DESIGN AND MANUFACTURE OF A HIGH-FREQUENCY ANNULAR ARRAY ULTRASOUND SYSTEM FOR MEDICAL IMAGING

by

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Abstract

This thesis presents the design of a high-frequency annular array ultrasound system suitable for medical imaging. To reduce the cost of the system, off-the-shelf parts were used whenever possible. The system consists of four main components: 1) a transmit beamformer, 2) a high voltage pulse generator, 3) an annular array transducer and 4) a receive beamformer. The transmit beamformer and pulser were designed for an 8-channel array but could be easily expanded for larger arrays. The pulser produces monocycle electrical pulses with centre frequencies that could be adjusted from 10-50 MHz and with amplitudes up to 90 Vpp. The annular array transducer has 12 equal area elements and a total active aperture of 6 mm. The transducer array produced pulses with a centre frequency of 20 MHz and 50% bandwidth. The resulting images had a lateral resolution of 172.5 µm at 10 mm and an axial resolution of 180 µm.

A new fabrication method was developed that makes it easier to build the array. The receive beamformer was based on a commercial 8-channel analog-to-digital converter. The digital signals were transferred to a laptop where the beamforming was performed in software. This avoided the need to develop custom hardware and allowed it to be reconfigured for different transducers by simply modifying the software. The beamformer used a new interpolation method that reduced the required sampling frequency while maintaining a satisfactory radiation pattern. The system produces images at 10 frames/sec.
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Finally, I would like to dedicate this thesis to my late grandmother, Bunny Cole, who was always supportive, even if she didn’t always know what it was I did. Love you, Nanny.
Statement of Originality

I hereby certify that all of the work described within this thesis is the original work of the author. Any published (or unpublished) ideas and/or techniques from the work of others are fully acknowledged in accordance with the standard referencing practices.

Holly Susan Lay

March, 2011
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Chapter 1

Introduction

Ultrasound imaging is an important form of medical imaging which has seen long-term usage in the medical industry due to its relative low-cost and non-invasive nature [1]. While MRI systems can cost upwards of $2 million, a conventional ultrasound system costs approximately a quarter of that, with additional cost savings possible, as outlined in this thesis [2]. Ultrasound also has the advantage of being a real-time imaging modality, which allows the operator to adjust the imaging parameters to optimize the resulting image. Additionally, it is suitable for real-time applications such as imaging blood-flow and guiding instruments during surgery.

1.1 Imaging with Ultrasound

Creating images using ultrasound is possible due to the reflection of sound waves when a change in material properties is encountered. The intensity of the reflection compared to the incident waves depends on the relative change in material properties, which can be represented as a ratio of their acoustic impedances, \(Z_i\). Given the impedance of the near-side material, \(Z_1\), and of the material on the far side of the transition, \(Z_2\), the relative intensity of the reflection with respect to the original pulse can be derived from Snell’s Law. The final ratio is:

\[
\frac{I_r}{I_i} = \left[ \frac{Z_2-Z_1}{Z_2+Z_1} \right]^2 \tag{Eqn. 1-1[3]}
\]

Thus, the amplitude of the reflected wave will be partially related to the relative density of the tissues under the surface. The other source of reflected energy is scatter from non-planar...
surfaces. By using a short burst of energy (pulse mode imaging), rather than a continuous wave (CW imaging), and assuming a constant wave velocity, the position of these reflectors can also be located based on total travel time.

The received set of reflected pulses from the internal reflections is known as amplitude-mode imaging, or A-mode. To build up an image from these A-mode lines, the scan-line must be shifted in a plane along the area of interest, creating a 2D brightness-mode or B-mode image. How this scanning is achieved depends on the type of ultrasound transducer used. The earliest transducers used a single dish-shaped piezoelectric element and the radiated ultrasound was focused using a lens, mirror or physical curvature of the device. To create the B-mode image, motors shift the transducer in either a linear or curved motion, creating a rectangular or arc-shaped image respectively. Transducers of this design are still in use today, as they have the advantages of being easy to manufacture and having simple electronic requirements. However, single element transducers are limited by a fixed focal point. This causes a degradation of resolution in the image the further away from the focal zone you get. In addition, a single element transducer will only produce a one-dimensional line of echoes. To produce a two-dimensional image, the transducer must be mechanically scanned.

To avoid some of the limitations of single element transducers, a transducer array can be used instead. Some of the first transducer arrays were annular arrays (Reid and Wild, 1958), consisting of concentric rings of elements which were focused electronically (Figure 1.1). These arrays are desirable to avoid the fixed focal point of the single element transducers.

Alternatively, a linear or a phased array can be constructed by using a transducer with a series of
rectangular elements aligned linearly along the plane of the desired image. Arrays then use electronic delays to focus the ultrasonic waves from the active elements to the desired depth along a given line in the image (Figure 1.2).

Figure 1.1 Concentric rings which form the active elements of an annular array

Figure 1.2 Physical characteristics of a linear or phased array along with the active aperture (shaded in grey) with respect to the active line and image plane

While linear and phased arrays have the advantage of being entirely electronically controlled, this introduces additional electrical complexity, as a separate transmit and receive circuit is required.
for each element of the array. Annular arrays supply a trade-off between linear arrays and single-element arrays as they have controllable focal depth, but only a small number of elements, minimizing electrical complexity.

Electronic focusing is produced in annular arrays with a set of pre-calculated delays. The pulses sent from the inner elements are delayed with respect to the outer elements, shaping the resulting wave-front into a concave shape. By changing the absolute delays, the radius of curvature of the wave-front can be altered, and the resulting focal point shifted towards or away from the transducer face. Only a single focal depth can be set for the initial pulse, as the transmit wave is produced in a single event. However, the received data can be analyzed one time-step at a time, allowing the receive focal depth to be slowly shifted to greater depths, improving the resolution of the final image.

Due to the axial symmetry of annular arrays, the beam can only be focused along the axis of imaging. Such arrays therefore require the same mechanical scanning as a single element transducer. To shift the focal point perpendicular to the axis of imaging, a linear or phased array is required.

Linear and phased arrays use similar techniques to shift the focal point across the array, but use slightly different strategies to build up the image. A phased array uses all its elements simultaneously to both send and receive the ultrasound energy and uses digital delays to change the angle of the scan line. This results in an arc-shaped final image. A linear array uses a subset of its elements for each scan-line, using electronic delays to focus the energy in a line.
perpendicular to the surface of the array. The active subset is then stepped across the array, producing a set of parallel scan-lines and resulting in a rectangular final image. Like the annular array, linear and phased arrays can use multiple transmit focal points and dynamic receive focusing to optimally focus the beam at each depth.

One of the downsides of linear and phased arrays is that they typically contain anywhere from 64 to 256 elements, while an annular array produces a comparable radiation pattern with as few as 7-15 elements. The large numbers of elements associated with a linear or phased array increases the number of connections required, as well as the complexity of the transmit and receive beamformer.

1.2 Image Resolution

When evaluating the quality of an ultrasound image, both the resolution and the contrast should be taken into account. The resolution of a B-mode image can be broken down into the lateral resolution and the axial resolution. Both resolutions represent the smallest distance two reflectors can be separated and still be distinguishable as separate points, and are measured in the plane of the image. The axial resolution is measured along the axis of the ultrasound beam and is a function of the length of the pulse. While this will naturally improve with increased frequency, which will be discussed in a later section, other factors can control this for a given centre frequency. The ideal ultrasound pulse can be modeled as a Gaussian or Hanning-windowed sine-wave. The axial resolution is obtained by measuring the points at which the amplitude of the envelope of the pulse drops to half the amplitude of the peak signal (-6 dB for a system that sends
and receives pulses) resulting in an ideal axial resolution of approximately twice the system wavelength.

Achieving this ideal response requires the design of a pulse generator that produces very short (broad-bandwidth) pulses as well as a broad bandwidth transducer. A narrower bandwidth transducer or pulse generator will cause longer-duration transmit pulses which will result in degradation of the axial resolution of the system.

The lateral resolution of a transducer array is determined by the ultrasound pulse bandwidth and frequency, the array geometry and by the ultrasound beamformer. The easiest way to determine the lateral resolution is to examine the two-way radiation pattern of the system. The two-way radiation pattern is obtained by stepping a point reflector across the array at the focal distance (either linearly or in an arc, as appropriate) and recording the maximum amplitude of the envelope of the received echo, as a function of the point reflector’s position (Figure 1.3).

![Figure 1.3 Example of a radiation pattern showing the lateral resolution of the main lobes and secondary lobes below 100 dB](image_url)

The lateral resolution is then a measure of the main beam width at the -6dB level. For an ideal transducer array, this should be approximately twice the f number. The f-number is the focal
distance divided by the width of the active aperture. In other words, a target placed a distance from the transducer equal to twice the active aperture would be said to be located at f/2. This terminology will be used throughout this thesis.

The radiation diagram can also be used to determine the peak and average level of secondary lobes. The level of these secondary lobes dictates the largest grey-scale range that can be used in the final display; if the display intensity is increased beyond this range, the brightest reflectors in the image will show “ghosts” at the positions of the side lobes. For clinical imaging purposes, it has been shown that secondary lobes should be suppressed by approximately 60 dB with respect to the main lobe.

1.3 High-frequency imaging

Traditional clinical ultrasound is usually performed with a centre frequency between 3-5 MHz. However, there has been significant research into ultrasonic devices operating at frequencies above 20 MHz, due to the resulting increase in resolution that is provided by the shorter wavelength [4][5][6][7][8]. As the signal attenuation with depth in tissue is frequency dependent, these higher frequency systems suffer a loss of penetration depth in exchange for increased image quality. Typical applications for high-frequency systems include skin, ophthalmic, intravascular, harmonic and small animal imaging [5][9][10][11][12].

The development of high-frequency array systems has lagged behind the development of single-element systems due to the increased difficulty in both designing electronics to operate at these
higher frequencies, and in manufacturing the transducers. Current commercially available analog to digital converters (ADCs) are typically limited to 150-200 MS/s. Although higher sampling rates are possible, they are only achieved at higher expense, lower dynamic range and often only a single device per package. With linear and phased array systems requiring ADCs on all their outputs, the cost and circuit board real estate requirements for these faster ADCs become prohibitive. With this limitation on sampling rate, digital up-sampling techniques become critical to achieve the required delay accuracy for correct beamforming.

Digitized data consists of data samples with an inter-sample time interval determined by the sampling frequency of the ADC used. As the focusing of ultrasound array often requires the application of delays smaller than this time interval, methods must be found to simulate the missing data between the sampled points. The process of calculating data points between the existing data is known as up-sampling.

While it is possible to perform beamforming on data sampled at lower frequencies using the simplest form of up-sampling, linear interpolation[13], many systems implement advanced algorithms to improve on this behavior[11]. Successful development of interpolation schemes with better accuracy will allow sampling rates closer to the minimum required to avoid aliasing.

Another potential electronic limitation is the maximum clock speed of the circuit boards and digital signal processing (DSP) chip. Most digital beamforming is performed on commercially available FPGAs or on custom VLSI chips. These commonly have maximum clock frequencies between 400 and 600 MHz, limiting the maximum sampling rate of incoming unprocessed data.
In addition, higher frequencies introduce additional design complexity to the connectivity between the ADC and the processing chip. At 25 MHz, any signal traces longer than 5-10 cm need to be impedance matched to avoid reflections and attenuation. Noise coupling from the clock and signal traces to the power supplies also increases with faster rise and fall times, requiring additional decoupling capacitors.

Equally difficult is the task of manufacturing ceramic transducers with centre frequencies above 20 MHz. A 20 MHz ceramic transducer must be lapped to a thickness of only 100 μm. Until the advent of kerf-less array designs, it was not possible to create linear or phased arrays above 30 MHz with bulk materials, as the dicing saws used to create the kerfs could not cut slots narrower than 20 μm. A 50 MHz linear array requires an element spacing of 30 μm (and a phased array 15 μm) to avoid unwanted regions of constructive interference in the radiation pattern called grating lobes. A standard blade would remove over two thirds of the material for the linear array, and would completely remove it attempting to cut the phased array. While kerfless design [14] and the development of micromachining techniques [15] has mitigated this limitation, making electrical connections to elements that small is still very challenging. Wire-bonding, which was used in early high-frequency array designs, requires a surface area at least 50 μm square on which to bond the wire [16]. This requires significant fan-outs from the array elements, increasing the total size of the final array. Researchers in this field are working on alternatives to wire-bonding, and have investigated approaches developed for integrated circuit field on flip-chip bonding techniques, such as solder bumps and conductive adhesives [17][18][19][20].
1.4 Low-Cost Systems

In addition to the current interest in high-frequency imaging, areas of research such as intravascular imaging and clinical set-ups have also encouraged the development of lower-cost systems. A conventional ultrasound system from Siemens or GE can easily cost as much as $500,000 [2]. This high price tag has led to an active market in reconditioned and second-hand units as many smaller research applications cannot afford a full-price unit. By taking advantage of developments in electronics developed both for the ultrasound as well as other markets, it is possible to produce a full system with a lower price tag.

Recent areas of research with potential to lower system implementation cost have included use of additional integration in commercially available Integrated Circuits (ICs)[21], the use of Field Programmable Gate Arrays (FPGAs) instead of custom built ICs[22][23][24] and the migration of processing from custom hardware low cost Graphic Processor Units (GPUs) in the host computer[2].

Recent developments in intravascular imaging are also driving interest in low-cost transducer manufacturing techniques. Each intravascular imaging probe must be discarded after use due to biological contamination concerns. This can significantly impact ultrasound imaging’s normally low day-to-day costs by adding a per-use cost. In particular, the same research into alternate inter-connection methods mentioned in the high-frequency imaging section can also contribute to lowering individual transducer costs.
1.5 Summary of Thesis Contribution

Many high-frequency systems (including all current commercial ophthalmic systems) use single element transducers rather than transducer arrays. These are inexpensive but feature poor depth of field and are difficult to use. Shifting to an annular array transducer could solve the depth of field of issues, but development faces two problems. One, high-frequency arrays are difficult and expensive to build. Second, the associated electronic beamformer systems make the systems too expensive for many of the specialized applications associated with high-frequency imaging.

Ultrasound transducer array systems operating at frequencies above 10 MHz are becoming more common in the commercial market, with most manufacturers producing transducers which operate up to 12-15 MHz. Transducer arrays operating above 15 MHz are less common, with the Acuson 18L6 (up to 18 MHz) and the ATL L19-5 (up to 19 MHz) featuring as those operating at close to 20 MHz. Arrays operating at higher frequencies are scarce, with the VisualSonics line of transducers (six arrays with centre frequencies between 14 and 50 MHz) representing the current high end of array transducers, but with a matching price tag. For example, a high-frequency array system from VisualSonics, the Vevo 2100, costs approximately $450,000. [25]

The system I am presenting in this thesis solves these issues with several innovative aspects. A new transducer design has been developed which is easy and inexpensive to fabricate. Connection is made to a kerfless ceramic transducer using a flexible circuit board. This circuit board also doubles as the front matching layer. Electrical connectivity is achieved using an anisotropic epoxy and a clever electrode grid system avoids alignment issues between the flexible circuit board and the ceramic.
In addition, I developed a low-cost and flexible dynamic receive beamformer based on commercially available circuit boards which connect to the PC via USB cable. By implementing the beamformer in software using a high-level programming language, a high degree of customizability is achieved. This type of software beamformer has only recently become feasible due to increasing USB bandwidth with the USB 2.0 standard and increasing CPU processing speeds.

The system is able to use commercially available circuit boards due to the implementation of a high-efficiency interpolation algorithm I used instead of a standard linear interpolation. This algorithm permits high-quality beamforming with a modest sampling frequency only 4 times the pulse center frequency.

Finally, I developed a low-cost, modular pulse generator and transmit beamformer capable of up to 90 Vpp pulses and an operating frequency range up to 50 MHz. Individual boards for each pulse output allows easy expansion for larger arrays, while allowing un-used channels to be removed, minimizing power requirements.

The completed system serves as a proof of concept for systems built using these techniques and proves their feasibility in keeping costs down and simplifying manufacturing while maintaining system performance.
1.6 Thesis Structure

The second chapter contains an introduction to the fundamentals of the math and physics used in the development of this system. It also contains a brief description of the individual components of the system along with their contribution to the completed system. As a number of modeling systems were used during the development.

The third chapter explains the design process used in the development of the pulse generator and transmit beamformer. Design criteria which impacted the final layout are explained as well as the trade-offs which were necessary. The performance of the pulse generator in software simulations is compared to the final pulses and the limits of the system are discussed.

The fourth chapter presents the annular array transducer and describes both the standard manufacturing methods as well as the innovative techniques used to replace them in this transducer. The arrays characteristics are presented as well as its performance in conjunction with the pulse generator and transmit beamformer.

The fifth chapter describes how commercially available circuit boards were implemented as the analog front-end of the transmit beamformer. The high-efficiency interpolation algorithm is explained in depth and compared to a linear interpolation algorithm. The software implementation of the algorithm and the receive beamformer is explored and its performance capabilities explained.
The sixth chapter discusses the integration of the three main components of the system and presents a system image as well as the overall system capabilities.

Finally, the seventh chapter summarizes the capabilities of the system and its functionality. Potential future areas of research to improve the system specifications are also presented.

1.7 Supporting Publications and Talks

1.7.1 Publications

(Not yet submitted) Lay, Holly; Lockwood, Geoffrey, “A Novel Low-Complexity, Low-Cost Beamformer Implementation”.

(Submitted) Lay, Holly; Simpson, Eric; Griffin, Greg; Lockwood, Geoffrey, “High-Frequency Annular Array Fabrication using a Flex Matching Layer”


Lay, H.S., Lockwood, G. R., “A 64-Channel Beamformer for 50 MHz Linear Arrays”,

1.7.2 Conference Presentations and Invited Talks

(PZT/Silicon) Transducer Arrays for High Frequency Medical Imaging”, 10th Annual
Ultrasonics Transducers Conference, Los Angeles, California, 2010 (Invited Talk)

Wall, K. & Lay, H. S. "Medical Ultrasound Beamforming Design and Implementation”. Third
Ontario Consortium for Small Animal Imaging High-Frequency Ultrasound Workshop, June
2007 (Invited Talk)

Lay, H.S. & Lockwood, G.R. “50 MHz Beamformer for Medical Imaging”, Presentation, 2006
Ultrasonics Biomedical Microscanning Conference.
Chapter 2

Background Material

2.1 Imaging Fundamentals

When an acoustic wave travelling through a medium encounters a change in material properties, be it speed of sound, elasticity or density, the waveform will be reflected, refracted and scattered from the resulting boundary, resulting in a backwards propagating echo wave. The intensity of the reflected wave can be calculated given the properties of the adjacent media. If the materials in the system have characteristic speeds of sound within a small range, assumptions can then be made about the nature of the interfaces as well as the depths at which the echoes occurred.

Acoustic imaging is well suited for use in the human body due to the uniformity of the speed of sound in human tissues. A table of speeds of sound (table 2.1) in mammalian tissue (pork and human) is shown. Though there is some variance depending on tissue type, the speed of sound in the body can normally be approximated to 1540 m/s [25]. Errors induced in images resulting from the variance in speed of sound are known as phase aberrations [25][26]. While the basic imaging techniques discussed here do not compensate for phase aberrations, there is significant work in the field focused on minimizing these errors.
The assumption of constant speed of sound allows several calculations to be performed, in particular the mapping of time to distance. Given a constant speed of sound, the depth of a given acoustic echo can be mapped using the simple kinematic relation:

\[ d = \frac{v \cdot t}{2} \]  

Eqn. 2-1

The factor of a half is caused by the fact that the time from the initial pulse to when the echo arrives is equal to the time it takes for the pulse to travel the depth into the tissue, reflect, then return, doubling the time in transit.

2.2 Envelope Detection

The pulse-echo waveform obtained from an ultrasound transducer can be represented as an amplitude-weighted sine wave.

\[ v(t) = a(t) \cdot \sin(2\pi ft + \theta) \]  

Eqn. 2-2
The grey-scale images used for medical imaging, however, are composed of logarithmically compressed relative pressure data. While the energy used is acoustic, the frequencies are in a range normally referred to as radio frequencies (RF) and, if electromagnetic waves, would fall in the HF band. To convert this RF waveform into intensity information, the amplitude, \( a(t) \), must be obtained. There are multiple methods of obtaining this information, given that the frequency, \( f \), of the wave is known from the originating pulse. The simplest method, and the one most commonly used, is to produce a quarter-period time-delayed version of the RF signal, either using hardware or software [27].

\[
v(t + \frac{T}{4}) = a \left( t + \frac{T}{4} \right) \sin \left( 2\pi f \left( t + \frac{T}{4} \right) + \theta \right) \quad \text{Eqn. 2-3}
\]

However, as the period, \( T \), is equal to the inverse of the frequency, \( f \), the second term of the sine equation can be reduced to \( 2\pi f / 4f \), or simply \( \pi / 2 \). This results in a sine to cosine shift, thusly:

\[
v(t + \frac{T}{4}) = a \left( t + \frac{T}{4} \right) \cos \left( 2\pi ft + \theta \right) \quad \text{Eqn. 2-4}
\]

If \( a(t) \) is assumed to shift slowly with respect to the centre frequency, \( f \), equations 2-2 and 2-4 can then be squared and summed to obtain:

\[
\left( v(t)^2 + v \left( t + \frac{T}{4} \right)^2 \right)^{1/2} = a(t) \quad \text{Eqn. 2-5}
\]

This is known as quadrature sampling and results in a reasonable approximation of the amplitude which can then be logarithmically compressed for grey-scale imaging.

### 2.3 Imaging Frequencies

The term ultrasound can refer to any sound above the human hearing range, or approximately 25 kHz. However imaging with ultrasound is traditionally done at frequencies in the 2 to 10 MHz
range [28]. The exact resolution of any given system can depend on multiple variables, however, due to the inverse relationship between centre frequency and the wavelength of the imaging pulse, for a given ultrasound system, the resolution will get smaller as the imaging frequency increases.

The wavelength of a propagating waveform in a material can be determined using a simple equation, \( \lambda = \frac{v}{f} \) [29], where \( \lambda \) is the wavelength, \( v \) is the speed of sound in the material and \( f \) is the centre frequency of the signal. Thus it can be shown that, for an ultrasound wave at 2 MHz travelling in the human body, the imaging pulse will have a wavelength of 0.77 mm. An imaging system functioning at this frequency could expect a resolution on the order of 1-2 mm [12]. While this is sufficiently small for larger organ imaging and obstetrics, the demand for systems that can be used for small animal, ophthalmic, skin and intravascular imaging has driven development of higher frequency imaging systems, above 20 MHz [9][10][12] [30] [31].

2.3.1 High-Frequency Imaging

While higher frequencies directly improve the image resolution of ultrasound systems, there are a number of trade-offs involved. The first, unavoidable trade-off is signal attenuation. The attenuation of sound in a given material is equal to \( \alpha(f) \) expressed in nepers/unit length. When used in ultrasound applications, this is normally expressed in dB/unit length, which is equivalent to \( 8.686 \times \alpha(f) \). Alpha is characteristic for each material, and for water is a function of \( f^2 \) [3]. However, for most human tissues the attenuation is only linearly proportional to \( f \) [31]. This means that, the higher the centre frequency being used, the shallower the possible image that can
be created. While the effects of this attenuation can be mitigated by increasing the power of the input signal, excessive power can cause localized heating at the contact point with the transducer, potentially injuring the patient. This sets a functional limit on the power that can be used in medical devices. Due to this limitation, higher-frequency ultrasound systems are typically designed for use either in areas close to the surface of the skin (needle placement, ocular and ear imaging) or for insertion into catheters to get the transducer closer to the area of interest.

2.4 Anatomy of an Ultrasound Imaging System

The basic structure of an imaging system can be seen in figure 2.1.

![Image of ultrasound imaging system components]

**Figure 2.1 Components of a Typical Ultrasound Imaging System**

The core of the system is the ultrasound transducer which transforms the electrical input signals from the pulser unit into acoustic pulses, and then converts the return echo pulses back into electrical signals. The received electrified signals are amplified and filtered before being digitized by the Analog to Digital Converters (ADCs). The digital signals are then processed and
displayed as an image. For transducer arrays, beamformers compensate for the time of flight difference between the various array elements and the focal zone. The transmit beamformer can also be located before the pulser in some configurations, while analog receive beamformers will be located before the ADC stage. The next few sections will describe each component of the imaging system in greater detail.

2.5 Transducers

Medical imaging using ultrasound requires the reliable production of short-time pulses at a known centre frequency and the ability to detect the resulting echoes. These are the primary functions of the transducer. Transducers consist of a piezoelectric element which transforms the electrical signals to an acoustic pressure wave and back again, plus supporting components. While there is some variation between devices, most complete transducers include the components seen in this diagram:

![Figure 2.2 Typical Assembly of an Ultrasonic Imaging Transducer](image)

All of these elements have important contributions to the production of a quality ultrasound image.
2.5.1 The Piezoelectric effect

The transformation of electrical signals into pressure wave and back again is achieved by the inverse and direct piezoelectric effect respectively.[31] In a piezoelectric material in its rest state, there are a large number of domains containing dipole moments whose directionality is determined by poling applied during manufacturing. When an electric charge is applied across the material via a set of electrodes, the domains re-orient themselves along the lines of the electric field, resulting in a physical expansion of the material (Figure 2.3(a)).[3] By applying a varying electrical charge across the material, a physical oscillation can be induced, resulting in a pressure wave.

When receiving echoes, the mechanical compression and expansion of the crystal domains result in a change in the electric field, which can be measured as a change in potential (Figure 2.3(b)).
Figure 2.3 Demonstration of the Piezoelectric Effect (a) The Inverse Piezoelectric Effect (b) The Direct Piezoelectric Effect

Standard ultrasonic transducers used in medical imaging consist of a piezoelectric material sandwiched between electrodes to apply and detect changes in the electric field. The most common piezoelectric material used in ultrasonic transducers is lead zirconate titanate or PZT. PZT is favourable for the use in ultrasound due to its high dielectric constant and relative ease of handling.

2.5.2 Material Characteristics

By applying shear/strain analysis to the physical distortions of a piezoelectric material, we can derive several material characteristics which can be used to compare various piezoelectric materials, as well as assist in the simulation of transducer designs.
Earlier in the thesis, the reflection intensity from a boundary between materials was described in terms of the characteristic acoustic impedance of each material. This quantity, $Z$, is defined as the ratio between the pressure applied to the material and the corresponding particle velocity induced by that pressure, and is measured in Rayls. It can also be calculated via the product of the material density and the sound velocity in the medium [31]. This value will be used particularly when determining the matching and backing layers best matched to the piezo-ceramic.

One of the most frequently quoted properties of a given piezoelectric material is the coupling factor, or $k_t$. The coupling factor, expressed as a unit-less ratio, is an expression of the ability of the material to transform energy from one form to another [32]. While there are several related, but not equivalent, definitions of $k_t$, one of the more common ones is “[…] equal to the square root of the fraction of energy converted from the electrical domain to the mechanical domain (or vice versa) in a single electromechanical cycle.” [31] Thus, a high coupling factor is highly desirable for imaging purposes, as the higher the coupling factor, the less electrical energy is required to obtain the same magnitude pressure wave.

### 2.5.3 Matching Layer

The characteristic acoustic impedance of PZT is approximately 30 MRayls. Tissue, by contrast, has an acoustic impedance on the order of 1.5 MRayls [33]. To avoid this reflection, one or more matching layers can be used. While a single matching layer with acoustic properties equal to the
geometric mean between the piezoelectric and the medium is quite common [34], multiple matching layers are also used [34]. The geometric mean of the impedances can be calculated as:

\[ Z_m = (Z_{pzt}Z_{issue})^{1/2} \]  

Eqn. 2.6[35]

Actual matching layers can vary from this value and should be modeled prior to implementation to best judge the acoustic transmission characteristics.

An ideal matching layer has a thickness equal to one quarter of the wavelength of the acoustic pressure wave. This thickness is a function of both the centre frequency of the acoustic wave as well as the speed of sound in the matching layer, forcing the matching layer thickness to be customized for each transducer design.

2.5.4 Backing Layer

Like the matching layer, the backing layer must be well matched to the piezoelectric material to avoid multiple reflection of the pressure wave. However, if a perfectly matched material is used as a backing layer, most of the acoustic pressure is lost into the backing, resulting in poor sensitivity for the transducer. Thus, there is a trade-off between excessive ringing in the piezoceramic and loss of sensitivity. In addition to meeting the trade-off requirements for acoustic matching, the backing layer must also be a good acoustic absorber to prevent energy from being reflected back to the piezoelectric after reflecting off the back of the transducer. While some researchers have put significant effort into the design of a good backing layer [34][36], if a simple solution is desired, the most common backing is a tungsten-loaded epoxy [29].
2.5.5 Interconnection Methods

Finally, the front and back electrodes must be connected in some manner to an external cable to allow connectivity to the rest of the imaging system. One method of doing this involves the bonding of miniature coaxial cables to the electrodes on the piezoelectric using wire-bonds. In addition to the difficulties inherent in wire-bonding were mentioned earlier, this also suffers from issues of physical bulk when arrays are considered. Even micro-coaxial cables become bulky when there are 64-256 of them, which has led to research into replacements, such as flexible circuit boards [37].

2.5.6 Focusing

In some systems a physical lens is included outside the matching layer to focus the beam. For a single-element system, either a lens or physical curvature of the piezoelectric is required to achieve a focused beam (Figure 2.4). While relatively accurate, physical focusing is fixed after manufacture, resulting in a relatively shallow depth of field and the inability to adjust the imaging depth for specific applications [38]. To avoid these limitations, a transducer array can be used instead.
2.6 Transducer Arrays

To avoid the difficulties inherent in mechanical focusing, an annular array design can be used. Here the single-element transducer is split into multiple concentric rings. The most common designs for these arrays feature a single monolithic electrode on the front face of the piezoelectric material to act as a ground. The piezoelectric is then divided into the concentric rings by cutting gaps all the way through the material and filling with a material selected for its ability to impede shear waves along the length of the transducer. These cuts are known as kerfs. The separated rings are then coated with individual electrodes. While the relative size of the electrodes is variable in these designs, it is advantageous to use what is called an equal-area layout. In this pattern, the array rings are designed to decrease in width as they increase in outside diameter so as to keep the total surface area of each ring equal. The main advantage of equal-array designs is that it produces the best radiation pattern using the fewest possible rings. A positive side-effect
of equal-area design is that the rings also possess approximately equal impedances, simplifying load calculations for the pulse generator and receive circuitry.

Annular arrays avoid the need for mechanical focusing due to the fact that each of the array rings can be stimulated with a different electrical pulse. By generating a pressure wave first from the outer-most ring and then successively inwards, the pressure wave can be shaped into a concave wave focused at a particular point. This is known as transmit beamforming, and will be discussed at greater length in the beamforming chapter.

Linear or phased arrays are used when the ability to control the formation of image lines electronically and the elimination of motors are of higher priority than electronic complexity. In these arrays, the same piezo-materials used in single element and annular array transducers are diced into a rectangular shape. The array elements are then sub-divided and electrodes attached, similar to an annular array, but rather than in rings the elements are laid out as rectangles aligned linearly. The spacing of the elements is determined by whether it is a linear array, with ultrasound beams only directed at right angles to the array face, or a phased array, which is steered at a 90 degree angle to the face.

The relative spacing of array elements dictates the angular position of the grating lobes. These are undesirable areas of constructive interference which can cause ghost images if present in the imaging area. Due to the steering of the phased array, the elements must be spaced half a wavelength or less apart to push the grating lobes 180° away from the main lobe. Linear arrays
can have elements only a wavelength apart, as the 90° grating lobes will not be rotated into the imaging area by off-angle steering.

While arrays allow a focused beamline to be produced at all desired position in the plane of the image in a manner similar to a mechanically scanned annular array, in order to focus along the width of the array, either physical shaping or a lens must be used. These arrays have the advantage of being easier to curve or fit with a lens than single element transducers, due to the curvature being limited to a single dimension.

2.7 Pulsers

The pulse generator, or pulser, is an electrical circuit used to excite the transducer. After receiving a trigger signal from the master system controller, the pulser emits a short pulse or series of pulses with a known amplitude and duration. These produce physical oscillations in the ultrasound transducer, which propagate into any neighbouring medium as acoustic waves. Pulsers are commonly designed with the ability to modify the amplitude in real time, as this affects the total signal in the system [39][40]. To maximize the bandwidth of the resulting acoustic pulse, high-quality pulsers produce pulses with very sharp voltage transition and a rapid return to zero. A longer return-to-zero period can result in extra ringing in the transducer, lengthening the resulting pulse and degrading the image resolution.

For transducer array systems, a separate pulser circuit is required for each array. To control the relative delays of the pulsers, a transmit beamformer is required. While single output pulsers are
commonly available on the commercial market, pulsters capable of outputting multiple pulses in response to a sequence of beamformed trigger signals are rare, and individual research groups commonly produce their own pulsters for internal use.

### 2.8 Beamformers

Beamforming refers to the process of adjusting the delays in electrical signals between the individual elements of an array to focus the acoustic beam at a given depth and location. To achieve this, the signals must be adjusted both during transmission and when processing the received signals. Since the theory involved is similar for both annular and linear arrays, it will be demonstrated with a linear array as its planar nature simplifies the diagrams.

![Figure 2.5 Beamforming of Signals to Focus Transducer Arrays](image-url)
The electronic focusing of an ultrasonic array starts with the pulses transmitted from the transmit beamformer. Figure 2.5a shows that if each element of the active aperture is pulsed simultaneously, the resultant wave is planar, with the wave front parallel to the face of the transducer. To shape the wave front, the pulses to the individual elements of the array are delayed with respect to each other (figure 2.5b). As most imaging systems are designed to focus the waves to a particular point, the outer elements of the aperture are pulsed first, followed in succession by the elements next to them until the centre element fires last. It should be noted that if the delays used are symmetrical around the centre element, the resulting wave will be focused at a point directly in front of the centre element. However, if the delays are increased on one side and decreased on the other, the wave can be focused along a line at an angle to the transducer. This technique is used to steer the phased array.

The highest resolution for an ultrasound system is obtained when all the elements are focused at a single point in the imaging field. The appropriate delays can be obtained using some simple geometry and knowledge of the speed of sound in the material and any curvature or lens on the array. First, each element is treated as a point source located at the geometric centre of the physical element. It can be shown that this is a reasonable assumption so long as the width of a given element is significantly smaller than the distance from the element to the desired focal target. With this simplification, the distance from each element to the target can be calculated. As the delays used in transmit beamforming are additive delays (i.e. each pulse is delayed with respect to the first pulse), the difference between these distances and the distance from the outermost element(s) is then calculated. This gives the difference in travel distance for the acoustic waves that must be corrected for using electronic delays (figure 2.6).
Figure 2.6 Diagram of the difference in travel distance between array elements

For the purposes of beamformer calculations, the speed of sound in the human body can be assumed to be equal to 1482 m/s. This can be approximated to 1500 m/s when designing a medical ultrasound system. Using this value, the differential distance values can be converted into delays quite simply.

2.9 System Modeling

To facilitate the design of an ultrasound system, it is standard to model the components in a computer aided design tool prior to manufacture. Which tools are used depends on the system elements being modeled and the accuracy of these models must be verified before the results can be used in the final design.

2.9.1 SPICE

There are many software tools available to model electronic circuits to determine their behavior under both standard and extreme conditions without risking destroying components. When
modeling discrete electronic circuit boards, the most common tool used is known as Simulation Program with Integrated Circuit Emphasis (SPICE). SPICE is a circuit simulation tool which has been in continuous development for over thirty years, and SPICE models for most electronic components are available directly from the manufacturers [41]. The most commonly used implementation of SPICE is pSPICE. This is a graphical program that allows designers to build up circuits in an identical manner to hand-drawn circuits. Using the manufacturers’ SPICE models for the individual components, the expected behavior of the circuit can be calculated for various inputs. SPICE was used to model the behavior of the pulser circuitry described in chapter 3, and will be discussed in greater detail at that point.

2.9.2 Two-Port Networks

While useful for modeling pure electrical circuits, pSPICE is of limited use in mixed models, such as systems involving acoustic components, such as ultrasound transducers. To model these components, there are two common solutions, depending on the desired level of analysis. For simpler analysis, the various components of the system can be represented using two-port networks [42]. A two-port network is a simplified model which replaces each discrete component in the system with a simple box with a known relationship between its input and output (figure 2.7).
There are multiple varieties of two-port network in use in circuit analysis. This thesis uses the transmission line, or ABCD-parameter model. For the given two-port variables, the ABCD-parameters can be defined as follows:

\[
\begin{bmatrix} V_1 \\ I_1 \end{bmatrix} = \begin{bmatrix} A & B \\ C & D \end{bmatrix} \begin{bmatrix} V_2 \\ I_2 \end{bmatrix}
\]

Eqn. 2-7

It should be noted that there are two generally used definitions for the ABCD-parameters, depending on the defined direction of \( I_2 \). To calculate the parameters of a network, it suffices to evaluate the circuit with the output ports first shorted together, then left open-circuited. This allows all four parameters to be calculated as a function of input and output variables.

\[
A = \frac{V_1}{I_2} \Bigr|_{I_2=0}
\]

Eqn. 2-8

\[
B = \frac{V_1}{V_2} \Bigr|_{V_2=0}
\]

Eqn. 2-9

\[
C = \frac{I_1}{V_2} \Bigr|_{I_2=0}
\]

Eqn. 2-10

\[
D = \frac{I_1}{I_2} \Bigr|_{V_2=0}
\]

Eqn. 2-11

The ABCD parameters are particularly useful when analyzing mostly linear circuits, as they can be chained together after calculation. Due to the chosen direction of \( I_2 \), it can be seen that when
two networks are connected together sequentially, \( V_1 \) of the second network is equal to \( V_2 \) of the first network and similarly \( I_2 \) of the first network is equal to \( I_1 \) of the second.

![Diagram of two networks connected sequentially](image)

**Figure 2.8 Connecting Two-Port Networks**

This allows the parameters for the second networks to be plugged into the output of the first, resulting in multiplication of the parameter matrices.

\[
\begin{bmatrix}
V_1 \\
I_1
\end{bmatrix} = 
\begin{bmatrix}
A_1 & B_1 \\
C_1 & D_1
\end{bmatrix} 
\begin{bmatrix}
V_2 \\
I_2
\end{bmatrix} \quad \text{Eqn. 2-12}
\]

\[
\begin{bmatrix}
V_2 \\
I_2
\end{bmatrix} = 
\begin{bmatrix}
A_2 & B_2 \\
C_2 & D_2
\end{bmatrix} 
\begin{bmatrix}
V_3 \\
I_3
\end{bmatrix} \quad \text{Eqn. 2-13}
\]

\[
\therefore \begin{bmatrix}
V_1 \\
I_1
\end{bmatrix} = 
\begin{bmatrix}
A_1 & B_1 \\
C_1 & D_1
\end{bmatrix} \begin{bmatrix}
A_2 & B_2 \\
C_2 & D_2
\end{bmatrix} 
\begin{bmatrix}
V_3 \\
I_3
\end{bmatrix} \quad \text{Eqn. 2-14}
\]

Thus a sequence of linear components with known parameters can be simulated numerically simply by multiplying together the parameters.

While these calculations do not include parasitic effects present in physical circuits, noise levels can be approximated by modeling each known noise source individually and recalculating the bulk two-port network with the relevant load conditions [43]. The additive effects of noise and their effects on system resolution will be discussed in greater detail in chapter 6, as their effect is most noticeable when all components of the system are brought together.
While two port networks were originally derived for electrical circuit analysis, an existing model can be modified to accommodate acoustic components using equivalent circuits [44], which were used when modeling the ultrasound transducer array.

Two port networks were used in the modeling of the transducer array output pulse shape and bandwidth, and the specific model elements will be explored in greater detail in chapter 4. The use of two-port networks to model electrical circuits is known as KLM modeling after the first group to present their use [45].

2.9.3 MatLab™

While KLM modeling can be performed in any mathematical environment, MatLab™ is an analytical software package well suited to simulations combining KLM circuit elements with desired software analysis. While the analog aspect of the pure circuit simulations would be better represented in a non-discrete environment such as Maple™, MatLab™ can more accurately represent the digital logic used to implement the software components of the complete ultrasound system.

MatLab™ is a high level programming language and simulation environment used extensively in the engineering community. The design of the environment allows various components of an ultrasound system to be simulated at varying levels of complexity, depending on what variables are being tested. As an example, the signal flow of the complete system could be tested by
representing the transducer array and the analog receive electronics with KLM models. The input can be a finely sampled data array from either a pulser design simulated in pSPICE or from a functional hardware device. The resulting output permits a good estimate for the expected losses in the system due to system components, and acoustic losses can be approximated by varying parameters of the KLM model.

MatLab™ can also be used to process data in real-time. By taking advantage of a reduced instruction set known as embedded MatLab, calculations can be done quickly and with low overhead. Using MatLab has the additional advantage of adaptability, due to the high-level nature of the coding environment; fine tuning of the model is easily and quickly implemented.
Chapter 3

Pulser and Transmit Beamformer

Every ultrasound imaging system, regardless of transducer technology, requires a good quality pulser circuit. The purpose of this circuit is to supply a high-voltage pulse which will initiate ringing at the resonant frequency of the transducer, resulting in the initial outbound pulse whose echoes will be resolved in the final image.

3.1 Types of Pulsers

The simplest form of pulser used in ultrasound systems takes the form of a simple switch connected to a high-voltage power supply and system ground. This allows the user to manually initiate a single pulse of the transducer. To better control the length of the voltage pulse, the switch can be designed to release after a set period of time, resulting in a voltage pulse with consistent characteristics. This will produce a short pulse of ultrasound centred on a frequency $f$, where $f = \frac{1}{2T}$, with the period, $T$, equal to the length of the pulse, provided $f$ is within the frequency bandwidth of the transducer.

While simple to implement, this type of pulser has an innate DC bias due to never going negative. This can result in undesired leakage current.[46] To avoid this leakage current, a bipolar arrangement can be used. Bipolar pulsers feature a trade-off between amplitude and pulse length. Due to the additional negative pulse applied to the transducer, the transducer will ring longer, resulting in a longer total pulse and lower axial resolution. However, by applying a second
driving pulse to the oscillation, the amplitude of the pressure wave will be approximately doubled, improving the signal level without increasing the noise. This makes the trade-off a viable option for pulser design, and allows some control over bandwidth and centre frequency by shifting the centre frequency of the excitation pulse.

To obtain a bipolar pulse, a multiplexer switch can be used with an additional connection to negative voltage. Alternatively, the internal architecture of the multiplexer can be replicated with discrete transistors, allowing greater control on the amount of amplification and switching time in the circuit. This latter method is preferred when designing high-frequency pulsers due to the requirement for switching times under ten nanoseconds.

An additional circuit known as a clamping circuit is also seen in some pulser designs [47]. This circuit uses a second set of transistors tied to the same output as the main pulser. However, where the main pulsing transistors are tied to positive and negative voltages, the clamping transistors are tied to ground. To use the clamping circuit, an activating pulse is sent to these grounding transistors just after the pulsing transistors are deactivated. The clamping transistors then supply an additional current path to ground, allowing the output to return to zero with less ringing, and a cleaner pulse.

While the basic design for a good quality pulser has not changed in decades, there have been significant advances in the quality of amplifiers for the initial stage as well as the power capacity and response time of the final output stage transistors. In addition, in the last few years single chip pulsers have become available commercially [48]. Currently available for output frequencies
up to 20 MHz, it is likely that subsequent generations will push this upper limit into the range of true high-frequency ultrasound. Until such chips are released, however, it is still necessary for designers of high-frequency systems to build their pulsers from discrete parts.

3.2 Design Criteria

There are only a few critical criteria for the design of a high-quality ultrasound pulser. First, the pulser must be capable of supplying sufficient power to the output to drive a $<100 \ \Omega$ transducer element with anywhere from 40-200 Vpp. The output stage therefore needs to switch between sinking up to 1 amp of current from the positive power supply to sourcing the same amount of current to the negative supply. For a high-frequency system (>20MHz), this change-over occurs in less than 25 ns. This leads to the second criterion of a rapid rise/fall time.

Specifically, the final amplification stage must have a rise/fall time from positive to negative full voltage of less than half the period of the pulse frequency. This stage must also recover from the negative half of the pulse as quickly as possible. This is important, as any over- or under-damping of the return to zero could result in excessive ringing of the transducer, lengthening the outbound pulse and degrading the system resolution. To facilitate this rapid return to zero, some pulser systems add a clamping circuit as described above. After the pulse, this circuit is turned on to pull the output channel to ground, acting as an additional current sink.
3.3 Circuit Design

While the basic circuitry involved in the implementation of a pulser is well-known and does not vary much from one application to another [46][49], the design must still be laid out and simulated to identify any issues with the amplification from signaling levels to high-voltage pulses, as well as to mitigate the noise from the output pulse. The parts selected must also be tested to determine if they will perform adequately under the specific loading conditions of the circuit.

3.3.1 Previous Pulser Design

When designing the pulser circuit, due to the commonality of the basic requirements, we started from an existing design which was known to work up to 50 MHz and 80 Vpp, supplying more than sufficient power and response time for our design requirements of 25 MHz and similar output voltage ranges. The pulser in question was designed by Jeremy Brown and can be seen in Figure 3.1.
Figure 3.1 Initial Bipolar Pulser Design as Laid Out in pSPICE for Simulation.[50]

It is a simple 3 stage pulser with a single op-amp gain stage and two transistor-based gain stages. The input to this circuit is a pair of short pulses at TTL voltage levels with the second pulse delayed by half a period with respect to the first. Not shown is a half-period delay stage used to drive the negative half of the pulse. In the original design, this delay was supplied with a carefully cut coaxial cable. While having the advantage of being extremely robust, this has the side effect of locking the pulser to a given centre frequency.

Examining the pulser circuit, it can be seen that the initial op-amp stage serves the dual purpose of a small amplification as well as inverting the pulse to the negative half of the circuit. The
second and third stages each add amplification with the final stage being built using high power transistors capable of sourcing and sinking the required current. The capacitors shown between the second and third stages act to preserve the DC bias of the final stage and were selected to achieve a knee frequency below the design frequency of the circuit.

3.3.2 Pulser Redesign

While it would have been desirable to change as little as possible from a design which was known to work, the transistors used in the positive half of the circuit have since become obsolete, necessitating a redesign. To help with the redesign, a pSPICE model of the old circuit was built up using transistor models supplied by the manufacturers. This allowed various replacement components to be tested against each other on a common platform.

Initially, the original design was preserved almost unchanged, with only the obsolete parts replaced with the closest analog parts available from the manufacturer. Although the specifications of the replacement parts were very similar to the previous parts, the simulated circuit displayed DC biasing issues, as well as inferior performance. It is unclear why the replacement parts did not perform at the same level as the obsolete parts, but the decision was made not to depend on behaviour that could not be replicated in the simulation environment.

Based on the specifications of the original circuit, several variations were simulated with parts currently available. In the process, the circuit was simplified to minimize possible behaviours not modeled in pSPICE. The final circuit was reduced to a two-stage system, as the driver transistors
proved sufficient for the necessary amplification as well as reducing the DC biasing requirements (Figure 3.2).

Figure 3.2 Revised pulser circuit with 2 stages and adjustable output voltage and input delay

The circuit was tested at drive voltages up to 100 Vpp and demonstrated a high tolerance for variation of the output drive voltage. The simulated performance of the pulser is shown at 50 MHz. The pulse shows more than adequate rise/fall times as well as minimal ring-down (Figure 3.3).
Unfortunately, many of the characteristics that can lead to poor ring-down performance are related to transistor behaviour that is not always well modeled, as well as the layout of the circuit board. In addition, the rise/fall times of the circuit can also be affected by the quality of the input pulses. During simulation, pulses with 1 ns rise and fall times were used. The true rise/fall times of the circuit cannot be seen until it is implemented in hardware. Thus, to properly assess the performance of the circuit, a test board was necessary.

### 3.4 Printed Circuit Board Design

To house the circuitry of the pulser design, as well as later circuits used for this thesis, printed circuit boards were used. A printed circuit board is simply a board of insulating material, originally rigid, but now also available in flexible options, alternating with layers of conductive metal. Connections are made from metal layer to metal layer by means of drilled holes plated in metal, known as vias. Printed circuit boards can have as many as 25 or more metal layers, or as few as one, however, due to the layering necessary for more metal layers, these multi-layered
boards are significantly more expensive to produce. On the other hand, simple one and two layer boards, which only feature metal on the exposed surface of the insulator, can be manufactured with a fairly simple set-up, allowing multiple test boards to be produced at low cost. The design of high-quality printed circuit boards has been heavily investigated due to their ubiquitous use in the computer industry and a number of rules of thumb have resulted for good quality results.

3.4.1 High-Voltage Design

While it is relatively simple to design a functional circuit board for low-power, low-frequency applications, increasing either the power and/or the frequency of the signals requires additional design rules to prevent noise and delay issues which will lead to inferior or, at worst, non-functional circuits.

High-voltage design is tricky in part due to the larger currents involved. The metal layers are commonly deposited in thicknesses of between 0.5-1 mil [51] (1/1000 of an inch) and have a maximum current tolerance per unit volume before trace heating and gap arching of current become concerns. The second critical aspect is that of signal coupling. This is when two traces on the circuit board experience parasitic capacitance and the signal from one appears in an attenuated manner as noise on the second line. While good circuit layout can decrease parasitic capacitance and heavily attenuate coupled signals, significantly more care must be taken when the signals being coupled are high-voltage, rather than the more typical 1-5 V, as the coupled signals are become respectively larger.
With these factors in mind, a few rules of thumb have been developed to maximize circuit performance working at high-voltages. The first rule of thumb is to divide the circuit as much as possible into high- and low-voltage areas. For the pulser design presented in this thesis, the division can easily be made between the first stage, at low voltages, and the second stage, at high voltages. This division is to avoid the parasitic coupling mentioned early, by separating sensitive low-voltage signal lines from the higher-voltage potential noise sources. A safe practice would be to separate the circuits onto individual boards, however the voltage levels are low enough that this is a trade-off with the desire to keep traces short, which will be expanded on in the discussion of high-frequency rules.

The second design rule of thumb is that of sufficient ground contacts and low resistance runs to them. When sinking large currents to ground, it is important that the current path be short and low resistance to avoid significant voltage drop across the trace. If the resistance of the trace is too high, the voltage at the contact point for the circuit components will suffer from a DC bias away from ground, impacting circuit performance. The easiest way to minimize this resistance is with a ground plane. To form a ground plane, an entire plane of metal within the circuit board is reserved solely for ground contacts, with only gaps for connections between the other layers. The ability to produce ground planes is one of the largest advantages of multi-layer printed circuit board design.

It should be noted that the same design rules that impact ground connection and minimizing resistance also apply to power connections. Unlike ground planes, however, power planes should
be restricted to the part of the circuit board dedicated to that voltage level as laid out in the first rule.

3.4.2 High-Frequency Design

Dealing with high-frequency, or RF signaling on a printed circuit board, like with high-voltage work, can lead to circuit behaviour not seen in simulation unless good design practices are maintained. Some RF design practices complement the high-voltage rules, while others directly contradict each other resulting in trade-offs having to be assessed.

The first rule of thumb is one in conflict with the high-voltage rules, which is to keep all signal traces as short as possible. This can cause difficulties with keeping proper high-voltage/low-voltage separation and leads to a compromise where traces are as long as they can be without requiring strict impedance and trace length control. The general rule for trace length beneath which RF effects are negligible is when the propagation time for a signal in the trace is less than half the rise or fall time of the signal being propagated [52]. For a copper trace exposed to air, such as traces on the surface of a circuit board, the propagation time is ~0.085 ns/in [53]. Therefore, for a signal with a rise time of 1 ns, the traces must be kept less than 5.8 inches in length to avoid RF effects. To provide a proper safety margin, this design will keep all traces under 2 inches in length. This precludes separating the high and low voltage components of the circuit onto two boards. Instead the design will do as much as possible to separate the different voltage zones spatially while maintaining short traces.
The second factor which affects RF circuit performance is that of ground bounce. Ground bounce takes the form of an attenuated form of the main RF signal, but at the ground contact of the circuit and is caused by the same long ground connections which cause DC bias concerns in high voltage applications. Ground bounce can then become coupled to all other components in the system, inducing considerable noise in the circuit [54]. Unlike in high-voltage applications, however, the high-frequency aspect of this coupled noise can be used to help isolate it. Good RF design calls for the use of multiple decoupling capacitors near all RF integrated circuits and near the power and ground inputs. The decoupling capacitors act as a low impedance path from power to ground for high frequency noise signals, minimizing coupled noise and prevent the noise from propagating around the ground and power planes [55].

Respecting these design rules of thumb can make the difference between performance resembling that seen in simulations, and noisy output signals with ringing issues which would propagate throughout the system. In this application, the need to observe both high-voltage and high-frequency rules will lead to some compromises which affected the final board design, as will be explained in the next section.

3.5 Pulser Board

Due to the high power requirements as well as rapid transition times, good circuit design requires a multi-layer circuit board with a good ground plane to sink the required currents. To design this board, OrCAD PCB Designer was used. This software is designed to work in unison with Pspice, making the conversion of the circuit to a board layout simpler.
To be able to test as many final circuit aspects in the test board as possible required the consideration of several additional factors, the first of which was the question of how to supply the delayed pulse to the negative half of the circuit. If the circuit’s frequency and voltage flexibility could be maintained from the computer model it would make the final pulser far more useful, with application to multiple systems. The final decision was to incorporate the generation of the half-period delay into the transmit beamformer circuitry, and test the pulser board with two pulses delayed a pair of function generators.

As this board was only intended for use in testing pulser functionality, it was manufactured in-house with only two metal layers. While the in-house manufacturing method limits the ability to create full power and ground planes, the pulser circuit is simple enough that all signal lines could be confined to the top layer of the board, leaving the bottom layer for ground and power routing. The significant saving in time and money made in-house manufacturing the correct decision for this board.

In addition to general circuit behavior, a critical question that the board was designed to answer was the question of noise coupling to adjacent output lines. To minimize the final size of the pulser and transmit beamformer circuit, it would be desirable to place as many pulser on the same board as possible. However, for the reasons given in the design rules section, there was concern that coupling would occur between output lines laid out too close to each other. Due to the delays between pulser outputs, the output lines are particularly sensitive to coupled noise. Any output coupled to a second line would result in a secondary pulse, either before or after the
main pulse, resulting in additional ringing in the transducer element. For this reason, the test board was laid out with two identical pulser circuits, so that signal coupling could be tested.

### 3.5.1 Test Board Manufacturing

The simple design of the pulser circuit allowed the test board to be manufactured in house. This was done by laying out the design in a software package (Lasi 7) which allows to-scale printing. By printing out the copper trace artwork onto clear acetate, copper-clad boards coated in photo-resist were used to manufacture the boards in a day. It should be noted that, because this process does not include a protective solder mask as commercial boards do, care must be taken to avoid shorting exposed traces between the pads. However, the exposure of the traces does facilitate debugging, as the voltage levels in the circuit can be probed at all points.

### 3.5.2 Isolated Pulser Testing

The pulser test board was populated with components and twisted pair cables were attached to the power and ground planes (Figure 3.4). The inputs were supplied from a Complex Programmable Logic Device (CPLD) programmed to output square pulses with delays of multiples of 5 ns, and a stack of DC power supplies. Once small errors in the copper and component placement had been corrected, a satisfactory pulse was obtained (Figure 3.5).
Figure 3.4 The prototype pulser board with two pulser circuits laid out with all components populated
Due to operating at the limits of the rise/fall time for the operational amplifiers, the positive pulse can be seen to have a longer duration than the negative pulse. However, this only results in a small distortion of the bandwidth and full frequency performance is still observed.

To test crosstalk between the channels, the CPLD was set to output two 5 ns trigger pulses, 5 ns apart. This delay is consistent with beamforming delays in a full ultrasound system and will show evidence of any undesirable interference due to the phase delay. When these delayed pulses were used as inputs to the 2 channel board, significant crosstalk was seen between the channels. This suggested that to avoid cross-talk in the final system, the channels would either need additional separation, increasing the size of the boards, or would need to be on separate circuit boards. Based on the existing size of the pulser circuit, the decision was made to isolate each
channel of the pulser on an individual circuit board and connect them together with a single backplane.

With the decision to separate the channels onto separate boards, investigation began into how to implement the transmit beamformer, and how much to integrate it into the pulser system.

### 3.6 Transmit Beamformer

There were several options available for how to implement the transmit beamformer. It could be implemented on a Field Programmable Gate Array (FPGA) chip using the Verilog programming language. This would have the advantage of being more readily modifiable; however the minimum pulse length and delay difference would be limited by the clock speed. In addition, circuit layouts for FPGAs feature significant complications and the small pitch of FPGA footprints can significantly increase manufacturing cost. FPGAs are also higher priced components, adding another potential cost issue.

A simpler solution to a full FPGA would be a CPLD such as the one used to test the original pulser boards. However, commercially available CPLD boards are limited to clock speeds of 100-200 MHz. As CPLD outputs are clock synchronized, this would limit delay resolution to 2.5-5 ns. This could result in a >5-10% phase error in the resulting pulses, depending on the desired delays. Since transmit beamformer delay is not the only source of delay error in a complete imaging system, and will be additive with other noise sources, this degree of delay is unsuitable for this application.
To obtain greater delay resolution, discrete delay stages were investigated. To be useful as a transmit beamformer, the delay chips would need both sufficient total delay as well as a small enough delay step. For the transducer presented in this thesis, the maximum total delay would be applied to the centre-most element when focusing close in to the array face. At 10 mm, this corresponds to a total delay of <200 ns. Determining the minimum step is slightly more complicated, as to achieve perfect time delays would require a minimum step on the order of 0.1 ns. However, the minimum step and total delay of a delay chip are related through the number of control bits used to set the delay. Most delay chips feature 8 or fewer control bits. This results in 256 delay steps, or a total delay equal to 256 times the minimum delay. This means that a delay chip capable of 0.1 ns delay steps would only have 25.6 ns total delay available. This leads to a trade-off between delay accuracy and the closest distance to the array at which the system can be transmit focused.

### 3.6.1 Transmit Beamformer Design

After comparing the available options, the decision was made to use the DS-1023 series supplied by Maxim Integrated Products. This family of ICs has two significant advantages: first, the line features products with multiple step sizes all supplied in the same package. This allows the same circuit board layout to be used regardless of the final design frequency, as the delay chips can simply be swapped out for chips with a higher total delay or lower step size as required. In addition, the chips have two outputs and two different output modes. In the standard mode, the
first output is the same as the input, except delayed by the desired time. The second output is a pulse which goes high when the input goes high and goes low when the output goes high.

The original design purpose of this output is to allow precise measurement of the output delay, as each chip is prone to internal delays in addition to the intentional delay. This functionality is not necessary in this application as the relative delay between chip is minor compared to the delay step size, which is the primary source of phase error. However, by using a delay chip with a fine delay step (the DS1023-25, with a 250 ps delay step), a pulse can be produced at this second output with precisely the desired pulse width for a given pulser frequency [56]. This significantly reduces the requirements on the input to the transmit beamformer, allowing a single low-frequency trigger pulse to act as input to the entire system.

The selected chips have the additional advantage that they are programmable either serially or in parallel. This is particularly useful for prototyping, as the chips can be set to parallel input for debugging, avoiding potential errors in the serial input sequence. The serial programmability is desirable for any commercial applications of this circuit, as it minimizes the number of input signals to the board and allows real-time control of the focal depth.

Implementing these chips, a transmit beamformer and pulser board was laid out using OrCAD PCB Designer. The DS1023-25 was implemented as the pulse source with a single coaxial input trigger. To increase the total delay range without sacrificing delay resolution, an additional delay chip was also implemented. The DS1100 series chips (Maxim Integrated Products) are tapped-delay devices with five available outputs with fixed delays between each [56]. The first and last outputs were connected to the inputs of a 2:1 multiplexer controlled by a switch. The multiplexer
allows a set coarse delay of 10-200 ns (determined by which chip in the DS1100 product line is inserted) to be added or removed with a switch change. The output of the multiplexer was then fed to two separate DS1023 delay chips. These supply the necessary input pulses to the existing pulser circuit (Figure 3.6).

Figure 3.6 Circuit layout for the transmit beamformer implementation, with DS1023 serving as delay/pulse generator chips and DS1100 supplying the coarse delay

The total delay of the resulting pulse is determined by the sum of the coarse delay setting, plus the delay of the positive pulse delay chip. The period is determined by twice the difference between the delay in the positive pulse and the negative pulse. If desired, this circuit can also produce negative pulses so long as the negative pulse delay is set to less than the positive delay.
3.6.2 Pulser/Transmit Beamformer Manufacturing

Due to the increase in the number of power supplies, and the two-layer limitation of the in-house manufacturing, the decision was made to keep the power connection points as close to the components as possible and make bulk connections via twisted pair cable. There were also concerns that modifying the existing pulser circuitry could lead to unforeseen noise issues, requiring a redesign for the final implementation.

The circuit board was designed with blocks of miniature dipswitches to control the delay chips as well as switch the coarse delay in and out. The pulser circuit was preserved unchanged from the prototype and the transmit beamformer circuitry added to the existing layout. Another addition to the board was the inclusion of a 2 kΩ potentiometer as a voltage divider between the delay chips and the pulser input. The divider was necessary as the pulses output from the DS1023 chips have a 0-5V output swing. By contrast, the pulser circuit requires a lower voltage input to avoid putting the op amps into non-linear behavior. By adjusting this divider, the relative amplitudes of the positive and negative pulses could be controlled, improving overall pulse shape.

The final component added to the boards was an expander at the output of the board. Shown in figure 3.1, the expander is a pair of diodes connected in opposing directions which only allow current flow for high voltage signals. Due to the 0.7 V turn on voltage of a standard diode, any signal applied to the output line less than 0.7 V above or below the ground level will see an open circuit. This allows the pulser to correctly output to the transducer, without loading the transducer during the low amplitude receive part of the imaging cycle. The corresponding
component for the receive side of the circuit is a limiter, and will be explained in greater detail in chapter 5.

The final layout of the board as displayed in the design software can be seen in figure 3.7. Some overlapping of components can be seen on the left side due to the DIP switches on the back of the board being visible as well as the delay circuitry on the front.

![Computer layout of the final transmit beamformer and pulser circuit](image)

**Figure 3.7 Computer layout of the final transmit beamformer and pulser circuit**

Due to the inner power and ground planes included in the final board design, the boards were manufactured commercially in a 6-layer board. The layer stack-up used is as shown in figure 3.8
and follows design guidelines for layout of ground and power planes. In particular, the pairing of ground planes with power planes creates additional decoupling capacitance between the ground planes and each of the power planes, which add to the capacitors on the board itself.

**Figure 3.8 The sequence of metal layers in the pulser and transmit beamformer printed circuit board (not to scale)**

The photograph of the final board with all components populated and twisted pair power supply cables trimmed to length is shown in Figure 3.9.
Figure 3.9 The transmit beamformer and pulser board in its final form with power cables and decoupling capacitors. Both coaxial and header pin inputs are mounted.

The final component to be built was the control board, which supplies the necessary trigger signal to the pulser/transmit beamformer boards, as well as the five necessary power lines and a solid, noise-resistant ground. Due to the low-frequency nature of these control signals, the critical characteristic of this board is trace width, not length. As the pulser boards themselves were built with a long rectangular shape, a back-plane design was used.

In this layout, a large board with traces for the input signals was manufactured with connectors mounted for each of the pulser boards. Once connected, the boards sit at a ninety degree angle to the control board, or backplane. This design has multiple advantages, including ease of expansion, as adding additional pulser channels would only require an expansion of the existing
backplane layout with minimal changes. In addition, stacking the pulser boards upright allows them to be packed together closer, minimizing the size of the final transmit system.

Due to the simplicity of the required layout for the backplane, a simple breadboard was acquired with traces laid out in a large, finger-like pattern. These fingers allow for the propagation of the global trigger and ground signals to all boards. As the power connections for the boards were already laid out for twisted pair connections, a set of power connectors were laid down at the top of the backplane, cutting down on the trace length and potential noise in these signals.

### 3.7 Single Channel Testing

After manufacturing, the new board proved fully functional, but with a small number of limitations induced by the delay chips. The performance is more than sufficient for the needs of this system, with excellent behavior at the system frequency (Figure 3.10).
The pulse generator chip performed extremely well, but was unable to produce pulses shorter than 5 ns. This is due a rise/fall time of approximately 2.5 ns. Any pulses shorter than 5 ns fail to reach full voltage, and do not trigger the chip output stage correctly. This places an upper limit on the circuit’s output frequency range at just over 50 MHz. While the frequency can be increased slightly over this limit by overlapping the positive and negative pulses, excessive overlap leads to break-down of the output pulse shape. In addition, the pulses would not propagate through a coarse delay of greater than 100 ns or a fine delay of greater than 200 ns.

While these limitations are well beyond the requirements of this project, it is possible that future delay chip developments will allow the same design to be implemented at even higher frequencies and greater delays without major modifications.
3.8 Full Transmit Beamformer and Pulser System

With these limitations in mind, the transmit beamformer and pulser were found to be more than adequate for the purposes of this system, and a full 8-channel implementation was built up. A full set of pulser boards were ordered with the final design with external mounting of all components to minimize noise due to hand soldering. Once these were received and installed on the backplane, the system was finalized by placing all circuitry in a customized box with holes for the input cables and output connectors. The output holes not only served to label the individual channels, but also acted as a stabilizer at the top of the long, thin boards, locking the structure together.

The result of this work is a fully programmable, 8-channel transmit beamformer and pulser which takes as input five power supplies and a low-frequency trigger signal, and outputs 8 independent, high-quality bipolar pulses with programmed relative delays suitable for focusing an annular array ultrasonic transducer at depths of 10-20 mm and at frequencies of 5-50 MHz. With a functional pulser and beamformer unit, the performance of the annular array transducer could be assessed.
Chapter 4

Ultrasound Transducer

The heart of any medical ultrasound system is the ultrasonic transducer which produces the ultrasonic waves used to image the body and receives the resultant echoes. The manufacture of higher frequency transducers (>20 MHz) follows many of the same principles that apply at lower frequencies; however the smaller acoustic wavelengths involved require modified techniques to shape the elements (<100 µm spacing) as well as to make connections with reduced element areas.

4.1 Transducer Characteristics

When designing transducers for medical imaging, particularly for high-frequency imaging, there are several characteristics that must be taken into account when laying down design guidelines, such as the resonant frequency, the aperture, number of elements and the acoustic impedance.

4.1.1 Resonance Frequency

The centre frequency of any transducer is determined by the thickness of the piezoelectric material and the speed of sound in the material. While electrical impulses of many frequencies will induce physical vibrations in the transducer, stimulating the transducer at its resonant frequency will result in a much stronger response. By treating the two faces of a transducer as
independent resonators, it can be shown that the fundamental resonance will occur at a frequency \( f \), such that:

\[
f = \frac{c_p}{2L}
\]

Eqn. 4-1

Where \( c_p \) is the wave velocity in the material and \( L \) is the thickness of the material [25]. Since the wavelength of an acoustic wave in a material is equal to the wave velocity divided by the frequency, \( \frac{c_p}{f} \), this results in a wavelength equal to twice the thickness of the material. Given a wave velocity of 4000 m/s in PZT [57], the half wavelength criterion results in a thickness of less than 100 microns for devices operating at greater than 20 MHz. This is one of many complications of manufacturing high-frequency transducers.

### 4.1.2 Aperture

The aperture of a transducer refers to the active area of the face of the transducer. For single element transducers and annular arrays, this is the full surface of the transducer. For linear arrays only a subset of the elements are active at any given time, resulting in an aperture less than the full face. The aperture of a device is selected based on the desired depth of field for the final image. The depth of field is normally expressed as a function of the aperture with f-numbers. A focal point spaced 3 aperture widths from the transducer face would be said to be at f/3. A common rule of thumb is to avoid imaging closer to the array face than f/2. Attempting to image closer than f/2 can lead to distortion in the image due to phase aberration of the echoes from the edge elements due to the large angles involved. One method of imaging closer to an array is to only active part of the normal aperture, reducing the effective f/2 depth.
4.1.3 Annular Array Elements

While the thickness of the piezoelectric material still determines the centre frequency of the ultrasound bandwidth in arrays, annular arrays add an additional design parameter in the number of rings. Several factors must be taken into account when deciding how many rings to use. First, enough rings must be included to achieve the desired focusing of the output pressure wave. If less than five rings are used, the pressure wave will only be weakly focused. The maximum number of rings is controlled both by ring thickness as well as electronic complexity. Due to the thinning of each ring necessary to maintain the equal-area criterion, if too many rings are used, the outer ring will become too thin, adding to manufacturing difficulty as well as changing the aspect ratio of the activated piezoelectric material. Additionally, each ring of the array requires its own transmit and receive electronics. If large numbers of rings are used, the connections required become cumbersome and the front and back-end electronics increase in complexity, size and power-usage. At a certain point the advantages over a linear or phased array are lost.

For the array presented in this thesis, 12 ring elements were selected as the thickness of the outermost ring was limited by the manufacturing process used for the flexible circuit board and a designed aperture of 6 mm, resulting in an f/2 distance of 12 mm.
4.1.4 Acoustic Impedance

As mentioned in chapter 2, propagation of the acoustic wave into the medium depends on a good acoustic impedance match between the ceramic and the imaging medium. Just as electrical impedance corresponds to a material’s resistance to electrical current, the characteristic acoustic impedance represents its resistance to acoustic pressure and is measured in Rayls [58]. The transmission coefficient between any two materials can be calculated using this impedance, labeled $Z$.

\[
\% \text{ transmitted} = \left[ \frac{4Z_1 Z_2}{(Z_1 + Z_2)^2} \right] \times 100 \quad \text{Eqn. 4-2[31]}
\]

It can be seen that as $Z_1$ approaches $Z_2$, the transmission approaches 100%. This leads to the requirement to match the acoustic impedances between any two abutting materials as best as possible.

4.2 Kerfs

One common feature in all array designs is the physical kerfs between the elements. These breaks in the piezoelectric material decrease the oscillation of neighbouring elements during both send and receive. Dicing saws are normally used to create kerfs, however they are only available in thicknesses down to 20 µm. As the stability of the pillars created is related to their thickness to height ratio, the inability to decrease the size of kerfs with decreasing wavelengths increases the difficulty to the point of infeasibility at frequencies above 20 MHz[31]. For this reason many alternatives to kerfed ceramic have been found to allow the manufacture of higher frequency arrays.
4.2.1 Composites

One common solution is to simply use a piezoelectric material that does not require kerfs. Many high-frequency (>20 MHz) devices have been developed using 3-1 composites [59][10][60]. A 3-1 composite is a material consisting of pillars of PZT ceramic embedded in parallel in an epoxy matrix. These can be manufactured in two ways. Dice and fill, in which slices very similar to kerfs are cut in the ceramic at various angles (using 90 degrees) and filled with epoxy to stabilize the material. Dice and fill suffers from the same limitations as kerfed ceramic, as a dicing saw must still be used to cut the gaps in the bulk material. An alternate method involves creating the ceramic pillars in a mold using a slurry of PZT powder and epoxy. While this avoids the difficulty of cutting the pillars apart, there are some difficulties in solidifying the slurry sufficiently so that it holds its shape after unmolding.

4.2.2 Kerfless Arrays

Previous work in our lab has shown that ultrasound transducer arrays can be constructed with only the defining electrodes, forgoing the kerfs entirely. Without the physical separation, a kerfless array will feature elements that have a slighter larger active area and a higher cross-talk [61][62]. However, as long as these factors are taken into account in the design of the full system, kerfless arrays have been successfully built and shown to have comparable performance to kerfed arrays [14].
Kerfless arrays are particularly useful in the design of annular arrays, as the circular kerfs required cannot be cut using a dicing saw. This made a kerfless design a simple choice for this project.

4.3 Interconnection Strategies

As well as incorporating the kerfless design, the transducer used for this project also implements a simplification of the standard manufacturing method leading to a design which is easier to fabricate and less prone to failures during the process. Significant research has been done recently into novel methods of connecting the transducer itself to the electrical circuitry needed to drive it. This system implements one of these methods while introducing a new addition to the process.

4.3.1 Alternative to Wire-Bonds

In a standard transducer, the electroded piezoelectric material is connected to a miniature coaxial cable or a flexible circuit board using wire bonds. Wire bonds are composed of thin tin wires approximately 25 µm in diameter, which are bonded to the metal of the transducer electrodes and contact pads either on a flexible cable or on the transducer casing. This is a time-consuming process and the wire bonds are prone to breakage if handled aggressively after bonding and prior to encapsulation [63].
To avoid the difficulties created by wire-bonding, the transducer presented in this chapter takes advantage of flip-chip bonding technologies developed for integrated circuit construction. To increase the amount of processing capability in a given area of circuit board, many manufacturers have begun producing what are known as system-on-a-chip. These products incorporate an integrated circuit core along with auxiliary circuitry in a single case. To provide the large number of connections thus required, the integrated circuit is fabricated with an array of contact pads on the underside of the chip. The connecting circuitry is designed with a matching set of contact pads on the surface. A layer of carefully laid out solder balls are laid out on the circuit pads using a mask and the integrated circuit is “flipped” onto the solder balls and the entire assembly is heated to complete the contact[64]. While this particular method requires a large amount of overhead and is extremely sensitive to the orientation and location of the integrated circuit prior to heating, less sensitive solutions have since been developed which are more suitable to ultrasound fabrication.

The solution selected for this particular transducer is an anisotropic conductive adhesive (ACA) manufactured by Delo, AC265 [19]. There has been considerable research into the development of ACAs due to their ability to replace the most costly mask and solder-bump process previously described[65]. Similar to the flip-chip bonding example, the conduction path is provided by metal spheres. In this case, hollow spheres which compress, rather than melt, make contact between the pads. In the case of the Delo ACA, these spheres are contained in a density matched epoxy which cures when heated, holding the spheres in their compressed state and bonding the layers together. This adhesive can be applied, without a mask, to the entire surface of the bonding substrate due to the cross-section of the ACA layer. When properly bonded, the spheres compress and form an
electrical conduction path so long as there is a conductive bond pad both above and below it. If either or both pads are not present, that particular sphere does not conduct. This, along with the insulating epoxy between spheres, ensures that conduction only occurs through the thin axis of the epoxy from conductive pad to conductive pad (Figure 4.1). This eliminates the need for a mask to align the solder balls.

![Diagram Demonstrating the One-Directional Conduction in a Correctly Bonded Layer of Anisotropic Epoxy](image)

**Figure 4.1 Diagram Demonstrating the One-Directional Conduction in a Correctly Bonded Layer of Anisotropic Epoxy**

While the ACA eliminates one aspect of flip-chip bonding that leads to manufacturing difficulties, it leaves the problem of alignment. Due to the small size of the ceramic wafer being used (< 1 cm square), precise alignment can be difficult without specialized equipment [66]. To avoid this requirement, our lab developed a novel method of laying out the electrode contact pads which makes the bonding stage orientation- and position-independent.

### 4.3.2 Ceramic Grid

Alignment is only an issue when using anisotropic epoxy due to the possibility of a single element on the ceramic being connected to two discrete electrodes on the connecting circuit or vise-versa, shorting the elements together. However, if only one of the two surfaces being bonded together features solid elements, the potential for short circuit is removed.
In the thesis transducer, the ceramic wafer was patterned with a grid of chrome/gold electrodes rather than a mirror image of the array pattern on the flexible circuit board. The squares in this grid were patterned so that, at their widest point, they are too small to bridge the spacing between the array elements on the flexible circuit board. This guarantees that each square in the grid will only make contact with a single annular array element. For this array, the grid squares are 30 µm on a side with a 15 µm gap between squares (figure 4.2). This results in a 42 µm diagonal width to correspond with the 50 µm gaps between the elements.

![Image](image.png)

Figure 4.2 The lapped ceramic square patterned with a chrome/gold electrode grid

### 4.3.3 Flexible Circuit Board

A flexible circuit board was laid out with a set of striplines to allow the rest of the ultrasound system to be connected to the transducer, and a 12-ring equal area annular array. The rings were split in a similar manner to the silicon array to allow the flexible circuit board to be manufactured with a single layer of copper. Previous research involving transducers manufactured with
polyvinylidene fluoride (PVDF) and PZT composites have incorporated flexible circuit boards bonded directly to the back in a similar manner to that used here[66][67]. In the case of the PVDF transducer, the difference in material properties between PVDF and PZT allowed the circuit board to be bonded with a simple epoxy layer, while composites required a similar ACA bonding technique to bulk ceramic.

While previous transducers based on this technique have placed the flexible circuit board on the back, analysis of the material properties of the polyamide used as the substrate of flexible circuit board suggested it would be well suited to act as the matching layer on the front of the transducer, provided it was manufactured at the correct frequency. With the active electrodes moved to the front of the transducer in this new design, the single electrode on the back of the transducer can be connected directly to the housing with a conductive bond to ground the unit. The backing layer is the same as in the standard transducer stack up.

4.3.4 KLM Model

To assess the effect of various backing material selection as well as the effect of the flexible circuit board matching layer, a KLM model was built up in MatLab™ (Figure 4.3)
The KLM model contains two separate stages. The first stage contains all the circuit elements between the pulse generator (Vs) and the impedance load of the water seen at the matching layer (Zw). The limiter circuit built into the receive beamformer shunts the receive side of the circuit to ground during transmit. The second stage contains the circuit elements from the matching layer to the input impedance of the ADC (Rl). The expander located at the output of the pulse generator board creates an open circuit during receive, preventing the pulser from loading the second stage.
In the transmit case, the pulser and its load are represented by $V_s$ and $R_s$ respectively. $Z_c$ represents the impedance of the cable used to connect the pulser to the transducer. The piezoelectric effect of the transducer itself is represented with a transformer with a coils ratio of 1 to $\Phi$, where $\Phi$ is directly related to the $k_t$ of the ceramic used. The signal path then branches to represent the dual acoustic waves, one travelling in the forward direction to the matching layer, and one travelling in the reverse direction toward the backing layer. All acoustic layers, including the piezo-ceramic are treated as transmission lines with lengths equal to their thickness in wavelengths and their impedance defined by their characteristic acoustic impedance. Finally the backing layer and the interface with the water are represented as impedance loads. Simulation of this half of the model allows the expected pulse at the face of the transducer to be calculated.

The receive portion of the model mostly replicates the forward case in reverse, but with the pulser and its output resistance replaced with the cable to the receive electronics and the input impedance of the first stage of the front-end receive electronics.

When simulating the KLM models, the pulse shape, frequency and amplitude of the input pulse can be modified, as can be the thickness of the ceramic and the impedance of the backing layer. All of these allow various transducer and pulser configurations to be tested before the transducer itself is manufactured, saving materials and time spent. It should be noted that by adjusting the transformer ratio in the model, the effect of loss of poling in the material (due to heating during manufacturing and mechanical stresses) can also be modeled and compared to experimental results.
4.4 Manufacturing

4.4.1 Flexible Circuit Board/Matching Layer

To manufacture this transducer the first step was to design and lay out the annular array rings and connecting strip line in a CAD program similar to the one used to design the pulser circuit. The desired ring widths were calculated using a MatLab script given a desired aperture of 6 mm and 12 rings. The program also included the 50 µm gaps between the rings required to avoid shorting during bonding. The resulting rings were laid out as C-shapes with 50 µm traces (Figure 4.4).

Figure 4.4 The annular array electrode pattern laid out on the flexible circuit board and the connecting traces

It was important to ensure that the traces were separated by the same minimum spacing as the array rings, as the portion of the traces within the rings could be shorted together the same way.
the rings could be if placed too closely together. These traces were extended 30 mm to a pad
layout designed to be soldered directly to a parallel cable connector designed for high-frequency
applications (Figure 4.5).

Figure 4.5 Circuit board connector at the end of the flexible circuit board used as the
matching layer and interconnection method of the annular array transducer

A plane of solid copper was also laid out to stop just short of both the annular rings as well as the
connector fan out. This copper plane performs two functions: first, after connection to the
transducer housing and the ground contact of the parallel connector, it supplies the necessary
ground return for the pulsing system. Second, once grounded, it supplies the ground plane for a
stripline layout. This is a well-known circuit trace design method to minimize parasitics in long
signal lines [68].

Once laid out, the design was sent to a flexible circuit board manufacturer with specifications for
a final board thickness of 25 µm. When the final boards were obtained they were discovered to
consist of polyamide with a characteristic speed of sound of 2435 m/s and 30-32 µm thickness.
To calculate the centre frequency, a modified form of equation 4-1 is used, allowing for that fact
that a matching layer should be optimized for propagation, rather than oscillation. For this application, the thickness should be a quarter of the acoustic wavelength, rather than half, like the resonator. Therefore, the modified equation is:

\[ f = \frac{c_p}{4l} \]  \hspace{1cm} \text{Eqn. 4-3}

By plugging in the speed of sound and the thickness into equation 4-3, the optimum frequency for acoustic propagation in this matching layer can be calculated. This calculation results in a centre frequency of 20.3 MHz.

4.4.2 Fan-out Board

With the flexible circuit board being manufactured, a second circuit was designed to connect the finished transducer to the pulser circuit as well as to the front-end circuitry of the receive circuit. This board is a simple fan-out board designed to interface between the parallel connector on the flexible circuit board and the SMA coaxial cables on the output of the pulser as well as the input to the receiver board. Due to its simplicity, this board was designed using the same in house design and manufacturing steps as used to create the pulser test board in the previous chapter. The resulting board features individual SMA connectors for each annular ring connected to SMA T-connectors. The board is also carefully labeled with the corresponding ring for each connector due to the non-intuitive layout caused by the combination of the flexible circuit board fan-out and the connector board fan-out (Figure 4.6).
4.4.3 PZT Wafer

To produce the grid electrode layout on the bulk ceramic, photolithography is used. For this purpose, a layout of the desired grid was produced in a CAD software package and submitted to the university of Alberta fabrication laboratory for mask production. A photo-lithography mask is a glass plate coated on one side with a metal layer patterned with the desired layout. For a positive mask, the metal protective layer is patterned on anywhere that metal is desired on the final sample. A negative mask features the reverse. For this application, a positive mask was obtained.

The bulk ceramic used for this transducer was TRS 5H PZT. The ‘H’ designation of this ceramic indicates that it is a ceramic designed for high-frequency applications, and that it contains a smaller grain size than the 5D ceramic available from the same manufacturer. Before photo-lithography, the ceramic was lapped down to a thickness of 105 μm which, using equation 4-1,
results in a 19 MHz centre frequency. As the ceramic is delivered with electrodes on both sides, lapping is performed prior to patterning the electrode grid so that one of the original electrodes can be used as the ground electrode.

To lap the ceramic, each sample was mounted to a glass carrier plate using a specialty mounting wax. The wax, sample and carrier plate are heated to 80 degrees, then the ceramic is applied on top of the wax and pressure is applied using a plastic wrap-covered squash ball. A squash ball is used to ensure even pressure across the entire surface of the ceramic wafer. This is critical, as any error in the flatness of the bond will result in a corresponding error in thickness across the wafer after lapping.

To thin the ceramic to the correct thickness a combination of a lapping stone and fine-grit slurry are used. For this transducer, a Logitech lapping system was used with a flat, round bottom plate and a 9 µm calcined aluminum oxide powder mixed 1:10 by volume with water. This set-up results in a lapping rate of ~5-10 microns per minute, allowing the final thickness to be well controlled. When the ceramic used in the transducer presented in this chapter was finalized lapping, a thickness of 113 µm had been achieved for a theoretical centre frequency of 18.6 MHz.

Once thinned, the wafer was sub-divided into 12.5 mm x 12.5 mm squares to match the size of the grids laid out on the pre-prepared photo-lithography mask. They were then removed from the glass lapping plate by heating the bonding wax. All photo-lithographic steps were performed on the bare wafers in a clean room to protect the chemical layers during the processing steps.
The wafers, now thinned and electroded on a single side, were cleaned using a sequence of acetone and methanol. Acetone is a more aggressive cleaner and removes any surface grit left by the lapping process as well as any debris picked up during transportation to the clean room. Methanol is used to remove the slight residue left by the acetone, which naturally evaporates at room temperature.

Once cleaned, a thin layer of photo-resist is applied to the bare side of the wafer using a spin coater. The appropriate spin rate for a given application can be determined using the specifications of the photo-resist. For this project, MaN-405 was used as it has a low enough viscosity to attain a one micron thickness with a rotation speed of 3000 RPM [69]. Optimum feature resolution is attained when the thickness of the photo-resist is as close as possible to the final metal thickness, thus a one micron thick layer is desired [70]. As the final electrode will be slightly thicker than half a micron, a one micron thick photo-resist layer is ideal.

Once the photo-resist layer had been heat-set, it was placed under the mask and exposed to ultra-violet light. The exposure time was again determined by the photo-resist specifications, however the exact exposure/second of the available light was not known, requiring some experimentation to determine a final exposure time of 15 seconds. After exposure, the excess photo-resist was removed using a developer, MF-319 [69]. For negative photo-resists, the ultra-violet light exposure causes the photo-resist exposed by the mask to harden, preventing it from being dissolved by the developer [64]. As the mask had been prepared so as to cover the areas where metal was to be deposited, these areas were washed away by the developing process.
Five hundred angstroms of chrome and five thousand angstroms of gold were laid down on the prepared wafers using an evaporator. The chrome layer is laid down first to increase the adhesion of the gold, which has low natural adhesion to bare ceramic. The thickness of the gold layer was determined through a combination of the minimum final cross-section to keep the final resistance of each ring as low as possible as well as controlling the total cost of the gold applied. Once evaporation was complete, a simple bath of acetone with ultrasonic vibration was sufficient to remove the rest of the photo-resist, leaving a clean, gridded ceramic (figure 4.2).

4.4.4 Assembly

With the ceramic wafers prepared and the flexible circuits manufactured, the final transducer could be assembled. The ACA used in this transducer must be kept at -10 degrees Celsius when not in use to avoid curing of the epoxy matrix [71]. To ensure even pressure during curing, a special clamp was constructed. A thick aluminum plate was used as a base to prevent movement of the system. This base supports two metal pillars attached to a top plate containing the clamping apparatus. Each side of the top plate contains a metal plunger assembly with weak springs designed to hold the ends of the plungers just above the base plate in their neutral position. This gap allows the inclusion of a glass lapping plate to be used as a transportation base for the completed transducer. The end of the plunger was machined into a square cross-section 7 mm on a side. This is just smaller than the final size of the ceramic wafer, allowing the alignment of the wafer on all sides to be verified prior to curing. The clamping jig also incorporates two screws above the plungers to increase the pressure applied gradually.
To glue the ceramic wafer to the flexible circuit board, the board was taped down flat to a glass carrier plate. The flatness of the board was checked using a flatness gauge and a marble flat calibrated to +/- 1 μm. An adhesive backed square of polypropylene approximately 1 mm in thickness was applied to the end of the plunger. This helps spread the pressure evenly across the entire ceramic to prevent cracking and maximize flatness in the final bond layer. To prevent sticking, a layer of aluminum foil was wrapped around the entire plunger assembly, making sure that the foil was flat against the polypropylene with no creases, again to avoid cracking.

Once the clamp was prepared, the ACA was allowed to come to room temperature so the epoxy would flow evenly from the syringe. The glass plate and circuit board were placed in the clamp with the plunger temporarily off centre. A small amount of ACA was then applied to the centre of the annular array rings. The ceramic wafer was carefully applied on top of the ACA with the gridded electrodes facing the ACA. Gentle pressure was used to compress the ceramic and ACA to a thickness of approximately a millimeter. The plunger was then compressed and the ceramic-ACA-circuit board sandwich centred under it. The plunger was slowly released to avoid cracking the thin PZT. Finally, a half-turn of the pressure screw was applied to achieve a pressure found in testing to result in optimum bonds. The final stack-up of materials in the clamp can be seen in figure 4.8.
The entire clamp assembly was then inserted in a laboratory oven and brought to a temperature of 180 degrees Celsius. As the ACA cures in 5 seconds at this temperature, the clamp was removed as soon as thermal equilibrium was reached. It is desirable to avoid protracted heating of the ceramic, as the material used has a Curie point of 210 degrees Celsius [72]. The Curie point refers to the temperature above which the piezoelectric domains in the ceramic become malleable. Ceramic manufacturers take advantage of this malleability by applying an electric field across the ceramic while holding the ceramic above the Curie point. This amplifies resonance along this axis and is known as poling. Raising a poled ceramic back above the Curie temperature can result in loss of poling, weakening the resonance [36]. This is reflected in a loss of $k_r$, which can potentially be modeled using the KLM models mentioned earlier.
4.4.5 Transducer Impedance

Once the clamp had cooled and been released, the centre frequency of the transducer was verified experimentally. This was achieved by connecting the exposed ground electrode of the ceramic to the ground plane on the flexible circuit board using a braided copper strap. To connect to the break-out board designed previously, a connector was soldered to the contacts at the other end of the flexible circuit board and another contact made to the ground plane. The impedance of the central array element was then measured using an impedance analyzer (HP 4396B, Hewlett-Packard Company, Palo Alto, CA). The impedance plot observed displays the overall capacitive impedance of the ceramic with two resonance peaks. These two resonance/anti-resonance pairs correspond to the resonance frequency of the ceramic itself as well as the resonance of the flexible circuit board (Figure 4.9).
Figure 4.8 The impedance of the transducer as seen at the end of the flex cable before a backing layer was applied

The lower frequency peak which can be seen at 17.2 MHz corresponds to the resonance frequency of the flexible circuit board, acting as a matching layer. This frequency correlates to the thickness of the received boards, which were measured at 30-32 μm in thickness, rather than the 25 μm ordered. In contrast, the higher frequency peak seen represents the centre frequency of the ceramic. At 23 MHz, this is higher than expected from the measured thickness of the lapped ceramic. This discrepancy may be due to an offset error in the thickness measurement caused by compression of the ceramic during measurement of the un-lapped and lapped ceramic.

This measurement also serves as a verification of the ACA bond, as the resonance peaks will not be visible if the bond is too thick to allow conduction through the metal spheres in the epoxy. A thick bond acts electrically as an insulator, resulting in a pure capacitive impedance plot. The
bond thickness can also be measured using a thickness gauge so long as a high quality measurement of the ceramic thickness is obtained beforehand. If the bond is $\leq 2 \, \mu \text{m}$, conduction should be possible in the thickness direction.

### 4.4.6 Backing Layer

Once conduction had been verified, the backing layer could be applied. To ensure a high-quality backing layer, a special housing was machined out of 5 mm thick stainless steel with a cylindrical hole 8 mm in diameter. By machining the hole larger than the diameter of the annular array, overlap of the metal housing with the active elements of the array is prevented so long as the housing is aligned correctly with the ceramic. For mounting to the testing rig, a second hole was machined with a metric thread. A stainless steel rod with a matching thread was also machined to secure the housing at a ninety degree angle to its face.

Prior to mounting the housing, the ground strap attached to the ceramic ground electrode was removed. If left in place, this strap could prevent the housing from being mounted parallel to the face of the transducer, making targeting the completed transducer more complex. The housing was aligned with the ceramic centred in the non-threaded hole and tacked in place. This was done using a conductive epoxy consisting of epoxy loaded with a high density of silver particles. Unlike the ACA, the volume fraction of silver particles in this epoxy is high enough to provide conductive channels in all directions due to particle on particle contact. Using this epoxy to attach the housing allows the entire housing to act as the ground contact for the transducer.
Additional conductive epoxy was used to reattach the ground strap to the outside of the housing, completing the circuit.

To achieve sufficient attenuation for the backing layer, a West System 2-part epoxy (102 and 105) was mixed with 3 μm tungsten powder to a 1:6 ratio by weight. Tungsten is commonly used in backing ultrasonic transducers due to its high density. A 1:6 ratio corresponds to an acoustic impedance of 10 MRayls [31]. As PZT-5H has an acoustic impedance of 36 MRayls [73], it is desirable to have as high an acoustic impedance as possible in the backing layer.

The marine epoxy has a “pot life” of 30 minutes to an hour, allowing the tungsten to be well blended with the epoxy matrix without fear of it curing in the mixing vessel. To prevent air bubbles forming between the epoxy and the ceramic during filling, the epoxy was drawn into a 30 ml syringe with a large bore tip. The choice to use a syringe places a limit on the amount of tungsten that can be mixed with the epoxy. While a higher ratio would result in a higher attenuation rate while still curing correctly, the resulting mix is too viscous to draw into a syringe, making it difficult to apply. By beginning the back-filling process with the syringe tip very close to the ceramic and slowly pulling back in a single motion, a smooth layer of epoxy was achieved.

The backing layer was cured for twenty-four hours, the impedance plots for the array rings were measured. Figure 4.10 shows that the backing layer has reduced the amplitude of the resonance peaks, but has otherwise not altered the plots significantly. The transducer was now ready for connection to the pulser system for testing (Figure 4.11).
Figure 4.9 Electrical impedance of the annular array rings after the backing layer was applied

Figure 4.10 The finished annular array transducer with backing layer and flexible circuit
4.5 Testing

Figure 4.12 shows the pulse echo response of the central element of the array and the corresponding magnitude spectrum. The pulse echo response was measured in a water tank by recording the reflection from a quartz flat placed 10 mm in front of the transducer. The transducer was excited using a mono-cycle pulse (30 Vpp) and the detected signal was recorded using the 50 Ohm input to a digital oscilloscope. The pulse spectrum was calculated in software.

![Pulse Echo Response and Spectrum](image)

**Figure 4.11** The pulse-echo response and bandwidth of the centre element of the annular array imaging a quartz flat at f/2 (10 mm)

The pulse has a center frequency of 22MHz and a -6dB fractional bandwidth of approximately 50%. For comparison, the KLM model prediction of the pulse echo response and pulse spectrum are shown in figure 4.13. There is reasonably good agreement between the KLM model prediction and the experimental result.
The insertion loss for the central element of the array was measured by placing a quartz flat in a water bath close to the face of the array (5 mm) and measuring the pulse echo response and also the excitation pulse using the 50 Ohm input of a digital oscilloscope. The insertion loss was calculated from the ratio of the magnitude spectrums of the pulse echo response and excitation pulse. Corrections were made for the reflection coefficient of the quartz flat as well as the attenuation of the water but diffraction losses were ignored. This resulting insertion loss is shown in figure 4.14. A minimum insertion loss of -22 dB was obtained at 24 MHz.

Figure 4.12 KLM modeled pulse-echo response and bandwidth of the annular array transducer
The cross-talk between adjacent elements in the array was measured in a water bath by exciting an array element using a 30 Vpp mono-cycle pulse and measuring the signal on the adjacent element. The cross talk ranged from -26 dB for the central elements to -45 dB for the outer elements. This is slightly better than the -20 dB level reported by Brown et al. for a kerfless PZT annular array[73] but much worse than the -40 to -50 dB levels reported by Chabok et al. for a 1-3 composite annular array[73].

Figure 4.13 Insertion loss for the annular array transducer as a function of frequency
The two-way radiation pattern of the array was measured in a water bath by scanning the array across a point target placed in front of the transducer. The point target was fabricated by drawing out a heated glass rod to produce a fiber with a thickness of ~125 µm, and then melting the tip of the fiber to produce a spherical knob. This was mounted vertically in a water tank with the stem of the glass target along the axis of the beam and the tip 10 mm from the transducer face. The inner 8 elements of the array were focused on this target using a transmit beamformer with delays produced using DS1023 digital delay chips (Maxim, Sunnyvale, CA). We were limited to 8 channels due to the pulser/transmit beamformer only hosting 8 channels. The transducer outputs were connected to the oscilloscope using a simple protection circuit consisting of a λ/4 coaxial cable terminated by a limiter. The output of the limiter was connected a 115 MHz low-pass filter (Mini-Circuits SBLP-117, Brooklyn, NY) and a 35 dB amplifier (Miteq, Hauppauge, NY) before terminating at the 50-Ω input to the oscilloscope. The response from each element of the array was recorded individually and exported to MATLAB™ for beamforming. The array was then scanned laterally across the point target and the process repeated at each step in the scan. The resulting two-way radiation pattern is shown in figure 4.15. The radiation pattern has a -6 dB main lobe width of 160 microns which corresponds to 2.1 wavelengths at 20 MHz Secondary lobes in the radiation pattern decrease smoothly to an amplitude of -50 to -60 dB with respect to the main peak.
The pulses received by each element when the glass target was positioned at the focal point of the array are shown in figure 4.16. The signals are shown after the receive beamforming delays were applied but before the signals were added. The amplitude of the signals from each element decreases smoothly from the central element to the outer element. This decrease in amplitude has been previously been reported for kerfless arrays [74]. The pulse echo response from the glass target is significantly longer than from the quartz flat (figure 4.12) due to reflections within the glass knob.

Figure 4.14 Radiation diagram for a point-target located at f/2
Figure 4.15 Beamformed pulses from a point target located at f/2 along with the resultant pulse (inset)

With the transducer and pulser/transmit beamformer fully functional, a discrete receive beamformer was required to complete the system.
Chapter 5

Receive Beamformer

Just as an ultrasound transducer requires a system of electronics to create the initial pressure wave used for imaging, a similar system is required to focus the signals detected by the transducer. This system is known as the receive beamformer system. In addition to the basic beamformer algorithm itself, which implements the delay, sum and any higher level processing of the received signals, the receive electronics also contain the analog front-end, an analog to digital converter and a computer interface.

5.1 Beamforming Algorithms

The basic functionality implemented in receive beamforming is known as delay-and-sum. Delay-and-sum implements either analog or digital delays in a similar manner to the transmit beamformer to the signals received by the transducer [75]. The delayed signals are summed together and the envelope obtained using either quadrature methods or heterodyning. When used with an annular array transducer, this system mimics the output of a single element transducer with the advantage of an adjustable focal depth. In early systems no additional processing was done on the signals due to limitations in hardware and computation speeds. As each element of a transducer requires its own calculation, bulky early electronics necessitated a simple beamformer to keep the final system manageable. These beamformers are still in use in applications where minimization of electronics is desired, such as in hand-held scanners or low-cost systems [76][21], or for synthetic aperture systems where low-complexity is desired for speed [77][78].
Early analog beamformers were limited to a single receive delay value by hardware [79]. The shift to digital delays in modern beamformers has allowed the use of different delays for each output pixel. This is known as dynamic receive focusing and is now considered standard practice. In this implementation, the depth of each final pixel in the image with respect to the individual transducer elements is calculated. The resulting delay values are used to determine the closest data point from each output signal. These data points are then summed to obtain the value of the given pixel, with each pixel calculated in turn. As the distance calculations can be performed ahead of time, dynamic receive focusing has minimal effect on the complexity of the final design versus fixed focusing, resulting in an increase in depth of field on the image for minimal additional computation.

It should be noted that in the above calculation, the summation for each pixel is only as accurate as the time delay between the optimum delayed data point and the closest sampled data point. This error is directly related to the ratio between the centre frequency of the ultrasonic pulse and the sampling frequency of the system. It is common in low frequency systems (under 1 MHz) to minimize this error by selecting a sampling rate significantly higher than the centre frequency ($f_{\text{sampling}} > 50$ MHz) [3]. To achieve an equally high centre to sampling ratio with a 20 MHz system would require a sampling rate of over 1 GHz. While analog to digital converters have been available at these frequencies for several years, they are significantly more expensive than those used in low-frequency systems and are single channel. That is to say, each chip can only process a single channel of electronic data. By contrast, ADCs capable of processing up to 8 channels simultaneously are available for low frequencies, at lower cost than the single channel,
>1 GHz chips. An 8-channel, 12-bit 80 MS/s ADC costs approximately $60 [80] while a single channel 12-bit 1 GS/s ADC costs over $750 [81]. The increase in cost per channel resulting from the use of high speed chips would lead to a prohibitive equipment cost in any system with a large number of channels.

To reduce the ADC sampling rate requirements of the system, several signal processing techniques have been used in previous systems. However, no matter what method is used, certain minimum criteria must be met. During the development of digital processing electronics, fundamental limits have been identified for the digitization process for the resulting digital data to have minimal error. The sampling frequency of the ADC must be higher than the Nyquist frequency. The Nyquist frequency is twice the highest significant frequency present in the signal to be sampled. The Nyquist condition is dictated due to the wrapping of spectral information above half the sampling frequency into lower frequency bands. This is known as aliasing and can result in signal distortion. For a 20 MHz signal with a bandwidth of 20 MHz (100%), this results in a Nyquist frequency of 60 MHz. While still higher than the sampling frequency used in low frequency systems, this is a much less stringent criterion than that of minimizing delay error.

To convert data taken at a near Nyquist sampling frequency to more finely sampled data, methods of various computational complexities are available. The simplest form of data conversion is linear interpolation. Linear interpolation’s advantage is its relative simplicity, allowing it to be implemented on systems with a limited number of arithmetic operations and those with less available storage. Linear interpolation takes each pair of adjacent data points and calculates the
slope between the two. Once this has been calculated, it follows that the first order approximation to any data point between the two can be calculated with the following formula:

\[ v(t) = \frac{v(t_2) - v(t_1)}{t_2 - t_1} \times (t - t_1) \]  

Eqn. 5-1

\( V(t) \) is the voltage at a given time step, \( t \) is desired time sample, \( t_1 \) is the time sample before the desired time sample and \( t_2 \) is the sample after the desired time sample. To save on computation time, only data points which will appear in the final image are computed. If additional processing will be done post-beamforming (Doppler flow, for example), more data points may need to be calculated. In hardware which cannot handle floating point arithmetic or with limited storage capabilities, up-sampling is commonly performed at a set rate, with the closest up-sampled data point used for beamforming. With slightly higher processing ability, a set of pre-calculated values based on the desired focal depths can be used to calculate the precise data point required for each pixel. While more accurate than the previous method, both methods suffer from the error induced by applying a linear spline to the pairs of data points. While higher order splines can be used to decrease this error, the increase in image resolution is minimal compared to the increase in computational complexity \[82\]. Alternate beamformer techniques include heterodyning \[83\], which is prone to noise and delta-sigma processing \[84\], which requires a higher sampling rate, but uses lower amplitude sampling.

5.1.1 Current Low-Complexity Beamformer Algorithms

Due to the ever increasing speed and dropping costs of consumer electronics, many groups have moved to beamformer implementations based on commercially produced chipsets. In particular, ADCs from Analog Devices and Texas Instruments and Field Programmable Gate Arrays
(FPGAs) from Xilinx have seen considerable use in recent work. While FPGA based systems have the advantage of parallel processing and can be placed adjacent to the ADCs, keeping digital noise to a minimum, they do face some limitations in terms of processing and on board storage. For this reason FPGA based systems use the same simple delay and sum architecture described earlier [22][78][85][86].

While prices for ADCs have dropped considerably in the last decade, there are still limitations on the upper range of sampling rates. For this reason most of the beamformer implementations are based on sampling ratios under 20. In other words, the sampling rate is no more than 20 times the highest frequency component of the sampled data. To compensate for this low sampling rate ratio, several algorithms are used, depending on the system architecture. For FPGA systems, the algorithms used are limited by the difficulty in performing non-integer math. This tends to preclude more complex algorithms involving FFTs or wavelets. The simplest systems use a basic linear interpolation scheme to obtain the desired data points[85][87][88]. While functional, the error rate increases significantly as the sampling ratio drops.

In the beamformer presented in this thesis, a modified curve fitting algorithm was applied. By making a few simple approximations and sampling at a particular frequency with respect to the centre frequency of the main pulse, a higher quality fit to the original curve can be achieved without increasing the computational complexity over linear interpolation.
5.1.2 Modified Curve-Fitting Approach

The time-domain signal of an ultrasound echo can be represented as an amplitude modulated sine wave of a given centre frequency.

\[ v(t) = a(t) * \sin(2\pi ft + \theta) \] \quad \text{Eqn. 5-2}

This form can be rearranged to replace the angle term theta with a second modulated cosine.

\[ v(t) = a(t) \sin(2\pi ft) + b(t) \cos(2\pi ft) \] \quad \text{Eqn. 5-3}

In equation 5-3, a(t) and b(t) are time-dependent terms which fully describe the pulse-echo response. To allow a curve to be fit with only two data points to solve with, the assumption is made that a(t) and b(t) are constant between any two samples. Any error resulting from this interpolation method is related to the accuracy of this assumption.

The computational simplicity of this fit is dependent on which data points are used for fitting. Using data points sampled at four times the centre frequency allows a significant simplification to the fit. In this system, the sampling rate of 80 MHz corresponds to 4 times the 20 MHz centre frequency of the transmitted pulses. The assumption is made that minimal frequency-shift occurs during acoustic transmit, resulting in a series of data samples at intervals of T/4. With this assumption, the two data points \( v(t_1) \) and \( v(t_2) \) in equation 5-3 can be simplified as \( v(0) \) and \( v(T/4) \). For \( t=0 \), equation 5-3 can be reduced to \( v(0) = b \cdot \cos(0) \), or \( b=v(0) \). A similar calculation for \( v(T/4) \) reduces the equation to \( a=v(T/4) \). With this simple fit, the calculation for each final data point is reduced to a single evaluation of equation 5-3. A further simplification can be applied due to fore-knowledge of the desired time values. This allows the sine and cosine
terms to be pre-calculated and stored in look-up-tables. The improvement in fit over a linear spline can be seen in Figure 5.1.

![Graph showing comparison of different methods]

**Figure 5.1 Comparison of Modified Curve-Fitting Algorithm to the Ideal Case and to a Linear Spline**

To compare the functionality of the curve fitting algorithm for beamforming purposes as compared to a linear interpolation, a linear array was simulated in MatLab™ and the output sampled at T/4 of the 20 MHz centre frequency. This output was then interpolated using a linear interpolation algorithm or the curve-fitting interpolation algorithm and the radiation pattern was plotted for both (Figure 5.2). A clear 10 dB improvement can be seen between the linear interpolation case and the curve-fitting case.
Figure 5.2 Radiation diagrams for a linear array focused at f/2, sampled at four times its centre frequency and interpolated using a linear fit and the curve fitting algorithm

5.2 Hardware Limitations

If the analog signals received at the ultrasound transducer could be perfectly digitized without loss of information, it would be possible to implement a gold standard beamformer with no additional errors [89]. However, in practice all digitization induces some error, at least in the amplitude domain, and quantifying these errors is important to the design of a receive beamformer which will produce images of acceptable quality.

5.2.1 Digitization Errors

The error induced by the digitization process is directly related to the amplitude digitization and sampling rate of the ADCs selected for the receive beamformer. To investigate the effects of
digitization on the final image, a simple linear array was simulated in MatLab. A linear array was used as the beam pattern is simpler to simulate than that of an annular array, and the off-axis dimension can be neglected for the purposes of this calculation. The simulation was done with a pulse centre frequency of 50 MHz. A single point target was then imaged at various positions along the aperture, y, at a fixed distance, x, from the array (Figure 5.3). The maximum amplitude of the beamformed A-scan lines was then plotted as a function of the target’s position on the y-axis.

Figure 5.3 Position of the Simulated Point Target With Respect to the Focal Point at y=0

The resulting plot is known as a beam diagram and can be used to assess the lateral resolution of a system as well as the dynamic range. This is because the plot shows what effect a strong echo source will have on the rest of the image. Any echoes which are weaker than the echo received from the strong source at a distance will not be visible in the final image.

To assess the effect of digitization on a system, the simulated return echoes were digitized at various time and amplitude settings and the resulting beam patterns plotted.
Figure 5.4 Effects of Sampling Rate on a 50 MHz Linear Array Radiation Diagram

Figure 5.5 Effects of Amplitude Quantization on a 50 MHz Linear Array Radiation Diagram
In these plots, the level of the pedestal lobes, the long tails to either side of the main peak, is a measure of the maximum dynamic range of a system featuring that level of amplitude sampling. Common design rules hold that 60 dB of dynamic range are necessary for diagnostic quality medical images. Using this as a benchmark, an amplitude sampling rate of at least 10 bits is necessary, with a larger range desirable.

5.3 Analog Front-End

The analog electronics necessary to clean the raw data signal from ultrasonic transducers are known as the front-end electronics. While the precise sequence of electronic stages and components vary from system to system, all systems require a protection circuit and amplification stages to deal with the 20+ dB difference in amplitude between the initial imaging pulse and the received pulses. In addition, filters are commonly incorporated to remove electronic noise outside the spectrum of the signal of interest.

5.3.1 Circuit Design

As mentioned in the transmit beamformer and pulser chapter, most ultrasound systems incorporate control circuits designed to prevent the receive electronics from being active during transmit and the transmit circuits being active on receive. While the circuit on the transmit side is called an expander, the receive equivalent is a limiter. There are two basic varieties of limiter used in receive circuitry, passive and active. A passive limiter consists of a pair of diodes, connected in opposing directions, like an expander, but tying the signal line to ground. Large
voltage signals, such as the transmit pulse, cause the diodes to conduct, grounding the receive circuit. To avoid shorting the transducer, a capacitor or quarter wavelength transmission line is placed between the transducer and the limiter [90].

While passive limiters work well at lower-frequencies, the rapid rise/fall time of higher-frequency systems can occasionally swamp the diodes, preventing them from switching on in time and overloading the receive circuit. In these cases, an active limiter circuit can be used. An active limiter circuit takes the form shown in figure 5.6.

![Active limiter circuit diagram](image)

**Figure 5.6 Active limiter circuit diagram**

The precise voltage and resistor values used can vary between designs, but the critical design characteristic is the separation of the input node from the output node. While the diodes on the input node in this design can still suffer from rise/fall issues, the diodes on the output node will
see the same delay in response, keeping the output node voltage from fluctuating too quickly. In
the simpler design, the input and output nodes are directly connected, allowing rapid fluctuations
to past directly through.

The first amplifier component in most front-end stages is a low-noise amplifier (LNA).
Amplification is necessary due to low amplitudes of received echoes in ultrasound systems. Even
with a high-quality matching layer, and a bright reflector in the imaging field, most systems
feature an approximately 15 dB loss in signal from transmit to receive, even compensating for the
loss of signal due to the imaging media. The loss in amplitude results in maximum return echoes
of $\sim 2\text{-}3\text{V}$. When performing in vivo imaging, however, most tissues reflect less of the signal
than the metal and quartz targets used for testing, leading to echoes of 10-100 mV [91]. Without
amplification, a 100 mV peak would require an electronic noise floor of <100 $\mu\text{V}$ to maintain the
60 dB dynamic range. Due to the front-end noise of the ADC circuitry, maintaining a noise floor
at this level would not be possible. To raise this noise floor requirement, an LNA gain setting of
anywhere between 15 and 30 dB can be used, depending on the rest of the system specifications.

As mentioned previously, acoustic waves are subject to depth dependent attenuation in human
tissue. This attenuation is due to absorption and scattering the ultrasonic wave off of sub-
wave-length scatterers in the material. As this attenuation is approximately constant through the
material, the relative losses can be considered to be linear with respect to depth [91]. Thus, if the
diffusion loss of a given material is known, it can be compensated for with an adjustable
amplifier. For this reason, the second stage in most receive systems is a voltage-controlled gain
amplifier (VGA) which is used as a Time-Gain Compensator (TGC) [92]. By applying a voltage
ramp to the control input of this stage with its lowest value triggered to the initial imaging pulse, the output signal from this stage will be corrected for imaging depth, evening out the image.

While called an amplifier, many VGAs result in minimal amplification, so if the system is being designed for low echo media, an optional amplification stage may be included beyond the VGA stage.

5.3.2 Electrical Noise

All electrical circuits produce noise in addition to the intended signals. Discrete components produce thermal noise with a known level proportional to their resistance. Integrated circuits will have noise induced by internal circuit parasitics as well as packaging effects. Most ICs manufacturers include specifications on the noise on their datasheets and will include circuit models to allow noise modeling. Thermal noise is called white noise, as it has a constant rms value across the entire frequency spectrum. White noise is a given for any specific electrical device, and can only be lessened by the selection of low-noise devices during the design process.

The presence of noise sources is the reason why it is critical for the first amplifier in the signal chain to be a low-noise amplifier. As not all noise sources in an amplifier are affected by the gain setting of the amplifier, they will be specified separately, with gain dependent noise specified as input-referred, and gain independent noise specified as output-referred [93]. In general, it is good circuit design to place the noisiest components as far along the signal chain as possible to avoid noise amplification.
5.3.3 Filters

The only components which should be placed after large noise generators are signal filters. Filters can be implemented in either the analog or the digital domain and act to attenuate signal levels in a given frequency range. A low-pass filter allows all frequencies below a given “knee” frequency to pass un-attenuated, while applying increasing attenuation as a function of frequency above the knee. The knee itself is defined as the frequency at which the filter attenuates the voltage signal by a factor of 2, or -6 dB. High-pass filters act as the reverse of low-pass filters, attenuating frequency content below the knee, rather than above, with the same definition of the knee frequency.

The appropriate selection of filter knee frequencies can improve the total dynamic range of the system by reducing the total noise in the system without impacting the signal amplitude. The dynamic range is a function of the signal to noise ratio (SNR).

5.4 System Specifications

The full receive beamformer system is pictured in figure 5.7. The system incorporates an analog to digital converter board, a data capture board and a PC controlling the sample transmission and performing software beamforming. The various components seen in the diagram are explored in detail in the next few sections.
5.4.1 AD9272

For the system presented in this thesis, a near-Nyquist sampling rate was used to keep costs low. Based on these requirements and the limits described earlier, an 8-channel ADC chip from Analog Devices (AD9272) was selected with a 12 bit amplitude sampling rate and a time-sampling rate of 80 MHz [92]. This satisfies both conditions for diagnostic quality images while keeping the cost per channel down. In addition to meeting the sampling requirements, the AD9272 was specifically designed with ultrasonic imaging systems in mind, and incorporates additional electronics for each channel to help shape the incoming pulses and minimize noise in the digitized image.
The AD9272 also includes an LNA capable of up to 21 dB of amplification, a VGA with -42 dB to 0 dB controllable gain and a 21-30 dB secondary gain stage in addition to the ADCs themselves [94]. An evaluation board for this chip is available from the manufacturer, which includes a capacitor and limiter on each input channel and an on board oscillator to set the sampling rate. As these components met all the requirements for our system layout, the decision was made to use the evaluation board with a few modifications. One limitation of the evaluation board is that while the AD9272 supplied with the board is capable of sampling at up to the 80 MHz required for the system, the oscillator on board is a 40 MHz model. Fortunately, an 80 MHz oscillator with the same footprint as the 40 MHz one was located and the lower frequency chip was replaced. It should be noted that this was only possible as the boards are supplied with oscillators below the design limits of the chip.

To prevent noise outside the frequency range of the signals corrupting the signal quality, the chip also contains both low and high-pass digital filters. The high-pass filter can be set for frequencies between 775 kHz and 11.5 MHz and the low-pass filters can be set from 9 MHz to 36 MHz. The filters, along with the LNA and secondary gain stage, are controlled via the same parallel connection containing the output digital signals. These external control signals, meeting a standard known as SPI, must be supplied by a secondary board using either digital or analog signals.
5.4.2 Over-voltage Input Protection

Unfortunately, while the evaluation board includes an onboard passive limiter, the diodes used have a limited current tolerance. This can lead to excessive current being shunted to the input to the analog front-end, which could cause damage to the circuitry if the induced voltages exceed the 5 V supply voltage of the evaluation board. Due to this limitation, the manufacturer recommends an active limiter stage [92]. The active limiter circuitry suggested features +/- 5 V power supplies, 5 kΩ resistors and a quartet of diodes. To implement this limiter without designing a new board for the ADC as well, a set of boards were designed to fit around the inputs to the evaluation board, protecting the inputs to the IC. A capacitor was also placed on each input before the active limiter to protect the pulse generator. The evaluation board with the additional limiter boards attached is shown in figure 5.8.
5.4.3 Gain +/- Setting

As the AD9272 chip was designed for ultrasound applications, the VGA component was designed with use as a TGC in mind. As Time-Gain Correction requires the gain of the amplified to increase then return to its lowest value in synchronization with the ultrasound pulser, an external control was placed on the board for this setting. As ultrasonic attenuation can be expressed as approximately linear in terms of dB/m, the control is set to be approximately linear in terms of V/dB (Figure 5.9).
As this system is not currently set up for TGC, this variable control was instead connected to a variable DC voltage source to allow control over the overall gain of the system. This allows the offset of the dynamic range depending on the amplitude of the signals from the medium being imaged.

5.5 Computer Interface

To supply the interface between the analog to digital converter board and the software beamformer, an HSC-ADC-EVALC board was acquired from Analog Devices. The board is
designed with a Field Programmable Gate Array (FPGA) Virtex 4, on board memory and a dedicated serial chip for communicating with the computer via high-speed USB cable [55] (Figure 5.10).

Figure 5.10 HSC-ADC-EVALC Communication Board for Interfacing between the ADC Circuit Board and the PC

The Virtex 4 chip was programmed to collect parallel data samples from all 8 channels and to upload samples via the high-speed USB interface chip. The VisualAnalog software package allows control of the USB communication and data transfer settings. To set the SPI-based settings on the ADC board, a FPGA bypass is available which allows direct communication with the ADC SPI ports via USB.
5.5.1 Software Beamformer

The modified curve-fitting algorithm described earlier was implemented in a minimum overhead MatLab function, running in real time. In standard MatLab coding, unlike in more standard coding languages, such as C or java, the code is compiled at run-time. This induces additional delay in computations as the computer must determine memory allocation and function calls while running in real-time. To remove these sources of delay, MatLab code may be pre-compiled. While pre-compiling applies additional constraints on the types of data structures and functions which can be used, the dis-allowed functions tend to be more computationally complex and should be avoided in code intended for rapid execution.

The beamformer code implements a pre-memory allocation method which pre-assigns memory blocks for all necessary computations. In addition, the cosine and sine terms for each pixel to be beamformed are pre-calculated when the code is initially booted, reducing just-in-time calculations further. This allows the code to be run within the MatLab environment without having to use additional code acceleration techniques. The final implementation can compute a full A-scan line in 0.2 ms, and can produce 200 x 200 pixel B-scans at a rate of 20 Frames/sec.

5.5.2 Filter and Gain Settings

Filter and gain settings for the ADCs can be changed at any time during imaging at the cost of a temporary drop in data transfer rate. This is done using SPIController, which supplies a GUI interface and an asynchronous USB connection to the ADC. As can be seen in the GUI (fig
settings such as the sampling frequency are selected with drop-down menus, with the filter corner frequencies defined as a function of the sampling frequency.

For this implementation the low-pass filter was set to a value equal to 1.1 times one third the sampling frequency, or 29.3 MHz. This is close to the highest possible setting due to the near Nyquist sampling rate. To filter noise outside of the bandwidth of the signal as much as possible while leaving the signals undistorted, the high-pass filter was programmed to the low-pass filter frequency divided by 6.0. This results in a corner frequency of 5.8 MHz. The gain settings for both the LNA and VGA were adjusted depending on the target being imaged, with both being set to maximum for small point targets such as those shown at the end of the chapter. This allowed optimal use of the dynamic range of the ADC.
5.5.3 Beamformer Interface

For each line to be beamformed, the VisualAnalog software collects 2048 samples for each of the eight channels. These samples are then transferred directly to MatLab which reads them into a set of pre-allocated arrays. Once beamforming is complete, the A-line is output to the screen as a logarithmically compressed envelope as a function of depth. The non-compressed beamformed line is also stored to a circular buffer to allow the creation of B-scan images.

When implemented on a 1.6 GHz CPU with 6 GB of RAM running MatLab 2010b, the software beamformer is capable of 20 Frames/sec with 200 line B-scan images.

5.6 Testing

The pulser and transducer arrangement used for testing the transducer was connected to the beamformer system using the same splitter board as used in testing the transducer in chapter 4. Due to the need to include a trigger signal for data processing, the 8th channel was removed from the system, resulting in a 7-channel annular array system. After powering up all components of the system, the SPIController software was booted up on the host PC, an i7 HP laptop running Windows 7 32-bit. This was used to set the filters and gain stages as described earlier. VisualAnalog was then turned on and used to program the FPGA on the HSC-ADC-EVALC board for simultaneous acquisition. The binary file writing program was loaded into the software and set on continuous refresh to begin data acquisition.
Once VisualAnalog was collecting live data from the transducer, the beamformer code was run in MatLab™. A-scan beamformed data was then displayed on a log-scale. The rf a-scan of a 100 micron wire target placed at f/2 (10 mm) can be seen in Figure 5.12. A second target at f/3 was ~1 mm off-axis, resulting in a low amplitude response. The image shows the noise floor is below 55 dB and that a well-defined peak is seen with a -6 dB resolution of 2.5λ.

Figure 5.12 The beamformed RF line produced by the MatLab beamformer imaging a wire target at f/2 and an off-axis wire target at f/3

Due to the through-put limitation of the USB cable, the complete analog front-end and receive beamformer is capable of 10 frames/sec of 200x200 pixel B-scans.
Chapter 6

System Integration

To assess the total system’s ability to produce medical quality images, it was desirable to produce a full B-scan image of a set of targets. However, the lack of a motorized mount for the transducer prevented the real-time production of these images. Instead, lines were obtained individually and formed into a grey-scale image with manual translation of the transducer.

6.1 Imaging Set-up

To create full B-scan images with the imaging system it is necessary to scan the transducer array linearly across the material to be imaged for the system to build up the A-scans that comprise a full image. The imaging tank used for assessing the system was designed with certain features to accommodate this need.

The entire imaging system is mounted to a large metal plate to prevent the transducer mount from shifting with respect to the imaging tank. The water tank is attached to an optical translation stage with a linear micrometer allowing precision movement with a resolution of 10 μm. By affixing all material to be imaged to the bottom of the water tank with vacuum grease, the position of the imaged material with respect to the transducer can be accurately mapped. The transducer is mounted to an optical stage capable of vertical adjustment so that the area of interest in the image can be moved into the focal zone. This stage is then mounted on a tilt stage with a
fine adjustment at right angles to the water tank translation and a tilt adjustment which allows the angle of the transducer face to be brought into alignment with the imaging plane.

6.2 Alignment

When mounting the housing on the transducer, considerable care was taken to ensure the axis of the mounting shaft was perpendicular to the face of the transducer. This was done to minimize the effects of misalignment. When the beam of an annular array is scanned along a plane to form a B-scan image, the one dimensional A-scans are assumed to have been acquired perpendicular to the translation axis, resulting in a rectangular image. If the face of the annular array is tilted away from this intended angle, the actual image formed will be trapezoidal, but will be displayed as if the array face had been correctly aligned. This can result in distortion of objects being imaged, as well as making areas of interest appear to be deeper in the tissue than they actually are. To prevent these types of errors, a flat quartz target was placed at the focal point of the array and the angle of the array face adjusted until the return pulse was fully focused when beamformed, indicating that no early or late acoustic waves were arriving at the outer rings.

6.3 Coupled Noise Sources

When connecting the components of the system together, care was taken to ensure that properly shielded cables were used, and that all ground lines were connected together to avoid floating lines. To minimize EM radiation, the high-voltage pulser units were encased in a grounded metal box connected to the system ground. The active limiter boards attached to the ADC board were
equipped with additional decoupling capacitors to prevent the received echoes from coupling to each other prior to digitization. Finally, the cables connecting the pulser to the transducer were bundled together with an insulating sheath, as were the cables to the ADC board.

6.4 Dynamic Range

As mentioned earlier, one of the measurements of the quality of an ultrasound system is the dynamic range. This is the range of logarithmically compressed amplitudes which can be displayed on a grey-scale image after final beamforming. There are two main elements which affect the dynamic range of a system. The first is the range of voltages which can be digitized by the combination of the analog to digital converter and any amplitude modulation being performed. In this system, each channel is digitized with a 12-bit ADC resulting in 4096 possible amplitude levels. However, the converter is prone to error in the final bit, resulting in an actual range of 2048 values. This still results in 66 dB of dynamic range per channel, with some additional range from constructive waveform interference from bright scatterers.

The other source of dynamic range limitation is the noise previously mentioned. If the SNR going into the ADC is too high, especially with coupled noise components, the resultant dynamic range will be less than it would be based purely on the ADC limits. As the dynamic range of the AD9272 chips used in this system is more than sufficient for medical imaging, the final system dynamic range was dictated by the noise levels in the testing rig.
With the noise mitigation strategies mentioned previously in place, a noise floor of 1 mV was obtained, allowing a dynamic range of 55 dB with a gain+/- input of 1.0V.

### 6.5 Imaging Targets

To properly assess the resolution of the system in both the axial and the lateral direction, a set of point targets were required. To ensure that the spacing and location of the targets were known for correlation with the obtained image, a metal frame was machined from a U-shaped piece of aluminum stock with mounting points on either arm of the “U” located 5 mm apart vertically and 1 mm apart horizontally. 100 µm-diameter wires were then stretched taut between the mounting points. Once secured, the entire construction was placed in the water tank with the wires stretching at right angles to the plane of the image to be formed, resulting in a circular cross-section in the plane of the image.

### 6.6 Wire Targets

The 100 µm wire target was imaged with a 400x400 pixel image by capturing 400 pixel a-scan lines using the beamformer described in chapter 5. The water tank was then translated 40 µm and another scan line acquired. This procedure was repeated until the full 400 scan lines had been captured. The resulting image is displayed with 40 dB of dynamic range (figure 6.1).
Figure 6.1 Grey-scale image of point targets at f/2, f/3 and f/4 with appropriate transmit focal depths and 40 dB of dynamic range

The f/2 point shows 2.5\(\lambda\) lateral resolution and 2.15\(\lambda\) axial resolution. The f/3 point has 3\(\lambda\) lateral resolution and 2.25\(\lambda\) lateral resolution. Finally, the f/4 point shows 3.5\(\lambda\) lateral resolution and 2.75\(\lambda\) axial resolution.
Chapter 7

Discussion and Future Work

This thesis describes the design and manufacture of a proof of concept low-cost, 20 MHz annular array ultrasound imaging system. The system obtained close to diffraction limited lateral resolution (173 μm at f/2), a 50 % fractional bandwidth and dynamic range greater than 60 dB.

7.1 Pulser and Transmit Beamformer Discussion

The goal of the pulse generator design was to obtain a low-cost solution that could be easily expanded to meet the requirements of different systems. This pulser produced features individual boards for each output pulse and a common backplane board, allowing simple addition and removal of outputs. The resulting pulse generators produced monocycle output pulses with centre frequency that could be adjusted between 20-50 MHz and peak to peak amplitude up to 100 V.

The design called for pulse generators capable of frequencies up to 50 MHz and amplitudes up to 100 Vpp. The pulser design met the frequency criteria by using rapid response operational amplifiers and pulse shaping chips that could produce 5 ns pulses.

The pulse generator also produced pulses with amplitudes up to 90 Vpp. While the output transistors used in the design were specified as capable of tolerating the full 100 Vpp in the original design and the pulser operated at full voltage in simulation, testing of the physical boards resulted in increased failure rates in boards operating at 100 Vpp. However, operation at 90 Vpp resulted in no such failures.
The transmit beamformer was designed to produce beamforming delays over the range from 0 to 255 ns in steps of 0.5 ns. For a 20 MHz centre frequency, the delay steps correspond to one hundredth of the ultrasound period. By switching the delay chips with others in the same product family a 0.25 ns delay resolution can be achieved, or up to >1500 ns delay range. The DS1023 is available in total delay ranges from 63.75 to 1275 ns and the DS1100 ranges from 20 to 500 ns, providing sufficient delay range for high-frequency imaging close to the array face [56][66].

7.1.1 Possible Future Research for the Pulse Generator and Transmit Beamformer

In its current implementation, the 8-bit delay values used to set the beamforming delays and frequency of the output pulses are inputted in parallel from physical DIP switches located on the back of each channel’s PCB. This solution worked well a prototype system, but prevented the creation of multiple transmit focal zones in real-time. However, the DS1023 chips used to create the delays fully support serial input with a simple communication protocol. A small modification to the circuits could allow the delays to be programmed from a single Complex Programmable Logic Device (CPLD) located on the backplane board. This would allow the same PC that houses the beamformer to also control the transmit settings and would permit real-time imaging using multiple transmit focal zones.
7.2 Transducer

A low-cost, straight-forward method for fabricating the transducer array was developed. The manufacturing process reduces the number of layers in the transducer stack by integrating the matching layer and the connective traces, simplifying production. In addition, the application of the anisotropic epoxy and clamping at room temperature allowed the transducer to be brought up to curing temperature for a minimal amount of time, reducing the risk of the ceramic de-poling.

7.2.1 Performance

The radiation pattern obtained from the glass point target placed 10 mm away from the array showed lateral resolution close to ideal at 2.4λ and side-lobes at 50-60 dB were demonstrated, meeting minimum requirements for medical imaging. The pulse generated from the central element had a 50% bandwidth and a 21 MHz centre frequency. Good agreement was found between the experimental pulse shape and that produced in simulations.

A -22 dB insertion loss was also found at a centre frequency of 23 MHz. The crosstalk was -26 dB, which is comparable to other kerfless designs [96].

7.2.2 Low-Cost Manufacturing

The manufacturing process used in the prototype transducer incorporated low-cost materials and minimized layers to keep overall costs down. Of the materials used in the production of this
transducer, the most expensive was the custom manufactured flexible circuit boards. When manufactured in a custom order batch of 20, the boards were purchased for less than $150 a board. The per board cost would drop significantly for larger board orders. The other components (ceramic, epoxy, tungsten) are standard materials in transducer manufacture with the exception of the anisotropic epoxy, which is used in very small quantities, keeping that cost outlay small as well.

The manufacturing process has been shortened and simplified by combining the connection cable and the matching layer into a single circuit board. In addition, the replacement of wire-bonding with an epoxy-bonding stage replaces a sensitive process prone to failure with a simple, easy to control process which can be done in large batches. In summary, a low-cost, simplified process has been developed which can be used for single transducer production in research labs as well as being easily expanded to large-scale production.

7.2.3 Future Transducer Design Research

Any future development on the annular transducer array would be best suited to making it more applicable for frequencies in the >40 MHz as well as additional cost saving measures. To this end, the first requirement for a higher frequency version of this transducer would be a thinner flexible circuit board to match the smaller quarter wavelength requirement. As the board used in this thesis was the thinnest standard board from the manufacturer, one of two solutions could be explored. The first would be to investigate ordering a custom circuit board from the manufacturer with a thinner base layer. This would allow greater control over the final thickness of the flexible
circuit board, allowing better control over the final frequency. The ability to control the thickness of the board to resolution of 2 μm, for example, corresponds to an accuracy of 1.25 MHz at 20 MHz and 5 MHz at 40 MHz. However, the 20% error in intended thickness obtained with these flexible circuits is a poor indication of the manufacturer’s ability to control the thickness of an even thinner board. An alternative approach would be to manufacture the flexible circuit in-house, using a thin film deposition technique and photo-lithography [53]. This could potentially allow the control over the film thickness to allow frequency tuning, as well as allowing some ability to adjust the acoustic impedance by experimenting with dopants, improving its matching layer qualities. However small-scale photo-lithographic processes can suffer from coarser detail resolution, limiting the minimum size of array elements.

To improve on the transducer’s performance at all frequencies, the ceramic grid dimensions could be adjusted to increase the area fraction of electroded surface without breaking the anti-shorting constraint. The stripline method use to lay out the connective traces could be modified to cut down on trace to trace noise. By introducing ground lines between each of the element traces, parasitic coupling would be reduced, potentially improving the transducer’s crosstalk [97].

7.3 Beamformer

The beamformer was an all software implementation using commercial circuit boards so as to keep costs down while maximizing the ability to customize the beamformer. While the beamformer functioned very well, with the alternate interpolation scheme showing a 10 dB improvement over linear interpolation in simulation and the software beamformer operating at
0.25 ms per rf line, the bandwidth limitations of the USB interface to the computer limited the system to a total frame rate of 10 frames/sec, rather than the desired 20 frames/sec. In all other respects the beamformer software and hardware behaved as expected and the system proved to be cost-effective and easy to implement and modify.

7.3.1 Analog to Digital Converter

The AD9272 evaluation board and the data transfer board were well suited to this use, with the correct combination of components to shape the input pulses without causing loss of dynamic range. The implemented interpolation scheme allowed near-Nyquist sampling rate and the bit rate was sufficient to maintain greater than 60 dB dynamic range without additional modulation. The digital filters were controllable within the desired range and helped minimize noise. The voltage controlled gain stage allowed 36.5 dB of gain within the linear region of the gain input, with fine enough control (2.85 dB/100 mV) to allow full use of the dynamic range.

The limiter circuits included on the board proved insufficient to fully protect the input to the ADC chip, however the additional active limiter circuits were simple to design and implement.

7.3.2 Evaluation Boards and Supplied Software

The evaluation boards were critical to keeping the overall system cost down while supplying the necessary hardware. Other than a change of oscillator the boards did not require board level work and functioned as desired when plugged in.
The VisualAnalog software which was supplied with the HSC-ADC-EVALC FPGA board was used only to interface to the evaluation board and the FIFO buffers containing the sampled data from the AD9272, as the GUI interface was found to be prone to significant slow-down if any calculations were performed within the environment. Transferring the data directly to MatLab™ for processing prevented this slowdown from happening, and allowed the interface to the board to run at the limit of the USB connection. USB 2.0 is theoretically capable of transfer speeds of up to 60 MB/s, which would have allowed 20 frames/sec beamforming. Communications overhead built into the protocol limits actual throughput to 30-40 MB/s, however, reducing the beamformer’s capabilities accordingly [98]. The 10 frames/sec achieved is still fast enough for real-time imaging and hardware upgrades could improve this.

7.3.3 New Interpolation Method

A lower-frequency ADC was selected for this beamformer to acquire a chip with 8 separate channels built into a single package. The curve-fitting interpolation scheme used allowed this decreased sampling rate due to its improvement over simple linear interpolation. By taking advantage of knowledge of the system centre-frequency as well as selecting a sampling frequency exactly four times the centre frequency, the modified interpolation approach used in this system allows near-Nyquist sampling without the associated rise in secondary lobe levels, at minimal additional computation cost. The resulting implementation in MatLab can beamform full 200 x 200 pixel images at 20 frames/sec. This implementation has the additional benefit of being easy
to modify due to being coded in a high-level programming language, needing no knowledge of the hardware structure or communication protocol.

The combination of the software beamformer and the hardware used in its implementation results in an ideal receive beamformer for rapid, low-cost implementations such as those in specialty clinics or in a research laboratory.

7.3.4 Potential Future Research into the Receive Beamformer

The fundamental bottleneck on the frame rate of the receive beamforming system is the bandwidth limitation of the USB 2.0 standard. USB 2.0 signaling requires a minimum dead time on the cable after the end of each transmission that cannot be eliminated, reducing the effective maximum bandwidth to 40 MB/s. To allow greater data throughput despite this limitation would require some additional processing to occur on the FPGA chip, increasing the development cost of the implementation. Some processing that could be implemented on the FPGA is the pre-selection of the samples required to interpolate the final pixels. As there are approximately 6 samples per pixel at the given sample rates and pixel spacing, reducing the data input to only the samples on either side for interpolation would permit a three times reduction in required samples. While this would allow 20+ frame/sec frame rates, it would be at the cost of losing the ability to easily change the focal distances during scanning. Another form of processing that would not have this impact would be implementing some form of compression of the signals. The 12-bit samples for areas of low reflection contain a large zero content which could be compressed using scheme designed for repetitive data streams, such as HASP [99]. This would increase the data
throughput for a given bandwidth without limiting the data output. The best solution would be the development of a similar evaluation board based on the new USB 3.0 standard. A proposed potential 10x bandwidth improvement would be more than sufficient for full real-time imaging [97]. As computer advances push the rapid spread of new communication standards, this is a likely future development and could produce a full real-time software beamformer with the presented system.

7.4 Conclusions

A high-frequency ultrasound imaging system has been developed which integrates low-cost manufacturing techniques with commercially available hardware and software. The completed system has a dynamic range of 55 dB, a lateral resolution of $2.4\lambda$ and an axial resolution of $2.5\lambda$. The receive beamformer is capable of imaging at a rate of 10 frames/sec, with software capable of 20 frames/sec. The entire system was implemented for less than $10k. Assuming a $40k overhead for marketing, development and manufacturing costs, the final system would cost only $50k, making it significantly cheaper than the VisualSonics system which costs ~$450k [2], while providing good system resolution.
References


Appendix

MatLab code for the beamformer implementation

*Realtimebeam.m*

```matlab
function [store,tinfo,time]=realtimebeam

%set function to return store of outputs, along with trigger info for those
%outputs. No inputs required.
%data=zeros(7,1864);
data=zeros(7,2600);
 fidt=fopen('trigger.hex','r');
 fid1=fopen('element1.hex','r');
 fid2=fopen('element2.hex','r');
 fid3=fopen('element3.hex','r');
 fid4=fopen('element4.hex','r');
 fid5=fopen('element5.hex','r');
 fid6=fopen('element6.hex','r');
 fid7=fopen('element7.hex','r');

%[aterm,bterm,atermq,btermq,lowp]=triglut;%pre-
%load delay LUT
[aterm,bterm,atermq,btermq,lowp,lowq,diff,delaytimes,delaytimesq]=triglut;

N=8192;%number of samples read into data set at a time
%N2=178;%number of pixels in final RF line
N2=length(aterm);
%store=zeros(40,355);%number of stored lines x length of line
%tinfo=zeros(40,1864);

%if changing number of rows, also change in if m> check below
m=1;
len=zeros(1,8);
%dt=12.5e-9;
dt=25e-9;

dBmax=20*log10(196662);%maximum output, set to 0 dB
%offset=-49146;%number of bytes to rewind for 8192 samples

%FS=stoploop('stop');
%adds stop button to graph, when pushed triggers end of while loop

t=[0:dt:(N2-1)*dt];
%t=t+10.025e-6;%compensate for dead zone
%t=t+1e-6;
d=t*1500/2;%converts time to distance
```
temp1=zeros(1,N2);
temp2=zeros(1,N2);
temp3=zeros(1,N2);
temp4=zeros(1,N2);
temp5=zeros(1,N2);
temp6=zeros(1,N2);
temp7=zeros(1,N2);

while (1)
    datat=fscanf(fidt,'%x',N);
    data1=fscanf(fid1,'%x',N);
    data2=fscanf(fid2,'%x',N);
    data3=fscanf(fid3,'%x',N);
    data4=fscanf(fid4,'%x',N);
    data5=fscanf(fid5,'%x',N);
    data6=fscanf(fid6,'%x',N);
    data7=fscanf(fid7,'%x',N);
    found=0;
    notfound=0;
    x=1;
    len(1,1)=length(datat);
    len(1,2)=length(data1);
    len(1,3)=length(data2);
    len(1,4)=length(data3);
    len(1,5)=length(data4);
    len(1,6)=length(data5);
    len(1,7)=length(data6);
    len(1,8)=length(data7);

    %check for full data sets
    if (min(len)<8192)
        found=1;
        notfound=1;
    end
    %if incomplete, jump to new data set

    while (found==0)
        test=datat(x);
        if (test > 6e4)%trigger amplitude
            found =1;
        end
        x=x+1;
        if (x>length(datat))
            notfound=1;%prevent beamforming this loop
            found=1;%end while loop
        end
    end
end
if ((notfound) == 0)
if ((length(data1)-x)>2600) %don't beamform with insufficient time
space

%beamform
data(1,:) = data1(x:x+2599);
data(2,:) = data2(x:x+2599);
data(3,:) = data3(x:x+2599);
data(4,:) = data4(x:x+2599);
data(5,:) = data5(x:x+2599);
data(6,:) = data6(x:x+2599);
data(7,:) = data7(x:x+2599);
data=data-2^15; %change from offset binary to signed integer
%tic
%[image,t]=beamformnew(data,lowp,aterm,bterm,atermq,btermq);

i=1:size(aterm,2); %calculate only desired pixels

%inphase term
    temp1=aterm(1,i).*data(1,lowp(1,i)) +
bterm(1,i).*data(1,lowp(1,i)+1); 
    temp2=aterm(2,i).*data(2,lowp(2,i)) +
bterm(2,i).*data(2,lowp(2,i)+1); 
    temp3=aterm(3,i).*data(3,lowp(3,i)) +
bterm(3,i).*data(3,lowp(3,i)+1); 
    temp4=aterm(4,i).*data(4,lowp(4,i)) +
bterm(4,i).*data(4,lowp(4,i)+1); 
    temp5=aterm(5,i).*data(5,lowp(5,i)) +
bterm(5,i).*data(5,lowp(5,i)+1); 
    temp6=aterm(6,i).*data(6,lowp(6,i)) +
bterm(6,i).*data(6,lowp(6,i)+1); 
    temp7=aterm(7,i).*data(7,lowp(7,i)) +
bterm(7,i).*data(7,lowp(7,i)+1); 
    I=temp1+temp2+temp3+temp4+temp5+temp6+temp7;

%quadrature term
    temp1=atermq(1,i).*data(1,lowq(1,i)) +
btermq(1,i).*data(1,lowq(1,i)+1); 
    temp2=atermq(2,i).*data(2,lowq(2,i)) +
btermq(2,i).*data(2,lowq(2,i)+1); 
    temp3=atermq(3,i).*data(3,lowq(3,i)) +
btermq(3,i).*data(3,lowq(3,i)+1); 
    temp4=atermq(4,i).*data(4,lowq(4,i)) +
btermq(4,i).*data(4,lowq(4,i)+1); 
    temp5=atermq(5,i).*data(5,lowq(5,i)) +
btermq(5,i).*data(5,lowq(5,i)+1); 
    temp6=atermq(6,i).*data(6,lowq(6,i)) +
btermq(6,i).*data(6,lowq(6,i)+1); 
    temp7=atermq(7,i).*data(7,lowq(7,i)) +
btermq(7,i).*data(7,lowq(7,i)+1); 
    Q=temp1+temp2+temp3+temp4+temp5+temp6+temp7;
I=I-mean(I); %subtract off any DC bias
Q=Q-mean(Q);
sum=(I.^2 + Q.^2).^0.5;

dB=20*log10(sum);
dB=dB-dBmax;
%end beamformnewfunction

plot(d,dB)
axis([7.5e-3,17.5e-3,-60,0])
drawnow
store(m,:)=sum;
tinfo=datat(x:x+1863)';
m=m+1;
if (m>40) % integer is number of stored lines also in tinfo and
store definitions above
    m=1;
end
end
fclose(fidt);
close(fid1);
close(fid2);
close(fid3);
close(fid4);
close(fid5);
close(fid6);
close(fid7);
end

Look-up-table creation code:

Triglut.m:

function [aterm,bterm,atermq,btermq,lowp,lowq,diff]=triglut
%returns look-up-table of cosine and sine values for a focused line
given
%an array of time delay values for ideal pixels in the time domain, as well as
%the centre frequency of the pulse

f=20e6;
lambda4=12.5e-9;

triggerdelay=1e-6;
samplet=12.5e-9;
pixelstep=25e-9; %half a wavelength
N=2640; % sample points
M=1300; % pixels in image

t=[triggerdelay:samplet:(samplet*(N-1)+triggerdelay)];

t2=[triggerdelay:pixelstep:(pixelstep*(M-1)+triggerdelay)];
dist=(t2/2)*1500;

array=[0.0004 0.0011 0.0014 0.0017 0.0019 0.0021 0.0022];

d=zeros(length(array),M);
delta=zeros(size(d));
deltat=zeros(size(d));
delaytimes=zeros(size(d));

for i=[1:M]
    [d(:,i),mid]=distance(array',0,dist(:,i));
    delta(:,i)=d(:,i)-max(d(:,i));
    deltat(:,i)=-delta(:,i)/1500;
end

for i=[1:length(array)]
    delaytimes(i,:)=t2+deltat(i,:);
end

delaytimesq=delaytimes+lambda4;

rdown=floor((delaytimes-triggerdelay)/samplet);
tdown=rdown*sampel+triggerdelay;

lowp=rdown+1;
diff=delaytimes-tdown;

aterm=cos(2*pi*f*diff);
fterm=sin(2*pi*f*diff);

rdownq=floor((delaytimesq-triggerdelay)/samplet);
tdownq=rdownq*sampel+triggerdelay;

lowq=rdownq+1;
diffq=delaytimesq-tdownq;

atermq=cos(2*pi*f*diffq);
ftermq=sin(2*pi*f*diffq);