A DESIGN ALGORITHM FOR ARBITRARY PHASE
AND AMPLITUDE SURFACE ACOUSTIC WAVE FILTERS

by

MARK S. SUTHERS, B.ENG., M.ENG.

A Thesis
Submitted to the School of Graduate Studies
in Partial Fulfilment of the Requirements
for the Degree
Doctor of Philosophy

McMaster University
1978
ARBITRARY PHASE AND AMPLITUDE

SURFACE ACOUSTIC WAVE FILTERS
DOCTOR OF PHILOSOPHY (1978)  McMaster University
(Electrical Engineering)  Hamilton, Ontario

TITLE:  A Design Algorithm for Arbitrary Phase and Amplitude
Surface Acoustic Wave Filters

AUTHOR:  Mark S. Suthers, B.Eng. (McMaster University)
           M.Eng. (McMaster University)

SUPERVISOR:  Professor C.K. Campbell
ABSTRACT

The subject of this study is the development of a computer-aided-design tool that filter designers could use to obtain nonlinear phase and nonsymmetric amplitude responses for surface acoustic wave devices. To accomplish this, a design algorithm was developed that uses a Fast Fourier Transform to realize the time domain transform of the frequency domain specifications required by the designer. From this time domain solution an interdigital transducer (IDT) is synthesized that will approximate the design specifications. The next step in the algorithm is to analyse the synthesized IDT using an accurate charge distribution model for the IDT (the charge distribution model is developed in this thesis which gives the frequency domain response for the IDT). The frequency response is then compared to the input specifications so as to generate an error function. This error function is then used to predistort the input frequency domain specifications and the whole synthesis/analysis procedure is repeated iteratively until an acceptable solution is obtained. The algorithm is programmed and used to design 2 filters (3 separate IDT designs) for the nonlinear phase and nonsymmetric amplitude specifications for television intermediate frequency (TVIF) specifications. Using these designs, two filters were fabricated and tested to show the validity of the algorithm.
# TABLE OF CONTENTS

**CHAPTER 1: INTRODUCTION**

1.1 Principles of Surface Acoustic Waves
1.2 Why SAW Filters?
1.3 Design Techniques
1.4 Arbitrary Phase
1.5 Arbitrary Amplitude
1.6 Building Block
1.7 Arbitrary Phase and Amplitude
1.8 Scope of Thesis
1.9 Description of Algorithm

**CHAPTER 2: SYNTHESIS OF INTERDIGITAL TRANSDUCER**

2.1 The Fast Fourier Transform
2.2 Input to the Fast Fourier Transform
2.3 Conditioning the DATA
2.4 Interpreting the FFT Output
2.5 Dirac-Delta Model
2.6 Limitations of Dirac-Delta Model

**CHAPTER 3: CHARGE DISTRIBUTION MODEL OF AN INTERDIGITAL TRANSDUCER**

3.1 Charge to Acoustic Deformation
3.2 Electrostatic Potential
3.3 Charge on a Finger
3.4 Finding the Charge Distribution Coefficients
3.5 The Integral Functions
3.6 Summary
### TABLE OF CONTENTS (continued)

<table>
<thead>
<tr>
<th>CHAPTER</th>
<th>TITLE</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>CHAPTER 4:</td>
<td>FREQUENCY RESPONSE OF AN INTERDIGITAL TRANSDUCER</td>
<td>42</td>
</tr>
<tr>
<td>4.1</td>
<td>General</td>
<td>42</td>
</tr>
<tr>
<td>4.2</td>
<td>Integral Solutions</td>
<td>43</td>
</tr>
<tr>
<td>4.3</td>
<td>The Acoustic Wave</td>
<td>44</td>
</tr>
<tr>
<td>4.4</td>
<td>Summary</td>
<td>45</td>
</tr>
<tr>
<td>CHAPTER 5:</td>
<td>NUMERICAL OPTIMIZATION</td>
<td>46</td>
</tr>
<tr>
<td>5.1</td>
<td>General</td>
<td>46</td>
</tr>
<tr>
<td>5.2</td>
<td>Error Function</td>
<td>46</td>
</tr>
<tr>
<td>5.3</td>
<td>Predistortion of Specifications</td>
<td>49</td>
</tr>
<tr>
<td>5.4</td>
<td>Summary</td>
<td>50</td>
</tr>
<tr>
<td>CHAPTER 6:</td>
<td>PROGRAMMING THE ALGORITHM</td>
<td>51</td>
</tr>
<tr>
<td>6.1</td>
<td>General</td>
<td>51</td>
</tr>
<tr>
<td>6.2</td>
<td>SAWINIT</td>
<td>51</td>
</tr>
<tr>
<td>6.3</td>
<td>SAWOPT</td>
<td>55</td>
</tr>
<tr>
<td>6.4</td>
<td>TAPEJOB</td>
<td>57</td>
</tr>
<tr>
<td>6.5</td>
<td>Summary</td>
<td>57</td>
</tr>
<tr>
<td>CHAPTER 7:</td>
<td>EXPERIMENTAL RESULTS</td>
<td>58</td>
</tr>
<tr>
<td>7.1</td>
<td>General</td>
<td>58</td>
</tr>
<tr>
<td>7.2</td>
<td>Filter Specifications</td>
<td>60</td>
</tr>
<tr>
<td>7.3</td>
<td>Single Bandpass Filter</td>
<td>60</td>
</tr>
<tr>
<td>7.4</td>
<td>Dual Bandpass Filter</td>
<td>64</td>
</tr>
<tr>
<td>7.5</td>
<td>Measuring the Responses</td>
<td>64</td>
</tr>
<tr>
<td>7.6</td>
<td>Summary</td>
<td>70</td>
</tr>
<tr>
<td>CHAPTER 8:</td>
<td>CONCLUSIONS</td>
<td>73</td>
</tr>
<tr>
<td>8.1</td>
<td>General</td>
<td>73</td>
</tr>
<tr>
<td>8.2</td>
<td>Synthesis</td>
<td>73</td>
</tr>
</tbody>
</table>
TABLE OF CONTENTS (continued)

8.3 Charge Distribution Model 74
8.4 Frequency Response 74
8.5 Optimization 75
8.6 Summary 75

APPENDIX I: SOLUTION TO CHARGE DISTRIBUTION INTEGRALS 77
  I.1 General 77
  I.2 Solution to I_10 77
  I.3 Solution to I_20 80
  I.4 Solution to I_30 85
  I.5 Summary 92

APPENDIX II: SOLUTION TO WAVE EQUATION INTEGRALS 93
  II.1 General 93
  II.2 Solution to I_10 93
  II.3 Solution to I_20 95
  II.4 Solution to I_30 96
  II.5 Summary 97

APPENDIX III: SAWINIT 98

APPENDIX IV: SAWOPT 120

APPENDIX V: TAPEJOB 141

APPENDIX VI: FABRICATION OF FILTERS 147

REFERENCES 151
# LIST OF FIGURES

<table>
<thead>
<tr>
<th>Figure</th>
<th>Description</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.1</td>
<td>Surface Acoustic Wave Filters</td>
<td>2</td>
</tr>
<tr>
<td>1.2</td>
<td>End-fire Antenna Array and Finite Impulse Digital Filters Analagous to Surface Acoustic Wave Filters</td>
<td>4</td>
</tr>
<tr>
<td>1.3</td>
<td>Interdigital Transducer Geometry Interacting with Piezoelectric Substrate</td>
<td>5</td>
</tr>
<tr>
<td>1.4</td>
<td>Obtaining IDT Geometry from Fourier Transform</td>
<td>9</td>
</tr>
<tr>
<td>1.5</td>
<td>Dual Bandpass Hermitian Specifications About $f = 2f_0$</td>
<td>13</td>
</tr>
<tr>
<td>1.6</td>
<td>Algorithm Block Diagram</td>
<td>15</td>
</tr>
<tr>
<td>2.1</td>
<td>Lowpass FFT Response</td>
<td>19</td>
</tr>
<tr>
<td>2.2</td>
<td>Bandpass FFT Response</td>
<td>20</td>
</tr>
<tr>
<td>2.3</td>
<td>Analogy to Intermediate Frequency Response</td>
<td>26</td>
</tr>
<tr>
<td>2.4</td>
<td>Effects of Discontinuities in Phase Response of Transform</td>
<td>27</td>
</tr>
<tr>
<td>3.1</td>
<td>Acoustic Wave Generation of IDT</td>
<td>30</td>
</tr>
<tr>
<td>3.2</td>
<td>Relaxation of Charge Distribution Coefficients</td>
<td>37</td>
</tr>
<tr>
<td>3.3</td>
<td>Charge Distribution Functions</td>
<td>41</td>
</tr>
<tr>
<td>5.1</td>
<td>Development of Error Function and Predistortion</td>
<td>47</td>
</tr>
<tr>
<td>5.2</td>
<td>Hard Limiting and Soft Limiting of Error Function</td>
<td>48</td>
</tr>
<tr>
<td>6.1</td>
<td>Flow Chart of SAWNIT</td>
<td>52</td>
</tr>
<tr>
<td>6.2</td>
<td>Flow Chart of SAWOPT</td>
<td>56</td>
</tr>
<tr>
<td>7.1</td>
<td>TVIF Filter Specifications</td>
<td>59</td>
</tr>
<tr>
<td>7.2</td>
<td>Mask Pattern for Single Bandpass Filter</td>
<td>61</td>
</tr>
<tr>
<td>7.3</td>
<td>Response of Single Bandpass Filter</td>
<td>62</td>
</tr>
</tbody>
</table>
LIST OF FIGURES (continued)

Figure                                      Page  
7.4  X-Y Plot of Single Bandpass Filter Response  63  
7.5  Hermitian Specifications for Dual Bandpass IDT  65  
7.6  Mask Pattern for Dual Bandpass Filter  66  
7.7  Response of Dual Bandpass Filter  67  
7.8  X-Y Plot of Dual Bandpass Filter Response  68  
7.9  Measurement System for Testing Filters  69  
7.10 Mask Pattern for Delay-Line Equalizer  71  
7.11 X-Y Plot of Delay-Line Filter Response  72
ACKNOWLEDGEMENTS

The author would like to express his appreciation for the consideration and supervision received from the supervisor, Dr. C.K. Campbell. The author also acknowledges the constructive criticism and interest of the other members of the supervisory committee: Dr. S. Haykin and Dr. J.P. Marton.

I would like to thank my colleagues: Messrs. P. Edmonson, P. Nanayakkara, J.P. Reilly and D.G. Seiler, who provided a stimulating work environment.

I also acknowledge: Mr. C. Goulson and the Canadian Institute for Metal Working for the use of their terminal; Mr. R. Martin and Mr. J. Jewell of the McMaster Computing Centre for their help with permanent file manipulation and management. The congeniality of all of the Computing Centre's operations staff is also very much appreciated.

Mr. R. Grant of the Electrical Engineering Department of the University of Waterloo is thanked for his help in the production of the photo-masks. The artwork of Mr. G. Kappel is much appreciated.

I also thank P. Dillon of the Word Processing Centre for the typing of this manuscript and finally I would like to thank my wife, Carol, for seeing me through this ordeal.
CHAPTER 1  
INTRODUCTION

1.1 Principles of Surface Acoustic Waves

Surface Acoustic Wave (SAW) devices (Figure 1.1) are essentially transversal filters which, in the transmitting mode, are analogous to end fire antenna arrays (Figure 1.2). The frequency response in the end fire direction is a function of the spacing between the elements and the relative amplitude and phase of the fields produced by each elemental source. This response is given by:

$$H(\omega) = \sum_{l=1}^{L} A_l e^{-j(k_x x_l + \phi_l)}$$  \hspace{1cm} (1.1)

where:  
- $H(\omega)$ is the frequency response along the x-axis
- $A_l$ is the relative amplitude of the $l^{th}$ element
- $k_x$ is the propagation constant along the x-axis
- $x_l$ is the position of the $l^{th}$ element measured from the summation point
- $\phi_l$ is the relative phase of the $l^{th}$ element.

Equation (1.1) is a summation of exponentials and it may be noted that there are no poles (frequency at which the denominator goes to zero) in the expression. This is due to the fact that the receiving array is not electrically connected to the transmitting array. This gives an increased degree of freedom from lumped element filter design.
Figure 1.1: SAW - Acoustic Wave Filters
In the receiving mode, the transducer is similar to a finite-impulse response (FIR) digital filter as seen in Figure 1.2 where the taps are allowed to vary in position and are weighted as to their relative sample amplitude. The response of this receiving system is given by an equation of the same form as (1.1).

Thus by varying the lengths of the transducer fingers (elemental amplitude or tap weight) and by varying the finger position (elements position in the array or sample time in the digital filter analogy), a desired frequency response may be met. The above analogies are the simplest explanations possible which explain how SAW devices produce filtering action.

1.2 Why SAW Filters?

This question is answered by the fact that these devices may be made economically, and small, and they do not require alignment once the device is made. Since the acoustic wavelength on the surface of the substrate is approximately $10^{-5}$ times the wavelength of the corresponding electro-magnetic wave, the principles of microwaves may be applied to the 10 MHz to 1 GHz band of frequencies. Also, the technology of production is planar lithography, i.e. micro-electronic fabrication procedures, which are well established as precise economical production techniques.

The physical structure of a surface acoustic wave (SAW) filter is shown in Figure 1.3. Since the waves are very short (35 μm at 100 MHz), it is necessary that the surface of the substrate be very smooth.
Figure 1.2: Antenna end-fire array and FIR digital filter analogies to SAW filters
Figure 1.3: Interdigital transducer interacting with piezoelectric substrate
Any irregularities in the surface will tend to scatter energy and thus degrade the device response. As the name implies, surface acoustic waves are mechanical in nature and therefore the electrical-to-mechanical and mechanical-to-electrical transduction must be accomplished simply and efficiently. The most obvious, and thus most common, solution to this transduction requirement is to use piezoelectric substrates. Since application of an electric field to a piezoelectric material causes a mechanical deformation to the material then it is seen from Figure 1.3 that the application of voltages to the aluminum fingers on the surface of the substrate will cause a periodic deformation. If the applied voltage is alternated in a sinusoidal manner then at those frequencies at which constructive interference occurs, acoustic radiation will develop causing mechanical waves to propagate along the surface. Since surface acoustic wave velocities are slower than bulk wave velocities, the wave deformation in the interior of the crystal tends to propagate ahead of the surface deformation for a given wave crest. This causes a "dragging" effect on the wave which tends to pull the wave toward the surface of the crystal and is the cause for approximately 90% of the acoustic energy being contained within 1.5 \( \lambda_s \) (where \( \lambda_s \) = surface acoustic wave length) of the surface [1].

From [2], the piezoelectric transduction process is given by the following pair of equations:

\[
[S] = [d] \cdot [E] + [s^E] : [T] \tag{1.2a}
\]

\[
[D] = \varepsilon_o \cdot [e^S] \cdot [E] + [e] : [S] \tag{1.2b}
\]
where: \([S]\) is the 1x6 stress vector measured in meters/meter i.e. dimensionless

\([d]\) is the 3x6 piezoelectric strain tensor measured in meters/volt

\([E]\) is the 1x3 electric field vector in volts/meter

\([s_E]\) is the 6x6 compliance tensor for constant \(E\) field measured in meters\(^2\)/Newton

\([T]\) is the 1x6 stress vector measured in Newtons/meter\(^2\)

\([D]\) is the 1x3 electric displacement density vector measured in coulombs/meter\(^2\)

\([\varepsilon^S]\) is the 3x3 relative permittivity tensor at constant stress (dimensionless)

\([\varepsilon_o]\) is the permittivity of free space

\([e]\) is the 6x3 piezoelectric stress constant tensor in coulombs/meter\(^2\)

The point to note in these equations (1-2) is that

\[ [S] = [d] \cdot [E] \quad \text{for } [T] = 0 \quad (1.3a) \]

\[ [D] = [e] : [S] \quad \text{for } [E] = 0 \quad (1.3b) \]

Thus from (1.3a), strain \((S)\) is a function of an applied electric field \((E)\) through the piezoelectric strain coefficients \((d)\). Therefore, the mechanical deformation is directly related to charge applied through metallized regions on the surface which gives the electrical to mechanical transduction process.

Similarly, from (1.3b) the electric displacement density \((D)\) is
a function of the mechanical strain ($S$) through the piezoelectric stress coefficients ($e$) thus relating the charge at the receiving transducer to the mechanical deformation caused by an incident wave.

1.3 **Design Techniques**

Referring to (1.1), it is noted that there is a strong resemblance to the frequency domain response being related to the spatial domain of the transducer fingers through the Fourier transformation in a truncated discrete form. Initially, attempts at a desired frequency domain response were obtained by evaluating the inverse Fourier transform. From the infinite time domain solution, a window function was applied to realize a finite solution. Then the finger length was taken as a function of the amplitude of the inverse Fourier transform (IFT) and the finger's position at the zero crossings of the IFT. The polarity was obtained from the phase response of the IFT. References such as [3], [4] and [5] illustrate this technique very well. This procedure is shown graphically in Figure 1.4. The variation in the length of the fingers, known as apodization, is equivalent to varying the amplitude of each finger of a constant overlap IDT since the acoustic amplitude in the far field is a function of the charge on the finger. Any technique of varying the charge on a finger in a manner equal to the amplitude of the inverse Fourier transform solution will give the desired results, but the most common technique is the use of apodization which is used in this study. Alternate schemes are: spectral weighting [6], series weighting [7] and
Rectangular linear phase frequency domain specs.

Inverse Fourier Transformation gives

Amplitude response

+ 

Phase Response

Spatial I.D.T. domain from fingers at phase=0 crossing; finger lengths proportional to amplitude response, polarity from slope and phase response.

Figure 1.4: Obtaining IDT geometry from Fourier transform
withdrawal weighting-[8].

Naturally, the next step from the classical Fourier transform design approach was to use a Fast Fourier Transform (FFT) [4], [9]. If the desired bandpass is symmetric about \( f_0 \) in amplitude and linear in phase, then the bandpass may be translated to baseband and the FFT applied for the band from 0 Hz to some frequency above the band-edge. The resulting time domain transform is finite in duration and of course an appropriate window function may be applied to reduce passband ripple and to suppress side lobes [9], [10].

It must be noted at this point that the above techniques have been restricted to linear phase/symmetric bandpass filter specifications. While this is often desirable, there are many filters that require different specifications.

1.4 **Arbitrary Phase**

Working with baseband specifications, it is possible to obtain a deviation from linear phase, if the specifications are Hermitian. This technique is illustrated in [3] and [11]. If the frequency domain specification is symmetric in amplitude and anti-symmetric in phase about the centre frequency (Hermitian specifications) then, these specifications may be Inverse Discrete Fourier Transformed (IDFT) to give real equally-spaced samples. Since, most SAW filters to date have been designed on the basis of equally-spaced fingers, then this type of response is quite easily obtained using either the classical transform or the IDFT.
The above technique does not give a truly arbitrary phase response; it is restricted to an anti-symmetric phase response.

1.5 **Arbitrary Amplitude**

Split electrodes have come to be used in general so as to eliminate ripple due to reflections which add constructively at certain frequencies [1]. Since the spatial sampling with split electrodes is twice that of single electrodes, it was noted in [12] that the amplitude response could be made non-symmetric by giving different lengths to each half of the split electrode. In effect, the synchronous frequency (frequency at which finger spacing resonates) is at twice the centre frequency \( f_0 \) of the desired response. There is also an image at the frequency \( f = 3 f_0 \). So long as the desired response and the image are constructed so as to be symmetric about the frequency \( f = 2 f_0 \), then the filter may easily be constructed using the previously discussed design techniques. Note that this response is also Hermitian (i.e. symmetric amplitude and anti-symmetric phase about the frequency \( f = 2 f_0 \)).

1.6 **Building Block**

The next obvious development to realize a given response was to design a bank of filters that would be driven and detected in parallel. Each filter would be a piece-wise fit to the desired response. These concepts (minus experimental results) are given in [13]. From these concepts, evolved the idea of combining the building blocks into a single IDT (interdigital transducer) pattern, which was shown in [14].
This technique has the advantage of being simple and effective in producing responses to specifications that are non-symmetric in amplitude and non-linear in phase; however, it is still a piece-wise fit.

1.7 Arbitrary Phase and Amplitude

The most elegant solution to the arbitrary Phase/Amplitude design problem is given in [15]. Essentially, this technique combines the concepts described in Sections 1.4 and 1.5. By using the split electrode concept as described before, a Hermitian function is constructed about \( f = 2 f_0 \) where \( f_0 \) is the centre frequency of the desired bandpass.

The desired amplitude response must be imaged at \( f = 3 f_0 \) such that a symmetric amplitude function is obtained about \( f = 2 f_0 \). Similarly, the desired phase response must be anti-symmetrically imaged at \( f = 3 f_0 \) so that the resulting specifications give a real equally spaced solution when the inverse Fourier transform is obtained. The other interdigital transducer (IDT) is designed to select only the desired bandpass and to reject the image bandpass. Of course the other IDT may be designed as a single bandpass linear phase filter or it may have a dual bandpass specification so long as the higher frequency image is at some frequency other than the upper image of the first IDT.

These concepts are illustrated in Figure 1.5 and will be further developed in Chapter 7 where these concepts are refined.
Specifications are Hermitian about $f=2f_o$
if for any $f_B$
the specification at $f=2f_o-f_B$ is $A+jB$ then,
the specification at $f=2f_o+f_B$ is $A-jB$

Figure 1.5: Dual bandpass Hermitian specifications about $f=2f_o$
1.8 Scope of Thesis

Since the development of surface acoustic wave (SAW) devices for filter applications, there has been a need to have a design technique that allowed for arbitrary amplitude and phase specifications. The easiest, and thus the first SAW filters obtained, had linear phase and symmetric amplitude responses; however, many systems require a nonlinear phase and a non-symmetric amplitude response. The exceptions to the rule of linear phase/symmetric amplitude responses for SAW's have been the "chirp" filters used in pulse compression for RADAR systems and some other concepts that have been explained in this chapter. The responses of these filters were generally limited to linear phase/non-symmetric amplitude or anti-symmetric phase/symmetric amplitude. The only publication known to this author that describes a design algorithm for arbitrary phase/amplitude specifications is given by Reference [16]. This algorithm required a "cut-and-try" optimization loop since no modelling of the device was included in the algorithm. It must be emphasized that no synthesis procedure presently available will give a SAW response as predicted by the synthesis. Thus the algorithm as described in this thesis involves: (a) synthesis (b) modelling of the synthesized IDT geometry (interdigital transducer) (c) obtaining the frequency response of the IDT based on the model (d) comparing the response to the desired specification and thus creating new pre-distorted specifications to be used in a new synthesis. Thus, the geometry of the transducer is obtained in an iterative multipass optimization loop. This procedure is shown in Figure 1.6 illustrating
A. Freq. Domain Specificat.
B. Generate equally spaced
   Samples for F.F.T.

C. Fast Fourier Transform
to Time Domain Samples
D. Sort + Window Samples

E. Dirac-Delta Model of
   IDT
F. Charge distribution Model

G. Far Field Frequency
   Response to SAW from
   Charge Model

Is this Solution
acceptable?

H. Error function to Pre-
distort Equally Spaced
Freq. Domain Specifications

I. Cut Mask and Build
   Device

Figure 1.6: Algorithm Block Diagram
the design procedure.

1.9 Description of Algorithm

The algorithm described in this thesis begins with the specification in the frequency domain of the insertion loss and the phase characteristics desired for the filter (Step A of Figure 1.6). From these criteria, a set of frequency domain equally-spaced samples is generated (Step B) which is stored for future reference. These complex numbers are then fed into a Fast Fourier Transform (FFT) routine (Step C). The output of the transform is then sorted and a Kaiser window function is applied (Step D). The next step gives the Dirac-delta-model ([1], [2], [3], [4]) of the IDT by selecting the position and length of each finger of the IDT in the time domain (Step E) which is stored on the system mass storage. From this model, the physical charge distribution model is obtained for the transducer (Step F) by solving the equations for the electrostatic solution for the metallized transducer pattern. This is done in a numerical iterative manner. Once the physical model has been obtained, then the far-field frequency response is obtained (Step G) which is stored on the system. Plots of the specifications, the ipodization (pattern showing finger lengths and positions) function and the responses are produced along with printout of the associated numerical values. If the response is not acceptable then the original specifications to the FFT are distorted in such a way such that the error is minimized (Step H) and the process is repeated from Step C. This process is repeated until an acceptable solution is
obtained at which point the mask for the IDT geometry may be made (Step I). From this point on, standard lithographic techniques are used to produce the SAW filter.

A description of the algorithm's significant elements is given including the following topics:

Chapter 2: the synthesis of the transducer geometry using a Fast Fourier transform technique

Chapter 3: development of the charge distribution model for the IDT that was synthesized in Chapter 2. The mathematical development is given in Appendix I.

Chapter 4: development of the far-field frequency response analogous to an end fire antenna array.

Chapter 5: development of the numerical optimization techniques that are used to minimize the design errors.

Chapter 6: notes on programming of the algorithm.

Chapter 7: experimental results of two attempts at television intermediate frequency (TVIF) filter specifications. The first filter has a single bandpass non-symmetric, nonlinear phase IDT in the device. The second filter uses a dual band Hermitian specification centred at 55 MHz for the IDT. Both filters have the same single bandpass rectangular linear phase receiving IDT.

Chapter 8: conclusions and some suggestions of how the algorithm might further be developed and refined.
CHAPTER 2
SYNTHESIS OF THE INTERDIGITAL TRANSDUCER

2.1 The Fast Fourier Transform

An efficient method for obtaining the time domain transformation from the frequency domain specifications is the Fast Fourier Transform (FFT) algorithm. This procedure is adequately described in many references such as [9] and [17]. The actual transform routine may be written by the filter designer or, if available, a computer system scientific subroutine package may be used.

Typically, a FFT works with lowpass specifications, and then the frequency-shifting theorem is applied to produce the desired bandpass response. However, the lowpass specification restricts the design to a symmetric amplitude and anti-symmetric phase response as is seen in Figure 2.1. To overcome this limitation, it is necessary to consider the desired bandpass and sample both below and above the centre frequency. To increase the sample density in the region of interest, it is necessary to frequency translate the desired bandpass such that only the adjacent stop bands will be considered in the transformation. This is illustrated in Figure 2.2. After the transformation, the sample rate of the time domain solution may be increased so as to bring the passband up to the desired frequency. However, for example, if the region from \( f = 1/2 f_o \) to \( f = 1 \ 1/2 f_o \) is transformed, then the mathematics imply image bandpasses at \( f = 0.0 \) Hz, \( 2 f_o \) etc. (where \( f_o \) is the centre
Figure 2.1: Lowpass FFT response when resonated at bandpass
Figure 2.2: Bandpass FFT response when frequency translated
frequency of the bandpass). In reality, the physical nature of the surface acoustic wave interdigital transducer (SAW IDT) prevents this from happening at the lower frequencies while the upper harmonics may be suppressed, if necessary, by simple RC external filtering techniques. This is usually not necessary as the amplifiers will be chosen such that they will pass only up to, and including, the frequencies of interest.

Thus, it is feasible to restrict the region of the frequency domain to be transformed and therefore reduce the required number of samples. This is necessary if the computation is to be kept to a reasonable amount of time and storage.

It should be noted at this time that the length of the IDT is determined by: the number of samples in the transformation and the band of the frequency domain. From [17] it is stated that:

\[ N \Delta t = T \]
\[ \Delta f = T^{-1} \]
\[ N \Delta f = \text{RLBW} \times \text{FO} \]

where: \( N \) is the number of samples in both the frequency and time domains
\( \Delta t \) is the time domain spacing
\( \Delta f \) is the frequency domain spacing
\( T \) is the period of the time domain being considered thus the IDT's length
RLBW is the relative frequency domain considered
FO is the centre of the frequency domain that is transformed.
It is easily seen that: \( T = N/(RLBW \times FO) \). By choosing RLBW to include only the passband and adjacent stopbands of interest, the length of the IDT corresponding to \( T \) may be chosen by selecting the appropriate value of \( N \). The length of the surface of the substrate covered by the IDT will be given by: \( L = v \cdot T \)

where: \( L \) is the length of the IDT
\( v \) is the surface acoustic wave velocity corrected for metallization
\( T \) is the period in time covered by the IDT.

2.2 **Input of Fast Fourier Transform**

In the programme written to illustrate this algorithm, the input DATA supplied by the designer is as presented below:

(a) IG is the power of 2 that specifies the number of sample points in the frequency and time domains which will be transformed. Typically \( IG = 6, 7, 8 \) such that there are \( N = 32, 64, 128 \) points.

(b) NP is the number of points in the frequency domain specified in terms of dB of insertion loss and degrees of phase deviation from linear phase.

(c) RLBW is the per unit bandwidth to be transformed. Typically this is from 0.5 FO to 2.0 FO.

(d) FO is the centre frequency of the region to be transformed and is usually chosen in the approximate centre of the desired bandpass. Units are MHz.
(e) next follow the NP specifications which are given as:
   (i) the frequency in MHz
   (ii) the insertion loss in dB
   (iii) the phase deviation in degrees.

The specifications are entered from the lowest frequency through to the highest frequency. All specified frequencies must be contained in the region to be transformed and a linear interpolation between points is assumed.

2.3 **Conditioning the DATA**

Having obtained the raw DATA, the frequency domain samples must be generated in the form required by the FFT process. The region to be transformed is divided into \( N = 2^{10} \) frequencies between \((1 - 1/2 \text{ RLBW})FO\) and \((1 + 1/2 \text{ RLBW})FO\). The points below the first specified frequency are given the same specifications of insertion loss and phase as the lowest supplied DATA point and the points above the highest specified frequency are given the highest frequency point's specifications. All of the points in between are given specifications obtained from a linear interpolation between the adjacent DATA points. Because of the discrete equal spacing of the specifications, it should now be obvious why the transformed region is minimized, otherwise an excessively large number of points would be required so as to accurately represent the desired continuous response by the sampled points.

Next, the specifications (which are stored on the system mass storage for future comparison) are transformed to the rectangular linear
co-ordinate system as a set of $N$ complex numbers. Since FORTRAN IV is the system language, the complex numbers are legitimate input arguments.

2.4 **Interpreting the FFT Output**

After applying the FFT routine, the resulting set of numbers (usually complex) must be sorted because of the "bit-reversed" order of the output as is shown in [17]. These time domain samples are then transformed into the linear-polar coordinate system with the amplitude vector allowed to have both positive and negative values so as to eliminate phase discontinuities at the nulls of the amplitude response. At these points, the phase changes typically by $180^\circ$ if the amplitude is restricted to positive values.

The phase of each one of the samples at this step in the process will be restricted to the range of $-90^\circ$ to $+90^\circ$. Therefore it is necessary to remove the discontinuous steps between samples. The result of this process is a continuously increasing phase response (with linear interpolation between sample points) from the first through the last sample point. The centre sample ($N/2$) is taken as $t = 0$ and is taken as the reference for the centre of the IDT to be developed from these samples.

It should be noted that since the bandpass being transformed is not at baseband, then the transformation will have a linear phase component as well as the phase deviation from linear phase which gives rise to the discontinuities mentioned above. This is analogous to the phase response of an intermediate frequency in communications theory and
is illustrated in Figure 2.3.

2.5 **Dirac-Delta Model**

Having conditioned the time domain response as described above it is now possible to synthesize the IDT. In principle, the fingers of the transducer are placed at any point in the phase response that has a phase of 0°, ±180°, ±360°, ±540°,... Those phases that are even multiples of 180° will have a positive pad connection and those with odd multiples of 180° will have a negative pad connection. The length of the finger overlap will be scaled according to the amplitude response at the appropriate phase position. The amplitude response is linearly interpolated between the evenly spaced samples of the Fast Fourier Transform solution.

It should be noted also that the negative amplitudes will be interpreted as a change of the rail connection that was obtained from the phase criterion. While this process seems more cumbersome than restricting the amplitude response to positive values and working with the discontinuous phase response, it actually reduces the number of samples that would be required to accurately establish finger lengths and positions in the region of the discontinuities.

The above procedure is shown schematically in Figure 2.4.

2.6 **Limitations of Dirac-Delta Model**

At this point the raw DATA of finger length, pad connection (polarity of finger) and the position in the time domain has been
Figure 2.3: Analogy to IF phase response
Figure 2.4: Effects of discontinuities in phase response of transform
obtained. In the past, most designers have simply translated this DATA to the surface spatial domain of the acoustic medium producing devices on this basis. Since that technique has produced experimental results with up to 15% deviation from the theoretical predictions, the next step in refining the design process is to use a more accurate analysis technique and from this, hopefully, to be able to design another transducer with updated input specifications such that the results of the next synthesis will be closer to the desired specifications. In the next chapter, a more accurate model will be described.
CHAPTER 3

CHARGE DISTRIBUTION MODEL OF AN INTERDIGITAL TRANSDUCER

3.1 Charge to Acoustic Deformation

It has been suggested [2] that the acoustic wave generated by a finger of an interdigital transducer (IDT) is related to the charge distribution on that finger by the following relationship:

\[ a_s = \frac{k_s}{2\sqrt{\omega \epsilon_r \lambda}} \int \left[ j \omega q(x_o) e^{\frac{-jk_x(x-x_o)}{\lambda}} \right] dx_o \]  \hspace{1cm} (3.1)

where: \( a_s \) is the acoustic wave deformation at \( x \)

\[ k_s = \frac{2\Delta V}{V_o} \] is the surface wave coupling constant

\( \omega \) is the frequency of excitation in radians/sec.

\( \epsilon_r \) is the effective dielectric constant

\( \lambda \) is the length of the finger

\( q(x_o) \) is the charge on the finger as a function of \( x_o \)

\( k_x \) is the SAW propagation constant in the \( x \) direction

\( x_o \) is defined as that region of \(-b \leq x \leq b\) on which the finger is deposited.

These quantities are depicted in Figure 3.1. The quantities on the right hand side of (3.1) are all known except for the charge distribution function \( q(x_o) \) and it is this quantity that needs to be
Surface Acoustic Wave Generation due to Distribution of Surface Charge.

Charge Distribution on Finger

Figure 3.1: Acoustic wave generation of IDT
established.

Before solving for \( q(x_o) \), it is necessary to point out some assumptions that are made with respect to SAW IDT's. Firstly, as has been stated in the introduction, the apodization (variable finger length) is taken to be equivalent to varying the applied voltage (and thus the charge since adjacent fingers form the plates of a capacitor) to each finger of a constant overlap IDT (i.e. all fingers have the same length). This approximation is quite accurate for longer (greater than 10 \( \lambda_s \))\(^{[18]} \) where \( \lambda_s \) is the SAW length and finger centres have approximately a \( \lambda_s/2 \) separation) fingers near the centre of the transducer. Those fingers that are shorter or near the ends of the array require correction factors to account for mechanisms such as diffraction. Secondly, so as to make the solution tractable, the electrostatic problem is taken as a two dimensional situation (\( y = q(x_o) \) as a function of \( x \)). That is, the charge distribution is assumed to be constant over the length of the finger to a first approximation. The end effects of the shorter overlaps will be compensated for empirically, by scaling at a later point.

3.2 Electrostatic Potential

The electrostatic potential, at some point \( x \) on the surface of the substrate, due to the charge \( q(x_o) \) of a finger centred at \( x = x_o = 0 \) is given to a first approximation by [18] as:

\[
\phi(x) = -\frac{1}{\pi \varepsilon_f} \int_A q(x_o) \ln |x-x_o| \, dx_o
\]
The X-Y plane is taken to be through the centre of the apodization of the IDT and it is assumed that the fingers are apparently of infinite length.

For any point \( x_p \) on the x-axis of the substrate, the voltage potential will be given by the summation of the electrostatic potentials of all the potentials for all the fingers in that IDT:

\[
V(x_p) = -\frac{1}{\pi e} \sum_{k=1}^{L} q_k(x_0) \ln |x_k - x_0| dx_0
\]

(3.3)

where the \( k \) subscript denotes the particular finger of the array whose charge effect is being evaluated. Also, \( x_k \) is the distance from the point \( x_p \) on the substrate to the centre of the \( k \)th finger, thus the origin of \( x_k \) is \( x = x_0 = 0 \) for each finger and is redefined for each finger.

3.3 Charge on a Finger

The charge distribution on a single metal strip has been shown to be given by [19] as:

\[
q(x_o) = \frac{C_o}{\sqrt{b^2 - x_o^2}}
\]

(3.4)

and it has been suggested [18] that the charge distribution of a finger in an IDT array can be modelled by:

\[
q(x_o) = \frac{C_0 + C_1 x_o + C_2 x_o^2}{\sqrt{b^2 - x_o^2}}
\]

(3.5)
where: \( C_0 \), \( C_1 \) and \( C_2 \) are constant coefficients that are a function of the finger's position in the transducer and \( b \) is 1/2 the finger width.

The linear term deviation, \( C_1 x_0 \) in (3.5), from the isolated strip distribution of equation (3.4) is caused by the fact that a finger nearer one end of an IDT than the other will have a skewed or non-symmetric charge distribution. The quadratic term corresponding to \( C_2 x_0^2 \) is due to the proximity of other fingers on either side of the finger under consideration which tends to attract more charge to the edges than would be the case with an isolated strip. Thus, the problem has now been reduced to finding the coefficients, \( C_0 \), \( C_1 \) and \( C_2 \), for each finger of an IDT.

3.4 Finding the Charge Distribution Coefficients

Since all the fingers of the IDT are connected in parallel to either the positive or negative rails, it is known that the voltage boundary condition will be \( \pm V/2 \) at any point on the fingers. By selecting the outer edges and the centre points of each finger, then 3 equations may be written for each finger at which the electrostatic potential may be calculated. The electrostatic potential will be a function of the charge distribution on all of the fingers in the IDT array and these charges may be arranged such that the potential due to the charges will equal the boundary value of the applied voltage (again, it is pointed out that the voltage on each finger is taken as the relative length of that individual finger multiplied by the applied voltage and the finger is assumed to have a length equal to the longest
finger overlap). Thus for each finger, there will be the following 3 equations:

\[ V_{ik} = \frac{1}{\pi \varepsilon_f} \sum_{l=1}^{L} C_{0l} \int_{-b_L}^{b_L} \ln \left| x_L - x^*_o \right| \frac{dx}{\sqrt{b_L^2 - x^*_o^2}} \]

\[ + C_{1k} \int_{-b_L}^{b_L} x_o \ln \left| x_L - x^*_o \right| \frac{dx}{\sqrt{b_L^2 - x^*_o^2}} \]

\[ + C_{2k} \int_{-b_L}^{b_L} x_o^2 \ln \left| x_L - x^*_o \right| \frac{dx}{\sqrt{b_L^2 - x^*_o^2}} \]

(3.6a)

where: \( i = 1, 2, 3 \)

\( k = 1, 2, 3, \ldots, L \)

putting:

\[ \phi_{0k} = \frac{1}{\pi \varepsilon_f} \int_{-b_L}^{b_L} \ln \left| x_L - x^*_o \right| \frac{dx}{\sqrt{b_L^2 - x^*_o^2}} \]

\[ \phi_{1k} = \frac{1}{\pi \varepsilon_f} \int_{-b_L}^{b_L} x_o \ln \left| x_L - x^*_o \right| \frac{dx}{\sqrt{b_L^2 - x^*_o^2}} \]

\[ \phi_{2k} = \frac{1}{\pi \varepsilon_f} \int_{-b_L}^{b_L} x_o^2 \ln \left| x_L - x^*_o \right| \frac{dx}{\sqrt{b_L^2 - x^*_o^2}} \]

(3.6b)
Then a set of equations may be written:

\[
\Phi_0 \cdot \begin{bmatrix} C_{01} \\ C_{02} \\ \vdots \\ C_{0L} \end{bmatrix} + \Phi_1 \cdot \begin{bmatrix} C_{11} \\ C_{12} \\ \vdots \\ C_{1L} \end{bmatrix} + \Phi_2 \cdot \begin{bmatrix} C_{21} \\ C_{22} \\ \vdots \\ C_{2L} \end{bmatrix} = \begin{bmatrix} V_1 \\ V_2 \\ \vdots \\ V_L \end{bmatrix}
\]  

(3.7)

where: \( \Phi_0, \Phi_1 \) and \( \Phi_2 \) are \( L \times L \) matrices and \( i = 1, 2, 3 \) corresponds to the 3 points on each finger i.e. the two edges and the centre where the potentials are calculated and then equated to the boundary values.

The authors of [8] wrote this set of equations as:

\[
\begin{bmatrix} C_{01} \\ C_{11} \\ C_{21} \\ \vdots \\ C_{0L} \\ C_{1L} \\ C_{2L} \end{bmatrix} \cdot \Phi = \begin{bmatrix} V_{11} \\ V_{21} \\ V_{31} \\ \vdots \\ V_{1L} \\ V_{2L} \\ V_{3L} \end{bmatrix}
\]  

(3.8)

where \( \Phi \) is a \( 3L \times 3L \) matrix.

Then, they inverted \( \Phi \) to obtain the coefficients of the charge distribution matrix. It is obvious that this technique is restricted to relatively small problems (they illustrated the process with a 13 finger transducer), thus large realistic IDT's must be solved in a different manner.

In (3.8) it is noted that \( V_{1i} = V_{2i} = V_{3i} \) for each finger, thus the set of equations may be arranged as in (3.7) and the left hand side may be evaluated for initial guesses of \( C_{0i}, C_{1i} \) and \( C_{2i} \) as:
\[
\begin{bmatrix}
C_{01} \\
C_{02} \\
\vdots \\
C_{0L}
\end{bmatrix} + \begin{bmatrix}
C_{11} \\
C_{12} \\
\vdots \\
C_{1L}
\end{bmatrix} + \begin{bmatrix}
C_{21} \\
C_{22} \\
\vdots \\
C_{2L}
\end{bmatrix} = \begin{bmatrix}
P_{11} \\
P_{21} \\
\vdots \\
P_{L1}
\end{bmatrix}
\] (3.9)

where \(i = 1, 2, 3\) which corresponds to the 3 points on each finger under consideration.

By comparing the calculated \(P_{\ell1}\) to the imposed boundary value \(V_{\ell}\), it is possible to relax the coefficients \(C_{0\ell}\), \(C_{1\ell}\) and \(C_{2\ell}\) in such a way that \(P_{\ell1}\), \(P_{\ell2}\) and \(P_{\ell3}\) are made to be equal to \(V_{\ell}\) within an acceptable error criterion. Once this has been accomplished, then applying the optimum values of \(C_{0\ell}\), \(C_{1\ell}\) and \(C_{2\ell}\) to (3.5) will give the charge distribution for each finger of the transducer.

The initial guess for \(C_{0\ell}\) is taken proportional to the corresponding fingers amplitude from the FFT solution while \(C_{1\ell}\) and \(C_{2\ell}\) are set to zero. Then the potentials \(P_{\ell1}\), \(P_{\ell2}\) and \(P_{\ell3}\) are calculated and the relaxation process is begun. Referring to Figure 3.2, the relaxation of \(C_{0\ell}\) is taken as:

\[
C_{0\ell}^{t+1} = C_{0\ell}^t [1 + B_0 (V_{\ell} - P_{\ell2}^t)]
\] (3.10)

where \(B_0\) is a constant that includes scaling and an over/under relaxation constant. The \(t+1\) subscript implies a newer value than the \(t\) subscript.

The initial value of \(C_{1\ell}\) is taken as:
The values of $C_{o1}$, $C_{11}$ and $C_{21}$ are relaxed in such a way so as to make the potentials $P_{11}$, $P_{12}$ and $P_{13}$, calculated from the charge distribution, approach the boundary value $V_1$.

Figure 3.2: Relaxation of charge distribution coefficients
\[ C_{1k} = \frac{P_{k3} - P_{k1}}{k_1} \]  

(3.11)

where \( k_1 \) is a relational constant. In the following passes \( C_{1k} \) is relaxed as:

\[ C_{1k} = C_{1k} \left[ 1 + B_1 \left( P_{k3} - P_{k1} \right) \right] \quad \text{at} \quad t+1 \quad \text{at} \quad t \]  

(3.12)

where \( B_1 \) is a constant including the under-relaxation constant.

The initial value of \( C_{2k} \) is taken as:

\[ \frac{V_k - 1/2 \left( P_{k3} + P_{k1} \right)}{K_2} \]  

(3.13)

where \( K_2 \) is the equation's relational constant. Thereafter \( C_{2k} \) is relaxed as:

\[ C_{2k} = C_{2k} \left[ 1 + B_2 \left( V_k - 1/2 \left( P_{k3} + P_{k1} \right) \right) \right] \quad \text{at} \quad t+1 \quad \text{at} \quad t \]  

(3.14)

where \( B_2 \) is the constant as before. No proof is offered in this thesis as to why the numerical technique causes the potentials to relax into their boundary values. The routine has shown itself to cause \( P_{k1} \), \( P_{k2} \) and \( P_{k3} \) to tend toward \( V_k \) for successive passes and produces deviations from \( V_k \) of less than 1% in 3 passes for those cases that have been investigated.

The only quantities to be established now are the integral functions \( 0_{11} \), \( 1_{11} \) and \( 2_{11} \). While the solutions to these functions must have been obtained by the authors of [18], neither the development nor the functions themselves have been presented in the literature.
3.5 The Integral Functions

Considering the equations (3.6b) and putting:

\[
I_{10} = - \int_{-b}^{b} \frac{\ln |x-x_0|}{\sqrt{b^2 - x_0^2}} \, dx_0
\]

\[
I_{20} = - \int_{-b}^{b} \frac{x_0 \ln |x-x_0|}{\sqrt{b^2 - x_0^2}} \, dx_0
\]

\[
I_{30} = - \int_{-b}^{b} \frac{x_0^2 \ln |x-x_0|}{\sqrt{b^2 - x_0^2}} \, dx_0
\]

where the subscript \( i \) has been dropped. Then it may be shown as is given in Appendix I that the solution to these integrals is:

\[
I_{10} = \pi \ln (2) \quad \text{for} \ x = 0
\]

\[
I_{10} = \pi \ln \frac{2}{1 + \sqrt{1 - (b/x)^2}} \quad \text{for} \ |x| > +b
\]

\[
I_{20} = 0 \quad \text{for} \ -b \leq x \leq +b
\]

\[
I_{20} = -\pi (\sqrt{1 - (b/x)^2} - 1) \quad \text{for} \ x > +b
\]

\[
I_{20} = +\pi (\sqrt{1 - (b/x)^2} - 1) \quad \text{for} \ x < -b
\]

\[
I_{30} = \frac{4}{\pi} \ln (z) - \frac{1}{2} \quad \text{for} \ x = 0
\]

\[
I_{30} = \frac{1}{2} \left( \frac{b}{x} \right)^2 \ln \left[ \frac{2}{1 + \sqrt{1 - (b/x)^2}} \right] + \sqrt{1 - (b/x)^2} - 1 - \frac{1}{2} \left( \frac{b}{x} \right)^2 \quad \text{for} \ |x| \geq b
\]

(3.17)
These functions are shown graphically in Figure 3.3. The results have been normalized such that the arguments of the logarithms are dimensionless and such that the integrals disappear as $x \to \infty$. This at first may seem strange since $x$ is a variable; however, it is noted that the resulting variable is $b/x$ and this is a variable of $x$.

3.6 Summary

Thus, by applying the results of the integral equations (3.17) to (3.3b) for the appropriate conditions of $x_L$, the potentials at each of the 3 points on each finger may be evaluated using (3.9) with the values of $C_{01}$, $C_{11}$, and $C_{21}$ being found as described in Section 3.4. Once this has been done, then an acceptable solution to the charge distribution on the IDT has been found (through (3.5)) and the results may now be applied to (3.1) so as to calculate the acoustic far field radiation and thus obtain the frequency response of the IDT. This will be described in the next chapter.
Figure 3.3: Potential functions due to charge distribution
CHAPTER 4
FREQUENCY RESPONSE OF AN INTERDIGITAL TRANSDUCER

4.1 General

Recalling equation (3.1)

\[ a_s = \frac{k_s}{2 \sqrt{\omega \epsilon \mu / L}} \int_{A} j\omega q(x_o) e^{-jk_s(x-x_o)} \, dx_o \]  \hspace{1cm} (4.1)

where the above quantities have been previously defined, it is now possible to calculate the acoustic deformation at some point \( x_p \). By using the superposition of the waves generated by each finger in the transducer, and applying the appropriate solutions to \( q(x_o) \) as established in Chapter 3. Then at some point \( x_p \) the response will be given by:

\[ H(\omega) = \sum_{i=1}^{L} \frac{a_s l_i}{s} + b \]

\[ = \frac{j\omega k_s}{2 \sqrt{\omega \epsilon \mu / L}} \sum_{i=1}^{L} \epsilon e^{-jk_s x_i} \cdot q_L(x_o) e^{jk_s x_o} \, dx_o \]

\[ = \frac{L}{k} \sum_{i=1}^{L} e^{-jk_s x_i} \cdot C_{ot} \cdot \frac{1}{\sqrt{v b_k - x_o^2}} \, dx_o \]

42
where all of the constants have been grouped into $k$. Thus the 3 integrals of (4.2) need to be evaluated.

4.2 Integral Solutions

From equation (4.2) the integrals that need to be evaluated are:

$$I_{10} = \int_{-b}^{b} \frac{e^{jkx_o}}{\sqrt{b^2 - x_o^2}} \, dx_o$$

$$I_{20} = \int_{-b}^{b} \frac{x_o e^{jkx_o}}{\sqrt{b^2 - x_o^2}} \, dx_o$$

$$I_{30} = \int_{-b}^{b} \frac{x_o^2 e^{jkx_o}}{\sqrt{b^2 - x_o^2}} \, dx_o$$

The solution to these integrals is given in Appendix II, and the results are:
\[ I_{10} = J_0 (k_x b) \]
\[ I_{20} = j b J_1 (k_x b) \]
\[ I_{30} = b^2 \left[ J_0 (k_x b) - (k_x b)^{-1} J_1 (k_x b) \right] \]  \hspace{1cm} (4.4)

where: \( J_0 \) and \( J_1 \) are Bessel functions of the 1st and 2nd kind

\( k_x \) is the propagation constant

\( b \) is 1/2 the finger width.

As before in Chapter 3, the normalization has been done with respect to \( x \).

4.3 The Acoustic Wave

In general, the frequency response of an IDT is evaluated at the centre of the transducer (which corresponds to \( t=0 \) and \( x=0 \)). However, since this algorithm deals with responses that are not the same in both directions, it is necessary to evaluate the response outside of the IDT in the direction in which the wave is desired to propagate. Substituting the solutions (4.4) into equation (4.1) gives:

\[
\begin{align*}
\alpha_s &= \frac{j k_x e^{-j k_x x}}{2 \sqrt{\omega \varepsilon_f L}} \left[ J_0 (k_x b), j b J_1 (k_x b), b^2 \left< J_0 (k_x b) - \frac{J_1 (k_x b)}{k_x b} \right> \right] \begin{bmatrix} C_0 \\ C_1 \\ C_2 \end{bmatrix} \\
&= \frac{j k_x \sqrt{\omega L} e^{-j k_x x}}{2 \sqrt{\varepsilon_f}} \left[ J_0 (k_x b), j b J_1 (k_x b), b^2 \left< J_0 (k_x b) - \frac{J_1 (k_x b)}{k_x b} \right> \right] \begin{bmatrix} C_0 \\ C_1 \\ C_2 \end{bmatrix} \\
&= \frac{j k_x \sqrt{\omega L} e^{-j k_x x}}{2 \sqrt{\varepsilon_f}} \left[ J_0 (k_x b), j b J_1 (k_x b), b^2 \left< J_0 (k_x b) - \frac{J_1 (k_x b)}{k_x b} \right> \right] \begin{bmatrix} C_0 \\ C_1 \\ C_2 \end{bmatrix} \hspace{1cm} (4.5)
\end{align*}
\]
Now by applying these results to (4.2) and doing the summation over the whole IDT under consideration, then the frequency domain response may be obtained.

4.4 Summary

Thus, the frequency response of the IDT will be given by

\[ H(\omega) = K \sum_{L=1}^{L} e^{-j k_x L} \left[ j \sqrt{\omega} c_{0L} j_0 \left( k_x b_L \right) \right. \]

\[ \left. - \sqrt{\omega} b_L c_1 j_1 \left( k_x b_L \right) + j \sqrt{\omega} c_{1L} b_L^2 j_0 \left( k_x b_L \right) - \frac{j_1 \left( k_x b_L \right)}{k_x b_L} \right] \]

(4.6)

Then \( H(\omega) \) is evaluated at each frequency \( \omega_i \) at which the specifications have been established. By comparing the specifications to the responses at these specific frequencies, an iterative design technique may be developed so as to make the response approach the specifications. This iterative scheme is described in Chapter 5.
CHAPTER 5
NUMERICAL OPTIMIZATION

5.1 General

The synthesis procedures available to designers at this time all involve delta functions in the time domain from which the finger geometry is established. Since delta functions imply filamentary line sources (or taps in the receiving mode), this delta model does not adequately describe the physical response of the interdigital transducers (IDT). Also, applying a window function causes distortion of the desired bandpass by broadening the bandpass and rounding the edges. It is mainly because of these reasons that an iterative design technique is required if the desired response is to be met more accurately.

5.2 Error Function

Referring to Figure 5.1 an error function is calculated by substracting the calculated response of the synthesised IDT from the original specifications. The amplitudes are left on the logarithmic scale so that the error function is not "hard limited" at zero amplitude. In Figure 5.2 it is seen that if the error function (times its over/under relaxation constant) is greater than the original specification and is added linearly to the specifications then negative amplitude specifications result. These would have to be interpreted as phase reversals of $180^\circ$ to the specifications which would cause...
Figure 5.1: Development of error function and predistortion of specifications
Linear System: Predistortion is done on linear scale

1. \[
\begin{align*}
\text{Imag} & \quad \text{SPEC}_1 \\
\text{Real} & \quad \text{ERR}_1
\end{align*}
\]

2. \[
\begin{align*}
\text{Imag} & \quad \text{SPEC}_2 \\
\text{Real} & \quad \text{ERR}_2
\end{align*}
\]

3. \[
\begin{align*}
\text{Imag} & \quad \text{SPEC}_3 \\
\text{Real} & \quad \text{ERR}_3
\end{align*}
\]

   \text{(no limiting)}

4. \[
\begin{align*}
\text{Imag} & \quad \text{SPEC}_4 \\
\text{Real} & \quad \text{SPEC}_3
\end{align*}
\]

   \text{(with limiting)}

   \text{(w. limit.)}

Logarithmic System: Predistortion is done on logarithmic scale

1. \[
\begin{align*}
\text{Imag} & \quad \text{SPEC}_1 \\
\text{Real} & \quad \text{ERR}_1
\end{align*}
\]

2. \[
\begin{align*}
\text{Imag} & \quad \text{SPEC}_2 \\
\text{Real} & \quad \text{ERR}_2
\end{align*}
\]

3. \[
\begin{align*}
\text{Imag} & \quad \text{SPEC}_3 \\
\text{Real} & \quad \text{ERR}_3
\end{align*}
\]

4. \[
\begin{align*}
\text{Imag} & \quad \text{SPEC}_4 \\
\text{Real} & \quad \text{SPEC}_4
\end{align*}
\]

   \text{(no limit.)}

   \text{(w. limit.)}

Figure 5.2: Hard limiting and soft limiting of error function
oscillations in the numerical process. By creating the error function on the logarithmic scale these problems are eliminated. The phase error function creates no such problem and is thus left in whatever angular linear measure is used in the routine. Thus, the error functions are given as:

\[ RMERR(\omega_1) = MDB(\omega_1) - RMDB(\omega_1) \]
\[ RPERR(\omega_1) = PHASE(\omega_1) = RPHS(\omega_1) \]  \hspace{1cm} (5.1)

where:
- \( RMERR(\omega_1) \) is the amplitude error in dB
- \( MDB(\omega_1) \) is the desired response in dB
- \( RMDB(\omega_1) \) is the IDT response in dB
- \( RPERR(\omega_1) \) is the phase error in degrees
- \( PHASE(\omega_1) \) is the desired phase in degrees
- \( RPHS(\omega_1) \) is the phase response in degrees

The error function is calculated at each \( \omega_1 \) of the input set of specifications.

5.3 Predistortion of Specifications

Once the error functions have been calculated, then the error, multiplied by an under-relaxation factor, is added to the specifications that are fed into the Fast Fourier Transform (FFT) synthesis process as described earlier. An under-relaxation factor of 0.5 has proved to be the most efficient since higher values (up to 2.0) caused oscillations in the resulting numerical solutions. Thus, the specifications that are fed to the synthesis routine are:
\[
\text{MAG}(\omega_{t}) = \text{MAG}(\omega_{t}) + 0.50 \times \text{RMERR}(\omega_{t}) \\
\text{t+1} \quad \text{t} \quad \text{t}
\]

\[
\text{PHSE}(\omega_{t}) = \text{PHSE}(\omega_{t}) + 0.50 \times \text{RPERR}(\omega_{t}) \\
\text{t+1} \quad \text{t} \quad \text{t}
\] (5.2)

where: \(\text{MAG}(\omega_{t})\) is the variable amplitude input specifications to the FFT in dB

\(\text{PHSE}(\omega_{t})\) is the variable phase input specifications to the FFT in degrees.

The subscript \(\text{t+1}\) implies the set of specifications that follow the pass where the \(\text{t}\) specifications were used. This process of predistortion is shown in Figure 5.1.

5.4 Summary

The iterative design process has been presented and because of the complexity of the problem, a totally automatic design algorithm is not suggested. It has proved expedient to build the system as an interactive computer-aided-design (CAD) technique which allows the designer to view the response of each synthesis. From these plots and numerical results, the designer is able to decide if the results are acceptable or if the procedure should be restarted with different design constraints.

It should also be noted that there is no need to optimize the phase criterion for amplitudes of less than \(-20\) dB for example. Normally the phase response is only of interest in the bandpass and not in the stopband.
CHAPTER 6

PROGRAMMING THE ALGORITHM

6.1 General

To illustrate the design algorithm described in this thesis, a series of programmes have been written in FORTRAN IV for the CDC-6400 computing system at McMaster University, Hamilton, Canada. The programmes have been written for "batch" submission but could be used in an on-line interactive mode. Once developed to the point of having the design system on-line, then a further breaking up of the programmes so that routines such as analysis could be accessed at any time would be advantageous to the designer.

6.2 SAWINIT

Referring to the flow chart in Figure 6.1 the input DATA required by SAWINIT has been described in detail in Chapter 2. Recalling Section 2.2, block 1 of Figure 6.1 requires: the power of 2 (IG) that determines the number of samples (N) in the FFT; NP which is the number of specifications; the number of samples in the FFT; NP which is the number of specifications; the relative bandwidth of the transformation (RLBW); the centre frequency (FO) and then follow the NPP specifications of FRSF (NP, 3).

Next in block 2 of Figure 6.1, the frequency domain from \( f = (1 - \text{RLBW}) \times \text{FO} \) to \( f = (1 + \text{RLBW}) \times \text{FO} \) is divided into \( N = 2^{\text{IG}} \) frequency
Figure 6.1: Flowchart of programme SAWINIT
points coded as FRQ(I). At each of these frequencies a desired specification MDB(I) for amplitude and PHASE(I) for phase is generated as a linearly interpolated fit to the DATA of FRSP. Those frequency sample points below FRSP(1,1) are given the specifications of FRSP(1,1) and those above FRSP(NP,1) takes its specifications.

For the first pass the input specifications to the FFT (block 3) are given the values of MAG(I) = MDB(I) and PHSE(I) = PHASE(I) which are then transformed into linear rectangular complex numbers for the transformation.

The output of the Fast Fourier Transform (FFT), is then sorted (block 4) so that the samples are in their correct order and so that t = 0 sec. is in the centre of the time domain solution i.e. the time domain solution goes from - T/2 to + T/2 and thus has its maximum amplitude near the centre of the domain.

Next a Kaiser-Bessel window function of:

$$ WT(I) = \frac{J_0 \left[ \left( \frac{16}{N} \right) \left( \sqrt{NH^2 - \left| NH - I \right|^2} \right) \right]}{J_0 \left( \sqrt{8.0} \right)} $$

is applied to the time domain samples [20] where:

- $J_0$ is the zero order Bessel function
- N is the number of samples in the transform
- $NH = N/2$
- I is the sample from 1 to N

This is done to reduce the side lobes at the expense of broadening the bandpass and rounding the edges of the bandpass. From Step 4 the finite time domain solutions to the frequency domain specifications are plotted
in Step 10.

Having obtained the time domain solutions, the next step is to synthesize the IDT geometry (Step 5). After removing the phase discontinuities as explained in Chapter 3, the linear phase component due to the centre frequency is added to the phase response of the time domain solution as:

\[ \phi(I) = 360.0^\circ \times FO \times DT \times \frac{I}{L} \]

where: FO is the centre resonant frequency in MHz
DT is the sample separation in micro-sec.
I is the sample number (I = 1, 2, ..., N)

Having changed the time domain samples as above then the finger lengths, positions and polarities are evaluated as described in Chapter 2 giving:

L is the number of fingers in IDT
LEN(I) is length of finger and sign gives polarity
TIM(I) is the time position in transducer structure relative to its physical position.

The basic IDT information of L, LEN and TIM are stored on the system mass storage (Step 11) so that the actual IDT geometry may be generated by the programme TAPEJOB if the designer wishes.

The next step of the charge-distribution modelling for the synthesized IDT is performed in Step 6. This procedure is described in detail in Chapter 3. The subroutine MODEL also calculates the acoustic response of each finger of the IDT from the charge distribution coefficients as described in (4.3) but without the phase delay \[ e^{-jkxL} \].
This information is presented as complex numbers in the array AR(L) which is fed to the next routine FRQRSP (Step 7). FRQRSP calculates the actual frequency domain response of the transducer in the forward direction. This response (RMDB and RPHS) is stored on the system mass storage (Step 12) for use in SAWOPT. The final step of SAWINIT is to generate the input specifications and the response in both plotted and numerical form. Also, the time domain solution is plotted. From this output, the designer can decide if the input specifications are valid or if the response meets his requirements. If further refinement of the design is wished then the programme SAWOPT is run as is described in the next section. The listing of SAWINIT is given in Appendix III.

6.3 SAWOPT

The programme SAWOPT (Appendix IV) is the same as the programme SAWINIT except for Steps 1 and 2. In Step 1, the desired specifications MDB and PHASE are obtained along with the predistorted specifications, MAG and PHSE, from the last pass of either SAWOPT or SAWINIT which ever preceded this run. The responses of the last synthesis, RMDB and RPHS, are compared to MDB and PHASE and then MAG and PHSE are distorted further as is described in Chapter 5. Then the whole process as in SAWINIT is repeated. This is shown in the flowchart of Figure 6.2.

The designer uses SAWOPT in successive passes in an attempt to minimize the error functions, RMERR and RPERR (Chapter 5), until a usable design is obtained. When RMDB and RPHS are acceptably small, the
Figure 6.2: Flowchart of programme SAWOPT
designer uses the programme TAPEJOB to generate the transducer mask cutting instructions. TAPEJOB is described next.

6.4 TAPEJOB

The programme TAPEJOB takes the information of L, LEN and TIM (described previously) and generates cutting instructions for a computer controlled cutting table available at the University of Waterloo. Since this is a programme unique to this University (and Waterloo), details of TAPEJOB will not be given. The programme listing is given in Appendix V.

Also a plot is generated by TAPEJOB from the cutting instructions so as to verify that the instructions are correct before cutting the mask.

6.5 Summary

Initially the designer enters his desired specifications into SAWINIT. He then uses SAWOPT interactively until an acceptable transducer has been designed. Then TAPEJOB is used to produce the cutting instructions for the IDT's mask.
CHAPTER 7

EXPERIMENTAL RESULTS

7.1 General

The programmes described in Chapter 6 have been applied to the design of two television intermediate frequency (TVIF) filter designs. Two such filters have been produced which involve the design of three IDT's. The one IDT, arbitrarily called the receiving IDT, is designed as a symmetric linear phase transducer and is responsible for obtaining the stop band insertion loss of 50 dB or better. This transducer is used as the receiver in both filters. The transmitting IDT of the first filter is a single bandpass with the non-symmetric and non-linear specifications of the TVIF. The second filter has a dual bandpass transmitting specification such that the lower bandpass has the required TVIF specifications (Figure 7.1). The upper bandpass is the image of the lower bandpass such that the specifications are Hermitian about a centre frequency of 55 MHz. As a result this IDT can be fabricated with single finger electrodes since the synchronous frequency (frequency at which periodic finger spacing resonates) is placed in the stop band. This allows for a relaxation on the tolerances of fabrication. The receiving IDT accepts only the lower bandpass.

The receiving IDT and the first transmitting IDT have split electrodes while the second transmitting IDT has single fingers.
Figure 7.1: TVIF Filter Specifications. Ideal—Solid Line; Typical Lumped Element Compromise—Dashed Line.
7.2 Filter Specifications

The ideal (solid line) and typical compromise (dashed line) filter specifications [21] and [22] are given in Figure 7.1. Since the specifications for TVIF standards were produced at a time when only lumped element filters were feasible, the phase response was allowed to be non-linear so that television receivers would not require expensive phase equalization networks. The transmitted signal of commercial television is predistorted so that the product of the predistorted signal and the TVIF will give a linear phase video signal to the monitor. The phase specification is actually given as group delay:

\[ D = -\frac{d\phi}{d\omega} \]  \hspace{1cm} (7.1)

where:
- \( D \) is the group delay in seconds
- \( \phi \) is the phase in radians
- \( \omega \) is the angular frequency in radians/second

but for the purposes of this thesis it has been converted to a phase specification as a deviation from linear phase in degrees.

7.3 Single Bandpass Filter

The first filter designed and built has a single bandpass specification. The solution of the 20th pass of SAWOPT for the transmitting IDT and the 15th pass of SAWOPT for the receiving IDT were used to produce the filter mask shown in Figure 7.2.

Photographs of the response of this filter are given in Figure 7.3 and an X-Y plot of the 10 MHz about the centre frequency is shown in.
Figure 7.3: Response of single bandpass filter
Figure 7.4: X-Y Plot of Single Bandpass Filter Response
Figure 7.4. It is noted that there is an error of approximately 250 kHz in the bandpass (too high) that is due to an error in the final reduction ratio. Also the stop band at $f = 41.25$ and $f = 47.25$ are not met. A redesign of the receiving transducer would resolve this problem.

7.4 Dual Bandpass Filter

The second filter built has a dual bandpass Hermitian specification (Figure 7.5). The receiving transducer accepts only the lower bandpass and rejects the upper. The 4th pass of SAWOPT for the transmitting IDT and as before the 15th pass of SAWOPT for the receiver were used to produce the second filter mask of Figure 7.6. Photographs of the response of the second filter are given in Figure 7.7 and an X-Y plot of the 10 MHz about the lower bandpass centre frequency is shown in Figure 7.8. As before the response is approximately 250 kHz high in frequency and the traps at $f = 41.25$ MHz and $f = 47.25$ MHz are not met because of the receiving transducer's response being too wide.

7.5 Measuring the Responses

An H.P. 8407A network analyser system (Figure 7.9) was used to obtain the filter responses. Since there is no normalisation in the system, the phase measurement requires an equal length in the reference channel. The only feasible way to account for the acoustic delay in the filter is to use a SAW delay line. Therefore, a delay line was constructed using two receiving transducers placed at the appropriate phase centres (corresponds to $t = 0$ or $r = 0$) for the receiving and
Figure 7.5: Hermitian Specifications for Dual Bandpass IDT
Figure 7.7: Response of dual bandpass filter.
Figure 7.8: X-Y Plot of Dual Bandpass Filter Response
Figure 7.9: Measurement system for testing filters
transmitting IDT's in the master mask. The delay line mask is shown in Figure 7.10 and the response is shown in the X-Y plot of Figure 7.11 for the bandpass region. Another approximate delay line was used to show the linear phase in Figure 6.11. The amplitude response is twice the response of a single receiving IDT and it is obvious that this IDT needs to be redesigned if the traps of the TVIF responses are to be met. The ripple in the phase response is due to electromagnetic feed through in the package for the approximate delay line (a dual inline package which did not provide adequate shielding).

7.6 Summary

The experimental results given in this thesis are used as illustrations of how the algorithm works. Further design work (redesign of receiving IDT) would be required to obtain a valid design that would meet the trap requirements. The filters that have been designed, fabricated and tested indicate the validity of the design process and considering the lumped-element compromise filter (dashed lines in Figure 6.1), the obtained responses compare quite favourably. The receiving IDT theoretical response predicted the wide bandpass as was obtained.
Figure 7.11: X-Y Plot of Delay Line Filter Response
CHAPTER 8

CONCLUSIONS

8.1 General

A design algorithm has been produced comprised of the following elements:

(a) Synthesis
(b) charge distribution model analysis
(c) frequency domain response
(d) numerical optimization of design to realize a desired response.

The algorithm has been programmed and the model has been developed mathematically and then some difficult filter responses were attempted.

The designed filters were fabricated and tested showing the algorithm, the programmes and the model to all be quite valid.

Because of the broadness of this investigation, most of the areas that go to make up the algorithm could not be investigated in as much depth as would be desirable. The following sections suggest areas of further research, development and refinement.

8.2 Synthesis

In the synthesis, further work needs to be done to establish the final effect of the windowing function. Some windows such as Bartlett, Hamming and Hanning [20] were investigated in the course of this study.
but the actual quantitative results of the various functions and their variable arguments have not been researched with respect to this design algorithm. A series of SAW filters need to be fabricated and measured using the algorithm with various windows to see what effect they have on the resulting device with respect to design time, band edge rounding, side lobe suppression and stop band insertion loss.

8.3 Charge Distribution Model

The charge model developed in this study is the solution to the model suggested in [18]. It has been used based on the claims of the authors of [18], however, further research on this model is warranted, particularly with respect to accounting for variable finger lengths.

The numerical solution to the charge distribution coefficients could be made much more efficient since only the nearby fingers will have a significant effect on a finger's charge distribution.

Also, it would be desirable if the variable finger length could be accounted for in the model rather than using the approximation in Chapter 3.

8.4 Frequency Response

The frequency response of the model does not include consideration of finger edge reflections. Since split electrodes or synchronous frequencies in the stop band have been used, this did not present a problem in this study. It is necessary under many other conditions, however, to use synchronous frequencies in the bandpass and
since lithography constraints at higher frequencies preclude the use of split electrodes, accounting for finger reflections is necessary if passband ripple is to be avoided.

Also, in much more advanced studies, diffraction in the far field could be included as well as accounting for the scattering at apodization cuts (finger ends).

8.5 Optimization

This area offers the greatest potential for further research. Attempts at optimizing the IDT geometry with respect to the desired and realized frequency responses were attempted in this study. While some positive results were obtained using this approach, the optimization was not very efficient since the response at all frequencies changed if one finger was moved or changed in length. Direct manipulation of the fingers is desirable but has not proved to be very efficient in terms of computer time. Perhaps using the transform algorithm as described in this thesis until very close results are obtained and then optimizing the IDT directly would prove effective.

Also, the possibility of using gradient search techniques should be investigated since the apodization function is well behaved near the solution.

8.6 Summary

There are at least four areas of significant potential for further research that will refine the design algorithm presented in this
thesis. Since surface acoustic wave filters offer such a significant improvement over lumped element filters in the 10 MHz to 1 GHz region, in size, ruggedness and realizable designs it is essential that effective and efficient design packages be provided to the design engineer that are easy to use and provide usable results without in-depth knowledge of how SAW devices work. The study presented in this thesis is an initial step in providing the design engineer with that facility.
APPENDIX I

SOLUTION TO CHARGE-DISTRIBUTION INTEGRALS

I.1 General

The integral equations of Chapter 3 to be solved are:

\[ I_{10} = - \int_{-b}^{b} \frac{\ln |x-x_0|}{\sqrt{b^2 - x_0^2}} \, dx_0 \quad (I.1a) \]

\[ I_{20} = - \int_{-b}^{b} \frac{x_0 \ln |x-x_0|}{\sqrt{b^2 - x_0^2}} \, dx_0 \quad (I.1b) \]

\[ I_{30} = - \int_{-b}^{b} \frac{x_0^2 \ln |x-x_0|}{\sqrt{b^2 - x_0^2}} \, dx_0 \quad (I.1c) \]

The table of integrals used in obtaining these solutions is Reference [23].

I.2 Solution to \( I_{10} \)

\[ I_{10} = - \int_{-b}^{b} \frac{\ln |x-x_0|}{\sqrt{b^2 - x_0^2}} \, dx_0 \]

using the transformations:

\[ x_0 = b \cos \theta \]
\[ dx_0 = -b \sin \theta \, d\theta \]
\[ \sqrt{b^2 - x_0^2} = b \sin \theta \]  \hfill (I.2)

(i) \( x = 0 \); making the substitutions of (I.2)

\[ I_{10} = -\int_0^\pi \ln |-b \cos \theta| \, d\theta \]
\[ = -2 \int_0^{\pi/2} \ln (b \cos \theta) \, d\theta \]
\[ = -2 \left[ \frac{\pi}{2} \ln (b) - \frac{\pi}{2} \ln (2) \right] \]

Thus

\[ I_{10} = \pi \ln (2/b) \text{ for } x = 0 \]  \hfill (I.3)

(ii) \( x = b \); making the substitutions (I.2):

\[ I_{10} = -\int_0^\pi \ln |b - b \cos \theta| \, d\theta \]

For \( \theta = \pi \) the argument of \( \ln \) is \( b - b \cos \pi = 2b \) and for \( \theta = 0 \) the argument is \( >0 \) over region of integration which gives:

\[ I_{10} = -\int_0^\pi \ln (b - b \cos \theta) \, d\theta \]
\[ = -\pi \ln \left[ \frac{b - \sqrt{b^2 - b^2}}{2} \right] \]

Thus

\[ I_{10} = \pi \ln (2/b) \text{ for } x = b \]  \hfill (I.4)
(iii) $x = -b$; making the substitutions of (I.2)

$$I_{10} = -\int_{0}^{\pi} \ln \left| -b - b\cos \theta \right| d\theta$$

$$= -\int_{0}^{\pi} \ln \left| (-1) (b + b\cos \theta) \right| d\theta$$

since $(b + b\cos \theta) \geq 0$ for $0 \leq \theta \leq \pi$ then:

$$I_{10} = -\int_{0}^{\pi} \ln (b + b\cos \theta) d\theta$$

gives

$$I_{10} = \pi \ln \left( \frac{2}{b} \right) \text{ for } x = -b \quad (I.5)$$

(iv) $x > b$; with substitutions (I.2)

$$I_{10} = -\int_{0}^{\pi} \ln \left| x - b\cos \theta \right| d\theta$$

since $x > b$ then argument of $\ln$ is $> 0$ thus

$$I_{10} = -\int_{0}^{\pi} \ln (x - b\cos \theta) d\theta$$

which gives

$$I_{10} = \pi \ln \left( \frac{2}{x + \sqrt{x^2 - b^2}} \right) \text{ for } x > b \quad (I.6)$$

(v) $x < -b$; with substitutions (I.2)

$$I_{10} = -\int_{0}^{\pi} \ln \left| x - b\cos \theta \right| d\theta$$
\[ I_{10} = - \int_{0}^{\pi} \ln |(-1)(x + b\cos \theta)| \, d\theta \]

taking \( x > +b \) then argument of \( \ln \) will be +ve over \( 0 \leq \theta \leq \pi \) if (-1) is removed giving

\[ I_{10} = \int_{0}^{\pi} \ln (x + b\cos \theta) \, d\theta \]

thus

\[ I_{10} = \pi \ln \left( \frac{2}{x + \sqrt{x^2 - b^2}} \right) \text{ for } x < -b \quad (I.7) \]

Thus the first integral \( (I_{10}) \) has been solved for all the points on the x-axis where the electrostatic potential will be calculated.

I.3 Solution of \( I_{20} \)

Solving for \( I_{20} \)

\[ I_{20} = - \int_{-b}^{b} \frac{x}{x_0} \ln |x - x_0| \, dx_0 \]

\[ = \int_{-b}^{b} \left[ \ln |x - x_0| \right] \cdot \left[ \frac{-x_0}{\sqrt{b^2 - x_0^2}} \right] \, dx_0 \]

\[ = \int_{-b}^{b} \frac{-x_0}{\sqrt{b^2 - x_0^2}} \ln |x - x_0| \, dx_0 \]

\[ = b \left( \frac{b}{\sqrt{b^2 - x_0^2}} \right) \ln |x - x_0| \, dx_0 \]
\[ I_{20} = \int_{-b}^{b} \frac{d \ln |x|}{\sqrt{b^2 - x^2}} \, dx \]

consider the transformation
\[ u = x - x_0 \quad x_0 = x - u \]
\[ du = dx_0 \quad dx_0 = -du \]

then
\[ I_{20} = -\int_{x-b}^{x+b} \frac{d \ln |u|}{\sqrt{(b^2-x^2) + (2x)u + (-1)u^2}} \frac{du}{du} \]

(i) \( x = 0 \); taking (I.8) with \( x = 0 \)
\[ I_{20} = -\int_{b}^{-b} \frac{d \ln |-x_0|}{\sqrt{b^2 - x_0^2}} \, dx_0 \]
\[ = -\int_{-b}^{b} \frac{d \ln (x)}{\sqrt{b^2 - x^2}} \, dx_0 \]
\[ = -\int_{-b}^{b} \frac{d \ln (x_0)}{\sqrt{b^2 - x^2}} \, dx_0 \]
\[ = -\int_{0}^{b} \frac{d \ln (x)}{\sqrt{b^2 - x^2}} \, dx_0 \]
\[ = -\int_{0}^{b} \frac{d \ln (x_0)}{\sqrt{b^2 - x^0}} \, dx_0 \]
\[ = -\int_{0}^{b} \frac{d \ln (x)}{\sqrt{b^2 - x^0}} \, dx_0 \]
\[ = \int_{0}^{b} \frac{d \ln (x)}{\sqrt{b^2 - x^0}} \, dx_0 \]
\[ = \int_{0}^{b} \frac{d \ln (x)}{\sqrt{b^2 - x^0}} \, dx_0 \]
Thus \( I_{20} = 0 \) for \( x = 0 \) \hspace{1cm} (I.11)

(ii) \( x = +b \); taking (I.10) and putting \( x = +b \) gives:

\[
I_{20} = \int_{b-b}^{b+b} \frac{b-b}{\sqrt{(b^2-b^2) + (2b)u + (-1)u^2}} \frac{d \ln |u|}{du} \, du
\]

\[
= - \int_{2b}^{0} \frac{\sqrt{2bu - u^2}}{u} \, du
\]

\[
= [\sqrt{2bu + \sqrt{2bu-u^2}} \ln \left( \frac{\sqrt{2bu} + \sqrt{2bu-u^2}}{u} \right)]_{0}^{2b}
\]

\[
= 0
\]

Thus \( I_{20} = 0 \) for \( x = +b \) \hspace{1cm} (I.12)

(iii) \( x = -b \); taking (I.10) and putting \( x = -b \) gives:

\[
I_{20} = - \int_{-b-b}^{-b+b} \frac{-b-b}{\sqrt{(b^2-b^2) - (2b)u + (-1)u^2}} \frac{d \ln |u|}{du} \, du
\]

\[
= - \int_{-2b}^{-\infty} \frac{\sqrt{-2bu - u^2}}{-u} \, du
\]

\[
= \int_{0}^{2b} \frac{\sqrt{2bu - u^2}}{u} \, du
\]

\[
= [\sqrt{2bu + \sqrt{2bu-u^2}} \ln \left( \frac{\sqrt{2bu} + \sqrt{2bu-u^2}}{u} \right)]_{0}^{2b}
\]

\[
= 0
\]
Thus \( I_{20} = 0 \) for \( x = -b \) \hspace{1cm} (I.13)

(iv) \( x > b \); taking (I.10) gives

\[
I_{20} = - \int_{x+b}^{x-b} \sqrt{(b^2-x^2) + (2x)u + (-1)u^2} \frac{d \ln |u|}{du} du
\]

\[
= - \int_{x+b}^{x-b} \frac{\sqrt{(b^2-x^2) + (2x)u + (-1)u^2}}{u} du
\]

\[
= - \sqrt{\frac{b^2-x^2}{x+b}} - \frac{B}{2} \int_{x+b}^{x-b} \frac{du}{\sqrt{U}} - A \int_{x+b}^{x-b} \frac{du}{u \sqrt{U}} \hspace{1cm} (I.14)
\]

where

\[
U = (b^2-x^2) + (2x)u + (-1)u^2
\]

\[
A = b^2 - x^2
\]

\[
B = 2x
\]

\[
C = -1
\]

From (I.14)

\[
\sqrt{\frac{b^2-x^2}{x+b}} = 0
\]

\[
\frac{B}{2} \int_{x+b}^{x-b} \frac{du}{\sqrt{U}}
\]

and

\[
= - \frac{3/2}{\sqrt{\pi}} \sin^{-1} \left[ \frac{-2u + 2x}{\sqrt{4Au + 4x^2}} \right]_{x+b}^{x-b}
\]

\[
= \pi x
\]

and

\[
A \int_{x+b}^{x-b} \frac{du}{u \sqrt{U}}
\]
\[\begin{align*}
&= \frac{(b^2 - x^2)}{\sqrt{x^2 - b^2}} \sin^{-1} \left[ \frac{2x^2 + 2b^2 - 2x^2}{|u| \sqrt{4b^2 - 4x^2 + 4x^2}} \right]_{x-b}^{x+b} \\
&= \pi \sqrt{x^2 - b^2}
\end{align*}\]

Thus combining the above terms into (I.14) gives

\[I_{20} = -\pi \left[ \sqrt{x^2 - b^2} + x \right] \text{ for } x > b \quad (I.15)\]

(v) \(x < -b\); taking (I.10) gives

\[I_{20} = - \int_{x+b}^{x-b} \frac{\sqrt{(u^2-x^2) + (2x)u + (-1)u^2}}{\sqrt{b^2-x^2}} \frac{d\ln|u|}{du} \, du - (|x|+b)\]

\[= - \int_{-u}^{\sqrt{u^2-b^2}} \frac{\sqrt{u^2-b^2}}{-u} \, du - (|x|-b)\]

\[= + \int_{|x|+b}^{u} \frac{\sqrt{u^2-b^2}}{u} \, du\]

which is the negative of case (iv). Thus

\[I_{20} = +\pi \left[ \sqrt{x^2 - b^2} - x \right] \text{ for } x < -b \quad (I.16)\]

Thus the second integral has been solved for all the points on the \(x\)-axis where the electrostatic potential was calculated.
I.4 Solution of \( I_{30} \)

Solving for \( I_{30} \)

\[
I_{30} = - \int_{-b}^{b} \frac{x_o^2 \ln |x-x_o|}{\sqrt{b^2 - x^2}} \, dx_o
\]

(i) \( x = 0 \); taking \( x = 0 \)

\[
I_{30} = - \int_{-b}^{b} \frac{x_o^2 \ln |x|}{\sqrt{b^2 - x^2}} \, dx_o
\]

\[
= - \frac{b^2 \pi}{2} \left[ \ln \left( \frac{2}{b} \right) - 1 \right]
\]

Thus \( I_{30} = \frac{b^2 \pi}{2} \left[ \ln \left( \frac{2}{b} \right) - 1 \right] \) for \( x = 0 \) \qquad (I.17)

(ii) \( x = +b \); taking \( x = +b \) then:

\[
I_{30} = - \int_{-b}^{b} \frac{x_o^2 \ln (b-x_o)}{\sqrt{b^2 - x^2}} \, dx_o
\]

\[
= - \left[ \frac{x_o}{\sqrt{b^2 - x^2}} \right] \cdot \left[ x_o \ln (b-x_o) \right] dx_o
\]

Integrating by parts gives:
\[
I_{30} = \left[ x_o \frac{\sqrt{b^2-x_o^2}}{b-x_o} \ln (b-x_o) \right]_b^{-b} - \int_{-b}^{b} \frac{\sqrt{b^2-x_o^2}}{b-x_o} [\ln (b-x_o) - \frac{x_o}{b-x_o}] \, dx_o.
\]

\[
= \int_{-b}^{b} \frac{x_o \sqrt{b^2-x_o^2}}{b-x_o} \, dx_o - \int_{-b}^{b} \frac{\sqrt{b^2-x_o^2}}{b-x_o} \ln (b-x_o) \, dx_o \quad \text{(I.18)}
\]

Taking one term at a time and applying the transformation:

\[
x_o = b \cos \theta \quad \text{dx}_o = -b \sin \theta \, d\theta
\]

\[
\sqrt{b^2-x_o^2} = b \sin \theta \quad \text{d}\theta = -\frac{dx_o}{\sqrt{b^2-x_o^2}}
\]

gives for the second term of (I.18):

\[
- \int_{-b}^{b} \frac{\sqrt{b^2-x_o^2}}{b-x_o} \ln (b-x_o) \, dx_o
\]

\[
= \int_{\pi}^{0} b^2 \sin^2 \theta \ln (b-b \cos \theta) \, d\theta
\]

\[
= b^2 \int_{\pi}^{0} \ln (b-b \cos \theta) \, d\theta - \int_{\pi}^{0} b^2 \cos^2 \theta \ln (b-b \cos \theta) \, d\theta
\]

\[
= -b^2 \int_{0}^{\pi} \ln (b-b \cos \theta) \, d\theta + \int_{-b}^{b} \frac{x_o^2 \ln (b-x_o)}{\sqrt{b^2-x_o^2}} \, dx_o
\]

\[
= -b^2 \pi \ln \left[ \frac{b + \sqrt{b^2-b^2}}{2} \right] - I_{30}
\]
\[ = - b^2 \pi \ln \left( \frac{b}{2} \right) - I_{30} \quad (I.19) \]

Now taking the first term of (I.18) and applying the transformation:

\[ u = b - x_0 \quad du = -dx_0 \]

\[ x_0 = b - u \quad dx_0 = -du \]

\[ \int_{b-x_0}^{b} \frac{\sqrt{b^2 - x_0^2}}{b - x_0} \, dx_0 \]

\[ = - \int_{b-u}^{0} \frac{(b-u) \sqrt{b^2 - b^2 + 2bu - u^2}}{u} \, du \]

\[ = - \int_{0}^{2b} \frac{\sqrt{2bu - u^2}}{u} \, du + \int_{0}^{2b} \frac{\sqrt{2bu - u^2}}{u} \, du \]

\[ = - \frac{b^2 \pi}{2} - b^2 \pi \quad (I.20) \]

Substituting the results (I.19) and (I.20) into (I.18) gives:

\[ I_{30} = - \frac{b^2 \pi}{2} - b^2 \pi \ln \left( \frac{b}{2} \right) - I_{30} \]

\[ 2I_{30} = - b^2 \pi \left[ \ln \left( \frac{b}{2} \right) + \frac{3}{2} \right] \]

thus \[ I_{30} = - \frac{b^2 \pi}{2} \left[ \ln \left( \frac{b}{2} \right) + \frac{3}{2} \right] \] for \( x = +b \) \quad (I.21)
(iii) \( x = -b \); taking \( x = -b \)

\[
I_{30} = - \int_{-b}^{b} \frac{x_o^2 \ln |(1) (b+x_o)|}{\sqrt{b^2 - x_o^2}} \, dx_o
\]

\[
= - \int_{-b}^{b} \frac{x_o^2 \ln |(1) (b+x_o)|}{\sqrt{b^2 - x_o^2}} \, dx_o
\]

transferring \( x = -b \) to \( x = +b \)

\[
I_{30} = + \int_{b}^{b} \frac{x_o^2 \ln (b-x_o)}{\sqrt{b^2 - x_o^2}} \, dx_o
\]

and reversing order of integration

\[
I_{30} = - \int_{b}^{b} \frac{x_o \ln (b-x_o)}{\sqrt{b^2 - x_o^2}} \, dx_o
\]

which is the same as for case (ii)

thus \( I_{30} = - \frac{b^2 \pi}{2} [\ln (\frac{b}{2}) + \frac{3}{2}] \) for \( x = -b \) \hfill (I.22)

(iv) \( x > +b \); taking the region of \( x > +b \)

\[
I_{30} = - \int_{-b}^{b} \frac{x_o^2 \ln |x-x_o|}{\sqrt{b^2 - x_o^2}} \, dx_o
\]
since \( x > b \) then \( x - x_o > 0 \) for region of integration

\[
I_{30} = - \int_{-b}^{b} \frac{x_o^2 \ln (x-x_o)}{\sqrt{b^2 - x_o^2}} \, dx_o
\]

\[
= - \int_{-b}^{b} [x_o \ln (x-x_o)] \cdot \left( \frac{x_o}{\sqrt{b^2 - x_o^2}} \right) \, dx_o
\]

Integrating by parts gives:

\[
I_{30} = [x_o \sqrt{b^2 - x_o^2} \ln (x-x_o)]_{-b}^{b} - \int_{-b}^{b} \frac{x_o^2}{\sqrt{b^2 - x_o^2}} \left[ \ln (x-x_o) - \frac{x_o}{(x-x_o)} \right] \, dx_o
\]

\[
= 0 - \int_{-b}^{b} \frac{\sqrt{b^2 - x_o^2}}{x - x_o} \ln (x-x_o) \, dx_o + \int_{-b}^{b} \frac{x_o \sqrt{b^2 - x_o^2}}{x - x_o} \, dx_o
\]

Taking the first term and substituting:

\[ x_o = b \cos \theta \]
\[ dx_o = -b \sin \theta \, d \theta \]

\[ \sqrt{b^2 - x_o^2} = b \sin \theta \]
\[ d \theta = -\frac{dx_o}{\sqrt{b^2 - x_o^2}} \]

gives:

\[
- \int_{-b}^{b} \frac{\sqrt{b^2 - x_o^2}}{x - x_o} \ln (x-x_o) \, dx_o
\]

\[
= \int_{-b}^{b} b^2 \sin^2 \theta \ln (x - b \cos \theta) \, d \theta
\]
\[ I_30 = b^2 \int_{x}^{0} \ln \left( x - b \cos \theta \right) \, d\theta - \int_{x}^{0} b^2 \cos^2 \theta \ln \left( x - b \cos \theta \right) \, d\theta \]

\[ = - b^2 \pi \ln \left[ \frac{x + \sqrt{x^2 - b^2}}{2} \right] - I_{30} \quad (I.23) \]

Taking the second term and substituting:

\[ u = x - x_o \quad \text{du} = -dx_o \]

\[ x_o = x - u \quad \text{dx}_o = -du \]

\[ \int_{b}^{x} \frac{x_o \sqrt{b^2 - x_o^2}}{x - x_o} \, dx_o \]

\[ = \int_{x}^{x+b} \frac{(x-u) \sqrt{b^2 - x^2} + (2x)u + (-1)u^2}{u} \, du \]

\[ = \int_{x}^{x+b} \frac{x \sqrt{b^2 - x^2} + (2x)u + (-1)u^2}{u} \, du \]

\[ - \int_{x}^{x+b} \frac{\sqrt{b^2 - x^2} + (2x)u + (-1)u^2}{u} \, du \]

\[ = \pi \left[ x \sqrt{x^2 - b^2} - x^2 - b^2/2 \right] \quad (I.24) \]

Combining the results (I.23) and (I.24) gives

\[ I_{30} = - b^2 \pi \ln \left[ \frac{x + \sqrt{x^2 - b^2}}{2} \right] - I_{30} + \pi \left[ x \sqrt{x^2 - b^2} - x^2 - b^2/2 \right] \]

\[ 2I_{30} = - b^2 \pi \ln \left[ \frac{x + \sqrt{x^2 - b^2}}{2} \right] + \pi \left[ x \sqrt{x^2 - b^2} - x^2 - b^2/2 \right] \]
Thus

\[ I_{30} = \frac{\pi}{2} \left\{ \frac{2}{x + \sqrt{x^2 - b^2}} + x \sqrt{\frac{x^2 - b^2}{2}} - \frac{b^2}{2} \right\} \]

for \( x > b \) \hspace{1cm} (I.25)

(v) \( x < -b \); considering the region where \( x < -b \)

\[ I_{30} = -\int_{-b}^{b} \frac{x_o^2 \ln |x-x_o|}{\sqrt{b^2 - x_o^2}} \, dx_o \]

\[ = -\int_{-b}^{b} \frac{b x_o^2 \ln \left| \frac{1}{(-1)} \left| \frac{|x|+x_o}{} \right| \right|}{\sqrt{b^2 - x_o^2}} \, dx_o \]

\[ = -\int_{-b}^{b} \frac{b x_o^2 \ln \left( \frac{|x|+x_o}{} \right)}{\sqrt{b^2 - x_o^2}} \, dx_o \]

Transferring \( x_o \rightarrow -x_o \)

\[ = -\int_{-b}^{b} \frac{b x_o^2 \ln \left( \frac{|x|}{} \right)}{\sqrt{b^2 - x_o^2}} \, dx_o \]

and reversing order of integration

\[ = -\int_{-b}^{b} \frac{x_o^2 \ln \left( \frac{|x|}{} \right)}{\sqrt{b^2 - x_o^2}} \, dx_o \]

Thus from case (iv):
\[ I_{30} = \frac{\pi}{2} \left[ b^2 \ln \left( \frac{2}{|x| + \sqrt{x^2 - b^2}} \right) + |x| \sqrt{x^2 - b^2} - \frac{b^2}{2} \right] \]

for \( x < -b \) \hspace{1cm} (1.26)

### 1.5 Summary

Of course all the above integrals must have a reference constant such that \( I_{10} = I_{20} = I_{30} = 0 \) at \( |x| = \pm \infty \). This is found to be the value of \( x \) thus the normalized solutions of the integrals are:

\[ I_{10} = \pi \ln (2) \]

for \( x = 0 \)

\[ I_{10} = \pi \ln \left( \frac{2}{1 + \sqrt{1 - (b/x)^2}} \right) \]

for \( |x| \geq b \)

\[ I_{20} = 0 \]

for \( x = 0 \)

\[ I_{20} = -\pi \left( \sqrt{1 - (b/x)^2} - 1 \right) \]

for \( x > b \)

\[ I_{20} = \pi \left( \sqrt{1 - (b/x)^2} - 1 \right) \]

for \( x < -b \)

\[ I_{30} = \frac{\pi}{2} \left[ \ln (2) - \frac{1}{2} \right] \]

for \( x = 0 \)

\[ I_{30} = \frac{\pi}{2} \left[ \frac{b^2}{x} \ln \left( \frac{2}{1 + \sqrt{1 - (b/x)^2}} \right) + \sqrt{1 - (b/x)^2} - 1 - \frac{1}{2} \right] \]

for \( |x| \geq b \)
APPENDIX II

SOLUTION OF WAVE EQUATION INTEGRALS

II.1 General

To integral equations of Chapter 4 to be solved are:

\[ I_{10} = \int_{-b}^{b} e^{j k x_0} \frac{1}{\sqrt{b^2 - x_0^2}} \, dx_0 \]

\[ I_{20} = \int_{-b}^{b} x_0 e^{j k x_0} \frac{1}{\sqrt{b^2 - x_0^2}} \, dx_0 \]

\[ I_{30} = \int_{-b}^{b} x_0 e^{j k x_0} \frac{1}{\sqrt{b^2 - x_0^2}} \, dx_0 \]  \hspace{1cm} (II.1)

II.2 Solution to \( I_{10} \)

\[ I_{10} = -\int_{-b}^{b} e^{j k x_0} \frac{1}{\sqrt{b^2 - x_0^2}} \, dx_0 \]

\[ \text{putting } x_0 = b \sin \theta \]

\[ \sqrt{b^2 - x_0^2} = b \sin \theta \]

\[ dx_0 = -b \sin \theta \, d\theta \]  \hspace{1cm} (II.2)
gives:

\[ I_{10} = \int_{0}^{\pi} e^{i k b \cos \theta} d\theta \]

\[ = \int_{0}^{\pi} \left[ 1 + j k b \cos \theta - \frac{(k b)^2}{2} - j \frac{(k b)^3}{3!} \cos^3 \theta 
+ \frac{(k b)^4}{4!} \cos^4 \theta + j \frac{(k b)^5}{5!} \cos^5 \theta \ldots \right] d\theta \]

\[ = \int_{0}^{\pi} \left[ 1 - \frac{(k b)^2}{2!} \cos^2 \theta + \frac{(k b)^4}{4!} \cos^4 \theta \ldots \right] d\theta \]

\[ + j k b \int_{0}^{\pi} \left[ \cos \theta - \frac{(k b)^2}{3!} \cos^3 \theta + \frac{(k b)^4}{5!} \cos^5 \theta \ldots \right] d\theta \]

\[ = [\pi - \frac{(k b)^2}{2!} \left( \frac{1}{2} \theta + \frac{1}{4} \sin 2\theta \right)_{0}^{\pi} + \frac{(k b)^4}{4!} \left( \frac{3}{8} \sin 2\theta - \frac{\sin 4\theta}{4} + \frac{\sin 6\theta}{32} \right)_{0}^{\pi} \ldots \]

\[ + j k b \left[ \sin \theta - \frac{(k b)^2}{3!} (\sin \theta)(\cos^2 \theta + 2) \right]_{0}^{\pi} + \frac{(k b)^4}{5!} \left( \frac{1}{5} \cos^4 \theta \sin \theta \right)_{0}^{\pi} \]

\[ + \frac{4}{5} \int_{0}^{\pi} \cos^2 \theta d\theta \ldots \]

\[ = \pi \left[ 1 - \frac{(k b)^2}{2} + \frac{(k b)^4}{4} \cdot \frac{3}{2} + \frac{(k b)^6}{6} \cdot \frac{5}{2} \cdot \frac{3}{4} \cdot \frac{5}{6} \ldots \right] \]

\[ + j k b \left[ 0 \right] \]

\[ = \pi \left[ 1 - \frac{(k b)^2}{2^{2}} + \frac{(k b)^4}{2^{2} \cdot 4^{2}} - \frac{(k b)^6}{2^{2} \cdot 4^{2} \cdot 6^{2}} \ldots \right] \]
Thus $I_{10} = \pi J_0(kb)$ is the solution to the first integral.

II.3 Solution to $I_{20}$

Using transformations (II.2):

$$I_{20} = \int_0^\pi b \cos \theta \ e^{jkbcos \theta} \ d\theta$$

$$= \int_0^\pi b \cos \theta \left[ 1 + jk \ cos \ \theta - \frac{(kb)^2 \cos^2 \theta}{2} - j \frac{(kb)^2 \cos^3 \theta}{3} + \frac{(kb)^4 \cos^4 \theta}{4} + j \frac{(kb)^5 \cos^5 \theta}{5} \ldots \right] \ d\theta$$

$$= \frac{1}{k} \int_0^\pi \left[ kb \ cos \ \theta - \frac{(kb)^3 \cos^3 \theta}{2} + \frac{(kb)^5 \cos^5 \theta}{4} \ldots \right] \ d\theta$$

$$+ \frac{1}{k} \int_0^\pi \left[ (kb)^2 \ cos^2 \ \theta - j \frac{(kb)^4 \cos^4 \ \theta}{3} + \frac{(kb)^6 \cos^6 \theta}{5} \ldots \right] \ d\theta$$

$$= \frac{1}{k} \left[ 0 + \frac{jkb}{k} \left[ \frac{1}{2} \ 0 + \frac{1}{4} \sin 2\theta \right] \right]\left[ \frac{(kb)^3}{31} \left( \frac{3\theta}{8} + \frac{\sin 2\theta}{4} + \frac{\sin 4\theta}{32} \right) \right]_0^\pi$$

$$+ \frac{(kb)^5}{51} \left( \frac{1}{6} \cos^5 \ \theta \sin \theta \right) \left[ \frac{5}{6} \int_0^\pi \cos^4 \theta \ d\theta \right] + \ldots$$

$$= j \ b \left[ \frac{kb\pi}{2} - \frac{(kb)^3}{31} \left( \frac{3\pi}{8} \right) + \frac{(kb)^5}{51} \left( \frac{5}{6} \left( \frac{3\pi}{8} \right) \right) \ldots \right]$$
\[
= j \pi b \left[ \frac{kb}{2} - \frac{(kb)^3}{2^2 \cdot 4} + \frac{(kb)^5}{2^2 \cdot 4^2 \cdot 6} + \ldots \right]
= j \pi b J_1(kb)
\]

Thus the solution to the second integral is
\[
I_{20} = j \pi b J_1(kb)
\]

III.4 Solution to \(I_{30}\)

Using the transformation (II.2)

\[
I_{30} = \int_0^\pi b^2 \cos^2 \theta e^{jkbcos \theta} d\theta
\]

\[
= \int_0^\pi b^2 (1 - \sin^2 \theta) e^{jkbcos \theta} d\theta
\]

\[
= b^2 \left[ \int_0^\pi e^{jkbcos \theta} d\theta - \int_0^\pi b^2 \sin^2 \theta e^{jkbcos \theta} d\theta \right]
\]

\[
= b^2 \pi J_0(kb) + \frac{1}{kb} \left[ \left. \int_0^\pi \text{bsin} \theta \text{e}^{jkbcos \theta} \right|_{-jkb}^{jkb} \right] d\theta
\]

\[
= b^2 \pi J_0(kb) + \frac{1}{kb} \int_0^\pi \text{bsin} \theta e^{jkbcos \theta} d\theta
\]

\[
= b^2 \pi J_0(kb) - \frac{\pi b}{kb} J_1(kb)
\]

Thus the solution to the third integral is:
\[
I_{30} = b^2 \pi \left[ J_0(kb) - \frac{1}{kb} J_1(kb) \right]
\]
II.5 **Summary**

The solution to the wave equation integrals in the far field (i.e., outside the transducer region) is:

\[ I_{10} = \tau J_0 (k b) \]
\[ I_{20} = j \tau b J_1 (k b) \]
\[ I_{30} = b^2 \tau [J_0 (k b) - \frac{1}{kb} J_1 (k b)] \]

where:  
- \( J_0 \) is the Bessel function of the first kind  
- \( J_1 \) is the Bessel function of the second kind  
- \( k \) is the propagation constant  
- \( b \) is 1/2 the finger width
APPENDIX III

SAWINIT

The following is a listing of the initial programme SAWINT. It is written in FORTRAN IV and operates under the NOS system for CDC-6000 computer systems. System libraries used are:

(1) SSPLIB - scientific subroutines package
(2) PLOTVER - Versatec plotting library package.

The files that are saved are: TAPE 1 and TAPE 2.
SAWINIT. THIS PROGRAM DOES 1ST PASS OF SAW DESIGN. SUTHERS
USER
CHARGE
FTN.
SAVE, INPUT=SAWINIT.

PROGRAM TST (INPUT, OUTPUT, TAPE5=INPUT, TAPE6=OUTPUT, TAPE1, TAPE2)

COMPLEX FFTS, V, RIDT, AR
REAL MAG
REAL LEN
REAL MDB, PI, IMAG, LENGTH, RELBP, IMAGL
DIMENSION AR(513)
DIMENSION V(500), XP(500), B(500), TIM(500), LEN(500), FRSP(30, 3), FREQ(C513), MDB(513), PHASE(513), X(513), Y(513), Z(513), RMDB(513), RPHS(513), CFPTS(513), RDT(513), MAG(513), PHSE(513)
DIMENSION C0(500), C1(500), C2(500)
CALL INPUT(N, NP, FRSP, RLBW, FREQ, MDB, PHASE, IG, FO, IC, ID)
T=N/(FO*RLBW)

C T IS THE PERIOD OF THE IDT IN MICROSECONDS.
DT=T/N

C DT IS THE TIME INTERVAL OF THE FFT SAMPLE POINTS.
SPBD =FO*RLBW
NZ = N+1
NH=N/2
NM=NH+1
PI = ACOS(-1.0)
IR=0
DO 10 I=1, NZ
IF (MDB(I).EQ.0.0.AND.IR.EQ.0) IR=I
MAG(I)=MDB(I)
AMP=10.0**(MDB(I)/20.0)
PHSE(I)= PHASE(I)
CALL RECT(AMP, PHSE(I), RR, RI)
FFTS(I)=CMPLX(RR, RI)
10 CONTINUE
IR=IR+3
CALL PLTFREQ(N,FREQ,MAG,PHSE)

C INITIALIZATION IS COMPLETE

CALL SAWFFT(FFTS,N,IG,X,Y,Z)
X(NZ)=6.0
Y(NZ)=Y(1)
Z(NZ)=Z(1)
CALL PLTFFT(NZ,X,Y,Z)
FC=FO-SPBD/2.0
DO 95 I=1,NH
Z(NM+I)=Z(NM+I)-I*360.0*DT*(FO-FC)
Z(NM-I)=Z(NM-I)+I*360.0*DT*(FO-FC)
95 CONTINUE
CALL PHSSTR(NZ,Y,Z)
CALL I0(8.0,DEN)
DO 100 I=1,NH
TM=FLOAT(NH**2-I**2)
ARG=(16.0/N)*SQRT(TM)
CALL I0(ARG,ANUM)
WT=ANUM/DEN
Y(NH+1+I)=Y(NH+1+I)*WT
Y(NH+1-I)=Y(NH+1-I)*WT
100 CONTINUE
CALL IDT(NZ,Y,Z,FO,SPBD,LEN,TIM,L)
F1=FREQ(NH-2)
F2=FREQ(NH+2)
VEL=3410
CALL MODEL(L,LEN,TIM,B,XP,T,F1,F2,VEL,C0,C1,C2,AR)
CALL PRQRSP(L,AR,TIM,FREQ,RMDB,RPHS,NZ,T,RIDT)
WRITE (1,20) NZ,L,IG,T,DT,SPBD,FO,FC,VEL
RMAX=RMDB(IR)
DO 50 I=1,NZ
RMDB(I) = RMDB(I) - RMAX
WRITE (1, 30) FREQ(I), MDB(I), PHASE(I), RMDB(I), RPHS(I), MAG(I), PHSE(I)
+ I)
IF (RMDB(I) .LT. -100.0) RMDB(I) = -100.0
50 CONTINUE
20 FORMAT (3I5, 6E10.5)
30 FORMAT (7E11.5)
WRITE (2, 81) L
DO 70 I = 1, L
XL = XP(I) - B(I)
XR = XP(I) + B(I)
WRITE (2, 80) LEN(I), XL, XR
70 CONTINUE
80 FORMAT (3E20.10)
81 FORMAT (15)
WRITE (6, 152)
DO 150 I = 1, N
WRITE (6, 151) FREQ(I), MAG(I), MDB(I), RMDB(I), PHSE(I), PHASE(I), + RPHS(I)
150 CONTINUE
CALL PLTFREQ(N, FREQ, RMDB, RPHS)
CALL PLOT (X(N), Y(N), 999)
STOP
END

SUBROUTINE INPUT(N, NP, FRSP, RLBW, FREQ, MDB, PHASE, IG, FO, IC, ID)
C THE INPUT SUBROUTINE READS IN THE SPECIFICATIONS IN THE FREQUENCY
C DOMAIN AND INITIALIZES THE INPUT VECTOR FOR THE FFT.
C N IS THE NUMBER OF POINTS FOR THE FFT WHERE N=2**IG. OUTPUT
C NP IS THE NUMBER OF SPECIFIED FREQUENCY POINTS. INPUT
C FRSP IS A 3 X NP ARRAY SPECIFICATION OF INPUT FREQUENCIES. INPUT
C SPBD IS THE SPECIFIED BAND WIDTH.
C FREQ IS THE VECTOR OF CALCULATED FREQUENCY POINTS FOR FFT. OUTPUT
C MDB IS THE VECTOR OF CALCULATED AMPLITUDES FOR FFT. OUTPUT
C PHASE IS THE VECTOR OF PHASE SPECIFICATIONS FOR FFT. OUTPUT
C IG IS THE POWER OF 2 SUCH THAT N=2**IG. INPUT, OUTPUT
C DIMENSION FRSP(30,3),FREQ(513),MDB(513),PHASE(513)
C REAL MDB
C READ (5,10) IG, NP, RLBW , FO
10 FORMAT (2I10,2F10.5)
C FO IS DESIGNATED AS CENTRE FREQUENCY.
C SPBD=FO*RLBW
C SPBD IS THE SPECIFIED BAND OF FREQUENCIES.
C DO 31 I=1, NP
31 CONTINUE
C READ (5,30) FRSP(I,1),FRSP(I,2),FRSP(I,3)
C FORMAT (3F10.2)
C FRSP(I,1) IS FREQUENCY IN MHZ.
C FRSP(I,2) DB OF INSERTION LOSS
C FRSP(I,3) IS PHASE SPECIFICATIONS IN DEGREES.
C N=2**IG
C DF=SPBD/N
C DF IS FREQUENCY INCREMENT BETWEEN POINTS
C K=2
C FO IS THE CENTRE OF THE SPECIFIED BANDPASS
C CALCULATE FREQUENCY POINTS.
C NZ=N+1
C NH=N/2
C NL=NH+1
C FREQ(NL)=FO.
C DO 49 I=1, NH
49 CONTINUE
C FREQ(NL-I)=FO-I*DF
C FREQ(NL+I)=FO+I*DF
C ENTER AMPLITUDE AND PHASE SPECIFICATIONS FOR EACH FREQUENCY.
DO 50 I=1,NZ
IF (FREQ(I).LT.FRSP(1,1)) MDB(I)=-FRSP(1,2)
IF (FREQ(I).GT.FRSP(NP,1)) MDB(I)=-FRSP(NP,2)
PHASE(I)=0.0

53 IF (K.GT.NP) GO TO 50
IF (FREQ(I).LT.FRSP(K-1,1)) GO TO 50
IF (FREQ(I).GT.FRSP(K,1)) GO TO 54
S=(FREQ(I)-FRSP(K-1,1))/(FRSP(K,1)-FRSP(K-1,1))
MDB(I)=FRSP(K-1,2)+(FRSP(K,2)-FRSP(K-1,2))*S
MDB(I)=-MDB(I)
PHASE(I)=FRSP(K-1,3)+(FRSP(K,3)-FRSP(K-1,3))*S
GO TO 50

54 K=K+1
GO TO 53
50 CONTINUE
FC=(FRSP(NP,1)+FRSP(1,1))/2.0
DO 60 I=1,N
IF (FREQ(I)-FC) GO TO 62,61,62
60 CONTINUE
62 IF ((FREQ(I)-FC).LT.(FC-.FREQ(I-1))) GO TO 61
GO TO 63
61 FC=FREQ(I)
IC=I
GO TO 65
63 FC=FREQ(I-1)
IC=I-1
65 ID=IC* ((FRSP(NP,1)-FRSP(1,1))/FC)
RETURN
END

SUBROUTINE SAWFFT(FFT1,N,IG,X,Y,Z)
COMPLEX FFT1,FFT2,W,WK,WP
REAL IMAG
DIMENSION FFT1(N),IFFT(513,2),FFT2(513),X(N),Y(N),Z(N)
PI=ACOS(-1.0)
L=N
C FAST FOURIER TRANSFORM FOR SAW DESIGN
C FREQ IS ARRAY OF FREQUENCIES OF POINTS OF FFT
C MDB IS ARRAY OF DB OF INSERTION LOSS FOR EACH FREQUENCY
C PHASE IS ARRAY OF PHASES FOR EACH FREQUENCY
C IFFT IS ARRAY KEEPING TRACK OF EXPONENTS FOR MULTIPLIERS W(J) IN
C THE FFT PROCESS
C FFT1 AND FFT2 ARE COMPLEX VECTORS OF FFT PROCESS
DO 10 I=1,513
DO 10 J=1,2
10 IFFT(I,J) = 0
C CALCULATE THE EXPONENT FOR THE MULTIPLIER FOR THE FFT PROCESS
AREAL = COS(2.*PI/N)
IMAG = SIN(2.*PI/N)
W = CMPLX(AREAL,IMAG)
DO 70 M=1,IG
L=L/2
NH = N/2
NM=NH+1
DO 50 K=1,NH
IFFT(K,2) = IFFT(2*K,1)
50 IFFT(K+NH,2)=IFFT(K,2) + L
C CALCULATE NEXT VECTOR IN FFT PROCESS
DO 60 K=1,N
WK = W**IFFT(K,2)
IF (IFFT(K,2).GT.IFFT(K,1)) GO TO 61
WP = WK*FFT1(K+L)
FFT2(K) = FFT1(K)+WP
GO TO 60
61 WP = WK*FFT1(K)
FFT2(K) = FFT1(K-L) + WP
60 CONTINUE
DO 70 I=1,N
FFT1(I) = FFT2(I)
IFFT(I,1) = IFFT(I,2)
CONTINUE
C
PUT OUTPUT VECTOR (DFT) IN CORRECT ORDER.
AMAX = 0.0
DO 82 K = 1, N
KI = 0
81
KI = KI + 1
IF (IFFT(KI,1).EQ.K-1) GO TO 80
GO TO 81
80
RR = REAL(FFT1(KI))
RI = AIMAG(FFT1(KI))
CALL POLAR (RR, RI, P, PA)
Y(K) = PL
Z(K) = PA
IF (Y(K).GT.AMAX) AMAX = Y(K)
X(K) = 9.0 * IFFT(KI,1)/N
82
CONTINUE
DO 83 I = 1, N
Y(I) = Y(I)/AMAX
83
CONTINUE
DO 93 K = 1, NH
YS = Y(K)
Y(K) = Y(K + NH)
Y(K + NH) = YS
ZS = Z(K)
Z(K) = Z(K + NH)
Z(K + NH) = ZS
93
CONTINUE
RETURN
END

SUBROUTINE IDT(NZ, Y, Z, FO, SPBD, LEN, TIM, LH)
REAL LEN
DIMENSION Y(NZ), Z(NZ), LEN(500), TIM(500)
NH=NZ/2
NM=NH+1
N=NZ-1
T=N/SPBD
DT=T/N

C       T IS THE IDT PERIOD IN MICRO-SECONDS
C       DT IS THE TIME INCREMENT OF THE FFT SAMPLES.
C       ADD IN THE LINEAR PHASE COMPONENTS OF THE FREQUENCY SHIFT.
ZM=Z(NM)
DO 40 I=1,NH
Z(NM+I)=Z(NM+I)-2M+ (I*DT*FO*360.0)
Z(NM-I)=Z(NM-I)-2M- (I*DT*FO*360.0)
40    CONTINUE

L=0
M=0
LH=1
LEN(250)=Y(NM)
TIM(250)=0.0
DO 10 I=1,NH
IF (L.EQ.249) GO TO 21
C       NP GIVES THE NUMBER OF FINGERS BETWEEN FFT POINTS.
NC=(+Z(NM-1+I)+0.5)/180.0
PHASL=180.0*NC
NP=(+Z(NM+I)-PHASL+0.5)/180.0
IF (NP.EQ.0) GO TO 21
C       OBTAIN TIMES, LENGTHS AND POLARITIES OF FINGERS.
DO 20 K=1,NP
L=L+1
LH=LH+1
PH=(180.0*(NC+K)-Z(NM-1+I))/(Z(NM+I)-Z(NM-1+I))
TIM(250+L)=(I-1+PH)*DT
SN=(-1.0)**(NC+K)
LEN(250+L)=SN*(Y(NM-1+I)+(Y(NM+I)-Y(NM-1+I))*PH)
20    CONTINUE
21    IF (M.EQ.249) GO TO 10
C       NP GIVES THE NUMBER OF FINGERS BETWEEN FFT POINTS.
NC = (-Z(NM+1-I)+0.5)/180.0
PHASL=180.0*NC
NP = (-Z(NM-I)-PHASL+0.5)/180.0
IF (NP.EQ.0) GO TO 10

C OBTAIN TIMES, LENGTHS AND POLARITIES OF FINGERS.
DO 30 K=1,NP
   M=M+1
   LH=LH+1
   PH=(180.0*(NC+K)+Z(NM+1-I))/(-Z(NM-I)+Z(NM+1-I))
   TIM(250-M)=-(I-1+PH)*DT
   SN=(-1.0)**(NC+K)
   LEN(250-M)=SN*(Y(NM+1-I)+(Y(NM-I)-Y(NM+1-I))*PH)
30 CONTINUE
10 CONTINUE
MN=250-M-1
DO 50 I=1,LH
   LEN(I)=LEN(MN+I)
   TIM(I)=TIM(MN+I)
50 CONTINUE
49 RETURN
END

SUBROUTINE FRQRSPL(V,TIM,FREQ,RMDB,RPHS,NZ,T,RIDT)

REAL LEN
COMPLEX RIDT
COMPLEX DEL,V
DIMENSION V(L),TIM(L),FREQ(NZ),RMDB(NZ),RPHS(NZ),RIDT(NZ)
N=NZ-1
NM=N/2+1

L IS THE NUMBER OF FINGERS IN THE IDT
LEN IS THE LENGTH OF FINGERS + POLARITY
TIM IS POSITION IN THE TIME DOMAIN OF FINGERS IN MICRO-SECONDS
FREQ IS ARRAY OF FREQUENCY POINTS FOR FFT
RMDB IS AMPLITUDE RESPONSE IN DB
C RPHS IS THE PHASE RESPONSE IN DEGREES
C NZ IS NUMBER OF SPECIFIED POINTS TO BE EVALUATED
  PI=ACOS(-1.0)
  RMAX=0.0
C DO FAR FIELD INTEGRATION
  DO 10 I=1,NZ
    RIDT(I)=CMPLX(0.0,0.0)
    DO 20 J=1,L
      ANG=2*PI*FREQ(I)*(-TIM(J)-T/2.0)
      DEL=CMPLX(COS(ANG),SIN(ANG))
      RIDT(I)=RIDT(I)+DEL*V(J)
10 CONTINUE
RRL=REAL(RIDT(I))
RSM=AIMAG(RIDT(I))
RMDB(I)=SQRK(RRLL**2+RSM**2)
RPBS(I)=(ATAN(RSM/RRLL))*(180.0/PI)
IF (RRLL.LT.0.0.AND.RSM.LT.0.0) RPBS=RPBS-180.0
IF (RRLL.GT.0.0.AND.RSM.LT.0.0) RPBS=RPBS+180.0
10 CONTINUE
RMAX=RMDB(NM)
C OBTAIN AMPLITUDE IN DB
  DO 30 I=1,NZ
    RMDB(I)=20*ALOG10(RMDB(I)/RMAX)
    IF (RMDB(I).LT.-100.0) RMDB(I)=-100.0
    RPBS(I)=RPBS(I)-RPBS(NM)
30 CONTINUE
C* FORMAT (1H0,5X,*FREQ =*,F5.1,* MHZ*,5X,*MDB =*,F10.3,* DB*,5X,
C*PHS =*,F10.3,* DEG*)
RETURN
END

SUBROUTINE MODEL(LZ,LEN,TIM,B,X,T,F1,F2,VEL,C0,C1,C2, AR)
  COMPLEX AR
  REAL LEN,I1,I2,I3
DIMENSION LEN(LZ), TIM(LZ), B(LZ), X(LZ), C0(LZ), C1(LZ), C2(LZ),
CV(500, 3), AR(LZ)
P1=ACOS(-1.0)
AL= ALOG(2.0)
L=LZ-1
VMAX=0.0
DO 10 I=1, LZ
IF (VMAX.LT.ABS(LEN(I))) VMAX=ABS(LEN(I))
IF (VMAX.EQ.ABS(LEN(I))) IMAX=I
X(I)=TIM(I)*VEL
V(I,1)=0.0
V(I,2)=0.0
V(I,3)=0.0
10 CONTINUE
B(1)=(X(2)-X(1))/4.0
DO 20 I=2, L
B(I)=(X(I+1)-X(I-1))/8.0
20 CONTINUE
B(LZ)=(X(LZ)-X(L))/4.0
C
FINGER WIDTH AND POSITIONS HAVE BEEN FOUND.

LH=LZ/2
DO 30 I=1, LZ
C0(I)=LEN(I)
C1(I)=0.0
C2(I)=0.0
30 CONTINUE

CALL ELPOT(LZ, X, B, C0, C1, C2, V(IMAX, 2), IMAX, 2)
FTR=ABS(LEN(IMAX)/V(IMAX, 2))
DO 41 I=1, LZ
C0(I)=C0(I)*FTR
41 CONTINUE

DO 42 I=1, LZ
CALL ELPOT(LZ,X,B,C0,C1,C2,V(I,1),I,1)
CALL ELPOT(LZ,X,B,C0,C1,C2,V(I,2),I,2)
CALL ELPOT(LZ,X,B,C0,C1,C2,V(I,3),I,3)
42 CONTINUE
BI30=0.5*ALOG(2.0)-0.75
ORC=1.5
DO 55 M=1,3
39 DO 43 I=1,LZ
C0(I)=C0(I)+ORC*(LEN(I)-V(I,2))/AL
C1(I)=C1(I)+ORC*(V(I,3)-V(I,1))/(2.0*B(I))
C2(I)=C2(I)+ORC*(V(I,2)-(V(I,1)+V(I,3))/2.0)/BI30
43 CONTINUE
CALL ELPOT(LZ,X,B,C0,C1,C2,V(IMAX,2),IMAX,2)
FTR=ABS(LEN(IMAX)/V(IMAX,2))
DO 50 K=1,LZ
C0(K)=C0(K)*FTR
C1(K)=C1(K)*FTR
C2(K)=C2(K)*FTR
50 CONTINUE
DO 51 I=1,LZ
CALL ELPOT(LZ,X,B,C0,C1,C2,V(I,1),I,1)
CALL ELPOT(LZ,X,B,C0,C1,C2,V(I,2),I,2)
CALL ELPOT(LZ,X,B,C0,C1,C2,V(I,3),I,3)
51 CONTINUE
55 CONTINUE
.52 FORMAT (1H0,5X,15,7E15.4)

C COEFFICIENTS HAVE BEEN FOUND AND PRINTED.
W1=SQRT(2.0*PI*F1)
DO 70 I=1,LZ
X1=2.0*PI*F1*B(I)/VEL
CALL BESJ(X1,0,BJ01,0.0001,IER)
CALL BESJ(X1,1,BJ11,0.0001,IER)
ARR=W1*C0(I)*BJ01+C2(I)*W1*B(I)**2*(BJ01-BJ11/X1)
ARI=W1*C1(I)*B(I)*BJ11
AR(I)=CMPLX(ARR,ARI)
70 CONTINUE

C ACOUSTIC RESPONSE FOR EACH FINGER HAS BEEN FOUND.
RETURN
END

SUBROUTINE ELPOP(LZ,X,B,C0,C1,C2,VK,IK,IN)

REAL I1,I2,I3
DIMENSION X(LZ),B(LZ),C0(LZ),C1(LZ),C2(LZ)
VK=0.0
AL=ALOG(2.0)
DO 41 J=1,LZ
XR=X(IK)+(IN-2)*B(IK)-X(J)
IF (XR.EQ.0.0) 43,48
48 BX=(B(J)/XR)**2
ARG=1.0-BX
IF(ARG.LE.0.0) GO TO 45
47 BS=SQRT(1.0-BX)
40 I1=ALOG(2.0/(1.0+BS))
I2=BS-1.0
I3=0.5*(BX*I1+BS-1.0-0.5*BX)
IF (XR) 42,43,44
41 CONTINUE
RETURN
42 VK = VK + C0(J) * I1 - C1(J) * I2 + C2(J) * I3
GO TO 41

43 VK = VK + C0(J) * AL
GO TO 41.

44 VK = VK + C0(J) * I1 + C1(J) * I2 + C2(J) * I3
GO TO 41

50 FORMAT (1H0, 5X, I5, 7E15.5)
51 FORMAT (1H0, 5X, I5, 3E15.5)
45 BS = 0.0
GO TO 40

END

SUBROUTINE PHSCPS(N, Z)

DIMENSION Z(N)
DO 10 K = 1, N
GO TO 11

12 Z(K) = Z(K) - 360.0
11 IF (Z(K) .GT. +181.0) GO TO 12
GO TO 13

14 Z(K) = Z(K) + 360.0
13 IF (Z(K) .LE. -179.0) GO TO 14
10 CONTINUE
RETURN
END

SUBROUTINE PHSSTR(NZ, Y, Z)

DIMENSION Y(NZ), Z(NZ)
NH = NZ / 2
NM=NH+1
DO 10 I=1,NH
14  DPHS=Z(NM+I)-Z(NM-1+I)
    IF (DPHS.GT.+90.0) GO TO 11
    IF (DPHS.LT.-90.0) GO TO 12
    GO TO 13
11  Z(NM+I)=Z(NM+I)-180.0
    Y(NM+I)=-Y(NM+I)
    GO TO 14
12  Z(NM+I)=Z(NM+I)+180.0
    Y(NM+I)=-Y(NM+I)
    GO TO 14
13  DPHS=Z(NM-I)-Z(NM+1-I)
    IF (DPHS.GT.+90.0) GO TO 15
    IF (DPHS.LT.-90.0) GO TO 16
    GO TO 10
15  Z(NM-I)=Z(NM-I)-180.0
    Y(NM-I)=-Y(NM-I)
    GO TO 13
16  Z(NM-I)=Z(NM-I)+180.0
    Y(NM-I)=-Y(NM-I)
    GO TO 13
10  CONTINUE
    RETURN
END

SUBROUTINE RECT(PL,PA,RR,RI)
C
C    RECT CONVERTS (PL,PA) TO (RR,RI)
C    PL IS THE POLAR CO-ORDINATE LENGTH.
C    PA IS THE POLAR CO-ORDINATE ANGLE IN DEGREES.
C    RR IS THE RECTANGULAR CO-ORDINATE REAL PART.
C    RI IS THE RECTANGULAR CO-ORDINATE IMAGINARY PART.
C
PI =ACOS(-1.0)
RI=PL*SIN(PI*PA/180.0)
RR = PL*COS(PI*PA/180.0)
RETURN
END

SUBROUTINE POLAR(RR, RI, PL, PA)

C POLAR CONVERTS (RR, RI) TO (PL, PA)
C RR IS THE RECTANGULAR CO-ORDINATE REAL PART.
C RI IS THE RECTANGULAR CO-ORDINATE IMAGINARY PART.
C PL IS THE POLAR CO-ORDINATE LENGTH.
C PA IS THE POLAR CO-ORDINATE ANGLE IN DEGREES.
PI = ACOS(-1.0)
PL = SQRT(RR**2 + RI**2)
IF (RR.EQ.0.0) GO TO 10
PA = ATAN(RI/RR)
IF (RR.LE.0.0 .AND. RI.GE.0.0) PA = PA + PI
IF (RR.LE.0.0 .AND. RI.LE.0.0) PA = PA - PI
PA = PA * 180.0 / PI
GO TO 15
10 IF (RI) 11, 12, 13
11 PA = -90.0
   PL = RI
   GO TO 15
12 PA = 0.0
   PL = 0.0
   GO TO 15
13 PA = +90.0
   PL = RI
15 RETURN
END

SUBROUTINE LBLF(YL)
CALL LETTER (19, 1.0, 1.5, 7.50, 19PHASE SPECIFICATION)
CALL LETTER (23, 1.0, 1.5, 1.2, 23AMPLITUDE SPECIFICATION)
YL=5.75
NA = 0055006000601440102B
DO 10 I=1,10
NA=NA+1000000000000B
YL = YL-0.4
CONTINUE
CALL LBPH(9.75)
RETURN
END

SUBROUTINE LBLT(YL)
CALL LETTER (15,1 ,0.0,1.5,1.25,15HIDT APODIZATION)
CALL LETTER (31,1 ,0.0,1.5,7.5,31HIDT DEVIATION FROM LINEAR PHA
CSE).
NA = 006000560071B
YL = 5.75
DO 10 I=1,10
YL = YL - 0.4
NA = NA - 0001B
CONTINUE
CALL LBPH(9.75)
RETURN
END

SUBROUTINE LBPH(YPH)
DIMENSION Y(5), NB(5)
NB(1) = 00550051007000600040B
NB(2) = 0055007100500040B
NB(3) = 00600040B
NB(4) = 0053007100600040B
NB(5) = 00530061007000600040B
Y(1) = 7.75
Y(2) = 8.25
Y(3) = 8.75
\[ Y(4) = 9.25 \]
\[ Y(5) = 9.75 \]
NC=0060B
DO 20 K=1,5
\[ YB=Y(K)+0.1 \]
CONTINUE
RETURN
END

SUBROUTINE PLTB(N)
N=N
CALL PLOT (0.0,0.0,3)
CALL PLOT (0.0,10.5,2)
CALL PLOT (8.6,0.0,3)
CALL PLOT (8.6,10.5,2)
CALL PLOT (1.5,1.75,3)
CALL PLOT (7.5,1.75,2)
CALL PLOT (7.5,5.75,2)
CALL PLOT (1.5,5.75,2)
CALL PLOT (1.5,1.75,2)
CALL PLOT (1.5,7.75,3)
CALL PLOT (7.5,7.75,2)
CALL PLOT (7.5,9.75,2)
CALL PLOT (1.5,9.75,2)
CALL PLOT (1.5,7.75,2)
CALL PLOT (1.5,8.25,2)
CALL PLOT (7.5,8.25,2)
CALL PLOT (7.5,8.75,2)
CALL PLOT (1.5,8.75,2)
CALL PLOT (1.5,9.25,2)
CALL PLOT (7.5,9.25,2)
DO 200 I = 1,10
\[ XA = I*0.6+1.5 \]
CALL PLOT (XA,1.75,3)
CALL PLOT (XA,1.55,2)
CALL PLOT(XA, 8.85, 3)
CALL PLOT(XA, 8.65, 2)
CONTINUE
DO 201 I = 1, 10
YA = I*0.4+1.75
CALL PLOT(1.5, YA, 3)
CALL PLOT(1.4, YA, 2)
CONTINUE
CALL PLOT(1.5, 1.75, 3)
RETURN
END

SUBROUTINE PLTIDT(TIM, LEN, L)
DIMENSION TIM(L), LEN(L)
REAL LEN
CALL PLOT(0.0, 0.0, 3)
CALL PLOT(0.0, 10.5, 2)
CALL PLOT(8.5, 10.5, 3)
CALL PLOT(8.5, 0.0, 2)
CALL PLOT(1.5, 2.0, 3)
CALL PLOT(1.5, 8.0, 2)
CALL PLOT(2.0, 8.0, 2)
CALL PLOT(2.0, 2.0, 2)
CALL PLOT(1.5, 2.0, 2)
CALL PLOT(7.0, 2.0, 3)
CALL PLOT(7.0, 8.0, 2)
CALL PLOT(7.5, 8.0, 2)
CALL PLOT(7.5, 2.0, 2)
CALL PLOT(7.0, 2.0, 2)
DO 10 I = 1, L
X=4.5+2.0*LEN(I)
Y=((TIM(I)-TIM(1))/(TIM(L)-TIM(1)))*6.0+2.0
IF (LEN(I).LT.0.0) GO TO 11
CALL PLOT(2.0, Y, 3)
CALL PLOT(X, Y, 2)
GO TO 10
CALL PLOT(7.0, Y, 3)
CALL PLOT(X, Y, 2)
10 CONTINUE
CALL PLOT(12.0, 0.0, -3)
RETURN
END

SUBROUTINE PLTFFT(NZ, X, Y, 2)
DIMENSION X(NZ), Y(NZ), Z(NZ)
CALL PLTB(128)
CALL LBLT(5.75)
N=NZ-1
M=N-1
XS=SCALE=(X(NZ)-X(1))/6.0
YS=SCALE=0.5
ZS=SCALE=540.0
X0=-1.5*XS
Y0=-3.75*YS
Z0=-8.75*ZS
XMIN=X(1)
YMIN=-1.0
ZMIN=-540.0
XMAX=X(NZ)
YMAX=1.0
ZMAX=+540.0
CALL PLTIN(XSCALE, YSCALE, X0, Y0, XMIN, XMAX, YMIN, YMAX)
DO 10 I=1, N
CALL PLTLN(X(I), Y(I), X(I+1), Y(I+1))
10 CONTINUE
CALL PLTIN(XSCALE, ZSCALE, X0, Z0, XMIN, XMAX, ZMIN, ZMAX)
DO 20 I=1, N
CALL PLTLN(X(I), Z(I), X(I+1), Z(I+1))
20 CONTINUE
CALL PLOT(12.0, 0.0, -3)
RETURN
END

SUBROUTINE PLTFRQ(N,FREQ,MDB,PHASE)

DIMENSION FREQ(N),MDB(N),PHASE(N)
DF=(FREQ(N)-FREQ(1))/(N-1)
XSCLAE = (FREQ(N)-FREQ(1)-DF)/6.0
YSCLAE = 180.0
XO = -1.5 * XSCALAE+FRQ(1)
YO = -8.75 * YSCALE
XMIN = FREQ(1)-DF
YMIN = -180.1
XMAX = FREQ(N)
YMAX = +180.1
M=N-1
CALL PLTIN(XSCALE,YSCLAE,XO,YO,XMIN,XMAX,YMIN,YMAX)
DO 60 I=1,M
   CALL PLTIN(FREQ(I),PHASE(I),FREQ(I+1),PHASE(I+1))
CONTINUE
60
   YSCALE = 25.0
   YO = -5.75*YSCLAE
   YMIN = -100.1
   YMAX=+20.0
   CALL PLTIN(XSCALE,YSCLAE,XO,YO,XMIN,XMAX,YMIN,YMAX)
   DO 70 I=1,M
       CALL PLTIN(FREQ(I),MDB(I),FREQ(I+1),MDB(I+1))
   CONTINUE
   CALL PLTIN(N)
   CALL LBLF(5.75)
   CALL PLOT(12.0,0.0,-3)
RETURN
END
APPENDIX IV
SAWOPT

The following is a listing of the optimization programme SAWOPT. It is written in FORTRAN IV and operates under the NOS system for CDC-6000 computer systems. System libraries used are:

1. SSPLIB - scientific subroutines package.
2. PLOTVER - Versatec plotting library package.

Files TAPE1 and TAPE2 are replaced while the old files TAPE1 and TAPE2 are renamed TAPE10 and TAPE20.
SAWOPT. THIS PROGRAM DOES SUCCESSIVE OPTIMISATIONS OF SAW DESIGN. SUTHERS
USER
CHARGE
FTN.
SAWOPT
SAVE, INPUT=SAWOPT.
PROGRAM TST (INPUT, OUTPUT, TAPE5=INPUT, TAPE6=OUTPUT, TAPE1, TAPE2,
+ TAPE10, TAPE20)

COMPLEX FFTS, V, RIDT, AR
REAL MAG
REAL LEN
REAL MDB, PI, IMAG, LENGTH, RELBP, IMAGL
DIMENSION AR(513)
DIMENSION V(500), XP(500), B(500), TIM(500), LEN(500), FRSP(20, 3), FREQ(C513), MDB(513), PHASE(513), X(513), Y(513), Z(513), RMDB(513), RPHS(513),
CFFTS(513), RIDT(513), MAG(513), PHSE(513)
DIMENSION C0(500), C1(500), C2(500)
READ (10, 20) NZ, L, IG, T, DT, SPBD, FO, FC, VEL
IR=0
DO 10 I=1, NZ
READ (10, 30) FREQ(I), MDB(I), PHASE(I), RMDB(I), RPHS(I), MAG(I), PHSE(I)+I
IF (MDB(I).EQ.0.0.AND. IR.EQ.0) IR=I
10 CONTINUE
IR=IR+1
PI = ACOS (-1.0)
N=NZ-1
NH=N/2
NM=NH+1
C
PREDISTORTION OF SPECIFICATIONS.

DO 205 I=1, N
RMERR=MDB(I)-RMDB(I)
RPERR=PHASE(I)-RPHS(I)
IF (MDB(I).LT.-20.0) RPERR=0.0

121
MAG(I) = MAG(I) + 0.50 * RMERR
PHSE(I) = PHSE(I) + 0.50 * RPERR
AMP = 10.0 ** (MAG(I) / 20.0)
CALL RECT(AMP, PHSE(I), RRE, RIE)
FFTS(I) = CMPLX(RRE, RIE)
CONTINUE

CALL SAWFFT(FFTS, N, IG, X, Y, Z)
X(NZ) = 6.0
Y(NZ) = Y(1)
Z(NZ) = Z(1)
CALL PLTFFT(NZ, X, Y, Z)
FC = FO - SPBD / 2.0
DO 95 I = 1, NH
Z(NM+I) = Z(NM+I) - I * 350.0 * DT * (FO - FC)
Z(NM-I) = Z(NM-I) + I * 350.0 * DT * (FO - FC)
CONTINUE
CALL PHSSTR(NZ, Y, Z)
CALL IDT(NZ, Y, Z, FO, SPBD, L, TIM, L)
F1 = FREQ(NH - 2)
F2 = FREQ(NH + 2)
VEL = 3410
CALL MODEL(L, L, TIM, B, XP, T, F1, F2, VEL, C0, C1, C2, AR)
CALL FRQRSR(L, AR, TIM, FREQ, RMDB, RPS, NZ, T, RIDT)
WRITE (1,20) NZ,L,IG,T,DT,SPBD,FO,FC,VEL
RMAX=RMDB(IR)
DO 50 I=1,NZ
RMDB(I)=RMDB(I)-RMAX
WRITE (1,30) FREQ(I),MDB(I),PHASE(I),RMDB(I),RPHS(I),MAG(I),PHSE(I)
   +I
   IF (RMDB(I).LT.-100.0) RMDB(I)=-100.0
   CONTINUE
50 CONTINUE
20 FORMAT (15,6E10.5)
30 FORMAT (7E11.5)
WRITE (2,81) L
DO 70 I=1,L
XL=XP(I)-B(I)
XR=XP(I)+B(I)
WRITE (2,80) LEN(I),XL,XR
70 CONTINUE
80 FORMAT (3E20.10)
81 FORMAT (I5)
WRITE (6,152)
DO 150 I=1,N
WRITE (6,151) FREQ(I),MAG(I),MDB(I),RMDB(I),PHSE(I),PHASE(I)
   +RPHS(I)
150 CONTINUE
151 FORMAT (1H0,5X,F8.2,* MHZ*,F9.2,* DB*,F9.2,* DB*,F8.2,
   +* DEG*,F8.2,* DEG*,F8.2,* DEG*)
152 FORMAT (1H1,9X,*FREQ*,9X,*MAG*,9X,*MDB*,9X,*RMDB*,9X,*PHSE*,9X,*RPHS*)
CALL PLTFREQ(N,FREQ,RMDB,RPHS)
CALL PLOT (X(N),Y(N),999)
STOP
END
SUBROUTINE PHSCPS(N,Z)

DIMENSION Z(N)
DO 10 K=1,N
GO TO 11
12 Z(K)=Z(K)-360.0
11 IF (Z(K).GT.+181.0) GO TO 12
GO TO 13
14 /
13 IF (Z(K).LE.-179.0) GO TO 14
10 CONTINUE
RETURN
END

SUBROUTINE SAWFFT(FFT1,N,IG,X,Y,Z)
COMPLEX FFT1,FFT2,W,WK,WP
REAL IMAG
DIMENSION FFT1(N),IFFT(513,2),FFT2(513),X(N),Y(N),Z(N)
PI=ACOS(-1.0)
L=N

C C C C C
C C C C C
C C C C C
C C C C C
C FAST FOURIER TRANSFORM FOR SAW DESIGN
C FREQ IS ARRAY OF FREQUENCIES OF POINTS OF FFT
C MDB IS ARRAY OF DB OF INSERTION LOSS FOR EACH FREQUENCY
C PHASE IS ARRAY OF PHASES FOR EACH FREQUENCY
C IFFT IS ARRAY KEEPING TRACK OF EXPONENTS FOR MULTIPLIERS W(J) IN
C THE FFT PROCESS
C FFT1 AND FFT2 ARE COMPLEX VECTORS OF FFT PROCESS
C C C C C
DO 10 I=1,513
DO 10 J=1,2
10 IFFT(I,J)=0

C C C C C
C C C C C
C C C C C
C C C C C
C CALCULATE THE EXPONENT FOR THE MULTIPLIER FOR THE FFT PROCESS
C AREAL = COS(2.*PI/N)
C IMAG = SIN(2.*PI/N)
C W = CMPLX(AREAL,IMAG)
DO 70 M=1,IG
  L=L/2
  NH = N/2
  NM=NH+1
  DO 50 K=1,NH
    IFFFT(K,2) = IFFFT(2*K,1)
    IPFFT(K+NH,2)=IFFFT(K,2) +L
  50 C
  CALCULATE NEXT VECTOR IN FFT PROCESS
  DO 50 K=1,N
    WK = W**IFFFT(K,2)
    IF (IFFFT(K,2) .GT. IFFFT(K,1)) GO TO 61
    WP = WK*FFT1(K+L)
    FFT2(K) = FFT1(K)+WP
    GO TO 60
  61 WP = WK*FFT1(K)
    FFT2(K) = FFT1(K-L) + WP
  60 CONTINUE
  DO 70 I=1,N
    FFT1(I) = FFT2(I)
    IFFFT(I,1)=IFFFT(I,2)
  70 CONTINUE
  C
  PUT OUTPUT VECTOR (DFT) IN CORRECT ORDER.
  AMAX=0.0
  DO 82 K=1,N
    KI=0
  81 KI=KI+1
    IF (IFFFT(KI,1) .EQ. K-1) GO TO 80
    GO TO 81
  80 RR=REAL(FFT1(KI))
    RI=AIMAG(FFT1(KI))
    CALL POLAR (RR,RI,PL,PA)
    Y(K)=PL
    Z(K)=PA
    IF (Y(K) .GT. AMAX) AMAX=Y(K)
    X(K)=0.0*IFFFT(KI,1)/N
  82 CONTINUE
DO 83 I=1,N
  Y(I)=Y(I)/Amax
  CONTINUE
DO 93 K = 1,NH
  YS = Y(K)
  Y(K)=Y(K+NH)
  Y(K+NH)=YS
  ZS = Z(K)
  Z(K) = Z(K+NH)
  Z(K+NH) = ZS
CONTINUE
RETURN
END

SUBROUTINE IDT(NZ,Y,Z,FO,SPBD,LEN,TIM,LH)

REAL LEN
DIMENSION Y(NZ),Z(NZ),LEN(500),TIM(500)
NH=NZ/2
NM=NH+1
N=NZ-1
T=N/SPBD
DT=T/N

T IS THE IDT PERIOD IN MICRO-SECONDS
DT IS THE TIME INCREMENT OF THE FFT SAMPLES.
ADD IN THE LINEAR PHASE COMPONENTS OF THE FREQUENCY SHIFT.
ZM=Z(NM)
DO 40 I=1,NH
  Z(NM+I)=Z(NM+I)-ZM+(I*DT*FO*360.0)
  Z(NM-I)=Z(NM-I)-ZM-(I*DT*FO*350.0)
CONTINUE
L=0
M=0
LH=1
LEN(250)=Y(NM)
TIM(250)=0.0
DO 10 I=1,NH
IF (L.EQ.249) GO TO 21
C NP GIVES THE NUMBER OF FINGERS BETWEEN FFT POINTS.
NC=(Z(NM-1+I)+0.5)/180.0
PHASL=180.0*NC
NP=(Z(NM+I)-PHASL-0.5)/180.0
IF (NP.EQ.0) GO TO 21
C OBTAIN TIMES, LENGTHS AND POLARITIES OF FINGERS.
DO 20 K=1,NP
L=L+1
LH=LH+1
PH=(180.0*(NC+K)-Z(NM-1+I))/(Z(NM+I)-Z(NM-1+I))
TIM(250+L)=(I-1+PH)*DT
SN=(-1.0)**(NC+K)
LEN(250+L)=SN*(Y(NM-1+I)+(Y(NM+I)-Y(NM-1+I)))*PH
CONTINUE
20 IF (M.EQ.249) GO TO 10
C NP GIVES THE NUMBER OF FINGERS BETWEEN FFT POINTS.
NC=(-Z(NM+1-I)+0.5)/180.0
PHASL=180.0*NC
NP=(-Z(NM-I)-PHASL+0.5)/180.0
IF (NP.EQ.0) GO TO 10
C OBTAIN TIMES, LENGTHS AND POLARITIES OF FINGERS.
DO 30 K=1,NP
M=M+1
LH=LH+1
PH=(180.0*(NC+K)+Z(NM+1-I))/(-Z(NM-I)+Z(NM+1-I))
TIM(250-M)=-(I-1+PH)*DT
SN=(-1.0)**(NC+K)
LEN(250-M)=SN*(Y(NM+1-I)+(Y(NM-I)-Y(NM+1-I)))*PH
CONTINUE
30 CONTINUE
10 CONTINUE
MN=250-M-1
DO 50 I=1,LH
LEN(I)=LEN(MN+I)
50 CONTINUE
TIM(I)=TIM(MN+I)
50 CONTINUE
49 RETURN
END

SUBROUTINE FRQ_RSP(L, V, TIM, FREQ, RMDB, RPHS, NZ, T, RIDT)

REAL LEN
COMPLEX RIDT
COMPLEX DEL, V
DIMENSION V(L), TIM(L), FREQ(NZ), RMDB(NZ), RPHS(NZ), RIDT(NZ)
N=N-1
NM=N/2+1

C L IS THE NUMBER OF FINGERS IN THE IDT
C LEN IS THE LENGTH OF FINGERS + POLARITY
C TIM IS POSITION IN THE TIME DOMAIN OF FINGERS IN MICRO-SECONDS
C FREQ IS ARRAY OF FREQUENCY POINTS FOR FFT
C RMDB IS AMPLITUDE RESPONSE IN DB
C RPHS IS THE PHASE RESPONSE IN DEGREES
C NZ IS NUMBER OF SPECIFIED POINTS TO BE EVALUATED
PI=ACOS(-1.0)
RMAX=0.0

C DO FAR FIELD INTEGRATION
DO 10 I=1,NZ
10 RIDT(I)=CMPLX(0.0,0.0)
DO 20 J=1,L
ANG=2*PI*FREQ(I)*(-TIM(J)-T/2.0)
DEL=CMPLX(COS(ANG),SIN(ANG))
RIDT(I)=RIDT(I)+DEL*V(J)
20 CONTINUE
RSRL=REAL(RIDT(I))
RSIM=AIMAG(RIDT(I))
RMDB(I)=SQRT(RSRL**2+RSIM**2)
RPHS(I)=(ATAN(RSIM/RSRL))*(180.0/PI)
IF (RSRL.LT.0.0.AND.RSIM.LT.0.0) RPHS=RPHS-180.0
IF (RSRL.GT.0.0.AND.RSIM.LT.0.0) RPHS=RPHS+180.0
CONTINUE
RMAX=RMDB(NM)
C OBTAIN AMPLITUDE IN DB.
DO 30 I=1,NZ
RMDB(I)=20*ALOG10(RMDB(I)/RMAX)
IF (RMDB(I).LT.-100.0) RMDB(I)=-100.0
RPHS(I)=RPHS(I)-RPHS(NM)
30 CONTINUE
C FORMAT (1H0,5X,*FREQ =*,F5.1,* MHZ*,5X,*MDB =*,F10.3,* DB*,5X,
C*PHS =*,F10.3,* DEG*)
RETURN
END

SUBROUTINE PHSSTR(NZ,Y,Z)
DIMENSION Y(NZ),Z(NZ)
NH=NZ/2
NM=NH+1
DO 10 I=1,NH
DPHS=Z(NM+I)-Z(NM-1+I)
IF (DPHS.GT.+90.0) GO TO 11
IF (DPHS.LT.-90.0) GO TO 12
GO TO 13
11 Z(NM+I)=Z(NM+I)-180.0
Y(NM+I)=-Y(NM+I)
GO TO 14
12 Z(NM+I)=Z(NM+I)+180.0
Y(NM+I)=-Y(NM+I)
GO TO 14
13 DPHS=Z(NM-I)-Z(NM+1-I)
IF (DPHS.GT.+90.0) GO TO 15
IF (DPHS.LT.-90.0) GO TO 16
GO TO 10
15 Z(NM-I)=Z(NM-I)-180.0
Y(NM-I) = -Y(NM-I)
GO TO 13
16 Z(NM-I) = Z(NM-I) + 180.0
Y(NM-I) = -Y(NM-I)
GO TO 13
10 CONTINUE
RETURN
END

SUBROUTINE RECT(PL, PA, RR, RI)
C
RECT CONVERTS (PL, PA) TO (RR, RI)
PL IS THE POLAR CO-ORDINATE LENGTH.
PA IS THE POLAR CO-ORDINATE ANGLE IN DEGREES.
RR IS THE RECTANGULAR CO-ORDINATE REAL PART.
RI IS THE RECTANGULAR CO-ORDINATE IMAGINARY PART.
PI = ACOS(-1.0)
RI = PL * SIN(PI * PA / 180.0)
RR = PL * COS(PI * PA / 180.0)
RETURN
END

SUBROUTINE POLAR(RR, RI, PL, PA)
C
POLAR CONVERTS (RR, RI) TO (PL, PA)
RR IS THE RECTANGULAR CO-ORDINATE REAL PART.
RI IS THE RECTANGULAR CO-ORDINATE IMAGINARY PART.
PL IS THE POLAR CO-ORDINATE LENGTH.
PA IS THE POLAR CO-ORDINATE ANGLE IN DEGREES.
PI = ACOS(-1.0)
PL = SQRT(RR**2 + RI**2)
IF (RR .EQ. 0.0) GO TO 10
PA = ATAN(RI / RR)
IF (RR .LT. 0.0 AND RI .GE. 0.0) PA = PA + PI
IF (RR.LE.0.0.AND.RI.LE.0.0) PA = PA-PI
PA = PA*180.0/PI
GO TO 15
10 IF (RI) 11, 12, 13
11 PA=-90.0
PL=RI
GO TO 15
12 PA=0.0
PL=0.0
GO TO 15
13 PA=+90.0
PL=RI
15 RETURN
END

SUBROUTINE MODEL(LZ,LEN,TIM,B,X,T,F1,F2,VEL,CO,C1,C2, AR)

COMPLEX AR
REAL LEN,I1,12,13
DIMENSION LEN(LZ),TIM(LZ),B(LZ),X(LZ),CO(LZ),C1(LZ),C2(LZ),
CV(500,3),AR(LZ)
PI=A COS(-1.0)
AL=A LOG(2.0)
L=LZ-1
VMAX=0.0
DO 10 I=1,LZ
IF (VMAX.LT.ABS(LEN(I))) VMAX=ABS(LEN(I))
IF (VMAX.EQ.ABS(LEN(I))) IMAX=I
X(I)=TIM(I)*VEL
V(I,1)=0.0
V(I,2)=0.0
V(I,3)=0.0
CONTINUE
10 B(1)=(X(2)-X(1))/4.0
DO 20 I=2,L

B(I) = (X(I+1) - X(I-1))/8.0
CONTINUE
B(LZ) = (X(LZ) - X(L))/4.0

C
FINGER WIDTH AND POSITIONS HAVE BEEN FOUND.

LH = LZ/2
DO 30 I = 1, LZ
C0(I) = LEN(I)
C1(I) = 0.0
C2(I) = 0.0
30 CONTINUE

CALL ELPO(T(LZ, X, B, C0, C1, C2, V(IMAX, 2), IMAX, 2)
FTR = ABS(LEN(IMAX)/V(IMAX, 2))
DO 41 I = 1, LZ
C0(I) = C0(I) * FTR
41 CONTINUE

DO 42 I = 1, LZ
CALL ELPO(T(LZ, X, B, C0, C1, C2, V(I, 1), I, 1)
CALL ELPO(T(LZ, X, B, C0, C1, C2, V(I, 2), I, 2)
CALL ELPO(T(LZ, X, B, C0, C1, C2, V(I, 3), I, 3)
42 CONTINUE
BI30 = 0.5 * ALOG(2.0) - 0.75
ORC = 1.5

DO 55 M = 1, 3
43 CONTINUE
CALL ELPOT(LZ,X,B,C0,C1,C2,V(IMAX,2),IMAX,2)
FTR=ABS(LEN(IMAX)/V(IMAX,2))
DO 50 K=1,LZ
C0(K)=C0(K)*FTR
C1(K)=C1(K)*FTR
C2(K)=C2(K)*FTR
50 CONTINUE

DO 51 I=1,LZ
CALL ELPOT(LZ,X,B,C0,C1,C2,V(I,1),I,1)
CALL ELPOT(LZ,X,B,C0,C1,C2,V(I,2),I,2)
CALL ELPOT(LZ,X,B,C0,C1,C2,V(I,3),I,3)
51 CONTINUE

55 CONTINUE

52 FORMAT (1H0,5X,I5,7E15.4)

C

COEFFICIENTS HAVE BEEN FOUND AND PRINTED.

W1=SQRT(2.0*PI*F1)
DO 70 I=1,LZ
X1=2.0*PI*F1*B(I)/VEL
CALL BESJ(X1,0,BJ01,0.0001,IER)
CALL BESJ(X1,1,BJ11,0.0001,IER)
ARR=W1*C0(I)*BJ01+C2(I)*W1*B(I)**2*(BJ01-BJ11/X1)
ARI=W1*C1(I)*B(I)*BJ11
AR(I)=COMPLX(ARR,ARI)
70 CONTINUE

C

ACOUSTIC RESPONSE FOR EACH FINGER HAS BEEN FOUND.

RETURN
END
SUBROUTINE ELPOT(LZ,X,B,C0,C1,C2,VK,IK,IN)

REAL I1,I2,I3
DIMENSION X(LZ),B(LZ),C0(LZ),C1(LZ),C2(LZ)
VK=0.0
AL=ALOG(2.0)
DO 41 J=1,LZ
XR=X(IK)+(IN-2)*B(IK)-X(J)
IF (XR.EQ.0.0) 43,48
48 BX=(B(J)/XR)**2
ARG=1.0-BX
IF(ARG.LE.0.0) GO TO 45
BS=SQR(T(1.0-BX))
40 I1=ALOG(2.0/(1.0+BS))
I2=BS-1.0
I3=0.5*(BX*I1+BS-1.0-0.5*BX)
IF (XR) 42,43,44
41 CONTINUE
RETURN

42 VK =VK +C0(J)*I1-C1(J)*I2+C2(J)*I3
GO TO 41

43 VK=VK+C0(J)*AL
GO TO 41

44 VK = VK +C0(J)*I1+C1(J)*I2+C2(J)*I3
GO TO 41

50 FORMAT (1H0,5X,I5,7E15.5)
51 FORMAT (1H0,5X,I5,3E15.5)
45 BS=0.0
GO TO 40

END
SUBROUTINE LBLF(YL)
CALL LETTER (19,.1,.0,1.5,7.50,19HPHASE SPECIFICATION)
CALL LETTER (23,.1,.0,1.5,1.2,23HAMPLITUDE SPECIFICATION)
YL=5.75
NA = 00550060006001440102B
DO 10 I=1,10
NA=NA+100000000000B
YL = YL-0.4
10 CONTINUE
CALL LBPH(9.75)
RETURN
END

SUBROUTINE LBLT(YL)
CALL LETTER (15,.1 ,0.0 ,1.5,1.25,15HIDT APODIZATION)
CALL LETTER (31,.1 ,0.0 ,1.5,7.5,31HIDT DEVIATION FROM LINEAR PHA
CSE)
NA = 006000560071B
YL = 5.75
DO 10 I=1,10
YL = YL - 0.4
10 NA = NA - 0001B
CONTINUE
CALL LBPH(9.75)
RETURN
END

SUBROUTINE LBPH(YPH)
DIMENSION Y(5), NB(5)
NB(1) = 00550061007000600040B
NB(2) = 0055007100600040B
NB(3) = 00600040B
NB(4) = 0053007100600040B
NB(5) = 00530061007000600040B
Y(1) = 7.75
Y(2) = 3.25
Y(3) = 8.75
Y(4) = 9.25
Y(5) = 9.75
NC=0060B
DO 20 K=1,5
YB=Y(K)+0.1
20 CONTINUE
RETURN
END

SUBROUTINE PLTB(N)
N=N
CALL PLOT (0.0,0.0,0,3)
CALL PLOT (0.0,0.1,0,5,2)
CALL PLOT (8.6,0.0,0,3)
CALL PLOT (8.6,0.1,0,5,2)
CALL PLOT (1.5,0.1,7,5,3)
CALL PLOT (7.5,0.1,7,5,2)
CALL PLOT (7.5,0.5,7,5,2)
CALL PLOT (1.5,0.5,7,5,2)
CALL PLOT (1.5,0.1,7,5,2)
CALL PLOT (1.5,0.7,7,5,3)
CALL PLOT (7.5,0.7,7,5,2)
CALL PLOT (7.5,0.9,7,5,2)
CALL PLOT (1.5,0.9,7,5,2)
CALL PLOT (7.5,0.7,7,5,2)
CALL PLOT (7.5,0.8,2,5,2)
CALL PLOT (7.5,0.2,2,5,2)
CALL PLOT (7.5,0.8,7,5,2)
CALL PLOT (1.5,0.8,7,5,2)
CALL PLOT (1.5,0.9,2,5,2)
CALL PLOT (7.5,0.9,2,5,2)
DO 200 I = 1,10
XA = I*0.6+1.5
CALL PLOT (XA,1.75,3)
CALL PLOT (XA,1.55,2)
CALL PLOT (XA,8.85,3)
CALL PLOT (XA,8.65,2)
200 CONTINUE
DO 201 I = 1,10
YA = I*0.4+1.75
CALL PLOT (1.5,YA,3)
CALL PLOT (1.4,YA,2)
201 CONTINUE
CALL PLOT (1.5,1.75,3).
RETURN
END

SUBROUTINE PLTIDT(TIM,LEN,L)
DIMENSION TIM(L),LEN(L)
REAL LEN
CALL PLOT(0.0,0.0,0,3)
CALL PLOT(0.0,10.5,2)
CALL PLOT(8.5,10.5,3)
CALL PLOT(8.5,0.0,2)
CALL PLOT(1.5,2.0,3)
CALL PLOT(1.5,8.0,2)
CALL PLOT(2.0,8.0,2)
CALL PLOT(2.0,2.0,2)
CALL PLOT(1.5,2.0,2)
CALL PLOT(7.0,2.0,3)
CALL PLOT(7.0,8.0,2)
CALL PLOT(7.5,8.0,2)
CALL PLOT(7.5,2.0,2)
CALL PLOT(7.0,2.0,2)
DO 10 I=1,L
X=4.5+2.0*LEN(I)
Y = ((TIM(I) - TIM(1)) / (TIM(L) - TIM(1))) * 6.0 + 2.0
IF (LEN(I).LT.0.0) GO TO 11
CALL PLOT(2.0, Y, 3)
CALL PLOT(X, Y, 2)
GO TO 10
CALL PLOT(7.0, Y, 3)
CALL PLOT(X, Y, 2)
CONTINUE
CALL PLOT(12.0, 0, 0, -3)
RETURN
END

SUBROUTINE PLTFPT(NZ, X, Y, Z)
DIMENSION X(NZ), Y(NZ), Z(NZ)
CALL PLTB(128)
CALL LBLT(5.75)
N = NZ - 1
M = N - 1
XSCALE = (X(NZ) - X(1)) / 6.0
YSCALE = 0.5
ZSCALE = 540.0
XO = -1.5 * XSCALE
YO = -3.75 * YSCALE
ZO = -8.75 * ZSCALE
XMIN = X(1)
YMIN = -1.0
ZMIN = -540.0
XMAX = X(NZ)
YMAX = 1.0
ZMAX = +540.0
CALL PLTIN(XSCALE, YSCALE, XO, YO, XMIN, XMAX, YMIN, YMAX)
DO 10 I = 1, N
   CALL PLTIN(X(I), Y(I), X(I+1), Y(I+1))
CONTINUE
CALL PLTIN(XSCALE, ZSCALE, XO, ZO, XMIN, XMAX, ZMIN, ZMAX)

10
18
DO 20 I=1,N
CALL PLTIN(X(I),Z(I),X(I+1),Z(I+1))
CONTINUE
CALL PLOT(12.0,0.0,-3)
RETURN
END

SUBROUTINE PLTFREQ(N,FREQ,DB,DEG)
DIMENSION FREQ(N),DB(N),DEG(N)
DF=(FREQ(N)-FREQ(1))/(N-1)
XSCLAE=(FREQ(N)-FREQ(1)-DF)/6.0
YSCALE=180.0
XO=-1.5*XSCLAE+FREQ(1)
YO=-0.75*YSCALE
XMIN=FREQ(1)-DF
YMIN=-180.1
XMAX=FREQ(N)
YMAX=+180.1
M=N-1
CALL PLTIN(XSCALE,YSCALE,XO,YO,XMIN,XMAX,YMIN,YMAX)
DO 60 I=1,M
CALL PLTIN(FREQ(I),DEG(I),FREQ(I+1),DEG(I+1))
CONTINUE
YSCALE=25.0
YO=-5.75*YSCALE
YMIN=-100.1
YMAX=+100.0
CALL PLTIN(XSCALE,YSCALE,XO,YO,XMIN,XMAX,YMIN,YMAX)
DO 70 I=1,M
DBA=DB(I)
DBB=DB(I+1)
IF (DBA.LT.-100.0) DBA=-100.0
IF (DBB.LT.-100.0) DBB=-100.0
CALL PLTIN(FREQ(I),DBA,FREQ(I+1),DBB)
APPENDIX V

TAPEJOB

The following is a listing of the programme TAPEJOB which generates the cutting instructions in the format requires by the mask cutting facility of Waterloo. It is written in FORTRAN IV and operates under the NOS system for CDC-5000 computer systems. System libraries used are:

(1) PLOTVER - Versatec plotting library package
TAPE8 is the information from TAPE2 or TAPE20 that results from SAWOPT or SAWINIT.

TAPE7 is the cutting instruction file.
TAPEJOB.
USER("SUTHERS")
CHARGE("SUTHERS")

SAVE,INPUT=TAPEJOB.

PROGRAM TST (INPUT,OUTPUT,TAPE5=INPUT,TAPE6=OUTPUT,TAPE8,TAPE7)
DIMENSION XL(500),XR(500),Y(500),XA(500),XB(500),XC(500),XD(500)
READ (8,10) N
I=63B
WRITE (7,35) I
FORMAT (A6)
WRITE (7,31)
WRITE (7,32)
WRITE (7,34)

YU=+16000.0
YL=-16000.0
DO 20 I=1,N
READ (8,30) Y(I),XL(I),XR(I)
B=0.0
XA(I)=8.0*(XL(I)-B)+0.5
XB(I)=8.0*(XL(I)+B)+0.5
XC(I)=8.0*(XR(I)-B)+0.5
XD(I)=8.0*(XR(I)+B)+0.5
Y(I)=Y(I)*15500.0+0.5
20 CONTINUE

CUT PADS AND BENCH-MARKS.

CALL PLTCUT(3,XA(I),YL)
CALL PLTCUT(2,XD(N),YL)
CALL PLTCUT(2,XD(N),-20000.0)
XP=XA(I)+4000.0
CALL PLTCUT(2,XP,-20000.0)
CALL PLTCUT(2,XP,-24000.0)
XP=XP+4000.0
CALL PLTCUT(2, XP, -24000.0)
CALL PLTCUT(2, XP, -32000.0)
CALL PLTCUT(2, XA(1), -32000.0)
CALL PLTCUT(2, XA(1), -16000.0)
CALL PLTCUT(3, XD(N), +16000.0)
CALL PLTCUT(2, XA(1), +16000.0)
XP = XA(1) - 4000.0
CALL PLTCUT(2, XP, -16000.0)
CALL PLTCUT(2, XP, -32000.0)
XP = XP - 8000.0
CALL PLTCUT(2, XP, -32000.0)
CALL PLTCUT(2, XP, -24000.0)
XP = XP + 4000.0
CALL PLTCUT(2, XP, -24000.0)
CALL PLTCUT(2, XP, -16000.0)
XP = XA(1) - 4000.0
CALL PLTCUT(2, XP, +16000.0)
CALL PLTCUT(2, XP, +20000.0)
XP = XD(N) - 4000.0
CALL PLTCUT(2, XP, +20000.0)
CALL PLTCUT(2, XD(N), +16000.0)
CALL PLTCUT(3, +20.0, -24000.0)
CALL PLTCUT(2, +20.0, -32000.0)
CALL PLTCUT(2, -20.0, -32000.0)
CALL PLTCUT(2, -20.0, -24000.0)
CALL PLTCUT(2, +20.0, -24000.0)
CALL PLTCUT(3, +4000.0, -27980.0)
CALL PLTCUT(2, +4000.0, -28020.0)
CALL PLTCUT(2, -4000.0, -28020.0)
CALL PLTCUT(2, -4000.0, -27980.0)
CALL PLTCUT(2, +4000.0, -27980.0)
CALL PLTCUT(3, +20.0, +24000.0)
CALL PLTCUT(2, +20.0, +32000.0)
CALL PLTCUT(2, -20.0, +32000.0)
CALL PLTCUT(2, -20.0, +24000.0)
CALL PLTCUT(2, +20.0, +24000.0)
CALL PLTCUT(3, +4000.0, +27980.0)
CALL PLTCUT(2, +4000.0, +28020.0)
CALL PLTCUT(2, -4000.0, +28020.0)
CALL PLTCUT(2, -4000.0, +27980.0)
CALL PLTCUT(2, +4000.0, +27980.0)

C PADS AND BENCH-MARKS ARE CUT.

C

GENERATE LINE-CUT INSTRUCTIONS.

DO 50 I=1, N
CALL PLTCUT(3, XA(I), YL)
CALL PLTCUT(2, XA(I), YU)
CALL PLTCUT(3, XD(I), YU)
CALL PLTCUT(2, XD(I), YL)
50 CONTINUE

C

GENERATE APODISATION-CUT INSTRUCTIONS.

CALL NEWPEN(5)
DO 60 I=1, N
  IF (Y(I) .LT. 0.0) GO TO 60
  BL = 4.0 * (XC(I) - XA(I))
  BT = 8.0 * BL
  IF (Y(I) .GT. BT) YT = Y(I) - (2.0 * BL)
  IF (Y(I) .LE. BT) YT = Y(I) - 2.0 * BL * (+Y(I)/(BL*4.0) - (+Y(I)/(8.0*8.0)))
  CALL PLTCUT(3, XA(I), YT)
  CALL PLTCUT(2, XD(I), YT)
  Y(I) = Y(I) + 500.0
60 CONTINUE

DO 70 I=1, N
  IF (Y(I) .LT. 0.0) GO TO 70
  BL = 4.0 * (XC(I) - XA(I))
  BT = 8.0 * BL
IF (Y(I).GT.BT) Y = Y(I) - (2.0*BL)
IF (Y(I).LE.BT) Y = Y(I) - 2.0*BL* (+Y(I)/(BL*4.0) - (+Y(I)/(BL*8.0))

CALL PLTCUT(3,XA(I),Y)
CALL PLTCUT(2,XD(I),Y)
Y(I) = Y(I) - 500.0

70 CONTINUE

DO 80 I = 1,N
IF (Y(I).GT.0.0) GO TO 80
BL = 4.0*(XC(I) - XA(I))
BT = -8.0*BL
IF (Y(I).LT.BT) Y = Y(I) + (2.0*BL)
IF (Y(I).GE.BT) Y = Y(I) + 2.0*BL* (-Y(I)/(BL*4.0) - (-Y(I)/(BL*8.0))

CALL PLTCUT(3,XA(I),Y)
CALL PLTCUT(2,XD(I),Y)
Y(I) = Y(I) - 500.0

80 CONTINUE

DO 90 I = 1,N
IF (Y(I).GT.0.0) GO TO 90
BL = 4.0*(XC(I) - XA(I))
BT = -8.0*BL
IF (Y(I).LT.BT) Y = Y(I) + (2.0*BL)
IF (Y(I).GE.BT) Y = Y(I) + 2.0*BL* (-Y(I)/(BL*4.0) - (-Y(I)/(BL*8.0))

CALL PLTCUT(3,XA(I),Y)
CALL PLTCUT(2,XD(I),Y)
Y(I) = Y(I) + 500.0

90 CONTINUE

WRITE (7,100)
CALL PLOT(0.0,0.0,999)

10 FORMAT (I5)
30 FORMAT (3E20.10)
31 FORMAT (*D90*)
32 FORMAT (*G1*)
33 FORMAT (*D1*)
34 FORMAT (*D2*)
100 FORMAT (*M30*)

STOP
END

SUBROUTINE PLTCUT(M,XD,YD)

IX=IFIX(XD)
IY=IFIX(YD)
YP=(YD+34000.0)/6400.0
XP=(XD+60000.0)/6400.0
CALL PLOT(XP,YP,M)
L=M-2
IF (L) 10,20,30
10 WRITE (7,1) IX,IY
RETURN
20 WRITE (7,2) IX,IY
RETURN
30 WRITE (7,3) IX,IY
RETURN

1 FORMAT (*X*,I7.7,*Y*,I7.7)
2 FORMAT (*X*,I7.7,*Y*,I7.7,*D1*)
3 FORMAT (*X*,I7.7,*Y*,I7.7,*D2*)

END
APPENDIX VI

FABRICATION OF FILTERS

Once an acceptable design for a receiving and a transmitting transducer had been obtained, the programme TAPEJOB generated the cutting instructions in the proper format and stored these instructions as a file on the system mass storage. These files were accessed through a facility of the Canadian Institute of Metal Working which translated the instructions onto paper tape. The paper tape was then taken to the University of Waterloo where the Rubylith master mask was cut for each transducer at 80x scale.

After peeling the Rubylith, the masks were taken to Hughes-Owens Reproduction Department in Hamilton where a 4:1 reduction was done of each mask giving a 20x negative transparency.

These transparencies were then assembled along with the multistrip coupler on a 50 cm x 25 cm master filter mask here at McMaster. This mask was then reduced 20:1 to give the final positive mask on a Kodak 5 cm x 5 cm high resolution plate. The camera system (in the McMaster Electrical Engineering Department) is an HLC Microkon 1600 Microphotography Camera system that has a resolution at 200 line pairs/mm, and a 2.5 cm square image area.

Having produced the final positive mask, it now remains to prepare the substrate. The lithium niobate (LiNbO$_3$) substrates which are 2.5 cm in the z-direction and 1.25 cm in the x-direction are
prepared as follows:

(1) scrub in trichloroethylene using cotton swabs for 2 minutes.
(2) rinse with trichloroethylene.
(3) scrub in acetone with cotton swabs for 2 minutes.
(4) rinse with acetone.
(5) scrub in distilled water with cotton swabs for 2 minutes.
(6) rinse with distilled water.
(7) place on vacuum chuck of Headway Research Inc. Model EC101 spinner, spin at 6000 rpm and rinse substrates with pure filtered methyl- or ethyl-alcohol for two minutes.
(8) place in clean-covered petri dish and bake in oven at 85°C for 15 minutes.
(9) inspect under microscope for surface contamination and repeat steps (1) through (8) if necessary (NOTE: all the above steps are done on a clean air bench in as dustfree an environment as is possible).
(10) if the substrate(s) are acceptably clean then they are placed in an Edwards Model 12E3/1670 coating unit and are subjected to ionic (argon) cleaning at as low an atmosphere as is required to maintain ionization for 15 minutes.
(11) then approximately 3000 to 5000 Å of aluminum is evaporated onto the substrate surface using tungsten filament sources. The deposition rate is not critical since aluminum adheres well to lithium niobate.
(12) allow the coated substrates to thoroughly cool in the evaporator
(4 hrs), then remove the substrates via clean covered petri dishes back to the clean air bench with the Headway spinner.

(13) place metallized substrate on vacuum chuck and spin at 6000 rpm and rinse surface with pure filtered methyl- or ethyl-alcohol and spin dry for 2 minutes.

(14) stop spinner and cover surface with filtered Shipley AZ-1350B positive working photo resist, then spin at 6000 rpm for 2 minutes.

(15) place resist covered substrates in petri dish and bake at 85°C for 15 minutes.

(16) the substrates are now ready to be exposed. Place final mask with the emulsion side down on top of the resist covered substrate using the 2.5 cm x 1.25 cm master mask outline for alignment with substrate edges.

(17) expose to UV light source for 1 minute.

(18) place exposed substrate in Shipley AZ developer and observe development (without agitation) until all exposed resist has been dissolved.

(19) rinse substrate with distilled water.

(20) place substrate in acid etch composed of 25 parts phosphoric acid \( \text{H}_3\text{PO}_4 \), 5 parts acetic acid \( \text{C}_2\text{H}_4\text{O}_2 \) and 1 part nitric acid \( \text{HNO}_3 \). Observe etching through microscope. Remove when etching is complete.

(21) rinse substrate with water and then remove unexposed photo resist with acetone.
(22) Rinse with distilled water and blow dry with dry nitrogen.
(23) Using RTV silicone based adhesive, glue substrate to 2.5 cm x 1.25 cm Tekform butterfly package.
(24) Using Kulicke and Soffa model 472 ultra-sonic bonder, bond rail connections to appropriate pin connections of case.
(25) Use RTV as acoustic absorber at ends of acoustic paths.
(26) Test device.
REFERENCES


