the azimuth of a distant source. Parameters in Fig. 3 are defined as follows:

- L - spacing between receiver elements (hydrophones),
- ψ - azimuth (DOA),

ϕ_1 - phase of the signal received at the left receiver 1,
 ϕ_2 - phase of the signal received at the right receiver 2.

The time delay τ between receipt of the signal at each receiver is:

$$\tau = \frac{L}{c} \sin \psi \quad (7)$$

and the corresponding differential phase $\Delta\phi$ is

$$\Delta\phi = \phi_1 - \phi_2 = 2\pi \frac{L}{\lambda} \sin \psi . \quad (8)$$

By measuring the differential phase $\Delta\phi$ we can obtain the azimuth ψ , that is:

$$\psi = \arcsin\left(\frac{\lambda}{2\pi L} \Delta\phi\right). \quad (9)$$

For $L \leq \lambda/2$ there is no ambiguity in Eq. (9). This condition is applied in USBL systems. However, system resolution can be increased for larger separation of the receivers. This will result in azimuth ambiguity that can, however, be resolved by unwrapping of phase $\Delta\phi$ according to Eq. (6).

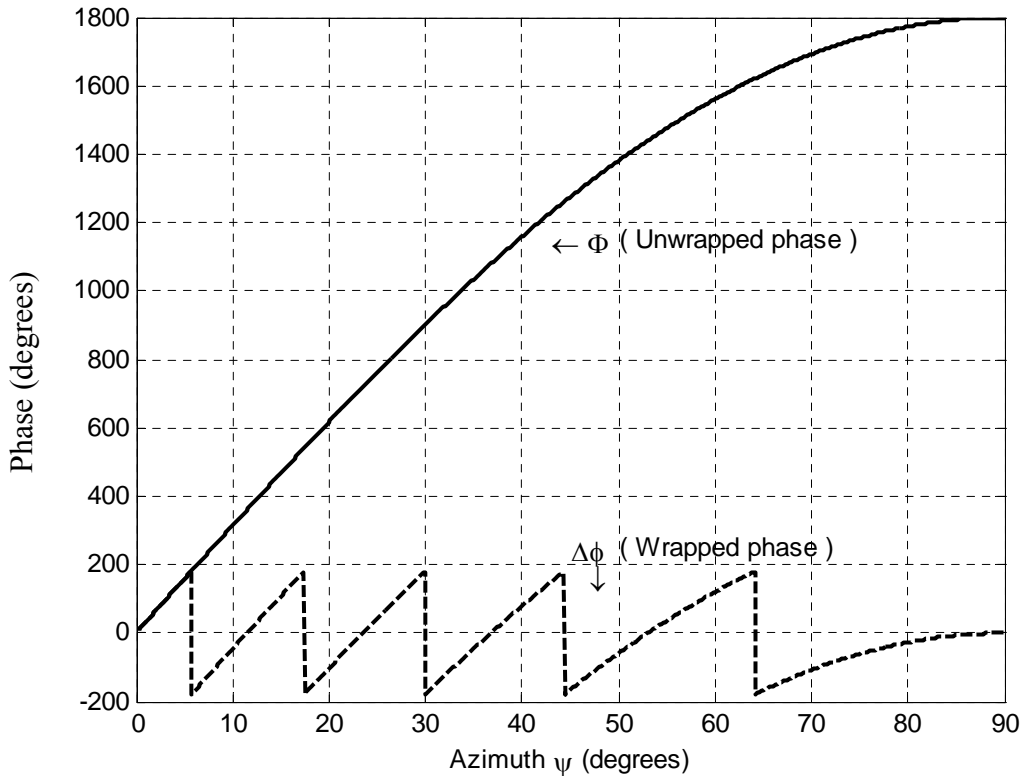


Fig.4 Phase Variations vs Azimuth

Fig.4 shows phase changes as a function of azimuth angle ψ for hydrophone spacing $L = 5\lambda$. We assume that the initial ROV azimuth is 0° (broad-side of the array) and that the azimuth

changes to 90° (end-fire of the array). As expected, the system resolution is best in the broad direction and deteriorates towards the end-fire direction. In other words, the same error in signal phase measurement results in a larger error in azimuth estimation for larger values of azimuth. If the signal is received by a second pair of hydrophones in an array orthogonal to the first, a signal source can be resolved also in elevation, thereby providing the full bearing of the ROV. Several factors will affect a differential phase measurement and, therefore, the range and bearing estimates. Among them are: uncertainty of sound velocity, its temporal and spatial variability, variable multipath and ambient noise. Some of them will have rapid temporal variability and their effects can be reduced by suitable averaging. In this paper, we concentrate only on ambient noise in order to determine theoretical error limits achievable with this method. As we will see later, the ambient noise is unlikely to be a factor limiting system accuracy.

2. QUADRATURE PHASE SENSITIVE DETECTION

Phase measurements can be made by a simple circuit with a zero-crossing detector and suitable logic gates. Such a digital phase detector works well when SNR is high, but has poor performance for low SNR [3]. A phase detector optimum for Gaussian white-band noise was proposed and implemented using digital signal processing [5]. For continuous operation, reduced complexity and for larger bandwidth we propose here a hybrid: analog-digital implementation using a quadrature phase sensitive detector (QPSD) in the context of our navigation system. Such a QPSD offers satisfactory performance in the presence of noise, as we will demonstrate by simulations.

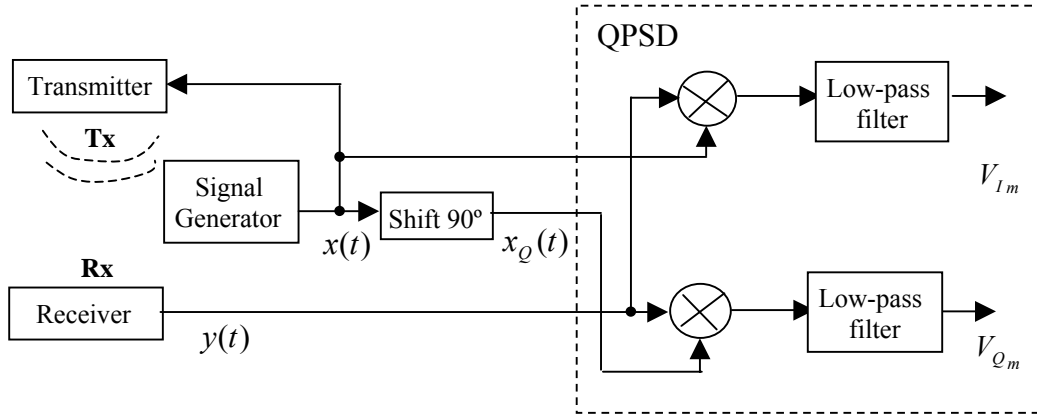


Fig.5 Quadrature Phase Sensitive Detection

Fig. 5 shows the block diagram of the QPSD used here for range measurements, where the reference signal and its quadrature are given by:

$$x(t) = A \cos \omega t \quad (10)$$

$$x_Q(t) = A \sin \omega t . \quad (11)$$

The phase modulated received signal $y(t)$ is:

$$y(t) = B \cos(\omega t + \phi(t)) + n(t) \quad (12)$$

where $n(t)$ is a zero-mean, white Gaussian noise, and $\phi(t) = \phi_0 + \Delta\phi$ is the phase delay of the received signal.

The received signal $y(t)$ and reference signal $x(t)$ are applied to the multiplier whose output V_{I_m} is given by

$$V_{I_m} = y(t) \cdot x(t) \quad (13)$$

from which

$$V_{I_m} = \frac{AB}{2} \cos \phi(t) + \frac{AB}{2} \cos(2\omega t + \phi(t)) + An(t) \cos \omega t. \quad (14)$$

Similarly $y(t)$ multiplied by $x_Q(t)$ yields:

$$V_{Q_m} = \frac{AB}{2} \sin \phi(t) + \frac{AB}{2} \sin(2\omega t + \phi(t)) + An(t) \sin \omega t. \quad (15)$$

The first terms in Eqs. (14) and (15) represent slow-varying signals (base-band signals) that can be separated from higher frequency components by suitable analog or digital low pass filters resulting in:

$$V_{I_m} \approx \frac{AB}{2} \cos \phi(t) \quad (16)$$

and

$$V_{Q_m} \approx -\frac{AB}{2} \sin \phi(t). \quad (17)$$

Consequently, the phase difference is calculated as:

$$\phi = \tan^{-1} \frac{\sin \phi}{\cos \phi} = \tan^{-1} \left(-\frac{V_{Q_m}}{V_{I_m}} \right). \quad (18)$$

3. SIMULATION RESULTS

Fig. 6 shows the block diagram of the ranging system using QPSD. It consists of a signal generator and front-end amplifier followed by a band-pass filter, QPSD and a processing unit. The signal generator drives an acoustic hydrophone and provides reference signals for QPSD. At the receiver side, after amplification and filtering, the input signal from the hydrophone is fed to QPSD circuits. The function of the front-end band-pass filter is to reject out-of-band noise. The filter center frequency is equal to the carrier frequency so that the signal is not affected. The bandwidth B of the filter must be large enough to accommodate any Doppler shifts due to ROV movements. In this paper, we assume $B/f_0 = 0.1$. Outputs of QPSD consist of two base-band signals proportional to $\cos \phi(t)$ and $\sin \phi(t)$, respectively. We assume here that low-pass filters have sufficient bandwidth not to affect the low pass signals of interest.

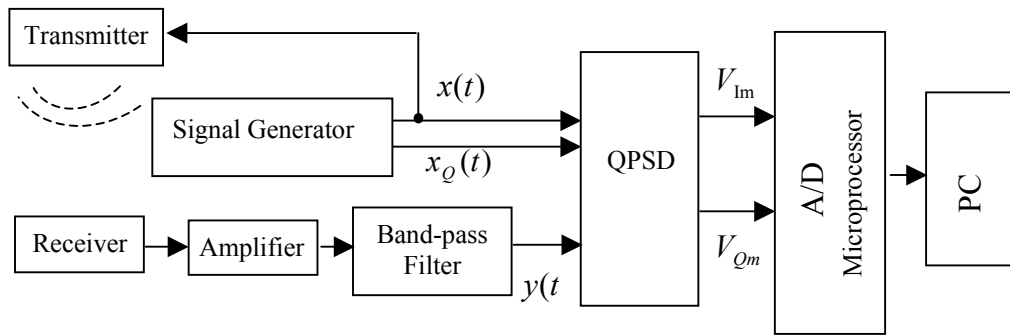
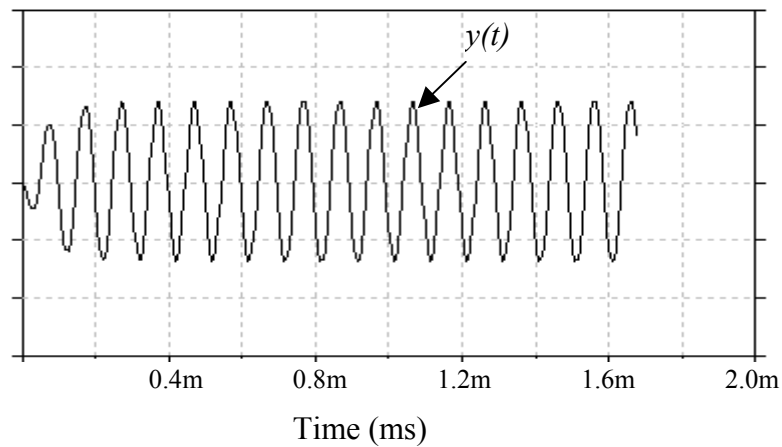
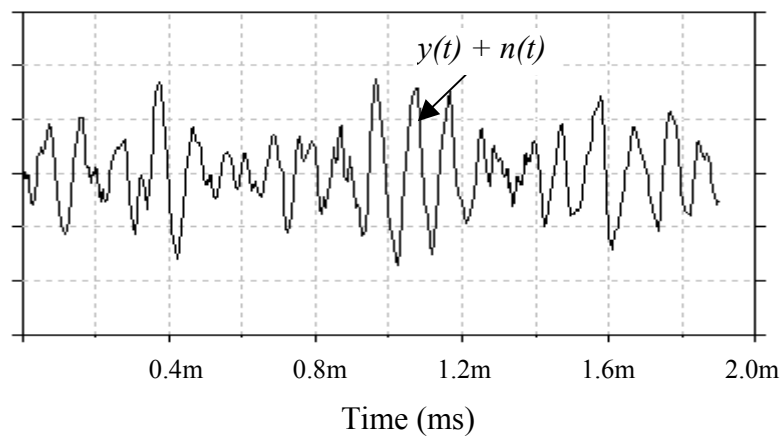


Fig.6 Block Diagram of the Ranging System

After performing the A/D conversion, the microcontroller calculates the phase difference ϕ using Eq. (18) and performs phase unwrapping. Finally, the PC will display the results.



(a)



(b)

Fig.7 (a) Signal and (b) Noisy Signal with SNR of 0 dB

We investigate the performance of the proposed method by simulations. The simulation was performed by combining the use of Matlab and Multisim software. First, analog circuits including the signal generator, filters, amplifier and QPSD were simulated in Multisim. Then the simulation results from Multisim were imported into Matlab through Excel. The data processing, including phase unwrapping and calculating the range was performed using Matlab. Different SNR levels were used in the simulations by adding white noise to received signals. The transmitted assumed frequency was $f = 10$ kHz and velocity $c = 1500$ m/s. Fig. 7a shows the noise-free received signal and Fig. 7b shows the noisy signal with SNR = 0 dB.

Fig. 8 shows simulation results for phase measurement with SNR of 0 dB and an unwrapping phase using Matlab. The estimated unwrapped phase and actual phase are indistinguishable because of the small phase error. Fig. 9 shows the root-mean-square (RMS) phase difference error versus the signal to noise ratio (SNR) varying from -10 dB to 20 dB. As we see, the RMS error of 14° in the phase difference estimate was achieved even with a poor SNR = 0 dB. This represents an equivalent range error of $(14^\circ/360^\circ)\lambda = 5.8$ mm for carrier frequency $f = 10$ kHz with corresponding wavelength $\lambda = 15$ cm in water. This range estimation accuracy outperforms all existing commercial USBL systems [2], but should be viewed as an upper theoretical limit only.

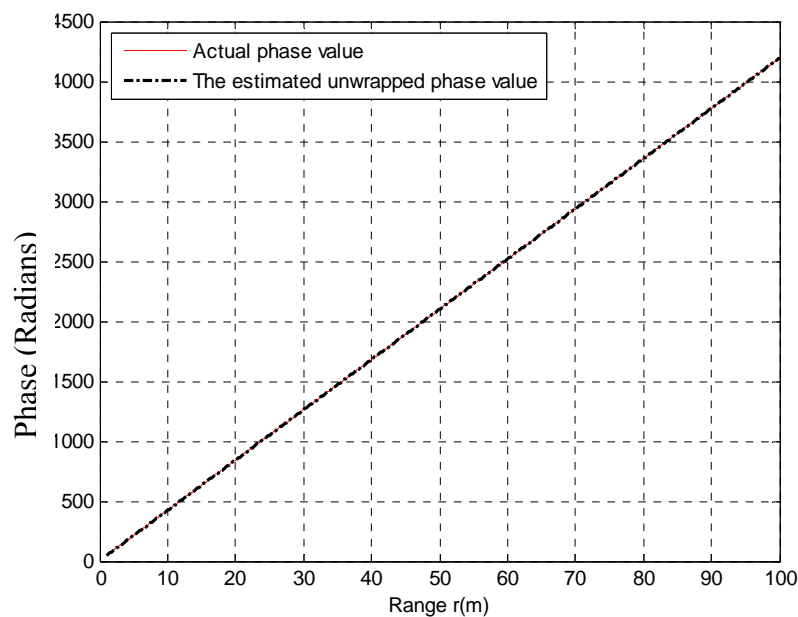


Fig.8 Unwrapped Phase with SNR of 0 dB

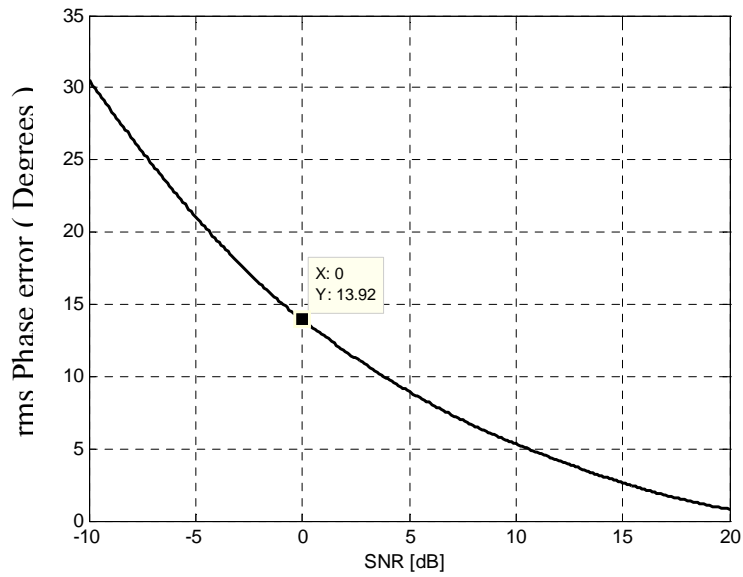


Fig.9 Variation of RMS phase Errors vs SNR

We can use QPSD and phase unwrapping also for target azimuth and elevation estimation. Fig.10 shows a schematic diagram of phase difference measurement for azimuth estimation. The phase difference $\Delta\phi$ is obtained in two steps. First, we measure the phase difference ϕ_1 and ϕ_2 with respect to the reference signal $x(t)$ and then calculate $\Delta\phi = \phi_1 - \phi_2$.

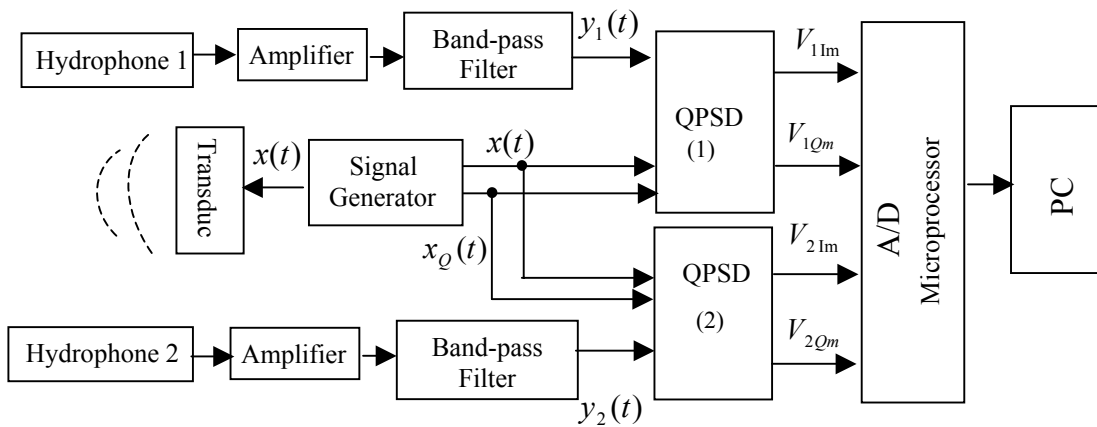


Fig.10 Quadrature Phase Sensitive Detection with Reference Signal

The advantage of this approach is that the reference signal sent through the tether cable is noise-free and only received signals are contaminated by the noise.

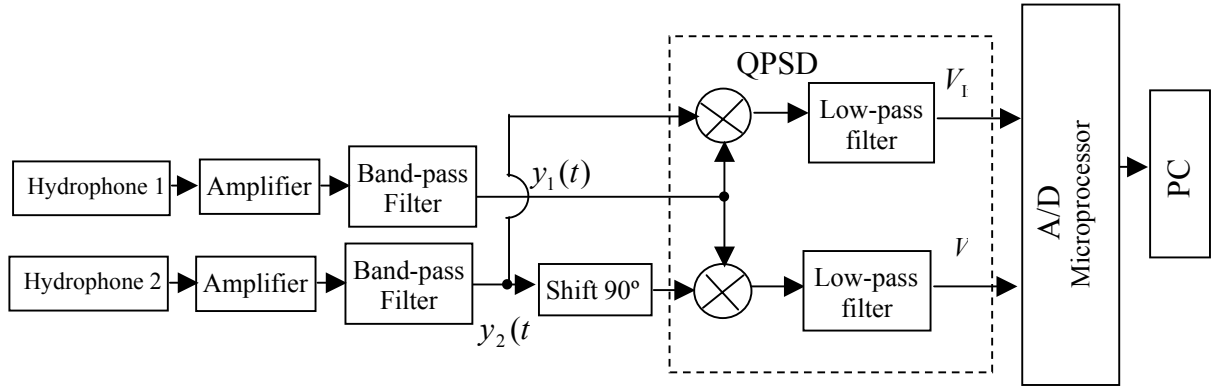


Fig.11 Quadrature Phase Sensitive Detection without Reference Signal

With the arrangement shown in Fig. 11, the phase difference $\Delta\phi$ can be measured directly without using a reference signal. The circuit is simpler, but leads to a larger phase error since both signals applied to the QPSD are contaminated by noise. For this case, the RMS phase difference error of 17° was obtained with the SNR of 0 dB. This error represents an equivalent angle error of $(17^\circ/\pi) \approx 5.4^\circ$ with the hydrophone spacing $L = \lambda/2$ and equivalent angle error of $(17^\circ/10\pi) \approx 0.5^\circ$ with the spacing $L = 5\lambda$. The important advantage, however, is that the reference signal is not required. This means that this method can be applied to trace the bearing of an autonomous underwater vehicle that does not require a tether cable.

4. CONCLUSIONS

An acoustic positioning method based on continuous phase measurement using quadrature phase detection and phase unwrapping has been described. Different SNR levels were used in simulations. High precision range estimates with theoretical RMS error of 5.8 mm and bearing estimate of 0.5° at frequency $f = 10$ kHz were achieved even with a poor SNR of 0 dB. Results show the potentially extremely high precision of the system both in range and in bearing estimations. These results indicate that actual system performance will be limited by other factors such as uncertainty, uniformity and variability of sound velocity and multipath.

5. APPENDIX : PHASE UNWRAPPING

In practice, the phase will be wrapped between $-\pi$ and π [6]. In a discrete-time sampled data system the received data can expressed as

$$y[n] = A \cos[\omega_s n + \phi] + w[n], \quad (1)$$

where the digital frequency $\omega_s = \omega T_s$, T_s is sampling frequency, $w[n]$ is white noise, phase difference $\phi = \frac{2\pi \cdot r}{\lambda}$ and r is the location of the receiver with reference to the transmitter and

λ is wavelength of the transmitted carrier. When $r > \lambda \Rightarrow \phi > 2\pi$, the phase difference can be registered only modulo 2π . The technique for phase unwrapping is to search the phase sequentially for jumps in phase greater than π , the assumption being that the phase changes at a rate slower than π radians per sample. These jumps are then corrected by adding a factor of 2π to all subsequent terms in the sequence. If phase ϕ_n is wrapped and phase Φ_n is unwrapped, we have:

$$\Phi_n = \phi_n + 2\pi k . \quad (2)$$

A straightforward method of phase unwrapping is to check the difference between the estimated ϕ_n and the last phase ϕ_{n-1} and then update k at each phase ϕ_n discontinuity. However, this method may fail to avoid sporadic false jump of phase due to noise in the system. On the other hand, based on continuity of unwrapped phase, the best estimate of k can be obtained by minimizing:

$$e(k) = (\Phi_{n-1} - (\phi_n + 2\pi k))^2, \quad k = M1, \dots, M2 \quad (3)$$

where $M1$ and $M2$ are determined by the initial and maximum phase difference between transmitted signal and received signal.

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