Development of a Sonar for Underwater Sensor Platforms and Surface Vehicles

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Abstract

In the frame of a thesis project the authors have developed software and methodologies for a sonar prototype. The ultimate goal is the installation of the sonar prototype on the department’s sensor platforms, which include AUVs and ASVs.

The thesis presents the intelligent sampling algorithms and DSP tools, e.g. a matched filter, that have been implemented. The sonar was successfully configured to work as an echo sounder. Tests were performed in the underwater laboratory at FOI and in-situ in the Stockholm archipelago. It was found that the sonar is capable of measuring distances with an average error of less than 1 cm in controlled experiments. The in-situ experiments in Baggensfjärden showed a echo detection rate of $\approx 90\%$.

Unfortunately, due to hardware limitations, attempts to implement DVL functions have failed. However, background theory and methodologies for velocity measurements, communication and positioning are briefly explained to provide a basis for future work.

The electronics currently are subject to improvements and a new version is expected to be ready for use soon. Future work will include analysis and integration of velocity sensing, communication protocols and positioning concepts.
B Hardware & Software

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1 Introduction

1.1 Sonar and Radar

In order for a vehicle to navigate safely in an unknown environment it needs to collect information about its position and its surroundings, i.e. possible obstacles, terrain etc. For aerial and ground vehicles, Radio Detecting and Ranging (radar) devices have been proved to be useful. However, radio waves are electromagnetic radiation which is usually damped to dissipation after approximately $1.5$ wave lengths due to the electrical conductivity of (salt) water (Lurton, 2010). This means that radio waves only penetrate into sufficiently deep water depths at (from an antenna design point of view) unfeasible long wave lengths. As a result, Sound Navigation and Ranging (sonar) devices have been developed in order to face the challenge of underwater ranging, communication and sensing. In contrast to radio waves, acoustic waves propagate well in the oceans, in extreme cases they can cover distances of hundreds and thousands of kilometers. However, sound waves in water travel at significantly lower speeds ($\approx 1480 \text{ m/s}$) than electromagnetic waves ($299,792,458 \text{ m/s}$, speed of light). This results in a relatively high latency. Typical frequencies are in the range $10 \text{ Hz}$ to $1 \text{ MHz}$.

1.2 What is a Sonar?

Sonar systems can be both, passive (only receiving) or active (transmitting and receiving), while most are active systems. The principal design layout of sonars, see Fig. 1.1, comprises a control unit usually equipped with a display, a transmitter, a transducer and a receiver. The transducer converts electric energy into acoustic energy and vice versa via piezo-electric elements (or similar elements). The transmitter provides and oscillating electric field (e.g. square waves) at certain frequencies. The receiver amplifies the electric signal and converts the analog signal into processable data. The control unit is responsible for data processing and filtering. While there usually is only one control unit, the transducers may be arranged as a phased array in order to achieve beam forming.

![Figure 1.1: Schematic design of sonar devices](image)
1.3 Application of Sonars

There is a wide variety of application for sonars: Military, fishing, oil & gas, oceanography, archaeology. In military applications, they are used to identify and locate enemy underwater vehicles such as submarines and also to spot underwater mines or other potential threats. In the fishing industry sonars are used to find shoals of fish. In the oil & gas industry and in marine archaeology they are usually used to provide acoustic images of the sea floor, and in oceanography the water properties are investigated acoustically.

Depending on the application different types of sonars are used. The most basic sonar is used for ranging purposes, i.e. measuring a distance between the sonar platform and the target (Downward Looking Sonar (DLS), Upward Looking Sonar (ULS) and Forward Looking Sonar (FLS)). For acoustic imaging Single-beam Echo Sounders (SBESs), Multi-beam Echo Sounders (MBESs), Side-scan Sonars (SSSs), Synthetic Aperture Sonars (SASs) and sub-bottom profilers are available. Doppler Velocity Logs (DVLs) and Acoustic Doppler Current Profilers (ADCPs) provide information of the relative motion between the platform and the acoustic target which can either be the sea floor or the water column itself. In the latter case ocean currents are analyzed. For wireless underwater communication acoustic modems can be used. However, they only offer limited data rates and ranges with comparably high latency (Sendra et al., 2016).

Sonar devices are typically mounted on underwater vehicles and surface vessels. For the latter sonars usually provide information on water depth and potential obstacles but they indeed are also used for tracking of fish and seismic exploration. The second - and probably largest - group of carriers are underwater vehicles such as Autonomous Underwater Vehicles (AUVs) and Remotely-operated Vehicles (ROVs) but also towed vehicles and buoys.

1.4 Project Scope

The goal of this thesis project is to develop and test sonar devices for subsequent use at the Centre of Naval Architecture at Kungliga Tekniska Högskolan (engl. The Royal Institute of Technology) (KTH). At KTH, miniature AUVs named Maribot are developed and continuously improved. Part of this improvement is the development of sonar devices. This thesis project deals with Digital Signal Processing (DSP) for sonar devices and software implementation, which are tailored to the Maribot’s system specifications. The scope eventually is to explore methodology and software solutions for the following sonar applications in medium water depths from 0 m to 1000 m:

1. Ranging
2. Velocity measurements
3. Two-node duplex communication
4. Long Base-line (LBL) positioning

Furthermore, it is intended to provide a solid foundation of knowledge in underwater acoustics in order to understand how the platform works and how the sonar’s performance
is influenced by the environment. Understanding of environmental parameters and sound propagation underwater are of great importance in order to track down errors and interpret ocean acoustical data from sonar devices.
2 Theory

2.1 Underwater Acoustics

Acoustic waves are mechanical waves caused by a deviation from the average of the ambient atmospheric pressure which eventually leads to compression and rarefaction of the medium. Since fluids cannot transfer shear stresses acoustic waves only propagate as longitudinal waves. They are characterized by:

1. Wave period $T$, $[T] = s$
2. Wave frequency $f = \frac{1}{T}$, $[f] = \text{Hz}$
3. Speed of sound $c$, $[c] = \frac{\text{m}}{\text{s}}$
4. Wavelength $\lambda = \frac{c}{f}$, $[\lambda] = \text{m}$

Underwater sound propagation is influenced by pressure $p$, salinity $\sigma$ and density $\rho$ of the water.

Wave equation

The wave equation in the three dimensional can be derived from fluid mechanics as

$$\Delta p = \frac{\partial^2 p}{\partial x^2} + \frac{\partial^2 p}{\partial y^2} + \frac{\partial^2 p}{\partial z^2} = \frac{1}{c^2(x, y, z)} \frac{\partial^2 p}{\partial t^2}$$

In this equation, $p$ denotes the acoustic pressure in the spatial domain $x, y, z$ and time $t$. $c$ is the speed of sound and $\Delta$ the Laplace operator. For sinusoidal waves this equation can be expressed as the Helmholtz equation with $c = \text{const.}$ and only one-dimensional it becomes

$$\frac{\partial^2 p}{\partial x^2} + \frac{\omega^2}{c^2} p = 0$$

where $\omega = 2\pi f_0$ is the circular frequency. The text book solution to this differential equation is

$$p(x, t) = p_0 \exp(j(\omega t - kx))$$

However, this solution is valid only for plane waves in $x$-direction. For spherical waves traveling in a three-dimensional space (polar) the solution is

$$p(R, t) = \frac{p_0}{R} \exp(j(\omega t - kR))$$

where $R$ is the radial distance to the point-source, $k = \frac{\omega}{c}$ is the wave number. $p0$ is the the pressure 1 m away from the source (as per convention). As one can see, the decrease
in acoustic pressure is inversely proportional to $R$. Plane waves are easier to model than spherical waves. Only for sufficiently large $R$ a wave front may be considered to be plane due to the low curvature (Lurton, 2010).

**Sound Intensity**

The acoustic power or sound intensity is defined as (F. Jensen et al., 2011; Lurton, 2010)

$$I = \frac{p_0^2}{2\rho c}$$  \hspace{1cm} (2.1)

where $p_0$ is the plane wave amplitude. Since $p_{RMS} = \frac{p_0}{\sqrt{2}}$, Eq. 2.1 can also be expressed as

$$I = \frac{p_{RMS}^2}{c\rho}$$  \hspace{1cm} (2.2)

where $p_{RMS}$ is the root mean square of the sound pressure and $u$ is the water particle velocity. The product $c \cdot \rho$ is known as the characteristic impedance $Z$.

The Sound Pressure Level (SPL) uses a logarithmic scale since it is more feasible. SPL is defined as

$$L_P = 20 \log\left(\frac{p}{p_{ref}}\right)$$  \hspace{1cm} (2.3)

As per convention, the reference sound pressure $p_{ref}$ underwater is $p_{ref} = 1\,\mu Pa$. The Sound Intensity Level (SIL) analogously is defined as

$$L_I = 10 \log\left(\frac{I}{I_{ref}}\right)$$  \hspace{1cm} (2.4)

where the reference sound intensity $I_{ref}$ can be calculated from

$$I_{ref} = \frac{p_{ref}^2}{\rho \cdot c}$$  \hspace{1cm} (2.5)

The reference distance $r_0$ typically is 1 m.

**Speed of Sound**

The speed of sound in water depends on three factors (Fofonoff et al., 1983):

1. Salinity $\sigma$
2. Pressure $p$ and thus depth $d$
3. Temperature $T$

From a physical point of view, the speed of sound is given as $c = \frac{E}{\rho}$. Medwin (1975) has derived a simple empirical expression for the approximation of speed of sound underwater:

$$c = 1449.2 + 4.6T - 0.055T^2 + 0.00029T^3 + (1.34 - 0.01T)(S - 35) + 0.016z$$  \hspace{1cm} (2.6)
One can see that $c$ varies linearly with depth and non-linearly with salinity and temperature. The speed of sound varies within the water column (speed gradient). The speed of sound profile often takes a $c$-shape and can also lead to the creation of shadows zones with low sound intensity and to caustics, area of high sound intensity. The speed of sound in salt water usually is the range of about $\approx 1450 \text{ m/s}$ to $1540 \text{ m/s}$. At water depths below 1000 m water temperature and salinity remain about constant and sound speed increases almost exclusively with depth due to an increase in hydrostatic pressure.

**Underwater waveguides**

The direction of sound propagation is determined by the gradient of the speed of sound. Locally, the speed of sound is changing most importantly with depth. Acoustics rays will always pass regions of varying sound speeds within the water column. They are refracted towards regions with lower speed of sound and away from regions with higher sound speeds.

Under certain circumstances, the presence of a sound speed minimum can lead to waveguide propagation. There is a number of different waveguides:

1. Underwater Sound Channel (USC), also Deep Sound Channel (DSC) or Sound Fixing and Ranging (SOFAR) channel
2. Surface Sound Channel (SSC)
3. Shallow Water Sound Channel (SWSC)

**Underwater sound channel**

The axis of the DSC is located at water depth $z_m$, which is the depth at which the sound speed profile has its minimum. Above $z_m$ the sound speed mostly increases due to an increase in temperature and below $z_m$ it increases mostly due to an increase in hydrostatic pressure. Depending on the location of $z_m$ and the sound speed gradient two different kinds of the DSC are distinguished. In the following explanation $c_0$ denotes the the speed of sound at depth 0 m (sea surface) and $c_h$ denotes the speed of sound at depth $h$ (seabed). The depth $z_c$, denotes either the depth at which the near-bottom sound speed reaches the magnitude of $c_0$ (if $c_0 < c_h$) or the depth at which the near-surface sound speed reaches the magnitude of $c_h$ (if $c_0 > c_h$).

The USC of the first kind occurs when $c_0 < c_h$ and is valid for sound rays that leave the source at grazing angles $\epsilon \leq \sqrt{\frac{2(c_0-c_m)}{c_m}}$. If the sound source is located close to $z_m$ sound rays are trapped within the USC and are protected from bottom reverberations. See Fig. 2.1.

The USC of the second kind occurs when $c_0 > c_h$ and allows the sound rays to reach the seabed (and consequently be backscattered) but they are prevented from reaching the sea surface. This kind of USC is more frequent in shallower water. See Fig. 2.2.

A special case of the USC is the USC with two axes. This kind occurs if there is a local sound speed maximum present between the sea surface and depth $z_m$. This usually happens when different water masses are mixing and leads to formation of DSC and SSC simultaneously.
In the USC acoustic waves can propagate over thousands of kilometers. The position of the USC at moderate latitudes is several hundred meters below the sea surface, down to 2000 m in tropical areas and rather close to the surface in polar regions (Kraus, 2016).

**Surface sound channel**

If there is no minimum in the sound speed profile but a maximum at depth \( z = h \), a SSC can occur. If the sound source is location between the surface and depth \( h \) and the grazing angles do not exceed a limit angle, the sound rays are trapped in the surface channel. No bottom reverberations are present. See Fig. 2.3.

**Shallow water sound channel**

In shallow waters sound rays can reach sea surface as well as seabed and be backscattered by both. This gives a wave guide with maximum transmission losses, see Fig. 2.4.
Antiwaveguide propagation

If the sound speed profile is monotonically decreasing sound rays are strictly refracted towards the seabed. In such case shadow zones are created which are out of reach for acoustic devices. The sound intensity in these regions is very low but not zero due to diffraction, scattering and reflections, see Fig. 2.5.

Geometrical spreading loss

The sound intensity decreases with increasing distance $R$. This is due to the fact that the sound pressure is acting on an ever increasing surface (spreading). At short ranges, geometrical spreading loss is the dominating constituent of all losses. The geometrical spreading loss for two-way spherical propagation (active sonars) is inversely proportional to the distance $R$

$$I \propto \frac{1}{R^4}$$

and for one-way spherical propagation (passive sonars) it is

$$I \propto \frac{1}{R^2}$$

Under certain circumstances, e.g. waveguide propagation, sound can be assumed to spread cylindrically - not spherically - in the far field and the proportionality for one way propagation then becomes

$$I \propto \frac{1}{R}$$
for passive sonars and

\[ I \propto \frac{1}{R^2} \]

for active sonars (Hansen, 2012). Fig. 2.6 shows how the area is increasing with increasing distance.

![Figure 2.6: Geometrical Spreading Loss, from www.sfu.ca](image)

**Reverberation**

The acoustic rays are reflected and refracted according to Snell’s law. A good explanation of Snell’s law can be found even on [www.wikipedia.org](http://www.wikipedia.org). The relationship between the angles are

\[ k = k_{P1} \cdot \sin(\theta_{P1}) = k_{S1} \cdot \sin(\theta_{S1}) = k_{P2} \cdot \sin(\theta_{P2}) = k_{S2} \cdot \sin(\theta_{S2}). \]

\( P_i \) is the incident pressure wave, that is either transmitted (index \( t \)) or reflected (index \( r \)) as a shear wave (\( PS \)) or a pressure wave (\( PP \)). As one can see, sound waves are refracted towards regions with low sound speed and away from regions with high sound speeds. Reflections often lead to multipath propagation, i.e. the transmitted sound wave is reflected on different surfaces and reaches the receiver of a number of different paths, leading to a number of undesired echoes of the transmitted signal. Depending on the surface roughness and the incident angle, reradiated sound waves are scattered more (rough surfaces) or less diffusely (smooth surfaces).

![Figure 2.7: Illustration of Snell’s law, source: www.wikipedia.org](image)
Surface backscattering
The difference in characteristic impedance between sea water \( (Z_{\text{water}} \approx 1\,540,000\, \text{Pa s m}^{-3}) \) and air \( (Z_{\text{air}} \approx 415\, \text{Pa s m}^{-3}) \) makes the sea surface a fairly good reflector of acoustic waves (Hansen, 2012). The amount of energy crossing the air-sea boundary is very small. When acoustic waves are reflected from the sea surface they are phase-shifted by \( \pi \text{ rad} \) (Etter, 1995). A rough sea surface due to waves or heavy rainfall causes the acoustic waves to be scattered, as well. The backscattering furthermore is assumed to be influenced by grazing angle and frequency (Kraus, 2016).

Bottom backscattering
Depending on the acoustic characteristics of the sea bottom (roughness, impedance etc.) acoustics rays are also back scattered from the sea bed. Bottom backscattering is strongly dependent on the grazing angle and frequency. Low frequencies can penetrate much deeper into the sea bed than high frequencies.

Volume backscattering
In the water column sound can be scattered from marine life and ocean fluctuations such as bubbles and ocean turbulence. How acoustic rays are backscattered within the water column depends on the target strength (a function of size, shape and impedance) and also on the angle of incidence.

Attenuation of Sound
Salt water as a medium is dissipative. Due to viscosity of salt water (above 100 kHz), relaxation (i.e. the conversion of acoustic energy into heat) of magnesium sulphate (above 100 kHz or 10 kHz to 100 kHz) and boric acid (above 1 kHz or \( \leq 10\, \text{kHz} \)) sound waves are attenuated (Lurton, 2010; Kraus, 2016). There are numerous empirical models available for the prediction of attenuation of underwater sound, e.g. Thorp, Schulkin-Marsh and Francois-Garrison Urick (1983) and Kraus (2016). The absorption of low frequency sound is comparably small. However, the magnitude of absorption increases with increasing frequency, see Fig. 2.8. In contrast to the sea surface, sound waves can penetrate into soft sea floors and consequently be absorbed significantly. Since the sea bed is a solid, also shear waves can develop. At long distances, losses in acoustic sound intensity result from attenuation and scattering losses. Sound attenuation in underwater environments is dependent on temperature, salinity, pressure and pH. If sound is propagating into the sea bed, it also is attenuated as a function of sediment type.

Noise
There are several noise sources in underwater acoustics. Noise can either be generated by the platform itself and by its sensors (self noise), or come from the environment (ambient noise). Ambient noise is that fracture of the received noise which is totally independent of noise generated by either the transmitting or receiving platform. The ocean itself is a noisy environment, human activities on and in the oceans even increase the noise levels.
All noise is superimposed onto the transmitted signal and eventually received together with the echo of the transmitted signal.

**Ambient noise**

In general, the sound intensity of ambient noise decreases with increasing frequency - a characteristic very favorable for the application of sonars. Background noise is introduced by a wide variety of sources (F. Jensen et al., 2011; Lurton, 2010; Havelock et al., 2008; Kraus, 2016).

1. In the lower range of 0.1 Hz to 20 Hz ocean turbulence and microseisms are possible sources
2. Ship traffic causes noise at frequencies around 20 Hz to 500 Hz
3. Wind-induced surface noise occurs at frequencies ranging from 0.3 kHz to 100 kHz
4. Thermal noise is present above 50 kHz (alternatively above 50 kHz or from 100 kHz to 1000 kHz)
5. Rainfall noise around 1 kHz to 5 kHz

There, however, are numerous more sound sources, some occurring rarely and some are almost omnipresent: Earthquakes, underwater volcanoes, biological activities etc.

### 2.2 Sound Navigation and Ranging

In this section it is described how sonars make use of the underlying physics described in the previous section.

**Piezoelectric Transducers**

Piezoelectric transducers convert electric energy into mechanical deformation and vice versa via piezo-electric ceramics. By applying an alternating electric potential on both ends...
of a transducer it can be excited to oscillate periodically. The mechanical deformations lead to pressure variations in the surrounding medium which eventually travel as sound waves.

**Piezoelectric ceramics**

The performance of transducers is related to the size and material properties. The bandwidth, resonance frequency and beam width are the three most important acoustic parameters. The bandwidth and resonance frequency can be estimated from the electro-mechanical property of the material, and are based on the principle of electro-mechanical effect. The beam width can be determined from the directivity pattern if frequency and transducer size are known, see Sec. 2.2. To avoid a reduction in directivity due to a large main lobe, the diameter of a circular transducer should be $D > 2\lambda$.

**Frequency and Bandwidth**

Electromechanical effect is the fundamental mechanism of piezoelectric transducers. Both, the material properties and the transducer’s dimensions influence the resonance frequency, bandwidth and sensitivity (transmit power) of the transducer. All these parameters are coupled. Because of their complexity, idealization and an appropriate oscillation model are required when analyzing the performance of the transducer. The idealization, see Fig. 2.9, of the transducer consists of the deformation model, the mechanical oscillation model, which describes the relationship between displacement of the piezoelectric material and thus the acoustic pressure, and the electrical equivalent circuit model that illustrates how the electrical signal and the displacement of the material are coupled. The details of these idealized models are described in numerous publications (H. Jensen, 1986; Sherman et al., 2013). The performance of the transducer can be estimated from idealized models with the Rayleigh-Ritz method or from Finite Element Method (FEM).

**Beam width**

The beam width of a transducer can be determined from the directivity pattern as described in Sec. 2.2. The simplest approximation for disk-shaped transducers is given by $\sin(\theta) = \frac{\lambda}{d}$ with $d$ as transducer diameter, $\lambda$ as the wave length of the sound wave (derived from frequency) and $\theta$ as the beam angle. In general, the larger the transducer diameter and the larger the frequency, the narrower the beam width.

**Directivity pattern and beam width**

The beam width, expressed as an angle, is defined at the point where the main lobe of the acoustic field is reduced by 3 dB from maximum intensity. The over all relationship between beam width, frequency and transducer size can be approximated as

$$\sin(\theta) \approx \frac{\lambda}{D} = \frac{c}{Df} \quad (2.7)$$
Deformation Model

\[ S = s^2T + d^2E \]
\[ D = dT + \varepsilon T E \]

Mechanical Model

\[ M\ddot{x} = -K_m\dot{x} - R\dot{x} + F + N_{em} V \]

Electrical Equivalent Circuit Model

\[ j\omega(M + M_r)u + (R + R_r)u + u/j\omega C_m^E = uZ_{mr}^E = NV + F_b \]

**Figure 2.9:** Electric Equivalent Model of a transducer

**Figure 2.10:** Examples: The influence of transducer diameter and frequency on beam width
According to Hansen (2012), for a disc array, it may be approximated as (in degrees)

$$\theta_{3dB} = \pm \frac{29.5\lambda}{D} = \pm \frac{29.5c}{Df} \quad (2.8)$$

For line arrays it is approximated as (in degrees)

$$\theta_{3dB} = \pm \frac{25.3\lambda}{L} = \pm \frac{25.3c}{Lf} \quad (2.9)$$

In general, higher frequencies result in narrower beams while lower frequencies result in wider beams. The directivity of a transducer expresses, when transmitting, the angular distribution of the acoustic energy radiated into the propagation medium and, when receiving, the electric response of the transducer as a function of the direction of arrival of the acoustic wave. The directivity pattern describes the angular distribution of energy as a function of the shape and size of the transducer and the frequency of the acoustic wave. For disk arrays the directivity pattern is

$$b(\theta) = \left[ \frac{2J_1\left(\frac{1}{2}kD \sin \theta\right)}{\frac{1}{2}kD \sin \theta} \right]^2 \quad (2.10)$$

For line arrays the directivity pattern is

$$b(\theta) = \left[ \frac{\sin(\frac{1}{2}kL \sin \theta)}{\frac{1}{2}kL \sin \theta} \right]^2 \quad (2.11)$$

Waveforms

The two basic waveforms used by sonars are

1. Continuous Waveform (CW) (narrowband)
2. Frequency-modulated Waveform (FM) Waveforms/Chirps (broadband)
   - Linear FM (LFM)
   - Hyperbolic (HFM)
   - Doppler Sensitive (DFM)

CW is a sinusoidal waveform with constant frequency. Mathematically it is expressed by

$$s(t) = A_0(t) \sin (2\pi f_0 t + \phi_0(t)) \quad (2.12)$$

where $A_0(t)$ is the amplitude with the time-dependency indicating the possibility of having envelope functions (e.g. Gaussian envelope). $f_0$ is the signal frequency, $t$ the time and $\phi_0(t)$ is the phase. The phase is often assumed to be 0 throughout the duration of the pulse while - in the simplest case. The envelope of the transmitted acoustic signal is determined by the electronic field that excites the transducer. The returning echo $r(t)$ will be altered in amplitude ($A_1(t)$), frequency ($f_1$) and phase ($\phi_1(t)$). Ambient noise additionally superimposes noise $n(t)$ on the signal

$$r(t) = A_1(t) \sin (2\pi f_1 t + \phi_1(t)) + n(t) \quad (2.13)$$
However, in order for a CW to be detectable it needs to carry sufficient acoustic energy. This can either be achieved by increasing the transmitter power or by lengthening of the pulses. Both solutions are limited, especially since a longer pulse significantly reduces the range resolution. Low-level multi-path echoes are likely to interfere with the direct path echo before it has been received completely. Unlike CWs, FMs can use the full bandwidth of the sonar which in turn allows to use longer pulses, if necessary. FMs are sinusoidal waveforms with a characteristically changing frequency (spread spectrum, broadband). The most common type FM today is the chirp. The advantage of FMs are an increased range detection capability. They however are disadvantageous to CWs when it comes to Doppler shift detection. Mathematically, up-sweeping (up chirp) and down-sweeping (down chirp) FMs are expressed as

\[
\begin{align*}
    s_u(t) &= A \cos \left[ 2\pi t \left( f_{\text{min}} + \frac{f_{\text{max}} - f_{\text{min}}}{2T} t \right) \right] \\
    s_d(t) &= A \cos \left[ 2\pi t \left( f_{\text{max}} + \frac{f_{\text{max}} - f_{\text{min}}}{2T} t \right) \right]
\end{align*}
\]

(2.14) and

(2.15) respectively. Here, \( A \) is the amplitude, \( t \) is the time, \( f_{\text{min}} \) and \( f_{\text{max}} \) denote the minimum and maximum frequencies of the spread spectrum and \( T \) is the chirp length. FMs have been presented already in the early 1990s by Austin (1994).

**Sonar Equation**

The sonar equation can be used to model/predict the performance of sonars. It helps to identify the required magnitude of the Detection Threshold (DT). The equation consists of the following parameters, all of them are expressed in decibels (dB):

1. Source Level (SL)
2. Transmission Loss (TL)
3. Noise Level (NL)
4. Directivity Index (DI)
5. Target Strength (TS)
6. Processing Gain (PG)

For passive sonars the equation is simpler since not all of the above parameters are used

\[
\text{DT} < \text{SL} - \text{TL} - (\text{NL} - \text{DI})
\]

(2.16)

The equation for active sonars, however, requires all parameters (Hansen, 2012; Lurton, 2010):

\[
\text{DT} < \text{SL} - 2\text{TL} + \text{TS} - \text{NL} + \text{DI} + \text{PG}
\]

(2.17)
The SL is equal to the SIL or SPL at 1 m distance at reference intensity $I_{\text{ref}}$ or reference pressure $p_{\text{ref}}$, see Sec. 2.1. Typical SL are around 100,000 Pa or 220 dB.

$$\text{SL} = 10 \log \left( \frac{I}{I_{\text{ref}}} \right) = 20 \log \left( \frac{p}{p_{\text{ref}}} \right) \tag{2.18}$$

The DI is defined by the ratio of the directional intensity $I_D$ and the omnidirectional intensity $I_O$

$$\text{DI} = 10 \log \left( \frac{I_D}{I_O} \right) \tag{2.19}$$

The TS is expressed by the ratio of scattered sound intensity $I_{\text{scat}}$ and incident sound intensity $I_{\text{inc}}$

$$\text{TS} = 10 \log \left( \frac{I_{\text{scat}}}{I_{\text{inc}}} \right) \bigg|_{r=r_{\text{ref}}} \tag{2.20}$$

TL describes the quantitative loss in sound intensity between the transmitter and receiver. It is the sum of attenuation losses and spreading losses. For a two-way flight the transmission loss is (Havelock et al., 2008)

$$2\text{TL} = 20 \log \left( \frac{I_S}{I_R} \right) \tag{2.21}$$

where $I_S$ is the far field intensity of the source at a reference point 1 m from the source and $I_R$ is the sound intensity at the receiver, expressed as the Equivalent Plane Wave Intensity (EPWI). Empirical expressions for the magnitude of TL can be found in literature (Lurton, 2010):

$$2\text{TL} = 40 \log(R) + 2\alpha R \tag{2.22}$$

**Time of flight measurement**

Sonar applications of active and passive sonars often are based on time of flight measurements. The time delay between transmitting and receiving sound waves is

$$\tau = \frac{2R}{c} \tag{2.23}$$

where $\tau$ is the time delay, $2R$ is the two-way distance to the target and $c$ is the speed of sound. In sonar measurements the time delay in general is known while the distance is unknown. Solving the above equation for $R$ gives the distance as a function of time delay $\tau$ and speed of sound $c$

$$R = \frac{\tau c}{2} \tag{2.24}$$

In practice in can be hard to retain control of all parameters. The sound speed profile, in general, is not known precisely and relies on measurements and empirical estimations. Moreover, the sensor platform is in constant motion due to currents and waves. Additional
sensors are used to account for this, especially for roll and pitch angles as well as heave motions.

**Ranging**

The time of flight measurements provides a simple tool to calculate distances (ranging). The minimum spacing between two echoes in order to still be detected separately is called the range resolution. For common CW pulses the range resolution $\delta R$ for a pulse of length $T_p$ is

$$\delta R = \frac{c T_p}{2}$$

From this equation we can see that shorter pings provide better range resolution, however, the transmitted acoustic energy also becomes less and therefore the echo is harder to detect. For FM pulses the range resolution is dependent on the bandwidth $B$

$$\delta R = \frac{c}{2B}$$

**Doppler shift**

When there is a relative motion between the transmitter or reflector of the initial signal and the receiver of the returning signal, the received echo has a different frequency than the initially transmitted signal. This difference between the initial frequency $f_0$ and the returning frequency $f_1$ is known as the Doppler shift $\delta f$

$$\Delta f = f_1 - f_0$$

(2.25)

A relative motion with decreasing distance (approach) causes a positive frequency shift while motion with increasing distance (departing) causes a negative frequency shift.

For active sonar systems where the transmitter also is the receiver, the relative Doppler shift becomes

$$\frac{\Delta f}{f} = \frac{2\Delta v}{c} \approx 0.695 \text{ Hz/kn/kHz}$$

(2.26)

with $c = 1480 \text{ m/s}$ and $\Delta v$ as the relative velocity. In order to determine the vehicle’s speed over ground, the Doppler shift is calculated as

$$\Delta f = 2f \frac{\Delta v}{c} \cos(\alpha)$$

(2.27)

where $f$ denotes the frequency of the transmitted signal, $\Delta v$ is the relative velocity, $c$ is the speed of sound and $\alpha$ is the angle between the direction of motion and the transmitted acoustic ray.

Acoustic devices that utilize the Doppler shift are ADCPs and DVLs. They can either be equipped with one transducer (single beam) or with an array of transducers (multi beam). The most common array consist of 4 transducers in so called Janus configuration (one beam in each direction: forward, backward, left, right). The beam width needs to be narrow, only a few degrees, in order to avoid interference with echoes from various
surfaces. If all transducers have free line of sight to the sea floor (bottom lock), velocity components in the cardinal directions can be calculated. However, one needs to take into account that the platform might be rolling and pitching. The working principle of such devices is as follows (Rudolph et al., 2012):

1. Send pulses along each transducer axis
2. Sample the returning signal(s)
3. Determine frequency shift between transmitted and received signals
4. Determine the velocity components and resultant velocity

![Figure 2.11: Example: Teledyne Workhorse Navigator DVL (©Teledyne Marine)](image)

**Communication**

Underwater communication is essential to send commands between different entities and also to report positions and other target values. Signals have to be modulated in order to be comprehensible as a message. The simplest case of communication is the exchange of binary data. Binary data only requires a differentiation between two different signals. The **coding** of binary data, i.e. either a 0 or a 1, can be achieved in various ways. Below, Phase Shift Keying (PSK), Frequency Shift Keying (FSK) and chirp modulation are presented. Future work should consider the following publications: (Benson et al., 2010; Kaminsky, 2006; Zhang et al., 2013)

**Phase shift keying**

Using PSK it is possible to efficiently express a quaternary numeral system, i.e. the digits 0, 1, 2 and 3. In underwater communications it, however, may become difficult to encode different phases if the originally transmitted phase has been altered, e.g. after reflection.
Frequency shift keying

Another method is to code a signal by compiling it from pulses of different frequencies. In principle, this enables the user to achieve numeral system of almost arbitrary order, only limited by available bandwidth and by the maximum allowable frequency steps.

Chirp modulation

A binary numeral system is most easily achieved through Chirp Slope Keying (CSK) with linear chirp slopes. A chirp is characterized by its frequency sweep which can either be increasing or decreasing. A chirp with increasing frequency is called an up-chirp while a chirp with decreasing frequency is called a down-chirp. Chirps are known to be detected rather easily and may be preferred to CW when it comes to underwater communications. Fig. 2.14 shows two examples of chirps, one up- and one down-chirping.
Kaminsky (2006) presented results of a comparison between CSK and PSK in the Rayleigh channel, showing that for higher Signal-to-Noise Ratio (SNR), i.e. ≥14.5 dB, CSK is less prone to erroneous underwater communication.

Positioning

Positioning can achieved by triangulating the vehicles position with the help of at least two (two-dimensional positioning, i.e. longitude and latitude) or three (three-dimensional, i.e. longitude, latitude and depth) acoustic beacons deployed at known positions. The beacons consist of transducers which need to detect a signal and, in case of detection, estimate the time of arrival of the signal. The positioning systems listed below are often used in combination with Inertial Navigation Systems (INSs) based on Inertial Measurement Units (IMUs). The additional use of baseline positioning increasing the navigation accuracy significantly. Fig. 2.15 shows the different positioning systems described below.
Long baseline

LBL positioning offers high accuracy positioning over long ranges. It relies on triangulation techniques. The beacons are usually deployed on the seabed and equipped with buoys that can also carry a Global Positioning System (GPS) (Gamroth et al., 2011).

(Ultra) Short baseline

Short Base-line (SBL) positioning uses a finite number of beacons that are usually attached to the mother ship. If the baseline is very short, e.g. in the case of acoustics transducer arrays, the system is called Ultra Short Base-line (USBL). The bearing in this case is estimated from the relative phase measured by each transducer. USBL easily leads to great errors in position estimation on long distances.

Acoustics Arrays & Beamforming

An acoustic array is an assembly of a finite number of transducers. They are most commonly arranged either as disc arrays, rectangle arrays or line arrays. An acoustic array is characterized by the geometrical assembly of its elements.

Beam forming is a spatial filtering technique associated with phased arrays. A beamformer enhances the sonars functionality by the use of directivity. Even though the underlying theory is identical, directivity of transmitted and received signals need to be explained separately.

For transmitted signals directivity means to control its preferred direction of propagation. This is achieved by transmitting the signal with phase shifts between each of the beamformer’s elements. This leads to either constructive or destructive interference and eventually to directivity of the transmitted signal, as Fig. 2.16 shows.

![Figure 2.16: Schematic of a phased array. Source: Wikipedia](image-url)
For received signals directivity means to determine the direction from which a signal is transmitted or reflected towards the receiver. The direction of emitting sound source or reflector is calculated from the time delay and/or phase shift which is recorded between each of the beamformers elements. This procedure is analogous to SBL and USBL positioning systems.

Besides directivity and beam steering beamformers also allow the user to have increased side lobe control. However, directivity of the main lobe and the side lobe levels are trading each other off.

Analogous to the single transducer beam pattern the beamformer’s beam pattern also is frequency dependent. The main lobe’s width increases with increasing frequency and decreases with decreasing frequency. So called grating lobes occur if the elements of the array are spaced (more than) one wavelength apart. This usually happens for very high frequencies (c.f. $\lambda = \frac{c}{f}$) and is called spatial aliasing.

**Sonars and Ice**

Publications on acoustic profiling of ice and on the deployment of acoustic devices in ice-covered waters are rather sparse. According to Melling (1998) ice profiling requires a narrow-beam transducer with a beamwidth $< 3^\circ$ at $-3$ dB in order to avoid biases that otherwise would be introduced into ice-draft statistics. It further is reported that typical sonar source levels are about 218.5 dB relative to $1 \mu Pa$ $1 m$ with an operating frequency of 400 kHz. For further reading, the authors refer to Langleben (1969), Langleben (1970), Greene et al. (1985), and Galloway et al. (1996) and others.

### 2.3 Digital Signal Processing

Upon receiving the echo signal with additive noise undesired frequencies can optionally be filtered with an FIR bandpass filter. In the next step the pre-filtered signal is compared to a template (replica) of the originally transmitted signal, a process which is known as matched filtering or correlation processing. The signal then can be squared point-wise in order to amplify peaks. After setting a detection threshold the returning echos can finally be detected. Fig. 2.17 shows this procedure.

**Matched filtering**

Matched filters perform a cross-correlation between the input signal and a replica of the transmitted pulse (template). If signal and noise are uncorrelated, the echo’s footprint on the filtered signal becomes very distinct and peaks are easily detectable. Matched filters enable us to detect returning echoes even at low SNR.

In DSP, a matched filter is most easily achieved in the frequency domain. It draws on the theorems of cross-correlation and convolution, see Sec. A.1 for background theory. If one of the input signals is time-reverted and complex conjugated, the cross-correlation integral and the convolution integral are equivalent. In the Fourier domain, complex conjugation is equivalent to time-reversal in the time domain.
2 Theory

Pulse transmission

Received signal $s(t)$

Optional: FIR-filter (bandpass)

Matched filtering

$\cdot^2$

Threshold

Detection

**Figure 2.17:** Digital Signal Processing in Sonar applications
In the following derivations,

- \( \mathcal{F} \) and \( \mathcal{F}^{-1} \) are the Fourier transformation and its inverse,
- \( * \) is convolution, \( \star \) is cross-correlation, \( (\cdot) \) is point-wise multiplication,
- \( \overline{h} \) denotes complex-conjugation,
- \( s(t) \) and \( h(t) \) are the input signal and template,
- \( S(f) \) and \( H(f) \) are their Fourier transforms.

Based on convolution, cross-correlation is defined as (Weisstein, 2017c)

\[
h(t) \ast s(t) = \overline{h}(-t) \ast s(t)
\]  
(2.28)

The convolution theorem shows that (Weisstein, 2017b)

\[
h(t) \ast s(t) = \mathcal{F}^{-1} \{ H(f) \cdot S(f) \}
\]  
(2.29)

Due to the conjugation property of Fourier transforms

\[
\mathcal{F}(\overline{h}(-t)) = \overline{H(f)}
\]  
(2.30)

we can combine Eq. 2.28, 2.29 and 2.30, yielding a new expression for the cross-correlation (cross-correlation theorem (Weisstein, 2017d))

\[
h(t) \ast s(t) = \overline{h}(-t) \ast s(t) = \mathcal{F}^{-1} \{ \overline{H}(f) \cdot S(f) \}
\]  
(2.31)

where \( \overline{H}(f) \cdot S(f) \) is the cross-spectral density of functions \( h(t) \) and \( s(t) \).

**Doppler Signal Processing**

In DVL applications the Doppler shift can be identified in different ways. First of all, the frequency of the transmitted signal \( f_0 \) needs to be known. Then, the following steps need to be undertaken:

1. Receive the echo
2. Post-process the signal (filters)
3. Determine frequency and other parameters
4. Compute velocity components

The main task is to determine the frequency of the received signal. There are several methods available in literature. They may either be based on time-domain or frequency-domain signal processing.

1. Frequency Spectrum (see Fast Fourier Transformation (FFT)) (Kay et al., 1981)
2. Zero-crossing interpolation
2 Theory

3. Non-linear square fitting (Rudolph et al., 2012)

4. Pulse Pair Processing (Lhermitte et al., 1984; Liu et al., 2009)

**Frequency Spectrum**

Fourier transformation can be used in order to compute the frequency spectrum of a signal. A distinct peak in the frequency spectrum can potentially show the dominant frequency of the received signal. As for digital signals, FFT algorithms can be implemented to derive the frequency spectrum of the echo. However, FFT yield discrete spectra, a fact that leads to problems related to frequency resolution (bins). The frequency resolution $\delta f$ of a spectrum with sampling rate $f_s$ and number of samples $N$ is

$$\delta f = \frac{f_s}{N} \quad (2.32)$$

For example, with a sampling rate $f_s = 2.8 \text{ MHz}$ and a number of samples $N = 4096$, the frequency resolution is $684 \text{ Hz}$. For a transmitted signal with center frequency $f_0 = 200 \text{ kHz}$ the corresponding velocity resolution is $2.5 \frac{\text{ m}}{\text{s}}$. For Doppler shift applications this resolution is insufficient. The frequency resolution can be increased (i.e. smaller bins) by either decreasing the sampling rate $f_s$ or increasing the number of samples $N$.

**Zero-crossing interpolation**

The frequency of the received signal can also be determined by identifying zero-crossings of the echo and calculating the zero-crossing period. The disadvantage of this method is the requirement of good SNR, since strong noise levels can shift the zero-crossings. Application of filters, e.g. finite impulse response (FIR) filters, may be required. Besides this, the advantage of zero-crossing based methods is the computational efficiency. They neither require much Random Access Memory (RAM) nor does it take much time to compute. However, errors still may arise from too high SNR. In some cases, the overall shape of the returning echoes can be too complex and distorted in order to be analyzed by simple and straight-forward zero-crossing algorithms.

**Non-linear least squares fitting**

The Doppler shift can be determined iteratively by non-linear least squares methods which find best-matches for frequency $f$, phase shift $\phi$ and amplitude $A$ of the returning signal. The frequency which matches the returning signal best, represents the Doppler shifted return frequency. This is an optimization problem using the method of non-linear least squares.

However, such parameter estimation methods may end up with a result that does not converge because the amplitude of the echo is not constant, which is influenced by the seabed and the directivity pattern of the transducer. The accumulation of this error will result in significant errors in navigation.

Fig. 2.18 illustrates the concept of non-linear least squares methods for Doppler shift detection. The underlying mathematics are described as follows.
The signals transmitted and received by the transducer are sinusoidal waves which can be written as

\[ s(t) = A \sin(2\pi ft + \phi) \]  

The received (digital) signal is discrete and in the form

\[ r(i) = A \sin \left( \frac{2\pi f_i}{f_s} + \phi \right) \]  

In Eq. 2.34, \( f_s \) is the sampling frequency and the three parameters that need to be solved for are: Amplitude \( A \), frequency \( f \) and starting phase \( \phi \). To determine these parameters, especially the frequency, an initial guess needs to be made. Substituting \( \Lambda = [\lambda_1, \lambda_2, \lambda_3] \) in Eq. 2.34 yields the optimization function

\[ f(\Lambda, i) = \lambda_1 \sin \left( \frac{2\pi \lambda_2 i}{f_s} + \lambda_3 \right) \]  

Then, the error between the assumption and the received signal is calculated as:

\[ R^2 = (r(i) - f(\Lambda, i)) \cdot (r(i) - f(\Lambda, i)) \]  

This error is to be minimized. Minimization is achieved numerically by iteration

\[ \Lambda_n = \Lambda_{n-1} + d\Lambda \]  

In Eq. 2.37, the difference between the two assumptions is:

\[ d\Lambda = [B^T B]^{-1} B^T (r(i) - f(\Lambda, i)) \]  

where \( B \) is the matrix of the partial derivatives of the assumed signal \( f(\Lambda, i) \).

The difference between the assumption and received signal will converge after several iterations and the unknown parameters can be solved. The frequency is determined as one of those parameters.

**Pulse-pair processing**

Pulse-pair processing is a method widely-used for acoustic velocity measurements in radar and sonar applications. In most cases, the magnitude of the Doppler shift is much lower than the centre frequency \( (\Delta f \ll f_0) \). A useful approach hence is to determine the phase derivative of two consecutive pulses rather than the frequency of the signal. In this processing method, the correlation function of the transmitted signal and received signal is calculated. The frequency of the received signal with Doppler shift is determined by calculating the phase derivative \( \frac{d\phi}{dt} \) of the correlation function:

\[ f = \frac{1}{2\pi} \frac{d\phi}{dt} \]  

However, in this method transmitted and received signals are required to be in complex form, which increases requirements towards the hardware. Fig. 2.19 illustrates the
Received Signal $s(i)$

Assumed Signal $f(\Lambda, i)$

Error Calculation $R^2$

Convergence criterion met?

No

Change the Assumption

$\Lambda_n = \Lambda_{n-1} + d\Lambda$

Yes

Doppler Shift $\Delta f$

Velocity $v$

Figure 2.18: Non-linear least squares method for Doppler detection

concept of pulse-pair processing.
2 Theory

Transmitted Signal
\[ x(t) = \text{Re}(t) + i \cdot \text{Im}(t) \]

Received Signal
\[ s(t) = \text{Re}'(t) + i \cdot \text{Im}'(t) \]

Transmitted Spectrum
\[ X(f) = \text{FFT}(x(t)) \]

Received Spectrum
\[ S(f) = \text{FFT}(s(t)) \]

Correlation Function
\[ y(t) = \text{IFFT}(Y(f)) \]

Phase Function
\[ \phi(t) = \arctan\left( \frac{\text{Im}(y(t))}{\text{Re}(y(t))} \right) \]

Frequency
\[ f = \frac{1}{2\pi} \frac{d\phi}{dt} \]

Figure 2.19: Pulse-pair processing method for Doppler detection
3 Hardware

Fig. 3.1 shows the printed circuit board (PCB) of the sonar developed by Cadson Production AB, Stockholm. A new version of the PCB is under development. It will significantly decrease in size and be designed to fit into one housing/casing together with a transducer. Besides, electro-technical performance will also be increased, allowing to transmit higher frequencies and output higher voltages.

![The sonar’s PCB](image)

**Figure 3.1:** The sonar’s PCB

3.1 Micro-control Unit

The built-in Micro-controller Unit (MCU) is a Microchip PIC32MZ2048EFH064 which is programmed with a Microchip MPLAB® ICD3 In-Circuit Debugger, see Fig. 3.2. In this project, the used software is the MPLAB® X IDE V3.55 with the XC32 V1.40 compiler. See Sec. B.1 for programming instructions. The source code may be available on request.

3.2 Analog-to-Digital Converter

The Analog-to-Digital Converter (ADC) is working with a reference voltage of 3.3 V and a base level of 0 V at 12 bit resolution, centered by the operating amperage in the amplifying stage at 2.048 V. However, this is subject to change.

An ADC performs discretization in amplitude, i.e. quantization, of analog signals. Due to a bit limit ADCs only have a limited resolution of $2^n$ steps where $n$ is the number of bits. For the given ADC that means $2^{12} = 4096$ steps (Bodén et al., 2010).
3.3 Transducer

The transducer used throughout most parts of this project is the model TD0200KA by Shenzhen Hurricane Tech. Co., Limited. It is of small size, measuring only 31 mm in height and 42 mm in outer diameter. Fig. 3.3 shows the transducer. Specifications can be found in Tab. 3.1.

3.4 Data Transfer

At the current stage of development data is transferred through a serial port. Connection is established by a USB 2.0 to UART cable (Model C232HD by Future Technology Devices International Ltd. (FTDI)). A serial port can be opened with a serial terminal software such as RealTerm (freeware), but also with MATrix LABoratory (MATLAB®) and similar software.

Sec. B.2 gives an overview of the most important commands to control the sonar. Commands (as well as data) are transferred through a serial connection with baud rate 921 600 bit/s.
### Transducer Specification

<table>
<thead>
<tr>
<th>Specification</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Material</td>
<td>PZT-4</td>
</tr>
<tr>
<td>Frequency</td>
<td>200 ± 4 kHz</td>
</tr>
<tr>
<td>Beam angle, in water (−3 dB)</td>
<td>14 ± 1 °</td>
</tr>
<tr>
<td>Launch Sensitivity</td>
<td>161 dB</td>
</tr>
<tr>
<td>Receive Sensitivity</td>
<td>−190 dB</td>
</tr>
<tr>
<td>Rated Power</td>
<td>20 W</td>
</tr>
<tr>
<td>Working Temperature</td>
<td>−20 to 70 °C</td>
</tr>
<tr>
<td>Weight</td>
<td>30 g</td>
</tr>
<tr>
<td>Max. Voltage</td>
<td>800 V</td>
</tr>
</tbody>
</table>

*Table 3.1*: Data sheet of transducer model TD0200KA by Shenzhen Hurricane Tech. Co., Ltd., as given by the producer.
4 Software Architecture

The source code is written in ANSI C and is structured as follows:

1. Inclusion of libraries and header files (#include)
2. Additional instructions to compiler (#pragma)
3. Definition of macros (#define)
4. Declaration of global variables
5. Definition of functions
6. Main program main()

Upon starting the device, the main program is executed. As a first step, the ports and peripheral pin select (PPS) are set up. Consequently, the universal asynchronous receiver/transmitter (UART) is initialized and the device is ready to receive the commands within an infinite loop. If a setting-related command is received, action is taken and a confirmation is send to the port. If a pulse train is requested, the frequency and voltage are set accordingly and the device is set up to first transmit the signal and then to listen. The device eventually starts sampling. The sampling function has been the main development site in this project. It contains the sampling algorithm as well as the DSP algorithms, including the matched filter.

4.1 Sampling Algorithm

The sampling algorithm is one of the most important parts of the source code. Since parallel computing is not possible with the present hardware and RAM is limited, a potentially returning echo is detected, sampled and stored in a circled buffer for DSP purposes. In the current version of the source code, a potentially returning echo is identified through a threshold detector.

Fig. 4.1 and 4.2 show how the sampling algorithm works.

Threshold Detector

The threshold detector uses the mean value of the ADC output to derive the upper and lower thresholds. This mean value is an initial guess or the mean value of the previous successful ping. The thresholds are based on the standard deviation of the noise. Upon deriving the thresholds the sampling begins. The sampling algorithm is configured to terminate (i.e. echo detected) when a certain number of exceedances above the upper and below the lower threshold have been recorded. These exceedances, however, need to be within a certain band. Otherwise the exceedance counters are reset.
4 Software Architecture

Start

Thresholds Up & Down

All counters = 0

Read ADC
Read Timer

Threshold criterion fulfilled?

yes

no

Enough exceedances?

yes

no

Threshold Exceedance Algorithm

See Flowchart: Threshold Exceedance Algorithm

Termination criterion

Loop counters ++

Buffer Full?

no

yes

Reset buffer counter

Termination?

no

yes

Stop

Figure 4.1: Flowchart describing the sampling algorithm
Figure 4.2: Threshold Exceedance Algorithm
Fig. 4.3 shows an example of the threshold detector. In this case, a total number of 32 exceedances above the upper and 32 exceedances below the lower thresholds have been aimed at. The maximum allowable distance between the first and last exceedance was set to 250 samples, leading to omittance of the first few exceedances (marked with circles).

Dead time

Just as any other mechanically elastic material that is subject to an oscillating exciting force, piezoceramics continue oscillating after the excitation force has been removed until its oscillations are fully damped. If this is not controlled, which is the case for this sonar, it can cause severe problems. A dead time, i.e. a state of idle of certain duration between transmission and recording, needs to be applied after the successful transmission of a ping. This is necessary because the ringing would otherwise cause an echo detection. Fig. 4.4 illustrates this dead time.

Circled Buffer & Rearrangement

The sampling algorithm stores the ADC data and the time data in a circled buffer. A circled buffer offers the possibility to sample data continuously (for unlimited time) and save data strings of certain length in an array. When sampling is finished, the starting point of the data series can be at any index of the circled buffer. Hence, the buffer has to be unfolded & rearranged afterwards. Fig. 4.5 shows this scheme.

4.2 Matched Filter

The theory behind matched filters was explained in Sec. 2.3. Fig. 4.6 shows how matched filtering is achieved computationally. The MCU already provides pre-installed and configured C libraries that also include DSP functions such as FIR filters and FFT algorithms. Unfortunately, an inverse FFT function is missing. However, an inverse FFT can easily be achieved by complex-conjugation of the Fourier transform of the complex-conjugate of the data under consideration.

\[ x = \text{conj} \left( \text{fft} \left( \text{conj} (X) \right) \right) \]

Fig. 4.6, again, shows how this is achieved with the MCU used.

Template Generation

As described in Sec. 2.3, the received signal is compared to a replica, the template, of the transmitted signal. With the current hardware it is not possible to log the transmitted signal. Therefore, and for reasons of simplification, the transmitted signal is modeled on the basis of the number of pulses and frequency of the transmitted signal (as per command). For CW signals it is described mathematically as

\[ y = \sin \left( \pi \frac{f_N}{N} t \right) \cdot \sin \left( 2\pi f_0 t + \pi \frac{\text{BW}}{T_p} t^2 \right) \]

(4.1)
Figure 4.3: Example of threshold detection on real signal (Depth: 44.5 m, Baggensfjärden. (a) shows the complete sampled signal, (b) shows a close-up view of the sampled threshold exceedances.
Figure 4.4: Dead time schematic

Figure 4.5: Circled buffer schematic
And for FM signals it is described mathematically as (note the factor $\frac{1}{2}$ in the first sine)

$$y = \sin\left(\pi \frac{f_s}{2N} t\right) \cdot \sin\left(2\pi f_0 t + \pi \frac{BW}{T_p} t^2\right)$$  \hspace{1cm} (4.2)

In these equations $f_s$ refers to the sampling rate, $N$ is the number of points of the echo, $t$ is the time, $f_0$ is the (starting) frequency of the transmitted pulse, $BW$ is the bandwidth (0 for CW) and $T_p$ is the pulse duration. The pulse duration is defined as

$$T_p = \sum_{i=1}^{np} \frac{1}{f + i \cdot \frac{BW}{np}}$$  \hspace{1cm} (4.3)

where $np$ is the number of pulses. The first sine multiplicand in Eq. 4.2 and 4.3 defines the overall envelope of the template. In case of CW the envelope is a half sine, in case of FM it is a quarter sine. The amplitude of the template is scaled to the peak amplitude of the received signal. The length of the CW template is fixed to 45 cycles, regardless of the length of the actual transmitted pulse. The reason for this is the uncertainty of the actual length of returning echoes. However, experiments have shown that 45 cycles provide a good fit for CW signals. The same approach, however, cannot be used for FM signals since the frequency is swept and needs to be identical for each cycle. Hence, for FM signals, the actual number of transmitted pulses is used to build the template. Due to hardware limitations FM signals could not be studied and used extensively. Fig. 4.7 shows an example of a CW template matching real data, Fig. 4.8 shows an example of a FM template matching real data.
Distortion Filter

A distortion filter is applied to the output of the matched filter. The reason for this distortion is the occurrence of multiple echoes after the first one, which possibly show greater correlation (due to larger amplitude). This would inevitably lead to an error in the distance measurement. Therefore each sample of the received signal is multiplied pointwise with the corresponding value from the linear distortion function, see Fig. 4.9. The authors are aware that applying a distortion also shifts the location of global maxima. Compared to multipath echoes, however, this resulting error is negligibly small. Fig. 4.10 shows an example of the distortion on real data.
4 Software Architecture

4.3 Self-adjusting Auto-pinging

When the sonar is put into auto-pinging mode it continuously sends out pings. Depending on if there is detection and on the echo amplitude, the number of transmitted pulses and the output voltage are adjusted. In case of unnecessarily high SNR the transmitted energy is decreased. In case of no detection or low SNR the transmitted energy is increased. Fig. 4.11 illustrates this behavior.
4 Software Architecture

start

listen() to noise

send_pulsetrain(np,f,U)

\[ \Delta A = \text{ADC Ideal} - \text{ADC Max} \]

Check if the max. ADC output is close to the ideal value

- \( \Delta A \in [-T, T] \)
  - yes
  - \( V \in (5, 280) \)
    - yes voltchange()
    - no
      - Condition 1 or 2
        - yes
          - \( U = 280, \ np < 25, \ \Delta A < -T \)
          - no
            - npulsechange()
          - \( U = 5, \ np > 15, \ \Delta A > T \)

Figure 4.11: Auto-adjustment of number of pulses and output voltage in auto-pinging mode
5 Simulations

In order to understand and interpret results obtained from experiments, certain aspects of underwater acoustics and related DSP have been simulated using MATLAB®. Namely, these are

1. Matched filter
2. Directivity pattern and beam width

5.1 Matched Filter

The performance of matched filters has been evaluated in MATLAB® in order to provide a theoretical measure of optimum echo detection. Fig. 5.1 shows how the matched filtering simulation works, the simulation parameters are given in Tab. 5.1. Fig. 5.2 shows a comparison of matched filter performance on FM and CW signals at different SNRs.

5.2 Directivity Pattern & Beam Width

Based on the theory as presented in Sec. 2.2, the actual beam pattern of the transducer TD0200KA was simulated. This gives important insight into the shape of main lobe and side lobes and can help interpreting echoes. The transducer used in this project is a disk-shaped transducer, its directivity pattern in transmitting or receiving mode can be estimated from Eq. 2.10. The directivity pattern is a function of the frequency $f$, the

<table>
<thead>
<tr>
<th>No.</th>
<th>FM</th>
<th>CW</th>
</tr>
</thead>
<tbody>
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</tr>
<tr>
<td>4</td>
<td>5</td>
<td>6</td>
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<tr>
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<td>$R_1$</td>
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<td></td>
</tr>
<tr>
<td>$R_2$</td>
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<td></td>
</tr>
<tr>
<td>$R_3$</td>
<td>15.3 m</td>
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</tbody>
</table>

Table 5.1: Matched Filter Simulation: Parameters
Figure 5.2: Matched Filter Simulation: Comparison between FM and CW signals
speed of sound \( c \), the diameter \( d \) and the shape of the transducer. Since \( c \) is dependent on the medium, the directivity pattern also is. As specified in Sec. 3.3, the diameter of TD0200KA is 42 mm and the centre frequency is 200 kHz. The directivity pattern is modelled using a speed of sound \( c = 1487 \, \text{m/s} \). Fig. 5.3 shows the simulated beam pattern in water.

**TD0200KA - Beam Width: 12.2°**

![Beam pattern diagram with specific angles and dB levels]

**Figure 5.3:** Calculated beam pattern of the transducer TD0200KA
6 Experiments and Results

6.1 Underwater Laboratory

Experiments have been performed at the underwater laboratory of the Swedish Defense Research Agency (Swedish: *Totalförsvarets forskningsinstitut*) (FOI) in order to achieve reference signals (echoes) from an acoustically nearly perfect environment and also to statistically evaluate the error in depth measurements. FOI disposes a water tank of dimensions 8 m x 4 m x 4 m (Length x Width x Depth) whose bottom and sides are equipped with ceramic tiles, granting perfect reflectivity of sound waves. The tank is resting on the rock beneath the building and hence is not connected to the building itself, making it independent of noise sources inside the building. Fig. 6.1 shows the results of a Conductivity, Temperature, Density (CTD) measurement which has been conducted to derive the speed of sound ($c$).

$c$ is calculated using Eq. 2.6. A relation to link conductivity to salinity (in ppt) can be found in IOC (2010). $c$ was consequently calculated as $1487 \text{ m/s}$. A series of depth measurements at a fixed depth of 3.71 m was conducted. For comparison, the distances (ranges) have been derived from the time at which the threshold is exceeded (triggered) and from the time at which the matched filter shows best correlation,
see Fig. 6.2. As one can see, the matched filter produces slightly larger deviations from the expected value. Currently, both distances are output to the user. Despite the more error-prone performance of matched filter based depth measurements the matched filter is considered to be an absolutely critical tool for verification of the presence of an echo.

Fig. 6.3 shows an example echo from the performed measurements. The signal was generated with 10 pulses at 195 000 Hz at output voltage of 5 V.

6.2 Stockholms Skärgård

Additional tests were performed on a motor boat (Buster Magnum) in the Stockholm archipelago for depths measurements and Doppler shift measurements.

Depth measurements

For depth measurements the transducer is supposed to aim straight downwards. Fig. 6.4 illustrates the experimental setup for ranging. For the tests the transducer was mounted onto aluminum frames (Bosch Rexroth) which in turn were fixed to the boat at the cleats with the help of wooden wedges.

The depth measurements were performed in Baggensfjärden. The tests were performed with the transducer described in Sec. 3.3 at resonance frequency, i.e. 200 kHz. Fig. 6.5 shows a number of example echoes that have been received. Fig. 6.5a shows a nearly perfect echo received in shallow water. Hence, the required voltage is very low. Fig. 6.5b shows similar results, but with echoes resulting from multipath propagation. In Sec. 4.2 it is described how the distortion filter ensures detection of the first echo despite the later echoes being larger in amplitude. The reason for this behaviour might be increased target strength due to higher reflectivity (e.g. rocks). Fig. 6.5c and 6.5d show echoes from increased water depth, close to the currently established maximum distance. They also show how much the echoes can vary in shape. Fig. 6.5e and 6.5f show two more examples with significantly decreased SNR due to weaker echo amplitude. The positive trend of the mean value in Fig. 6.5e is related to the electronics, it can potentially cause problems in echo detection and has not yet been resolved.

In addition, the auto-pinging mode was used to statistically evaluate the sonar’s performance. Unfortunately, reference data from the test sites is missing and thus the actual depth measurements cannot be validated. However, they are in fairly good agreement with the authors’ expectations according to nautical charts. Fig. 6.6 and 6.7 show a number of successful depth measurements in the top graphs and the according number of pulses and voltage in the bottom graphs. The success rate for these measurements lies between 90% to 95%. Especially Fig. 6.6 shows exceptionally good results with only few false detections.

Doppler shift

For Doppler shift measurements (s. DVL) the experimental setup had to be changed. The transducer has been immersed at the bow of the boat pointing forward at an angle of 30°. This setup is illustrated in Fig. 6.8. The tests have been performed in Halvkakssundet & Skurusundet.
Figure 6.2: Statistical Evaluation of Range Measurements at FOI. Target Range: 3.71 m. Comparison between Threshold (TH) and Matched Filter (MF) based Computations.
6 Experiments and Results

Unfortunately, it was not possible to detect constant frequency shifts at constant speeds. These errors in frequency determination result from various circumstances. The angle between the transducer axis and the sea bed may have significant impact on the echo and may lead to errors in frequency determination. In order to reliably trace the error sources, experimental setups with higher degrees of control are required. This, however, is difficult to achieve in the real environment due to uncertainties in sea bed characteristics and the dynamic motions of the boat.

Fig. 6.9 shows a scatter plot with the frequencies of echoes that have been received during a cruising speed of $\approx 4\text{ kn}$. The echoes have been recorded and the frequencies were calculated from the zero-crossings of the bandpass filtered signals (FIR filter). Zero-crossings were computed in the vicinity of maximum amplitude of the echoes. The FIR filter is centered at 200 kHz with a bandwidth of $\pm 10$ kHz. From Fig. 6.9 it can be seen that it is hardly possible to relate the echo frequency to the speed of the boat, i.e. Doppler shift detection is not possible. For reference, the echo frequency for $\approx 0\text{ kn}$ has been calculated as 199.61 kHz. This frequency is the average frequency of a pre-study performed at FOI. The results of this pre-study are presented in Fig. 6.10. The frequency at $\approx 4\text{ kn}$ theoretically is 199.88 kHz. It is assumed that there are multiple sources of error that alter the output frequency, and hence the Doppler shift cannot be detected.

6.3 Known Issues

The sonar developed within the frame of this thesis project is a prototype. Despite greatest efforts by the authors there are still issues that were not resolved completely. Fig. 6.11a shows some pre-recorded noise that is recorded prior to transmission of a pulse.
6 Experiments and Results

Figure 6.5: Example echoes from in-situ experiments in the Stockholm archipelago
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Figure 6.6: Depth measurements #1

Figure 6.7: Depth measurements #2
**Figure 6.8:** Experimental setup for Doppler shift measurements

**Figure 6.9:** Results from the Doppler shift measurements
6 Experiments and Results

Figure 6.10: Frequency analysis of transmitted pulses of current hardware

(a) Frequency analysis from 198.5 kHz to 201.5 kHz
(b) Closeup of frequency analysis showing 200 kHz and 200.5 kHz

Train. Based on this noise the standard deviation of the noise is calculated and used for the threshold definition. As one can see, there is a sudden jump in noise Root Mean Square (RMS) towards the end of the recording. The noise level at the end supposedly is correct. Fig. 6.11b shows the behavior of the ADC in the beginning of a measurement, after transmission of a pulse. This behavior is likely to be linked to the switching of states from transmission to sensing. Unfortunately, this behavior can potentially trigger the threshold detector and cause false echo detection under certain circumstances and in connection with ringing. Fig. 6.11c shows another problem which is purely related to the electronics. It is assumed to be a kick-off of the Metal–Oxide-Semiconductor Field-Effect Transistor (MOSFET) drivers, causing two consecutive spikes. When closely spaced, they can also trigger the threshold detector. Because of the large area covered by the spikes also the matched filter can be triggered. This will definitely need adjustment of threshold settings. The authors, however, are optimistic to resolve this issue with an updated version of the hardware.
Figure 6.11: Known issues at the current stage of development
7 Conclusion

An echo sounder for depth ranging has been developed and tested successfully. Tests at FOI have shown that the sonar can measure depths with an average deviation of $<1$ cm in a controlled setup. Furthermore, in-situ experiments have shown that the sonar also works under realistic and non-ideal conditions. During tests in the Stockholm archipelago echo detection rates of up to 95% have been achieved in water depths of up to 55 m, which currently poses the limit in depth ranging. The authors, however, expect that improvements on the electronics (PCB and transducers) will increase the sonar’s depth rating.

For the continuation of the sonar project a static installation of the transducer inside the boat’s hull should be considered. This would significantly simplify the conduction of in-situ experiments. In order to validate data obtained with the sonar, a commercial echo sounder could be run in parallel and the acquired data could be compared. GPS should be used to relate and compare the data.

More extensive Doppler shift measurements should be performed in experiments with a higher degree of control, possibly in a laboratory environment (e.g. at FOI). A steady relative motion is favorable, as well as flat and high reflective surfaces.

In general, a solid foundation of methodologies in underwater acoustics has been established. Future work can be based on the results presented in this thesis.
References


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Appendix
A Theory

A.1 Expected Value, Variance, Covariance, Correlation

In statistics, the expected value $E$ is defined as:

$$E(X) = \sum_x x \cdot p(x)$$

where $p(x)$ is the probability of $x$). The variance is defined as the squared deviation from the mean $\mu = E[X]$:

$$\text{var}(X) = E[(X - \mu)^2] = \sum_x (x - \mu)^2 p(x)$$

and is often denoted $\sigma^2$. The standard deviation $\sigma$ is the square root of the variance $\sigma^2$. The covariance is defined as follows:

$$\text{cov}(X,Y) = E[(X - \mu_x)(Y - \mu_y)]$$

A positive covariance indicates a positive trend in the data (increasing $y$ for increasing $x$) while a negative covariance indicates a negative trend in the data. $\mu_x$ and $\mu_y$ denote the mean values of $x$ and $y$, respectively. Correlation is defined as follows:

$$\text{corr}(X,Y) = \frac{\text{cov}(X,Y)}{\sqrt{\text{var}(X)\text{var}(Y)}}$$

Correlation is often denoted $\rho$ and can only take values from $-1 \leq \rho \leq 1$. When analyzing time series, the terms auto-correlation and cross-correlation often are used. Auto-correlation is the correlation of one signal with itself, cross-correlation is the correlation of two different signals.

A.2 Convolution & Cross-correlation

Convolution is defined as

$$[f * g](t) = \int_{-\infty}^{\infty} f(\tau)g(t - \tau)d\tau$$
And through Eq. 2.28 cross-correlation is defined as

\[ (f \ast g)(t) = \int_{-\infty}^{\infty} f(-\tau)g(t - \tau)d\tau \]

For references see Weisstein (2017a) and Weisstein (2017c).

### A.3 Fourier Transformation

The Fourier transformation is defined as follows:

\[ x(\omega) = \int x(t) \cdot \frac{1}{\sqrt{2\pi}}e^{-i\omega t} dt \]
B Hardware & Software

B.1 Using the ICD3 In-Circuit Debugger

This is a step-by-step guide about how to setup the MPLAB® X Integrated Development Environment (IDE) with XC32 Compiler and correct MCU on a Windows computer. The XC32 should be installed with v1.40.

1. Create new project

2. Choose Microchip Embedded and Standalone Project. A working makefile is generated by the IDE.

3. Specify Device Family and Device (PIC32MZ2048EFH064)
4. Select Tool (ICD3)

5. Select Compiler (XC32)

6. Specify Project Name and Folder
7. Add Existing Item (select source code) to Source Files

8. After plugging in the ICD3 In-Circuit Debugger right-click on the project and select properties
9. Configure the project properties to use the plugged-in ICD3 with XC32 compiler

10. The device is now ready to be programmed. The tools *Build Main Project*, *Debug Main Project* and *Make and Program Device Main Project*, among others, are found in the top tool bar.

### B.2 Important Commands

**Sonar communication protocol, 27/06/2017**

Prefix is always address of the sonar.

**ST**, set measure start delay in microseconds.

**Example:**

Send \(\rightarrow\) $000$ST9999$

Response \(\rightarrow\) "Sampling delay 9999 microseconds"
AR, toggle raw output of signal
Example:
Send -> $000$AR$
Response -> "RAW output enabled" or "RAW output disabled"

PT, send pulsetrain (number of pulses, frequency, voltage)
pulses 1–1000, frequency 1000–500000, voltage 5 – 280V
Example:
Send -> $000$PT15:200000:150$
Response -> Range or raw output

CT, send chirptrain (number of pulses, voltage, from_frequency, to_frequency)
pulses 1–1000, voltage 5 – 320V, from_frequency 1000–500000, to_frequency 1000–500000
Example:
Send -> $000$CT20:150:195000:205000$
Response -> Range or raw output

SP, set speed of sound in m/s
Example:
Send -> $000$SP340$
Response -> "Speed of sound set to 340"

TE, threshold exceedance level factor in percent (multiplier for noise RMS if AT=1)
Example:
Send -> $000$TE750$
Response -> "Threshold exceedance level set to 750"

AP, toggle autopinging on/off
Example:
Send -> $000$AP$
Response -> "Auto-pinging enabled" or "Auto-pinging disabled"

MF, matched filter threshold
Example:
Send -> $000$MF250$
Response -> "MF Threshold set to 250"

VT, set ADC output tolerance for voltage changing and number of pulse changing
Example:
Send -> $000$VT50$
Response -> "Amplitude Tolerance set to 50"
IA, set ideal ADC output value of the received signal
Example:
Send -> $000$IA250$
Response -> "Ideal Amplitude set to 250"

MR, toggle the average of certain number of ranges
Example:
Send -> $000$MR$
Response -> "Range–averaging disabled" or "Range–averaging enabled"

NP, set the number of ranges for average
Example:
Send -> $000$NP10$
Response -> "Number of pings set to 250"

PORTSPEED is 921600 bps for UART.
System voltage shall be 12V.