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Ultrasonic Digital Communication System for a Steel Wall Multipath Channel: Methods and Results

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ULTRASONIC DIGITAL COMMUNICATION SYSTEM FOR A STEEL WALL MULTIPATH CHANNEL: METHODS AND RESULTS

by
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Approved:

A handwritten signature in dark ink, appearing to read "Gary J. Saulnier", is written over a horizontal line.

Professor Gary J. Saulnier, Thesis Adviser

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ABSTRACT

As of the development of this thesis, no commercially available products have been identified for the digital communication of instrumented data across a thick (~ 6 in.) steel wall using ultrasound. The specific goal of the current research is to investigate the application of methods for digital communication of instrumented data (i.e., temperature, voltage, etc.) across the wall of a steel pressure vessel. The acoustic transmission of data using ultrasonic transducers prevents the need to breach the wall of such a pressure vessel which could ultimately affect its safety or lifespan, or void the homogeneity of an experiment under test. Actual digital communication paradigms are introduced and implemented for the successful dissemination of data across such a wall utilizing solely an acoustic ultrasonic link. The first, dubbed the “single-hop” configuration, can communicate bursts of digital data one-way across the wall using the Differential Binary Phase-Shift Keying (DBPSK) modulation technique as fast as 500 bps. The second, dubbed the “double-hop” configuration, transmits a carrier into the vessel, modulates it, and retransmits it externally. Using a pulsed carrier with Pulse Amplitude Modulation (PAM), this technique can communicate digital data as fast as 500 bps. Using a CW carrier, Least Mean-Squared (LMS) adaptive interference suppression, and DBPSK, this method can communicate data as fast as 5 kbps. A third technique, dubbed the “reflected-power” configuration, communicates digital data by modulating a pulsed carrier by varying the acoustic impedance at the internal transducer-wall interface. The paradigms of the latter two configurations are believed to be unique. All modulation methods are based on the premise that the wall cannot be breached in any way and can therefore be viably implemented with power delivered wirelessly through the acoustic channel using ultrasound. Methods, results, and considerations for future research are discussed herein.

1. Introduction

Presently, there are no identified commercially available means—apart from communication lines fed through drilled holes—for communicating instrumentation data into and out of thick-wall (~ 6 in.) steel pressure vessels. This poses a significant problem, as breaching the wall in any way can reduce the life and safety of the vessel significantly or cause pressure leaks. Modifying the vessel to allow the through-wall passage of data lines may also be unattractive in certain instances where maintaining experiment pressure and homogeneity are of utmost concern. Such issues seriously reduce, if not prevent, the ability to conduct any type of instrumentation within the vessel. Therefore, there is a real need to implement a non-destructive instrumentation system that is capable of measuring parameters in an environment that is high pressure, high temperature, or toxic (an environment with such characteristics is herein referred to as *volatile*) and communicating them through a steel wall.

In an effort to circumvent this limitation, the goal of this project is to develop a method of data transmission across a thick-wall vessel using ultrasonic transducers. In such a fashion, an acoustic communications link is attempted where hardwire or radio frequency (RF) links are not feasible. There are also several environmental considerations which dictate the methods used for the final implementation. Specifically, the volatile environment within the vessel poses a threat to any type of sensitive communication devices placed within. For this reason, signals originating from within the vessel which may require the use of fixed oscillators, for example, are unachievable. The presence of an erratic and non-ideal measurement environment calls for a minimal solution to an advanced communication problem within the vessel. We ultimately require a robust measurement and communication link that is reliable, accurate, and easily implemented.

As a result of these stipulations, primary investigations within the project include communication schemes where carrier signals originate wholly external to the pressure vessel. The two carrier modes utilized in this project are Pulse Modulation (PM) and Continuous Wave (CW) Modulation. Each mode is used because of the high oscillatory nature of the ultrasonic transducers which are tuned for a specific center frequency. The

practical application of the ultrasonic communication scheme requires the generation of these signals external to the vessel. In PM mode, large power sources are necessary to ensure desirable peakedness and duration of the emitted signal. In addition, in CW mode stable oscillators are necessary to maintain the tuned center frequency of each transducer. Clearly large voltage sources are unavailable within the volatile vessel environment without the use of fed-through power. Signal oscillators—which are notoriously unstable at high temperatures—cannot viably exist within this environment either. Therefore, moving the generation of said carrier signals to the stable and controlled environment external to the vessel is not only attractive but necessary. Consequently, modulation of the carrier signal occurs internal to the vessel to thereafter convey the measured data externally. The amount of power necessary to modulate the signal appropriately is minimal and can therefore ultimately be transmitted wirelessly using acoustic power delivery.

The development of this project evolved through two major research phases, both of which were aimed at demonstrating the potential of a standalone ultrasonic communication system for pressure vessel use. The first phase of research revolved around developing a primitive proof of concept system to demonstrate the achievability of an ultrasonic communication link and to justify further research in the area. It also facilitated the development of core communication processing concepts to be utilized in later, and potentially more sophisticated instantiations. This bottom-up approach focused on developing an ultrasonic communication link using one pair of coaxially mounted transducers affixed to a steel block of rectangular dimensions. In this implementation—herein dubbed the “single-hop” configuration—a CW burst signal is both synthesized and modulated on the fly using Differential Binary Phase-Shift Keying (DBPSK). Instrumented data is then transmitted by one UT and received by another to be demodulated and decoded for display to the user.

The second major phase of research sought to more accurately emulate the true vessel configuration in the communication paradigms implemented. More specifically, the obvious downside to the single-hop configuration is that it models a configuration where carrier signals are wholly synthesized and modulated internal to the vessel. As discussed earlier, origination of the oscillator signal within the vessel is not an attractive

means for the true vessel implementation due to the volatile environment present. It is, however, a necessary scheme to demonstrate fundamental achievability of the concept. Therefore, in the latter phase of research developed paradigms focused on locating the carrier generation external to the vessel. In the “double-hop” configuration, two sets of coaxially mated transducers are mounted such that the first set transmits the carrier into the vessel and receives it, while the other retransmits the modulated carrier externally. Based on the deterministic modulation of the signal received by the second UT in the path and subsequently fed to the third UT (both internal to the vessel), instrumented data is conveyed across the wall and reconstructed outside of the vessel.

The research into this area studied both the use of On-Off Keyed (OOK) Pulse Amplitude Modulation (PAM)—where a pulse train is gated to allow individual pulses to pass. The research also investigated the use of DBPSK previously utilized in the single-hop configuration. While both methods were successful, the use of least-mean squared (LMS) adaptive noise suppression increased the data bit rate (DBR) ten times from approximately 500 bits per second (bps) to 5 kbps when using DBPSK in the double-hop configuration. While the double-hop configuration boasts the features of increased data transmission rates (with the use of adaptive suppression) and a CW carrier format which may be easier for harnessing wireless power, its drawbacks include an increased number of transducers, more complex signal processing techniques, and increased computational overhead.

Many of the advanced modulation techniques are readily implemented in the demonstration systems by way of a personal computer and data acquisition (DAQ) system. While both convenient and appropriate for initial research, the end-goal system will ultimately not allow the luxury of this equipment internal to the vessel. Consequently, the overriding prospect is to move towards a paradigm where the complexity of the electronics inside of the vessel is minimized. Doing so not only minimizes the cost and logical control necessary, but also expands the number of components that may be available for the required temperature and volatility of the internal vessel environment. Moreover, an additional requirement is to minimize the number of ultrasonic transducers (UT) necessary to run an ultrasonically-communicated, data instrumentation system within the desired constraints. Again, the current research focuses on the demonstrable

achievability of the techniques more so than application in the true volatile vessel environment. This area is left for future research.

In an effort to reduce the number of transducers utilized in the system to the minimum necessary, experimentation with a hybrid single-hop (reflected-power) setup is made. To reiterate, the problem with the original single-hop setup is that it requires signal *origination* from within the vessel rather than simply modulating a carrier injected from external to the vessel. In order to circumvent this limitation a communication scheme which stems from an empirical result is implemented. In the “reflected-power” configuration, a single set of coaxially mounted transducers is attached to the vessel as they are in the single-hop configuration. The major difference is that in the reflected-power configuration, the carrier signal is transmitted into the vessel and the receiving transducer’s terminals are shorted to provide modulation. Doing so effectively modifies the acoustic impedance at the interface and causes a measurable amount of the signal to be reflected back to the source. This result is exploited to produce a digital modulation scheme.

In the reflected-power mode, the external ultrasonic transducer is excited with a repeating analog pulse at a fixed repetition rate. At the instant the emitted pulse is arriving at the mate UT on the opposite side of the wall (internal to the vessel), the receiving transducer’s terminals are either left open or are shorted together. The result is an appreciable change in the signal thereafter received at the originating transmit UT external to the vessel. The reflected signal produces measurable, constructive interference within subsequent echo pulses. Mapping the deterministic change between the shorted and un-shortened states to 0’s and 1’s readily facilitates a binary modulation scheme. The obvious benefit of this system is the reduction in the number of transducers necessary from four to two. Moreover, it eliminates the need to modulate the sinusoidal carrier in the vessel (to create the $0^\circ / 180^\circ$ phase shift or ‘0’ / ‘1’ transitions) which requires the use of a mixer or other similar device. Instead, a simple diode switching circuit is utilized to short the receiving UT’s terminals and modify the acoustic impedance at the interface. The drawback with this method is in the limitations of trying to harness enough usable power delivered ultrasonically from a pulse train to supply the internal electronics necessary.

Each demonstration system for the aforementioned configurations was first developed on a fixed, rectangular steel block to ensure repeatability of the experiment. Thereafter, the concept was attempted on an actual decommissioned pressure vessel to mimic the practical application of the techniques. With the assistance of the DAQ system as the development suite, rapid reconfiguration is achievable which precipitated the notable success of the concept. The demonstration systems developed are able to transmit and receive a known set of test data (e.g., a measured voltage) using solely an acoustic link through a solid steel wall. The received data is then appropriately reconstructed to display the originally measured data to the user. Given the evident success of the concepts attempted, it is clear that a fully developed communications system with dedicated electronic devices for the purpose of ultrasonic data communication is achievable. Moreover, the straightforward implementation of each technique paves way for the acoustic delivery of power levels suitable to enable signal modulation as necessary. A detailed discussion of the methods and results for this research—carried out in two major phases—is the subject of this thesis and is subsequently discussed. Generally speaking, the discussion follows the evolution of the concept over the course of the project.

2. Historical Review

After exhaustive product searches it is clear that an ultrasonic communications system for the aforementioned specific purpose is not in use; if it is, it is likely in limited esoteric deployment. The lack of commercial availability justified the investigative research and development within the area. The concept of ultrasonic communication is by no means novel, but the specific medium and signaling paradigms that have been implemented are felt to make the current research unique and fundamental in its own rite. As a result, there may be a subtle overlap of *concept* in the prior art but much less so in the *application* and method for use with steel pressure vessels. For example, Binary Phase-Shift Keying (BPSK) has been in widespread use for more than three decades for digital communication over analog channels in some computer modems [1]. Least-mean squared adaptive suppression is also a fundamental tenet in the equalization of stochastic channels [2]. However, the double-hop configuration—which uses these common methods as facilitative tools—has, as of the writing of this thesis, remained undiscovered in the literature, public documentation, and patent searches and is therefore believed to be unique. The same could be claimed about the reflected-power configuration as described herein.

Application	Average Range
Upper limit of human hearing	16 kHz
Defoaming and degassing	2-30 kHz
Ultrasonic metal working and welding	16-25 kHz
Control applications	16-45 kHz
Ultrasonic cleaning	20-40 kHz
Nondestructive testing (NDT)	1-10 MHz

Table 1 - Ultrasonic frequency ranges and typical applications.

As Table 1 suggests, ultrasonic applications have had widespread implementation within industry for some time [3]. Specifically, acoustic communication was first introduced for control installations in the mid-twentieth century. The first primitive acoustic remote controller, the Space Command controller, was introduced by Zenith in

the 1950's. The utilization of acoustic electronics perhaps first met critical necessity in World War II, when systems were developed to facilitate underwater object detection, navigation, and communication. Similar to the current installation for the pressure vessel, the necessity for said acoustic systems stemmed from the fact that water is a poor medium for wireless modes such as radio frequency (RF) or optical communication. When properly coupled to the medium and processed, an acoustic link can be made as reliable as a hard-wired installation. Later developments in acoustic communication around the same time led to computer data modems which were acoustically coupled to a telephone and achieved data rates around 300 bits per second. These modems were subsequently replaced by the direct-couple data modems which we are so familiar with today [4]. Many of the aforementioned applications progressed toward ultrasonic ranges, where interference from other ubiquitous sources is less likely. In addition, due to the intrinsic material properties of many media some ultrasonic frequencies propagate more readily than those in the audible range.

Communicating data through analog channels using transmitting and receiving ultrasonic transducers has also been in use for some time. For example, in one published patent a claim is made to the use of an ultrasonic link to facilitate communication from telephone handsets within the cabin of an automobile to a radio frequency transmitter mounted elsewhere in the vehicle [5]. Further claim is established wherein full-duplex communication can be accomplished by utilizing a multiplexer and ultrasonic communication at several unique frequencies. Another patent claim along the same lineage includes the use of an omnidirectional ultrasonic transducer to facilitate wireless communication between a telephone handset and an earpiece [6]. Additionally, many inventions have used gating functions to vary the pulsewidth of the transmitted ultrasonic signals for translating binary signals to analog and vice versa (e.g., [7] and [8]). In [8], the data-carrying capacity of each received signal is augmented through the use of varying frequencies, pulsewidths, and duty cycles which have been predetermined. The receiver also seeks to minimize the effects of multipath interference by disabling the receive function immediately following the receipt of the dominant line-of-sight (LOS) signal.

In the actual implementation of a similar device, Holm, et al. developed a system for airborne ultrasonic communication using user tags and detectors with Ethernet attachments [9]. Their system achieved a functional range of 10-20 meters and found the channel bandwidth to be slightly less than that utilized for human speech. One patented method communicates bidirectionally across a coupling medium using in a system with embedded controllers and actuators [10]. In a similar fashion, this system can deliver power from an external source acoustically across a coupling medium without the connection of wires [11]. Similarly, a wide range of commercially available devices for underwater diver-to-diver communication utilize an acoustic link in the ultrasonic frequency range. For example, Ocean Technology Systems' MKII-BUD (Model # 900264-015) & MKII-BUDS (Model # 900264-016) Buddy Phone integrated mask system uses a 32.768 kHz upper single sideband (USB) ultrasonic carrier and has an audio bandwidth of 2.7 kHz [12]. Some may question the safety of ultrasonic signals in close proximity to life. Incidentally, Gerasimov and Bender reported that ultrasonic communication is safe for humans if intermittent, low-intensity, and in the range of 20-500 kHz. Above this frequency range, there is no reported threat to the health and safety of life if emissions are of reasonable intensity [13].

Some inventions utilize multiple ultrasonic transducers centered at different frequencies to convey a plurality of data. For example, in [14] a method is introduced whereby an array of ultrasonic transducers centered at several unique frequencies are utilized to transmit varying, predetermined serial streams of permuted frequency bursts. Rather than the arrangement of bits being transmitted with a single transducer, an arrangement of bits—which are mapped to one of many frequencies—is transmitted. In this fashion, it is claimed that signals are less prone to interference from reverberation within the medium. In addition, as a result of the increase from simple binary to higher order (i.e., M-ary) bits the data-carrying capacity of each word is increased and hence, data transmission time is decreased. Homogeneous ultrasonic transducer arrays are also widespread in their use for many reasons; mainly to aid in beam focusing in a large field. For instance, Webb and Wykes developed an enhanced ultrasonic transducer array for beam forming in airborne ultrasound [15]. The two also introduced a technique that would allow the separation between transducers to be less than the traditionally accepted

limit of one-half the wavelength of the signals in use (i.e., higher field resolution). The general technique has been vastly developed and is generally well-documented in the literature (e.g., see [16], [17], [18], or [19]). This technique may be useful for augmenting bandwidth and data transmission speed or in focusing emissions for power delivery in future investigations of the current research.

Prior work also deals with the transmission of ultrasonic signals into a network of inter-digital transducers (IDT) on a piezoelectric substrate using coded burst pulses with surface acoustic waves (SAW). For example, one invention deals with the identification coding of such devices to facilitate a scheme such as the familiar code-division multiple access (CDMA) regime [20]. The inventor of the aforementioned work has another issued patent dealing with the coding modem for such a network, whereby a baseband communication signal can be kept secret [21]. In one paper by the inventor published in the literature, a phase-shift keying (PSK) method was demonstrated using a configuration utilizing three IDT SAW devices: two for differential inputs and one for their corresponding peak output [22]. In a later publication by the same author, an ultrasonic modem using the Binary Phase Shift Keying (BPSK) technique was introduced for use with Lamb acoustic wave IDT devices [23]. In the latter, the Lamb wave device takes as its input the carrier and the binary message signal. An analog switch with a 50% duty cycle provides a one-bit phase delay between the IDT devices at 0° and 180° relative phase transitions to be sent from the output.

The technique of acoustic communication may arguably be most developed—when viewed in the light of publication within the literature—for underwater channels. For instance, many developments have military applications in the command and control of aquatic weapons. One such application elicits the use of ultrasonic transducers to convey control information in serial data format for prelaunch weaponry control [24]. Many specialized transducer applications have been developed for the underwater channel. For example, Siwapornsathain and Lal presented a micromachined piezoelectric transducer (PZT) exclusively for battery-powered underwater communications [25]. In the underwater communications channel, carrier frequencies for signals are often in the audible range, but may very well be ultrasonic. Furthermore, many researchers have developed working models of the underwater communications channel (e.g., as multi-

path Rician channels). As a result of the multipath phenomena and non-stationarity of the channel, most practically applied systems for underwater acoustic communications involve channel equalization in some form. Stojanovic surveyed contemporary advances in underwater acoustics and noted that with the use of coherent detection and channel equalization, reliable communication is achievable over ranges of several miles [26].

A great deal of research has been attributed to the development of a model for the underwater channel (e.g., [27], [28]) which is unstationary and, not unlike the steel pressure vessel, suffers from multipath interference. Adaptive techniques are most often used when multipath, Doppler shift, and signal-to-noise ratio (SNR) parameters reach unacceptable levels. If an acoustic channel is highly unstationary, phase incoherent modes—such as FSK—are most attractive. If phase coherent communication is achievable in a stationary or near-stationary channel, adaptive equalization typically provides the best SNR results and consequently, better data rates [29]. Innumerable simulations and exhibitions within the literature have verified this claim (e.g., [30], [31], & [32]). Yeo, et al. presented an improved Recursive Least Squares (RLS) algorithm for unstationary underwater acoustic channels [33]. Their improved RLS algorithm exhibited demonstrably faster convergence in the time-variant channel when compared to the standard RLS and LMS algorithms.

Clearly, if the signal is at all determinable, maximizing the received SNR brings with it the benefits of increased data capacity and transmission speeds through the channel. To this end, the literature is filled with the development of techniques and methodologies to increase signal efficacy in the time-variant multipath channel. For instance, Huang, et al. developed a novel hybrid Multiple Frequency-Shift Keying – Differential Phase-Shift Keying (MFSK-DPSK) scheme for long range underwater communication [34]. Their work demonstrated an achieved increase in data rate while in the same distance range or an increase in range while maintaining the same data rate when compared to a single, fixed-frequency carrier. The results of their experimentation show that the method results in decreased intersymbol interference (ISI) and increased communication reliability. Some techniques are empirically derived or computationally complex in nature, but most unequivocally point to the aforementioned method of channel estimation and equalization. The work of Lee, Son, and Kim is similar to many

other's and is one such more advanced technique which uses time reversal convolution (i.e., matched filtering) of the channel impulse response with the received signal [35].

Even in the most rudimentary installations—i.e., using a fundamental modulation scheme without source and / or channel encoding—it appears that reliability over great distances in the underwater acoustic channel is realizable using ultrasound. In one communication link from the surface of the ocean to an underwater observatory over 2 kilometers deep, FSK is used for control operations while PSK is utilized for data uplink at rates of 300-5000 bps using the File Transfer Protocol (FTP) [36]. In this configuration the link protocol is time-division multiple access (TDMA), controlled by the unit at the surface. Additionally, van Gijzen and van Walree developed a robust acoustic communication scheme for a shallow-water channel using high bit rate BPSK signals [37]. From review of the literature, it appears that the shallow-water communication channel is the hardest to develop for as the delay spread of multipath signals becomes less due to the approach of the reflective surface and bottom to one another. Consequently, as the delay spread of the multipath interference begins to decrease it becomes harder to “gate and ignore” received interference at higher bit rates. Specifically, at higher bit rates the bit width becomes less and is more greatly affected by the multipath interference on a bit-to-bit basis. The delay spread of multipath interference for the shallow-water channel is typically on the order of 20 milliseconds and perhaps most closely emulates the channel presented by the pressure vessel. As this spread increases, a *differentially*-encoded phase modulation scheme might prove best. Regardless, van Gijzen and van Walree augmented their scheme with the leading transmission of a pseudorandom, source-encoded, learning sequence and least-squares estimation to avoid multipath fading of the BPSK carrier. They achieved data rates up to 4 kbps and, for due reason, anticipated better results with the use of DBPSK encoding.

In one project developed by the National Security Program office of the Department of Energy, a system was developed for digital communication using commercially available ultrasonic transducers at frequencies around 40 kHz [38]. In their implementation, the researchers utilized LabVIEW VI's along with a DAQ system and ancillary signal processing electronics to transmit binary words in a binary FSK mode. An earlier implementation of their research utilized FM carriers as well. The goal of the research

was to investigate the use of ultrasonic signals coupled to pipes of various materials—both empty and filled with water—or to the air to transmit and receive digital messages and images. The initial work of the researchers led to the successful development and patenting of a device for the ultrasonic transmission of speech signals through air [39]. The project ultimately led to a fully functioning system for digital communication using ultrasound. The functioning system was capable of sending monochrome images at 569 pixels per second and ASCII text at 85 characters per second. This equates to a data transmission rate of 680 bps.

One might argue that a pipe is analogous to the pressure vessel in its geometry. A key differentiator in this development from the current research is the direction of data transmission. The current research seeks to develop data transmission diametrically through a thick-wall vessel whereas the aforementioned research developed a system for the circumferential dissemination of signals *along* thin-wall pipes. The fundamental difference between the two is that the current research seeks to facilitate the simplistic, non-destructively implemented data transmission from within a volatile environment where the choice of survivable electronics is highly limited. In addition, while the DOE researchers pay mention to the potential use of Phase-Shift Keying (PSK) in future research, it is absent from their study. Moreover, they utilized only basic analog filters whereas the current research necessitates the use of adaptive digital filtering for interference suppression to mitigate multipath reflections. The importance of the latter was emphasized by Li, Harrold, and Yeung in research they presented in 1997 [40]. In the study, the researchers exhibited the effect of multipath interference through water-filled pipes on a primary link signal. Unsurprisingly, the researchers concluded that as a result of the ISI and fading due to multipath, low rate signaling was necessary to maintain data communication. Specifically, the aforementioned researchers utilized FSK modulation with a transmission rate of 600 bps. Most importantly, they underscored the significance of channel equalization and interference suppression in increasing the transmission rate and reliability; a result which the current research so clearly depicts.

It is evident that a fair amount of signal processing is necessary to sustain reliable communications in the multipath channel regardless of the signaling method used. Left to the effects of these channels, the data-carrying signal is of little utility. As a result,

signal pre-processing methods are of significant importance as well. This determination was emphasized by Zheng, et al. in a work published in 1996 [41]. In their study, the researchers investigated the use of an M-ary Differential Phase Shift Keying (MDPSK) data transmission system using ultrasound in the underwater channel. Due to the highly multipath nature of underwater propagation, the researchers exploited the non-linear characteristics of the channel in a manner that aided in the multipath-minimized signal propagation. The multipath effects arise from signal propagation and echoing from the surface, bottom, and intermediate obstacles of the body of water. Therefore, in a large field a signal emanating from a highly-directional ultrasonic transducer with minimized sidelobes is preferred. The researchers achieve a signal with these characteristics by simultaneously injecting two high intensity carrier signals of differing frequencies into the medium. As a result of the near-field, non-linear interaction of the signals with the water, a signal at the difference frequency is generated whose far-field sidelobes decrease exponentially with distance. The resulting beamwidth is much narrower than that which would have resulted from a transducer centered at a single frequency.

As a product of the aforesaid application, the magnitude of the multipath fading at the receiver is minimized. The resulting beamforming can be visualized as a “virtual array” of transducers and is therefore analogous in effect to the other array methods presented earlier. The researchers implemented the system in an M-ary DPSK paradigm which is robust in the presence of multipath (e.g., Doppler phase) noise. This is generally the case for the differential phase-shift paradigms: bit-transitions are encoded in the *relative* change in phase as opposed to the absolute change [42]. This stems from the assumption that phase noise is uniform, independent, identically distributed, and will consequently affect received phase equally. This effect is overcome by the differential phase encoding. Similar to the majority of the models of the multipath underwater channel which identify the large propagation field and slow oceanic wave conditions as justification, Zheng et al. claim that the channel is slowly time-varying (i.e., approximately stationary) and therefore, the effect on *differential* phase transitions is minimal. Multipath fading is therefore very slow. Thus, any interference which could affect decoding is typically inconsequential when viewed from a differential encode-decode standpoint. For the vessel channel, multipath interference is of short and fixed—that is,

it is a random function but not a random *process*—duration in each realization under identical conditions. As a result, bit transitions become greatly affected as the bit duration decreases. Similarly, due to the fixed duration of the multipath interference, the vessel channel can also be viewed as approximately stationary. Regardless, in the underwater channel Zheng and his team achieved data rates using ultrasound of 1 kbps, 2 kbps, and 3 kbps in binary, 4-ary, and 8-ary DPSK configurations, respectively.

The multipath phenomenon is inherently random and is generally difficult to model with a high degree of accuracy. We can therefore make statistical claims about the anticipated results, but in almost all cases channel postulates are in no way deterministic. The degree to which knowledge of the anticipated channel is well formulated and understood determines how well other schemes such as parameter estimation may work. Some researchers have attempted maximum likelihood (ML) communication schemes, which requires a fair amount of knowledge—such as the joint probability density function (PDF) of the random variables modeling the channel—for satisfactory results [43]. There are certainly many variations on the theme and the robustness of each depends on the nature of the channel and signaling paradigm utilized. [44] demonstrates one such investigation into the efficacy of typical ML algorithms when applied to various mismatched channels. In most implementations, the channel is highly dynamic and therefore some type of adaptive filtering is advantageous. A survey of the literature demonstrates that many researchers enhance their acoustic digital communication systems by implementing adaptive channel equalization as in the work by van Gijzen and van Walree cited earlier.

The usage and understanding of adaptive filter theory is generally well-understood. Adaptive filtering schemes have reached widespread implementation within the communications field and with the current availability of high-speed processing systems they have become a paramount tenet in achieving fast and error-free digital communication over multipath channels. The commonality of the underwater channel for ultrasonic digital communication rears here as well. For example, Stojanovic, Catipovic, and Proakis presented one theoretically-optimal multichannel, joint-equalization scheme that achieved substantially improved results for phase-coherent PSK binary telemetry data in the underwater acoustic channel [45]. The researchers

underscore the importance of adaptive equalization to overcome multipath interference which may spread over several tens of data symbols at higher data rates and is of utmost concern in phase-coherent installations. Their study utilized linear decision feedback equalization (DFE) and a recursive least squares (RLS) algorithm in a multichannel configuration for optimal results. In addition, their configuration was enhanced through the use of a digital phase-locked loop (PLL) which enables phase tracking in rapidly fluctuating multipath interference. Clearly, better performance comes with the tradeoffs of algorithmic complexity and processing time.

While there are several adaptive filtering algorithms that may be of use in the acoustic channel, the optimal selection depends upon the specific characteristics of the channel and computational complexity; the latter especially so in real-time installations. In one installation for a remotely-controlled, underwater vehicle, data communication rates up to 10 kbps were achieved over a range of one kilometer. The system employed synchronization word bursts for use with the LMS algorithm to mitigate the effect of multipath fading in the underwater channel. Improved through the use of Reed-Solomon source coding and matched filters, the discrete FM signals successfully carried digital video via the acoustic link [46]. In [47], researchers showed that for simple channels LMS and RLS algorithms are of comparable efficacy. However, when the channel becomes more unstationary RLS adaptation provides enhanced tracking capability over LMS adaptation which suffers from noise due to tap weight misadjustment. The drawback is the increased computational complexity in the use of RLS over LMS suppression.

Divergence of the LMS algorithm during times of rapid change is always of concern. Therefore, another design consideration revolves around computational speed and complexity verses stability. To combat this, the researchers in the previous citation proposed a method whereby the RLS algorithm is used for the early phase of adaptation and is subsequently switched to the LMS algorithm. The result is increased stability of the adaptation process—as the channel becomes less stationary—and a decrease in computational complexity. In addition, the final level of suppression was fairly comparable to the fundamental LMS algorithm. Unlike the current channel, however, the researchers denote that because of the highly unpredictable nature of the channel a rubric

for each received data word may be necessary to determine the appropriate adaptation algorithm utilized. As a result of the fixed nature of the vessel channel, it is assumed to be stationary between realizations and the LMS algorithm is hence of suitable application since the rapid-tracking capabilities of the RLS algorithm are not justified.

The use of adaptive equalization and interference suppression is extensively cited within the literature and texts. Developments heretofore have revolved around the use of both Wiener and Kalman filtering for both the continuous- and discrete- time cases. With modern advancements in DAQ and digital signal processing (DSP) techniques, most practical implementations of the algorithms are hence in discrete-time and utilize the sample- rather than the time- space. Adaptive filters, in their most direct form, are implemented as operations that take linear combinations of signal observations as their parametric inputs. However, as a result of the adaptive nature of said filters they are intrinsically non-linear; due to their data dependence, they do not follow first principles of superposition. A rubric for measuring the quality of an adaptive filter algorithm can stem from the following key aspects [2]:

1. Rate of Convergence – A measure proportional to the number of algorithm iterations necessary to converge to the closest achievable Wiener solution in the mean-square sense when a stationary input is applied to the filter.
2. Misadjustment – A measure of the deviation of the achieved mean-squared error to the ideal minimum mean-squared error.
3. Tracking – A measure of the filter's ability to follow statistical variations given a non-stationary input.
4. Robustness – A measure of the satisfactory performance of the filter when non-ideal signals are applied as inputs.
5. Computational Requirements – A measure of operations, memory usage, etc. for an implemented algorithm.

6. Structure – A measure of the relative complexity of an algorithm’s organization such as parallelism, modularity, etc.
7. Numerical Properties – A measure of the filter’s robustness in the presence of quantization error and finite word length effects.

As a result of their rapidly reconfigurable nature, adaptive filters are often utilized when filtering in the presence of noise in a non-deterministic channel. Channels may be random, time-varying, unknown at the time of development, or a combination of all three. Furthermore, adaptive filtering is a commonly utilized method for implementation within multipath channels in many realms of communications. Since the propagation paths of multipath interference are intrinsically random and received signals are highly correlated to signal input, adaptive filter algorithms are an attractive method for interference suppression in the presence of multipath noise. The LMS algorithm is notably one of the most popular used for this purpose (e.g., see [48] or [49]). Although effective in its base form, there are several variants and hybrids of the algorithm which yield benefits in terms of the measures of quality listed above (e.g., see [50], [51], or [52]). One such variant is the normalized LMS, or NLMS, algorithm where a faster rate of convergence is achieved with the simple addition of a factor proportional to the power content of the signal under measurement. Channel equalization using adaptive filtering techniques provides the key advantage of increased signal-to-noise ratio (SNR) in the presence of multipath noise.

In conclusion, it is clear that there are some similarities in the concepts introduced by a portion of the discovered prior art compared to the current research. Based upon aspects such as the system configurations and the channel over which data is transmitted, however, there are demonstrable differences. Additionally, an aspect of the current implementation which utilizes adaptive interference suppression to circumvent multipath interference within a pressure vessel, has many unique features namely: the communication of a carrier originating external to the vessel for modulation within it, to subsequently convey instrumented data back externally to the vessel. From the survey

of the literature, overlap of core concepts is apparent and many of which have been introduced as fundamental tenets within areas of acoustic communication. Ultrasonic communication has been in use for some time and has found greatest acceptance in media where other modes are ineffectual (e.g., in water, along pipes, etc.). A few conceptual and practical devices for ultrasonic communication (both analog and digital) encompassing key areas of interest were discovered and introduced. Digital communication over the acoustic channel has also seen widespread implementation in command / control and data telemetry applications. Unsurprisingly, primary communication modulation and data-encoding techniques such as M-ary PSK and FSK find their place in acoustic communication. Differential encode-decode techniques offer increased efficacy where coherent modes are not possible because of variations due to random phase noise. It was discovered that acoustic communication techniques using ultrasound have been vastly developed and published for the underwater channel, where modes such as RF and optical links are of limited usefulness. The vessel channel is analogous to the underwater channel in that it too suffers from the multipath phenomenon albeit more slowly fading due to the lack of Doppler effects in the physically fixed steel wall channel. Finally, it was found that adaptive filtering provides rapidly reconfigurable interference suppression and an increase in the system SNR in the presence of random fading or interference due to multipath noise. Examples of such algorithms were presented for the general communications channel as well as for specific use in the acoustic channel.

3. Channel Characteristics and Approximate Model

To demonstrate the feasibility of a system capable of communicating data across a steel wall using ultrasound, two test appliances are used. Per the project sponsor's specifications, communication across walls of both rectangular and cylindrical geometries is investigated. The rectangular geometry consists of a steel block with transducers permanently affixed in a coaxial fashion on either side of the block. The cylindrical geometry consists of an actual decommissioned pressure vessel. For the vessel, mounting appliances were manufactured to facilitate the temporary affixation and ready adjustment of the ultrasonic transducers in use. Each of the two channels suffers from multipath effects that pose a detriment to trans-wall communication. The general approach in the development of the proof-of-concept systems was to first utilize the rectangular block because of the permanency of the alignment of each transducer and the repeatability of the results. Thereafter, each test system was demonstrated on the vessel as to mimic a possible end-goal implementation.

3.1 Rectangular Test Apparatus

The preliminary investigative work was conducted on a steel rectangular block with dimensions 45.10 x 14.85 x 7.55 cm. This test block was used initially because of its deterministic acoustic paths as well as its portability. In addition, the planar surfaces provide a simplistic method of attachment for each ultrasonic transducer (UT). Prior to use, the block was machined and finished for smoothness. The center locations for each transducer were scribed onto the block to prevent misalignment in the coaxial directions when affixed. Each UT was permanently affixed to one side of the block using a two-part epoxy once alignment was ensured. The same procedure was performed to mount an identical mate transducer directly across from, and coaxially with, each transducer on the opposite face. Figure 1 below depicts the described setup on the rectangular test apparatus. An aerial view can be seen below in Figure 2 as well.

Figure 1 - Image depicting the rectangular test apparatus (left) along with equipment (right).

Figure 2 - Image depicting the rectangular test apparatus (aerial view).

In total, three coaxially-mated sets of transducers are affixed to the rectangular test apparatus:

1. one pair of Krautkramer / GE Inspection Technologies model # 261-043-CR 1.0 inch, 1.0 MHz, Gamma series (medium damping) contact transducers (pictured in the foreground [to the left in] of Figure 1 [Figure 2]),¹
2. one pair of Krautkramer / GE Inspection Technologies model # 260-043-CR 1.0 inch, 0.5 MHz, Gamma series (medium damping) contact transducers (pictured above [to the left of] the 1.0 MHz Gamma transducers in Figure 1 [Figure 2]), and
3. one pair of Panametrics model # V102-RB 1.0 inch, 1.0 MHz, Videoscan series (high damping) contact transducers (not pictured in Figure 1 [Figure 2], however positioned above [to the left of] the 0.5 MHz Gamma transducers).²

All work in the current research utilizes a center frequency of 1.0 MHz because of the availability of transducers and since it was observed that there was less attenuation through the media in use at this frequency. Therefore, the 0.5 MHz Gamma transducers described above are not utilized for the configurations described herein. Since these transducers are permanently affixed to the block, a repeatable attenuation factor was determinable across the channel. Hereafter, unless specifically stated otherwise, the “channel” refers to the path a signal takes through and including each contact transducer and through the medium. That is, the channel attenuation includes that of the transmit UT, the medium, and the receive UT. Utilizing a sinusoidal carrier fixed at 1.0 MHz, the measured channel attenuation through the 1.0 MHz Gamma transducers is -41.85 dB and through the 1.0 MHz alpha transducers is -53.63 dB.

¹ See Appendix A.1 for transducer compliance data.

² See Appendix A.2 for transducer compliance data.

Figure 3 - Capture of multipath channel reflections for the rectangular test apparatus.

As mentioned earlier, the channels (i.e., the rectangular and cylindrical test apparatus) are non-ideal in that they suffer from severe multipath reflections. This is of significant importance as it ultimately leads to multipath fading phenomena which adversely affect the signals of interest. Again, the rectangular test apparatus is nice for multipath propagation analysis in that the geometry is fixed, regular and to some degree, the dominant reflective signals are deterministic. Since the end-goal implementation of the current research seeks to employ such a communication system for any solid thick-wall of interest, it is unnecessary to develop a fully deterministic model of the channel. Since the medium is not well-understood at the development stage, utilizing adaptive filtering algorithms is the most logical approach to this problem which is exemplified herein. Regardless, gaining a baseline understanding of the rectangular test apparatus is

useful for extension to other media. Consequently, a representation of the multipath phenomenon for the rectangular geometry can be visualized in Figure 3 above.³

The representation shown in the figure is of a one-way pulse transmission between the two 1.0 MHz Gamma probes attached to the rectangular block. For this test, a Krautkramer-Branson USIP-12 Non-Destructive Testing (NDT) pulser-receiver is utilized in the dual transducer mode (i.e., simultaneous pitch and catch). A 1 MHz pulse at a fixed repetition rate is fed into one UT while the received signal is captured in the figure from its trans-wall mate. It can be seen in the figure that due to the fixed, rectangular geometry of the block the reflections occur at regular intervals. Specifically, the main LOS signal can be seen at time 25 μ s which corresponds to the direct length of the path. Subsequent reflections seen at odd multiples of the path length (3L, 5L, 7L, ...) correspond to the signal being reflected off of the receiving face of the block, back off of the originating face, and finally to the receiving face multiple times. Additional received pulses from undetermined locations occur at the notations. Note that they are undetermined, but of regular interval. Due to their proximity to the main pulse, it is speculated that these are the result of reflections off of the back wall of each UT. Similar results are seen with the cylindrical vessel, which may point to transducer—not geometric—effects.

In terms of the main pulse reflections, the sources of the multipath phenomena can be better understood by visualizing possible acoustic paths as shown in Figure 4 below. Of course, it is clear that an infinite number of acoustic paths are possible. The paths noted in the analysis in the figure are those which have propagation times in close proximity to the main pulses and are of limited attenuation. The figure demonstrates dominant paths and their locations within the steel block. Clearly, as the geometry of the channel becomes more complex, obtaining a deterministic analysis of the paths becomes less achievable. This is certainly the case where the complexity of such an analysis is exacerbated by the geometry of the cylindrical test apparatus. A description of each of the dominant paths is presented in Table 2 below. In addition, Table 3 denotes each dominant path along with their associated propagation times to the receiving transducer.

³ Reflection path and transducer directivity analysis contained herein courtesy of Seth Bard (MANE RPI '05) and Matthew Wetzel (MANE RPI '06).

Of key importance is the intensity of each received signal, normalized to the direct propagation path. Data is presented for both the 0.5 MHz and 1.0 MHz Gamma transducers. Clearly the data shows that the 1.0 MHz Gamma transducer has more focused directivity over the 0.5 MHz probe. Appendix B contains directivity plots for the 0.5 MHz and 1.0 MHz Gamma probes.

Figure 4 - Diagram of select acoustic paths in the rectangular test apparatus.

Acoustic Path	Description
Direct	A straight line longitudinal wave from the transmitter to the receiver.
Top / Bottom Longitudinal	Longitudinal waves reflecting midway along the top or bottom surfaces.
Side Longitudinal	A longitudinal wave reflecting midway along the side wall.
Top / Bottom Shear	Shear waves generated at a calculated distance beyond the midway point along the top or bottom surfaces.
Side Shear	A shear wave reflecting at a calculated distance beyond the midway point along the side wall.
Top & Bottom Longitudinal	Longitudinal waves reflecting at $\frac{1}{4}$ and $\frac{3}{4}$ of the length along the top and bottom surfaces.
1 st Direct Reflection	The first echo along the direct path returning to the receiver.
2 nd Direct Reflection	The second echo along the direct path returning to the receiver.

Table 2 - Acoustic paths and description of their nature.

Acoustic Path	0.5 MHz Gamma Probe		1.0 MHz Gamma Probe	
<i>Name</i>	<i>Time (μs)</i>	<i>Normalized Intensity</i>	<i>Time (μs)</i>	<i>Normalized Intensity</i>
Direct	26.0	1.0000	26.0	1.0000
Top / Bottom Longitudinal	29.2	0.0306	29.2	0.0041
Side Longitudinal	64.7	0.0036	32.8	0.0042
Top / Bottom Shear	54.2	0.3418	54.2	0.0006
Side Shear	100.0	0.0094	57.6	0.0090
Top & Bottom Longitudinal	37.2	0.0170	37.2	0.0001
1 st Direct Reflection	78.0		78.0	
2 nd Direct Reflection	130.0		130.0	

Table 3 - List of acoustic paths, propagation times, and normalized intensity for Gamma probes.

It turns out that the reflections due to the initial, dominant propagating pulse is of little significance from a communications standpoint. As a result of the high oscillatory nature of the probes and the reflective nature of the channel, ringing within the apparatus is of considerable amplitude and duration well after the aforementioned dominant acoustic paths. These properties are clearly visible in Figure 5 below. In this test, a 1.0 MHz pulse is transmitted along the path of the 1.0 MHz Gamma transducers through the channel. Note that the buffered (amplified) display output of the NDT pulser is used to capture the signals with the DAQ. Therefore, the amplitudes shown in the plot are not representative of the actual signal. For instance, the generated pulse is plotted on the same scale as the channel response and can be seen to the left on the figure. This pulse is attenuated by approximately 40 dB prior to being displayed on the plot. Henceforth, this buffered acquisition is generally the case for experimentation.

Figure 5 - Response of the rectangular channel to a pulse through the 1.0 MHz probes.

Note that apart from the dominant path reflections described earlier in Table 2 and Table 3, significant ringing exists in the channel for over 2 milliseconds. Thereafter, the ringing follows an exponential decay for nearly an additional 8 ms. Comparing this result with the aforementioned, it is clear that there are other effects within the channel that are not well understood. This is of little consequence from a signal processing standpoint, as processing can simply wait for channel rest or employ adaptive algorithms as described herein. The exponential decay—described more analytically in subsequent sections—within the channel can be explained by the incremental attenuation of the

medium. However, the sustained oscillation following the dominant reflective pulses is not well understood. It is hypothesized that the result is due to the tight coupling (due to the epoxy bond) of each transducer to the acoustic medium. One transducer may produce a complimentary oscillatory effect with its mate. This postulate may be supported in part by the result obtained for the cylindrical vessel later, which shows less oscillatory time in response to a pulse. The “looser” acoustic coupling, curved geometry, and high incremental attenuation of the Lexan transition pieces may mitigate such a complementary oscillatory effect.

3.2 Cylindrical Test Apparatus

After design configurations were verified on the fixed rectangular test block, they were then tested on an actual decommissioned pressure vessel with an inner diameter of 12.5 cm and an outer diameter of 24 cm. The main reason for demonstrating achievable results on this cylindrical test apparatus is that it most closely emulates what may be a typical medium for the final installation of such a system. A directivity analysis was not conducted for this apparatus due to its complex geometry. The vessel is, however, similar to the rectangular apparatus in that it also suffers from multipath phenomena. Four Panametrics-NDT Accuscan “S” A539S, 0.5 inch, 1.0 MHz, lightly damped contact transducers were obtained for use with the vessel in the various configurations.⁴ From a reconfiguration standpoint, permanently affixing said transducers to the vessel is not practical. Therefore, appliances were manufactured that allowed the temporary attachment of the transducers to the vessel with Lexan transition pieces machined to fit the curvature of the vessel. Images of these appliances can be seen in Figure 6, Figure 7, and Figure 8 below.

⁴ See Appendix A.3 for transducer compliance data.

Figure 6 - Overall view of the pressure vessel and connected transducers.

Figure 7 - Zoomed aerial view of two transducers on the inside of the vessel.

Figure 8 - Zoomed view of two outer transducers affixed to the vessel.

Note that due to the fact that the transducers are not permanently attached, there is some difficulty in reaching precise alignment. Consequently, due to this and other factors (e.g., depletion of the viscous couplant used over time) results are not strictly repeatable over time. Foregoing a comprehensive directivity analysis, the one-way transducer-to-transducer response can be seen below in Figure 9. Note the similarity of the exponential decay to that of the rectangular channel in Figure 5. One noticeable difference is the lack of oscillatory ringing for a period as lengthy as that of the rectangular channel. Again, it is postulated that the “loose” acoustic coupling to the media, the high attenuation of the Lexan transition pieces, and the curved geometry may downplay what is a complementary transducer oscillatory effect. Note that the response seen in Figure 9 has a time scale double that of Figure 5 above. This discrepancy may prevent the reader from noticing the slightly exaggerated oscillatory response for the rectangular apparatus. Nonetheless, even when compared on the same time-scale the disparity is quite large—almost 2 ms longer for the rectangular apparatus.

Figure 9 - Response of the cylindrical channel to a pulse through the 1.0 MHz probes.

3.3 Towards an Approximate Channel Model

Intuition may bring one to believe that the communication challenge may be wholly an acoustic problem. In fact, there are many parallelisms between acoustic and electromagnetic (EM) theory which provide analogous links between the two disciplines. If we form the analog of the UT as a general communications transmitter and the longitudinal acoustic wave (LAW) emitted by it as a transverse electromagnetic (TEM) wave, then clearly we move into a scheme that is the familiar communication channel. From a communications design standpoint, this is a necessity before developing a suitable channel model. In fact, to this day one of the open challenges in ultrasonics is the development of a fully deterministic transducer and channel model.

While the mathematical tensor formalism for piezoelectric transducers is well developed in many fundamental texts in ultrasonic theory, much of it can be simplified when the medium of interest is isotropic [53]. Both test appliances are characteristically isotropic throughout: the rectangular appliance consists of a steel alloy and the cylindrical vessel consists of inconel alloy 600 (a nickel derivate). The simplification of said formalism stems from the condition that there can be no more than two independent elastic constants in an isotropic medium. In Kino's text [53]—a purported classic within the ultrasonic community—the development of wave equations in isotropism is presented.

The tensor motion in the x, y, and z directions are given as:

$$\rho_{m_0} \frac{\partial^2 u_x}{\partial t^2} = \frac{\partial T_{xx}}{\partial x} + \frac{\partial T_{xy}}{\partial y} + \frac{\partial T_{xz}}{\partial z}, \quad (3.3.1)$$

$$\rho_{m_0} \frac{\partial^2 u_y}{\partial t^2} = \frac{\partial T_{xy}}{\partial x} + \frac{\partial T_{yy}}{\partial y} + \frac{\partial T_{yz}}{\partial z}, \quad (3.3.2)$$

$$\rho_{m_0} \frac{\partial^2 u_z}{\partial t^2} = \frac{\partial T_{xz}}{\partial x} + \frac{\partial T_{yz}}{\partial y} + \frac{\partial T_{zz}}{\partial z}, \quad (3.3.3)$$

or alternatively, in vector form:

$$\nabla \cdot \vec{\mathbf{T}} = \rho_{m_0} \frac{\partial^2 \mathbf{u}}{\partial t^2}, \quad (3.3.4)$$

where the T_{ij} quantities are the stress tensors in the appropriate directions, ρ_{m_0} is the mass density of the medium, and the u_i quantities are the particle displacements in the appropriate directions.

By observation, this bears resemblance to the wave equation derived from Gauss' law as put forth in EM theory given for time-harmonic fields as:

$$\nabla \cdot \vec{\mathbf{E}} = \frac{\tilde{\rho}_v}{\varepsilon}, \quad (3.3.5)$$

or equivalently, for charge-free media:

$$\nabla^2 \vec{\mathbf{E}} - \gamma^2 \vec{\mathbf{E}} = 0, \quad (3.3.6)$$

with:

$$\gamma^2 \triangleq -\omega^2 \mu \varepsilon_c, \quad (3.3.7)$$

where $\vec{\mathbf{E}}$ is the vector electric field phasor, $\tilde{\rho}_v$ is the charge density phasor, ε is the permittivity, and γ is the complex propagation constant (with radian frequency ω , magnetic permeability μ , and complex electrical permittivity ε_c) of the medium [54]. Note that the phasor quantities are periodic functions in time (i.e., time-harmonic).

Along these lines, it is often argued that an acoustic-electrical analog can be totally formulated to develop an equivalent circuit for the UT in general based on the similarities of the acoustic and TEM waves. That is, the degree to which an acoustical

wave can be viewed as an electrical wave in one dimension and vice versa dictates the appropriate use of such a model. The observed similarities in the wave equations are better explained for this analog by proofs utilizing Poynting's theorem, which demonstrates energy conservation for the traveling waves. Specifically, the energy of a propagating wave through a closed contour must be conserved during conversion from electrical to acoustical and vice versa. Hence, an analog is formed between the spatial displacement versus time and electrical displacement versus time to create a one-dimensional model. Therefore, particle velocity is analogous to electric current and mechanical stress is analogous to electric voltage [53]. There are two historically competing models developed independently by Redwood [55] and Mason [56]. Each of these models develops the electrical analog of the ultrasonic transducer as a two-port model. An example of this type of formulation was attempted based on the work Maione, et al. [57] and can be found in Appendix C.⁵ Models utilizing this method, based on a survey of the literature, appear to be in widespread use. For example, Zipparo investigated the development of a computer program to model PZT transducers based on the Mason model [58]. These models provide a nice vehicle for simulating the ultrasonic PZT transducer.

However, these models are cumbersome to work with and often deviate from practical results due to indefinite assumptions made in their development.⁶ Moreover, while they provide excellent models for the PZT transducer device itself, the model of the propagating medium (i.e., the acoustic channel) is not well developed. In addition, from a signal processing standpoint a fundamental understanding of the manifestation of the signals of interest is of more importance than precise numerical agreement. This is especially true in deciding which signal processing techniques may be of greatest efficacy. Thus, a simpler approximation is attempted to gain an appreciation for the characteristic appearance of signals transmitted through the channel. This one-dimensional approximation also provides ready characterization in most standard

⁵ PSPICE model and simulation courtesy Matthew Wetzel (MANE RPI '06).

⁶ An interesting paper by Whitworth claims precise numerical agreement by avoiding small calculation errors cited elsewhere within the literature. See: "Discussion of One-D Piezoelectric Transducer Models with Loss." *IEEE Transactions on Ultrasonics, Ferroelectrics, and Frequency Control*, 48, No. 3 (May 2001), 844-846.

computational programs. It should be emphasized that the following development is simply an approximation to empirical results and should be utilized more for characterizing results useful for the signal processing problem rather than a wholly deterministic solution. Appendix D.1 contains the Matlab code used to obtain the plots in this section.

3.3.1 Line-of-Sight Channel Model

The preliminary search for a suitable channel model should begin with the direct, line-of-sight (LOS) transmission path. However, even the simplistic nature of the single-hop path is pervaded by multipath effects. The nature of the acoustic path is that it is highly reflective off of each face of the test apparatus. Clearly, the characteristics of the reflection process are determined by the intrinsic properties of the material and shape of the medium. Furthermore, we are dealing with piezoelectric transducers (PZT) which are highly resonant at their center frequencies [53]. It is plausible that the acoustic coupling of the propagating signal from the medium to the sink transducer further distorts the signal due to reverberations being retransmitted into the channel by it.

Figure 10 - Impulse response decomposition of the line-of-sight transmission path.

The LOS transmission path can consequently be viewed as the passing of the transmitted signal through a cascade of three transfer functions: that of the transmitting transducer $[h_{TX}(t)]$, that of the apparatus channel $[h_{CH}(t)]$, and that of the receiving transducer $[h_{RX}(t)]$. An overview of this decomposition can be seen in Figure 10 above. The convolution of these three system impulse response functions fully characterizes the LOS transmission path. That is,

$$h(t) = h_{TX}(t) * h_{CH}(t) * h_{RX}(t), \quad (3.3.8)$$

is the overall LOS channel impulse response.

3.3.2 Transducer Model Approximation

Since the PZT ultrasonic transducer (UT) is characteristically resonant at its center frequency, we would expect that the frequency spectrum for it would be similar to a realized band-pass filter (BPF). While the manufacturing variability between brands and makes of each UT is potentially very large, little information has been found with respect to formalizing the frequency response of a particular transducer. Realized spectrums, which characterize the post-manufacturing response of a UT, are generally provided with ultrasonic transducers. Of particular importance to the current instance is that of the Panametrics-NDT A539S 0.5-inch contact transducers utilized for the cylindrical test apparatus communication system because of the likelihood of their use in future implementations.

Transducer Number:	1	2	3	4
Center Frequency (f_c)	0.93 MHz	0.92 MHz	0.92 MHz	0.95 MHz
Relative -6 dB Bandwidth	36.90%	37.26%	39.54%	40.27%
Approximate -6 dB Bandwidth	343.2 kHz	342.8 kHz	363.8 kHz	382.6 kHz

Table 4 - Transducer spectrum measurements.

In each test implementation, identical transducers are used at each side of the channel. Incidentally, the specification sheets provided for each UT from the manufacturer in Appendix A provides the frequency response as measured per ASTM E1065

requirements [59]. The response in the frequency domain can be approximated by a Gaussian distribution centered at the tuned frequency, with peaked kurtosis comparable to the aforementioned BPF realization. Specifically, for the four Panametrics-NDT A539S contact transducers utilized the values in Table 4 above are reported. It is observable that even in this small batch of UT measurements, there is quite a bit of process variability. Therefore, to develop an idealized model and facilitate computational ease we can therefore state that in (3.3.8),

$$h_{TX}(t) = h_{RX}(t) \xleftrightarrow{\mathfrak{F}} H_{TX}(f) = H_{RX}(f). \quad (3.3.9)$$

Hence, the impulse response of an individual UT can be denoted by $h_T(t)$. That is:

$$h_{TX}(t) = h_{RX}(t) = h_T(t). \quad (3.3.10)$$

Furthermore, the following assumption will herein be asserted:

$$f_{c1} = f_{c2} = f_{c3} = f_{c4} = f_c, \quad (3.3.11)$$

where f_c is, in general, the system center frequency. Consequently, to maximize system throughput we shall set the carrier frequency to equal that of f_c . It is clear from the data presented in Table 4 that the center frequency of each transducer deviates somewhat from the ideal (i.e., 1.00 MHz). In general, the assumption is made that this is not the case.

It is also evident from the manufacturer's datasheets that the actual frequency spectra may suffer from considerable skew when assumed to follow the Gaussian distribution. Therefore, to complete the idealization of the LOS transmission path transfer function, it is assumed that the distribution has no skew. We can conclude that the frequency spectrum for each transducer can be ideally modeled with the Gaussian distribution as [60]:

$$H_T(f) = \frac{1}{\sqrt{2\pi}\sigma} e^{\frac{-(f-f_c)^2}{2\sigma^2}} \quad -\infty < f < +\infty. \quad (3.3.12)$$

The ideal transducer transfer function should have unity gain at its center frequency. To ensure this, an arbitrary multiplicative constant β will be added as:

$$H_T(f) = \beta \frac{1}{\sqrt{2\pi}\sigma} e^{\frac{-(f-f_c)^2}{2\sigma^2}} \quad -\infty < f < +\infty \quad (3.3.13)$$

For unity at the center of the Gaussian, the function:

$$H_T(f_c) = \beta \frac{1}{\sqrt{2\pi}\sigma} = 1 \quad (3.3.14)$$

results. By inspection:

$$\beta = \sqrt{2\pi}\sigma \quad (3.3.15)$$

which implies that (3.3.12) could be simplified, but it will be left in its current form to ease transformation back to the time domain with transform properties. Although not explicitly specified, it seems as though the frequency spectra provided in the datasheets are normalized to unity at their peak.

The center frequency f_c is observed to be the mean of the distribution, that is:

$$f_c = \mu. \quad (3.3.16)$$

Noting that the -6 dB bandwidth is equivalent to approximately a 50% decrease in the spectrum from its maximum, it is possible to determine the variance (σ^2) of the spectrum in terms of the -6 dB bandwidth.

Assume that the distribution has zero mean. Equation (3.3.12) then becomes:

$$H_T(f) = \beta \frac{1}{\sqrt{2\pi}\sigma} e^{\frac{-(f-0)^2}{2\sigma^2}} = \beta \frac{1}{\sqrt{2\pi}\sigma} e^{\left(\frac{-f^2}{2\sigma^2}\right)} \quad -\infty < f < +\infty. \quad (3.3.17)$$

If the spectrum is assumed to be identically distributed, then the upper- and lower- -6 dB frequencies are related to the 6 dB bandwidth (denoted as BW) as:

$$BW = f_u - f_l \quad (3.3.18)$$

The maximum (peak) of the spectrum occurs at the DC frequency (when the mean is assumed to be zero). The approximate ratio of the spectrum evaluated at DC and the upper- -6 dB frequency is equivalent to:

$$\frac{H_T(f_u)}{H_T(0)} = \frac{\beta \frac{1}{\sqrt{2\pi}\sigma} e^{\left(\frac{-f_u^2}{2\sigma^2}\right)}}{\beta \frac{1}{\sqrt{2\pi}\sigma} e^{\left(\frac{-(0)^2}{2\sigma^2}\right)}} = \frac{\beta \frac{1}{\sqrt{2\pi}\sigma} e^{\frac{-f_u^2}{2\sigma^2}}}{\beta \frac{1}{\sqrt{2\pi}\sigma}} = e^{\frac{-f_u^2}{2\sigma^2}} \approx \frac{1}{2} \quad 0 \leq f_u < +\infty. \quad (3.3.19)$$

Solving for the variance:

$$\sigma^2 \approx \frac{-f_u^2}{2 \ln\left(\frac{1}{2}\right)} = \frac{f_u^2}{2 \ln(2)} \quad 0 \leq f_u < +\infty. \quad (3.3.20)$$

In general, zero skew implies:

$$|f_u - f_c| = |f_l - f_c| = \frac{BW}{2}. \quad (3.3.21)$$

Therefore, we can write the final approximation of the variance for this idealization in terms of the -6 dB bandwidth as:

$$\sigma^2 \approx \frac{BW^2}{8 \ln(2)}. \quad (3.3.22)$$

For real systems, we note that the frequency response must be an even function [64]. Therefore, to translate (3.3.13) into a practical function, (3.3.17) is convolved in the frequency domain with a cosine carrier at the transducer center frequency to mirror the function about the DC line. Hence, we obtain:

$$H_T(f) = \{\delta(f - f_c) + \delta(f + f_c)\} * \beta \frac{1}{\sqrt{2\pi\sigma}} e^{\left(\frac{-f^2}{2\sigma^2}\right)}. \quad (3.3.23)$$

Convolution in the frequency domain translates to multiplication in the time domain, hence given:

$$e^{-\pi(bt)^2} \xleftrightarrow{\mathfrak{F}} \left(\frac{1}{b}\right) e^{-\pi\left(\frac{f}{b}\right)^2}, \quad (3.3.24)$$

$$\begin{aligned} h_T(t) &= \{\cos(2\pi f_c t)\} \beta e^{-2\pi^2(\sigma t)^2} \\ &= \{\cos(2\pi f_c t)\} \sqrt{2\pi\sigma} e^{-2\pi^2(\sigma t)^2}, \quad -\infty < f < +\infty, \sigma \neq 0 \end{aligned} \quad (3.3.25)$$

3.3.3 Channel Model Approximation

The isotropic medium is often characterized in terms of its fundamental acoustic parameters. While the propagation of an acoustic wave through an isotropic medium is a complex interplay of constituent channel and wave parameters over three dimensions [53], a rudimentary communications model is sought. More specifically, from a communications standpoint we are most concerned with the one-dimensional signal model

(i.e., voltage vs. time) so that we can gain a fundamental understanding of the characteristics of the received signal to be utilized in signal processing. Along these lines, the use of the incremental attenuation per unit length of the medium can be used to formulate the basic model.

Many sources provide the basic acoustic material parameters of an isotropic medium in tabulated form (e.g., see [61]). Of particular importance are the incremental attenuation (typically given in dB/cm) and acoustic wave velocities within the medium (typically given in cm/s). For the latter, parameters are given for both the longitudinal and shear waves. In the current implementation, UT's operating in the longitudinal mode are utilized and, for the model idealization, it is assumed that no mode conversion between longitudinal and shear waves occur.

Observe first that the attenuation per distance is given by:

$$Attenuation_{dB} = \frac{-20 \cdot \text{LOG}_{10} \left(\frac{x}{x_0} \right)}{d}, \quad d \neq 0, \quad (3.3.26)$$

where x is the magnitude of the signal at distance d from the source, and x_0 is the magnitude of the originating signal. A form of this equation that readily permits manipulation with the Fourier transform from the frequency domain is attractive. Therefore, noting that in general [62]:

$$\text{LOG}_{10}(a) = \frac{\ln(a)}{\ln(10)}, \quad (3.3.27)$$

equation (3.3.26) can be written as:

$$Attenuation_{dB} = -\frac{20}{d} \cdot \frac{\ln \left(\frac{x}{x_0} \right)}{\ln(10)}, \quad d \neq 0. \quad (3.3.28)$$

Rearranging, we obtain:

$$\ln \left(\frac{x}{x_0} \right) = \ln(x) - \ln(x_0) = -\frac{Attenuation_{dB} \cdot d \cdot \ln(10)}{20}, \quad x_0 \neq 0. \quad (3.3.29)$$

Solving for x ,

$$\begin{aligned}
x &= e^{\frac{-\text{Attenuation}_{dB} \cdot d \cdot \ln(10)}{20} + \ln(x_0)} \\
&= e^{\frac{-\text{Attenuation}_{dB} \cdot d \cdot \ln(10)}{20}} e^{\ln(x_0)} \\
&= x_0 \cdot e^{\frac{-\text{Attenuation}_{dB} \cdot d \cdot \ln(10)}{20}} .
\end{aligned} \tag{3.3.30}$$

An equivalent form, which incorporates the wave speed (v) of the medium, yields a function of time:

$$x(t) = x_0 \cdot e^{\frac{-\text{Attenuation}_{dB} \cdot (t \cdot v) \cdot \ln(10)}{20}} . \tag{3.3.31}$$

If we set:

$$\alpha = \frac{\text{Attenuation}_{dB} \cdot v \cdot \ln(10)}{20} , \tag{3.3.32}$$

we can write an equivalent form as:

$$x(t) = x_0 \cdot e^{-\alpha t} . \tag{3.3.33}$$

Clearly, this is a multiplicative function—that is, x_0 is the signal that is given as the input to the channel. For real, causal systems $t \geq 0$. Hence, we can use the Fourier transform for translation into the frequency domain [63]:

$$x(t) = x_0 e^{-\alpha t} u(t) \xleftrightarrow{\mathfrak{F}} X(f) = x_0 \frac{1}{\alpha + j2\pi f} . \tag{3.3.34}$$

3.3.4 Combined LOS Approximation

Given the transducer transfer function generalized and approximated by (3.3.12) the development of the overall LOS transfer function given by (3.3.8) can be attempted. The convolution operator is commutative [64], so given the general transducer impulse response $h_T(t)$, (3.3.8) may be written as:

$$h(t) = h_T(t) * h_{CH}(t) * h_T(t) = [h_T(t) * h_T(t)] * h_{CH}(t) . \tag{3.3.35}$$

The system function for the transducer is approximated by the Gaussian and the convolution of Gaussian with another Gaussian (in this case itself) will yield another Gaussian. Since convolution in the time domain is equivalent to multiplication in the frequency domain, we may equivalently write:

$$h_T(t) * h_T(t) \xleftrightarrow{\mathfrak{F}} H_T(f) \cdot H_T(f), \quad (3.3.36)$$

which yields a generalized transducer-to-transducer response of:

$$\begin{aligned} H_{T-T}(f) &= \left(\left\{ \delta(f - f_c) + \delta(f + f_c) \right\} * \beta \frac{1}{\sqrt{2\pi}\sigma} e^{\left(\frac{-f^2}{2\sigma^2}\right)} \right) \\ &\quad \left(\left\{ \delta(f - f_c) + \delta(f + f_c) \right\} * \beta \frac{1}{\sqrt{2\pi}\sigma} e^{\left(\frac{-f^2}{2\sigma^2}\right)} \right) \\ &= \left(\left\{ \delta(f - f_c) + \delta(f + f_c) \right\} * \left(\beta \frac{1}{\sqrt{2\pi}\sigma} e^{\left(\frac{-f^2}{2\sigma^2}\right)} \right)^2 \right) \quad (3.3.37) \\ &= \left(\left\{ \delta(f - f_c) + \delta(f + f_c) \right\} * e^{\left(\frac{-f^2}{\sigma^2}\right)} \right) \end{aligned}$$

Using the properties of the Fourier transform, this can be written in the time-domain as [65]:

$$\mathfrak{F}^{-1} \left\{ \left\{ \delta(f - f_c) + \delta(f + f_c) \right\} * e^{\left(\frac{-f^2}{\sigma^2}\right)} \right\} = \sqrt{\pi}\sigma \cos(2\pi f_c t) e^{-(\pi\sigma t)^2}. \quad (3.3.38)$$

As it turns out, the exact form in (3.3.38) requires a large inverse scaling factor to result in an approximation comparable to the actual results obtained empirically. This can stem from the normalization of the transducer impulse response (i.e., to obtain unity gain at its center frequency) as mentioned earlier as well as other reasons. The adjustment term will be added in (3.3.38) as κ :

$$h_{T-T}(t) = \kappa \sqrt{\pi}\sigma \cos(2\pi f_c t) e^{-(\pi\sigma t)^2}. \quad (3.3.39)$$

All other parameters remained the same. Taking $\kappa = \frac{1}{100\sqrt{\pi}\sigma}$ and the bandwidth (as derived from Table 4 above) to be 300 kHz, the result seen in Figure 11 is determined as the cascaded transducer-to-transducer impulse response. Again that kappa (κ) is characteristic to the specific medium in use.

Figure 11 - Cascaded transducer-to-transducer impulse response (bandwidth = 300 kHz).

A plot of the frequency spectrum of the result shown in Figure 11 can be seen in Figure 12 below as a confirmation of the result. For emphasis, note that the -6 dB bandwidth corresponds approximately to the bandwidth of 300 kHz specified.

Figure 12 - Cascaded transducer-to-transducer frequency response (bandwidth = 300 kHz).

To better approximate the empirical result obtained, an input pulse similar to that of an idealized pulse from a device such as the Krautkramer-Branson NDT pulser is formulated as:

$$x_{pulse}(t) = 200 \{u(t - 0.5 \times 10^{-6}) - u(t + 0.5 \times 10^{-6})\}, \quad (3.3.40)$$

where $u(t)$ is the unit step function, or, equivalently:

$$x_{pulse}(t) = \begin{cases} 200, & |t| \leq 0.5 \mu s \\ 0, & else \end{cases}. \quad (3.3.41)$$

This is simply a 200 V ideal pulse that is 1 microsecond in duration (i.e., at a frequency of 1.0 MHz). A plot of (3.3.40) and (3.3.41) can be seen in Figure 13 below.

Figure 13 - Plot of the idealized pulse from the NDT device.

The next step is to model the attenuation within the channel. Due to the incremental nature of the attenuation, the function of the channel's attenuation is dependent on distance and is therefore multiplicative as in (3.3.33). Therefore, the overall, one-way impulse response of the channel is given by:

$$\begin{aligned}
 h_{channel}(t) &= \left\{ \kappa \sqrt{\pi} \sigma \cos(2\pi f_c t) e^{-(\pi \sigma t)^2} \right\} e^{-\alpha t} \\
 &= \left\{ \frac{1}{1000} \cos(2\pi f_c t) e^{-(\pi \sigma t)^2} \right\} e^{-\alpha t}, \quad t \geq 0
 \end{aligned} \tag{3.3.42}$$

Note that $t \geq 0$ to obtain a real, causal channel response. The finalized approximated channel response to a pulse similar to that from the NDT pulser (i.e., not the *impulse* response) is obtained by:

$$y_{pulse}(t) = x_{pulse}(t) * h_{channel}(t) \tag{3.3.43}$$

which was computed numerically and can be seen in Figure 14 below. Note that the values used for the incremental attenuation and wave speed for the rectangular apparatus channel were empirically determined to be approximately $2.75 \frac{dB}{cm}$ and $5.789 \times 10^5 \frac{m}{s}$, respectively. Again, these parameters are used to determine α above.

Figure 14 - Approximate one-way channel response to a pulse.

3.3.5 Combined LOS Approximation with Elementary Multipath

Given the approximate channel impulse response developed in (3.3.42) above, a generalized function for the dominant channel multipath response can be defined. From the aforesaid characteristics of the LOS channel, it is clear that the channel has a dominant reflection path which is along the direct, horizontal path between two coaxially mounted transducers. The clearest demonstration of this can be observed in Figure 3

earlier. For the most basic multipath process, it is assumed that reflections occur only in the horizontal direction off of the faces at the transducer-apparatus interface. Therefore, the received multipath signal for the dominant reflection process can be given as a function of the initial dominant LOS signal followed by an infinite number of its time-delayed images as [66]:

$$y_{multipath}(t) = y(t) + \sum_{n=1}^{\infty} y(t - n\tau), \quad (3.3.44)$$

where τ is the time it takes for an acoustic signal to traverse to the receiving face, back to the originating face, and finally back to the receiving face. Clearly, this parameter is a function of the wave speed (v) within the channel:

$$\tau(v) = \frac{2l}{v}, \quad (3.3.45)$$

where l is the length of the channel.

Figure 15 - Channel pulse response with dominant multipath effect.

Note that (3.3.44) is the case where lossless reflections are considered. However, the dominant multipath phenomenon also follows an exponential decay. Therefore, (3.3.44) is updated by absorbing the dominant LOS signal into the summation as:

$$y_{\text{multipath}}(t) = \sum_{n=0}^{\infty} e^{-\eta t} y(t - n\tau) \quad (3.3.46)$$

where the attenuation parameter η is characteristic to the channel. This value was empirically determined to be 7800 s^{-1} for the rectangular test apparatus. Utilizing the aforementioned wave speed parameter and shifting $y(t)$ such that $t = 0$ corresponds to the origination of the pulse at the transmit side of the channel, Figure 15 above is determined. In comparison to Figure 3 earlier, it can be seen that this approximation bears close resemblance to the actually determined response. Note that the current approximation does not include the echoes which are of undetermined origin; although a simple modification to the channel model could readily account for this.

Figure 16 - DBPSK data burst received through simulated multipath channel.

Figure 16 above demonstrates the usefulness of this type of approximated model. Using a model along these lines, the simulated transmission of a random DBPSK data burst (see subsequent sections for more detail on this mode) through a channel with an impulse response as that of (3.3.46) is obtained. Note the similarity to the actual results depicted in Figure 36 later. Phase transitions cause the transients that can be seen in the figure. This model can be appropriately augmented to model channels and signaling paradigms of interest. In fact, this type of approximation provided initial simulation results to merit study in the area.

Figure 17 - Channel pulse response with dominant multipath effect and $N(0,0.5)$ Gaussian noise.

Note that the result in Figure 3 also includes acoustic noise of relatively low power. This can be *ostensibly* modeled as Gaussian noise with equivalent noise power (vari-

ance) solely for demonstration and approximation purposes as in Figure 17 above. Consequently, (3.3.46) can be updated to reflect this noise component as:

$$y_{\text{multipath}}(t) = y(t) + n(t) + \sum_{n=1}^{\infty} e^{-\eta t} y(t - n\tau), \quad (3.3.47)$$

where $n(t)$ is zero-mean Gaussian noise.

3.3.6 Combined LOS Approximation with Extended Multipath

Clearly, the multipath phenomenon is of greater complexity than the model in (3.3.46) suggests. There are an infinite number of possible acoustic paths which are random and depend on many factors (e.g., geometry of the medium, material properties, etc.). A key difference is that, in actuality, the acoustic paths are free to deviate from the single direct, LOS path. There are an infinite number of ray traces within the propagating medium rather than the earlier simplification of one dominant LOS path. In addition, not only is the multipath effect of infinite summation, it is of infinite recursion—that is, signals received through multiple paths also suffer from multipath effects ad infinitum. The problem is exacerbated by the fact that the associated propagation paths are random in nature.

As a result, an exact deterministic channel model is practically intangible. It is evident that an approximate model development can be made by simply estimating the envelope of the channel response. With this information, a random variable representing the delay spread of the multipath images can be utilized. Towards this end, probabilistic models are most often used to model the multipath channel. Because the spatial separation of the receiver and transmitter in this installation are fixed, there is little short-term fading due to Doppler variations. Instead, long-term or “slow” fading models are better utilized. A major characteristic of slow fading is that within small relative time intervals, the channel impulse response remains approximately constant. Over larger time intervals, however, environmental factors (e.g., temperature) can affect the channel response. This effect is apparent in the results with the use of LMS suppression, where periodic re-adaptation over long time periods is necessary. In typical slow fading multipath systems, it has been shown that the fading follows a log-normal distribution

[67]. Visual inspection of the channel results confirms the progression to this type of a model. Consider the probability density function of the log-normal random variable in Figure 18 below with expected value (μ) equal to 0.⁷

Figure 18 - PDF of the log-normal random variable with $\mu=0$.

Specifically, the PDF for a standard log-normal expression is a function of the path SNR per symbol (γ) which is typically given as [68]:

$$f_{\gamma}(\gamma) = \frac{10}{\sqrt{2\pi\sigma\gamma\ln(10)}} \exp\left[-\frac{(10\log(\gamma)-\mu)^2}{2\sigma^2}\right], \quad (3.3.48)$$

where the mean (μ) and variance (σ^2) are that of $10\log(\gamma)$. This probabilistic approximation could be utilized to develop a more rigorous model of the multipath channel in the current application with the choice of appropriate parameters characteristic to the

⁷ From <http://en.wikipedia.org/wiki/Lognormal> (under GNU public license).

channel. However, a comprehensive development and evaluation is beyond the scope of the current work. The previous discussion is to serve as a demonstration of underlying channel properties and to gain an appreciation of the extensive multipath interference.

4. Transducer Mounting and Signal Routing Methods

Given the anticipated volatile nature of the environment internal to the pressure vessel, the development of a simplistic modulation scheme was sought. The high pressure and high temperature environment would prove perilous to any type of sensitive communication blocks which are typically utilized for some of the more complex communication modes. As a result, methods that minimize the complexity of components on the inside of the vessel are the most attractive. The system development evolved through two major phases. The first phase of the project revolved around demonstrating the feasibility of an acoustic communications link through a steel block. The second phase of the project sought to move towards an implementation that would more closely represent a typical end-goal installation on a pressure vessel. The major difference between the two phases is apparent in the sole origination of signals from “inside” of the vessel (i.e., oscillation and modulation) in the first phase whilst the second phase originated signals external to the vessel and simply modulated them internally.

4.1 Single-Hop Configuration

To demonstrate feasibility in the first phase of the project, the rectangular test apparatus was utilized to develop a proof-of-concept system. This system utilizes a single pair of the coaxially mounted transducers on the rectangular test apparatus. Again, the block allows us to mimic the wall of a typical pressure vessel. This configuration allows the experimentation with a one-way trans-wall—i.e., “inside” to “outside”—communication link. In this configuration, the sensor measurements are made and conveyed externally by a carrier signal *originating* internal to the vessel. For example, the signal flow can be viewed in Figure 19 below. A sensor provides some arbitrary signal on the inside of the vessel to be conveyed. This signal must be encoded into a usable digital format and thereafter, this information is used to modulate the carrier originating in the vessel.

Figure 19 - Rudimentary single-hop configuration.

In order to minimize the complexity of any electronics contained internal to the vessel, the progression of development sought to deviate from the aforementioned “single-hop” configuration. Generating a stable carrier in a high temperature environment is a near impossibility. In the rudimentary system, generation and modulation are all accomplished by a data acquisition (DAQ) system prior to the signal ever reaching the transmit UT. While the DAQ provides great flexibility in terms of generating very accurate arbitrary signals on the fly, this type of signal generation—especially at high temperatures—is not feasible. Moreover, the power requirements to transmit a trans-wall signal are too large to be considered achievable within the vessel. While the rudimentary system serves as an excellent proof of concept, it is clearly limited in its practicality. External to the vessel there are much fewer limitations on the device size and power in use.

Therefore, any system where the carrier signal originates on the outside of the vessel is highly attractive. Moving the carrier generation process external to the vessel undoubtedly minimizes complexity on the inside. While there is always a desire to strive for efficiency, the power of the signal can more readily be increased externally if communication needs warrant it. If a signal is sent into the vessel, then data must be conveyed on it for its return-trip back out. Encoding and modulation must take place regardless of the characteristics of the system in use. Therefore, this single modification to the system provides a more realizable scheme for carrier generation in the controllable environment external to the vessel.

4.2 Double-Hop Configuration

During the second phase of research a “double-hop” configuration was developed which encompasses these techniques. Using both continuous wave (CW) and pulse signaling techniques, a system was successfully developed to communicate data from inside of the vessel to a data logger on the outside. In this configuration, carrier generation does not originate on the inside of the vessel as it did in the single-hop configuration earlier. In the double-hop configuration, four transducers are affixed to the steel wall: two are mated coaxially to convey signals into the vessel and two are mated coaxially to convey signals outward. The generated carrier signal is transmitted across the wall into the vessel, modulated based on the sensor reading, and retransmitted to the outside of the vessel using another transducer pair.

Once received on the outside, the sensor data is recovered and displayed to the user. Hence, signals are simultaneously transmitted across the channel in both directions (i.e., full-duplex mode). Transmitting and receiving simultaneously through the same channel brings with it inherent signal interference issues. Two techniques are used to mitigate these effects: in the CW mode, adaptive suppression techniques are utilized while in the pulse mode the pulse repetition rate is kept low enough to allow interfering echoes to fade.

Figure 20 - Overview of the double-hop system.

Figure 20 above shows a block diagram of the double-hop system concept. Herein, the external transmitting transducer will be referred to as the *source* transducer, its receiving mate the *internal receiving* transducer, the next the *internal transmitting* or *retransmitting* transducer, and the final one the *external receiving* or *sink* transducer. Due to power losses from having a second acoustic (return) path and inefficiencies in the system for modulating the signal (signal variation), more signal power is required than in the single-hop system. The double-hop path losses result in a sacrifice of power efficiency. However, the important feature of this configuration is that the signal originates from, and its power requirements are, external to the vessel.

Double-hop communication systems that operated in pulse and CW modes were successfully implemented and demonstrated. The double-hop topology can be seen in the pictures of the overall view (Figure 6), the aerial view of the internal transducers (Figure 7), and the aerial view of the external transducers (Figure 8) of the vessel double-hop configuration earlier. Of special importance is the custom-machined appliance to allow the non-permanent mating of the internal transducers to the vessel surface in Figure 7 on page 28. The spring-loaded pressure against each transition piece allows fast reconfiguration of transducer mounting positions.

4.3 Reflected-Power Configuration

In another configuration, a hybrid single-hop system was developed that can communicate using a single pair of UT's. In contrast to the rudimentary single-hop system presented in Figure 19 above, carrier generation originates external to the vessel in this reflected-power configuration. Using a continuous analog pulse train, the system reflects power back to source transducer to as a means of modulation. This is accomplished by changing the acoustic impedance of the internal UT by way of shorting its terminals. From transmission line theory, the amount of transmitted energy (or power) that is transferred to the receiving load from a transmission line is maximized when the line and load are of equal impedances or, in other words, the impedances are matched.

This concept holds true for both electromagnetic and acoustic waves, alike [53,54]. In the acoustic case, we specifically refer to the acoustic impedance of a medium or object. Conversely, if the impedances of the transmission line and loads are not matched, some energy (power) is reflected back towards the source. The fraction of the signal energy that is reflected from the load back to the source is a function of how well matched the acoustic impedances are for the medium-transducer interface.

Figure 21 - Reflected-power configuration.

The reflected-power configuration utilizes a technique by which data is communicated through a steel wall by intentionally varying the acoustic impedance at the internal transducer-wall interface in order to alter the power reflected back from an externally applied source. This acoustic impedance is changed in response to the data that is to be transmitted from the inside to the outside of the wall. In the demonstration of this technique, the modulated reflections come from an ultrasonic transducer that is mounted on the inside wall and whose acoustic impedance is varied by changing its electrical load (i.e., through shorting). The result is an appreciable change in the received signal at the originating UT after each pulse is emitted.

This reflected-power communication system uses a single pair of transducers, one mounted on the inside and a second on the outside of the wall or vessel. The configuration overview is shown in Figure 21. The “outside” portion of the system alternates between a transmit mode, in which a pulse is sent into the wall and a receive mode, in which the reflected pulse is detected. The “inside” portion of the system either

shorts or opens the electrical terminals on the internal transducer, depending on whether a binary '0' or '1' is being sent, thereby changing the size of the reflected pulse (see Figure 22) in response to the data. To reduce complexity and to enable high-speed data communication, this switching operation is implemented as a simple diode switch as described later.

Figure 22 - Shorting of the receiving transducer using a mechanical switch.

4.4 Implementation

The three aforementioned generalized configurations were implemented successfully as demonstration systems. Each system was affixed to both the original steel test block and to the provided pressure vessel and are generally effective in their use. Detailed explanations and system overviews are provided in the subsequent sections.

5. System Process and Signaling Overview

Each of the aforementioned transducer configuration schemes were successfully developed into demonstration systems. The demonstration systems are implemented around a National Instruments NI PCI-6115 data acquisition card that is operated and programmatically controlled using the LabVIEW software development suite. These components are installed on a Dell Optiplex GX280 personal computer, with the collection herein referred to as the “DAQ system.” The DAQ system affords a great deal of flexibility in terms of generating arbitrary test signals. Coupled with the LabVIEW program, the PCI-6115 card can serve as an arbitrary function generator whose output can be updated on the fly. Consequently, many types of signals can be generated, including modulated pulses and CW.

The NI PCI-6115 DAQ board is an expensive purchase so a single DAQ setup is used for the demonstration systems. Consequently, one PCI-6115 implements both the “inside” and “outside” signal processing. That is, for demonstration purposes the available functionality of the card is utilized to also modulate signals internal to the vessel as standalone circuitry would in a true installation. While this is acceptable in a demonstration system, a real system would clearly require separate processing for each side of the channel. A single DAQ card offers plenty of I/O ports for use. Consequently, it enables the computer to measure parameters to be conveyed, encode to transmit, and receive to decode using the single DAQ system. It is emphasized, however, that no synchronization or communication of data occurs within software subsystems. All data is solely conveyed acoustically through the channel using ultrasound.

Furthermore, both the NI BNC-2110 DAQ I/O breakout box and the UT contact transducers are outfitted with BNC connectors with interface impedances of 50Ω , so they can be easily interfaced using standard coaxial cable connectors. Consequently, in most instances when the DAQ system needed to interface with a UT, a direct connection was made. An image of the I/O breakout box can be found below in Figure 23.⁸ An

⁸ Image courtesy National Instruments (<http://www.ni.com/>).

enlarged technical drawing of the available ports on this device has been reproduced from its datasheet in Appendix E.1 [69].

Figure 23 - National Instruments DAQ I/O breakout box.

5.1 General System Process

Figure 24 is a block diagram that illustrates the general operation concept of the system. This figure demonstrates the typical software-hardware and data processing progression. Through the DAQ system, a voltage is measured and displayed on the computer screen. The purpose of this voltage is not only to provide a tangible example of how information is conveyed across the channel but also to mimic a typical transducer voltage whilst measuring temperature, flow rates, etc. In each of the demonstration systems, the user has the option of choosing the acquisition of an actual dynamic voltage to be measured externally or a constant “virtual” voltage generated within software. This voltage (in decimal form) is then converted to a binary representation that is as accurate as the acquisition (i.e., 12 bits for the NI-6115 card). This binary representation is also displayed on the monitor for the user to see.

The binary word is then encoded using the method of choice. An algorithm is implemented to monitor whether the channel is ready to have more data conveyed through it; if it is, the data is transmitted. Again, note that there is no cross-channel interaction internal to the computer; it merely serves as an acquisition and display unit.

That is, the only knowledge of the information being conveyed on the receive side is that which passes acoustically through the steel wall physical barrier. On the opposite side of the steel wall, the receive algorithm is continuously monitoring the UT transducer for incoming data. If present, the DAQ captures the full transmission so that it may be analyzed in software. Once captured, the data is decoded using a demodulation scheme appropriate for the signaling method used. The decoded data is converted from its binary representation to the original decimal representation; both are sent to the computer display. By having this data presented on the display, the user can visually verify that the transmitted and recovered data are in fact the same.

Due to limitations of the NI-6115 DAQ card and the computer operating system, it was found to be difficult to implement a continuous transmit and receive function utilizing circular buffers at the acquisition speeds necessary. This specific problem stemmed from the necessary generation of a carrier signal at a frequency which challenged the limits of the card (i.e., 1 MHz). Continuous processing would be possible with a true physical implementation utilizing a digital signal processor, for example. Instead, triggered acquisitions are utilized to ensure synchronous and stable hardware data acquisitions. Therefore, coded bursts of binary words are created and transmitted followed by the triggered acquisition—and subsequent processing—of the entire word by the receive function. Note that there is no loss of generality in the signal processing techniques by doing this. A brief pause is injected into the process after the signal recovery to allow the channel to come to rest. The purpose of this delay is two-fold. From fundamental knowledge of acoustic transmissions, it is evident that there will be a multitude of reflections that need to die down within the steel block (i.e., multipath). Additionally, this software pause allows the system buffers to be adequately purged to prevent data overflow. After this pause, the process continues until the user requests it to stop. Note that while there are certain nuances that are unique to the signaling paradigm utilized, this general global process model can be applied in each instance.

Figure 24 - Overall system process concept.

5.2 Signaling Methods

The NI PCI-6115 DAQ can output one analog signal at a maximum sampling rate of 4 MS/s or two analog signals with a maximum sampling rate of 2.5 MS/s. On the acquisition side, the DAQ can receive up to 4 analog signals each with a sampling rate of up to 10 MHz. The sampling rate for the analog output limits the highest signal frequency that can be generated to approximately 2 MHz. Within this limit, however, the DAQ provides considerable flexibility in the type of signal that is generated. There are two major choices to be made regarding the type of signaling to be used in each configuration. The first choice is between continuous wave (CW) and pulsed signaling—due to the resonant nature of each transducer—and the second choice is between analog and digital modulation. Both choices impact system performance and implementation complexity.

5.2.1 Continuous Wave vs. Pulsed Signaling

Continuous wave (CW) communication signals convey information by modulating a sinusoidal carrier. The information can be placed on this carrier by modifying the carrier amplitude, frequency, phase, or some combination of these three. Continuous wave carriers offer the benefits of simple generation and low peak-to-average power ratio. In addition, since the carrier can have constant frequency it can be set at the transducer's center frequency for maximum throughput. CW signaling is often used for ultrasonic communication utilizing fixed transducers.

Pulsed signaling generally utilizes short-duration, large amplitude pulses to convey information. Pulsed signals are useful for transducers if you need to have a signal propagate for a long distance, overcome attenuation within a channel, or characterize deflections in a material. While most UT transducers are designed to handle large-voltage ($\sim 100 V_{pp}$) pulses, the instantaneous power demands for the transmitter are very high. As a result, these types of signals need some type of amplification between the DAQ and the transducers if generated by the DAQ system. As a substitute for pulse generation by the DAQ system, the Krautkramer-Branson USIP-12 pulser made avail-

able for the project could be utilized. Information is sent using pulses by modifying their amplitude, duration and / or position.

5.2.2 Analog vs. Digital Modulation

In analog modulation, the “continuously-variable” analog signal—i.e., the measured value—proportionally modulates the CW carrier or pulse. For instance, in the case of amplitude modulation (AM), the amplitude of the CW carrier or pulse is determined by the analog value that is being transmitted. In a noiseless environment, analog modulation can theoretically convey values with infinite precision. In a practical system, however, the presence of noise limits the signal-to-noise ratio (SNR) of the demodulated signal. Figure 25 shows a block diagram of a system for analog modulation in which a sensor voltage is placed on a CW or pulsed carrier and fed to the ultrasonic transducer.

Figure 25 - Typical analog modulation configuration.

With digital modulation, the measured value is first sampled and quantized by an analog-to-digital converter (ADC). With amplitude modulation in this case, the amplitude of the CW carrier or the pulse would now have a finite set of possible values; the size of the set being determined by the resolution of the ADC. The demodulator decides which of these values was sent. In a well-designed system, it is possible to make the probability of making an error in estimating which quantization level was sent by the transmitter as small as desired, though error-control coding may be required. In such a

system, the SNR of the received, demodulated signal is determined by the quantization noise introduced by the ADC. That is, the introduction of noise into the system is mainly due to quantization error. Figure 26 is the block diagram of a system for binary digital modulation. An ADC processes the sensor voltage and converts it into a serial stream of binary data that are, in turn, fed to the modulator.

Figure 26 - Typical binary digital modulation configuration.

5.2.3 Binary Phase-Shift Keying (BPSK)

Digital CW phase modulation and specifically, binary phase-shift keying (BPSK), was selected for use in the configurations where pulse modulation is not utilized. In general, analog modulation tends to be more susceptible to noise and other impairments. Analog amplitude modulation also has the disadvantage of having sensitivity to variations in acoustic attenuation, which are indistinguishable from variations in

the sensor voltage being transmitted. Analog phase and frequency modulation schemes are more difficult to generate, particularly when high-precision is desired. Simple, low-complexity frequency and phase modulators tend to have linearity problems that distort the transmitted message (i.e., the sensor data). Digital modulation is generally less sensitive to noise, since the demodulator only needs to decide between a finite set of levels (rather than an infinite continuum), but introduces the added complexity of an ADC [70].

In the configurations where it is utilized, CW modulation is often preferable over pulse modulation for several reasons. First, CW modulation readily allows the adjustment of the average amount of transmit power associated with each bit of information by varying the duration of the transmit symbols in addition to adjusting the input power. Secondly, interference suppression techniques can be applied to CW modulation to mitigate the effects of the multipath propagation introduced by the channel as described earlier. Interference suppression was not used in the initial proof-of-concept phase (i.e., single-hop), but it proved invaluable for later double-hop configurations. Finally, CW modulation, with its lower peak-to-average power ratio, makes it possible to deliver higher average power. One future enhancement to the communication system might be the continuous delivery of electrical power to the internal sensors through ultrasound. Utilizing rectification and filtering, this will be made easier through the use of CW signals over trying to harness the power transmitted by pulses.

In the general case of M -ary PSK, the phase of the CW carrier is shifted among M unique levels to send information. Using this methodology, in BPSK 0's and 1's are mapped into phase shifts of 0° and 180° , respectively. This provides a maximum relative phase shift of 180° . Figure 27 below shows a BPSK signal at the top and the binary data stream at the bottom; notice the phase reversal at the $0 \rightarrow 1$ and $1 \rightarrow 0$ transitions. Figure 28 is the block diagram of a BPSK modulator in which an ADC samples and quantizes the sensor voltage, producing a serial binary stream of 1's and 0's which are, in turn, converted into $+1$'s and -1 's and multiplied by the sinusoidal carrier.

Figure 27 - A typical binary phase-shift keying (BPSK) signal.

Figure 28 - BPSK modulation configuration.

Note from the figure that the BPSK signal can be expressed mathematically as:

$$\text{BPSK} = b_k \cos(2\pi f_c t + \phi_c), \quad b_k = \{-1, +1\}. \quad (5.2.1)$$

Since the negation of a sinusoid is equivalent to the phase reversal of the phasor, (5.2.1) can be equivalently written as:

$$\text{BPSK} = \begin{cases} \cos(2\pi f_c t + 180^\circ + \phi_c), & b_k = -1 \\ \cos(2\pi f_c t + \phi_c), & b_k = +1 \end{cases} \quad (5.2.2)$$

where b_k are the data bits (after conversion to +1's and -1's), f_c is the carrier frequency and ϕ_c is an arbitrary phase shift of the transmitter. At the receiver side, another arbi-

trary phase offset is introduced due to random phase shifts within the channel. Therefore, (5.2.2) can be updated to reflect the received signal as:

$$\text{BPSK} = \begin{cases} \cos(2\pi f_c t + 180^\circ + \phi_c + \phi_n), & b_k = -1 \\ \cos(2\pi f_c t + \phi_c + \phi_n), & b_k = +1 \end{cases} \quad (5.2.3)$$

where ϕ_n is the phase shift introduced by the channel. Note that the multipath phenomenon within the channel exacerbates this noise and it is ultimately introduced in the demodulation of the signals of interest. In addition, this method requires synchronous demodulation which adds undue complexity to the receiver. Synchronous demodulation is also susceptible to phase variations within the channel. For these reasons, differential encoding is more attractive and is robust in the presence of phase noise [71]. Finally, note that the 180° phase shift is introduced by multiplying the carrier by -1 , making the modulator easy to implement using a standard mixer.

5.2.4 Differential Binary Phase-Shift Keying (DBPSK)

While the BPSK modulator is easy to implement, the demodulator can be more difficult. A synchronous BPSK demodulator is shown in Figure 29 below. It is assumed that an additional phase shift has been added to the BPSK signal, accounting for the ϕ_n phase offset term. This demodulator requires a local carrier reference that is locked in both phase and frequency to the carrier for the BPSK signal. The received BPSK signal is multiplied by the local carrier reference and lowpass filtered to remove the component at $2f_c$ —a frequency component that arises from the sine squared law.⁹ A threshold device is then used to decide whether it is more likely that a $+1$ or -1 was transmitted.

⁹ Recall that $\cos^2 u = \frac{1 + \cos(2u)}{2}$.

Figure 29 - Synchronous BPSK demodulator configuration.

The need for a coherent (phase and frequency locked) local reference makes the demodulator complex. A phase-locked loop (PLL) is commonly used for this purpose. A way to simplify the demodulator is to use differential BPSK (DBPSK), sometimes called simply DPSK, in place of BPSK [472]. In this case, information is encoded in the phase *transitions* of the modulated signal rather than the absolute phase values. That is, the data is conveyed in the *change in phase* that is transmitted in the carrier rather than simply changing the phase of the carrier when a symbol is transmitted. For instance, consider that the phase of the carrier at any given instant is denoted as θ_i . The carrier phase at the instant before that can then be denoted as θ_{i-1} . Therefore, the change in phase can be denoted as:¹⁰

$$\Delta\theta_i = \theta_i - \theta_{i-1} \quad (5.2.4)$$

Rather than mapping 0's and 1's to a particular phase shift, they are now mapped to a particular *change* in phase shift of the carrier. Consider, for example, the following mapping of 0's and 1's denoted as b to the change in phase shift $\Delta\theta_i$:

¹⁰ Discussion adapted from: <http://www.ee.byu.edu/ee/class/ee444/newlectures/dbpsknotes.pdf>.

b	$\Delta\theta_i$
0	π
1	0

Table 5 - Mapping of the change in phase shift to binary bits.

Note that, in general, an assumption must be made for the initial bit. That is, since the carrier phase for each bit interval is related to the phase during the interval before it, some arbitrary value must be assumed for the first bit. To maintain the same level of bit precision, the binary word is simply padded. Thereafter the difference in the carrier phase shift is governed by (5.2.4), thus it follows that the phase of the carrier at any given instant is:

$$\theta_i = \theta_{i-1} + \Delta\theta_i. \quad (5.2.5)$$

Consider the following sample binary sequence:

$$1\ 0\ 0\ 1\ 1\ 1\ 0\ 0\ 0\ 1\ 0\ 1\ 0, \quad (5.2.6)$$

the carrier phase shift, θ_i , is then tabulated as:

i	b_i	$\Delta\theta_i$	$\theta_i = \theta_{i-1} + \Delta\theta_i$
0	1	0	0
1	0	π	π
2	0	π	0
3	1	0	0
4	1	0	0
5	1	0	0
6	0	π	π
7	0	π	0
8	0	π	π
9	1	0	π
10	0	π	0
11	1	0	0
12	0	π	π

Table 6 - DBPSK encoding example.

Assuming that the bit interval equals one cycle of the carrier signal and imposing the phase shifts for the DBPSK signal on the carrier, you obtain the upper plot shown in Figure 30. The lower plot in this figure shows the BPSK signal for the same data sequence. Observe that while the two signals look very similar they convey the same binary word differently (i.e., in bit transitions versus value).

Figure 30 - Carrier signals of DBPSK versus BPSK.

One way to flip the phase of a sinusoid is to multiply the signal by -1. Therefore, by mapping our 0's to -1 and 1's to +1, we can simply multiply the carrier by the ± 1 symbols generated by a differential encoder as mentioned earlier. Now the traditional binary bits (i.e., 0's and 1's), represented earlier as b_i , are translated to -1's and +1's and denoted as b_k . An encoder able to differentially encode these symbols is easily implemented as a one bit duration time-delayed multiplication as shown in Figure 31.

Figure 31 - Differential encoder as a time-delayed multiplication.

Figure 32 is a block diagram of a differential detector for recovering the binary information bits from the DBPSK signal. The received signal is multiplied by a version of itself that is delayed by one symbol interval to reproduce the baseband data. Note that when this baseband data modulates a sinusoidal carrier, performing this operation on the received signal will generate carrier- plus double- frequency terms as mentioned earlier. Hence, the output of this decoder is lowpass filtered with a frequency cutoff low enough to recover the baseband data and reject these terms. The output thereafter is typically referred to as the decision variable. A subsequent threshold device, generally comparing this decision variable to a threshold of zero, determines the most likely transmitted bit as discussed for the BPSK method earlier.

Figure 32 - Differential decoder as a time-delayed multiplication.

To illustrate the operation of the differential encoder and differential detector, consider the earlier binary pattern (b_i) given as:

$$1\ 0\ 0\ 1\ 1\ 1\ 0\ 0\ 0\ 1\ 0\ 1\ 0. \quad (5.2.7)$$

This time, mapping all 1's $\rightarrow +1$ and all 0's $\rightarrow -1$, the pattern becomes:

$$+1\ -1\ -1\ +1\ +1\ +1\ -1\ -1\ -1\ +1\ -1\ +1\ -1. \quad (5.2.8)$$

This string is the serial input b_k to the differential encoder above. Now assume its output d_i is the input to the differential decoder shown. The input to the encoder, output from the decoder, and intermediate steps are shown in Table 7. Observe that in the two shaded columns, the input bits and the detected output bits are equal.

i	b_i	ENCODER			DECODER			b_i
		b_k	$d_{(i-1)}$	d_i	d_i	$d_{(i-1)}$	b_k	
0	1	+1		+1	+1		+1	1
1	0	-1	+1	-1	-1	+1	-1	0
2	0	-1	-1	+1	+1	-1	-1	0
3	1	+1	+1	+1	+1	+1	+1	1
4	1	+1	+1	+1	+1	+1	+1	1
5	1	+1	+1	+1	+1	+1	+1	1
6	0	-1	+1	-1	-1	+1	-1	0
7	0	-1	-1	+1	+1	-1	-1	0
8	0	-1	+1	-1	-1	+1	-1	0
9	1	+1	-1	-1	-1	-1	+1	1
10	0	-1	-1	+1	+1	-1	-1	0
11	1	+1	+1	+1	+1	+1	+1	1
12	0	-1	+1	-1	-1	+1	-1	0

Table 7 - DBPSK example with differential detection.

With the exception of the pulsed carrier implementations, the demonstration systems use DBPSK to demonstrate the ability to communicate effectively through the steel channels. Both the modulator and demodulator are implemented with the NI PCI-6115 DAQ using LabVIEW VI's.

6. Implementation and Results

This section provides an overview of the implementation of the aforementioned system configurations. Each configuration was successfully implemented and demonstrated using the DAQ system with appropriate ultrasonic transducers on the rectangular test apparatus and pressure vessel. Each configuration has its advantages and disadvantages as discussed earlier. Firstly, the rudimentary single-hop configuration is meant to serve solely as a proof-of-concept system to demonstrate the achievability of acoustic communication utilizing ultrasound. It did, however, enable the early development of many of the programmatic algorithms (e.g., DBPSK encode / decode) necessary for subsequent installations. The double-hop and reflected-power configurations more closely emulate a typical end-goal implementation for a pressure vessel.

The double-hop configuration was first developed and implemented utilizing a pulsed carrier. Utilizing a mixer, pulses arriving at the internal receiving transducer are gated and passed to the internal retransmitting transducer to convey binary data across the channel. Of course, pulsed carrier techniques are unattractive because of their instantaneous power demands as noted earlier. However, over the single-hop configuration the double-hop configuration is attractive in that it offers the potential for power delivery when utilized with a CW carrier. A precursory demonstration was developed to exhibit this potential. Subsequently, a full communication system was developed for the CW double-hop configuration. Due to the multipath interference within the channel, full-duplex operation can not be achieved without the use of adaptive filtering. In addition, when utilizing adaptive interference suppression algorithms in this configuration with CW mode, a ten-fold increase in the data rate to 5 kbps from the early limit of 500 bps in the single-hop configuration is achieved. This result is achievable even when the transducer pairs were in close proximity to each other (~10 cm separation).

Finally, the reflected-power configuration was attempted and implemented successfully. The clear benefit to utilizing this technique is the reduction in the number of required transducers from two pairs to one. However, this technique requires a sizeable signal to be transmitted into the channel for it to work effectively. Therefore, a pulsed

carrier scheme was successful whereas a CW scheme was not. An overview and results for this and the aforementioned configurations are discussed herein.

6.1 Single-Hop Configuration

As noted earlier, because of the available I/O processing on the DAQ system, the required circuitry internal to the vessel can also be emulated by it. To reiterate, the only interaction between the “inside” and the “outside” is acoustically with an ultrasonic link. The goal of the single-hop configuration was to develop a proof-of-concept prototype. Not only does the prototype demonstrate the achievability of ultrasonic communication, it allows for the early development of many of the techniques utilized in later configurations. An overview of the interface of the DAQ subsystem to the single-hop configuration can be seen in Figure 33 below.

Figure 33 - Overview of the single-hop configuration and the DAQ interface.

The core system process flow in this and subsequent configurations was presented earlier in Figure 24 on page 60. The main challenge in this initial configuration was to first develop the DBPSK algorithms.

6.1.1 System Configuration

DBPSK Signal Generation

As noted earlier, the DAQ has a maximum output sampling rate of 4 MS/s when only one output is utilized. To interface with the 1 MHz transducers, we need to produce a DBPSK signal with a carrier frequency of 1 MHz. With the 4 MS/s sampling rate, the maximum number of points that can be generated in one cycle of the 1.0 MHz carrier is:

$$\left(4 \frac{MS}{s}\right) \times T_{carrier} = \left(4 \frac{MS}{s}\right) \times \left(1 \frac{\mu s}{cycle}\right) = \boxed{4 \frac{samples}{cycle}}. \quad (6.1.1)$$

Although the first of the four samples in a cycle can be placed anywhere in the first quarter of the cycle, the sampling arrangement shown in Figure 34 was selected. The actual signal is approximated better by a triangle wave, but the transmitting transducer will filter the signal to suppress the harmonics and produce the required sinusoid. Note that the difference between the left and right signals is a phase shift of 180° , i.e. one represents the carrier with a 0° phase and the other with a 180° phase. Either of these two sections (cycles) of the carrier can be concatenated arbitrarily to construct a sinusoid composed of 0° or 180° phase shifts.

Figure 34 - Sampled discrete-time sinusoid transmitted by the DAQ.

For this initial evaluation configuration, adaptive interference suppression is not used to mitigate the effects of reflections within the block which introduce multipath distortion into the received signal. Instead, the bit duration is set sufficiently large to allow the reflections to “settle” (i.e., a standing wave situation develops before a bit decision is made at the receiver). Based on the observed duration of the reflections, a bit duration of 2 ms is appropriate. With a 1 MHz carrier, this interval corresponds to 2000 cycles of the carrier. If there are 4 points per cycle, the number of points that must be sampled in each transmitted bit is therefore:

$$p_{transmit} = (2000 \text{ cycles}) \times \left(4 \frac{pts}{cycle} \right) = 8000 \text{ pts} . \quad (6.1.2)$$

It then follows that the transmission bit rate is:

$$f_{bit} = \frac{1}{T_{bit}} = \frac{1}{2 \text{ ms}} = \boxed{500 \text{ bps}} \quad (6.1.3)$$

The LabVIEW virtual instrument (VI) performs the differential encoding of the information bits to be transmitted and subsequently instructs the DAQ to output 8000 samples of the 1 MHz carrier with the appropriate phase for each encoded data bit.

DBPSK Signal Reception and Detection

The signal from the receive transducer is sampled at the DAQ’s maximum input sampling rate of 10 MS/s to avoid aliasing. To implement the differential detector, there must be a delay of one bit interval, i.e. 2 ms, which corresponds to

$$p_{receive} = \left(\frac{10 \frac{MS}{s}}{4 \frac{MS}{s}} \right) \times (8000 \text{ pts}) = 20000 \text{ pts} \quad (6.1.4)$$

Since the carrier is a sinusoid and it is multiplied by a delayed copy of itself in the decoding process, a signal at twice the carrier frequency will be produced along with the desired DC signal (i.e., binary word). Hence, in this case the transmitted carrier of 1.0

MHz will be multiplied to obtain a signal at 2.0 MHz.¹¹ Once the received signal is captured and this delayed multiplication occurs, it is necessary to implement a lowpass filter at a relatively low frequency to obtain the original binary data back. It is important to discard the frequency component at 2.0 MHz. To minimize noise and optimize performance, the bandwidth of this lowpass filter should be as small as possible while passing the recovered baseband data stream. This is done with a 3rd order Butterworth filter implemented in the LabVIEW VI with a bandwidth of 5 kHz.

6.1.2 Experimental Results

The first tests established that the DAQ was capable of arbitrarily flipping the phase of a sinusoid on the fly. The success in doing so is visible in Figure 35, which shows the waveform observed at the receive transducer when the waveform applied to the transmit transducer undergoes a phase reversal. Note the 180° phase reversal at the location of the arrow. The variations in the envelope of the carrier after the phase reversal are the result of the multipath reflections in the block, since the phase change initiates a new transient within the channel that takes time to acquiesce.

The next step was to send 12 bit data words through the steel block. Words of 12 bits in length were chosen as this is the resolution of the DAQ. A sequence of 12 bits is differentially encoded, the encoded bits modulate the 1 MHz carrier, and the result is fed to the transmit transducer. Carrier generation and modulation occur in software prior to being conveyed to the buffer of the DAC of the DAQ. The transmitted carrier has a peak-to-peak voltage of 20 volts. Figure 36 shows the signal obtained at the receive transducer. It is clear that the phase reversals in the carrier cause large transients, which are to be expected because of the nature of the channel. However, the envelope becomes relatively constant during the last one-half to one-third of the data bit and this region is used to make data bit decisions.

¹¹ This is due to the sine-squared law noted earlier. Reference the earlier Signaling Methods section for more information on implementing DBPSK encoding and decoding.

Figure 35 - Phase reversal in sinusoidal carrier at an arbitrary time.

Figure 36 - Envelope of received single-hop DBPSK signal.

Figure 37 shows a typical signal recovered by the multiplication of the received signal by a by the same signal delayed by one symbol duration from the differential decoder. Note that the envelope of this signal profiles the binary data that is transmitted, namely 1 1 0 1 1 0 1 1 0 0 0 0. A fast Fourier transform (FFT) of this signal can be seen in Figure 38, which shows the spectrum of the data near DC, a significant component near 1 MHz and a larger component near 2 MHz. The 2 MHz signal is the double frequency term that is generated by the multiplication and the 1 MHz component is leakage of the received signal through the detector due to the presence of some DC component in that signal.

Figure 37 - DBPSK signal after differential decoder.

After passing this signal through the 5 kHz low pass filter, a result similar to that shown in Figure 39 is obtained. This result is for the case when the binary word 1 0 0 0 1 0 1 0 1 1 0 0 is sent. To extract the data bit decisions, this signal now needs to be sampled and compared to a threshold of zero volts. This current algorithm performs a crude timing recovery operation to set the sampling points. That is, the VI looks for the input signal to surpass a designated level and uses this event as a trigger to initiate the sampling of the signal at points thereafter equivalent to one bit length apart (i.e., 20000 points). Sample points above zero are assigned a data bit of 1 and those below zero are

assigned a data bit of 0. Figure 40 demonstrates where the demodulated signal might be sampled. The recovered binary words are finally sent through a DAC to ascertain the originally transmitted voltage.

Figure 38 - FFT of DBPSK signal after differential decoder.

Figure 39 - Recovered word from DBPSK signal.

Figure 40 - Recovered word from DBPSK signal with typical sample points noted.

6.1.2.1 Temperature Measurement System

A test configuration was developed to enable a demonstration whereby actual temperature measurements could be transmitted through the rectangular test apparatus using ultrasound. Figure 41 below shows the process overview of the measurement system. A thermistor in a bridge configuration is placed in a beaker of water which can be heated using a hot plate. The voltage from the bridge is sampled using one of the ADC inputs on the DAQ and converted into a temperature value using the known temperature versus voltage characteristic curve of the measurement system. A binary representation of the voltage is fed to the transmit transducer using DBPSK encoding as mentioned earlier and received on the other side of the block where the temperature value is detected and displayed.

Although a single DAQ is used to implement both “sides” of the system, separate VI’s are used and the only communication path between them is acoustically through the steel block. Figure 42 shows the VI for the “inside” which allows the user to transmit a

dial-in voltage or the actual instrumented temperature measurement. Likewise, Figure 43 shows the VI for the “outside” that reports the data received through the ultrasonic communication link. Again, the overall operation of the system is as shown earlier in Figure 24 on page 60.

1. Voltage across thermistor acquired by DAQ
2. Binary representation transmitted by DAQ (using DBPSK encoder)
3. Signal received by DAQ and decoded (using DBPSK decoder)
4. Voltage reconstructed from binary word
5. Temperature vs. Voltage curve checked
6. Voltage and Temperature displayed on computer

1. Select actual (acquired) or simulated (user defined) voltage
2. User defined voltage entered here
3. Voltage selected in (1) is displayed
4. Binary representation of (3) is shown as LEDs
5. User can adjust update delay
6. Statistics
 - a. Bit rate (Set at 500 BPS)
 - b. Update delay (s)
 - c. Update rate (Hz)

1. Reconstructed voltage displayed
2. Approximate temperature (°F)
3. Approximate temperature (°C)
4. Binary representation of (1) is shown as LEDs
5. LED will light if signal lost
 - a. Cable disconnected
 - b. Very high attenuation
6. Statistics
 - a. Received signal amplitude (V)

Figure 43 - VI for the "outside" of the single-hop configuration.

6.2 Double-Hop Configuration

The double-hop configuration is implemented as described earlier in the general overview of Section 4.2 earlier. Again, the double-hop configuration utilizes two pairs of UT's affixed to the medium of interest. A carrier is transmitted from outside of the vessel into it, modulated, and subsequently retransmitted across the medium to the outside where it is demodulated. The double-hop configuration is implemented in two different carrier modes: pulsed and CW. The implementation according to the carrier type utilized is discussed aptly below.

6.3 Pulsed Double-Hop Configuration

In the pulsed double-hop mode a technique called Pulse Amplitude Modulation (PAM) is used to vary the signal retransmitted externally from the vessel in response to the sensor data. Generally speaking, PAM involves varying the amplitude of a pulse to convey a message signal. In an analog PAM system, the pulse amplitude can take on a continuum of values, where the pulse amplitude is a linear function of the message

signal amplitude at the instant that the pulse is sent. In a digital PAM system, generally called an Amplitude-Shifted Keying (ASK) system, the amplitude of the pulses are taken from a finite set, where a particular amplitude corresponds to one or more bits of message data. In the double-hop system however, binary information is sent using two pulse amplitudes: zero to send a binary ‘0’ and non-zero to send a binary ‘1.’ This special variety of ASK is fittingly named On-Off Keying (OOK). An OOK signal is detected by simply determining the presence or absence of a pulse in the reference time intervals. Figure 44 shows the OOK signal when the 4-bit binary word “1011” is transmitted.

Figure 44 - Using On-Off Keying (OOK) for digital modulation.

Consider the system concept presented in Figure 45. In this general implementation, a pulse train is generated outside the vessel and applied to the source transducer at the lower-left. Inside of the vessel, sensor data is generated and encoded into an appropriate binary format. Pulses received by the internal receiving transducer (bottom-right) are then sent through a gating mechanism that allows pulses to pass or not pass as determined by the sensor data. That is, the sensor data is utilized to formulate a binary word for OOK as above. The gated pulses are fed to the internal transmitting transducer on the upper-right in the figure. The retransmitted, modulated pulses are received at the sink transducer at the upper left and, subsequently, demodulated to recover the sensor data.

Figure 45 - Pulsed double-hop system concept.

Pulses from the source transducer—without the modulation due to the sensor data—will also arrive at the sink transducer via an acoustic path. These are appropriately referred to as multipath echoes. However, due to the near-zero delay in the electrical segment of the retransmitting path, the path from source to sink transducers that goes through the internal transducers has the shortest delay. Consequently, the modulated pulses arrive at the sink transducer before the multipath echoes. A minimum spacing between transducer pairs, i.e. between source / internal receiving and internal transmitting / sink transducer pairs is required to enable the modulated pulse to finish before the arrival of the first reflected pulse due to multipath. In this fashion, pulses will either pass (binary ‘1’) to the sink transducer outside of the vessel—at the upper left in Figure 45—or not (binary ‘0’).

The gating function can be performed using a switch, or alternatively, the carrier signal can be multiplied by the binary data. This can be seen in Figure 46, where switched-gating is presented on the left and multiplied-gating on the right. The multiplication is generally performed using a mixer, so the latter scheme can also be referred to as “mixed” pulse gating. Multiplying by a ‘1’ yields the pulse at the output and a ‘0’ produces no voltage at the output. Again, the presence or absence of each pulse can be associated with a binary ‘1’ or ‘0,’ respectively. This is illustrated in Figure 47 below.

Figure 46 - Switched- versus multiplied- (mixed) pulse gating.

Figure 47 - Multiplying the pulse stream by binary data.

One drawback of using a pulsed technique is that a large amount of power is delivered to the originating transducer in a short period of time to produce the pulses. This is in contrast to the CW mode, where the instantaneous power demands are low while the average power delivery over time is high. However, in cases where a large CW signal is still needed to overcome attenuation through the channel, heat dissipation may become an issue for a continuous carrier. In such cases, a pulsed technique may be more attractive.

6.3.1 System Configuration

The pulsed double-hop demonstration system was implemented using the DAQ. The block diagram of the system is shown in Figure 48. Pulses are generated at regular time intervals using a Krautkramer-Branson USIP-12 Non-Destructive Testing (NDT)

pulser-receiver and fed to the source transducer. The DAQ acquires samples of the sensor data using one of its analog-to-digital converter (ADC) channels and converts these samples into digital form.

The binary data representing these sensor data samples are output to a mixer on the “inside” which either passes (binary ‘1’) or suppresses (binary ‘0’) a pulse received by the internal receiving transducer for retransmission. Alternately, software generated voltages can be used in place of actual sensor data for testing purposes as it is in Section 6.1 (Single-Hop Configuration). In this case, the user can manually adjust the “acquired” voltage that will be encoded to properly gate the pulses internal to the vessel. After the pulses are gated and retransmitted across the wall as shown in the diagram, the next step in the communication process is to acquire the pulses received by the external sink transducer. These pulses are acquired by the DAQ using a second ADC channel and signal processing occurs internal to the computer to determine if a ‘0’ or a ‘1’ is sent from inside the vessel.

Figure 48 - DAQ subsystem in overall pulsed double-hop configuration.

The signal processing procedure can be seen in Figure 49 on the following page. Note that this diagram is simplified and is meant to only demonstrate the concept. First, the received signal during the interval encompassing the defined pulse position is sampled and captured by the system. Next, the absolute value of these samples is taken and the result is fed through a low pass filter to remove high frequency components of the signal and to create a single smooth signal for further processing. This signal is then compared to a threshold to decide whether or not a pulse is present in the time interval. If the signal is above a threshold value, then it is assumed that a pulse exists and a '1' was conveyed from the inside or vice versa. This process continues for each bit in the binary word representing a single sensor reading (12 bits). As the last step in the process, the captured word undergoes a software digital-to-analog (DAC) conversion to determine the sensor voltage sample. This sensor value is decoded and displayed to the user as it was earlier. This demonstration system was successfully used on both the test block and cylindrical vessel.

Figure 49 - Received pulse signal processing for the double-hop configuration.

6.3.2 Experimental Results

Before implementing this method using the DAQ system, a simple experiment was performed using the test block to demonstrate the feasibility of the technique. The test block has three pairs of transducers mounted coaxially: two pairs centered at 1.0 MHz and the third pair at 0.5 MHz as discussed in Section 3.1 on page 19 (Rectangular Test Apparatus). A manual double-hop system, in which a jumper cable connecting the inside receiving and inside transmitting transducers is either inserted or removed, was tested using the four 1.0 MHz transducers. The results of this initial experiment using the NDT pulser-receiver can be seen in Figure 50. The signals at the sink transducer are shown on the right side of the figure for the case when the jumper cable is in place (top) and removed (bottom). Note that a significant and noticeable pulse is present in the circled region when the jumper cable is connected. The received pulses at a larger delay are due to direct acoustic coupling between the source and sink transducers. These pulses are produced by the many reflective paths (multipath) within the block.

Figure 50 - Initial test using the pulsed double-hop configuration.

For the system implemented with the DAQ, results were first obtained using the 1-inch 1.0 MHz transducers on the rectangular test block and a mixer to modulate the pulses as noted earlier. Figure 51 shows the originating pulse plotted on the same graph as the received signal. The waveform was captured from within LabVIEW. The originating pulse, seen on the left side of the plot, is the signal generated by the Krautkramer-Branson USIP-12 NDT pulser-receiver that is applied at the source UT to produce an ultrasonic carrier pulse. In the specific case of Figure 51, a binary ‘0’ is sent, hence no pulse is sent (i.e., allowed to pass) by the inside transmit UT. The only signal received, seen on the right side of the plot, is the superposition of the multiple reflected source pulses (i.e., multipath). The first of these pulses arrives at the receiving external transducer approximately 60 μ s after the pulse is applied to the source UT.

Figure 51 - Pulsed double-hop received voltage (no pulse) for the test block.

Figure 52 shows the corresponding waveform when a binary ‘1’ is mixed with the pulse after it is received internally. Rather than canceling the pulse as in the case of a binary ‘0,’ the binary ‘1’ allows the pulse to pass and be retransmitted back to the outside. This pulse is received during the interval 50-60 μ s after the pulse is applied to the source UT. Again, notice the direct acoustic multipath signal arriving shortly thereafter.

Figure 52 - Pulsed double-hop received voltage (gated pulse) for the test block.

For an applied pulse, the ringing in the channel lasts for a considerable period of time before following an exponential decay. The channel perturbation is of considerable amplitude for over 8 milliseconds (reference Figure 5 earlier). Clearly, if a second pulse were to be applied before this ringing dies out, it would be difficult or impossible to accurately detect the presence or absence of the second pulse. Hence, the pulse repetition rate and, consequently, the data communication signaling rate are limited in this configuration. This point is further demonstrated by Figure 53 and Figure 54, which show expanded views of the ungated / gated pulse transmission responses of Figure 51 and Figure 52. These two figures show more of the response just after the gated pulse position. It is evident from these plots, together with the response in Figure 5, that a significant time period must exist between successive pulses to sustain a reliable communication link.

Figure 53 - Pulsed double-hop received voltage (no pulse) for the test block, extended time.

Figure 54 - Pulsed double-hop received voltage (gated pulse) for the test block, extended time.

Figure 55 - Extracted signals and filter / threshold for the rectangular test block.

After extracting the portion of the signal that corresponds to the modulated pulse, the next steps include taking the absolute value of the signal, feeding it through a LPF,

and finally making '0' / '1' decisions by threshold. The LPF operation provides a clean pulse to use in the threshold algorithm. The results of these operations can be seen in Figure 55. The plots on the left of the diagram are of the pulse position (no pulse on top, pulse on the bottom) and the plots on the right are of the LPF outputs for each case. The decision threshold is depicted by the background shading on the figures. The results shown above were obtained using the test block with the 1.0-inch 1.0 MHz transducers.

Similar results were found when the setup was moved to the cylindrical vessel. In this case the 0.5-inch 1.0 MHz transducers were used. Recall the response of the channel to a pulse can be seen in Figure 9 earlier on page 30. The channel disturbance appears to decay more quickly in the vessel. Notice that this is due, in part, to the lack of oscillatory ringing which was seen in Figure 5 on page 26. This may be a consequence of the use of the Lexan transition pieces utilized (which have large incremental attenuation) as well as the medium of differing material and shape. The cylindrical vessel has a unique response in that two pulses are received at the sink transducer when a connection is made between the two internal transducers and a single pulse is transmitted into the vessel. The contrast can be viewed between the pulse and no-pulse cases shown in Figure 56 and Figure 57, respectively. The presence of this double pulse does not change the signal processing; the digital signal processing routines are carried out on only one of the pulses. The extracted signal and accompanying filter / threshold can be seen in Figure 58.

Figure 56 - Pulsed double-hop received voltage (no pulse) for the pressure vessel, extended view.

Figure 57 - Pulsed double-hop received voltage (with pulse) for the pressure vessel, extended view.

Figure 58 - Extracted signals and filter / threshold for the pressure vessel.

Note that the LabVIEW interfaces are very similar to those presented in Figure 42 and Figure 43 of Section 6.1.2.1 on page 80 (Temperature Measurement System). Hence, they are not presented again to avoid redundancy. Empirical results indicate that in both instances of the test block and vessel, achievable communication data rates are 500 bits per second (BPS) and below. To reiterate, larger transmission rates require

higher pulse repetition rates. This results in heightened interference due to direct acoustic coupling and subsequently, diminished reliability of the communication.

6.4 CW Double-Hop Configuration

The double-hop configuration utilizing a continuous wave (CW) carrier was also investigated. This method has the same transducer mounting topology as the pulsed case, as is shown in Figure 59. The CW sinusoidal carrier—centered at the resonant frequency of the transducers for maximum signal emission—enters the channel by the source transducer on the lower left of the figure. After its coaxial mate receives the carrier signal, the internal electronics modulate the carrier with the sensor data, in this case using Differential Binary Phase-Shift Keying (DBPSK), before sending it to the inside transmitting UT. The modulated signal is received at the external sink transducer, demodulated, and decoded.

Figure 59 - CW double-hop configuration.

As in the case of pulse modulation, reflections in the block result in a direct acoustic path between the external source and sink transducers within the channel. This situation is illustrated in Figure 60. The basic idea is that along with direct-path, line-of-

sight (LOS) propagation between coaxially mated transducers, a theoretically infinite number of time-delayed versions of the signal of varying attenuation arrive at the sink transducer due to internal reflection. In the pulse case, it was possible to avoid these multipath pulses through time gating. In the CW case, however, the carrier is applied continuously and a standing-wave arrangement results within the block. The net result is that a carrier-only, fixed-frequency signal—having an amplitude and phase determined by the multipath arrangement—is added to the modulated signal at the external sink transducer. This interfering signal can be very large relative to the modulated signal and, therefore, must be suppressed to reliably estimate the sensor data that is being conveyed from the inside to the outside of the wall. A technique known as adaptive interference suppression is utilized for this purpose.

Figure 60 - CW double-hop configuration with multipath.

6.4.1 System Configuration

Consider the system block diagram of the general filtering scheme shown in Figure 61 [2]. Note that with the DAQ system, we are working in discrete-time so that we are concerned with a signal at sample time n rather than continuous time t (i.e., $x[n]$ vs. $x(t)$). In this system, the received signal $x[n]$ is initially passed through a linear filter. After the filter operates on the signal, the result is the filtered signal output $y[n]$. The

difference between $y[n]$ and the so-called desired response, $d[n]$, is denoted as the estimation error $e[n]$. The goal is to adapt the linear filter in a way that minimizes $e[n]$.

Figure 61 - Block diagram of the general filtering scheme.

For this application, as shown in Figure 62, $d[n]$ is the output of the sink transducer consisting of the modulated signal added to an amplitude- and phase- altered copy of the un-modulated carrier, while $x[n]$ is a copy of the un-modulated carrier signal obtained directly from the carrier source. Since the channel is dynamic in nature and may deviate from one realization to the next, it is desirable to utilize an adaptive linear filter. An adaptive filter is one in which tap weight vectors are recursively updated to minimize the estimation error. Actually, the term adaptive “linear” filter may be considered a misnomer in some instances. As a result of the recursive tap weight updating, the filter is generally non-linear when looked at over its entire adaptation life. However, the filter is linear at each *realization* of the tap weight vector by its nature. Hence, the linear filter will adjust the amplitude and phase of $x[n]$ such that the subtraction process will remove the un-modulated carrier from $d[n]$ at each sample instant. This is the general linear estimation problem [73]. The error signal, $e[n]$, is the recovered modulated signal that is fed to the demodulator to recover the sensor data.

Figure 62 - Linear filtering scheme in the double-hop configuration.

6.4.1.1 Least Mean-Squared (LMS) Adaptive Filtering

Adaptive filtering is a commonly used technique in areas of signal processing where channel parameters, while known in a general way, may vary from one channel realization to the next. In the double-hop configuration we know that the interference that is added to the modulated signal is a sinusoid at the carrier frequency. However, the amplitude and phase of that sinusoid will vary from installation to installation and, also, over time for a given installation. Consequently, a fixed linear filter cannot be used to process the reference signal from the carrier source; instead, an adaptive filter is needed. The Least Mean-Squared (LMS) algorithm is a commonly used adaptive algorithm because of its effectiveness and low implementation cost function. This algorithm operates to minimize the error signal, $e[n]$, in a mean-square sense. The linear filter has the structure of a transversal filter, also commonly referred to as a tapped delay-line filter, which applies tap weights to delayed versions of the reference signal and sums the weighted values to produce the filter output.

For the double-hop system, the filter is required to adjust the amplitude and phase of a sinusoid of known frequency. This operation can be accomplished using a second order transversal filter, i.e. a filter with two taps, if the taps are separated in time by a delay equal to a 90° phase shift at the carrier frequency. Said more precisely, we seek the linear combination of orthogonal bases of the interferer that minimizes the error. This property results from the trigonometric identity given by:

$$R \cos(\omega t + \phi) = A \cos(\omega t) + B \sin(\omega t), \quad (6.4.1)$$

and the fact that:

$$\sin(\omega t) = \cos(\omega t - 90^\circ). \quad (6.4.2)$$

Therefore, it appropriately follows that:

$$R = \sqrt{A^2 + B^2}, \quad (6.4.3)$$

and

$$\phi = \tan^{-1}\left(\frac{B}{A}\right). \quad (6.4.4)$$

The coefficients A and B correspond to the filter *tap weights* which control the amplitude of the two signals that are ultimately added together after the transversal filter.

This weighted sum of sine and cosine signals will be adjusted and subtracted from the desired signal (with multipath interference) to effectively cancel the interference. The principle behind the use of this 2-by-1 or 2nd order transversal filter is that the interfering noise will be a function of both in-phase and quadrature components at the receiver [70]. In fact, while the transversal filter can generally have an arbitrary order or number of taps, it has been shown that for a wide-sense stationary (WSS) channel (i.e., one where the mean of the noise is constant and its autocorrelation is shift-invariant), the solution for the adaptive transversal filter is precisely a 2nd order solution [2,43].

Consider the general 2nd order transversal filter show in Figure 63 below. Following from left-to-right, the reference carrier signal $x[n]$ is first sent into the transversal filter. The carrier signal—multiplied by a tap weight w_1 —is then delayed by a specified amount, denoted by z^{-1} , and multiplied by a tap weight w_2 . The two results are then added together and their result is subtracted from the desired response signal $d[n]$, which consists of the modulated signal and multipath interference. The estimation error signal, $e[n]$, is the recovered modulated signal with the interference suppressed.

Figure 63 - General 2nd order transversal filter.

It follows that an adaptive algorithm is necessary to properly adjust the weight vector $\vec{w}$ such that the resulting estimation error yields only the modulated signal when given $d[n]$. The three-dimensional quadratic error surface is the surface of the estimation error output over all possibilities of the two tap weights w_1 and w_2 (or in general, the weight vector $\vec{w}$). This general quadratic surface has a global minimum at a single point where the error output is ideally zero (see Figure 64).¹² The “method of steepest descent” is a phrase commonly given to the adaptation algorithm that works to find the gradient of the error surface when it is equal to zero. Specifically, the tap weight vector for the next sample instant is given by:

$$\vec{w}(n+1) = \vec{w}(n) - \mu(n) \frac{\partial J(n)}{\partial \vec{w}(n)}. \quad (6.4.5)$$

Where $J(n)$ is the error function. Clearly, when the derivative of the error function is zero, the tap weight vector $\vec{w}$ reaches a steady-state value. Recall that in three-dimensional calculus the gradient operator (∇) computes the measure of rate of change in three dimensions; it will be zero at the minimum on the quadratic surface.

Figure 64 - Typical quadratic error surface over the weight vector $\vec{w}$.

¹² Adapted from <http://www.science.gmu.edu/~bchouikh/csi801/hwkprobl/Ex4p.html>.

Unfortunately, while the steepest descent algorithm is a great theoretical model it requires prior knowledge of the correlation and cross-correlation matrices of the error surface. The error surface in our implementation is inherently dynamic and this information is unknown. Therefore, it is preferable to use a stochastic gradient algorithm which provides a random-walk (rather than deterministic) down the error surface. One type of commonly used stochastic gradient algorithm is the LMS algorithm. The LMS algorithm uses instantaneous samples of the signals in its calculations, rather than their statistics, to form an instantaneous estimate of the mean-squared error (MSE) cost function. While the variance of these samples can be large in general, the adaptive-feedback nature of the algorithm causes the variance to decrease exponentially over time. The LMS algorithm gets its name from computing the minimal or “least” value of the squared mean of the estimation error signal $e[n]$ [74]. The algorithm does so by estimating the gradient vector of the error signal at sample n and recursively seeking a minimal value. In equation form, the MSE *cost function* is given as:

$$J(\vec{\mathbf{w}}) = \frac{1}{2} E[|e(n)|^2], \quad (6.4.6)$$

where $E[\cdot]$ is the expected value (i.e., the average in probabilistic terms) of the square of the error signal. This effectively computes the average of the *energy* (i.e., 2nd order statistic of the random variable) of the error signal at some instant in sample n . That is, the energy of a signal is proportional to the square of the voltage [63]:

$$\mathcal{E}(t) \propto V^2(t), \quad (6.4.7)$$

so the gradient of the cost function will effectively provide a measure of the rate of change of the energy of the remaining signal over the error surface. When the gradient of the cost function is minimal, i.e.:

$$\nabla(J(\vec{\mathbf{w}})) = 0, \quad (6.4.8)$$

the energy (power) of the interfering signal is minimal. Again, it is necessary to perform an instantaneous estimate of the gradient given a single sample point. This is a recursive operation which has been shown to satisfy the difference relation:

$$\vec{\mathbf{w}}[n+1] = \vec{\mathbf{w}}[n] + \mu \vec{\mathbf{u}}[n]e[n], \quad (6.4.9)$$

for real signals. This stems from making the substitution of the LMS cost function (6.4.6) into (6.4.5). Note that (6.4.9) is in vector notation due to there being more than

one tap within the transversal filter. Therefore, the difference equations for updating the values of the individual taps at a particular sample instant n are:

$$\begin{aligned} w_1[n+1] &= w_1[n] + 2\mu x[n]e[n] \\ w_2[n+1] &= w_2[n] + 2\mu x'[n]e[n] \end{aligned} \quad (6.4.10)$$

where the carrier reference signal $x[n]$ has been substituted for the general reference signal $u[n]$ and $x'[n]$ is the reference signal input shifted by 90° as in Figure 63.

The value μ is known as the convergence or step parameter, which controls how quickly the LMS algorithm will converge to an optimal solution in the LMS sense. Simply stated, a large μ will cause rapid convergence while a small μ slower convergence. In the latter case, more resolution allows a better adaptation solution, i.e. less variation of the tap weights around their optimal value after convergence, often called “misadjustment noise,” with the tradeoff being more time required in the adaptation process. Moreover, a convergence parameter that is too large is apt to cause divergence from the desired solution. The aforementioned algorithm will successively update the weight values of each tap in the transversal filter to minimize the power of the error signal. In this fashion, the LMS algorithm will seek an optimal adaptation solution.

The action of the adaptive algorithm is to remove the portion of $d[n]$ that is correlated with $x[n]$. Consequently, it is important that the part of $d[n]$ that is to be preserved (i.e., the carrier signal modulated by the sensor data) be uncorrelated with the unmodulated carrier interfering signal. If Differential Binary Phase-Shift Keying (DBPSK) modulation is used and the binary data is random with a uniform distribution, i.e.:

$$P('1') = P('0') = \frac{1}{2}, \quad (6.4.11)$$

where $P(x)$ indicates the probability of x occurring, the signals will be uncorrelated. This property results from the fact that the DBPSK signal will not have any power at the carrier frequency under these conditions. Since the ‘1’ and ‘0’ bits are mapped to +1 and -1, the statistical mean (i.e., expected value) will be zero. However, with real (redundant) sensor data, the adaptive filter is likely to partially suppress the carrier modulated by the sensor data since this condition cannot be guaranteed. Redundant sensor data is guaranteed without the use of source coding or data compression.

As a result, the adaptive algorithm is operated only during a *training* phase, during which the modulated carrier signal is not transmitted. During this period, $d[n]$ contains only the multipath signal and the adaptive algorithm finds the weight vector needed to suppress this signal. After training, the adaptive algorithm is stopped but the weight vector is retained. The modulated carrier containing the sensor data is then sent and the adaptive filter continues to suppress the multipath signal while passing the modulated carrier. Since the multipath conditions change very slowly (i.e., the weight vector does not need to change rapidly), training periods are only required occasionally.

Figure 65 - CW double-hop LMS adaptation process.

To review, consider the LMS adaptation process shown in Figure 65. To setup and adaptively equalize the channel, the following procedure occurs:

1. REST PHASE: At time $t = 0$, the channel is initially at rest.

2. TRAINING PHASE: At time $t \approx 1$ millisecond (shortly after the program starts), the carrier is transmitted into the channel, but is not modulated. Therefore, the only signal received at the sink transducer is the multipath component. The LMS suppression algorithm works to suppress the multipath noise. Multipath components of the signal have been reduced to a nominal level using the LMS algorithm. When the average voltage of the noise reaches a set threshold level, the weight values w_1 and w_2 are no longer updated. Their values are held in memory and used for cancellation in the DATA PHASE.
3. DATA PHASE: At time $t \approx 2.5$ seconds, the adaptation process has completed and data transmission begins—i.e., modulation occurs within the vessel. The filter weight vector determined in the ADAPTATION PHASE is held constant during this phase. Multipath interference is canceled, while the DBPSK-modulated signal is left for demodulation.
4. RE-TRAINING PHASE: The signal-to-noise ratio (SNR) is continuously being monitored during communications. While the LMS algorithm is ideal for a stationary channel—where the noise at time t is statistically identical to the noise at another time $t + \tau$ for any τ [75]—it is practically the case that the noise characteristics vary over time. To account for this, when the SNR level falls below a threshold suitable to sustain communication data transmission pauses and steps 1-3 reoccur.

The time values presented above are based on experimental results that will be presented later. Times may be longer or shorter for different implementations.

6.4.1.2 Normalized Least Mean-Squared (NLMS) Algorithm

To account for lengthy convergence times given a fixed convergence parameter μ , researchers sought to develop an LMS algorithm where μ changed appropriately as the algorithm ran. Specifically, an algorithm whereby μ varied in inverse proportion to the energy of the reference signal $x[n]$ is attractive. This makes sense from an intuitive standpoint—when the reference sinusoid reaches its largest values, we would like the weight vector to change more slowly to prevent it from diverging from the “steady-state” value. Equivalently, when the reference signal is at smaller energies larger changes in the tap vector is permissible since divergence is less likely. While this is just an intuitive

postulate, the formal development involving the method of Lagrange multipliers for complex weight vectors has been presented in reference items such as [2]. The phrase Normalized Least Mean-Square (NLMS) was coined for the adaptive algorithm satisfying these properties [76].

The NLMS algorithm varies only slightly from the LMS algorithm presented earlier. Specifically, the tap-weight update equation for real signals takes on the slightly modified form:

$$\vec{\mathbf{w}}[n+1] = \vec{\mathbf{w}}[n] + \frac{\mu}{\|\vec{\mathbf{u}}[n]\|^2} \vec{\mathbf{u}}[n]e[n] \quad (6.4.12)$$

Where $\|\vec{\mathbf{u}}[n]\|$ is the Euclidean norm of the reference signal. It is evident that the μ from the original LMS update equation (6.4.9) is now normalized by the energy (power) of the reference signal. To circumvent computational limitations of dividing by a near-zero number when $\|\vec{\mathbf{u}}[n]\|^2$ becomes very small, a positive constant a is added to (6.4.12) as:

$$\vec{\mathbf{w}}[n+1] = \vec{\mathbf{w}}[n] + \frac{\mu}{a + \|\vec{\mathbf{u}}[n]\|^2} \vec{\mathbf{u}}[n]e[n] \quad (6.4.13)$$

without loss of generality. Note that while this algorithm may be more attractive in general, it does have the drawback of increased computational complexity.

As the experimental results demonstrate, the implementation of the NLMS algorithm is highly successful; so much so that it led to a ten-fold increase in the communication rate from 500 bps to 5 kbps compared to the earlier mentioned methods. In addition, as compared to the LMS algorithm the NLMS algorithm provides comparable suppression in approximately ten-times fewer samples.

6.4.2 Experimental Results

In the first experiment with the adaptation algorithms, the initial investigation was made with the LMS algorithm in use with the single-hop configuration shown in Figure 19. That is, the transmitted carrier is used as the reference signal and the received signal is used as the desired signal $d[n]$. Arbitrarily choosing $\mu = 0.001$ achieved the error signal shown in Figure 66 for this single-hop setup. After convergence, the error

signal is reduced to approximately 3 mV peak, which is about equal to the noise floor of the DAQ system. We would not expect the system noise floor to be effectively canceled since it is uncorrelated to the reference signal $x[n]$.

Figure 66 - 100-point RMS voltage after 200 iterations of the LMS algorithm for $\mu = 0.001$.

To gain an appreciation for the use of different *fixed* (i.e., not normalized) values for μ in the double-hop fashion, the following experiment was performed:

1. Initially, the channel is at rest.
2. The carrier is started and transmitted in the single-hop fashion.
3. A value of μ is chosen by the equation:

$$\mu = (n+1) \times (1 \times 10^{-6})$$

where n is the trial number ($0 \leq n \leq 4200$).

4. The adaptation runs for 2000 points in an attempt to cancel the received carrier.

5. The windowed root mean-squared (RMS) voltage value is computed over the final 100 points of the error signal $e[n]$ and recorded.

The results of this experiment can be seen in Figure 67. It is evident that there is some critical $\mu = \mu_0$, for which the power of the error signal is minimized within the number of iterations chosen [2]. This value of μ makes the best compromise between speed of adaptation and misadjustment noise, i.e. the variation of the weights after convergence. While it is difficult to see in the graph, the minimum occurs near the 335th trial, with $\mu = 0.000335$. For very small values of μ , the error remains high because the algorithm does not have enough time to converge within the 2000 sample period. For values of μ above the optimal value, the error increases due to increased misadjustment noise.

Figure 67 - 100-point RMS voltage after 2000 iterations of the LMS algorithm for varying μ .

Figure 68 shows a set of results for a similar experiment for the case where the algorithm is allowed to run for 4000 iterations. In addition, to reach the minimum value of μ more quickly the update equation is now changed to:

$$\mu = (n+1) \times (1 \times 10^{-6}) + 0.0003$$

In this instance, the graph has been zoomed in around the saddle point on the graph. The same value of μ produced the minimum error signal as in the earlier experiment.

Figure 68 - 100-point RMS voltage after 4000 iterations of the LMS algorithm for varying μ .

Choosing $\mu = 0.000335$ —which resulted in the minimum in the charts of Figure 67 and Figure 68—yields the LMS suppression result in Figure 69 for the double-hop configuration. In this case the desired signal, $d[n]$, is the carrier signal that has been collected at the sink transducer and results from the carrier signal that is injected at the source transducer. Both transducers are on the same side of the block. Clearly, the algorithm is effective; the output error of the operation (originally over 200 mV peak), is reduced to approximately 10 mV peak in 2252 iterations. To prevent memory buffer overflow, a software-induced delay results in about 1 millisecond per iteration.

The next step in the experimental process is to determine the effectiveness of the LMS suppression when data is being communicated from the inside to the outside. To simplify the experiment, DBPSK is not initially utilized but instead CW OOK is used—a

180° phase shift is introduced to transmit a ‘1’ while *no carrier* is sent to transmit a ‘0.’ This is in contrast to DBPSK which would use a 0° phase shift to transmit a ‘0.’ CW OOK is less optimal from a communications point-of-view than DBPSK, having a greater susceptibility to noise, but is useful to show the advantages of the adaptive suppression. Figure 70 exhibits the result of this operation with (lower trace) and without (upper trace) the use of LMS adaptive suppression. Note that this modulation scheme is not as easily discernable without LMS suppression. The SNR is found to be about 15.92 dB with the LMS suppression.

Figure 69 - 100-point RMS voltage after LMS adaptation in the double-hop configuration.

Figure 70 - Double-hop OOK communication – with (bottom) and without (top) LMS suppression.

Next, the NLMS algorithm was implemented in the same double-hop configuration. The NLMS algorithm can be made to converge much more quickly than with the ordinary LMS algorithm without large misadjustment noise. Consider the graph shown in Figure 71, which shows the tap weight vector verses the iteration number. The tap weights vary very little after the first few hundred iterations of the NLMS algorithm. The adaptation process for the double-hop channel using the NLMS algorithm is shown in Figure 72. Note that this graph is shown in discrete-time (or sample-time) to emphasize that the algorithm is a point-by-point routine. In this case, the NLMS algorithm reached an offset error comparable to what the LMS algorithm of Figure 69 obtained—in over 1000 samples—with only 175 samples.

Figure 71 - Tap weight vector verses iteration number for NLMS algorithm.

Figure 72 - NLMS adaptation process (discrete time) for the double-hop channel.

Next, CW OOK modulation with interference suppression provided by the NLMS algorithm was performed as it was for the LMS algorithm in Figure 70 above. As is demonstrated in Figure 73, the NLMS algorithm yields even better results than with the LMS algorithm in the same configuration. Specifically, the NLMS algorithm yields an SNR—on average—of 20.91 dB. We can conclude, therefore, that the NLMS algorithm is inherently more attractive; it offers fewer adaptation iterations and better SNR results. The tradeoff is increased computational overhead which is transparent to the user in this installation. To complete experimentation, modulation is converted from OOK to DBPSK. Now, the data bits representing the sensor voltage are differentially encoded, the ‘0’ bits are converted into -1’s, the ‘1’ bits are converted into +1’s, and these ± 1 symbols are used to phase modulate the sinusoidal carrier as discussed in Section 5.2.4 (Differential Binary Phase-Shift Keying (DBPSK)) on page 66 earlier.

Now, an encoded bit of ‘0’ corresponds to a carrier phase shift of 180° and an encoded bit of ‘1’ corresponds to a carrier phase shift of 0° . Also, the full carrier path is implemented, with carrier generation outside, data modulation inside, and reception of the modulated signal outside of the vessel. Sensor data is represented by 12 bit binary words and each word is sent as a burst due to the limitations of the DAQ mentioned earlier. The received waveforms before and after NLMS suppression are shown in Figure 74. Note that unlike the case with CW OOK modulation, the signal is present during the whole burst because the carrier is always transmitted and only its phase is altered to convey the information.

Figure 73 - Double-hop communication with and without NLMS suppression.

Figure 74 - Double-hop DBPSK communication with and without NLMS suppression.

It is interesting to note that in all the waveforms that show the modulated signal with interference prior to interference suppression (Figure 70, Figure 73, and Figure 74), it is possible to discern the information by observing the envelope of the waveform. This occurs because the modulation of the carrier on the inside of the vessel in both the OOK case (in which the carrier is allowed to pass or not) and the DBPSK case (in which the carrier phase is shifted) alters the signal that is added to the multipath component and hence changes the amplitude of the composite signal. A simple receiver, then, could be constructed using an envelope detector. In viewing the figures, it is clear that the envelope will have a large DC term due to the interference but this DC value can be removed using a training phase. As in the training phase for the LMS or NLMS adaptive filters, no signal would be sent during training. The DC level at the output of the envelope detector would be sampled, stored and, during the data transmission phase, this DC level would be subtracted from the envelope detector output.

The primary drawback of using envelope detection, a form of *non-coherent* demodulation, is that it is less optimal in terms of performance than differential detection (for DBPSK), meaning that it has greater susceptibility to noise. Generally speaking, differential detection has greater efficacy over its non-differential counterpart in all possible modulation methods [68]. In a communication environment in which signal power is limited, this can be a critical limitation of the technique. In this application, however, the signals may be very large, particularly if we are delivering electrical power into the vessel using the carrier signal. Therefore, the power inefficiency of envelope detection may not be important while its relative simplicity remains very attractive.

The NLMS algorithm has been successfully implemented in a full communication system using the DAQ. The mainstay of the software interface is the calibration routine—a screenshot of which can be seen below in Figure 75. When run, the calibration routine will stop modulation and signal feed-through internal to the vessel. During this period, the NLMS algorithm will run and attempt to suppress the received signal (i.e., the interference). As an adaptation performance measurement, the variance (over 100 points) is computed for the error signal $e[n]$. The variance is a suitable proportional measure of the random noise power [43]. It was empirically determined that calibration has reached its best-case level when $\sigma^2 = 3 \times 10^{-7}$.

A plot of the variance versus sample iteration number can be seen in the upper right of the image in Figure 75. The calibration routine can be run manually, but in the automatic implementation the routine will run first at startup and thereafter, whenever the SNR becomes too low for communication to continue reliably. The figure also shows the error signal variance as well as the tap weight values as a function of iteration number. Figure 76 shows a screenshot of the communication interface for the double-hop configuration. In this screenshot, the received signal before and after interference suppression is shown in the upper-left and upper-middle plots, respectively. The upper-right plot is of SNR versus the transmitted binary word number. The SNR varies over time, due to changes in the direct acoustic path, which necessitates the re-adaptation of the tap weights periodically. If the channel was unchanging, the SNR would remain relatively constant over time and re-adaptation would not be necessary. Finally, a plot of the transmitted voltage (in this case random values are sent) is presented with the decoded bitstream and other data of interest. This data includes the transmitted voltage, SNR, and most importantly, data transmission speed. The important result here is that the use of the NLMS algorithm results in a ten-fold increase in the usable transmission speed from 500 bps to 5 kbps.

Figure 75 - Calibration routine for NLMS calibration routine in for double-hop setup.

Figure 76 - Communication interface for the double-hop DBPSK configuration.

6.5 Reflected-Power Configuration

As overviewed earlier in Section 4.3 (Reflected-Power Configuration) on page 54, the reflected-power configuration reduces the number of necessary transducers from two pairs to one. By systematically varying the acoustic impedance at the channel-transducer interface on the “inside” of the vessel, it is possible to reflect a minute, but measurable amount of power back to the source. Again, this configuration is attractive in its minimization of required transducers. It does, however, reduce the potential for continuous power delivery trans-wall. Regardless, the current research sought to investigate the technique and develop a suitable demonstration system. The signal processing in this configuration is of reduced complexity, approaching the techniques utilized for the pulsed double-hop configuration. As a result of the considerable attenuation through the acoustic channel, a substantial amount of power is required in the originating signal. As a result, the CW mode is practically unachievable in the current instance even with the use of adaptive suppression. Therefore, the use of a pulsed carrier is the only demonstrable method in this configuration.

6.5.1 System Configuration

Recall from the earlier discussion of this configuration that this configuration follows the following system process:

1. A pulse is transmitted through the channel.
2. The terminals of the receive transducer are either shorted or left open-circuited to convey a ‘1’ or ‘0,’ respectively.
3. The DAQ captures a section of interest as received from the originally transmitting transducer.

This process continues until a binary word (in this case 12-bits) is formulated. Once completed, the global process continues to repeatedly convey the instrumented sensor data across the channel until the user requests it to stop. This methodology can be visualized in Figure 77 below, as repeated from Figure 21 earlier.

Figure 77 - Reflected-power configuration.

Note that the overall signal processing in this configuration is no different than that presented in Section 6.3 (Pulsed Double-Hop Configuration) on page 82 earlier. To avoid redundancy, this information will not be presented again. However, in contrast this configuration does not use mixed gating as it does in the double-hop configuration. The necessity in this configuration is effectively to “short” the terminals of the transducer on the inside of the vessel. Figure 78 shows the simple diode switch circuit which is utilized for this purpose. Since a mechanical switch such as a relay is both difficult for logic to drive directly and is slow in its operation, a solid-state switch is the most attractive solution for switching. In addition, we overridingly seek to minimize complexity where possible. A switch meeting an acceptable level of simplicity with these desirable properties can be achieved with this diode switching circuit. Note that the capacitor and resistor are used to block DC current from the transducer and limit current draw from the driving circuit.

Figure 78 - Shorting of the receiving transducer using a simple diode switch.

With this circuit, the binary signal can be modeled as the DC power source shown in the figure. If the DC voltage value for a binary ‘1’ is large enough to forward bias the diode (>0.7 V typ.), a very low impedance (nearly shorted) condition will occur between the terminals of the transducer. On the other hand, when the DC voltage for a binary ‘0’ is low enough to reverse bias the diode, an open-circuit condition will arise. In the current implementation, therefore, having ‘1’ and ‘0’ switched as +1.0 V and -1.0 V for the binary word encoded internally is suitable for operation. Hence, the acoustic impedance at the internal channel-transducer interface is deterministically modified in relation to the binary word representing the measured sensor data.

6.5.2 Experimental Results

Figure 79 exhibits the fundamental results for the reflected-power mode. In the figure, the originating pulse can be seen as the first signal (on the left) in the two plots. Thereafter, the first reflected pulse received (by the originally transmitting transducer) can be seen in both the transducer open-circuit (top) and short-circuit (bottom) conditions. While the difference between the two conditions is subtle, the shorted condition is noticeable in the constructive interference of the ripple occurring just after the first reflected pulse in the 64-70 μ s range. This subtle change is appreciable and deterministic enough to reliably detect. It should be reemphasized that the aforementioned plot shows the transmitted response of the external transducer sending a pulse, followed thereafter by the reflected pulse received by the same transducer. Therefore, the delay in receiving the returning pulse should be twice the transit time across the wall in one direction. We expect—due to the nature of the acoustic channel—that there will be returning pulses via the LOS transmission path (reference Figure 3 in Section 3.1). This is consistent with the response visible in Figure 79.

Figure 79 - Originating and reflected pulses – transducer open-circuit (top) and closed-circuit (bottom).

Similar to the pulsed double-hop mode presented earlier, the region of interest is gated such that the change is observable in the windowed area. The result of this operation can be seen in the left two plots of Figure 80. In addition, the region of interest is sent through a rectification function and a low-pass filter (LPF) to extract the envelope and remove high-frequency components. The result is a single pulse which can be compared to a threshold to determine the state of the internal transducer and, consequently, whether a '1' or '0' was sent. The output from the LPF operation can be seen in the right two plots of Figure 80. Note that the threshold region for determining whether a binary '1' was transmitted is represented by the shaded region on these plots.

Figure 80 - Gate, filter, and threshold for the reflected-power mode.

Each plot in Figure 80 shows the windowed 4 μ s portion of the latter portion of the first reflected pulse. The top row of the figure shows the result when the internal transducer is left in the open-circuit configuration. Notice that while there is some ripple, the filtered signal does not pass the threshold set (currently 50 mV). The communication software controlling the DAQ system concludes, therefore, that a binary ‘0’ was sent. In the bottom row, the result of the internal transducer operating in the short-circuit mode is shown. It is observable that the stronger reflection causes the filtered signal to fall well above the threshold. In a similar fashion, the system concludes that a ‘1’ is sent.

While a slight change is noticed using the 0.5-inch transducers on the pressure vessel, it is not substantial enough to sustain reliable communications. The reduction in the size of the change in the reflected pulse between the open and short conditions is likely due to the larger beam width of the smaller transmitted transducers at the opposite

side of the acoustic channel. Better instrumentation could enable communication even with the observed smaller change, while focusing the acoustic power could result in a larger change. For these initial experiments, however, signals and data are collected using the 1.0-inch transducers on the test block to simulate the wall of the vessel. Note that the repetition rate of the pulses sets the bit rate for the system and this rate is limited by the duration of the response to a single pulse (i.e., the multipath response), since the reflected pulses must be allowed to attenuate before the next pulse is applied for reliable data detection. For the test setup with the steel block, reliable communication rates of approximately 300 bps are achievable.

Theoretically, the reflected-power technique could also be used with CW signals. Here, varying the load on the inside transducer will vary the standing wave arrangement that is produced in the wall due to the CW carrier and reflections. However, since the external transducer transmits continuously in the CW case, it is more difficult to detect the variations of the standing wave arrangement. Attempts to isolate a usable signal were not immediately successful. A hybrid double-hop/reflected-power system suggested later for future study may provide a way of gaining the simplicity of the reflected-power system and the performance of the double-hop system with CW signaling.

7. Discussion and Conclusions

The current research effort successfully demonstrated the ability to communicate through a solid steel wall using ultrasound. Specifically, digital data representing measured sensor voltage levels are conveyed across the channel of interest solely utilizing an acoustic link and digital modulation techniques. In the demonstration systems, the user could choose to acquire true analog voltage data or data simulated by the software developed. Several methods were attempted and successfully demonstrated in actual communication systems. The use of the DAQ system facilitated the rapid development of each of the configurations and in every instance, approximated the results achievable with dedicated electronic circuits developed and utilized for this specific purpose. The success and results from the current research not only demonstrate the achievability of such a system, but also exhibit the potential for increased data rates and multiplexed measurements. Moreover, as a guiding constraint in the development methodology, the minimization of complexity makes the techniques demonstrated suitable for use in an actual volatile pressure vessel environment. Specifically, initial empirical results exhibit the potential for acoustic power delivery at levels appropriate for the operation of such minimalist circuitry.

The overriding difficulty with implementing any communication method for the acoustic channel is the presence of multipath interference. It was clearly demonstrated that this problem severely limits the reliable communication of data across the steel wall channel. Therefore, the multipath interference must be addressed whether by more advanced signal processing methods or by simply waiting for the interference to acquiesce. In either case an initial model is desired that can characterize the channel and communication paradigms in a simulated environment. A simplified channel model approximation was presented that enables the simulation of many of the communication paradigms of interest. This type of model was utilized as an initial test bench prior to the realization of the configurations discussed. As it was discussed earlier, this model is very limited in that the multipath interference phenomenon is, for the most part, random in nature. Therefore, a deterministic channel model is consequently intractable. A more advanced model could be investigated, perhaps utilizing the concept of random log-

normal multipath interference which the current results appear to approximate. However, this is beyond the scope of the current research. With a baseline understanding of the multipath problem, it was possible to develop and implement communication schemes for the steel wall acoustic channel.

The primary objective in the current study was to demonstrate the achievability and practicality of ultrasonic digital communication across the steel wall channel. To this extent, the initial demonstration was made using the single-hop configuration as discussed in Section 4.1. A simple digital communications technique, namely DBPSK, was used for this demonstration and the implementation was performed using a personal computer and a commercially available DAQ system. One-way communication, from the “inside” to the “outside” of the vessel was shown. Achievable data rates were around 500 bps. For an application where the complexity of the electronics internal to the vessel is limited, whether due to space, power, or environmental concerns, the approach used for this demonstration may not be feasible. The primary difficulty is that the internal system must generate the excitation for the transmit transducer, requiring the use of an oscillator for a CW system or a pulser for a pulsed system. Not only may this topology become unstable in the volatile pressure vessel environment, but the amount of power required to surmount the attenuation of the acoustic channel is not practically achievable. Therefore, while the single-hop configuration demonstrated achievability, it lacks utility. A system whereby the complexity of signal generation of suitable power levels is minimized is desirable for the end-goal implementation. Additionally, power amplification and stable carrier generation is attainable in the controllable environment external to the environment.

Consequently, the next proposition was for the use of a double-hop configuration in which the carrier excitation is delivered from the “outside” to the “inside” of the vessel where it is modulated to impart the sensor information, and, subsequently, fed back to the “outside.” This signaling topology was discussed in Section 4.2 and was also successfully implemented and demonstrated utilizing the DAQ system. Two methodologies were introduced: one utilizing a gated or “mixed” pulsed carrier technique and another utilizing a CW technique. While the pulsed carrier technique employed simplistic signal processing algorithms, it was demonstrated that it may not be the best suited

for the multipath vessel channel. Due to multipath effects, the technique is limited to a primitive “transmit and wait” process whereby multipath reflections are allowed to acquiesce prior to sampling the received data. Achievable data rates in this configuration were limited to about 500 bps. While this configuration utilizes simplified detection algorithms, it does not address the multipath interference as well as the CW double-hop configuration does.

The CW double-hop configuration requires more sophisticated signal processing techniques, but the obvious gain is in the increase of data transmission rates from 500 bps to 5 kbps. In addition, the CW double-hop configuration offers the advantage of low instantaneous power requirements for carrier delivery. Conversely, the CW carrier allows large average power to be delivered over time. Since the CW carrier is more readily rectified, it is best suited for wireless power delivery as well. While the use of a continuous carrier is exacerbated by the multipath interference phenomenon, the use of LMS adaptive suppression can mitigate the effects. This stems from the fact that the interference arriving at the receiving transducer from random multipath routes is a sum of in-phase and quadrature-phase basis components. The interference can consequently be “subtracted out” by appropriately selecting the LMS tap weight vector with the adaptive algorithm. The experimental results presented in Section 6.4.2 on page 106 earlier demonstrates how successful the use of this algorithm can be. Utilizing the basic LMS algorithm, an SNR of approximately 15.92 dB was achieved after about 2252 iterations of the adaptation algorithm. It was also shown that the NLMS algorithm achieved comparable results in over 90% fewer iterations.

In an effort to reduce the number of transducers necessary from two pairs to one, the reflected-power configuration was introduced and implemented as discussed in Section 6.5 earlier. In this configuration, it was shown that by varying the acoustic impedance at the transducer-wall interface internal to the vessel, a minute (but discernable) change in the received signal could be measured at the originating transducer external to the vessel. Mapping the change in the signal to bit states, a binary modulation scheme was formed. Aside from the advantage of fewer transducers, an added advantage is the simplified implementation internal to the vessel. Specifically, the change in acoustic impedance at the interface was achieved by shorting the leads of the

internal transducer with a solid-state switch. The successful configuration utilized a pulsed carrier. Unfortunately, due to the attenuation through the channel the measurable change is very small and also falls to the perils of multipath interference. As a result, this configuration was the slowest implementation—achieving a data transmission rate of no more than 300 bps. The attenuation through the channel is an area that must be addressed in future investigations of this research. This is the only configuration where this limiting factor prevented the successful implementation on the cylindrical pressure vessel. Moreover, it was mentioned earlier that this configuration could theoretically be implemented using a CW carrier instead of a pulsed one. However, the amount of attenuation through the channel prevents this configuration also; the calculated level of change is well below the noise floor of the multipath interference. Note that while it may increase the complexity of the system process, it may theoretically be possible to implement adaptive interference suppression for the channel when using a CW carrier in this reflected-power configuration. Again, the severity of channel attenuation prevented the realization of a complete demonstration from investigation with this technique.

As a result of the experimental verification of increased data rates in the CW double-hop configuration, it is clearly the best choice when viewed in light of performance measurements. In addition, the delivery of large amounts of average acoustic power over time makes the use of a CW carrier attractive for conversion to electrical power within the vessel. The drawback is, however, that two transducers are required internal to the vessel instead of one. Therefore, the reduction to a single transducer internal to the vessel as in the reflected-power configuration is attractive when considering the real estate necessary for the implementation. A system that could combine the primary advantage of the reflected-power technique (i.e., a single transducer on the inside) with those of the double-hop technique (i.e., a CW carrier and high data rate) is shown in Figure 81. This hybrid double-hop / reflected-power system combines the arrangement of the two earlier configurations. The arrangement and operation of the two external transducers mimics that of the double-hop configuration while the operation of the single internal transducer is identical to that of the reflected-power configuration. The originating external transducer would transmit a CW carrier into the vessel while the acoustic impedance of the internal transducer-wall interface is varied as it was in the

reflected-power configuration (i.e., by varying its electrical load). Thereafter, the external receiving transducer captures the incoming signals as it has in each of the other configurations.

Figure 81 - Hybrid CW double-hop / reflected-power system.

It is suggested that this configuration be investigated in future research because of the advantages mentioned. Based upon the experimental results obtained heretofore, it is believed that this configuration is not only achievable, but the best compromise across all design facets. Rote envelope detection may be possible in this configuration as it was in the reflected-power configuration. This is, however, dependent upon the SNR of the received signal and multipath interference. This is clearly the least complex of the possible signal processing algorithms that could be utilized. To more aggressively address multipath interference at the receiving UT, the use of adaptive interference suppression as in the double-hop configuration can be utilized if necessary. Along with the reduction in the number of internal transducers necessary, this would also be an attractive configuration from a power delivery standpoint. Between the conveyance of instrumented data, power could be harnessed through rectification of the received CW carrier. Angled transition pieces were machined for use as in Figure 81 above—with the launch angles appropriately acoustically engineered—and an investigative trial was

conducted. It was determined, however, that the channel attenuation is currently too severe to measure an appreciable change.

As the earlier discussions allude to, channel attenuation is a considerable limitation in the current research. Again, note that the “channel” refers collectively to the attenuation through the transducers and their interfaces with the wall, as well as the wall itself. From empirical observations, it is clear that there is a significant amount of acoustic attenuation at each of the transducer-wall interfaces. This is a function of transition pieces, acoustic couplant, mounting methods, etc. and is for the most part, an acoustic engineering challenge. Therefore, a discussion of these considerations is beyond the scope of the current work. Note that in many of the aforementioned configurations, simply amplifying signals or more aggressively filtering them was suitable to achieve a demonstration system. Nonetheless, this is perhaps the most overriding issue which must be addressed prior to continuing future developments in the area. From an efficiency standpoint, near-ideal acoustic matching should be a guiding objective. This may make more advanced arrangements possible from a signal processing standpoint. In addition, smaller transducers may be utilized if greater throughput can be achieved. This is clearly attractive from a size-of-implementation consideration. Most importantly, however, greater efficiency enhances the capacity for wireless, acoustic power delivery into the vessel. For example, a simple diode-bridge circuit was utilized to rectify the acoustic signal received from the trans-wall path to power an LED in one experiment. This was only achievable, however, with the use of a large RF power amplifier at the transmit side. Clearly, the current channel attenuation results in a highly inefficient power delivery arrangement.

Finally, it is clear from the current research that digital communication systems through a thick steel wall utilizing ultrasound are not only theoretically possible, but practically achievable. This result has been demonstrated through the implementation of actual communication systems utilizing a commercially available DAQ system in the single-hop, double-hop, and reflected-power configurations introduced earlier using either a pulsed or CW carrier where appropriate. The multipath interference phenomenon is of significant consideration in the application of an ultrasonic digital communication scheme for the steel wall. Unless aggressively addressed with signal

processing schemes such as adaptive interference suppression, achievable data transmission speeds are greatly limited. While pulsed carrier modes allow the channel attenuation to be overcome in many situations, a CW carrier is more attractive for acoustic power delivery as mentioned. If the issue of acoustic matching is appropriately addressed, it is believed that the efficiency of the system will increase greatly. Doing so would not only more readily facilitate power delivery, but would also allow the investigation of more robust and attractive hybrid configurations. If the restrictions discussed in this thesis are appropriately addressed, based upon experimental results there is no doubt that standalone systems for the specific purpose of trans-wall data communication and acoustic power delivery can be realized.

PARENTHETICAL ABBREVIATIONS

A

ADC - Analog-to-Digital Converter (Conversion).....	62, 63, 64, 80, 86
AM - Amplitude Modulation.....	62
ASK - Amplitude-Shifted Keying	83

B

BPF - Bandpass Filter.....	34, 35
BPS - Bits Per Second	94
BPSK - Binary Phase-Shift Keying.....	6, 9, 11, 63, 64, 65, 66, 67, 69, 70
BW - Bandwidth.....	36

C

CDMA - Code-Division Multiple Access	9
CW - Continuous Wave xii, 1, 2, 3, 53, 54, 57, 61, 62, 63, 64, 72, 73, 82, 85, 95, 96, 104, 109, 110, 113, 120, 125, 127, 128, 129, 130, 131, 132	

D

DAC - Digital-to-Analog Converter (Conversion).....	87
DAQ - Data Acquisition (System)..	3, 5, 11, 16, 26, 52, 57, 58, 59, 61, 71, 72, 73, 74, 75, 76, 80, 85, 86, 89, 90, 96, 107, 113, 116, 120, 124, 126, 127, 131
DBPSK - Differential Binary Phase-Shift Keying	2, 95, 103, 105, 109, 110, 113, 116
DFE - Decision Feedback Equalization.....	15
DSP - Digital Signal Processing (Processor).....	16

E

EM - Electromagnetic.....	30
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F

FFT - Fast Fourier Transform.....	78, 79
FTP - File Transfer Protocol.....	11

I

IDT - Inter-digital Transducer	9
ISI - Intersymbol Interference	10, 12

K

KBPS - Kilobits Per Second	117
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L

LAW - Longitudinal Acoustic Wave	30
LMS - Least Mean-Squared....	99, 102, 103, 104, 105, 106, 107, 108, 109, 110, 111, 112, 113, 116
LOS - Line-of-Sight.....	7, 23, 33, 34, 35, 39, 44, 48, 96, 122
LPF - Low-Pass Filter.....	92, 93, 123

M

MDPSK - M-ary Differential Phase-Shift Keying	13
MFSK-DPSK - Multiple Frequency-Shift Keying - Differential Phase-Shift Keying....	10
ML - Maximum Likelihood.....	14
MSE - Mean-Squared Error.....	102

N

NDT - Non-Destructive Testing	23, 85, 90
NLMS - Normalized Least Mean-Squared.....	105, 106, 112, 113, 115, 116, 117, 118

O

OOK - On-Off Keying.....	3, 83, 109, 110, 111, 113, 116
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P

PAM - Pulse Amplitude Modulation	3, 82, 83
PDF - Probability Density Function	14, 49
PLL - Phase-Locked Loop.....	15, 67
PM - Pulse Modulation	1
PSK - Phase Shift Keying.....	9, 11, 12, 14, 18, 64

PZT - Piezoelectric Transducer	9, 32, 33, 34
<i>R</i>	
RF - Radio Frequency.....	1, 7, 18, 131
RLS - Recursive Least Squares	10, 15, 16
RMS - Root Mean-Squared	107, 108, 109, 110
<i>S</i>	
SAW - Surface Acoustic Wave	9
SNR - Signal-to-Noise Ratio	10, 18, 49, 62, 63, 105, 110, 113, 117, 128, 130
<i>T</i>	
TDMA - Time-Division Multiple Access.....	11
TEM - Transverse Electromagnetic	30, 31
<i>U</i>	
USB - Upper Single Sideband	8
UT - Ultrasonic Transducer .. 2, 3, 4, 19, 21, 23, 30, 31, 34, 35, 52, 54, 55, 57, 59, 61, 82, 90, 95, 130	
<i>V</i>	
VI - (LabVIEW) Virtual Instrument.....	11, 75, 76, 78, 80, 81, 82
<i>W</i>	
WSS - Wide-Sense Stationary	100

APPENDIX A.1: Compliance Sheets for the 1.0 MHz Gamma Probes

GE Inspection Technologies

Krautkramer Transducer Certificate of Conformity Compliant to ASTM E-1065 Guidelines

Date: 2/18/2005
Product Code: 113-261-043
Serial Number: 014P73
Description: FDCP-XDCR ORRH 1X1 BNC

Test Setup--
Test Target: 2" 303 Steel
Energy: 4
Impedance: 250 Ω

Waveform is the first Return Echo from the indicated target.

UTA S/N: MM00113
UTA Cal Due Date: 5/5/2005
O-scope SN: B010804
O-scope Cal Due Date: 10/31/2005

Software: FNT0040 Rev G

Test Data--
Sensitivity(Abs dB): -29.20 dB

Pulse Duration@-20dB: 2.74E-06 Sec
Peak Frequency: 1.07E+06 Hz
Center Frequency: 1.05E+06 Hz
% Bandwidth@-6dB: 60.0 %

The Ultrasonic Transducer listed in this document has been performance tested and meets all manufacturing specifications. It performed as designed and specified on the applicable style of Krautkramer instrumentation.

Inspector: JPT

The accuracy of the transducer described above has been confirmed by factory standard test equipment and laboratory reference standards traceable to the National Institute of Standards and Technology. This facility's Quality System is registered to ISO 9001:2000, and is compliant to MIL-STD-45662A and ANSI/NCSS Z540-1-1994.

GE Inspection Technologies, LP
50 Industrial Park Rd.
Lewistown, PA 17044
Tel.: 717.242.0327
Fax: 717.242.2606
GEInspectionTechnologies.com

021-247-287, Rev D

Krautkramer Transducer Certificate of Conformity

Compliant to ASTM E-1065 Guidelines

Date: 2/14/2005
Product Code: 113-261-043
Serial Number: 014LK8
Description: FDCP-XDCR CRRHP 1X1 BNC

Test Setup--
Test Target: 2" 303 Steel
Energy: 4
Impedance: 250 Ω

Waveform is the first Return Echo from the indicated target.

UTA S/N: MM00113
UTA Cal Due Date: 5/5/2005
O-scope SN: B010804
O-scope Cal Due Date: 10/31/2005

Software: FNT0040 Rev G

Test Data--
Sensitivity(Abs db): -29.20 dB

Pulse Duration@-20dB: 2.48E-06 Sec
Peak Frequency: 1.03E+06 Hz
Center Frequency: 1.03E+06 Hz
% Bandwidth@-6dB: 57.0 %

The Ultrasonic Transducer listed in this document has been performance tested and meets all manufacturing specifications. It performed as designed and specified on the applicable style of Krautkramer instrumentation.

Inspector: JPT

The accuracy of the transducer described above has been confirmed by factory standard test equipment and laboratory reference standards traceable to the National Institute of Standards and Technology. This facility's Quality System is registered to ISO 9001:2000, and is compliant to MIL-STD-45662A and ANSI/NCCL Z540-1-1994.

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Lewistown, PA 17044
Tel.: 717.242.0327
Fax: 717.242.2606
GEInspectionTechnologies.com

APPENDIX A.2: Compliance Sheet for the 1.0 MHz Videoscan Probes

TRANSDUCER DESCRIPTION

PART NO.: V102
 SERIAL NO.: 515536
 DESIGNATION: CONTACT
 FREQUENCY: 1.00 MHz
 ELEMENT SIZE: 1 in. DIA.

TEST INSTRUMENTATION

PULSER/RECEIVER: PANAMETRICS 5052UA #1
 DIGITAL OSCILLOSCOPE: LeCroy LT342 / SIN: LT34202249
 TEST PROGRAM: TP103-3 VER. 10547G
 CABLE: RG 174/U LENGTH: 4FT

TEST CONDITIONS

PULSER SETTING: ENERGY: 1 ; DAMPING: 200 ohm
 RECEIVER SETTING: ATTN: 38dB ; GAIN: 40dB
 TARGET: 5 in. SILICA
 JOB CODE: TP200

MEASUREMENTS PER ASTM E1065

WAVEFORM DURATION:
 -140B LEVEL -- 1.32US
 -200B LEVEL -- 1.40US
 -400B LEVEL -- 7.72US
 SPECTRUM MEASUREMENTS:
 CENTER FREQ. --- 0.97MHz
 PEAK FREQUENCY -- 0.94MHz
 -60B BANDWIDTH -- 97.275 %

COMMENTS:

* MEASUREMENT NOT PERFORMED OR MAY BE UNRELIABLE.

** ACCEPTED.

TECHNICIAN: (3) DATE: 02-11-2005

APPENDIX A.3: Compliance Sheets for the 1.0 MHz Accuscan Probes

PANAMETRICS-NDT™
A Business of R/D Tech Instruments Inc.
Tel: 781-419-3900
www.panametrics-ndt.com

TRANSDUCER DESCRIPTION

PART NO.: A5365
SERIAL NO.: 510952
DESIGNATION: CONTACT
FREQUENCY: 1.00 MHz
ELEMENT SIZE: .5in. DIA.

TEST INSTRUMENTATION

PULSER/RECEIVER: PANAMETRICS 9000A UA #2
DIGITAL OSCILLOSCOPE: LeCroy LT342 / SN: LT3420114
TEST PROGRAM: TP103-3 VER. 10577B
CABLE: RG 174U LENGTH: 4FT

TEST CONDITIONS

PULSER SETTING: ENERGY: 1; DAMPING: 200 OHM
RECEIVER SETTING: ATTN: 44dB; GAIN: 40dB
TARGET: 1IN. LUCITE
JOB CODE: TP200

MEASUREMENTS PER ASTM E1065

WAVEFORM DURATION:
-140dB LEVEL --- 3.960 US
-20dB LEVEL --- 4.700 US
-40dB LEVEL --- 3.360 US
SPECTRUM MEASUREMENTS:
CENTER FREQ. ----- 0.93 MHz
PEAK FREQUENCY -- 0.9 MHz
-80dB BANDWIDTH --- 36.9 %

COMMENTS:

** ACCEPTED

TECHNICIAN (S)

Law Bonilla

DATE: 11-24-2004

TP103X Rev. A TEG.Dwg# 16150

A Business of R/D Tech Instruments Inc.

Tel: 781-419-3000
www.panametrics-ndt.com

TRANSDUCER DESCRIPTION

PART NO.: A538S
SERIAL NO.: 510979
DESIGNATION: CONTACT
FREQUENCY: 1.00 MHz
ELEMENT SIZE: .5 in. DIA.

TEST INSTRUMENTATION

PULSER/RECEIVER: PANAMETRICS 5030A UA #2
DIGITAL OSCILLOSCOPE: LeCroy LT342 / SN: LT34201114
TEST PROGRAM: TP103-3 VER. 1057B
CABLE: RG 174U LENGTH: 4FT

TEST CONDITIONS

PULSER SETTING: ENERGY: 1; DAMPING: 200 OHM
RECEIVER SETTING: ATTN: 40dB; GAIN: 40dB
TARGET: 1 IN. LUCITE
JOB CODE: TP200

MEASUREMENTS PER ASTM E1065

WAVEFORM DURATION:
-140dB LEVEL --- 4.000 US
-200dB LEVEL --- 4.760 US
-600dB LEVEL --- 9.680 US
SPECTRUM MEASUREMENTS:
CENTER FREQ. ----- 0.92 MHz
PEAK FREQUENCY -- 0.88 MHz
-60dB BANDWIDTH --- 37.25 %

COMMENTS:

** ACCEPTED

TECHNICIAN (6)

Luis Bonilla

DATE: 11-24-2004

SIGNAL WAVEFORM

FREQUENCY SPECTRUM

A Business of R/D Tech Instruments Inc.

Tel: 781-419-3900
www.panametrics-ndt.com

TRANSDUCER DESCRIPTION

PART NO.: A3385
SERIAL NO.: 510673
DESIGNATION: CONTACT
FREQUENCY: 1.00 MHz
ELEMENT SIZE: .5 in. DIA.

TEST INSTRUMENTATION

PULSER/RECEIVER: PANAMETRICS 5030AUA #2
DIGITAL OSCILLOSCOPE: LeCroy LT342 / SN: LT34201114
TEST PROGRAM: TP103.3 VER. 10577B
CABLE: RG 174U LENGTH: 4FT

TEST CONDITIONS

PULSER SETTING: ENERGY: 1; DAMPING: 200 OHM
RECEIVER SETTING: ATTN: 44dB; GAIN: 40dB
TARGET: 1 IN. LUCITE
JOB CODE: TP200

MEASUREMENTS PER ASTM E1065

WAVEFORM DURATION:
-14DB LEVEL --- 3.580 US
-20DB LEVEL --- 4.960 US
-40DB LEVEL --- 9.660 US
SPECTRUM MEASUREMENTS:
CENTER FREQ. ----- 0.92 MHz
PEAK FREQUENCY -- 0.85 MHz
-60DB BANDWIDTH --- 38.54 %

COMMENTS:

** ACCEPTED

TECHNICIAN (6)

John Bonilla

DATE: 11-24-2004

SIGNAL WAVEFORM

FREQUENCY SPECTRUM

PANAMETRICS-NDT™

A Business of R/D Tech Instruments Inc.

Tel: 781-415-3000
www.panametrics-ndt.com

TRANSDUCER DESCRIPTION

PART NO.: A539S
SERIAL NO.: 510906
DESIGNATION: CONTACT
FREQUENCY: 1.00 MHz
ELEMENT SIZE: .5in DIA

TEST INSTRUMENTATION

PULSER/RECEIVER: PANAMETRICS 5030A UA #2
DIGITAL OSCILLOSCOPE: LeCroy LT342 / SN: LT34201114
TEST PROGRAM: TP103.3 VER. 105778
CABLE: RG 174/U LENGTH: 4FT

TEST CONDITIONS

PULSER SETTING: ENERGY: 1; DAMPING: 200 OHM
RECEIVER SETTING: ATTN: 44dB; GAIN: 40dB
TARGET: 1 IN. LUCITE
JOB CODE: TP200

MEASUREMENTS PER ASTM E1065

WAVEFORM DURATION:
-140dB LEVEL --- 3.440 US
-20dB LEVEL --- 4.160 US
-60dB LEVEL --- -9.760 US
SPECTRUM MEASUREMENTS:
CENTER FREQ. ----- 0.95 MHz
PEAK FREQUENCY -- 0.9 MHz
-60dB BANDWIDTH --- 40.27 %

COMMENTS:

** ACCEPTED

TECHNICIAN (6)

Luis Bonilla

DATE: 11-24-2004

APPENDIX B.1: Directivity Plots for the 1.0 MHz Gamma Probes

Figure 82 - Directivity plot for the 1.0 MHz Gamma probe.

Figure 83 - 3D Visualization of the 1.0 MHz Gamma probe intensity.

APPENDIX B.2: Directivity Plots for the 0.5 MHz Gamma Probes

Figure 84 - Directivity plot for the 0.5 MHz Gamma probe.

Figure 85 - 3D Visualization of the 0.5 MHz Gamma probe intensity.

APPENDIX C.1: PSPICE Transducer Model

Figure 86 - PSPICE transducer model.

APPENDIX C.2: PSPICE Transducer Simulation

Figure 87 - Input pulse and response of the PSPICE model.

APPENDIX D.1: Matlab Code Utilized to Obtain Plots

```
clear all
close all

fs=10000000;           % Set sample rate
ts=1/fs;               % Sample period is the inverse of sample rate

t=[-1e-5:ts:1e-5];     % Build time sample vector
t_squared=t.^2;         % Performed computation of time squared for efficiency

BW=300000;             % Set 3 dB bandwidth of transducer for model
sig=BW/sqrt(8*log(2));  % Calculate sigma

f_c=1e6;               % Set the center frequency of transducer for model

h_both=zeros(size(t)); % Initialize the vector for the transducer-transducer
response

kappa=1/(100*sig*sqrt(pi)); % Set value for the constant gamma

% Define the transducer-transducer impulse response
h_both=(sig*sqrt(pi)*kappa)*(exp(-(t_squared.*(sig^2)*(pi^2)))).*cos(2*pi*f_c*t);

% plot(t,h_both,'k','LineWidth',2)
% set(gcf,'units','inches','Position',[2.5 5 9 7]);
% set(gca,'XMinorTick','on','YMinorTick','on','LineWidth',2,'FontSize',13);
% title('Transducer-to-Transducer Impulse Response');
% xlabel('Time');
% ylabel('Amplitude');
% grid on;
% axis([-1e-5 1e-5 -1.1e-2 1.1e-2]);

h_both_freq=fft(h_both).*t; % Find the frequency response
h_both_freq_centered=abs(fftshift(h_both_freq)); % Rotate (center) the FFT
l=length(h_both_freq_centered)-1; % Find length of frequency response
df=fs/l; % Find frequency increment
f=[0:df:fs]-fs/2; % Build frequency sample vector

% plot(f,h_both_freq_centered,'k','LineWidth',2)
% set(gcf,'units','inches','Position',[2.5 5 9 7]);
% set(gca,'XMinorTick','on','YMinorTick','on','LineWidth',2,'FontSize',13);
% title('Transducer-to-Transducer Frequency Response');
% xlabel('Frequency (Hz)');
% ylabel('Magnitude');
% axis([-1.6e6 1.6e6 0 2e-6]);

sz1=size(t);
sz1=sz1(1,2); % Find the size of the time vector
sz1=ceil(sz1/2); % Find the middle of the time vector (t=0)

pulse_length=ceil(1e-6/(2*ts)); % Find half of the pulse width

pulse=zeros(size(t));
pulse(sz1-pulse_length:sz1+pulse_length)=200; % Define the pulse

% plot(t,pulse,'k','LineWidth',2)
% set(gcf,'units','inches','Position',[2.5 5 9 7]);
% set(gca,'XMinorTick','on','YMinorTick','on','LineWidth',2,'FontSize',13);
% title('Ideal Pulse from NDT Device');
% xlabel('Time');
% ylabel('Amplitude');
% grid on;
% axis([-1e-5 1e-5 -30 230]);

% Find the output of the transducer-transducer impulse response to pulse as input
y=conv(h_both,pulse);
```



```

sz1=size(y);
sz1=sz1(1,2);           % Find the size of y

l=(sz1-1)*ts/2;
t=[-l:ts:l];           % Update time vector

% plot(t,y,'k','LineWidth',2)
% set(gcf,'units','inches','Position',[2.5 5 9 7]);
% set(gca,'XMinorTick','on','YMinorTick','on','LineWidth',2,'FontSize',13);
% title('Channel Pulse Response with Exponential Attenuation');
% xlabel('Time');
% ylabel('Amplitude');
% grid on;
% axis([0 1.5e-5 -3 3]);

ATT=2.74606;
v=5.789e5;              % Set the attenuation of the channel in dB/cm
alpha=(1/20)*ATT*v*log(10); % Define the attenuation factor

x=zeros(size(t));       % Initialize the vector

sz1=size(x);
sz1=round(sz1(1,2)/2);  % Determine the size of half of x vector

x=(y.*exp(-alpha*t));   % Define the output of the channel with attenuation
x(1:sz1-1)=[];          % Set values to 0 prior to t=0 (for causal response)

sz1=size(x);
sz1=sz1(1,2);
t2=[0:ts:(ts*(sz1-1))]; % Determine a suitable time vector for plotting
% plot(t2,x)
% axis([-0.5e-5 1.5e-5 -1.25 1.25]);

t=[0:ts:260e-6];
y_multipath=zeros(size(t)); % Create a large vector for the multipath response

sample_advance=round((52e-6)/ts); % Delay of reverberation in channel

sz2=size(t);           % Find out the size of the original time vector
sz2=sz2(1,2);

% See how many reflections will fit in the plot of the vector
reflections_that_fit=floor(sz2/sample_advance);

i=1;
k=1;
while i<=reflections_that_fit % Put in as many reflections as will fit
    y_multipath(k:(k+sz1-1))=x.*exp(-7800.*t(k:(k+sz1-1))); % Attenuate reflections
    k=sample_advance*i;
    i=i+1;
end

k=floor((26e-6)/ts); % See how many samples are necessary for initial delay

y_multipath=circshift(y_multipath,[0 k]); % Circularly shift for initial delay

% plot(t,y_multipath,'k','LineWidth',1.3)
% set(gcf,'units','inches','Position',[2.5 5 9 7]);
% set(gca,'XMinorTick','on','YMinorTick','on','LineWidth',2,'FontSize',13);
% title('Multipath Channel Pulse Response');
% xlabel('Time');
% ylabel('Amplitude');
% grid on;
% axis([0 2.6e-4 -2 2]);

gauss_noise=normrnd(0,0.05,1,sz2); % Create N(0,0.5) Gaussian noise
y_multipath_noise=y_multipath+gauss_noise; % Add noise to multipath response

plot(t,y_multipath_noise,'k','LineWidth',1)
set(gcf,'units','inches','Position',[2.5 5 9 7]);
set(gca,'XMinorTick','on','YMinorTick','on','LineWidth',2,'FontSize',13);

```

```
title('Multipath Channel Pulse Response with N(0,0.5) Gaussian Noise');  
xlabel('Time');  
ylabel('Amplitude');  
grid on;  
axis([0 2.6e-4 -2 2]);
```

APPENDIX E.1: BNC-2110 Front Panel Technical Drawing

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