Enhancement of guided wave tomography for pipelines via improved transduction and signal deconvolution

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**ENHANCEMENT OF GUIDED WAVE
TOMOGRAPHY FOR PIPELINES VIA
IMPROVED TRANSDUCTION AND SIGNAL
DECONVOLUTION**

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ENGINEERING**

A thesis submitted to the Nanyang Technological University in
partial fulfilment of the requirement for the degree of

Doctor of Philosophy

2021

Statement of Originality

I hereby certify that the work embodied in this thesis is the result of original research, is free of plagiarised materials, and has not been submitted for a higher degree to any other University or Institution.

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Mahadhir Hamzah Caffoor

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Summary

Structural health monitoring (SHM) as opposed to conventional NDT methods involve the placement of permanent sensors and improve the reliability, cost efficiency and allows for continuous monitoring of non-uniform degradation. SHM monitoring can be categorised into local or area monitoring but area monitoring has the benefit of providing a better baseline to assess the structural integrity of the pipeline in cases where critical parts are unknown because local monitoring does not provide enough of a representation to assess the integrity of the pipeline in its entirety. In area monitoring SHM, methods such as visual or thermal cameras are limited to surface breaking defects which are unable to quantify thickness from the inner pipe wall. Vibration techniques are a possible method although they are insensitive. This leaves acoustic emission (AE) or guided waves (GWs). However, AE requires continuous monitoring without interruption due to the nature of the defect propagation that is random and irreversible. This has implications on battery life and therefore the use of GWs provide a more suitable approach.

Guided Waves (GWs) are bulk ultrasonic waves that interact with the boundaries of the structural wall and travel along the length of the structure. Most cases of GWs are confined to the low frequency range, where attenuation is low, allowing long sections of structures, e.g. pipelines to be screened. Although this method allows a quick assessment of a large area with defects, it is difficult to obtain the information on the size of the defect. Alternatively, a pitch-catch arrangement between two arrays of transducers, can be combined with tomographic principles, over a shorter distance, to provide more accurate sizing of defects. This concept, referred to as GW tomography (GWT), and still poses challenges for implementation in industry.

The strong attenuation of fundamental modes in liquid loaded pipelines and the need to use voltage due to the intrinsic safety of operation around flammable pipelines have significant implications on the need for an improve Signal to Noise Ratio (SNR).

To increase the Signal to Noise Ratio (SNR) and increase the effective distance between arrays, stacking of PZT discs are investigated and the parameters investigated to optimize the design are the mode amplitude, bandwidth, and mode purity. It was observed that the designed transducer with 10 PZTs stacked provides the optimum central frequency, nearest to 0.5 MHz·mm for GWT applications. However, the bandwidth decreases with increasing height of the stack. This drop in bandwidth with increasing number of PZTs stacked is compensated for by the addition of an Alumina composite backing mass.

While increasing mode amplitude provides the possibility of using a lower voltage of excitation, a deconvolution technique using a Linear Frequency Modulated (LFM) signal is investigated to improve the SNR. The technique involves excitation of an LFM signal and deconvolution of the response to a toneburst. While this method has been previously used to optimize the number of cycles and frequency of tonebursts used, the application here is to reduce the voltage. During this study however, a trend of increasing input SNR with a plateauing output SNR is discovered. This is achieved by using the inherent convolution of a rectangular filter with the chirp frequency response. It was found that as the input SNR of the excited LFM signal increases, the noise after deconvolution (attributed only to the ringing effect) increases with the increased signal amplitude, resulting in an unchanging SNR. This area can be used to lower voltage while maintaining SNR.

Finally, this thesis discusses the reliability and integrity of the application of the designed stacked transducer and lowering of voltage through deconvolution in an on-site application to validate the findings. Full Waveform Inversion (FWI) based GWT is used to provide a reconstruction of measured signals and defects on the gas pipeline were monitored for a period of 5 months. A simple statistical model is then incorporated to monitor the growth of selected defects beyond the 5-month period to predict the possible failure or when intervention techniques such as lowering of pipeline pressure is needed to extend the life of the pipeline.

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Chapter 1: Introduction

1.1 Background

Pipelines are considered critical components in most industries. For example, pipelines are used as integral parts in different streams in the oil and gas industry; they are utilised to extract products from a well, for multi-phase separation in the upstream, transportation and storage in the midstream and refinement in the downstream. With the energy needs of the industry set to increase in the coming years [1], pipeline sustainability and integrity have become key focus areas in the past decade due to numerous pipeline ruptures that have led to income and environmental losses, property destruction and even death [2]. Pipeline failures result in significant environmental impacts even in low-risk scenarios where pipelines operate outside the vicinity of human activities. High-risk pipelines operating near human activities can cause death in certain cases. These failures are often a result of malfunction, corrosion, material and weld failures or incorrect operation [3, 4]. Corrosion is regarded as the major cause of pipeline ruptures [4, 5] and tends to occur at pipe supports and bends, as shown in Fig. 1.1.

Figure 1.1: Corrosion on pipelines. The major corrosion hotspots in pipelines are found near the pipe features [6].

The use of pipeline inspection gauges or pipeline intervention gadgets (PIGs) is regarded as a non-destructive testing (NDT) method that has been successfully employed in detecting corrosion in pipelines. PIGs are cylindrically shaped gauges that use either ultrasound (UT), magnetic flux leakage (MFL), eddy current or all three simultaneously [7-9]. However, normal operations need to be shut down to allow PIGs to run within the pipe, as shown in Fig. 1.2. Moreover, several pipelines are ‘unpiggable’ due to the lack of pressure in pushing the PIG through the pipeline, back-to-back elbows and tight elbow bend radii [10].

Figure 1.2: Schematic of a PIG inside a pipeline [11].

With most designed pipelines nearing the end of their life cycle [1], the industry focus has shifted to acquiring new methods of detecting problem areas and providing solutions to maintain the integrity of ageing pipelines through life extension without the disruption of normal operation. For example, in the North Sea, most pipelines designed for 20–25 years have been in operation for 20 years [5, 12]. This situation has resulted in the need for highly reliable monitoring of pipelines in a paradigm referred to as structural health monitoring (SHM). In comparison with conventional NDT, SHM uses permanently attached sensors to enable measurements and therefore provides a more reliable indication of the change in structures [13].

The rationale behind the replacement of conventional NDT with SHM is driven by various factors and can be summarised as follows [14]:

- To improve the reliability of measurements by removing coupling variability through permanent sensors.
- To reduce the costs of obtaining repeated access to measure points in pipelines through scaffolding or stripping of insulation, etc.
- To continuously monitor problem areas with non-uniform degradation.

Pipeline integrity management (PIM) encompasses the entire process of assessing pipeline integrity. The process is achieved by identifying defects and monitoring defect growth through SHM. Subsequently, mitigation techniques are implemented through repair or pressure downrating for life extension. In PIM, the ability of pipeline operators to implement prudent strategies for pipeline life extension depends largely on the accuracy and reliability of inspection methods. This notion further explains the shift from conventional NDT to highly reliable SHM.

Frameworks for inspections have been developed over the years on the basis of operational experience, corporate mandate and asset management, which differ from operator to operator. A common feature of these frameworks is regular monitoring or predetermined inspection intervals [1], and they all revolve around accurately identifying the maximum depths of defects due to corrosion. Defect depth is a key parameter that determines the current state of health of pipelines and their remaining residual life.

Inspection frameworks for pipelines that use SHM can be categorised into two main groups: (1) local monitoring and (2) area monitoring [13]. Local monitoring involves point monitoring and encompasses a range of ultrasonic and electromagnetic NDT methods. These conventional non-destructive evaluation (NDE) methods can be regarded as SHM by permanently attaching sensors onto a pipe at spaced intervals along the pipeline length. Local monitoring is suitable only when the possible critical areas are known and the degradation of corrosion is uniform throughout the pipeline structure, such that point measurements provide a sufficient assessment of pipeline integrity. Area monitoring is preferred in cases where critical parts are unknown, the degradation areas are small and the sample points are too few to provide a reliable estimate of defect severity [13].

Area monitoring in SHM requires sensors to provide a significant area of coverage per sensor. Visual and thermal imaging are possible methods, but as explained in [13], they are confined to surface-breaking defects or near-surface defects, respectively. Another method that could be used is vibration monitoring, which is generally insensitive to the detection of localised damage [13]. This situation leaves two possible methods, namely, acoustic emission (AE) and guided wave (GW). AE involves testing of stress waves generated by a propagating crack or defect growth. These stress waves propagate as

GW. However, AE occurs randomly in time and is intermittent and irreversible [15]. Any monitoring with AE must be continuous, which presents implications on the battery life of permanently installed systems. In contrast with AE, GW can be switched off and on at regular intervals.

GWs are bulk ultrasonic waves that interact with the boundaries of the pipe wall and travel along the length of the pipe wall as opposed to conventional UT (Fig. 1.3). Successful commercial applications have been developed in rail and pipeline applications by using deployable arrays coupled with the various methods available for accurate sizing; this approach allows for fast screening to determine the loss in the cross-sectional area [16-18]. The use of GW technology has provided an alternative for point inspections to identify problem areas whilst reducing the down time of operations.

Figure 1.3: Conventional UT wave path (left) and ultrasonic GW path (right).

Most of the applications of GW in the current industry are localised in the low-frequency range where the attenuation is low, thereby allowing long pipeline sections to be screened. Although this method allows a quick assessment of a large area with corrosion defects, the accuracy of defect sizes is too low for practical implementation with industry guidelines, such as DNV-RP-F101 [19].

Current GW long-range screening systems cannot provide accurate estimations of defect depth in the presence of complex morphologies [20, 21]. Meanwhile, the constant group velocity method [22, 23] provides a medium-range solution over a distance of several pipe diameters for a pitch catch arrangement of a pair of transducers; however, this approach lacks spatially localised information to quantify the maximum defect depth of irregularly shaped defects. In the 1990s, Hutchins et al. [24] realised that combining multiple transmissions between transmitters and receivers and tomographic techniques increases the amount of spatially localised information. This concept is now referred to as GW tomography (GWT).

Several imaging algorithms have been investigated for GWT over the years. These algorithms include ray tomography [25], diffraction tomography (DT) [26], hybrid algorithm for robust breast UT tomography (HARBUT) [27] and full-wave form inversion (FWI) [28]. RT ignores diffraction, which severely limits its resolution. DT is limited by linear scattering models, such as Born or Rytov approximation [20]. Although this method is suitable for low-contrast defects where the phase is small and the noise is low due to phase unwrapping, it is limited for high-contrast defects where Born approximation is violated. HARBUT and FWI provide a good resolution. The strong attenuation of fundamental symmetric and asymmetric modes in liquid loading compared with non-loaded cases at low frequencies still poses a challenge due to the low signal-to-noise ratio (SNR), but quantitative estimations of defect depth using FWI have been successfully demonstrated by Rao et al. [29].

Inversion in the low-frequency range is accurate due to increased scattering from corroded surfaces at high frequencies, which limits the propagation distance for inspection [30]. The attenuation is also large at high frequencies. Although the range of GWT is limited in distance, GWT can be used in pipeline areas that are known to

undergo the largest amount of corrosion, which can be used as a sample location to map the integrity of the entire pipe network through hardware application in generating GW and software application in post-processing signals. A schematic of a short-range monitoring system to be used with GWT is shown in Fig. 1.4.

Figure 1.4: Schematic of GWT application in pipelines consisting of a hardware portion with GW generation and a receiving system and a software portion applying tomographic principles.

As shown in Fig. 1.4, the receiving and transmitting arrays are placed in the area of interest. GWs are excited by the transmitting array and received by the receiving array. The pipeline with a diameter that is much larger than its wall thickness can be unwrapped, transforming the two ring arrays into two parallel linear ones. The configuration shown in Fig. 1.4 is a typical cross-hole tomography setup that has reduced spatial information due to its limited view configuration. For pipelines, this limited view can be addressed with the use of high-order helical modes (to be elaborated in the literature review). The motivation for the improvement of the system in Fig. 1.4 is presented in the following section.

1.2 Thesis Motivation

The first motivation for this thesis is related to the effective range and reconstruction accuracy of GWT, which is sensitive to the SNR of measured signals. This situation makes SNR improvement desirable. At a low frequency, SNR is lower than that in the high-frequency case for two reasons. Firstly, the attenuation is acceptable for gas carrying pipelines, but an increase in attenuation in liquid-loaded cases results in a decreased SNR and monitoring distance. Short-range monitoring with GWT can be implemented on liquid-loaded pipelines by incorporating complex velocities into the frequency domain [31, 32]. However, the strong attenuation of selected GW modes requires SNR improvement for reconstruction accuracy. Secondly, low-frequency UT transducers using piezoelectric materials have low amplitudes. This notion is explained by the fact that a wave will transmit a large amount of energy per second if it has a high frequency simply because numerous waves pass at any given period. At high frequencies, the transducer that is needed to generate the out-of-plane mode should have an estimated diameter of 3–5 wavelengths with respect to the wave being generated [17, 33]. However, at low frequencies, the wavelength is too large to make this scenario practical, especially for applications on pipes. A typical commercial low-frequency transducer is shown in Fig. 1.5. The transducer is too large to be used in an array implementation to satisfy the Nyquist criterion, where the wavefields are required to be sampled at spatial intervals of not more than $\lambda/2$ [34]. However, point sources have been shown to be sufficient in generating the out-of-plane mode at low frequencies. However, point sources result in a low mode amplitude because of their small surface area of contact. A low mode amplitude translates to a low SNR. The improvement of mode amplitude is therefore an important design consideration in increasing the effective range and reconstruction accuracy and although work has been done to investigate

improved mode amplitude, the amplitude can be improved further. This is especially important considering the need to have lower excitation voltages which is described in the next motivation.

Figure 1.5: A commercial low-frequency transducer with a central frequency of 100 KHz.

The second motivation of this thesis is the need to use low excitation voltage on-site. The effective range of monitoring and reconstruction accuracy can be improved by increasing the excitation voltage. However, industrial applications prefer low voltages for several reasons. In a monitoring case wherein a power source is unavailable, battery life is critical and could be extended by low excitation voltages. Pipelines carrying flammable loads or located in high-risk areas also require low excitation voltages as a general standard of safety practice. However, low-voltage generation imposes significant challenges. The low SNR of the measured signals is one of the main challenges in low-voltage generation and current methods to retain the SNR after voltage reduction has not yet been investigated.

The last motivation of this thesis is to assess the reliability of the designed GWT system with improved SNR at a low voltage and provide a methodology for predicting the

probability of failure with industry guidelines, such as DNV-RP-F101 [19]. Most guidelines provide uncertainty to common NDT methods but none to GWT because of the lack of data testing on a series of pipe wall thicknesses. This situation limits the uncertainty estimation in calculating the allowable pipeline pressure undergoing corrosion. However, Rao et al. [35] provided sufficient error analysis for defect depths on a 6–10 mm wall thickness, which was used to estimate the uncertainty in measurement accuracy.

1.3 Objectives

The following thesis objectives are formulated to address the associated issues highlighted in the motivation.

1. Design a novel transducer that is effective at low frequencies and low voltage inputs.
2. Implement a novel method by applying a linear FM-coded signal with a signal deconvolution algorithm to improve SNR with a low input voltage.
3. Develop a methodology to implement the GWT system for assessing pipeline structural integrity through a widely adopted industrial guideline.

1.4 Outline Structure

This thesis has six chapters.

Chapter 2 provides a literature review on GWT and low-frequency transducers. The reasons for choosing the operating frequency, wave mode, imaging algorithm and signal processing technique are reviewed.

Chapter 3 presents parametric studies for designing a low-frequency transducer with optimised mode amplitude, bandwidth and mode purity. This chapter also presents the

operating frequency range of the transducer in GWT applications. An epoxy alumina composite backing mass is explored to widen the bandwidth for a range of frequencies for use with GWT.

Chapter 4 presents the design of coded excitation of ultrasonic GW signals by using linear frequency modulation (LFM) to improve the SNR for low voltage excitation. Firstly, the use of signal deconvolution as a processing technique to improve SNR is presented. Secondly, a novel study is carried out to demonstrate a nonlinear SNR improvement after deconvolution. The proportional relationship between the input SNR and voltage is used to reduce the input voltage of the GW signals whilst maintaining the same SNR.

Chapter 5 analyses the system stability of the monitoring performed over several months on a pipeline under severe corrosion conditions in an industrial site. A risk assessment based on a corrosion guideline and probabilistic estimation of defect growth is carried out to provide a useful analysis for rectifying and optimising maintenance schedules. The uncertainty of GWT based on the accuracy of the results and the defined resolution limit is discussed, and a conservative uncertainty is implemented in a predictive allowable corrosion model [19] to provide a new approach for implementing GWT in a PIM strategy. Lastly, the probability of failure and downrating pressure is presented.

Chapter 6 concludes this report by providing a summary of the work in this thesis, the overall contributions and possible future endeavours in improving the designed system and its possible application with big data.

Chapter 2: Literature Review

This chapter starts with the review of the fundamental equations of ultrasonic GW propagation in waveguides. Tomographic principles and the choice of fundamental mode for use with GWT at a low frequency are reviewed. Subsequently, piezoelectric UT transducers are employed to generate this fundamental mode, and the associated issues in designing the transducer for a low voltage input are introduced. Afterwards, the use of pulse compression as a signal processing technique to improve SNR for a low voltage input is reviewed.

Ultrasonic GWs are bulk waves that travel within elastic layers that are subjected to boundary conditions [36]. They can be represented as the superposition of bulk waves that interact with the boundaries through reflection and refraction under a phenomenon similar to the total internal reflection seen in fibre optic cables [20].

The characteristics of GWs have been investigated in detail in extant literature [37-42]. Therefore, only fundamental equations are reviewed in this chapter.

2.1 GW theory in plates and pipes

Bulk waves can either be longitudinal (L) or shear (S) and characterised by the displacement direction being in the same direction as the propagation (L) or orthogonal to the direction of propagation (S), respectively. This results in a Helmholtz decomposition given by

$$\mathbf{u} = \nabla\phi + \nabla \times \Psi, \quad (2.1)$$

where $\mathbf{u}$ is the displacement vector, $\nabla\phi$ represents the gradient operator of a scalar potential (L component) and $\nabla\times\Psi$ represents the curl of the vector potential (S component). Substituting Eq. (2.1) into the governing Navier–Stokes equation for elastic waves results in

$$(\lambda' + \mu')\nabla\nabla\cdot\mathbf{u} + \mu'\nabla^2\mathbf{u} + \rho\mathbf{f} = \rho\ddot{\mathbf{u}}, \quad (2.2)$$

and disregarding body forces yields

$$\nabla^2\phi = \frac{1}{c_L^2}\ddot{\phi}. \quad (2.3)$$

The dilation motion ϕ governs the longitudinal waves and

$$\nabla^2\Psi = \frac{1}{c_S^2}\ddot{\Psi}, \quad (2.4)$$

where the rotational motion Ψ governs shear waves. Here,

$$c_L = \sqrt{\frac{\lambda' + 2\mu'}{\rho}} \text{ and } c_S = \sqrt{\frac{\mu'}{\rho}}.$$

In Eq. (2.2), λ' and μ' represent Lamé constants related to Young's modulus E and Poisson's ratio ν through Eq. (2.5). ∇ and ∇^2 are the gradient and Laplace operator, respectively.

$$\lambda' = \frac{E\nu}{(1+\nu)(1-2\nu)}, \text{ and } \mu' = \frac{E}{2(1+\nu)} \quad (2.5)$$

The general solution for Eqs. (2.3) and (2.4) is

$$\phi, \Psi = A e^{i(k_{L,S}z - \omega t)}, \quad (2.6)$$

where A is an initial wave amplitude constant, ϕ_0 and Ψ_0 are scalar and vector potentials and ω is the angular frequency. The longitudinal and shear wave numbers, $k_{L,S}$, can be expressed as

$$k_{l,s}^2 = \frac{\omega^2}{c_{l,s}^2}. \quad (2.7)$$

Substituting Eq. (2.6) into Eqs. (2.4) and (2.3) and coupling them with boundary conditions (Eq. (2.8)) result in frequency-dependent wave motions, which can either be symmetric or asymmetric depending on the particle motions through the thickness and the combination of the top and bottom surfaces reflecting L and S waves (Lamb A and S modes) or just the top and bottom S waves (SH mode) superimposing one another. These waves are referred to as GWs.

$$\sigma^t=0 \text{ and } \sigma^b=0, \quad (2.8)$$

where σ^t and σ^b are traction vectors given at the top and bottom surfaces.

Different guided modes, such as shear horizontal (SH), symmetric (S) and anti-symmetric (A), can be obtained for a material with a finite thickness, and the frequency dependencies of their velocities are referred to as dispersion curves. Further details can be found in [20, 42, 43].

Phase velocity is the velocity of a mode at a given frequency. Group velocity is the velocity of a group of waves propagating at similar frequencies and corresponds to the energy of a multi-frequency wave packet [43]. The phase and group velocities of the SH mode are shown in Eqs. (2.9) and (2.10) and can be solved analytically because the scalar potential term (Eq. (2.3)) disappears.

$$c_p(fd) = \pm 2c_s \left(\frac{fd}{\sqrt{4(fd)^2 - p^2 c_s^2}} \right), \quad (2.9)$$

$$c_g(fd) = c_s \sqrt{1 - \frac{(p/2)^2}{(fd/c_s)^2}}, \quad (2.10)$$

where p is any non-negative integer and d is the thickness of the free plate. Different modes are obtained for different orders of p . Eqs. (2.9) and (2.10) indicate that the group and phase velocities of the SH_0 mode ($p=0$) are constant for all frequency–thicknesses.

However, the dispersion relations for Lamb modes are complex and solved numerically. A typical dispersion curve for Lamb waves in a 10 mm aluminium plate solved by a numerical software, DISPERSE [44], is shown in Fig. 2.1. S_n and A_n represent symmetric and asymmetric modes, respectively, for Lamb waves, and the subscript n refers to the mode order. For each mode, whilst the frequency approaches infinity, the velocity approaches the Rayleigh wave velocity of the plate.

Figure 2.1: Dispersion curves of Lamb waves in a 10 mm aluminium plate: phase velocity (a) and group velocity (b).

GWs in pipes are similar, and several studies have explored their similarities [45, 46]. The curvature of the pipe can be disregarded, and the unwrapped pipe can be analysed as a plate when the conditions are satisfied [45]. The conditions are presented in Eq. (2.11).

$$R \gg d \quad (2.11a)$$

$$2\pi r \gg \lambda \quad (2.11b)$$

The frequency–thickness dependence of the GW velocities observed in the dispersion curves is the basis of tomographic algorithms using the forward model [20, 28, 43]. Observation of the 2D wave equations shows that the behaviour of GWs can be approximately described and interpreted to produce wall thickness loss maps over several pipe diameters. This task is achieved through the implementation of tomographic algorithms that consider refraction and diffraction effects through a model-based inversion method [27] referred to as GWT.

2.2 Guided wave tomography (GWT)

This section provides a brief review of literature on several primary tomography principles for thickness mapping.

2.2.1 Ray Tomography

The literature for tomography has two main segments. The first one is medical imaging, which applies a filtered back projection algorithm. Convolution back-projection algorithms are suitable for only two scanning geometries. The first one is parallel beam projection (Fig. 2.2(a)), and the second one is fan beam projection (Fig. 2.2(b)).

Figure 2.2: (a) Parallel beam projection and (b) fan beam projection.

The second segment of tomography is geophysical imaging, in which parallel or fan beam projection is impossible to utilise. Consequently, the development in this field has focused on the use of iterative methods. An example of an iterative method is reviewed in Section 2.2.1.2.

2.2.1.1 Convolution Back-projection Algorithm

The filtered back-projection algorithm developed for straight ray medical tomography assumes that ray paths are not influenced by inhomogeneities in the specimen. The algorithm proceeds in two steps: filtering and back projection. The algorithm applies the Fourier slice theorem [47], which is valid for the parallel beam theorem and only in the case of straight ray tomography. The Fourier transform of a parallel projection is limited to a bandwidth of Ω , which is a function of the sampling in the spatial domain through

$$\Omega = \frac{\pi}{\Delta t}, \quad (2.12)$$

where Δt is the distance between each transducer. A schematic of the theorem is shown in Fig. 2.3.

Figure 2.3: Schematic of the Fourier slice theorem.

The projections are calculated at as many angles of θ as possible to obtain an estimate of the 2D Fourier transform of the object. The object is then recovered by taking the inverse Fourier transform of this estimate.

Parallel projection scanning geometry has been used effectively by several researchers [48-50]. However, parallel beam projection presents a challenge for fixed transducers in SHM. The required projection angles of θ are typically obtained by rotating either the object or the transducers. The required number of transducers for a fixed array is impossible to implement. Likewise, parallel scanning geometry is impossible to implement on pipelines.

2.2.1.2 Iterative Algorithm

Iterative algorithms can be used as alternatives. Iterative algorithms were introduced by Hounsfield in 1972 for the first X-ray computed tomography scanner [51]. These

algorithms were later developed for geophysical imaging. A simplified case of an iterative algorithm consisting of one transmitter and one receiver is presented below. A mesh of cells is drawn between the two transducers to find the ray path, as shown in Fig. 2.4.

Figure 2.4: Mesh of squares generated between a transmitter and a receiver in iterative algorithms.

If the ray does not pass through the cell, then the velocity is set to zero. If the velocity of the ray is assumed constant in each cell, then the time of flight of the wave is given by

$$T_{\text{estimate}} = \sum_{i,j} \frac{d_{i,j}}{V_{\text{gr},i,j}}, \quad (2.13)$$

where d represents the cell length and V_{gr} is the group velocity within the cell. The time of flight estimate, T_{estimate} , is then compared with the measured value, and the

group velocity in each cell is updated until the error between the estimated and measured time of flight is minimised. The principle of this iterative algorithm is outlined with the time of flight principle, but it can be applied with other parameters used in tomography, such as attenuation and frequency centroid shift [24, 52].

Mckeen and Hinders [53] investigated the difference between parallel projection scanning using filtered back projection and cross-hole scanning (Figure 2.5) with an iterative processing method. Cross-hole scanning is a practical approach for SHM in pipelines.

Figure 2.5: Cross-hole scanning geometry for transducer arrays. The signals are excited by one source and measured by all the receivers across a region of interest.

The advantage of cross-hole scanning geometry is that it only requires access from two sides of the region of interest. However, ray density varies within this region, and the rays do not pass through the region of interest from all orientations compared with the rotating parallel beam scanning geometry [43]. This limited view problem reduces reconstruction quality.

However, reconstruction accuracy can be improved. Frequency compounding has been found to improve image quality by enhancing the contrast between the defect region and its surroundings [54, 55]. This enhancement is achieved by reconstructing the image for different input signals centred at multiple frequencies.

For straight ray tomography, ray theory is valid only when the size of the object, d , is larger than wavelength λ of the signal and width L_F of the first Fresnel zone. When ray theory is invalid, scattering from the object superposes with the direct ray at the receiver. This condition modifies the arrival time and amplitude of the received signal. At the midpoint between the receiver and source, the Fresnel zone is estimated to be $\sqrt{\lambda L}$. A schematic of the scattering effect is shown in Fig. 2.6.

Figure 2.6: Schematic of the scattering effect from the object.

Three approaches can be considered when ray theory is violated. The first approach is decreasing the propagation distance to reduce the Fresnel zone. The second approach is to increase the defect size. Thirdly, the operation frequency can be increased until ray theory is valid. These approaches are typically not of interest in practical applications.

Alternatively, algorithms that consider scattering can be used in the low-frequency region.

2.2.2 Diffraction tomography (DT)

Diffraction tomography (DT) considers the scattering of the wave from the object in reconstruction. The scattering wave is produced by the interaction of the incident wave with the object. Straight ray tomography and DT are equivalent when the wavelength approaches zero [26].

Figure 2.7: (a) Array of transducers receives wavefields that propagate in a straight line from the source through the object. (b) The incident wavefield from the source produces scattered wave fields when interacting with the object. The array of transducers receives the total wavefield, which is the superposition of the incident wavefield from the source and scattered wavefields.

However, the phase and amplitude of the scattered field (Fig. 2.7) is used in DT reconstruction, as opposed to the total field in straight ray tomography. The array of transducers measures the total field shown in Fig. 2.8(b). An approximation of the wave

equation is then utilised to calculate the scattered field. Born and Rytov approximations are commonly used, and their expressions are presented as Eqs. (2.14) and (2.15), respectively.

$$U_a = U_t - U_i, \quad (2.14)$$

$$U_a = U_i \log \frac{U_t}{U_i}, \quad (2.15)$$

where U_a is the scattered field, U_t is the total field and U_i is the incident field of the wave propagating in the region of interest when inhomogeneities do not exist.

Figure 2.8: (a) An array sends an incident wave from every direction, r_0 , and the scattered field is detected from any direction r . (b) 2D K space showing the mapping of the scattered field measured in direction r from the incident wave direction, r_0 . The solid and dashed semicircle arcs correspond to the transmitted and reflected directions, respectively [43].

The scattered field is processed in the spatial frequency domain and denoted as K space (Fig. 2.8). The scattered field in direction r is mapped in the K space in the direction $k(\mathbf{r} - \mathbf{r}_0)$, where k is the wavenumber at the frequency of interest. A disk with radius $2k$ is

populated with information corresponding to a transducer source change because the excitation direction, $\mathbf{r}_0$, varies. This disk is referred to as the Ewald limiting disk [56].

Transmission mode DT populates the disk with radius $\sqrt{2k}$, and reflection mode DT populates the disk between $\sqrt{2k}$ and $2k$ in the K space [57]. This condition implies that the reflection mode contains components with a high spatial frequency. The loss in low-frequency components results in an accurate reconstruction of the defect boundary only [58]. The cross-hole configuration provides similar results as reflection mode DT due to missing information from a limited view configuration [59]. As mentioned in the previous section, this condition can be improved by using certain methods, such as frequency compounding.

DT considers the effects of diffraction but is limited by linearised scattering models, such as Born and Rytov approximations, without considering multiple scattering effects. Born approximation is suitable for low- and small-contrast defects where the phase shift is small, and Rytov approximation is limited to simple and low-contrast defects due to phase unwrapping constraints [60]. Other hybrid algorithms that use a combination of ray and DT, such as HARBUT, are also limited by Born approximation.

2.2.3 Full Waveform Inversion (FWI)

A model that considers multiple scattering was recently applied in corrosion thickness mapping [28, 29, 35, 61-63]. Full waveform inversion (FWI) using the iterative method was first used in geophysics for seismic imaging [64]. FWI improves reconstruction by minimising residuals between modelled and observed data over multiple frequencies, as shown in the schematic below.

Figure 2.9: Schematic of the FWI model. The model consists of two loops that are iterated either by setting a minimum convergence of error between the measured and modelled wavefields or setting a maximum number of iterations.

In FWI, several models match the data with reasonable tolerance due to the local minima in the misfit function, which can be eliminated by linearising the inverse model through the forward operator [65]. Non-unique solutions are also caused by cycle-skipping, which can be mitigated through multiscale approaches [66].

The frequency step for iterations in loop 2, which is shown in Fig. 2.9, is selected according to the wavenumber k distribution of the scattered wavefield in DT [67]. Transmission wavenumbers are limited to the $\sqrt{2}k_1$ radius in the K space. This factor is always included in a large wavenumber radius at the next frequency iteration f_2 within the circle of radius $\sqrt{2}k_2$. Fig. 2.10 shows the K space of two consecutive frequency iterations. The relationship at two consecutive frequency iterations (f_1 and f_2) should follow $k_2 \leq \sqrt{2}k_1$ to account for the reflection wavenumbers.

Figure 2.10: k_1 is the wavenumber at a low frequency, and k_2 is the wavenumber at the next frequency for iteration.

2.2.4 Current Limitations of GWT and Improvement of SNR

The resolution is generally limited by the wavelength used based on the diffraction limit of $\lambda/2$ [68, 69]. This limit originates from the assumption on the inversion of the Helmholtz equation, which assumes that GW modes behave as acoustic waves (phase and group velocities) associated with local thickness. The resolution decreases to $1.5\text{--}2\lambda$ in GWT algorithms, such as DT and FWI, for test data that use an elastic model. Although the plates are assumed to be uniform, it is not always the case. However, any varying thickness change is assumed to be gradual. Considering that thickness is also the parameter being measured, this situation introduces a limitation, beyond which the equation becomes invalid. This limit is related to the wavenumber and associated wavelength.

Resolution is commonly defined in GWT as the maximum wavelength of a wave number component that is accurately reconstructed. Resolution can be improved by

several approaches. In Born approximation, the forward scattering models disregard the effects of multiple scattering, thereby linearising the model. In inversion, the object function can be determined from the information within the Eswald limiting disk by setting the information outside to zero. This regularisation can extrapolate the object function and eventually lead to an improved resolution [70]. However, analytical continuation is sensitive to noise; thus, a stable solution to the inverse problem cannot be provided for measurements with low SNR [71-73]. Hence, SNR should be enhanced to achieve an improved resolution.

The use of high-order helical modes to provide substantial spatial information is another alternative to increase the reconstruction accuracy in a pipe structure. This method involves helically separating high orders of wave propagation in pipelines by extending the unwrapped pipe in the positive and negative x-directions (Fig. 2.11). The data acquisition time should be considered to obtain information on high-order helical modes that will arrive much later. Another challenge in using high-order helical modes is the increased attenuation. High-order helical modes have a lower SNR than the first arrival due to the long propagation distance. Therefore, improving the amplitude of A_0 mode transduction increases the SNR and accuracy of reconstruction by enhancing the SNR of high-order helical modes.

Figure 2.11: (a) 3D physical model showing the transmitter and receiver arrays and high-order helical paths; (b) order 0 is the unwrapped section of the pipe, and other orders are added to both directions along the x axis to replicate the cyclic characteristic of the pipe. In the pipe, the measured signals at R_0 correspond to the superposition of the signals from each helical path on the plate. The arrows present various propagation paths between a transmitter and a receiver [61].

Artefacts due to noise may be present in the reconstruction image of experimental measurements. For example, FWI uses the iterative method, and reconstruction images from completed experimental measurements [28, 61, 62] show some artefacts that could have been due to several reasons, such as measurement positioning error and slight anisotropy of the material [43]. However, the presence of these artefacts can also be attributed to noise, which again makes the improvement of SNR desirable.

Another limitation is that the uncertainty in measurements for GWT must be recorded for different pipe wall thicknesses, defect sizes and operating frequencies to predict the

allowable pressure in pipelines as described by industrial standards, such as DNV-RP-F101 [19]. The guideline requires the uncertainty to be approximated in predicting the allowable corrosion pressure. Two inspection methods are presented in the guideline: MFL and convention ultrasonic testing (UT). When applying GWT in a scenario where data uncertainties for a range of wall thicknesses and defect sizes have not been provided by previous authors, estimation of these uncertainties should be carried out with care and caution to ensure that they are neither conservative nor too moderate. This notion will be further elaborated in Chapter 5.

2.3 Frequency and Mode Selection for GWT

In GWT, a steep velocity gradient of the mode in the dispersion relation is advantageous because a small change in wall thickness results in a large velocity change in the mode. This relation between the sensitivity of mode velocity to wall loss allows an accurate evaluation of corrosion damage even under various experimental uncertainties, such as noise [27]. The frequency is limited to 2 MHz/mm to avoid high-order modes, which would otherwise complicate the received signal. This frequency region has three guided modes, namely, SH_0 , S_0 and A_0 . However, only S_0 and A_0 are dispersive and have thus been the focus in GWT.

Belanger [43] identified the following areas for the selection of the optimal ultrasonic guided mode for tomography: (1) excitability and detectability and (2) attenuation due to fluid loading. Excitability corresponds to the ratio between surface displacement in a given direction and surface excitation force in the same direction and location (referred to hereafter as mode purity) [43]. Excitability can be separated into two excitation directions. The first considers out-of-plane excitation force, and the second considers

in-plane excitation force. When normal excitation force is used, the A_0 mode is excited with an amplitude of 40 dB above S_0 in the dispersive frequency range of A_0 (0.45 MHz.mm to 0.55 MHz.mm). In the dispersive frequency range of S_0 (1.5 MHz.mm to 2 MHz.mm), however, A_0 and A_1 modes are excited by normal force with a high amplitude comparable with S_0 . Consequently, the propagation distance must be sufficiently long to ensure that the signals separate and the S_0 mode is pure. When in-plane excitation force is applied, the amplitudes of A_0 and S_0 modes are comparable in the dispersive frequency range of A_0 and S_0 , respectively. Fig. 2.12 shows the excitability of A_0 , S_0 , A_1 and S_1 modes from out-of-plane and in-plane excitation.

Figure 2.12: (a) Out-of-plane/normal and (b) in-plane excitability of A_0 , A_1 , S_0 and S_1 lamb wave modes for a plate [43].

This situation implies that a transducer exciting the A_0 mode with normal force is preferable in obtaining the optimal mode purity for convenient signal processing and to prevent overlapping modes from corrupting the first arrival signal.

Huthwaite [27] correlated the reconstructed thickness error to the velocity gradients by comparing the mode sensitivity to thickness change with the insensitivity to the sources of uncertainty to select the optimal mode. Small errors from the reconstructed thickness were obtained for the A_0 mode at all remnant thickness values but only for shallow defects in the S_0 mode [27]. This feature is advantageous in implementing GWT for corroded pipe guidelines in industries [19, 74, 75] by providing an accurate allowable corrosion pressure during inspection and a probability of failure with small uncertainty.

Figure 2.13: Effects of attenuation on fundamental Lamb modes for a 10 mm steel plate embedded in water [76].

Belanger and Cawley [30] and Huthwaite [76] showed that when no fluid loading exists, the A_0 mode should be used. However, A_0 is attenuative at a low frequency with fluid loading. The A_0 mode attenuation is approximately 37 dB/m with respect to a 10 mm steel plate at 0.5 MHz.mm, as shown in Fig. 2.13. This means that after 1 m, the

amplitude of the A_0 mode will be equivalent to that of the S_0 mode if both modes are excited with the same energy. The attenuation of the A_0 mode limits the distance between the receiver and transmitter when the A_0 mode is excited. Improving the transducer excitability for the A_0 mode could make it an acceptable choice, even in the liquid-loaded case.

In this thesis, the A_0 mode is used in the frequency operating at approximately 0.5 MHz.mm. Deconvolution is adopted to improve the transduction of the A_0 mode by increasing SNR. This step allows the high-order helical modes, which would otherwise be buried in noise, to be used in reconstruction.

With the inversion algorithm, mode and operating frequency selected, the next few sections cover the literature on hardware improvement. Firstly, the literature on the current low-frequency A_0 mode UT transducer is reviewed. Secondly, possible signal processing techniques to improve the SNR for low-frequency A_0 mode transduction are discussed.

2.4 Piezoelectric Transducers for Low-frequency Applications

Successful systems for short-range monitoring must be able to excite the selected mode with high mode purity so that signals may be interpreted reliably. Lowe and Cawley [77] reported that test sensitivity is a function of the signal-to-coherent noise ratio, which means that the generation of unwanted modes should be reduced because they produce coherent noise in the system.

One of the traditional ways to generate GW modes at a given frequency is to have periodically varying forces or displacement within the transducer, with the period equal to the desired wavelength; examples of this are piezoelectric transducers, certain

EMATs and interdigital transducers [17, 33]. This thesis focuses on the use of piezoelectric transducers.

At low frequencies, particularly below the A_1 mode cut-off frequency, wavelength controlled transduction is unnecessary, and a point source is sufficient to generate the anti-symmetric (A) and symmetric (S) modes [78].

The use of a small dimensioned A_0 low-frequency UT transducer has previously been investigated [78], with the transducer geometry selected based on the purity and amplitude of a particular GW and its frequency bandwidth. However, the transduction can be further improved. Increasing the A_0 mode amplitude without sacrificing the mode purity is desirable to improve transduction under high attenuation in certain cases, such as fluid loading.

The downside of point sources is their lower mode amplitude compared with that in conventional transducer designs. Low-frequency transducers are typically too large to generate acceptable mode amplitudes. A small transducer geometry should be maintained to satisfy the Nyquist criterion. The criterion states that wavefields must be sampled at spatial intervals not more than $\lambda/2$ [34]. In a cross-hole geometry on a pipeline, two arrays of transducers are implemented. This feature implies that each transducer should be small to provide sufficient spatial information for defect reconstruction within the monitoring region.

A way to keep the transducer geometry small whilst providing a large mode amplitude is to stack PZTs together and ensure in-phase vibration based on the d_{33} effect [79]. This concept will be further elaborated in Chapter 3. A schematic of a stacked PZT transducer is shown in Fig. 2.14. The diameter of each PZT is 10 mm with a thickness of 2 mm. Such transducers have been successfully implemented in AE, vibration control and

energy harvesting applications [80, 81]. However, the use of stacked PZTs in GWT is focused on improving the mode amplitude whilst minimising the transducer geometry and keeping the operating frequency of the transducer within the dispersive range of the dispersion relation. An increased mode amplitude increases the propagation distance in fluid loading and provides an improved SNR on high-order helical mode waveforms whilst maintaining mode purity (excitability).

Figure 2.14: Stacked PZT excited in parallel to ensure in-phase vibration. The PZTs are stacked in the excitation direction. The excitation direction is denoted as the 3-direction.

The stacked PZT design provides increased amplitude in constant environmental noise. A possible issue for stacked PZT transducers is the associated dielectric loss at high voltages, which increases heat generation [82, 83]. A temperature increase in practical applications, such as pipelines carrying flammable contents, is unacceptable for safety reasons. The epoxy bondings between PZT layers are sensitive to temperature and may cause transducer degradation. Hence, low voltage excitations below 50 V are generally preferred.

2.5 Pulse Compression

Apart from a new transducer design to improve amplitude, a signal processing technique using chirp deconvolution can be employed to reduce the noise and voltage whilst maintaining the signal amplitude. A literature review of linear pulse compression and deconvolution is conducted below.

2.5.1 LFM Signal

Pulse compression is the correlation of any frequency modulated waveform with a matched or mismatched filter. The simplest case is the LFM signal (chirp), where its instantaneous frequency is a linear function of time. Fig. 2.15 presents a chirp signal with a bandwidth ranging from 30 KHz to 110 KHz and a tone burst signal with a central frequency of 70 KHz.

Amongst all compression waveforms, the LFM or chirp signal has been studied the most extensively due to its symmetric properties and ease of generation. Moreover, LFM is insensitive to frequency shifts, which is a desirable property in conventional UT imaging.

Figure 2.15: Chirp signal with a bandwidth of 30 KHz to 110 KHz and duration of 429 μ s. The chirp is 10 times longer than the 70 KHz three-cycle tone burst shown in a solid line.

LFM signals can be emitted by UT transducers and subsequently compressed to provide a similar axial resolution and contrast whilst increasing SNR by 10 dB to 20 dB [84]. Misaridis [84] showed that LFM signals have optimal and robust features for conventional UT imaging.

2.5.2 Issues Associated with LFM

Detectability and resolution are the two issues associated with the modulated signal. The former can be optimised by maximising SNR. The filter output should have a good axial resolution. Both issues can be addressed by applying a filter.

Sidelobes are introduced during pulse compression and lead to self-noise or self-clutter [85]. In LFM signals, Fresnel integral oscillations also contribute to sidelobes. These

integrals are inherent in the analysis of the frequency equations of the chirp signal and are stable when the time–bandwidth products of LFM signals are large. Fresnel integral oscillations will be further explained in Chapter 4. Large sidelobes result in a loss of information because imaged weak scatters are not distinguishable from large ones within their vicinity.

Sidelobes can be eliminated in two ways. Firstly, matched filtering can be used. A matched filter has the same response as the input signal, except for a loss factor k representing the attenuation of the input waveform and a translation in time representing the delay between the input and output signals. However, the matched filter is not optimised for the detection of scatterers in the self-clutter [84]. The sidelobes of the matched filter masks weak echoes. Consequently, designing suitable waveforms with low sidelobe properties in matched filtering is a challenge in coded signal imaging.

Alternatively, sidelobes can be suppressed through a mismatched condition. This task is achieved through windowing of the filter [84]. The effect of different windowing functions and their associated sidelobe properties can be found in [86]. A Hann window is typically used.

Figure 2.16: The SNR and width comparison of the mainlobe for the matched filter (auto correlation) and unmatched filter (Hann window) on a “Chirp” excitation.

The mismatched filter has two main drawbacks, which are shown in Fig. 2.16. The first drawback is the widening of the mainlobe. In UT, the widening does not exceed two times the width of the unmatched mainlobe even when a mismatched filter is applied; accordingly, an acceptable resolution is achieved in most cases [84, 87]. Secondly, mismatched filtering results in SNR degradation of approximately 1.5 dB to 3 dB compared with the matched case [84], which is acceptable in most SHM applications.

2.5.3 LFM in GWT

Resolution requirements favour the transmission of short pulses because a shortened pulse length provides good axial resolution for time of flight applications. Although short pulses can be used to satisfy the diffraction limit, they are low-energy pulses. This feature can be problematic, especially for UT transducers with a small-geometry low-frequency A_0 mode and low mode amplitudes. Moreover, the dispersive and multi-modal nature of GWs require that the excitations are of a wide bandwidth to begin with. Therefore, short pulses are preferable.

As previously mentioned, resolution requirements can be addressed with pulse compression. This notion implies that a long chirp pulse can provide high power for optimal SNR compared with a short pulse whilst achieving the resolution of a short pulse through pulse compression.

Alternatively, deconvolution is a method that can be used to retain the benefit of a high-powered long chirp signal with a resolution similar to that of a three-cycle tone burst. Deconvolution is achieved by using a long chirp for transmission and data collection and altering the received signal to that of a transmitted 3–5 cycle tone burst [88]. A schematic of the deconvolution process is shown in Fig. 2.17.

Figure 2.17: S_c is the transmitted chirp, S_{tb} is the transmitted tone burst, R_c is the received propagation of the chirp and R_{tb} is the tone burst propagation obtained from deconvolution.

In imaging algorithms where complex velocity maps are evaluated to reconstruct the wall thickness, deconvolution can increase SNR and remove artefacts due to noise.

2.5.4 Benefits of Deconvolution

Deconvolution provides an SNR improvement that is proportional to the length of the chirp used, as shown in Fig. 2.18 [88]. It also provides benefits apart from SNR improvement. The operating frequency and number of cycles for the excited tone burst can be optimised in one measurement through the generation of a chirp signal with a bandwidth range that includes the frequencies of interest. The measured chirp response can then be deconvoluted to tone bursts with different central frequencies and a varying number of cycles. An alternative to this process is the generation of tone bursts with different frequencies and number of cycles and obtainment of measurements for each case, which is time consuming.

Figure 2.18: The output SNR of the deconvoluted tone burst improves with a factor proportional to $\sqrt{N}$, where N is the ratio of chirp length to tone burst length [88].

Additionally, sending a long signal reduces the power requirements by allowing voltage reduction through signal lengthening. However, no study has investigated the relationship of the input chirp SNR with SNR improvement. The gating in the frequency domain intuitively results in a nonlinear change in SNR improvement because the input of the SNR chirp increases as noise becomes negligible. This situation could result in a limit on SNR improvement with changing voltage but could allow voltage reduction without any change in SNR at this limit. This work will be presented in Chapter 4.

2.6 Summary

On the basis of the literature mentioned above, the thesis discusses the optimisation of low-frequency, out-of-plane transducers to be used with GWT for the detection of corrosion in gas pipelines. The out-of-plane mode has been demonstrated [43, 60, 76, 78] as the appropriate mode for GWT in the frequency range of interest at 0.5 MHz.mm. Although FWI is utilised as an imaging tool in this work, a high-amplitude, low-frequency transducer is designed to be used in conjunction with any GWT algorithm. Low-frequency UT transducers with small diameters have been designed to provide transduction of acceptable mode purity [17, 33, 78], but the mode amplitudes are low and can still be improved especially when considering the need to reduce voltage on-site.

SNR is then further improved by using modulated signals through a deconvolution technique introduced by Michaels [88]. Although SNR improvement through deconvolution has been investigated, deconvolution is applied here to investigate the possibility of reducing voltage whilst maintaining the improved SNR. SNR improvement is especially important when using the A_0 mode transducer in fluid loading applications due to the higher attenuation of the A_0 mode compared with that of the S_0 mode at low frequencies. Additionally, the possibility of reducing the voltage through deconvolution decreases transducer degradation by preventing the large dielectric losses and heat generation inherent in a PZT stack. Deconvolution of the signals to a tone burst is implemented, and a voltage reduction analysis is presented in Chapter 4.

Chapter 3: Optimisation of Low-frequency Ultrasonic Transducers for Guided Wave Tomography

3.1 Introduction

Good transduction (large amplitude and high SNR) is required in GWT for accurate image reconstruction. Conventional transducer designs require the transducer diameter to be larger than the wavelength generated to produce narrow beams that provide detailed reconstructions. However, this situation brings significant challenges in the low-frequency region because the required transducer sizes are unsuitable for the monitoring of pipelines with small diameters as it needs to satisfy the Nyquist criterion, which requires wavefields to be sampled at spatial intervals of not more than $\lambda/2$ [89]. These large transducers are infeasible based on pipe curvature. By contrast, small diameter transducers have small transmission coefficients and thus generate low mode amplitudes. In the presence of a large amount of environmental noise, small-diameter transducers result in poor SNR, thus reducing the monitoring distance. A potential method to increase the generation amplitude whilst keeping the small size of the transducer is to use stacking PZT discs. In this study, out-of-plane motion is used for the generation of the A_0 -guided mode, and the reason for this selection has been discussed in Section 2.3. The following parameters are utilised to select an optimal design of the PZT stack to be used for GWT: A_0 mode amplitude, mode purity and transducer bandwidth.

A larger displacement can be generated in these stacks compared with a singular PZT disc. At low frequencies below 100 KHz, A_0 mode attenuation reduces the monitoring range because the mode attenuates and is buried in noise. The A_0 mode attenuation is large under liquid loading. Consequently, the improvement of mode amplitude increases

the propagation distance. The displacement of the A_0 -guided mode can be increased with an increasing number of PZT layers. However, an increase in the number of PZT layers decreases the resonance frequency of the stack. Consequently, the optimum number of PZT layers for generating the largest A_0 -guided mode amplitude with a resonance frequency close to 0.5 MHz·mm is investigated.

Mode purity is related to mode excitability at low frequencies. Further details can be found in Section 2.3. Mode purity is important for low-frequency, out-of-plane excitations. In the 0.5 MHz·mm region, the S_0 mode arrives earlier than the A_0 mode over short propagation distances. In the presence of defects, mode conversion modifies the received signal, thus complicating the processing of the measured signals. Therefore, the highest mode purity possible is desired. In stacked PZT transducers, the relationship of A_0 mode purity with increasing stack layers must be investigated, and the optimum value should be selected.

In GWT applications, the dispersive properties of the A_0 mode are used in the reconstruction of a defect image. Consequently, a wide bandwidth over the range of 0.5 MHz·mm is desirable. This chapter investigates the relationship between the number of layers of the PZT stack and its effect on the mode amplitude, mode purity and bandwidth of the stack frequency response.

The chapter starts with a brief outline of the theory of a stacked PZT transducer. Thereafter, a finite element (FE) method is described by applying electrostatic and solid mechanics modules in a commercially available software. Mode amplitude and purity are studied with respect to the stacking layers of PZT discs. Subsequently, backing mass dimensions and materials are investigated to widen the transducer bandwidth. The FE model is validated through carefully designed experiments. Afterwards, the transducer

limitation in GWT applications is discussed. The application of the designed transducer has been published in [63].

3.2 Stacked Actuator

The piezoelectric effect is a phenomenon wherein certain materials can generate an electric charge in response to an applied stress, and the effect is reversible. This phenomenon occurs in two ways: directly (a piezoelectric material generates voltage through an applied stress) or indirectly (an applied voltage produces displacement). In UT transducers, these generated vibrations allow for the transmission of ultrasonic waves into the medium. Piezoelectric materials act as transmitters and receivers based on direct and indirect modes. The most common of these materials is lead–zirconate–titanate, which is commonly referred to as PZT.

PZT vibrates with a wavelength that is twice that of its thickness at its resonance frequency. This characteristic poses an issue at very high frequencies, where the material becomes too thin and fragile. However, a thick and durable PZT can be used at low frequencies. A limitation of PZT is that it is sensitive to shear forces perpendicular to the polarisation direction. Consequently, PZT at low frequencies is not only thick in the polarisation direction but also wide to prevent shear breakage. The large dimensions of low-frequency transducers make them infeasible in a pipe array arrangement due to the pipe curvature. A method that can reduce the possibility of shear breakage is the use of flexure guides, which guide the vibration of PZTs when they are excited. This circumstance allows PZT discs to remain thick in the polarisation direction but narrow on the plane perpendicular to the excited direction. However, a reduction in PZT width reduces the charge generated. Alternatively, multiple layers of PZT with narrow widths

can be stacked together and excited in parallel to improve the amplitude of the generated wave. The theory of a stacked PZT actuator is presented below.

3.2.1 Piezoelectric Theory

The piezoelectric effect can be represented by the strain–charge equation, coupling Hooke’s law with electric field equations. These equations are readily available elsewhere, and only key equations are presented below for convenience.

$$D_i = e_{ij}^\sigma E_j + d_{im}^d \sigma_m \quad (3.1)$$

$$\varepsilon_k = d_{jk}^c E_j + s_{km}^E \sigma_m \quad (3.2)$$

In matrix form, Eqs. (3.1) and (3.2) can be presented as

$$\begin{bmatrix} \mathbf{D} \\ \boldsymbol{\varepsilon} \end{bmatrix} = \begin{bmatrix} \mathbf{e}^\sigma & \mathbf{d}^d \\ \mathbf{d}^c & \mathbf{s}^E \end{bmatrix} \begin{bmatrix} \mathbf{E} \\ \boldsymbol{\sigma} \end{bmatrix}, \quad (3.3)$$

where D is the electric field displacement (Coulomb/m²); $\boldsymbol{\varepsilon}$ is the strain vector; $\boldsymbol{\sigma}$ is the stress vector; e^σ is the dielectric permittivity (Farad/m); d^c is the piezoelectric coefficients (m/V) and s^E is the elastic compliance (m²/N) or inverse of stiffness. These terms consist of matrices that are presented in Appendix A. In PZTs, the poling axis is generally denoted as 3, and the planes normal to the poling axes are denoted 1 and 2 (Fig. 3.1). The 4, 5 and 6 directions depict the shear axes to represent the angular change in the plane parallel to each rotation. For example, the shear of the 2–3 plane is represented by the 4-direction. In a stacked actuator, a factor of n , which denotes the number of layers, can be placed as an amplifier in the equation because of linearity. Fig. 3.1 shows the convention of number 3 to represent the polarisation direction that generates the A₀ mode in this chapter.

Figure 3.1: Diagram of a stacked PZT with labelled axes.

In theory, stacked PZTs increase the d_{33} effect [79]. Using the increase in ceramic thickness in the direction of the applied electric field can result in an increase of axial strain rates, as can be seen in Eq. (3.4). The first number in the subscript represents the direction of the applied electric field, and the second number represents the direction of measurement.

3.2.2 Stacked PZT as a Transmitter

As shown in Eq. (3.2), a thick PZT material as a transmitter reduces the strain and vibration amplitude. The achieved displacement change can be high for the stacked actuator by reducing the thickness of the PZT in the 3-direction, stacking it to a height equivalent to that of a very thick PZT and exciting each layer with a parallel voltage.

When the poling axis is 3, the piezoelectric coefficient matrix, d , decreases to

$$\mathbf{d} = \begin{bmatrix} 0 & 0 & 0 & 0 & d_{15} & 0 \\ 0 & 0 & 0 & d_{24} & 0 & 0 \\ d_{31} & d_{32} & d_{33} & 0 & 0 & 0 \end{bmatrix},$$

where the coefficients d_{31} , d_{32} and d_{33} correlate the normal strains to the electric field in the poling direction, E_3 . Coefficients d_{15} and d_{24} correlate the shear strains on the 1–3 and 2–3 planes, respectively. When the applied stress is zero and the applied electric field is in the 3-direction, Eq. (3.2) is deduced to

$$\epsilon_3 = n d_{33} E_3, \quad (3.4)$$

where n is the number of layers of PZT in the stack. This means that a stacked actuator has a strain of up to n times larger than that of a thick PZT of equivalent height, translating to an amplitude of vibration of equivalent improvement.

3.2.3 Stacked PZT as a Receiver

The applied electric field of the PZT stack as a receiver becomes zero, and Eq. (3.1) is reduced to

$$\begin{bmatrix} D_1 \\ D_2 \\ D_3 \end{bmatrix} = n \begin{bmatrix} 0 & 0 & 0 & 0 & d_{15} & 0 \\ 0 & 0 & 0 & d_{24} & 0 & 0 \\ d_{31} & d_{32} & d_{33} & 0 & 0 & 0 \end{bmatrix} \begin{bmatrix} \sigma_1 \\ \sigma_2 \\ \sigma_3 \\ \sigma_4 \\ \sigma_5 \\ \sigma_6 \end{bmatrix}. \quad (3.5)$$

Electric displacement matrix D is then related to charge q by

$$q = \iint [D][\delta A], \quad (3.6)$$

where δA is the areas of contact perpendicular to the electric field on the 1–2 plane.

The voltage output can be obtained from Eq. (3.6), which demonstrates that the reduction in the area of contact reduces the generated charge. However, the reduction in the generated charge for the PZT stack as a receiver can be compensated for by the increased electric displacement of the PZT stack.

The following sections apply the theory for the design of a stacked PZT transducer at a low frequency to obtain a large A_0 mode amplitude and improve SNR. In this chapter, the polarisation direction that generates the A_0 mode is denoted by the 3-direction.

3.3 Numerical Modelling

Numerical modelling is carried out in COMSOL 5.2 to simulate the frequency response of a stacked PZT transducer with different numbers of PZT-5a discs. Fig. 3.2 shows the

schematic of the model. A 10 mm aluminium plate is modelled with a transducer placed in the middle to generate out-of-plane vibrations. A 0.4 mm alumina matching layer is used to reduce the impedance mismatch between the PZT layers and the aluminium plate. The monitor points are placed 5 mm away from the transducer: one on the surface and the other in the middle of the plate. Absorbing regions are modelled at both ends to attenuate edge reflections [90]. The mesh is made up of squares and triangles, with the largest one being 1 mm to ensure that more than 30 elements per wavelength at the highest frequency are exploited for accurate modelling. Each PZT layer is excited by a potential difference of 1 V in the 3-direction.

Figure 3.2: Illustration of a 2D FE model used in COMSOL for testing the out-of-plane and in-plane spatial frequency responses of stacked PZT transducers.

Out-of-plane displacement, which is dominated by the A_0 guided mode at a low frequency, is monitored in the model. The plane strain condition is considered in the 2D model.

3.3.1 Absorbing Layer and Material Properties

The following equations were used in a classical study of an absorbing region [91].

$$u(x,t)=e^{[i(kx-\omega t)]}e^{-ikx}, \quad (3.7)$$

where k is the propagation wavenumber; z is the propagation direction; t is time and ω is angular frequency. The decay term $\exp[-ikx]$ attenuates waves within the absorbing region.

This is achieved by linearly increasing the viscoelasticity of the plate for every column of nodes. In COMSOL, this task can be achieved by introducing a complex Young's modulus, E , for the material properties of the absorbing layer.

The complex Young's modulus can be introduced as

$$E_i=E \left(1+i\beta \left| \frac{x-l}{l} \right| \right). \quad (3.8)$$

The optimum value for β in this model is approximately 0.3 for a selected length of the absorbing layer. The length l of the absorbing layer is set to be 250 mm on each side.

Table 3. 1: Materials Used in the Simulation

Material	Young's Modulus (GPa)	Density (kg/m ³)	Poisson's Ratio
Brass	108	8400	0.333
SiC Foam	2.5	258	0.222
Aluminium	73	2780	0.333
Alumina	300	3860	0.21
Carbon steel	200	7850	0.295

Table 3. 2: Material Properties of PZT-5a

Density (Kg/m ³)		7750
Relative	$\epsilon_{33}^s/\epsilon_0$	826.6
Permittivity	$\epsilon_{11}^s/\epsilon_0$	919.1
Elasticity Matrix (10 ¹⁰ Pa)	C_{11}^E	12.03
	C_{33}^E	11.08
	C_{12}^E	7.52
	C_{13}^E	7.52
	C_{14}^E	2.11
	C_{66}^E	2.26
Coupling matrix (C/m ²)	ϵ_{11}	-5.35
	ϵ_{33}	15.78
	ϵ_{15}	12.29

The materials used in the simulation are presented in Tables 3.1 and 3.2, respectively. The material used for the matching layer is alumina due to its low stiffness. When the stiffness of the matching layer is lower than the stiffness of the medium and PZT, the PZT stack can be modelled as a simple mass spring system with a resonant frequency given by

$$\omega_n = \sqrt{\frac{E_L}{(n\rho_{pzt}l_{pzt})L}}, \quad (3.9)$$

where E_L is the Young's modulus of the matching layer; n is the number of layers of the PZT stack; ρ_{pzt} is the density of each PZT layer; and l_{pzt} is the thickness each PZT layer. As indicated in Eq. (3.11), an increase in matching layer length reduces the resonant frequency of the PZT stack. This frequency shift can be compensated for by stacking few PZT discs to select the preferred frequency operating point instead. However, a decrease in number of layers in the PZT stack reduces the mode amplitude, as shown in Eq. (3.4). This scenario will be discussed in Section 3.6.

3.4 Parametric Studies of stacking PZT discs

Several studies have been conducted on a typical backing mass, PZT and matching layer stack with similar dimensions [33, 78, 92]. However, the objective of this chapter is to design a low-frequency A_0 mode to be used at low voltages. Previous designs have mode amplitudes that are still too low, resulting in poor SNR at reduced voltages. Stacking PZT discs increases mode amplitude, thus improving the transduction distance and SNR at low voltages.

For this reason, a stacked PZT design is used, and a parametric study is conducted on the optimum number of PZT layers (n) to ensure that it operates near a central frequency of 0.5 MHz·mm with a reasonable bandwidth and good mode purity. For a 10 mm plate, the frequency of interest is 50 KHz. Further details on the mode selection for GWT can be found in Chapter 2. The goal of designing the transducers is to eventually implement them in an array design for application on a pipeline monitored by GWT. This will be presented in Chapter 5.

To observe the increased d_{33} effect in Eq. (3.4), two models are implemented in the simulation: 10 PZTs with a thickness of 2 mm and diameter of 10 mm stacked together and a 20 mm thick PZT with a 10 mm diameter (A). The frequency response of each transducer is measured from the top plate and presented in Fig. 3.3. The results in Fig. 3.3 show that the stacked PZT in the simulation exhibits the d_{33} effect without changing the resonance frequency in comparison with stack A. The amplitude change at the resonance frequency is close to being 10 times more, which is proportional to the number of layers, as previously shown in Eq. (3.4).

Figure 3.3: Numerical simulation showing the frequency response of two transducer designs: 10 2 mm thick PZTs stacked together and a 20 mm thick PZT (design A), both with an equivalent diameter of 10 mm.

The diameter of both PZTs are 10 mm. A PZT can be modelled as a capacitor. The principal equation for capacitors in terms of geometry is

$$C = \epsilon \frac{A_{pz}}{l_{pz}}, \quad (3.10)$$

where C is capacitance (F); ϵ is PZT permittivity (F/m) and A is the area of PZT perpendicular to the direction of the electric field (m^2). As indicated by Eq. (3.10), capacitance C increases when the cross-sectional area A is increased whilst keeping the thickness the same. An increase in capacitance reduces the bandwidth of the frequency response.

A narrow bandwidth is undesirable because it limits the range of wall thicknesses for the transducer to operate on based on the GWT criterion of $0.5 \text{ MHz}\cdot\text{mm}$. This condition will be further investigated in Section 3.8. Moreover, a large transducer is difficult to apply on the curvature of the pipeline. Hence, the diameter of the PZT stack is set to 10 mm.

3.4.1 Mode amplitude

A stack design is adopted based on the theory that the phase vibration of PZTs will superimpose, giving rise to more energy released for propagation on the plate. To investigate the optimum number of layers in the PZT stack, this study focuses on three main areas: mode amplitude, bandwidth and mode purity. Increasing the number of layers in a PZT stack whilst keeping the thickness of each layer the same increases the mode amplitude but reduces the resonant frequency. Therefore, a mode amplitude comparison is conducted to observe the frequency response of PZT stacks with different numbers of layers to identify the most suitable PZT stack for use in the frequency region of about 50 KHz.

Figure 3.4: FE simulation on PZT stacks with 4, 6, 8, 10 and 12 PZT layers. The frequency response of the out-of-plane A_0 mode is measured at a frequency interval of 1 KHz over 100 KHz from the top-plate monitoring point.

Fig. 3.4 shows that the resonance frequency of the out-of-plane mode decreases with an increasing stack (n value), and the mode amplitude increases. This result is expected given that a high stack vibrates with a large strain at its resonance, causing a dramatic decrease in impedance [93]. The increase in the number of layers in the stack also reduces the resonant frequency. For a 10 mm aluminium plate, the GWT operating point of 0.5 MHz·mm equates to 50 KHz. The PZT stack that has its resonance close to this value is the 10 stack PZT ($n=10$). However, transducers consisting of 8 and 12 PZTs in the stack may also be used because the operating frequency of 50 KHz lies within their 6 dB bandwidth, which is defined as the frequency range at half the maximum mode amplitude at resonance. Optimisation of transducer bandwidth will be further investigated in the next section.

3.4.2 Bandwidth Analysis of PZT Stack Designs

Although the A_0 mode amplitude comparison shows that a PZT stack design with 10 layers should be selected, the bandwidth of each stack must be investigated to ensure that the GWT operating point lies within the 6 dB bandwidth of the central frequency.

The bandwidth is inversely related to the capacitance of a PZT stack. The voltage across each PZT layer in the PZT stack is the same. Stack capacitance can then be modelled as the cumulative capacitance of all the capacitors in parallel, and a factor n is added in Eq. (3.10) as follows:

$$C = n\epsilon \frac{A_{pz}}{l_{pz}}. \quad (3.11)$$

The equation manifests that an increase in the number of layers results in an increase in stack capacitance, which reduces the overall bandwidth of the PZT stack linearly by a factor of n . This trend can be observed in Fig. 3.5.

Figure 3.5: 6 dB bandwidth of transducers with 6, 8, 10 and 12 stacked PZT discs.

The bandwidth in this thesis is calculated by measuring the frequency range at half the maximum mode amplitude (6 dB bandwidth). The low- and high-frequency points at this amplitude are referred to as f_{low} and f_{high} , respectively, and the difference between them is taken as the 6 dB bandwidth of the PZT stack. The bandwidth is then plotted against an increasing number of PZT layers. It can be deduced from Fig. 3.4 and 3.5 that the operating frequency of 50 KHz is found only in the 6 dB bandwidth of the 8, 10 and 12 stack layers. The increase in PZT layers results in bandwidth reduction due to an increase in capacitance, which is predicted by Eq. (3.11). The PZT stack with the highest sensitivity and a frequency of 50 KHz located approximately in the centre of its corresponding 6 dB bandwidth is the PZT stack with 10 layers of PZT discs.

Figs. 3.4 and 3.5 demonstrate that an inverse relationship exists between mode amplitude and bandwidth with increasing PZT layers. The increase in mode amplitude should be a priority for SNR improvement. Although ensuring that the bandwidth of the transducer encompasses 0.5 MHz·mm should be a priority for use with GWT, the bandwidth can be widened through alternative methods, such as the addition of a backing mass. A narrow bandwidth loses information from mode dispersion, which in turn results in an inaccurate reconstruction of defects. A backing mass is applied to further widen the bandwidth, as described in Sections 3.5 and 3.9.

3.4.3 Mode Purity

At this operating point, the out-of-plane excitation would generate both A_0 and S_0 modes. Considering that the S_0 wave is faster than the A_0 wave, it needs to be suppressed to maximize the energy received in the A_0 mode. Consequently, mode purity must be considered. The ratio of the amplitude A_0/S_0 is used to quantify mode purity. This ratio can be obtained by calculating the ratio between the maximum in-plane and out-of-plane displacements in the middle of the plate at resonance. Based on the mode shapes

of A_0 and S_0 modes, the mid plate is the point where the out-of-plane power flow mode is attributed only to the A_0 mode and the in-plane power flow mode is attributed only to the S_0 mode. The relationship between mode purity and increasing number of PZT layers is plotted in Fig. 3.6.

It can be seen from the figure that mode purity increases with increasing stack size. As shown in [94], the excitability of a particular mode by a pure out-of-plane force (E_3) is given by

$$|E_3| = \frac{k\omega}{8} m_3^2, \quad (3.12)$$

where the out-of-plane direction is represented by the subscript 3; k is the wavenumber of the considered mode, ω is the angular frequency of the out-of-plane excitation force and m_3 is the power flow normalised surface displacement component of the mode shape.

Figure 3.6: Ratio of the maximum out-of-plane and in-plane displacements (A_0 and S_0 mode amplitudes, respectively) or mode purity obtained at the resonant frequency for 4, 6, 8, 10 and 12 PZT layers.

In the low-frequency region below 100 KHz, the value of m_z for the A_0 mode increases rapidly with a decreasing frequency–thickness product, whereas that for the S_0 mode diminishes. Consequently, the excitability of the A_0 mode over the S_0 mode increases at low frequencies. This trend can be observed in Fig. 2.12 in Section 2.3. Increasing the number of layers in the PZT stack reduces the resonant frequency of the stack, thus increasing the mode purity of the A_0 mode. The 10 and 12 stacks have bandwidths that cover the GWT operating point of 50 KHz. However, increasing the number of layers in the PZT stack from 10 to 12 is unnecessary because it only provides a slight improvement to mode purity but not much improvement to the displacement measured at 50 KHz. Moreover, a reduction in bandwidth occurs. Therefore, using a stack with 10 PZT layers is preferred and used in this study.

3.5 Backing Mass and Matching Layer Dimensions

The PZT stack with 10 PZT layers is selected from the previous section based on the highest A_0 mode amplitude at 50 KHz with reasonable mode purity. However, Eq. (3.11) demonstrates that the increased number of layers increases capacitance, thus reducing the actuator bandwidth. The addition of a backing mass with a similar impedance to PZT-5a can potentially improve the bandwidth but with a reduced mode amplitude. A study on the addition of a brass backing mass to the stacked actuator is performed to investigate the effect of bandwidth improvement and increase in the number of PZT layers.

3.5.1 Effects of Adding a Brass Backing Mass to the PZT Stack Transducer

The added brass backing mass is used to damp out small, spurious waves whilst lowering the resonance frequency. As indicated in Fig. 3.7, the out-of-plane frequency response is obtained from the numerical simulation of a PZT stack with six PZT layers,

demonstrating the increase in mode amplitude whilst decreasing the 6 dB bandwidth with a 6 mm brass backing mass.

The added backing mass increases the amplitude by 28.99% and reduces the resonance frequency and bandwidth by 26.58% and 17.95%, respectively. The decrease in bandwidth for the PZT stack with backing mass is expected considering the larger impedance mismatch between the stacked PZTs and the brass backing mass as opposed to the PZT stack without backing mass.

Figure 3.7: Out-of-plane frequency response obtained from the top-plate monitoring point in the numerical simulation of a PZT stack with six PZT layers with (solid line) and without a 6 mm backing mass (dashed line).

The addition of a backing mass improves the transducer bandwidth by absorbing the resonator back emission [95]. However, in this case, brass is not attenuative enough for a PZT stack design because of the larger back emission compared with that of a single vibrating PZT. This makes the use of a brass backing mass to improve bandwidth ineffective. However, the reduction in frequency and improvement in A_0 mode

amplitude due to this impedance mismatch could reduce the number of layers needed in the stack to have an equivalent mode amplitude at 50 KHz with a slightly widened bandwidth.

PZT stacks with six PZT layers and a brass backing mass with thicknesses varying from 3 mm to 12 mm are modelled, and their frequency responses are shown in Fig. 3.8. The stack with a 9 mm backing mass has a resonance frequency close to the 50 KHz frequency region. This result can also be deduced based on the total height of the stack, with the height of each design in Fig. 3.7 being 21.4 and 20.4 mm. Both stacks provide two frequency responses with similar resonance frequencies on the basis of Eq. (3.9) because the density of brass and PZT-5a are similar.

Figure 3.8: Frequency response of the A_0 mode obtained from the numerical simulation of a PZT stack with six PZT layers with 3 mm (solid black line), 6 mm (dashed blue line), 9 mm (dashed-dotted green line) and 12 mm (dotted red line) brass backing masses.

The bandwidth reduction from the addition of the brass backing mass and that from increasing the number of PZT layers in the stack are compared to select the design with the wider bandwidth. In Fig. 3.9, the bandwidth of PZTs with different layers and without backing mass is compared with that of a six-stack PZT with backing masses of different thicknesses. The figure indicates that increasing the thickness of the brass backing mass reduces the bandwidth of the stack. This is because an increase in brass backing mass thickness increases the capacitance of the stack, thereby reducing the bandwidth. The six-stack PZT with a 9 mm backing mass has a bandwidth comparable with that of the 10-stack PZT with no backing mass.

Figure 3.9: Bandwidth comparison for a six-stack PZT with 3, 6, 9 and 12 mm backing mass thicknesses (solid line) and bandwidth of the frequency response of PZT stacks with an increasing number of PZT layers without a backing mass (dashed line).

Although the bandwidth increases with the selected backing mass, the mode amplitude decreases significantly, as shown in Fig. 3.10. Therefore, different transducer designs should be selected depending on the applications.

Figure 3.10: Out-of-plane frequency response for a PZT stack with 10 PZT layers and no backing mass (dash-dot line) and a PZT stack with six PZT layers and a 9 mm backing mass (dash line). The mode amplitude at resonance is large for the 10-stack PZT.

Increasing the thickness of the backing mass leads to optimal mode purity, as shown in Fig. 3.11. This scenario is expected because the backing mass damps out spurious modes. However, when the mode purity for the stack with and without the brass backing mass is compared, the result indicates that the mode purity of the stack without a backing mass is larger than that with a backing mass. This result suggests that the impedance mismatch between the brass backing mass and PZT stack layers reduces the ability of the backing mass to damp out spurious modes.

The reduction in A_0 mode amplitude at resonance and bandwidth and mode purity for the PZT stack with six layers and a 9 mm backing mass is undesirable. Therefore, a PZT stack with 10 PZT layers and without a backing mass is still preferred.

Figure 3.11: Mode purity obtained by using the ratio of maximum out-of-plane and in-plane displacements measured at the monitoring point in the middle of the plate for a PZT stack with 10 PZT layers and varying backing mass lengths between 3 and 12 mm.

3.6 Optimization of a Matching Layer on the PZT Stack Transducer

This section discusses the selection of a proper matching layer to enhance the transduction efficiency and increase the transducer bandwidth. In this study, aluminium oxide (alumina) is selected as the material for the matching layer because its acoustic impedance is between the PZT and aluminium plate. Numerical simulations are conducted to assess the appropriate thickness of the alumina matching layer.

The intensity transmission coefficient for the propagation of acoustic waves from the PZT into the medium at normal incidence is

$$T_I = \frac{4Z_p Z_m}{(Z_p + Z_m)^2 \cos^2 \theta + (Z_L + Z_p Z_m / Z_L)^2 \sin^2 \theta}, \quad (3.13)$$

where Z_p , Z_m and Z_L are the acoustic impedances of the PZT, medium of propagation and matching layer, respectively, and

$$\theta = \frac{2\pi L}{\lambda_L}, \quad (3.14)$$

where L is the thickness of the matching layer and λ_L is the wavelength at the frequency of interest.

Eqs. (3.13) and (3.14) show that the optimum matching layer thickness is typically $\lambda_L/4$. When Z_L is the geometric mean of Z_p and Z_m ($\sqrt{Z_p \times Z_m}$) and $T_I=1$, a perfect transmission is achieved. Therefore, in the frequency of interest (50 KHz) for A_0 mode propagation on a 10 mm plate wherein the wavelength ranges from approximately 20–36 mm, the optimum thickness is roughly 5–9 mm. However, a thick matching layer reduces the resonant frequency of the PZT stack. This situation can be addressed by using only a few layers of PZTs in the PZT stack to increase the resonant frequency. However, similar to what has been indicated in previous discussions, this will cause a

significant reduction in mode amplitude. Therefore, thin matching layers are applied in our study.

Figure 3.12: Frequency response of the out-of-plane mode obtained from a numerical simulation of a PZT stack with 10 layers and 0.4 mm (solid black line), 0.5 mm (dashed line), 0.8 mm (solid green line) and 1.2 mm (dotted red line) alumina matching layers.

In Fig. 3.12 shows the frequency response of the out-of-plane A_0 mode for PZT stacks with 10 layers of PZT and with the matching layer thickness varying from 0.4 mm to 1.2 mm. The transmission coefficient and resonant frequency are similar under small changes in the matching layer dimension. The A_0 mode amplitude at 50 KHz varies minimally with changing thicknesses of the matching layer. Therefore, the thickness of the matching layer is kept to the thinnest available matching layer of 0.4 mm.

3.7 Experimental validation

Figure 3.13: Experimental setup for testing a 10-stack transducer.

An experiment is conducted on a 10 mm thick aluminium plate to validate the FE model of the PZT stack. A five-cycle, Hann-windowed tone burst signal is used and excited at frequencies from 10 KHz to 100 KHz with a 10 KHz step. The ultrasonic waveform is generated with TiePie Handyscope HS3. A laser vibrometer (Fig. 3.13) is used to pick up the response at the top of the plate 5 mm away from the edge of the PZT stack seen in Fig. 3.14. The frequency response is obtained by measuring the maximum amplitude of the first arrival wave packet. A time gate is applied to separate the first arrival wave packet for the A_0 mode. The data are normalised for comparison with the simulation results, and good agreement is observed (Fig 3.15).

Figure 3.14: Ten-stack transducer used in the experiment.

Figure 3.15: Frequency response of the out-of-plane displacement of the PZT stack with 10 PZT layers obtained from the experiment (solid black line) and simulation (dotted blue line).

3.8 Limitation of Transducer Design

Whilst the transducer design in this chapter is selected based on a 10 mm thick aluminium plate, practical applications might require the use of a transducer in different structures. An increase in plate stiffness theoretically reduces the amplitude of the A_0 mode generated at resonance because of the high attenuation experienced by wave propagation. In addition, resonance of the PZT stack occurs when the generated wavelength is twice the plate thickness. For plates with a small thickness, the generated wavelength is smaller than that for thick plates. Small wavelengths are associated with high frequencies, which increase the resonance frequency of the PZT stack.

In this section, the selected PZT stack with 10 PZT layers is modelled with different plate thicknesses. The 6 dB bandwidth of the frequency response should be within the region of interest for GWT, being $0.5 \text{ MHz}\cdot\text{mm} \pm 10\%$.

3.8.1 Plate Thickness and Stiffness

The frequency response for a PZT stack with 10 layers is obtained from numerical simulations in a similar process as that in the previous sections. This task is performed to investigate the changes in the mode amplitude and bandwidth for different plate thicknesses and properties. Plate thickness is varied between 5 and 20 mm at a 2.5 mm interval. Two material properties for the plate are also investigated for a comparison with the mode amplitude for different material stiffness values.

3.8.1.1 Mode Amplitude

Figs. 3.16 and 3.17 present the frequency response of the out-of-plane displacement obtained from the numerical simulations of a transducer with 10 PZT layers on aluminium and carbon steel plates with different thicknesses. The figures demonstrate that the thick plates result in a low mode amplitude and reduced resonance frequency. The amplitude of the A_0 mode on the carbon steel plate is smaller than that on the aluminium plate because of the large impedance mismatch between the transducer and carbon steel plate. The resonant frequency decreases with increasing plate thickness, suggesting a limitation for the transducer.

Figure 3.16: Numerical simulation of a 10-stack PZT on a range of aluminium plate thicknesses between 5 and 20 mm at 2.5 mm intervals to obtain different out-of-plane frequency responses from the top-plate monitoring point.

Figure 3.17: Numerical simulation of a 10-stack PZT on a range of carbon steel plate thicknesses between 5 and 20 mm at 5 mm intervals to obtain different out-of-plane frequency responses from the top-plate monitoring point.

The frequency response of the transducer changes with different plate thicknesses. If the operating frequency of 0.5 MHz·mm is beyond the 6 dB bandwidth, then the transducer cannot be used for such plate thickness. This limit considers the case where the environmental noise is half the maximum mode amplitude at resonance. In this case, the signal generated for frequencies outside the 6 dB bandwidth is buried in noise.

In the following section, this limitation is visualised by investigating the bandwidth and tomography frequency (0.5 MHz·mm) that change with changing plate thickness.

3.8.1.2 Bandwidth

The 6 dB bandwidth of the PZT stack with 10 PZT layers is obtained with respect to plate thickness. The result for aluminium and carbon steel is shown in Figs. 3.18 and 3.19, respectively.

Figure 3.18: 6 dB bandwidth for the out-of-plane frequency response obtained for the PZT stack with 10 PZT layers on 5, 7.5, 10, 12.5, 15, 17.5 and 20 mm aluminium plate thicknesses are plotted with the frequencies of the dispersive region of the A_0 mode for different plate thicknesses.

Fig. 3.18 shows that the best area of operation lies within the intersection points of 0.55 MHz.mm and the upper bandwidth (f_{high}) and 0.45 MHz.mm and the lower bandwidth (f_{low}). This condition results in an area of operation between 8.4 and 16 mm wall thicknesses for aluminium.

Figure 3.19: 6 dB bandwidth for the out-of-plane frequency response obtained for the PZT stack with 10 PZT layers on 5, 10, 15 and 20 mm carbon steel plate thicknesses are plotted with the frequencies of the dispersive region of the A_0 mode for different plate thicknesses.

Fig. 3.19 shows that the stack design has an area of operation between 10.3 and 14.2 mm wall thicknesses for steel. The effective range for carbon steel is smaller than that for aluminium. This notion indicates that the range of operation decreases due to the increase in material stiffness, a condition that is expected considering that rigid plates result in less energy propagation from the transducer into the plate. Therefore, the bandwidth should be extended to increase the effective range of the transducer on stiff materials.

3.9 Bandwidth Widening with composite backing

The use of a composite material for the backing mass is an alternative way to widen the bandwidth. In this case, epoxy and alumina bits are mixed at a 3:1 ratio and moulded into a 5 mm thick layer to be used as a backing mass. This mixture can provide an optimal impedance match to the PZT stack, thus reducing reflections from the back emission. The alumina bits in the epoxy act as scatterings to further attenuate the back emissions. Fig. 3.20 shows the normalised amplitude of the A_0 mode frequency response of the PZT stack with 10 PZT layers with and without the epoxy composite backing mass.

Figure 3.20 A_0 mode frequency response of the PZT stack with 10 PZT layers with (dotted red line) and without the epoxy alumina composite backing mass (solid black line).

The bandwidth is widened to 61 KHz with the backing mass. The widened bandwidth increases the effective range of wall thicknesses that the transducer can operate on. The dispersive region of the A_0 mode is within a large range of wall thicknesses. The transducer can operate on a wide range of thicknesses without greatly sacrificing the A_0 mode amplitude. A wide bandwidth is also desirable in exciting short pulses with good axial resolution. The bandwidth of the transducer with the epoxy alumina backing mass is 61 KHz and approximately 1.5 times larger than that of the 10-stack PZT. The expected range of wall thicknesses for the PZT stack with the composite backing layer is at least twice the range of the PZT stack without any backing layer and will be used in Chapters 4 and 5.

3.10 Summary

This chapter presented the design of a stacked PZT transducer to be used in obtaining measurements for GWT. The stacked PZT layers provide satisfactory A_0 mode amplitude and mode purity. The optimisation of the A_0 mode amplitude involved sacrificing the optimal mode purity and was accompanied with a reduction in bandwidth for the out-of-plane mode. Firstly, the mode amplitude increased with increasing number of PZT layers in the stack. Secondly, the mode purity increased with increasing number of PZT layers. Thirdly, the bandwidth decreased with the increasing stack. The PZT stack with 10 PZT layers was selected because its resonance frequency was the closest to the region of 0.5 MHz·mm. The reduced bandwidth was addressed through the addition of a backing layer.

The addition of a brass backing mass did not improve the bandwidth probably due to the large reflections from high back emissions because of an impedance mismatch between the brass and stacked PZTs. The bandwidth was narrow when the impedance mismatch was large. This mismatch increases the A_0 mode amplitude at a reduced resonance frequency. These results have also been shown in a previous study [78]. The bandwidth of the six-stack PZT with a 9 mm backing mass was slightly larger than that of the 10-stack PZT at a similar resonance frequency. However, the mode amplitude was reduced. For this reason, the 10-stack PZT was preferred over the six-stack PZT with a 9 mm brass backing mass. Alternatively, the use of an alumina–epoxy composite provided a good impedance match and increased dampening, as shown in Section 3.8. The obtained 6 dB bandwidth was approximately 1.5 times wider than the bandwidth of a 10-stack PZT without sacrificing the mode amplitude.

The stacked PZT design is relatively flexible for applications within a range of wall thicknesses. However, this range is limited and is reduced for materials with high stiffness. Therefore, bandwidth becomes the limiting factor. Any further increase in mode amplitude reduces the range of wall thicknesses where the dispersive frequency region of the A_0 mode is still within the 6 dB transducer bandwidth. In this work, the addition of a composite backing mass increased the effective range of wall thicknesses for which transducers can operate on when used for GWT. Consequently, the 10-stack PZT with a composite backing mass was deemed to be versatile for on-site applications where pipe wall thicknesses may vary depending on the operating requirements. The application of this transducer has been published, and an on-site application of this transducer is presented in Chapter 5.

Modulated signals can be efficiently used with the increase in bandwidth because transducers act as filters; moreover, having a bandwidth that is smaller than that of the modulated signal reduces the amount of energy propagating into the plate. The use of modulated signals is essential in further improving the SNR of the designed transducer whilst retaining its high mode amplitude. This concept will be discussed in the following chapter.

Chapter 4: SNR Improvement and Power Reduction through Deconvolution

4.1 Introduction

A pure guided wave (GW) mode with a wide bandwidth is preferred in ultrasonic guided wave tomography (GWT) applications. This is achieved through the optimisation of transducer design, as discussed in Chapter 3. However, the design constrains the SNR of the signal due to its filtering effect. This situation can be improved by deconvolution methods. In practice, the power of the signal is also restricted, especially in the oil and gas industry. One of the methods that can be used to improve SNR with reduced power consumption is the utilization of coded pulse compression. This process involves sending a frequency-modulated signal.

The linear frequency modulated (LFM) signal or ‘chirp’ is a common excitation signal initially developed for radar systems and applied in ultrasonic systems in the 1970s [96]. This signal has been recently used in ultrasonic NDT [97-99] and medical imaging [100, 101]. With a chirp signal, the excitation voltage is reduced whilst the energy of the excitation signal is maintained by increasing the signal length.

In most ultrasonic applications, chirp signals are used for TOF extraction by cross-correlating the excitation signal and measured response (matched filter) either in its raw form or by applying a Hilbert transform first to determine the arrival time of response signals [102]. Discrete wavelet transform can also be applied to denoise a noisy signal before the cross-correlation [103]. However, use of chirp signals in GW applications is further limited by the dispersive and multi-modal nature of the modes that require narrowband excitations to ensure mode purity whilst limiting dispersive effects [104, 105].

In most applications for obtaining TOF information, dispersion can be compensated for by various signal processing methods [97, 106, 107]. However, in GWT, dispersion properties are used to reconstruct the defect geometry and are therefore desirable. The deconvolution method provides the benefit of SNR improvement through the excitation of long chirps, which are then deconvoluted into the toneburst signal. Accordingly, the wide bandwidth with improved SNR is retained.

It should also be noted that in industrial applications, the voltage of the excitation signals should be reduced for either safety or energy conservation purposes. The stacked PZT transducers designed in Chapter 3 are sensitive to high voltages, which result in heat generation and dielectric losses, thereby leading to transducer degradation.

This chapter does not aim to design a chirp signal, which has already been extensively covered in literature, but to investigate the effect of low voltage excitation of a chirp signal on SNR improvement.

The rest of the chapter is organised as follows. The fundamental theory of the LFM signal for SNR improvement is reviewed. The experimental setup for GWT is presented in Section 4.3. The deconvolution process is subsequently outlined, and the SNR improvement is obtained for high- and low-voltage cases. Numerical simulations are also conducted to investigate the effect of changing the excitation voltage on the output SNR of the deconvoluted toneburst. The trend is then explained in the discussion section, and the factors that affect the observed trend are identified.

4.2 Theory of SNR Improvement using the LFM Signal

The LFM waveform in the time domain can be defined as follows:

$$u(t) = Ax \left(\frac{t}{T} \right) e^{j\pi \left(2f_c t \pm \frac{B}{T} t^2 \right)} = Ax \left(\frac{t}{T} \right) e^{j\Theta(t)} \quad -\frac{T}{2} \leq t \leq \frac{T}{2}, \quad (4.1)$$

where A is the signal amplitude; f_c is the central frequency of the chirp signal; B is the bandwidth; T is the time duration, and B/T is the slope of frequency modulation.

$x(t)$ is defined as any weighted function. In this chapter, $x(t)$ is defined as a Hann window where

$$x(t) = \sin^2\left(\frac{\pi t}{T}\right). \quad (4.2)$$

The instantaneous LFM frequency is given by

$$f(t) = \frac{1}{2\pi} \frac{\delta\Theta}{\delta t} = f_c \pm \frac{B}{T} t \quad -\frac{T}{2} \leq t \leq \frac{T}{2}. \quad (4.3)$$

A chirp signal in the time domain can be described in its Euler form as

$$S_c(t) = A e^{j(2\pi f_0 t + \frac{\pi B}{T} t^2)}, \quad (4.4)$$

where A is the amplitude related to the voltage powering the system; f_0 is the central frequency and B is the bandwidth.

The obtainment of the frequency spectrum of the chirp signal has been discussed in literature and will be reviewed here for convenience. The frequency spectrum of the chirp signal is expressed as follows:

$$S_c(\omega) = \int_{-\frac{T}{2}}^{\frac{T}{2}} A e^{j(2\pi f_0 t + \frac{\pi B}{T} t^2)} e^{-j2\pi f t} dt. \quad (4.5)$$

The resulting integral form can be solved by completing the square to produce

$$S_c(\omega) = A e^{-j\frac{\pi(f-f_0)^2 T}{B}} \int_{-\frac{T}{2}}^{\frac{T}{2}} e^{j\frac{\pi}{2} \left(\sqrt{\frac{2B}{T}} t + \frac{f-f_0}{\sqrt{\frac{B}{2T}}} \right)^2} dt,$$

$$\text{where } x = \left(\sqrt{\frac{2B}{T}} t + \frac{f-f_0}{\sqrt{\frac{B}{2T}}} \right)^2.$$

The expression can then be presented as

$$S_c(\omega) = A \sqrt{\frac{T}{2B}} e^{-j \frac{\pi(f-f_0)^2 T}{B}} \int_{x_2}^{x_1} e^{j \frac{\pi}{2} x^2} dx, \quad (4.6)$$

where the limits of the integral are defined as follows:

$$x_1 = \sqrt{\frac{D}{2}} \left[1 - \frac{2(f-f_0)}{B} \right]$$

and

$$x_2 = -\sqrt{\frac{D}{2}} \left[1 + \frac{2(f-f_0)}{B} \right],$$

where D is the time–bandwidth product.

Eq. (4.6) can then be expressed in sinusoidal form through Euler relations as follows:

$$S_c(\omega) = A \sqrt{\frac{T}{2B}} e^{-j \frac{\pi(f-f_0)^2 T}{B}} \left(\int_0^{x_1} \cos \frac{\pi}{2} x^2 + j \sin \frac{\pi}{2} x^2 dx - \int_0^{x_2} \cos \frac{\pi}{2} x^2 + j \sin \frac{\pi}{2} x^2 dx \right),$$

where

$$\int_0^x \cos \frac{\pi}{2} x^2 dx = \sum_{n=0}^{\infty} \frac{(-1)^n \left(\frac{\pi}{2}\right)^{2n} x^{4n+1}}{(2n)!(4n+1)},$$

$$\int_0^x \sin \frac{\pi}{2} x^2 dx = \sum_{n=0}^{\infty} \frac{(-1)^n \left(\frac{\pi}{2}\right)^{2n+1} x^{4n+3}}{(2n+1)!(4n+3)}.$$

Finally, an expression for the Fourier transform of a chirp signal is obtained and shown as

$$S_c(\omega) = A \sqrt{\frac{T}{2B}} e^{-j \frac{\pi(f-f_0)^2 T}{B}} [Z(x_1) - Z(x_2)], \quad (4.7)$$

where Z(x1) and Z(x2) are the Fresnel integrals from x1 to 0 and x2 to 0, respectively.

The limits of the Fresnel integral is shown in Eq. (4.8) and visualised in Fig. 4.2.

$$\lim_{x \rightarrow \infty} \cos \frac{\pi}{2} x^2 = 0.5, \text{ and } \lim_{x \rightarrow \infty} \sin \frac{\pi}{2} x^2 = 0.5 \quad (4.8)$$

Figure 4.1: Fresnel integral limits for $\cos 0.5\pi x^2$ (dash-dot line) and $\sin 0.5\pi x^2$ (solid line)

The limits shown here represent the dependence of the Fresnel integrals on the time bandwidth product D . A large D value means that oscillations are stable as the integral approaches 0.5. This situation signifies that large time–bandwidth products are desirable because the large oscillations in the Fresnel integral increase the sidelobe generation that masks weak scatters in the vicinity of huge ones. These sidelobes also result in large spectral leakages of the chirp in the frequency domain, which are undesirable when applying a passband filter, an inherent step in the deconvolution process.

The magnitude of the chirp frequency spectrum in Eq. (4.7) can be presented as

$$|S_c(\omega)| = A \sqrt{\frac{T}{2B}} |Z(x_1) - Z(x_2)|. \quad (4.9)$$

4.2.1 Hann-windowed Chirp

The frequency spectrum of a Hann-windowed chirp can be obtained from a similar process as above. The time domain and frequency spectrum of the Hann-windowed chirp are shown below.

$$S_c(t) = 0.5A(1 - 0.5(e^{j\frac{2\pi}{T}t} + e^{-j\frac{2\pi}{T}t}))e^{j(2\pi f_0 t + \frac{\pi B}{T}t^2)} \quad (4.10)$$

Similar to the derivation of Eq. (4.7) above,

$$S_c(\omega) = \frac{A}{2} \sqrt{\frac{T}{2B}} \left(e^{-j\frac{\pi(f-f_0)^2 T}{B}} \int_{x_2}^{x_1} e^{j\frac{\pi}{2}(x)^2} dx - \frac{1}{2} e^{-j\frac{\pi(30f-31f_0)^2 T}{900B}} \int_{w_2}^{w_1} e^{j\frac{\pi}{2}(w)^2} dw - \frac{1}{2} e^{-j\frac{\pi(30f-29f_0)^2 T}{900B}} \int_{v_2}^{v_1} e^{j\frac{\pi}{2}(v)^2} dv \right), \quad (4.11)$$

where

$$w = \left(\sqrt{\frac{2B}{T}} t + \frac{30f-31f_0}{30\sqrt{\frac{B}{2T}}} \right)^2, \text{ and } v = \left(\sqrt{\frac{2B}{T}} t + \frac{30f-29f_0}{30\sqrt{\frac{B}{2T}}} \right)^2,$$

$$w_1 = \sqrt{\frac{D}{2}} \left[1 - \frac{2(30f-31f_0)}{30B} \right], \text{ and } v_1 = \sqrt{\frac{D}{2}} \left[1 - \frac{2(30f-29f_0)}{30B} \right],$$

and

$$w_2 = -\sqrt{\frac{D}{2}} \left[1 + \frac{2(30f-31f_0)}{30B} \right], \text{ and } v_2 = -\sqrt{\frac{D}{2}} \left[1 + \frac{2(30f-29f_0)}{30B} \right].$$

The frequency spectrum can then be presented as

$$S_{c,Hann}(\omega) = \frac{A}{2} \sqrt{\frac{T}{2B}} \left(e^{-j\frac{\pi(f-f_0)^2 T}{B}} [Z(x_1) - Z(x_2)] - \frac{1}{2} e^{-j\frac{\pi(30f-31f_0)^2 T}{900B}} [Z(w_1) - Z(w_2)] - \frac{1}{2} e^{-j\frac{\pi(30f-29f_0)^2 T}{900B}} [Z(v_1) - Z(v_2)] \right). \quad (4.12)$$

4.2.2 Hann-windowed Toneburst — three cycles

The Hann-windowed toneburst can be mathematically presented in the time domain as follows:

$$S_{tb}(t) = 0.5A(1 - 0.5(e^{j\frac{2\pi}{T_{tb}}t} + e^{-j\frac{2\pi}{T_{tb}}t}))e^{j(2\pi f_0 t)}. \quad (4.13)$$

For a three-cycle toneburst, Eq. (4.13) becomes

$$S_{tb}(t)=0.5A(1-0.5(e^{\frac{2\pi f_0}{3}t}+e^{-\frac{2\pi f_0}{3}t}))e^{j(2\pi f_0 t)}.$$

The frequency spectrum of the three-cycle Hann-windowed toneburst is

$$S_{tb}(\omega)=\frac{AT_{tb}}{2}\text{sinc}(\pi(f_0-f)T)-\frac{AT_{tb}}{4}\text{sinc}\left(\frac{\pi(4f_0-3f)T}{3}\right)-\frac{AT_{tb}}{4}\text{sinc}\left(\frac{\pi(2f_0-3f)T}{3}\right). \quad (4.14)$$

4.2.3 Quantifying SNR Improvement

The wave propagation convolution in the time domain can be expressed as follows:

$$r_{tb}(t)=[s_{tb}(t)+n(t)]*h(t). \quad (4.15)$$

Assuming a linear time-invariant (LTI) system, it can be presented in the frequency domain as

$$R_{tb}(\omega)=[S_{tb}(\omega)+N(\omega)]H(\omega)=[\underbrace{S_{tb}(\omega)H(\omega)}_{\text{Signal}}+\underbrace{H(\omega)N(\omega)}_{\text{Noise}}], \quad (4.16a)$$

$$R_c(\omega)=[S_c(\omega)+N(\omega)]H(\omega)=[S_c(\omega)H(\omega)+H(\omega)N(\omega)]. \quad (4.16b)$$

Eqs. (4.16a) and (4.16b) show the reduction in noise whilst maintaining the signal amplitude. The gain in SNR can be expressed as

$$SNR_g=20\log\left(\frac{R_c(\omega)}{R_{tb}(\omega)}\right)=20\log\left(\frac{S_c(\omega)H(\omega)}{H(\omega)N(\omega)}\cdot\frac{N(\omega)H(\omega)}{S_{tb}(\omega)H(\omega)}\right)=20\log\left(\frac{S_c(\omega)}{S_{tb}(\omega)}\right). \quad (4.17)$$

At $\omega=\omega_0$, the last two terms in a Hann-windowed chirp and Hann-windowed toneburst disappear, and Eqs. (4.12) and (4.14) are reduced to

$$S_{c,hann}(\omega_0)=\frac{A}{2}\sqrt{\frac{T_c}{2B}}|[Z(x_1)-Z(x_2)]|$$

and

$$S_{tb}(\omega_0)=\frac{AT_{tb}}{2},$$

Figure 4.2: Numerical solution of Fresnel integrals in functions $Z(x1)$ and $Z(x2)$ at varying frequencies.

respectively. From Fig. 4.2, the average magnitude of $[Z(x1)-Z(x2)]$ is $\sqrt{2}$ and the magnitude at ω_0 is assumed to be 1.55, corresponding to 70 KHz. The SNR gain (SNR_g) can then be expressed as a function of toneburst length (T_{tb}), bandwidth of the chirp signal and ratio of the length of the chirp signal with respect to the toneburst signal (k).

$$\text{SNR}_g = 20 \log \left(\frac{S_c(\omega)}{S_{tb}(\omega)} \right) = 20 \log \left(1.275 \sqrt{\frac{k}{2T_{tb}B}} \right) \quad (4.18)$$

This SNR gain can be modified by selecting the optimal values of k , the length of the toneburst signal and the bandwidth of the chirp signal (Appendix B).

4.3 Experimental Setup and Selection of the Chirp Signal

In this chapter, the bandwidth of the chirp is set to 30 – 110 KHz for experiments and simulations with a central frequency of 70 KHz. The selected frequency reflects the steep gradient of the A_0 mode for applications in GWT. The optimum bandwidth of the chirp should be similar to the 12 dB bandwidth of the transducer developed in Chapter 3. The comparable bandwidths of the excitation signal and transducer can provide the optimum SNR [108] and allow for the use of the excitation signal efficiently.

Figure 4.3: Experimental setup where the receiver and transmitter arrays are placed 580 mm apart. The arrays are connected to an array controller for excitation and data acquisition.

The length of the chirp signal used in this work is 429 μ s to ensure sufficient data collection in the data collection system used (OPEN system from Lecoer Electronique). The data acquisition time is set to 1000 μ s. The pipe is sufficiently long to prevent reflections from the pipe edge during experiments. The two arrays of 24 transducers developed in Chapter 3 are placed 580 mm apart on a 12-inch nominal diameter steel pipe with 10 mm wall thickness (Fig. 4.3). The distance between the arrays and each pipe end is 700 mm.

4.4 Deconvolution of an LFM

In this chapter, the chirp is used as an excitation signal, and the received signal is deconvoluted to a toneburst signal [88] for post-processing in GWT. The deconvolution method has a gating process that removes parts of the signal spectrum (in our study, below 30 KHz and above 110 KHz). The amplitude for all signals are referenced to a voltage excitation of 50 V peak to peak (± 25 V) and the length of the chirp is ten times the length of the toneburst unless stated otherwise.

Figure 4.4: (a) Chirp excitation $S_c(t)$ and (b) the three-cycle toneburst excitation $S_{tb}(t)$.

In the deconvolution process, the transmitted toneburst signal (Fig. 4.4(b), solid signal) and chirp signal (Fig. 4.4(a), dotted line signal) are used to obtain a transfer function $G(\omega)$, as shown in Fig. 4.5, on the basis of the following. Given that the desired excitation is a toneburst T_{tb} , the corresponding response R_{tb} is obtained by propagation on the plate described by transfer function $H(\omega)$ and is presented as follows:

$$R_{tb}(\omega) = H(\omega)T_{tb}(\omega). \quad (4.19)$$

The chirp response R_c is obtained as follows:

$$R_c(\omega) = H(\omega)T_c(\omega). \quad (4.20)$$

The deconvoluted toneburst $R_{tb,deconv}$ is then derived by combining Eqs. (4.19) and (4.20) and expressed as Eq. (4.21).

$$R_{tb,deconv}(\omega) = \frac{R_c(\omega)}{T_c(\omega)} T_{tb}(\omega) = R_c(\omega)G_0(\omega) \quad (4.21)$$

Figure 4.5: Transfer function G_0 (solid black line) is obtained from the division of toneburst $S_{tb}(\omega)$ (dash-dotted black line) by chirp $S_c(\omega)$ (dotted black line) frequency responses. The gating process is carried out by applying a rectangular filter in the deconvolution to obtain the transfer function $G_0(\omega)$.

Transfer function G_0 in Fig. 4.5 incorporates the plate material, attenuation and distance of propagation and is assumed to remain constant regardless of the type of excitation

used. This condition means that by using chirp excitation T_c , the corresponding chirp response, R_c , is utilised to obtain the desired excitation response, which in this case is the toneburst response (R_{tb}) (Eq. (4.21)). This process is referred to as deconvolution in this chapter.

4.4.1 Comparison of Deconvoluted and Original Toneburst

A comparison is performed between a deconvoluted toneburst response ($R_{tb,deconv}$) and a typical toneburst response (R_{tb}) obtained from the experimental measurements described in Section 4.3. This comparison is carried out to determine if the method is reasonable. The response of chirp signal excitation R_c and toneburst signal excitation R_{tb} are shown in Fig. 4.6.

Figure 4.6: Chirp response $R_c(t)$ and toneburst response $T_{tb}(t)$ obtained from the deconvolution. Transmitter 12 and receiver 12, which are on a straight line on the pipe, are used from the transmitter and receiver arrays, respectively.

Chirp response R_c is then deconvoluted through Eq. (4.21) to obtain the deconvoluted toneburst $R_{tb,deconv}$. A comparison of the deconvoluted toneburst and toneburst response is shown Fig. 4.7.

Figure 4.7: The response for both the deconvoluted toneburst $R_{tb,deconv}(t)$ (solid line) and toneburst $R_{tb}(t)$ (dotted line) responses for an acquisition time of 1000 μs .

The deconvolution process results in a deconvoluted toneburst response ($R_{tb,deconv}$) of a similar arrival time with a toneburst response (R_{tb}). Both signals are normalised to compare SNR, and the deconvoluted toneburst is found to have an improved SNR of 3 dB. The signals are similar but scaled differently due to the effect of only taking one-sided fast Fourier transform (FFT) in the deconvolution. This situation can be resolved by either normalising the amplitudes or incorporating a scale based on Parseval's theorem in the deconvolution method.

The deconvolution is performed only on a portion of the single-sided FFT. Accordingly, the symmetric property of FFT and the power loss from frequency bins outside of the gating function should be considered. Here, a rectangular gate is used due to the continuous and smooth nature of the windowed chirp response (R_c).

The scaling is dependent on the utilised bandwidth, which translates to the number of frequency bins used over the total number of frequency bins, corresponding to the power of signal and noise that is gated and the total power of the signal and noise respectively. A multiplication factor can be incorporated into the inverse FFT (IFFT) to account for this loss in information.

According to Parseval's theorem for any FFT algorithm,

$$\Delta t \sum_{i=1}^N V_i^2 = \frac{\Delta t}{N} \sum_{i=1}^N |H_i|^2, \quad (4.22)$$

where V_i indicates the real voltage samples in the time domain and H_i is the complex FFT values.

Eq. (4.22) shows that gating the FFT of the time domain signals reduces the overall power of the signal. This means that a scaling factor should be incorporated to compensate for the power loss during deconvolution. For example, the power of the gated signal in the frequency domain is $\frac{\Delta t}{N} \sum_{N_{y,30}}^{N_{y,110}} |H_i|^2$ when gating the signal in the frequency domain is from $N_{x,30\text{KHz}}$ to $N_{y,110\text{KHz}}$, denoting the frequency bins of 30 and 110 KHz, respectively. To retain the original power of the signal, the gated signal is scaled by a factor proportional to the ratio of full FFT and gated FFT, as shown in Eq. (4.23).

$$\text{Scf} = \frac{\frac{\Delta t}{N} \sum_{i=1}^N |H_i|^2}{\frac{\Delta t}{N} \sum_{N_{y,30}}^{N_{y,110}} |H_i|^2} \quad (4.23)$$

The value of S_{cl} can generally be approximated by a factor of 2 when the noise is low, but Eq. (4.23) is used in the deconvolution technique for this chapter.

Figure 4.8: Scaled compensation for deconvoluted toneburst $R_{tb,deconv}(t)$ (solid black line) and toneburst $R_{tb}(t)$ (dash-dotted red line) responses.

The noise removal in the deconvolution process is the result of the inherent gating process in the frequency domain that removes energy and high-frequency noise from the signal, thereby improving SNR. Deconvolution incorporates an inherent passband in the frequency domain due to the numerical artefacts formed in the division of the frequency response of the chirp ($S_c(\omega)$) and toneburst excitation ($S_{tb}(\omega)$). Fig. 4.9 shows the power spectral density (PSD) estimate of the different passband filters applied to the deconvolution of the chirp response. N denotes the frequency bin at 20 MHz. Without a passband filter, the noise generated outside the bandwidth of interest is comparable to the power of the signal. Without a passband filter, the numerical artefacts from the

division of toneburst $S_{tb}(\omega)$ and chirp $S_c(\omega)$ excitation signals outside of the bandwidth of interest introduces substantial noise into the final deconvoluted toneburst $R_{tb,deconv}(t)$. This condition drastically reduces SNR through deconvolution. Hence, the use of a passband filter is critical in achieving SNR improvement through deconvolution.

Figure 4.9 PSD is obtained for the deconvoluted toneburst $R_{tb}(t)$ with no passband filter (solid black line) and with 0–133.33 KHz (dotted red line), 20–120 KHz (dash-dot black line) and 30–110 KHz (blue dash line) passband filters. 20 MHz represents the entire frequency spectrum.

4.4.2 SNR Comparison

SNR in this chapter is obtained through the amplitude of the signal in the time domain.

A Hilbert envelope is applied, and the signal is taken as the magnitude of the first arrival A_0 mode. Noise is taken as the maximum magnitude between the cross talk and the first arrival (i.e. 80–120 μ s). The equation for SNR is

$$SNR = 20 \log \left(\frac{A_{Signal}}{A_{Noise}} \right). \quad (4.24)$$

The two signals previously shown in Fig. 4.8 are similar, except for the high frequency and incoherent noise that are absent from the deconvoluted toneburst ($R_{tb,deconv}(t)$) due to gating. An SNR improvement of approximately 3 dB can be observed from the respective Hilbert envelopes shown in Fig. 4.10.

Figure 4.10: Hilbert envelope applied to the signals in Figure 4.6 compared with the SNR of the deconvoluted toneburst ($R_{tb,deconv}(t)$) and experimental toneburst ($R_{tb}(t)$) of useful signals up to 600 μ s. A 3 dB improvement in SNR is shown.

Although the SNR improvement is reasonable at high voltages, the objective is to reduce the excitation voltage whilst maintaining SNR through deconvolution. This topic will be covered in the following section.

4.5 Voltage Reduction

In industrial applications, the overall power consumption of permanent monitoring systems, especially around gas pipelines, follow intrinsic safety standards [109]. The current and voltage of a system should be below fluid ignition curves to prevent sparks from causing explosions. A frequently used power limitation is 1.3 W. Ideally, voltage should be reduced whenever possible. In this section, the impact of low-voltage excitation of a chirp signal on SNR improvement is investigated.

A reduction in power when using a toneburst results in a low amplitude and therefore a low SNR against systemic and environmental noise.

As presented in the previous section through Parseval's theorem, the power of the signal is proportional to its amplitude squared in the time domain.

$$P = \Delta t \sum_{i=1}^N V_i^2 = \frac{\Delta t}{N} \sum_{i=1}^N |H_i|^2 \dots\dots\dots(4.25)$$

The voltage requirements of the system can be reduced significantly by using a chirp signal at a low amplitude and then deconvoluting to a toneburst. This is done by reducing the amplitude of the chirp signal used and increasing the length of the signal in the time domain. Fig. 4.11 presents 5 V (peak-to-peak voltage unless stated otherwise) chirp excitation $S_{c,5v}(t)$, 50 V toneburst excitation $S_{tb,50v}(t)$, 5 V toneburst excitation $S_{tb,5v}(t)$ and 50 V chirp excitation $S_{c,50v}(t)$.

Case 1

In the first case, transfer function $G_{5v-50v}(\omega)$ is obtained from the division of the frequency response of the 50 V toneburst excitation $S_{tb,50v}(\omega)$ and the 5 V chirp excitation $S_{c,5v}(\omega)$. This transfer function is deconvoluted with the chirp response $R_{c,5v}(\omega)$ obtained from the 5 V chirp excitation $S_{c,5v}(\omega)$ used in the experiment to obtain the deconvoluted toneburst ($R_{tb,5v-50v,deconv}(t)$).

Case 2

In the second case, transfer function $G_{50V-50V}(\omega)$ is obtained from the division of the frequency response of the 50 V toneburst excitation $S_{tb,50V}(\omega)$ and 50 V chirp excitation $S_{c,50V}(\omega)$. This transfer function is deconvoluted with the chirp response $R_{c,50V}(\omega)$ obtained from the 50 V chirp excitation $S_{c,50V}(\omega)$ in the experiment to obtain the deconvoluted toneburst $R_{tb,50V-50V,deconv}(t)$.

Case 3

In the third case, transfer function $G_{5V-5V}(\omega)$ is obtained from the division of the frequency response of the 5 V toneburst excitation $S_{tb,5V}(\omega)$ and 5 V chirp excitation $S_{c,5V}(\omega)$. This transfer function is deconvoluted with the chirp response $R_{c,5V}(\omega)$ obtained from the 5 V chirp excitation $S_{c,5V}(\omega)$ in the experiment to obtain the deconvoluted toneburst $R_{tb,5V-5V,deconv}(t)$.

The selected values are arbitrary to determine if SNR can be maintained or improved when the voltage of the signal is reduced.

In Fig. 4.11(b), the deconvoluted toneburst responses for one of the receivers in the receiver array are shown for all the three cases. The responses shown are $R_{tb,5V-50V,deconv}(t)$ (black dotted line), $R_{tb,50V-50V,deconv}(t)$ (black dashed lined) and $R_{tb,5V-5V,deconv}(t)$ (solid blue line).

In all the cases, the transmitter and receiver are on a straight line in the axial direction of the pipe directly opposite one another. In Case 1, the deconvoluted toneburst, $R_{tb,5V-50V,deconv}(t)$, has the same SNR as the deconvoluted toneburst, $R_{tb,5V-5V,deconv}(t)$, in case 3. The two signals are exactly the same when normalised because the problem is reduced to a scaling problem.

Figure 4.11: (a) 5 V chirp excitation (dash-dotted line), 50 V toneburst excitation (solid red line), 5 V toneburst excitation (dotted line) and 50 V chirp excitation (solid blue line). (b) Normalised comparison of deconvoluted toneburst responses obtained in the experiment for the 5 V–5 V (solid blue line), 5 V–50 V (black dotted line) and 50 V–50 V deconvolution (black dashed line). Only the first arrival is shown for each case. (c) SNR comparison of 5 V–5 V (cross marker) and 50 V–50 V (square marker) deconvolution. The improvement is large at a low voltage corresponding to a low input SNR of chirp excitation, $S_c(t)$.

The transfer function in deconvolution is scaled by a factor proportional to amplitude change A (and by definition, voltage change) in the transmitted toneburst $A.S_{tb}(t)$ used in the deconvolution.

However, a comparison of 50 V–50 V deconvolution with 5 V–5 V deconvolution from cases 2 and 3, respectively, shows that SNR is larger for the latter. Meanwhile, a comparison of the deconvoluted toneburst responses from cases 2 and 3 with the corresponding toneburst responses $R_{tb,50V}(t)$ and $R_{tb,5V}(t)$, respectively, indicates that the improvement in SNR through deconvolution is smaller in the 50 V case. This result is observed in Fig. 4.11(c), which shows the SNR of the deconvoluted toneburst response and its corresponding toneburst response for the same excitation voltage at 5 and 50 V, respectively. The SNR is obtained from the signal for the first 350 μs to consider only the first arrival. In this case, only the fundamental helical order is included, which refers to the first arrival helical mode in a pipe configuration that travels the shortest distance between the transmitter and receiver.

Without deconvolution, a reduction in voltage by 10 times results in an SNR drop from 25.8 dB to 11.1 dB. This phenomenon is shown in Fig. 4.11(c). However, by deconvoluting the signal, the SNR of the signal at 5 V can be improved to 28.5 dB (improvement of 16 dB). Deconvolution at 50 V has a smaller improvement of approximately 3 dB to 39.8 dB, suggesting a relationship between the SNR gain obtained through deconvolution and the initial SNR of the chirp signal used before deconvolution. The small SNR gain at a high voltage can be related to inherent ripples formed due to the gating function applied in the deconvolution method. These ripples are possibly significant at high voltages because they increase with increasing input SNR of the chirp. The next few sections focus on the work performed to validate this hypothesis.

4.6 2D Wave Simulation of Propagation for Chirp and Toneburst Excitations

A Matlab simulation is performed to propagate signals in a pipeline array of 24 transducer transmitters and 24 transducer receivers and further investigate the trend of SNR improvement in deconvolution.

The propagation of a plane wave in a flattened and replicated domain can be described as

$$u=AC(\omega)e^{i(k_x r)} \quad (4.26),$$

where r is the distance between the transmitter and receiver; k_x is the wavenumber magnitude in the x -direction; C is an arbitrary complex function representing the excitation signal in the frequency domain and A is the amplitude of the excitation signal.

The replicated domain [110] is introduced to acquire information on high-order modes. In pipelines, the modes travel on a straight line (order 0) and helically around the pipeline. The number of turns represents the order, and the sign represents the direction (+1 denotes one counter-clockwise turn and -2 denotes two clockwise turns). Fig. 4.12 shows a pipeline where the excitation from one transducer source results in the propagation of multiple helical modes that travel around the pipeline. The number of modes that can be seen is a function of distance between the two arrays and acquisition time.

Figure 4.12: 3D physical model showing the transmitter and receiver arrays and high-order helical paths; (b) order 0 is the unwrapped section of the pipe, and other orders are added to both directions along the x axis to replicate the cyclic characteristic of the pipe. In the pipe, the measured signals at R_0 correspond to the superposition of the signals from each path on the plate. The arrows present various propagation paths between a transmitter and a receiver [61].

The pipeline can be unwrapped into a plate (Fig. 4.12(b)), and high orders can be obtained based on the helical path distance through the replicated domain. The straight path distance between each selected transmitter and receiver for each order [111] is

$$r_{m,q,R} = D \sqrt{1 + \left(\frac{L}{D}\right)^2 \times \left(m + \frac{(R-q)}{n}\right)^2}, \quad (4.27)$$

where r is the distance between the receiver and transmitter; subscripts m , q and R represent the helical order, source number and receiver number, respectively; D is the distance between arrays; L is the perimeter of the unwrapped pipe and n is the number of receivers in the array.

Fig. 4.13(a) shows the chirp excitation for each transducer at time $t=0$. The propagation of each signal is a function of the distance between the transducer and receiver, and it increases for high-order helical modes. In an acquisition time of 1000 μs , five orders (i.e. 0, +1, -1, +2 and -2) can be obtained.

The response of the signal obtained based on the number of orders selected is

$$R_{c/t,m} = \sum_{-m}^{+m} U_{c/t,m}, \quad (4.28)$$

where subscripts c and t represent a chirp or toneburst excitation.

This scenario is visualised in the time domain in Fig. 4.14(a) for five helical modes. The number of orders that can be obtained is dependent on the initial length of excitation of the chirp, the data acquisition time and chirp bandwidth. The signal is highly dispersive when the bandwidth is large. This situation may cut off the +2 and -2 high-order modes at the 1000- μs mark. In GWT, the use of a higher number of orders increases the accuracy of reconstruction for limited view problems as is the case for arrays on a pipeline. The reduction in the number of orders due to deconvolution is therefore undesirable and should be avoided by selection of an optimum data acquisition and signal bandwidth for a particular chirp length.

Figure 4.13: Separation of the experimental time traces. The measured time traces are replicated four times (from -2nd to $+2\text{nd}$ order) in the (a) initial chirp excitation and (b) measured chirp response after propagation.

Figure 4.14: Time trace for receiver 12 when the source is 12. (a) Chirp response, including orders 0, $+1\text{th}$, -1th , $+2\text{th}$ and -2th . (b) Noise added with $\text{SNR} = 12.4 \text{ dB}$.

Noise with a mean (μ) of 0.2 and standard deviation (σ) of 0.05 is subsequently added to the chirp and toneburst responses (Figs. 4.15(a) and 4.15(b)) to simulate environmental noise. The signal is then deconvoluted to observe the SNR improvement.

The deconvoluted toneburst ($R_{tb,deconv}(t)$) is compared with a toneburst ($R_{tb}(t)$) that is propagated using the same method above with the same amount of added noise. The SNR of the toneburst response ($R_{tb}(t)$) is 9.1 dB, whereas the deconvoluted toneburst ($R_{tb,deconv}(t)$) has an SNR of 36.2 dB. The time traces for these signals are shown in Fig. 4.15 (b).

Figure 4.15: (a) Toneburst with SNR=9.1 dB and (b) chirp response with SNR = 12.1 dB (red dotted line).

4.6.1 SNR Improvement with a Chirp Signal

Laboratory experiments indicate that as the voltage decreases (the SNR input decreases), the improvement through deconvolution increases. This occurrence is attributed to the convolution of the signal into noise in the frequency domain due to the filtering effects from Fourier transform. The passband filter causes jump discontinuity in the frequency domain. The inverse Fourier transform of the frequency response of the deconvoluted toneburst, $R_{tb,deconv}(\omega)$, results in the formation of ringing artefact in the time domain, which is proportional to the signal amplitude in the frequency domain. The observed ringing artefact is the direct result of convoluting a rectangular passband filter in the frequency domain. The inverse Fourier transform of a rectangular filter is a Sinc function in the time domain, with the ripples of the function forming throughout the data acquisition time. A discussion on ringing artefact and the factors that contribute to it will be presented in Section 4.6.2.

Signals are propagated with amplitudes varying from 0.05 V to 25 V at 0.05 V increments to investigate under a wide spread of data whilst keeping the noise constant during the simulation of the changing initial SNR of the chirp used. The SNRs before and after the deconvolution are recorded. Fig. 4.16 presents the SNRs for the simulated chirp, toneburst signals and deconvoluted tonebursts as functions of input SNR.

Figure 4.16: SNR for the simulated toneburst response R_{tb} (dash-dotted line), simulated deconvoluted toneburst response $R_{tb,deconv}$ (solid black line) and simulated chirp response R_c (dotted line) and the theoretical SNR of the original toneburst response R_{tb} (dashed line).

The SNR gains of the simulated toneburst and chirp responses are compared to validate the simulated propagation in Matlab. The SNR gain in the simulation shows good agreement with the theoretical prediction obtained in Eq. (4.18). Notably, the theoretical prediction is based on a fixed frequency of 70 KHz, but in reality, a frequency band of 30–110 KHz exists. Therefore, a slight difference in SNR improvement is observed in the simulation.

Figure 4.17: Theoretical SNR (dash-dotted) and Simulated SNR (solid line) of the toneburst and chirp response signals are measured for different input voltages, and their corresponding gain is plotted.

The improvement is large for small input SNRs after deconvolution, and this improvement decreases with increasing initial SNR, which was previously observed in the experimental measurements. The SNR of the deconvoluted toneburst, $R_{tb,deconv}(t)$, plateaus as the input SNR increases. Deconvolution causes a reduction in SNR as the input SNR increases beyond the intersection point between the deconvoluted toneburst SNR curve, $R_{tb,deconv}(A,t)$, and the toneburst SNR curve, $R_{tb}(A,t)$, corresponding to an SNR input of 40 dB. This phenomenon suggests a limitation of this method. The limit is dependent on the intersection point between the deconvoluted toneburst response, $R_{tb,deconv}(t)$, and the toneburst response, $R_{tb}(t)$, curves. However, the input SNR is rarely, if never, this large in practical applications. Therefore, this limit is not a consideration for this case.

Figure 4.18: SNR for the toneburst response (R_{tb}) and deconvoluted toneburst response ($R_{tb,deconv}$). The experimental data (seen in Figure 4.12) validate the simulation.

The experimental data from Fig. 4.11(c) are fitted into Fig. 4.18, and they show consistency with the simulated data. The corresponding SNR_{input} values for the experimental toneburst SNR at 5 and 50 V are 8.2 and 37.4 dB, respectively. The trend observed in the experiment is in good agreement with the expected improvement obtained in the simulation.

Fig. 4.18 shows the possibility of reducing the voltage when the input SNR of the chirp is lower than 40 dB. Assuming that the voltage and amplitude are directly proportional, which is a reasonable assumption in an LTI system, the x axis in Fig. 4.18 is

interchangeable between the voltage and input SNR (linear relationship). The voltage can be reduced from 50 V to 10 V (almost 80% reduction in voltage) whilst maintaining a similar SNR output.

4.6.2 Discussions

The plateauing effect in Fig. 4.18 is believed to be due to the rectangular passband filter applied in the frequency domain. The rectangular filter causes a jump discontinuity in the frequency domain response. Ripples or ringing artefacts form in the time domain due to the convolution of the rectangular filter with the frequency response of the chirp, $R_c(\omega)$. Accordingly, an additional sinc function is formed in the time domain. A rectangular filter in the frequency domain can be expressed as

$$H_r(f) = A \left[\text{rect} \left(\frac{f}{2B} \right) \right], \quad (4.29)$$

where A is the filter amplitude corresponding to the maximum amplitude of the deconvoluted toneburst frequency response at central frequency, $R_{tb,deconv}(\omega_0)$, and B is the bandwidth of the filter. The inverse Fourier transform can be expressed as

$$h_r(t) = F^{-1}[H(f)] = 2AB \text{sinc}(2Bt). \quad (4.30)$$

The ripples in the sinc function decrease with a wide filter, corresponding to the increasing frequency of the sinc function in the time domain where the ripples are small at a fixed recording time. The ripples also increase with a large amplitude filter, corresponding to the amplitude of the deconvoluted toneburst response at the central frequency of 70 KHz ($R_{tb,deconv}(\omega_0)$), which is denoted by A in Eq. (4.30).

Fig. 4.19 presents the frequency response of the deconvoluted toneburst signal, $R_{tb,deconv}(t)$, for $k=1$ and $m=1$. The jump discontinuity is the cut-off point in the frequency response due to the passband filter. This discontinuity causes ringing artefacts to form in the time domain after IFFT due to the convolution of a rectangular filter.

Figure 4.19: Ripples or ringing artefacts formed from the inverse transform of the rectangular filter in (b) are seen in (a). The ringing artefacts are formed for the entire acquisition period.

The amplitude of the formed ringing artefacts is proportional to that of the signal terms in the frequency domain. The passband filter is adjusted to demonstrate this effect. The frequency response of the deconvoluted toneburst, $R_{tb,deconv}(\omega)$, is plotted for 20–120 KHz, 30–110 KHz and 40–100 KHz passband filters and is shown in Fig. 4.20.

Figure 4.20: Frequency response of the deconvoluted toneburst, $R_{tb,deconv}(\omega)$, for 20–120 KHz (dashed black rectangle), 30–110 KHz (solid black rectangle) and 40–100 KHz (dashed red rectangle) passband filters.

For high SNR inputs where the added white noise is negligible, the noise generated after deconvolution is formed exclusively due to the ringing artefacts. The chirp response is deconvoluted for 40–100, 30–110 and 20–120 KHz passband filters in two cases to validate this idea. In the first case, noise is added before deconvolution. In the second case, the clean chirp response is deconvoluted, and the noise generated after deconvolution is attributed to the ripples formed due to the application of a rectangular filter in the frequency domain. Noise is measured for increasing amplitudes of the chirp, $S_c(t)$, and toneburst, $S_{tb}(t)$, in the time domain after deconvolution. Fig. 4.21 presents the three plots.

Figure 4.21: Noise measured after deconvolution for a chirp response with and without noise plotted for increasing amplitudes of excitation for 20–120, 40–100 and 30–110 KHz.

In Fig. 4.21, the noise after deconvolution for signals with and without noise merges at high initial amplitudes of the chirp. SNR is represented by the inverse gradient of each line and plateaus for high input SNR of the chirp response, $R_c(t)$, based on the linearity of each line. As the amplitude increases (corresponding to an increase in SNR input of the chirp), the output noise measured after deconvolution of a chirp signal with and without input noise becomes equivalent. This phenomenon suggests that when the ringing artefact formed is larger than the deconvoluted noise, the noise generated after deconvolution is attributed to only the ringing artefact. Fig. 4.20 illustrates that the noise generated in the time domain increases as the amplitude increases. This occurrence is due to the increase in the height of the filter of the chirp signal, $R_c(\omega)$, with increasing

amplitude. The height of the rectangular filter increases, and the ripples formed from the IFFT of the filter increases.

The gradient of each line is the inverse of the output SNR. The inverse of the slope for the linear trends observed in Fig. 4.21 causes the SNR plateau seen in Fig. 4.18. The noise measured after deconvolution is attributed only to the ringing artefacts when the added white noise is negligible. This occurrence is observed for high initial SNR inputs where the output SNR remains the same even with increasing input SNR (increasing amplitude). The formed artefacts are proportional to the signal amplitudes in the frequency domain. The noise increases with the increase in SNR input, resulting in a plateau of the SNR output for the deconvoluted toneburst, $R_{tb,deconv}(t)$.

This plateau can be observed by plotting the logarithmic output SNR of the deconvoluted toneburst, $R_{tb,deconv}(t)$ with respect to the input SNR of the chirp response, $R_c(t)$, (Fig. 4.22). The plateau starts later for wide passband filters. Wide filters form small ringing artefacts. Consequently, narrower filters are required to increase the ringing artefacts until noise from the environment is negligible. The reason to increase the ringing artefacts is so that the plateau region of interest starts earlier, corresponding to the possibility of further lowering the voltage of the input (equivalent to lower input SNR).

The deconvoluted noise observed in Fig. 4.21 is obtained by the deconvolution of a signal of random noise without any signal wave packets. The noise after deconvolution for signals with and without noise merges at high initial amplitudes of the chirp as mentioned previously. SNR is represented by the inverse gradient of each line and plateaus for high input SNR of the chirp response, $R_c(t)$, based on the linearity of each line. This trend also shows that the plateau begins late for systems with large noise. As

such, there is a trade-off with lowering the voltage through deconvolution and the obtaining a higher magnitude output SNR. For this chapter, the lowering of voltage is given priority alongside the output SNR. As such, the 30 – 110 KHz bandwidth is preferred with respect to the 20 – 120 KHz bandwidth. However, to prevent a further decrease in the output SNR, the 40 – 100 KHz bandwidth is not selected

Figure 4.22: Output SNR of the deconvoluted toneburst, $R_{tb,deconv}(t)$, for 20–120 KHz (dotted line with diamond markers), 40–100 KHz (dash-dotted line with x markers) and 30–110 KHz (dashed line with circle markers) passband filters.

Section 4.4.1 showed that the length (or time duration) of the chirp signal is 10 times longer than the toneburst used in deconvolution. Changing the number of helical orders m , that can be obtained for a fixed acquisition time would change the maximum SNR improvement because the rectangular filter is tall for a high number of helical orders. The increase in helical orders increases the height of the rectangular filter in the

frequency domain. For a high number of helical orders, many signal terms are present, which result in a large amplitude of the chirp frequency response. A tall rectangular filter increases the ringing artefacts formed from IFFT and observed in the time domain. This situation results in a decrease in the observed plateau SNR as shown in Fig. 4.23. The figure plots the output SNR of each deconvoluted toneburst, $R_{tb,deconv,m}(t)$, against the input SNR of the chirp response, $R_c(t)$. The increase in helical orders, m , reduces the output SNR as expected. Fig. 4.23 shows that with a chirp length of $k=10$, increasing the number of orders (helical modes) that can be seen reduces the SNR improvement after an initial SNR input of 15.6 dB or $x=6.1$ where the curves start to plateau (denoted by the vertical black dotted line). This portion is where the signal amplitude is six times larger than the maximum noise amplitude before deconvolution. When $m=5$, the plateau is 39.4 dB, and the maximum SNR values for $m=3$ and 1 are 41.4 and 42.4 dB, respectively.

Figure 4.23: Simulated deconvoluted toneburst with three (dotted line), five (solid line) and one (dash-dotted line) number of helical orders, m .

Although reducing the number orders improves SNR, defect reconstruction accuracy decreases due to the limited view problem for a cross-hole geometry. For this reason, the number of orders should be the maximum for a given acquisition time. The maximum number of orders, which is affected by toneburst length, propagation distance and acquisition time, can be selected easily.

Chirp length must also be monitored apart from toneburst length. If five orders are required, then five orders of the chirp signal would be obtained for a given acquisition time. For the selected toneburst length, a chirp with length $k=10$ (10 times longer) or $429\text{ }\mu\text{s}$ and a bandwidth between 30 and 110 KHz can satisfy the criteria of obtaining five helical orders. This notion means that the acquisition time and selected toneburst, $k=10$, should be the upper limit for the distance between arrays.

The deconvolution limit decreases with a decrease in the maximum SNR obtained in the region of interest. This situation is due to the curves of the deconvoluted toneburst, $R_{tb,deconv}(t)$ and toneburst response $R_{tb}(t)$, intersecting earlier, as was previously shown in Fig. 4.18. Such a notion implies a reduction in the region of interest up to the point where the maximum SNR obtained through deconvolution is equivalent to the SNR of the toneburst response. (intersection point)

If limited view is not a criterion, then decreasing the number of orders at high input SNR is preferred. However, SNR improvement through deconvolution remains constant only when the system input SNR is within the region of interest.

4.7 Summary

The chapter presented a means to optimise high SNR by using deconvolution with low voltage excitation. The SNR improvement at low input SNR is larger than that at high input SNR. This endorses the method of deconvolution for low input SNR cases, which are either inherent (due to large system noise) or externally forced (by reducing the voltage).

The initial SNR input is important because it affects the location of the deconvolution improvement in a trend that increases the SNR improvement of the deconvoluted toneburst with increasing input SNR. This SNR improvement approximately plateaus after an input SNR of 15 dB. However, a slight voltage reduction in this region is still possible before this starting limit. The plateau in SNR is due to the passband filter used in deconvolution. The inverse transform of a rectangular filter in the frequency domain can be represented as a sinc function in the time domain. The ripples or ringing artefacts caused are proportional to the filter width and height. A wide passband filter reduces the ringing artefacts in the time domain, and a tall filter increases the ringing artefacts in the time domain.

Two scenarios could pose an issue in GWT. Firstly, a low maximum SNR could be obtained at high frequencies due to high-order GW modes in the response. These signal components increase the amplitude of the response in the frequency domain. Accordingly, a tall rectangular filter is needed, resulting in large ripples that reduce the maximum SNR obtained after deconvolution. High frequency is not an issue for low-frequency A_0 mode transduction, where the A_0 mode generated is almost pure in out-of-plane excitations.

Secondly, the obtained maximum SNR in liquid loading could be low because of the generation of quasi-Scholte waves in the response, which increases the signal components within the bandwidth for a fixed acquisition time. On-site application in a water-loaded pipeline is presented in the next chapter to investigate the possibility of low-frequency voltage reduction through deconvolution in liquid loading.

This chapter discussed the possible factors that affect the trend observed in SNR improvement through deconvolution for different input SNRs. This study translates into the varying input voltages and corresponding SNR improvement from the deconvoluted toneburst response. The region of interest is where the formed ringing artefacts are larger than the environmental noise in the signal before deconvolution. The SNR of the deconvoluted toneburst plateaus in this region, which indicates that any reduction in voltage in the region has no effect on the output SNR. The next chapter will apply the findings of the current chapter with a voltage reduction through on-site deconvolution application.

Chapter 5: System Reliability and Stability on-site and Guideline Implementation

5.1 Introduction

This chapter focuses on the application of the transducers designed in Chapter 3 and the implementation of deconvolution to maintain SNR whilst reducing the voltage. Low voltage deconvolution was presented in Chapter 4. Permanent monitoring of structures can reduce the down-time in inspection whilst providing the owners a risk management of their structures and extending the structures' life.

A permanent monitoring system requires a test of system stability over time and an assessment of the reproducibility and accuracy of the results. Result accuracy represents the confidence in the measured data. Most inspection tools have a degree of inherent uncertainty, which must be accounted for in any model prediction. The recorded data can be put into a framework with ASME B31G or DNV RPF101 guideline codes [19, 74] that translate the thickness loss due to corrosion to a change in the allowable operating pressure. Although this framework is applied to a water pipeline by using the FWI algorithm, it can be eventually implemented with any pipeline with GWT inversion algorithm. Here, an assessment is performed with FWI for a water pipeline; such an assessment can be reproduced for high-class pipelines in an offshore setting of varying uncertainties.

Analysis of the pipeline on-site shows that voltage can be reduced by almost half through deconvolution without any loss in SNR. The use of deconvolution to improve SNR intuitively means that voltage can be reduced, but in these cases, the SNR achieved after deconvolution is decreased. By applying the work performed in Chapter 4, a

similar SNR is achieved on-site for the high- and low-voltage chirp generation cases after deconvolution.

In this chapter, the on-site setup is presented, and the stability of the transducers designed in Chapter 3 is monitored over five months of measurement. The stability is analysed in terms of the process capability index (C_{pk}). The improvement in SNR through deconvolution is then presented in terms of an improvement in the C_{pk} value. Subsequently, the benefit of deconvolution in improving SNR and extending the life of the prototype (array of transducers and signal processing through deconvolution) is reviewed. Thickness construction from measurements obtained on-site is then presented. The DNV allowable corrosion pressure guideline is outlined and applied to the region with the largest thickness loss over the five months of monitoring. Finally, the probability of failure of the pipeline is predicted.

5.1.1 On-site Setup

Figure 5.1: Two sets of arrays, each with 24 transducers that are spring-loaded to ensure the generation of UT signals in the pipe wall, are presented. The pipe is above a fire water line and is water-loaded with a pipe wall thickness of 10.5 mm. It is located at Jurong Island, Singapore.

The two sets of arrays are connected to a Lecouer system, which in turn is linked to a central processing unit. The water-loaded pipe here results in a lower SNR compared with that of the signals observed in the laboratory experiment due to the attenuation of GWs into the water medium and the generation of quasi-Scholte waves. The SNR is measured similarly to that described in Chapter 4.

5.2 Stability of Transducers and Process Stability

5.2.1 Stability of Transducers

In this section, thermal fluctuations and transducer degradation is investigated because the prototype is left under all weather conditions over time. Figs. 5.2(a)–5.2(c) show two instances one month apart, where the tone burst response is recorded for a receiver and source. Although the two signals are not the same, they are similar. The difference in the results are believed to be from the slight degradation of the transducers over time because they are kept in the pipeline regardless of the weather condition and from the fluctuating environmental noise at the moment of measurement.

Figure 5.2: Measured response for (a) Source 1, receiver 28; (b) source 1, receiver 30; (c) source 6, receiver 48; and (d) average SNR variation of source 1 for the first month of measurement (December).

Fig. 5.2(d) shows the average SNR difference for each of the receivers for Source 1, with all the receivers having no more than a 5 dB variance from the average SNR change for all the receivers. The average SNR difference is close to 0 dB.

5.2.2 Process Stability

The distribution can be approximated to a normal one to be analysed for statistical stability. The average SNR fluctuation for all sources is obtained by averaging the average SNR change for each source. Fig. 5.3(a) shows the average SNR fluctuation for each source and the average SNR fluctuation for all sources. The average drop in SNR is 0.034dB, almost negligibly. The variance is obtained for the SNR fluctuation of each source with respect to the average SNR fluctuation for all sources and is used as a

measure of the degree of variability in the process and is shown in Fig. 5.3 (b). An average variance of 2.19 dB is observed and is depicted in Fig. 5.3 (b).

Figure 5.3: Averaged SNRs for all receivers for one source. The average SNR fluctuation for two consecutive months for each source is compared. (a) Ave. drop in SNR = 0.034 dB. (b) The variance observed between the two months of measurement is $\sigma_{ave} = 2.19$ dB.

In order to analyse the process stability, the SNR obtained for each source-receiver (24x24 data points) is plotted into a distribution to monitor the process for transducer stability. Fig. 5.4 shows the cumulative probability for the distribution of SNR for all the data points in the first month of measure and indicates that the distribution of the collected data shows good agreement with a normal distribution. Accordingly, the data is approximated with a normal distribution and used to analyse process capability.

Figure 5.4: Cumulative probabilities for the average SNR observed for the first (thick black line) and second (dashed black line) months of measurement plotted against the corresponding normal fit distributions (dotted and dashed-dotted lines, respectively).

Process capability, which is defined in this chapter as a statistical measure of the inherent process variability for the average SNR change for different months of measurement, is analysed using a six-sigma approach to determine if a process is in statistical control (low variability). The process capability index, which is the statistical measure of the process capability is defined as

$$C_{pk} = \min\left(\frac{\bar{x} - LSL}{3\sigma}, \frac{USL - \bar{x}}{3\sigma}\right). \quad (5.1)$$

Here, the upper sigma limit (USL) is not defined (one-sided specification), thereby reducing Eq. (5.1) to

$$C_{pk} = \frac{\bar{x} - LSL}{3\sigma}, \quad (5.2)$$

where the lower sigma limit (LSL) is defined as 6 dB (the signal is twice the noise).

A higher value of C_{pk} is equivalent to a system with a better average SNR output. A drop in C_{pk} values between 2 consecutive months indicates SNR degradation. Process capability is defined as

$$C_p = \frac{USL - LSL}{6\sigma}. \quad (5.3)$$

A value of $C_p = C_{pk}$ means that the process is centred over the mean. In terms of SNR, this indicates an average output SNR for each source and receiver is equivalent to the mean SNR output of the system. However, in this chapter, we compare only C_{pk} values as a measure of improvement.

Figure 5.5: Normal distribution obtained from the average SNR from all receivers for varying sources of 1–48 for the first (dash-dotted thick black line) and second (dashed black line) months of measurements. The thick and thin vertical black lines represent the mean of the second and first months of measurement respectively.

Stable processes are defined by the reproducibility of each distribution over a fixed mean. In Fig. 5.5, the normal distribution for the SNR obtained from measurements for the first and second month is shown. The transducers are generally stable and reproduce nearly the same readings at two separate occasions one month apart, with the C_{pk} values in December and January being 0.387 and 0.382, respectively, as is shown in Fig. 5.5. However, an average mean SNR deterioration of 0.03 dB is observed, which is acceptable considering that the transducers are fixed and susceptible to changing weather conditions. As such, the measurement process is considered stable.

The minimum value for C_{pk} should be 1, where the mean of the process is $+3\sigma$ from the defined LSL of 6dB. $C_{pk} < 0$ indicates failure while increase positive values of C_{pk} indicate improvement in SNR. The low C_{pk} values obtained from Fig. 5.5 suggest a need for deconvolution for SNR improvement and is presented in Section 5.3.

5.3 Deconvolution Stability

The deconvolution method presented in the previous chapter is applied to increase SNR, thereby increasing C_{pk} . The tone burst recorded in January is deconvoluted with a chirp. The tone burst length used on-site is 28.9 μ s, and the corresponding chirp factor, k , is 15, which is the maximum length for seeing all five high-order helical modes. Fig. 5.6 shows that both data points are approximated to normal distributions for analysis and shows good agreement.

Figure 5.6: Cumulative probabilities for the average SNR observed for the deconvoluted tone burst (dash-dotted line) and tone burst (black line) responses for the first month of measurement plotted against the corresponding normal fit distributions (dashed and thick black line, respectively).

Fig. 5.7 presents an improvement of 7.04 dB or an increase from 0.39 to 0.52 in process capability index through deconvolution. This improvement in deconvolution is summarised in the following section for all months that the prototype is analysed.

Figure 5.7: Normal distribution obtained from the average SNR from all the receivers for varying sources of 1–48 for the deconvoluted tone burst (thick black line) and tone burst (dashed black line) responses for the second month of measurement.

The improvement in process capability through SNR improvement is acceptable, but an inadvertent benefit of extending the life of the prototype can be obtained and is highlighted in the following section.

5.4 Life Extension of the Prototype

Figure 5.8: Average SNR for each month of measurement for toneburst and deconvoluted toneburst responses. The minimum SNR of 6dB is indicated by the blue dashed line.

In Fig. 5.8, the mean SNR obtained from measured toneburst responses and deconvoluted toneburst responses through chirp excitation is presented for each month of measurement. The average SNR for the tone burst responses drops by -0.034 dB from 14.95 dB in December to 14.92 dB in January and then by -0.28 dB to 14.64 dB in February (blue dotted line). In March, the tone burst SNR improves by 0.17 dB to 14.81 dB and then drops to 14.24 dB in April. The deconvolution improvements in January, February, March and April are 7.04, 6.99, 6.94 and 7.02 dB, respectively, in comparison with the average SNR from the tone burst response for each corresponding month. The transducers monitored over six months do not show any major degradation.

Instead, they show a gradual SNR drop over time with slight fluctuations due to changing environmental conditions at the point of measurement. The stability of system SNR over time shows its suitability as a permanent monitoring system. However, SNR is expected to decrease to below 6 dB over time due to transducer degradation under unpredictable weather conditions.

Deconvolution can be implemented in two ways here. Firstly, deconvolution is a means to extend the life of the prototype by increasing the average SNR, keeping it away from the 6 dB minimum. The minimum is set to ensure that the majority of signals for each month of measurement is at least twice the magnitude of the system noise. The second one, as was previously described in Chapter 4, is to lower the voltage used whilst maintaining the improved SNR and process capability. The latter is presented in Section 5.5.

5.5 Low Voltage Generation

As mentioned in the previous chapter, deconvolution can be used to reduce the voltage of the generating system whilst maintaining SNR. The input SNR of the chirp is within the region of interest and corresponds to a tone burst response of approximately 15 dB. During on-site application, the voltage is reduced through trial and error. The voltage for the chirp on-site is 70 V peak to peak. The normal distribution for the deconvoluted toneburst data obtained from a 30V and 70V chirp excitation is presented in Fig. 5.9. Fig. 5.9 shows that a similar SNR and process capability index, C_{pk} can be obtained when the voltage is lowered from 70 V to 30 V.

Figure 5.9: Normal distribution of the SNR output from the 30 V deconvoluted tone burst response (thick solid line) and 70 V deconvolution (thick dotted line). The vertical dash-dotted line represents the mean for both distributions.

The voltage of the chirp and not that of the tone burst determines the SNR after deconvolution. This statement implies that the SNR of a 30 V chirp deconvoluted to a 50 V tone burst is similar to the SNR of a 30 V chirp deconvoluted to a 30 V tone burst, as mentioned in the previous chapter. The mean SNR in Fig. 5.9 is 21.3 dB.

It should be noted that the maximum SNR obtained through deconvolution is half of that observed in the laboratory experiment on an empty pipe (Chapter 4). With liquid loading, the formation of quasi-Scholte waves contributes to the signal terms within the bandwidth. The increase in signal components increases the maximum amplitude of the chirp frequency response, requiring a tall rectangular filter in deconvolution. This condition increases the ringing artefact observed in the time domain and reduces the maximum SNR obtained after deconvolution. Furthermore, the A_0 mode attenuation in liquid loading is larger than that for an empty pipe. Hence, this reduction in the maximum SNR is expected.

Fig. 5.10 shows all the C_{pk} values for each month of measurement for the tone burst and deconvoluted tone burst responses. A high value corresponds to good SNR and is therefore desirable.

Figure 5.10: Summary of process capability values for the tone burst (dotted black line with circle markers) and deconvoluted tone burst (black line with star markers) responses at 50 V. The process capability of the 30 V deconvoluted tone burst is denoted by a diamond marker for the last month (April) of measurement.

The degradation of the transducers can be clearly observed in terms of process capability. Here, the benchmark is for C_{pk} to be above 0. A negative value indicates ‘failure’, where the SNR is below 6 dB. The low voltage generation of 30 V has a similar C_{pk} value as the 70 V generation and reaffirms the findings presented in Chapter 4. In the next section, the pipeline life prediction is presented to use the process measurements described previously to predict pipeline failure.

5.6 Pipeline Life Prediction and Intervention (DNV-RP-F101)

Assessing the reliability of the pipeline system is critical for system confidence. A risk assessment or inspection guideline can be formulated after validating the reliability of system data. For the purposes in this section, the pipeline is assumed to have only internal loading. With the values of pipe wall loss in the hotspot of the monitoring area, a means of applying the data to a corrosion assessment with recommendations is the next step in providing a complete solution for industries. Only failure modes from corrosion are considered in the following sections. Other modes of failure must be considered in situations wherein external loading is an issue. The guideline implemented here involves summarising the design of software and hardware for permanent monitoring with GWT.

A few guidelines can be used to assess corrosion on a pipe with regard to operating pressure and burst capacity [19, 74, 75, 112]. Two guidelines will be discussed here. The first one is ASME-B31g [74], which uses flow stress theory to account for large-scale yielding and a semi-empirical fracture mechanics formula to calculate the remaining allowable pressure strength of corroded pipelines. However, a large safety factor of 1.1 is used in the governing equations, making this approximation highly conservative. Moreover, the equations apply only to corruptions with smooth contours and low stress concentrations. The second guideline, DNV-RP-F101 [19], provides capacity and acceptance equations (covered in a later section) on the basis of a large number of FE analyses and full-scale burst tests. The limitation is that the equations are used for rectangular-sized defects only, making this approximation conservative. However, for complex-shaped defects, a methodology can be used to split the defects into several rectangles and make the calculated allowable pressure less conservative.

This recommended practice is available to any user and provides a good estimate of the actual failure pressure of a corroded pipeline [113]. The DNV guideline will be used in the following assessment, and the theory will be presented in Section 5.6.3.

Uncertainty is critical in the incorporation of any assessment, and it can be categorised into two main types. The first one involves the material properties at the defect location, which is unknown. This uncertainty can be addressed with the incorporation of a conservative safety factor. The selected safety factor must not be too conservative because this is undesirable, and selection of the safety factor can be obtained in the guideline. The second one is the uncertainty from the type of ultrasonic measurement. FWI based GWT is a relatively new tool for thickness reconstruction, and its accuracy has not been widely investigated. Data sets are available to form a baseline [35]. These data sets are confined to 3–6 mm depth for frequencies below 120 KHz. This concept will be addressed in the following section.

5.6.1 Analytical Uncertainty in FWI and Uncertainty Estimation

The uncertainties involved in ultrasonic measurements have been incorporated into industrial guidelines. For example, in DNV-RP-F101, the uncertainties in UT and MFL measurements are included in the calculation of the allowable corrosion pressure, P_{corr} .

In FWI, these uncertainties stem from depth errors from inversion imaging that are related to the resolution limit. The resolution limit is defined for depth errors, $e_{\text{depth}} < 10\%$, where e_{depth} is shown in Eq. (5.5), and for defect depths of up to 6 mm. FWI performs similar to DT and HARBUT, where defects of up to 1.2λ could be accurately reconstructed for the majority of defect depths between 3 and 6 mm [35]. In Eq. (5.5), T_{min} and T_{meas} represent the original and reconstructed thickness, respectively. This

initial error is compensated for by the incorporation of safety factors according to the DNV guideline.

$$e_{\text{depth}} = \frac{T_{\text{min}} - T_{\text{meas}}}{T_0} \quad (5.5)$$

Errors in measurement in FWI can be a result of different scattering behaviour of GWs with varying thickness and acoustic waves with diverse velocities. The 2D approximation used in FWI leads to inconsistency with signals measured in a 3D elastic model, which is a good representation of reality [60]. This situation constrains the reconstruction accuracy because of the similar resolution limits obtained from FWI, DT and HARBUT. The uncertainty is estimated from the work performed by Rao [28, 29, 35]. However, the study was limited to a wall thickness of 10 mm, and the performed simulation was limited by a defect range of 1–6 mm. Although the study was conducted on plates for 10 mm, errors were analysed as a function of nominal thickness and provided a good approximation for an estimation of uncertainty for a wall thickness of 6.35 mm, as in the case of the pipeline onsite.

For frequencies between 60 and 80 KHz, errors in depth are up to 20% for defect depths up to 6 mm. A large frequency equates to a small error, and a shallow defect equates to a small error, as shown in Fig. 5.11(a). The ratio of top and bottom defect diameters also affects resolvability. Although e_{depth} is consistent for most changing surface diameter defects investigated in [35], the shape of the defect (ratio of top to bottom diameter of the defect) affects whether the defect is underestimated or overestimated. In on-site applications where the original thickness is unavailable, a conservative approach should be adopted to ensure that defect overestimation and underestimation are considered. The larger error between the two is used in uncertainty estimation.

Figure 5.11: (a) Maximum depth error of reconstruction from the elastic model for a 10 mm plate as a percentage of nominal thickness. (b) Cross section along the central line of the reconstructed defects for varying depths at 60 KHz from (a) [35].

For defects with a small bottom surface to top surface ratio, such as pitting corrosion defects, the largest error recorded is a 30% underestimation for defects with a surface diameter close to the resolution limit (1.2λ for a 10 mm plate). For a larger bottom surface to top surface ratio (as the defect approaches a square defect shape), such as gradual shallow surface corrosion defects, overestimation of up to 10% occurs due to low-pass filtering in the algorithm that removes spatial components above a certain frequency [114]. This trend is shown in Fig. 5.12, where the negative errors are overestimations.

Figure 5.12: (a) Maximum depth error for flat-bottomed defects with different ratios of bottom diameter (W_2) to surface diameter (W_1) using data from the elastic model for depths of (a) 6 mm and (b) 3 mm [35].

FWI based GWT has a diffraction-limited resolution proportional to the wavelength used in reconstruction. The depth error changes for different shaped defects, frequency and depth. These errors increase with defect depth and a reduction in frequency primarily due to the increase in resolution at high frequencies. For FWI, the parameters mentioned must be adjusted accordingly to achieve a resolution and minimum depth error of 10% at the most. However, uncertainty can be introduced on site to account for large errors. Considering that the frequency used in the inversion is 60 KHz with defect depths not larger than 6 mm, errors of up to a maximum of 30% can be expected. However, these errors are highly conservative. In general, defects reconstructed on site are Hann-windowed. Approximating these as flat bottom defects, the ratio of bottom to top surface diameters is much less than 0.5.

However, the error must be estimated when the shape of the defect is unknown, as in the case of an application on-site without additional methods, such as UT, or an in-line Pig to validate the results. The assumption is that the defects observed in the inversion have been successfully reconstructed with an error that has to be accounted for. This notion means that the surface diameter of defects can be larger or smaller than the resolution limit seen in the laboratory. A conservative approach is to use the maximum allowable error from the data recorded previously in the laboratory on an open pipe and a liquid-loaded pipe for the smallest surface diameter defect available [29, 35] and to use this as a variation of thickness loss.

The ratio of bottom surface to top surface is less than 0.5 based on the reconstruction shape and assuming flat bottom defects for ease, which is acceptable because the guideline assumes rectangular defects. This result indicates a general maximum underestimation of up to 30% for a surface diameter defect of 1.2λ and defect depth of 6 mm, as shown in Fig. 5.12(a) (defect at resolution limit). This underestimation error for a defect larger than the resolution limit is reduced to 17.5% and 14% for a surface diameter defect of 1.8 and 2.25λ , respectively. The error decreases for shallow defects. However, the errors will be large in a liquid-loaded pipe with attenuation, as shown in [29], due to the large difference in amplitude between waves that travel directly from the source to the receiver and those that are scattered from the defect.

The reconstruction (Appendix D) presents a mixture of defects larger and smaller than the resolution limit. All defect depths are below 3.3 mm. The error is estimated at 20% by using data from Fig. 5.12(b) and accounting for liquid-loaded attenuation.

The e_{depth} estimated here is 20% to account for the liquid loading and attenuation, which is intentionally made larger than that in reality, and to consider defects smaller than the resolution limit seen in the laboratory with large errors. The normal spread of uncertainty accounts for slight changes in pipe wall thicknesses and errors due to a mixture of defects larger and smaller than the resolution limit. On the conservative side, a normal distribution of d/t is assumed, with e_{depth} of 20% equating to $\pm 0.20t$ in the relative sizing accuracy (DNV, 2017).

Table C-2 in Appendix C illustrates that the standard deviation of wall loss thickness to wall thickness, $\text{StD}[d/t]$, at the 80% confidence level is 0.16.

5.6.2 Safety Factor

The safety factor to incorporate depends on the loading of the pipe and its location. Pipelines are classified into safety classes, namely, low, normal and high, each with a decreasing target for annual probability of failure (Table C-1 in Appendix C). This notion indicates that the probability of failure is high when the class is low because of the low risk consequence of the pipe failing. The safety factor accordingly correlates to this idea with high safety factors for high-risk pipelines. The safety factors used in the guideline account for corrosion depth (γ_d), longitudinal corrosion model prediction (γ_m), circumferential corrosion model prediction (γ_{mc}) and longitudinal stress for circumferential corrosion (η).

The safety factor, γ_d , is taken from Table C-3 in Appendix C in accordance with DNV-RP-F101, and the fractile value, ε_d , is calculated based on the following equation. The partial safety factors γ_m and γ_{mc} and ξ are taken from Tables C-4, C-5 and C-6,

respectively, because FWI or any GWT imaging method is considered an absolute ultrasonic measurement.

$$\varepsilon_d = -1.33 + 37.5 \left(\text{StD} \left[\frac{d}{t} \right] \right) - 104.2 \left(\text{StD} \left[\frac{d}{t} \right] \right)^2, \quad (5.5)$$

where $\varepsilon_d = 0.5449$, $\gamma_d = 1.18$, $\gamma_m = 0.85$, $\gamma_{mc} = 0.85$, $\eta = 1.00$ and $\xi = 0.90$.

5.6.3 Guideline Theory for Single and Interacting Defects

A defect assessment is conducted in accordance with DNV-RP-F101 in Section 5.6, and additional information can be found there. However, the principal equations are included here for convenience. The defects can be treated as rectangular or complex. The latter is less conservative and more desirable than the former in industrial applications and is used here.

A defect must satisfy the following conditions for it to be considered isolated.

- 1) The circumferential angular spacing between two defects, φ , must satisfy

$$\varphi > 360 \sqrt{\frac{t}{D}}. \quad (5.6)$$

- 2) The axial spacing between adjacent defects, s , must satisfy

$$s > 2\sqrt{Dt}. \quad (5.7)$$

For a 12 inch nominal diameter pipe (D) and wall thickness (t) of 6.35 mm, defects greater than 88 mm apart in the axial direction and 52° apart in the circumferential direction are treated as isolated defects. Defects that are close to one another should be treated as a single defect and projected onto a single line to form a new combined length as is shown in Fig. 5.13.

Figure 5.13: Formation of a composite defect with two interacting defects.

Interacting defects can be divided into individual ones. The allowable corroded pipe pressure for each defect, P_i , up to the n th defect is

$$P_i = \gamma_m \frac{2t f_u}{(D-t)} \frac{(1 - \gamma_d (\frac{d_i}{t})^*)}{(1 - \frac{\gamma_d (\frac{d_i}{t})^*}{Q_i})} \quad i=1 \dots N, \quad (5.8)$$

where l is the longitudinal length of the corroded region and the length correction factor

$$Q_i = \sqrt{1 + 0.31 \left(\frac{l_i}{\sqrt{D}t} \right)^2}$$

and the relative defect depth

$$\left(\frac{d_i}{t} \right)^* = \left(\frac{d_i}{t} \right)_{\text{meas}} + \varepsilon_d StD \left[\frac{d_i}{t} \right].$$

The allowable corroded pressure of a single defect corrosion subjected to longitudinal compressive stresses and internal pressure should be calculated based on the following equations

$$\sigma_A = \frac{F_x}{\pi(D-t)t}, \quad \text{and}$$

$$\sigma_B = \frac{4M_y}{\pi(D-t)^2 t},$$

where F_x and M_y are the externally applied force and externally applied bending moment on the pipe, respectively. The combined longitudinal stress is

$$\sigma_L = \sigma_A + \sigma_B. \quad (5.9)$$

Appendix E provides additional details on the derivation of the longitudinal stress. The allowable corroded pipe pressure corrected with the longitudinal stresses is

$$P_{\text{corr, stress}} = \gamma_m \frac{2t f_u}{(D-t)} \frac{\left(1 - \gamma_d \left(\frac{d_i}{t}\right)^*\right)^H}{\left(1 - \frac{\gamma_d \left(\frac{d_i}{t}\right)^*}{Q_i}\right)} \quad i=1 \dots N, \quad (5.10)$$

where i represents each defect and the corrective factor to account for stresses

$$H = \frac{1 + \frac{\sigma_L}{\xi f_u A_r}}{1 - \frac{\gamma_m}{2\xi A_r} \frac{\left(1 - \gamma_d \left(\frac{d_i}{t}\right)^*\right)}{\left(1 - \frac{\gamma_d \left(\frac{d_i}{t}\right)^*}{Q_i}\right)}}$$

and

$$A_r = 1 - \frac{d}{t} \Theta,$$

where A_r is the circumferential reduction factor and Θ is the ratio of the circumferential length of the corroded region c to the nominal outside circumference of the pipe.

The minimum value between P_i and $P_{\text{corr, stress}}$ is taken as the allowable corrosion pressure

$$P_{\text{corr}} = \min[p_i, p_{\text{corr, stress}}]. \quad (5.11)$$

Regarding defects that are interacting, the combination length from all combinations of defects (n to m) can be calculated along with the corrected effective depth of the combined defect as follows:

$$l_{nm} = l_m + \sum_{i=n}^{i=m-1} l_i + s_i \quad (5.12)$$

and

$$d_{nm} = \frac{\sum_{i=n}^{i=m} d_i l_i}{l_{nm}}. \quad (5.13)$$

The allowable corroded pipe pressure of the combined defect from n to m (P_{nm}) is

$$P_{nm} = \gamma_m \frac{2t f_u}{(D-t)} \frac{(1 - \gamma_d (\frac{d_{nm}}{t})^*)}{(1 - \frac{\gamma_d (\frac{d_{nm}}{t})^*}{Q_{nm}})}, \quad n, m = 1 \dots N, \quad (5.14)$$

where the length correction factor for a defect combined from adjacent defects n to m in a colony of interacting defects

$$Q_{nm} = \sqrt{1 + 0.31 \left(\frac{l_{nm}}{\sqrt{Dt}} \right)^2},$$

the average relative defect depth for a defect combined from adjacent defects n to m in a colony of defects

$$\left(\frac{d_{nm}}{t} \right)^* = \left(\frac{d_{nm}}{t} \right)_{\text{meas}} + \varepsilon_d \text{StD} \left[\frac{d_{nm}}{t} \right]$$

and

$$\text{StD} \left[\frac{d_{nm}}{t} \right] = \frac{\sum_{i=n}^{i=m} l_i \text{StD} \left[\frac{d_i}{t} \right]}{l_{nm}}.$$

The allowable corroded pressure is the minimum corroded pressure of all the defects and combination of defects.

$$P_{\text{corr}} = \min[p_1 p_2 \dots p_N \dots p_{nm}] \quad (5.15)$$

The acceptable measured defect size can be found by setting the known operational pressure P_{op} to provide

$$\left(\frac{d}{t}\right)_{\text{meas}} = \left(\frac{1}{\gamma_d}\right) \cdot \frac{\left(1 - \frac{P_{\text{op}}}{P_0}\right)}{\left(1 - \frac{P_0}{Q}\right)} \cdot \varepsilon_d \cdot \text{StD}\left[\frac{d}{t}\right], \quad (5.16)$$

where

$$P_0 = \gamma_m \frac{2t f_u}{(D-t)}$$

The allowed maximum defect size to thickness ratio $\left[\frac{d}{t}\right]_{\text{max}}$ is used to monitor defect growth. If the number is exceeded, then an alarm to trigger intervention or reduction in the allowable operating pressure is required.

5.6.4 Analysis of Results

The defect area selected for analysis is the most severe point (deepest defect) in the area of coverage in permanent monitoring. The defect is circled in Fig. 5.14, with the axial and circumferential directions of the pipe denoted by black dashed lines.

Figure 5.14: Thickness loss contour for the region of monitoring between the 2 arrays. Region of interest is circled with the axial and circumferential projection lines (black dashed lines) selected for the cross section. The defect of interest represents the deepest defect in the area of monitoring.

The circumferential direction of the pipe is used to assess defects that are interacting with the axial projection line. These defects satisfy Eq. (5.6). The defects present in both axes are shown in Fig. 5.15 for the first recorded measurement. The region of interest is circled for the cross sections.

The figure demonstrates that only two defects in the axial projection line are interacting (defects 1 and 2). This defect is monitored over five months to assess the burst pipe pressure reduction and allowable corrosion pressure P_{corr} . The number of defects increases over each month of measurement, along with their respective depths. As new defects form, an increasing number of defects are considered to be interacting. Appendix D illustrates the new defects and groups of interacting defects for each month of measurement.

Figure 5.15: Defects in the axial and circumferential projection lines in Figure 5.14 are shown on the right and left, respectively. The defect of interest is in dark grey.

Table 5. 1: Identifying interacting defects/first month measurement

Axis Type	Circumferential Defects					Longitudinal Defects		
Max Length (mm)	$r\Phi_0 = 138.2$					$S_0 = 88$		
Defect	A-X	B-X	X-C	X-D	X-E	1-X	2-X	X-3
Distance between Defects (mm)	$\Phi_{AX} =$ 242.17	$\Phi_{BX} =$ 169.76	$\Phi_{XC} =$ 315.07	$\Phi_{XD} =$ 480.43	$\Phi_{XE} =$ 536.2	$S_{1x} =$ 109.89	$S_{2x} =$ 89.98	$S_{x3} =$ 100.6

Table 5.1 indicates that the two defects that are interacting are axial defects 1 and 2 where the longitudinal distance between the centre of both defects is less than 88mm.

Each allowable corrosion pressure, $P_{corr,n}$, where n is the defect number, is calculated individually and then together for interacting defects to obtain the minimum by using

Eqs. (5.11) to (5.15). The lower value between the two is selected. Fig. 5.16 shows the assessment of allowable corrosion pressure for each month over the span of five months.

Figure 5.16: Maximum allowable corrosion pressure, P_{corr} , over the five months of monitoring obtained from Eq. (5.15).

The allowable corrosion pressure for all months of monitoring is determined by the interaction amongst defects 1, 2 and 4, except for the first month wherein defect 4 has yet to form (Appendix D). The implemented model is highly conservative, assuming that each defect is rectangular. This characteristic combined with an incorporated uncertainty of 20% provide sufficient room to determine an acceptable corrosion pressure for the pipeline. The current operating pressure of the pipeline is 12 bar, which is 12.42% of the last recorded P_{corr} in April. Therefore, no intervention is required as of yet. The maximum allowable pressure under corrosion may be used as an intervention tool to ensure that the operating pressure of the pipeline remains below it to prevent any

leakage. This condition should be followed in high and normal classes of pipelines where leakages result in disastrous consequences. The next section covers the remaining life assessment of the pipeline and the contribution of permanent monitoring to an accurate but conservative estimate.

5.7 Remaining Life Assessment through Corrosion Rate Estimation

The advantage of permanent monitoring using FWI or any other GWT inversion algorithm is its ability to provide a complete depth map of the area of inspection. This feature allows for the observation of the corrosion rate in the area of monitoring, which should remain the same given that the area is under the same environment and loading condition. However, small differences should be expected depending on the defect location because of small changes in bending stress and longitudinal loading. The average corrosion rate used in a predictive statistical model is more accurate when the number of measured points is large. The corrosion rate should then follow a normal distribution, with the uncertainty represented in the form of standard deviation. The respective corrosion rate for the defects is shown below.

Defects that form after a few months of measurement have minimal data for the estimation of the corrosion rate. The uncertainty is high, but these defects will have increasingly accurate and stable corrosion rates as additional data are obtained. Fig. 5.17 demonstrates that defect x (defect of interest) has the highest corrosion rate. The defect is also the deepest point in the area of monitoring and will fail (reach 80% wall loss) before the other defect points of measurement.

Figure 5.17: Defect corrosion rate obtained from the average increase in each defect depth for each month of measurement. The changes in corrosion rate stabilises as additional months of measurement are included. The defect corrosion rates for defects 1 (dash-dotted line with a circle marker), 2 (dotted line with an x marker), 3 (black line with a square marker) and x (thick black line with an x marker) are obtained for all months of measurement. The defect corrosion rates for defects 4 (black line with a circle marker), 5 (black line with an x marker) and 6 (black line with a diamond marker) can only be obtained for the last 3, 2 and 1 month, respectively.

For this reason, a corrosion rate with an estimated uncertainty based on a normal distribution can predict the change in wall thickness over time.

The minimum defect depth at which $P_{\text{corr},x}$ equals operating pressure P_{op} is also shown in Fig. 5.16 and can be estimated from Eq. (5.8) by replacing P_{corr} with the operating pressure of the pipe and taking into consideration the average change in axial length of the corrosion, $l_{x,\text{ave}}$, over the period of monitoring.

From Eq. (5.8), we have

$$\left(\frac{d_{x,max}}{t}\right)^* = \frac{(a-1)}{\left(\frac{a}{Q_{x,ave}}-1\right)\gamma_d}, \quad (5.17)$$

where

$$a = \frac{P_{o,p}(D-t)}{\gamma_m 2t f_u},$$

$$Q_{x,ave} = \sqrt{1 + 0.31 \left(\frac{l_{x,ave}}{\sqrt{Dt}}\right)^2}.$$

This expression provides the maximum defect depth of

$$d_{x,max} = t \left[\left(\frac{d_{x,max}}{t}\right)^* - \epsilon_d StD \left[\frac{d_{x,max}}{t}\right] \right]. \quad (5.18)$$

Figure 5.18: Defect X growth rate.

For the next month of measurement, we can be reasonably sure (approximately 70%) that the depth of the predicted defect is not more than 1 standard deviation from the

predicted value. However, a domino effect of increased uncertainty will occur in the subsequent months. Such an occurrence is the basis of corrosion rate uncertainty. There is a risk of an intervention that is either too early or too late without correcting the predicted growth of the defect with measurements,

Based on a statistical model, the wall loss reaches past the 80% mark in the middle of June and July. However, without a proper assessment either visually or through another measurement, we have no certain way of verifying this. This situation makes monthly measurements highly desirable, but it introduces an inspection schedule. Moreover, the pressure downrating point is expected to be early, that is June with an uncertainty of 32%, to prevent pipe bursts (pitting corrosion that leads to a leak). However, the minor leaks that occur for low-class pipelines that are not operational in terms of turnover are usually allowed to fail first before intervention. The method used alongside GWT allows the maintenance team to be aware of such incidents that are incoming to efficiently prepare an intervention schedule.

A probability analysis is used to determine the probabilities of failure and downrating. This process is carried out to determine the risk of letting the defect grow on the basis of the provided model. In the onshore industry, presenting probabilities is desirable because it allows pipe owners to implement their own solutions at a time when the risk of failure exceeds their acceptable probability of failure which may or may not be in accordance with the DNV code. The probability is calculated for each depth uncertainty through Eq. (5.19) and shown in Fig. 5.19.

$$\text{Prob}(X \leq c) = \frac{1}{\sigma\sqrt{2\pi}} \int_{-\infty}^c e^{-\left(\frac{1}{2}\right) \cdot \left(\frac{x-\mu}{\sigma}\right)^2} dx \quad (5.19)$$

Figure 5.19: Probability of failure after downrating the pressure and probability of failure for the predicted months of growth for defect x. The acceptable probability according to the DNV guideline is 0.01.

The probability of failure after downrating is expected to be smaller than that without downrating the pressure. According to the DNV guideline, the allowable probability of failure of a low-risk pipeline is 0.01. The probability of failure is exceeded in both cases by June and should therefore be monitored closely in the coming months of measurement.

5.8 Summary

This chapter extensively covered the stability of the transducers and deconvolution process by defining the spreads of respective data in terms of a normal distribution and introducing the process capability index, C_p , as a means of comparison.

The C_{pk} value for the first two months of measurement obtained by using a tone burst was 0.387 and decreased to 0.382, signifying a degradation in SNR. Through deconvolution, the process capability was increased to 0.52. The improvement in SNR of 7 dB is consistent with that in a previous work using deconvolution [88]. The average SNR of 21.3 dB was also maintained after voltage reduction from 70 V to 30 V through deconvolution. This finding reaffirms the conclusion that deconvolution can be used in two ways: (1) to increase the life of the prototype by improving SNR through deconvolution, keeping it away from the 6 dB minimum average SNR, and (2) to lower the voltage whilst maintaining the improvement in SNR through deconvolution compared with the conventional original tone burst.

A risk assessment was then performed alongside the use of inversion through (but not limited to) FWI. The allowable pressure under corrosion, P_{corr} , was determined using the deepest defect with the largest corrosion rate. A predictive mode provided the generation of failure probability and downrating based on the defect of interest. However, several months of monitoring showed new defect growths that were not considered after terminating the monitoring. This finding emphasises the benefits of monthly monitoring.

If the pipeline operating pressure is higher than the maximum allowable corrosion pressure in a low-class pipeline, the pipeline can be monitored rigorously through visual

inspections. Meanwhile, for a normal- to high-class pipeline, a pre-emptive pipe clamp can be placed on the pipeline.

The specification of this pipeline is API5L Grade B (GR.B) and ERW schedule 20 ($\times 20$) with a material specification of ASME B36.10M. The pipeline's current operating pressure is 12 bar, with a pipeline wall thickness of 6.35 mm (0.25 in.) The methodology employed in this chapter can be used for high-class pipelines with high restrictions or low failure probabilities. A GWT thickness map can work with an algorithm that employs this analysis to generate a probability of failure map, which can trigger alarms when the acceptable probability of failure is exceeded, which is defined by either the owners or certification bodies. However, it is still the opinion of this author that permanent short-range monitoring and long-range intermittent monitoring should be used concurrently to accurately determine the structural health of pipelines.

6. Conclusions

6.1 Thesis Review

In this thesis, a novel transducer for low frequency A_0 mode wave generation with improved SNR is designed for Guided Wave tomography, and advanced signal processing techniques based on linear frequency modulation was developed to further enhance the SNR of measurements. The reliability of the system is assessed through on-site monitoring of a pipeline for several months.

Low frequency guided wave transducers tend to be bulky and wide, which prevents use in an array configuration on pipelines due to its size resulting in lesser transducers in the array, and due to the pipe curvature of small diameter pipelines. Reducing the size of the transducer results in low amplitude signals which lead to poor SNR. This affects the accuracy of images through the formation of artifacts in thickness reconstruction. The stacked design could increase the energy output of low frequency piezoelectric generators in guided wave applications. Consequently, the SNR of output measurements can be improved. Additionally, the A_0 mode purity, amplitude and bandwidth of the corresponding frequency response are investigated for increasing stack sizes. It was shown that any further increase in stack height beyond 10 2 mm PZTs resulted in only a small improvement in mode amplitude and mode purity, suggesting the optimised design is to use 10 PZTs. Moreover, increasing the stack height of the PZT reduces the central frequency of the generator, which should be considered as a critical parameter in selecting an optimal stack height. In Chapter 3, this central frequency is 0.5MHz.mm. The bandwidth of the stacked PZT is eventually widened with a composite of Alumina and epoxy in a 1:3 weight ratio. The widening of bandwidth has two main benefits for GWT applications. Firstly, a wider bandwidth around a central frequency of

0.5MHz.mm is critical in capturing the dispersive region of the A_0 mode. Secondly, in deconvolution, presented in Chapter 4, a wide bandwidth transducer is necessary when the coded signal has a large bandwidth to avoid spectral leakage and masking of weak scatterers.

The second novelty in this thesis is a signal processing method based on deconvolution and was demonstrated to improve SNR and reduce the voltage of excitation. The use of linear frequency modulated signal or chirp have generally been isolated for the auto correlation of the received signal to improve resolution, whilst reducing sidelobes. Here, the application focused on reverting the chirp signal back to a toneburst, which was previously done by [88]. The study has demonstrated improved SNR proportional with different chirp signals. The main finding of this chapter is that any further increase in input SNR of the chirp excitation results in an eventual plateau of output SNR for the deconvoluted toneburst. The reason for this plateau is due to the inherent convolution of the rectangular passband filter with the frequency response of the transmitted chirp. The inverse fourier transform of this convolution results in the formation of ringing artifacts in the time domain for the length of the data acquisition period that increases with the increase in the height and width of the rectangular filter used in the convolution.

It was then shown that any increase in the number of helical orders increases the maximum amplitude of the transmitted signal, thereby increasing the noise formed and reducing the maximum output that can be obtained through deconvolution. However, in limited view GWT, the maximum number of helical orders for a given acquisition time is critical in achieving a more accurate thickness reconstruction from the measurements. An alternative to compensate for the drop in maximum output SNR achieved from the deconvoluted toneburst is by widening the rectangular filter used in deconvolution. However, widening of the rectangular filter used results in the plateau region starting

for much higher values of input SNR. The consequence of this, as discussed previously, is not being able to lower the voltage of excitation while keeping the output SNR the same, for lower input SNR cases.

The plateauing of the output SNR with increasing input SNR seen in Chapter 4 leads to a possible reduction in voltage used in the application. The application of low voltage deconvolution is seen in Chapter 5 where a drop in voltage from 70V to 30V is implemented with both deconvoluted tonebursts achieving the same output SNR of 21.3dB.

In Chapter 5, the overall permanent monitoring system was also tested for reliability. The transducers monitored over 6 months did not show any major degradation, although it showed a gradual SNR drop over time with slight fluctuations due to changing environmental conditions at the point of measurement. The prototype is thus suitable for permanent monitoring, but the degradation of transducers should be monitored yearly to ensure that the SNR remains above the 6dB cutoff (where signal is double the noise). The deconvolution process was also shown to be a stable process over the months of measurement. The deconvolution method showed 3 main benefits on-site.

- The output SNR of the deconvoluted toneburst is approximately 7dB higher than the output SNR of the toneburst response.
- The voltage of excitation for the chirp used initially was 70V. The voltage of the chirp was eventually reduced to 30V with both the 70V and 30V deconvolution achieving an SNR of approximately 21.3dB.
- This increase in SNR through deconvolution extends the life of the prototype against transducer degradation, measured by the gradual drop in SNR for each passing month of measurement. The limit for the prototype life is arbitrarily set

to an output SNR of 6dB. When the output SNR is lower than 6dB, the prototype has to be put through maintenance to repair the wear in the matching layer etc.

Finally, an example of a risk assessment for use of GWT with an industry guideline, DNV-RP-F101 is presented. The first step involved estimating the uncertainty of the method of inspection and then applying that uncertainty in the guideline. The defects selected in the inversion are the defects with the deepest points and the probability of failure for the coming months are estimated through normal distribution statistics. While the uncertainty increases in the estimation of failure for future months, it highlights one of the benefits of permanent monitoring systems that is keeping uncertainty in future predictions low. The study shows how GWT can be incorporated in industry, and add value to asset owners looking into permanent monitoring systems.

6.2 Future Work

The final risk assessment using the guideline can be further developed by incorporating imaging inputs that investigate all the defects in the area of coverage. In Chapter 5, the axial and circumferential projections are obtained for the region of interest corresponding to the deepest defect. However, an algorithm can be eventually developed to take in the entire reconstruction as an input and analyse all interacting defects within the imaging area to calculating the allowable operating pressure for the coming months.

Figure 6.1: The reconstruction thickness can be meshed into a matrice. probability of failure for each mesh point can be through the methodology presented in chapter 5. Each month of monitoring will generate a 3-D probability of failure chart that can monitor when the probability exceeds a specified value. For low class pipelines, $P(\text{failure}) < 10^{-3}$. If exceeded, an alarm can be triggered.

A probability map can then be generated as shown in Fig. 5.1. If the probability calculated is lower than the acceptable probability of failure, as defined in Table C-1 in Appendix C, for the particular class of pipeline being inspected, an alarm can trigger to signal maintenance of the pipeline.

A digital twin can be created for the physical entity inspected and can then be updated during each month of monitoring to reduce the uncertainty and recalibrate future predictions of failure.

Additionally, more work should be done to tabulate uncertainties for GWT for different wall thicknesses and frequency, including higher operating frequencies and on different

carbon steel grades according to API guidelines to reduce conservativity in estimating uncertainty. Large uncertainties tend to increase the probability of failure unnecessarily which is undesired for owners when trying to implement maintenance and visual inspection schedules.

Improvement of transducer bandwidths for low frequency A_0 transducers should also be studied to allow wider chirp signals to be used. A wider chirp would allow a shorter toneburst pulse to be obtained in deconvolution. This is due to the requirement in deconvolution for the bandwidth of the toneburst excitation to be within the bandwidth of the chirp excitation in obtaining the transfer function. These wide bandwidths are also desirable in making these transducers versatile in being implemented in a range of wall thicknesses for 0.5MHz.mm operating frequencies in GWT and for alternative applications that rely solely on obtaining TOF through the cross-correlation method.

Finally, in terms of the deconvolution method, different passband filters can be designed instead of the conventional rectangular filter used here. For example, a transition area can be incorporated between the passband and stopband. Alternatively, a Bessel or Butterworth filter can be used and the effects of the different filters on the improvement in SNR through deconvolution should be investigated. It is noted that any filter used represents a finite passband that form ringing artifacts in the time domain. However, a smoother filter might possibly reduce the magnitude of these artifacts and increase the maximum output SNR that can be achieved through deconvolution.

A. Strain Charge Equation Matrices

The following matrices for the strain-charge equations are readily available in any engineering textbook but are also presented in IEEE standard. (1987) The piezoelectric matrix, $\mathbf{d}_{jk}^c$, can be represented as follows,

$$\mathbf{d} = \begin{bmatrix} d_{11} & d_{12} & d_{13} & d_{14} & d_{15} & d_{16} \\ d_{21} & d_{22} & d_{23} & d_{24} & d_{25} & d_{26} \\ d_{31} & d_{32} & d_{33} & d_{34} & d_{35} & d_{36} \end{bmatrix}^T \quad (\text{A.1}),$$

where d_{31} , d_{32} and d_{33} relate to the normal strain in the 1,2 and 3 directions respectively.

Here, d_{33} relate to the strain in the direction of polarization.

The compliance matrix is of the form

$$\mathbf{S}^E = \begin{bmatrix} S_{11} & S_{12} & S_{13} & S_{14} & S_{15} & S_{16} \\ S_{21} & S_{22} & S_{23} & S_{24} & S_{25} & S_{26} \\ S_{31} & S_{32} & S_{33} & S_{34} & S_{35} & S_{36} \\ S_{41} & S_{42} & S_{43} & S_{44} & S_{45} & S_{46} \\ S_{51} & S_{52} & S_{53} & S_{54} & S_{55} & S_{56} \\ S_{61} & S_{62} & S_{63} & S_{64} & S_{65} & S_{66} \end{bmatrix} \quad (\text{A.2})$$

The permittivity matrix is

$$\mathbf{e}^\sigma = \begin{bmatrix} e_{11}^\sigma & 0 & 0 \\ 0 & e_{22}^\sigma & 0 \\ 0 & 0 & e_{33}^\sigma \end{bmatrix} \quad (\text{A.3})$$

The stress vector is written as

$$\boldsymbol{\sigma} = \begin{bmatrix} \sigma_1 \\ \sigma_2 \\ \sigma_3 \\ \sigma_4 \\ \sigma_5 \\ \sigma_6 \end{bmatrix} \quad (\text{A.4})$$

B. Parameter trends for SNR gain using chirp

Increasing the toneburst length and bandwidth of chirp used reduces the SNR (Figure B.1 a & b) while increasing the number of times the chirp is longer than the toneburst improves SNR (Figure B.1 c). However, for a data acquisition of 1000 μs , based on 42.9 μs toneburst, the value of k should not be more than 10 because we lose out information on higher order helical modes, which poses an issue in the limited view problem in GWT.

From Eqn. (5.14),

$$\text{SNR}_g = 20 \log \left(\frac{S_c(\omega)}{S_{tb}(\omega)} \right) = 20 \log \left(1.275 \sqrt{\frac{k}{2T_{tb}B}} \right)$$

Parameter trends for SNR gain using chirp

Figure B.1 a) SNR gain for different toneburst lengths, Ttb (s). b) SNR gain for different bandwidth of chirp used in Hz. C) SNR gain for different values of k , the ratio between the length of the chirp to toneburst.

C. Tables of safety factors and fractile values

Table C-1

Safety Class and target annual failure probability for Ultimate Limit State (ULS)	
Safety Class	Indicating a target annual failure probability of:
High	$< 10^{-5}$
Normal	$< 10^{-4}$
Low	$< 10^{-3}$

Table C-2

Standard Deviation and Confidence Level		
Relative Sizing Accuracy	Confidence level	
	80%	90%
± 0.0 of t	StD[d/t] = 0.0	StD[d/t] = 0.0
± 0.05 of t	StD[d/t] = 0.04	StD[d/t] = 0.03
± 0.1 of t	StD[d/t] = 0.08	StD[d/t] = 0.06
± 0.2 of t	StD[d/t] = 0.16	StD[d/t] = 0.12

Table C-3

Partial Safety Factor and Fractile value				
Inspection sizing accuracy, StD[d/t]	ε_d	Safety Class		
		Low	Normal	High
(exact) 0.00	0.0	$\gamma_d=1.00$	$\gamma_d=1.00$	$\gamma_d=1.00$
0.04	0.0	$\gamma_d=1.16$	$\gamma_d=1.16$	$\gamma_d=1.16$
0.08	1.0	$\gamma_d=1.20$	$\gamma_d=1.28$	$\gamma_d=1.32$
0.16	2.0	$\gamma_d=1.20$	$\gamma_d=1.38$	$\gamma_d=1.58$

Table C-4

Partial Safety Factor γ_m for pipelines			
Inspection method	Safety Class		
	Low	Normal	High
Relative	$\gamma_m=0.82$	$\gamma_m=0.77$	$\gamma_m=0.73$
Absolute	$\gamma_m=0.85$	$\gamma_m=0.80$	$\gamma_m=0.75$

Table C-5

Partial Safety Factor γ_{mc} and η		
Safety Class	Factor γ_{mc}	η
Low	$\gamma_{mc}=0.85$	$\eta=1.00$
Normal	$\gamma_{mc}=0.80$	$\eta=0.90$
High	$\gamma_{mc}=0.75$	$\eta=0.80$

Table C-6

Usage factors, ξ	
Safety Class	Usage Factor, ξ
Low	$\xi=0.90$
Normal	$\xi=0.85$
High	$\xi=0.80$

D. Reconstruction and analysis of defects

The following figures presented are thickness reconstructions using FWI for toneburst signals. The transducer arrays are placed 400 mm apart.

D.1 December Measurements - 4 single axial defects, 1 interacting (1-2)

Figure D1: a) FWI December toneburst reconstruction of 12 inch water loaded pipe on-site b) circumferential thickness reconstruction c) axial thickness reconstruction

Table D1-December Distance between defects

Axis Type	Circumferential Defects					Longitudinal Defects			
Max Length (mm)	rΦ ₀ =138.2					S ₀ =88			
Int. Defect	A- X	B- X	X -C	X -D	X -E	1- X	2- X	X -3	1-2
Distance between defects (mm)	Φ _{AX} =242.17	Φ _{BX} =169.76	Φ _{XC} =315.07	Φ _{XD} =480.43	Φ _{XE} =536.2	S _{1x} =109.89	S _{2x} =89.98	S _{x4} =100.6	S ₁₂ = 14.03
Axial Length of Defects (mm)									
L ₁ =68.83		L ₂ =9.27		L _x =35.74		L ₃ =80.48			
Defect Depth (mm)									
d ₁ =1.29		d ₂ =0.12		d _x =1.51		d ₃ =1.3			

D.2 January - 5 single axial defects, 2 interacting (4-x/1-2-3)

Figure D2: a) FWI January toneburst reconstruction of 12 inch water loaded pipe on-site b) circumferential thickness reconstruction c) axial thickness reconstruction

Table D.2-January Distance between defects (x-4/1-2-4)

Axis Type	Axial defects					
Max Length (mm)	S ₀ =88					
Int. Defect	1-X	2-X	X-3	4-X	1-2	1-4
Distance between defects (mm)	S _{1x} =114.85	S _{2x} =91.25	S _{x3} =102.39	S _{x4} =45.62	S ₁₂ =17.24	S ₁₄ =45.36
Axial Length of Defects (mm)						
L ₁ =69.23		L ₂ =9.02	L ₄ =25.2	L _x =31.57	L ₃ =79.58	
Defect depth (mm)						
d ₁ =1.51		d ₂ =0.3	d ₄ =0.35	d _x =2.05	d ₃ =1.34	

D.3 February – 6 single axial defects, 3 interacting (4-x-5/1-2-4/3-5)

Figure D3: a) FWI February toneburst reconstruction of 12 inch water loaded pipe on-site b) circumferential thickness reconstruction c) axial thickness reconstruction

Table D.3-February Distance between defects

Axis Type	Axial defects								
Max Length (mm)	S ₀ =88								
Int. Defect	1-X	2-X	X-3	4-X	X-5	1-2	1-4	3-5	
Distance between defects (mm)	S _{1x} =116.55	S _{2x} =92.98	S _{x3} = 101.46	S _{x4} =46.89	S _{x5} =59.87	S ₁₂ =13.51	S ₁₄ =42.91	S ₃₅ =30.99	
Axial Length of Defects (mm)									
L ₁ =69.9		L ₂ =10.6		L ₄ =28.87		L _x =31.52		L ₅ =15.1	L ₃ =78.94
Defect depth (mm)									
d ₁ =1.50		d ₂ =0.46		d ₄ =0.39		d _x =2.28		d ₅ =0.14	d ₃ =1.32

D.4 March- 7 axial defects (3-5-6/4-x-5-6/1-2-4)

Figure D4: a) FWI March toneburst reconstruction of 12 inch water loaded pipe on-site b) circumferential thickness reconstruction c) axial thickness reconstruction

Table D.4-March Distance between defects

Axis Type	Axial defects						
Max Length (mm)	S ₀ =88						
Int. Defect	4-X	X-5	X-6	1-2	1-4	3-5	3-6
Distance between defects (mm)	S _{x4} =47.06	S _{x5} =58.69	S _{x6} =25.38	S ₁₂ =11.1	S ₁₄ =38.6	S ₃₅ =28.02	S ₃₆ =72.17
Axial Length of Defects (mm)							
L ₁ =70.59	L ₂ =11.9	L ₄ =30.14	L _x =28.02	L ₅ =17.7	L ₆ =7.93	L ₃ =79.04	
Defect depth (mm)							
d ₁ =1.49	d ₂ =0.52	d ₄ =0.45	d _x =2.44	d ₅ =0.27	d ₆ =0.12	d ₃ =1.35	

D.5 April – 7 single axial defects (3-5-6/4-x-5-6/1-2-4)

*Figure D5: a) FWI April toneburst reconstruction of 12 inch water loaded pipe on-site
b) circumferential thickness reconstruction c) axial thickness reconstruction*

Table D.5-April Distance between defects

Axis Type	Axial defects						
Max Length (mm)	S ₀ =88						
Int. Defect	4-X	X-5	X-6	1-2	1-4	3-5	3-6
Distance between defects (mm)	S _{x4} =47.24	S _{x5} =56.8	S _{x6} =23.62	S ₁₂ =11.94	S ₁₄ =37.69	S ₃₅ =28.66	S ₃₆ =71.93
Axial Length of Defects (mm)							
L ₁ =71.4	L ₂ =11.15	L ₄ =31.59	L _x =28.67	L ₅ =18.58	L ₆ =8.49	L ₃ =78.03	
Defect depth (mm)							
d ₁ =1.98	d ₂ =0.75	d ₄ =0.64	d _x =3.3	d ₅ =0.43	d ₆ =0.22	d ₃ =1.74	

E. Compressive Axial Stresses on Pipe

The material properties of the pipeline are presented in the table below.

Table E.1 Material Properties

Material properties used in bending moment calculations	
Pipe wall thickness, t (mm)	6.35
Pipe outer diameter, D_{outer} (mm)	323.85
Density of water (kg/mm^3)	1.5E-6
Density of pipe (kg/mm^3)	7.85E-6
Acceleration due to gravity (m/s^2)	9.81
Specified Minimum Yield Strength, SMYS (N/mm^2)	245
Specified Minimum Tensile Strength, SMTS (N/mm^2)	415
Internal Operating pressure (N/mm^2)	1
Weight of Prototype (N)	235.44
Length of pipe span between 2 supports, L (mm)	9000
Weight of loaded pipe per unit length (N/mm)	1.6067
Poisson's Ratio, ν	0.303

The assumptions made in bending moment stress calculations for the carbon steel pipeline used in chapter 5 are as follows.

1. The material is isotropic and homogeneous.
2. Plane sections remain plane in bending.
3. The stresses do not exceed the elastic limit of the material
4. The weight of the prototype is negligible compared to the entire loading of the pipeline. (weight of the prototype is only 1.63% of the segment of loaded pipeline)
5. The beam is initially straight
6. The maximum moment is used to calculate stresses in the pipeline guideline as a conservative approach even though the location of the defect and the location of maximum bending stress are essentially different.
7. The pipe is fully loaded with water.
8. There are no externally applied forces ($F_x=0$) Thermal stresses are neglected under the assumption that the installation temperature is similar to the operational temperature and that thermal stresses are only significant in welded supports, which is not the case here.

Figure E.1: Loading on pipeline. The hoop stress and axial stress due to internal pressures are ignored due to the negligible values compared to the axial compressive stress due to bending.

For a cylindrical thin walled pipe ($t < \frac{D}{10}$), the hoop stress and axial stress due to internal pressure are shown below.

$$\sigma_{\text{hoop}} = \frac{Pr_{\text{ID}}}{t} \quad (\text{E.1})$$

$$\sigma_{\text{axial,int pres}} = \frac{Pr_{\text{ID}}}{2t} \quad (\text{E.2})$$

The effects of Eqn. (E.2) however are only present in the cylindrical case for end caps but the standing assumption is that bends in pipelines, which act as endcaps are commonplace and result in tensile longitudinal stress. The Poisson effect from hoop stress is also present and results in an accompanying compressive displacement along the longitudinal axis. The pipe however, does not physically contract, resulting in tensile stress.

For an operating pressure of 10bar or 1N/mm², the value of the hoop stress, σ_{hoop} , is simply the ratio of the inner radius of the pipe with the pipe wall thickness.

The poisson effects should be superimposed considering that the longitudinal stress due to bending already causes a tensile stress in the longitudinal axis. For ease, we define the longitudinal axis of the pipe as x and the circumferential axis as y. The strain in the x direction can then be determined from the equation as follows.

$$\epsilon_x = \frac{1}{E} (\sigma_{bending} - \nu \sigma_{hoop}) \quad (E.3)$$

The convention however is important in that the expansion in the hoop direction results in a tensile stress in the longitudinal axis. The top pipe experiences compression which results in an decrease in longitudinal compressive stress. The bottom of the pipe undergoes tensile stress. The hoop stress adds to the cumulative stress on the bottom of the pipe. Equation E.4 and E.5 below depicts the respective compressive and tensile stresses in the pipe.

$$\sigma_{longitudinal, compressive} = \sigma_{bending} - \nu \sigma_{hoop} - \sigma_{axial, int pres} \quad (E.4)$$

$$\sigma_{longitudinal, tension} = \sigma_{bending} + \nu \sigma_{hoop} + \sigma_{axial, int pres} \quad (E.5)$$

From Fig. E.1, based on the loading, the bending moment diagram (BMD) and Shear force diagrams (SFD) are presented below in Fig. E.2.

Figure E.2: Bending Moment Diagram (BMD) and Shear Force diagram (SFD) of the pipeline.

From the figure above, the maximum stress due to bending occurs at the midpoint of the pipeline. The stresses due to bending can be calculated from the equation below or approximated by the equation in the DNV guideline written here for convenience but also presented in chapter 5 as Eqn. (5.9).

$$\sigma_{\text{bending, DNV}} = \frac{4M}{\pi(D_{\text{outer}} - t)^2 t} \quad (\text{E.6})$$

$$\sigma_{\text{bending, theory}} = \frac{Mc}{I} \quad (\text{E.7}),$$

where

$$I_y = \frac{\pi}{4} (r_{\text{OD}}^4 - r_{\text{ID}}^4)$$

$$c = r_{\text{OD}}$$

The results of Eqn. (E.6) and (E.7) are shown in Fig. E.3 below.

Figure E.3: Axial stress due to bending based on the DNV guideline and bending stress theory.

The theory is shown to be more conservative because it takes the outer diameter of the pipe whereas the DNV guideline uses the mean diameter. Nevertheless, the conservative value from the theory is used.

From Eqn. (E.4) and (E.5), the value of the cumulative longitudinal stress on the pipe is then,

$$\sigma_{\text{cumulative, compressive}} = \sigma_{\text{bending}} - \nu \sigma_{\text{hoop}} - \sigma_{\text{axial, int pres}} = 13.31 \text{ N/mm}^2$$

$$\sigma_{\text{cumulative, tensile}} = \sigma_{\text{bending}} + \nu \sigma_{\text{hoop}} + \sigma_{\text{axial, int pres}} = 40.14 \text{ N/mm}^2$$

The guideline in DNV follows the conservative approach of compressive longitudinal stress and neglects effects from internal pressure. Compressive stress is taken over tensile stress because the former reduces burst capacity. The compressive stress, σ , only reflects the effects of bending. In reality, this stress is much lower as is shown above and both the calculated compressive stress, $\sigma_{\text{cumulative, compressive}}$ and σ_{bending} will be used to assess the allowable operating pressure under corrosion, P_{corr} , for comparison.

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