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SECOND ORDER EFFECTS

IN A

CODED SURFACE ACOUSTIC WAVE LABEL SYSTEM

by

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SYNOPSIS

A system for the remote interrogation of objects using passive electronic labels has been under investigation in the Electrical Engineering Department of the University of Adelaide for some time.

The interrogation system operates by transmitting a short burst of radio frequency energy, some of which is absorbed by a passive label in the vicinity of the transmitter. A short time later the label re-transmits a small amount of energy in the form of binary coded pulses. This weak signal is processed by a receiver to extract the binary code which uniquely identifies the label. The system is based on binary coded surface acoustic wave delay lines used in the passive labels.

Various delay lines operating at 50 megahertz have previously been constructed and tested in the Department. Although a complete interrogation system was not built for these delay lines, much valuable information was obtained which may be applied to binary coded delay lines in general. In particular, it was found that the binary code generated by a delay line may, for some codes, be corrupted by a "second order" response of the delay line. This aspect is undesirable for the correct interrogation of labels.

Recently an interrogation system and passive labels operating at 200 megahertz have been built. It was considered that the second order responses of some delay lines may be significant enough to interfere with the desired operation of the system. This thesis represents the author's investigation of second order effects caused by electro-acoustic regeneration in the 200 megahertz system.

The thesis describes the basis for two computer models of the system which were developed by the author to examine system operation with delay lines containing varied codes.

The first model uses a frequency domain approach where frequency response functions are obtained for each section of the system. The product of these functions gives the total system frequency response which is transformed to the time domain using a Fast Fourier Transform. The resultant time functions (the total system time response, the second order response and the receiver outputs) are plotted. This model is designed specifically to simulate the 200 megahertz 20-bit system.

The second model simulates the system in the time domain. The total system time response is derived by the convolution of time responses for each section of the system. A method has been developed for simplifying the calculations required for multiple convolutions. The resultant computer model is much faster than the frequency

domain model and is used interactively to provide receiver outputs. The second order response can be made available readily. The time domain model provides a fast and convenient method of evaluating the performance of various coded delay lines.

A comparison is made between the frequency domain model, the time domain model and results obtained from the 200 megahertz system in laboratory investigations.

Methods are proposed to minimise both second order responses in delay lines and intersymbol interference between code bits caused by narrow system bandwidth. The proposals are tested using the computer models with encouraging results. The theory for the time domain model is extended to suggest some consequences of second order responses in a phase-encoded system.

Finally, current photo-fabrication methods are briefly examined for their suitability for delay line production in view of proposals for minimising second order responses.

PREFACE

This thesis contains no material which has been accepted for the award of any other degree or diploma in any University. To the best of the author's knowledge and belief, this thesis contains no material previously published or written by another person, except where due reference is made in the text.

(David M. Green)

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GLOSSARY OF COMMON SYMBOLS

N	Number of fingers in an interdigital transducer (IDT)
W	Finger overlap width in an IDT (also referred to as transducer aperture)
L	Spacing between centres of adjacent fingers of an IDT
v	SAW propagation velocity
Δ	SAW piezoelectric coupling coefficient (also referred to as $\Delta v/v$)
ε_o	Permittivity of free space
ε_p	Effective permittivity of SAW substrate $= (\varepsilon_{xx}\varepsilon_{zz} - \varepsilon^2_{xz})^{\frac{1}{2}}$ for x,y,z axes system
ω	Radian frequency
β	Propagation constant (or wave number) $= \omega/v$
f_o	fundamental frequency of IDT operation (hence $\omega_o = 2\pi f_o$)
λ	SAW wavelength on substrate $(= 2\pi/\beta)$
*	convolution operator <u>or</u> complex conjugate (depending on context)

CHAPTER 1

THE SURFACE ACOUSTIC WAVE DELAY LINE

1.0 INTRODUCTION

Elastic waves which propagate along the surface of a solid were first noted by Lord Rayleigh in 1885 while studying the propagation of earthquake waves. Although the study of these elastic waves, or surface acoustic waves, proceeded over the years following Rayleigh's work, practical use was not realised until the early 1960s. At this time the interdigital transducer[1] was proposed by several researchers for producing surface acoustic waves on piezo-electric substrates. The interdigital transducer still remains as the basis for many surface acoustic wave (SAW) devices including filters, delay lines and oscillators.

Since the introduction of the interdigital transducer, a wide range of SAW devices using various substrates and semiconductor technology have been produced[2,3]. These devices range from commercially available television I.F. filters to convolvers, correlators and Fourier transformers used in applications including signal processing and spread-spectrum communications.

The history of the development of surface acoustic waves and their applications has been widely documented and

summarized by Burgess[4]. Likewise, the mathematical analysis of SAW propagation is not essential to the development of this thesis and has also been well documented.

The theory and analysis in this thesis is based on the operation of the interdigital transducer and its admittance characteristics as proposed by Burgess and Cole[5,6]. The theory is used as the basis to perform analysis of coded SAW delay lines operating in a 200 megahertz 20-bit coded label system.

1.1 THE INTERDIGITAL TRANSDUCER

On piezo-electric substrates, a common method of producing surface acoustic waves is to apply an electrical signal to metal electrodes which have been deposited on the substrate surface. The most widely used of these electrode arrangements is the interdigital transducer, so called due to its pattern of interleaved electrode 'fingers' (Figure 1.1). The interdigital transducer (IDT) is an electro-acoustic transducer capable of both producing and detecting surface acoustic waves on piezo-electric substrates. Its characteristics are highly dependent on its size and geometry.

IDTs are produced on crystal surfaces using standard microcircuit photo-lithography techniques. Briefly, the substrate is polished, cleaned and a thin layer of aluminium is deposited on the surface. The aluminium layer is coated with photo-resist and exposed to ultraviolet light through a

Figure 1.1 Basic geometry of an interdigital transducer with 5 fingers

mask containing the IDT pattern. Unexposed resist and corresponding aluminium are chemically removed leaving the desired IDT metal pattern on the surface of the substrate.

All current 200 megahertz SAW devices have been fabricated at the microcircuit laboratory of the University of New South Wales using a microcomputer-controlled "step-and-repeat" camera exposure system. At the beginning of this project the author developed two computer programs to assist in the manufacture of the SAW devices. The first program produced a 250-times full size metallisation pattern of the device using the University of Adelaide Calcomp plotter. The resultant plots were photo-reduced at the University of N.S.W. for use as masks in the exposure stage of device fabrication. The second program produced a sequence of instructions (on paper tape) for the controller of the "step-and-repeat" camera system to produce devices through multiple exposures using several simple rectangular pattern masks. These programs are listed in Appendix B.

The most satisfactory method of fabrication employed a multiple exposure approach using the "step-and-repeat" system and a mask containing the pattern of one complete IDT. This method was favoured for its greater pattern accuracy and definition and was used for the production of the prototype 200 megahertz coded delay lines.

On a piezo-electric crystal, the mechanical stress present in a surface acoustic wave is accompanied by an electric charge distribution and electric field. When a

voltage is applied to an IDT, an electric field and charge distribution are created at the crystal surface and a corresponding mechanical stress is produced which then propagates away from the transducer along the crystal surface. This is the mechanism of production of surface acoustic waves by an IDT. Conversely when a surface acoustic wave passes through an IDT, the electric field which propagates with the wave interacts with the metal fingers of the IDT to produce an electrical signal at the IDT terminals. Thus a surface acoustic wave may also be detected by an IDT.

To understand the characteristics of an IDT, it is necessary to examine the interaction of the IDT with the electric field/charge distribution associated with a surface acoustic wave. Consider a uniform IDT of infinite width on a piezo-electric substrate as shown in Figure 1.2 and with a unit impulse voltage applied to its terminals. A measure of the resultant charge distribution on the crystal surface is given by the transducer's SOURCE FUNCTION. For an infinite uniform transducer, the source function can be shown to be[4]

$$f(x) = \sum_{n=0}^{\infty} \left[\frac{\varepsilon_p}{\varepsilon_o} + \frac{\beta}{B_n}\right] \frac{\pi}{L} \frac{P_n(\cos(\pi a))}{K\left(\cos\left(\frac{\pi a}{2}\right)\right)} (-1)^n \cos(B_n x) \qquad (1.1)$$

where a = metallisation ratio (the ratio of finger width to gap finger periodicity)

P_n = Legendre polynomial of the first kind

K = complete elliptic integral of the first kind

Figure 1.2 Source function, f(x), for an infinite IDT and modified source function, f'(x), for narrowband applications.

L = spacing between finger centres

ε_p = effective permittivity of the substrate

ε_o = permittivity of free space

B_n = $(2n+1)\pi/L$

β = wave number = ω/v

It can be seen that the source function comprises a fundamental and a series of odd spatial harmonics. Furthermore if $a = 0.5$ the third harmonic disappears leaving only the fundamental and fifth and higher odd harmonics. When considering the operation of the SAW device in the present system, only the fundamental of the source function needs to be considered due to bandwidth limitations. The modified source function (fundamental only and with $a = 0.5$) now becomes:

$$f(x) = \frac{\pi(\varepsilon_p+\varepsilon_o)}{1.854\ \varepsilon_o L} \cos\left(\frac{\pi x}{L}\right) \qquad (1.2)$$

This also assumes that $\beta = B_o$ for narrow bandwidth operation. Thus a good approximation for the source function of an infinite uniform IDT is a sinusoid function.

However a practical IDT has both finite width and length. Restricting the length of the IDT means that the source function will be zero some distance away from the IDT giving rise to a rectangularly enveloped sinusoid as the source function (Figure 1.3).

A useful concept of the source function is to consider

Figure 1.3 Source function for a 5-finger interdigital transducer

it as the impulse response of an electrical-to-acoustical transfer function representing the IDT. An electrical impulse applied to the IDT produces a rectangularly enveloped sinusoid charge distribution which propagates away from the IDT on the substrate surface. At a given point on the surface the charge distribution varies with time and is a replica of the source function (Figure 1.4).

The derivation of expressions for transadmittance (and input admittance) relies on the assumption that the source function of a finite length IDT is simply a truncated version of the source function of an infinite IDT. Equivalent expressions for admittances using this assumption have been produced by many researchers including Burgess[4,5] (based on Engan[14]), Emtage[10], Bahr and Lee[15], Auld and Kino[16], Engan[11] and Wagers[17].

Truncation of an infinite source function ignores "end effects" in finite length IDTs. These end effects cause the source function to be reduced at the ends of an IDT. The result is that admittances in practical transducers are smaller than values from calculations based on truncated source functions. Calculations which allow for end effects have been produced by Hartmann and Secrest[18], Smith and Pedler[19], and Wagers[17].

Hartmann and Secrest proposed a technique for calculating electrostatic functions for finite IDTs thereby allowing a more detailed harmonic analysis of transducer operation. Given more realistic source

Figure 1.4 Propagation of charge distribution for an electrical impulse applied to an interdigital transducer

functions, methods developed by Wagers for finger withdrawal analysis seem to be ideally suited for admittance calculations involving finite IDTs.

However end effects become less significant as the number of fingers in an IDT increases, as expected, and as demonstrated by Smith and Pedler. Based on these results it is the author's opinion that admittances calculated in this thesis for narrowband fundamental operation of 11-finger IDTs differ from actual values by less than a few percent.

1.2 TRANSADMITTANCE

Once a surface acoustic wave has been launched, an IDT may be used for detecting the wave. In this role, the characteristics of the IDT are also determined by its source function. Of more interest than the electro-acoustic characteristics of a single IDT is the electrical interaction of a transducer pair. This characteristic comprises the electrical-to-acoustical conversion of an IDT, propagation and the acoustical-to-electrical conversion of a second IDT.

A useful measure of this interaction between two IDTs is the TRANSADMITTANCE[5] and is the function relating the voltage input at a transducer to the current produced in a second, short-circuited, transducer (Figure 1.5).

Burgess[4] showed that the transadmittance of an IDT pair is given by:

$$Y_{21}(\omega) = \frac{-|\omega|\Delta\varepsilon_o{}^2 W}{\varepsilon_p+\varepsilon_o} \int_{-\infty}^{\infty} f_1(x)e^{j\beta x}\,dx \int_{-\infty}^{\infty} f_2(x)e^{-j\beta x}\,dx \qquad (1.3)$$

Figure 1.5 Transadmittance, $Y(\omega)$, between the interdigital transducers

where Δ = coupling coefficient = $\Delta v/v$

W = the overlap length of fingers in the IDTs

β = wave number

ω = radian frequency

The transadmittance is a product of two integrals involving the source functions of the transducers ($f_1(x)$ and $f_2(x)$). It should also be noted that the integrals are Fourier integrals. Further manipulation of equation 1.3 yields a more useful form.

Assume that the first IDT is centred on x_1 and its source function is non-zero in the interval $x_1 - \Delta x_1 < x < x_1 + \Delta x_1$. Likewise the second IDT is centred on x_2 and its source function is non-zero in $x_2 - \Delta x_2 < x < x_2 + \Delta x_2$. Equation 1.3 may be rewritten as:

$$Y_{21}(\omega) = K \int_{x_1-\Delta x_1}^{x_1+\Delta x_1} f_1(x) e^{j\beta x}\, dx \int_{x_2-\Delta x_2}^{x_2+\Delta x_2} f_2(x) e^{-j\beta x}\, dx \qquad (1.4)$$

where $$K = \frac{-|\omega| \Delta \varepsilon_o^2 W}{\varepsilon_o + \varepsilon_p}$$

By changing the variables of integration in the two integrals, equation 1.4 becomes:

$$Y_{21}(\omega) = K \int_{-\Delta x_1}^{\Delta x_1} f_1(x-x_1)e^{j\beta x}e^{-j\beta x_1}\, dx \cdot \int_{-\Delta x_2}^{\Delta x_2} f_2(x-x_2)e^{-j\beta x}e^{j\beta x_2}\, dx$$

$$= K\, e^{j\beta(x_2-x_1)} \int_{-\Delta x_1}^{\Delta x_1} f_1(x-x_1)e^{j\beta x}\, dx \cdot \int_{-\Delta x_2}^{\Delta x_2} f_2(x-x_2)e^{-j\beta x}\, dx \tag{1.5}$$

Note that $f_1(x)$ is a function centred on x_1 and hence $f(x-x_1)$ is the same function centred on zero (similarly for $f_2(x)$, x_2 and $f_2(x-x_2)$). Therefore equation 1.5 states that the product of the Fourier integrals of source functions $f_1(x)$ and $f_2(x)$ is equal to the product of the Fourier integrals of $f_1(x-x_1)$ and $f_2(x-x_2)$ multiplied by a "phase" term, $e^{j\beta(x_2-x_1)}$, which takes account of the spatial separation of the two IDTs.

Now let $f_1'(x) = f_1(x-x_1)$

= source function of first IDT translated to zero

and $f_2'(x) = f_2(x-x_2)$

= source function of second IDT translated to zero

Also let $F_1(\beta)$ = Fourier transform of $f_1'(x)$

and $F_2(\beta)$ = Fourier transform of $f_2'(x)$

The first integral in equation 1.5 is simply $F_1(-\beta)$ and the second integral is $F_2(\beta)$ and hence equation 1.5 is rewritten as:

$$Y_{21}(\omega) = KF_1(-\beta)F_2(\beta)e^{j\beta(x_2-x_1)} \tag{1.6}$$

It should also be noticed that:

$$\begin{aligned} Y_{12}(\omega) &= KF_2(-\beta)F_1(\beta)e^{j\beta(x_1-x_2)} \\ &= Y_{21}(-\omega) \\ &= Y^*_{21}(\omega) = Y_{21}(\omega) \end{aligned}$$

From equation 1.6 the time domain form of the transadmittance can be predicted. The frequency domain transadmittance is the product of three main terms; two Fourier transforms and a phase term. Transforming to the time domain shows that the time domain transadmittance is the convolution of the IDT source functions (one of which is reversed in time) and a time delay (the propagation time between IDTs). For two identical IDTs whose source functions are rectangularly enveloped sinusoids, the time domain transadmittance is a triangularly enveloped sinusoid delayed by the propagation delay between IDTs. Section 1.4 will show how this time response is used to generate a binary coded SAW device.

1.3 INPUT ADMITTANCE

The input admittance (or self admittance) of an IDT results from the interaction of an IDT with the surface acoustic waves produced by the same IDT. If a voltage impulse is applied to an IDT, a charge distribution is set up under the IDT according to its source function. This results in two waves propagating away from the IDT in opposite directions. It is the interaction of these two waves with the IDT which is responsible for the input admittance. This mechanism is similar to both the launching and detection of waves in an IDT pair (thus producing the transadmittance function) except that it occurs within a single IDT.

In fact, the time domain input admittance is equal to twice the transadmittance of an identical IDT pair with zero spatial separation, except that the input admittance is causal (i.e. zero for negative time). It follows from this definition that the frequency domain input conductance (the real part of the input admittance) is given by the transadmittance of two identical spatially coincident IDTs. Furthermore, due to the causality constraint, the input susceptance is the Hilbert transform of the input conductance.

To support this definition of the time domain input admittance, consider Figure 1.6. An IDT is placed on a substrate with two identical IDTs placed symmetrically on either side. The distance between the IDT centre lines (centres of symmetry) corresponds to a propagation delay

Figure 1.6 Derivation of input admittance in the time domain

of T. If a voltage impulse is applied simultaneously to IDT 2 and IDT 3 at time $t = -T$, then at $t = 0$ there are two identical waves (although propagating in opposite directions) beneath IDT 1. This situation is identical to one where an impulse is applied only to IDT 1 at $t = 0$. Therefore the current in IDT 1 is identical in both cases.

The current in IDT 1 is given by the transadmittances between the IDTs as:

$$i_1(t) = y_{11}(t) = [y_{21}(t) + y_{31}(t)] * \delta(t+T) \quad \text{for} \quad t \geqslant 0$$

$$= 2y(t) * \delta(t+T)$$

since $y_{21}(t) = y_{31}(t) = y(t)$ (say)

$$= 2y(t+T) \qquad \text{for} \quad t \geqslant 0$$

and $$= 0 \qquad \text{for} \quad t < 0$$

Thus the corresponding frequency domain input admittance is given by

$$Y_{11}(\omega) = G(\omega) + jB(\omega) \tag{1.7}$$

where $G(\omega) = Y(\omega) =$ Fourier transform of $y(t+T)$

$B(\omega) = H(\omega) * Y(\omega)$

$H(\omega) =$ the Hilbert Transform function $= \frac{2}{\omega}$

In addition to the input admittance due to the interaction of the IDT with the waves which it produces, each IDT displays a capacitance due to simple electrostatics. This static capacitance is calculated separately

from the IDT input admittance and is described further in section 2.3.

1.4 N-Ports, 2-Ports and 1-Ports

Consider a general SAW device with n IDTs whose terminals are available as n electrical ports (Figure 1.7). The electrical response of this n-port can be represented in the frequency domain by:

$$\begin{bmatrix} I_1 \\ I_2 \\ I_3 \\ \cdot \\ \cdot \\ \cdot \\ \cdot \\ \cdot \\ I_n \end{bmatrix} = \begin{bmatrix} Y_{11} & Y_{12} & \cdot & \cdot & \cdot & \cdot & \cdot & Y_{1n} \\ Y_{12} & Y_{22} & & & & & & Y_{2n} \\ Y_{13} & & \cdot & & & & & Y_{3n} \\ \cdot & & & \cdot & & & & \cdot \\ \cdot & & & & \cdot & & & \cdot \\ \cdot & & & & & & & \cdot \\ \cdot & & & & & & & \cdot \\ \cdot & & & & & & & \cdot \\ Y_{1n} & Y_{n2} & \cdot & \cdot & \cdot & \cdot & \cdot & Y_{nn} \end{bmatrix} \begin{bmatrix} V_1 \\ V_2 \\ V_3 \\ \cdot \\ \cdot \\ \cdot \\ \cdot \\ \cdot \\ V_n \end{bmatrix} \qquad (1.8)$$

where $I_1, I_2, \ldots, I_n$ are the currents in the n ports and $V_1, V_2, \ldots, V_n$ are the voltages at the n ports. If $n-1$ ports are connected in parallel (e.g. $i = 2,..,n$ say) to form a 2-port network, then equation 1.8 is reduced to

$$\begin{bmatrix} I_1 \\ I_t \end{bmatrix} = \begin{bmatrix} Y_{11} & Y_{t1} \\ Y_{1t} & Y_{tt} \end{bmatrix} \begin{bmatrix} V_1 \\ V_t \end{bmatrix}$$

where $$I_t = \sum_{i=2}^{n} I_i$$

and $$V_t = V_i \quad \text{for } i = 2,3, \ldots n$$

Figure 1.7 Generalised electrical n-port

Furthermore the total transadmittance between the two ports is given by

$$Y_{1t} = Y_{t1} = \sum_{i=2}^{n} Y_{1i}$$

= the sum of individual transadmittances between port 1 and ports 2 to n in the n-port.

In the time domain,

$$y_{1t}(t) = y_{t1}(t) = \sum_{i=2}^{n} y_{1i} \ .$$

As the transducers which comprise the second port are located at differing distances from the transducer forming port one, the time domain transadmittance comprises a series of pulses occurring at times which correspond to the spacings between the first IDT and the IDTs forming the second port.

Another interpretation is that an impulse applied to port one launches two waves. One wave propagates to the end of the substrate and is lost (artificial means are used to suppress this wave since unwanted reflections may occur at the edge of the substrate). The other wave propagates towards the other IDTs forming port two. As the wave passes under each IDT a current is produced in the short-circuit termination of port two. The exact nature of the current waveform depends on individual IDTs as described in section 1.2, but the timing of the pulses is solely dependent on the physical separation of IDTs on the substrate. This is the basis of a coded SAW delay line, and is diagramatically shown in Figure 1.8. A binary code may be generated by

Figure 1.8 Example of 2-port delay line response

placing IDTs at certain locations on the substrate. This will result in pulses occurring at corresponding time instants. The presence or absence of pulses at certain times correspond to ones and zeroes of the binary code.

Considering the general n-port again, a one-port network can be constructed by connecting all ports in parallel. The response of the one-port is:

$$I = YV \tag{1.9}$$

where

$$I = \sum_{i=1}^{n} I_i$$

$$V = V_i \quad \text{for} \quad i = 1,2 \ldots n$$

and

$$Y = \sum_{i=1}^{n} \sum_{j=1}^{n} Y_{ij} \tag{1.10}$$

$$= \text{sum of all admittances}$$

The time response of this network is more complicated than the 2-port since an impulse applied to the one-port will cause all IDTs to launch waves which will interact with all other IDTs. In order to readily identify the binary code of a coded delay line, one of the IDTs is placed some distance from the others on the substrate (Figure 1.9). This separate IDT is referred to as the "transmit IDT" and the remainder as the "coding set" since it is the coding set IDTs which are placed in such a way as to generate the code.

When an impulse is applied to the one-port, all IDTs launch waves. A complicated response arises due to the

Figure 1.9 Example of binary code generation with SAW one-port

interaction of the coding set with itself. The last pulse due to this interaction occurs at a time which corresponds to the maximum distance between any two IDTs in the coding set. At some later time, the wave launched from the transmit IDT reaches the coding set and produces a response as described for the 2-port. However for the one-port, the amplitude of the pulses are double those of the 2-port since as the wave from the transmit IDT reaches each IDT in the coding set, a wave launched by an IDT in the coding set arrives at the transmit IDT (thereby doubling the response).

From equation 1.10, the one-port admittance is:

$$Y = \sum_{i=1}^{n} \sum_{j=1}^{n} Y_{ij} \quad \text{(capacitance not included)}$$

$$= Y_{11} + \sum_{i=2}^{n} \sum_{j=2}^{n} Y_{ij} + \sum_{i=2}^{n} \left(Y_{1i} + Y_{i1}\right)$$

The code is due to the term $\sum_{i=2}^{n} \left(Y_{1i} + Y_{i1}\right)$ but this is equal to $2 \sum_{i=2}^{n} Y_{1i}$ since $Y_{1i} = Y_{i1}$.

1.5 SECOND ORDER RESPONSES

In a practical system the SAW one-port is connected to a network which has a finite non-zero impedance. In the case where the one-port is connected to a voltage source, V, with a source impedance, Z, (shown in Figure 1.10), circuit analysis shows that the current into the one-port is

$$I = VY(1 + ZY)^{-1} \tag{1.11}$$

Figure 1.10 Response of a SAW one-port driven from a source with finite impedance

Since the one-port admittance, Y, is usually very small, the product ZY is significantly less than unity and the binomial expansion for $(1+ZY)^{-1}$ enables equation 1.11 to be approximated by:

$$I = V(Y - ZY^2 + Z^2Y^3 - Z^3Y^4 + \ldots.) \qquad (1.12)$$

It is seen that a non-zero source impedance has introduced higher order response components (compared with equation 1.9). The second order response is due to the term ZY^2, and in the time domain represents the ideal one-port response convolved with itself and then with the source impedance. This auto-convolution of the ideal response can (for some codes) produce significant responses and may interfere with the desired code to be produced by the delay line.

The higher order terms of equation 1.12 are produced by a mechanism known as electro-acoustic regeneration. For a non-zero source impedance, any current flowing in the one-port terminals due to the desired code response produces a voltage at the terminals. This voltage causes a secondary launching of waves on the substrate which are responsible for the second order response. This mechanism of electro-acoustic regeneration is repeated to produce the other higher order responses.

Although all higher order responses have the potential to interfere with the desired response, the second order response is considered to be the most significant and is therefore investigated in detail in this thesis.

1.6 THE CODED LABEL SYSTEM

The prototype system comprises a transmitter, coded SAW labels, a receiver and receiver processing circuits. A block diagram is shown in Figure 1.11.

The transmitter outputs 5 cycles of 200 megahertz which passes through the transmit antenna and some of this energy is received by the label antenna. A small amount of this energy is then re-transmitted from the label and passes through the receive antenna to the receiver where the signal is processed to extract the label's code.

In the frequency domain, the overall system response is the product of the frequency responses of the individual sections comprising the total system.

i.e. $G_s(\omega) = G_1(\omega)G_2(\omega) \ldots$

where $G_s(\omega)$ = total system frequency response

$G_i(\omega)$ = individual frequency responses

Correspondingly in the time domain, the overall system response is the convolution of the impulse responses of individual sections of the system

$$g_s(t) = g_1(t) * g_2(t) * g_3(t) * \ldots$$

where $g_s(t)$ = total system time response

$g_i(t)$ = individual time response

Figure 1.11 Block diagram of coded SAW delay line system

Both approaches have been used to model the 200 megahertz prototype system with the time domain approach giving a better insight into system response as well as being a more efficient computing algorithm.

1.7 SYSTEM SPECIFICATIONS

1.7.1 Transmitter

The transmitter local oscillator generates a frequency of 200 000 768 hertz of which 5 cycles are output via a power amplifier and the transmit antenna. The transmit frequency is chosen to give an intermediate frequency of 768 hertz in the receiver. The transmit pulse is repeated at regular intervals.

1.7.2 Transmit and Receive Antennae

Both the transmit and receive antennae are tuned for a centre frequency of 200 megahertz and a quality factor (Q) of 5.

1.7.3 Label Antenna

The label antenna comprises a loop of conductor forming the inductance with capacitance added to produce resonance at 200 megahertz. (Note that part of the capacitance is contributed by the SAW delay line). Resistance is added to the loop to give an overall Q of 5.

1.7.4 SAW Delay Line

The physical placement of IDTs on the substrate is chosen to give a binary code as described in section 1.4. The bits are required to be spaced 37.5 nanoseconds apart resulting in a spacing of 118.5 μmetres between IDT centres on the substrate. (Propagation velocity for quartz is 3160 metres per second).

Section 1.4 also mentioned that the coding set IDTs are placed some distance away from the transmit IDT. This is to ensure that the binary code occurs well clear of any response due to the interaction of the coding set with itself. The coding set may be placed such that a "zone of silence"[7] (or interval of no response) occurs before the code commences. This zone of silence is chosen to be one bit position and ensures that no pulses occur in the bit position before the first code bit (see Figure 1.12). Unfortunately this does not ensure that the second order response will be zero in the zone of silence.

Other specifications for the SAW delay line are given in Figure 1.13.

1.7.5 Receiver (Figure 1.14)

The signal from the receiver antenna is amplified and fed to a series of mixers (one mixer per bit of code). The amplifier following the antenna is effectively turned off before each transmit pulse and restored just before the code commences to prevent overload from the transmit pulse.

Figure 1.12 Layout for an M-bit coded SAW one-port

Centre frequency	200 Megahertz
Substrate	Quartz (Y-cut)
Propagation Velocity	3160 metres/second
Coupling Coefficient	-9.3×10^{-4}
Code Bit Separation	37.5 nanoseconds
Spacing between transducer centre lines	118.5 microns
Fingers per transducer	11
Width of finger ($\lambda/4$)	3.95 microns
Spacing of fingers, L($\lambda/2$)	7.90 microns
Finger overlap width, W(70λ)	1106 microns
Zone of Silence	1 Bit

Figure 1.13 Delay Line Specifications

Figure 1.14 Receiver Components

Each mixer is also fed with a gated burst of local receiver oscillator. The bursts are timed to coincide with the pulses in the received signal representing bits in the code. Thus each mixer acts as a correlation detector for a single different bit. The local oscillator bursts are 5 cycles (25 nanoseconds) of 200 000 000 hertz thereby creating an intermediate frequency (i.f.) of 768 hertz at the outputs of the mixers. The outputs are then filtered and down-converted once more by another local oscillator/ mixer combination to produce a second i.f. of 32 hertz. The 32 hertz signals are passed through envelope detectors and then integrators.

The outputs of the integrators are called the channel outputs and are a measure of the bits in the code. Further circuitry is responsible for the decision as to whether the output from a channel represents a one or zero in the code. The integration of the channel signals continues until an overload condition is detected or a specified time limit expires.

The mean signal level (MSL) is the average of two channel outputs corresponding to two bits in the code which are always set to one. The MSL is a measure of the signal level produced by 'ones' in the code. Other levels, which are defined as specified percentages of the MSL, are used to decide the value of the code bits. These levels are the unity-upper-level (UUL), the unity-lower level (ULL) and the zero-upper-level (ZUL). If a channel output is between ULL and UUL then the code bit is decoded as a 'one'. If a

Figure 1.15 Receiver channel outputs

channel output is less than ZUL then the corresponding bit is decoded as a zero. Channel outputs not falling within these intervals are interpreted as errors by the logic (see Figure 1.15).

CHAPTER 2

FREQUENCY DOMAIN MODELLING

2.0 INTRODUCTION

As mentioned in the previous chapter, two approaches to modelling the coded SAW label system have been developed. This chapter deals with the frequency domain approach and its eventual computer implementation. The frequency response of each part of the system is considered and then combined to give an overall response. The frequency response is transformed using a Fast Fourier Transform to give the equivalent time response. Several practical considerations relating to this computation are discussed.

2.1 TRANSADMITTANCE

In section 1.2 it was shown that the frequency domain transadmittance is the product of the Fourier transforms of the IDT source functions and a phase term produced by the propagation delay. The source functions and their Fourier transforms will now be examined in more detail.

From Figure 2.1 it can be shown that the source function for an N-finger IDT is given by

Figure 2.1 Source function for an N-finger transducer

$$f_i(x) = \frac{\pi(\varepsilon_p+\varepsilon_o)}{1.854\ \varepsilon_o L} \cos \beta_0\left(x + \frac{(N-1)L}{2}\right) \tag{2.1}$$

for x in the interval $-NL/2 \leqslant x \leqslant NL/2$ and zero elsewhere.

This source function is for a transducer which is symmetrical about $x = 0$. This axis of symmetry will be referred to as the transducer's centre line.

Taking the Fourier Transform of this source function gives

$$F_i(\beta) = A \int_{\frac{-NL}{2}}^{\frac{NL}{2}} \cos\left[\beta_o x + \beta_o \frac{(N-1)L}{2}\right] e^{-j\beta x}\, dx$$

$$= A \int_{-\infty}^{+\infty} \mathrm{rect}\left(\frac{x}{NL}\right) e^{-j\beta x}\, dx \; * \int_{-\infty}^{+\infty} \cos[\beta_o x + \pi(N-1)] e^{-j\beta x}\, dx$$

where $A = \dfrac{\pi(\varepsilon_p + \varepsilon_o)}{1.854\ \varepsilon_o L}$

and $\mathrm{rect}(y) = 1$ for $-\frac{1}{2} \leqslant y \leqslant \frac{1}{2}$

$= 0$ elsewhere.

Thus the Fourier transform of a rectangularly enveloped sinusoid is the convolution of the Fourier transforms of the sinusoid and its envelope. The two Fourier integrals will be evaluated separately.

$$F\left[\mathrm{rect}\left(\frac{x}{NL}\right)\right] = NL \quad \mathrm{sinc}\left(\frac{NL\beta}{2\pi}\right)$$

and $$F\left[\cos[\beta_o x + \pi(N-1)]\right] = \frac{1}{2}\left[\delta(\beta-\beta_o)e^{j\pi(N-1)} + \delta(\beta+\beta_o)e^{-j\pi(N-1)}\right]$$

where $F[\quad]$ denotes the Fourier transform

and $$\mathrm{sinc}(y) = \frac{\sin(\pi y)}{(\pi y)}$$

Convolving the two transforms produces

$$F_i(\beta) = \frac{ANL}{2}\left[e^{j\pi(N-1)}\ \mathrm{sinc}\left(\frac{NL}{2\pi}(\beta-\beta_o)\right) + e^{-j\pi(N-1)}\ \mathrm{sinc}\left(\frac{NL}{2\pi}(\beta+\beta_o)\right)\right]$$

$$= A\left[e^{j\pi(N-1)}\ \frac{\sin\left(\frac{NL}{2}(\beta-\beta_o)\right)}{(\beta-\beta_o)} + e^{-j\pi(N-1)}\ \frac{\sin\left(\frac{NL}{2}(\beta+\beta_o)\right)}{(\beta+\beta_o)}\right] \tag{2.2}$$

The Fourier transforms of the source functions are evaluated using this expression and the total transadmittance of an IDT pair is given by

$$Y_{21}(\omega) = KF_1(-\beta)F_2(\beta)e^{j\beta(x_2-x_1)}$$

where $$K = \frac{-|\omega|\Delta\varepsilon_o^{\ 2}}{\varepsilon_p + \varepsilon_o}$$

and $$\beta = \omega/v\ .$$

In the frequency domain computer model, this calculation is performed by FORTRAN routines XAD and SF (Appendix A). SF evaluates the Fourier transforms of the source functions (i.e. $F_1(-\beta)$ and $F_2(\beta)$) while XAD completes the computation of $Y_{21}(\omega)$ and also sums all transadmittances from all possible IDT pairs on the delay line. Hence for a one-port XAD calculates the sum:

$$\text{total of transadmittances} = \sum_i \sum_j Y_{ij}(\omega) \qquad i \neq j$$

2.2 INPUT ADMITTANCE

The calculation of the input admittance of an IDT is based on the fact that the real part of the input admittance is equal to the transadmittance of two spatially coincident IDTs. Furthermore, since the input admittance is causal in the time domain, the real and imaginary parts of the input admittance are related by the Hilbert Transform.

Burgess[4] calculated the input admittance by using algorithms developed for transadmittance calculation to evaluate the input conductance in the frequency domain. The Hilbert Transform operation required to yield the input susceptance was achieved by transforming the conductance to the time domain via Fast Fourier Transform methods. The time domain conductance was multiplied by $-j\,\mathrm{sgn}(t)$ (the time domain equivalent of the Hilbert Transform) and then transformed back to the frequency domain to give the input susceptance.

However, by making use of an approximation, both the input conductance and susceptance may be calculated in the frequency domain without the need to transform to the time domain in order to implement the Hilbert Transform.

It can be shown that if the conductance, G, is of the form

$$G = G_o\left(\frac{\sin x}{x}\right)^2 \tag{2.4}$$

then the susceptance, B, is given by

$$B = G_o\left(\frac{\sin 2x - 2x}{2x^2}\right) \tag{2.5}$$

where G_o is a constant.

(A proof of this relationship showing that G and B are related by the Hilbert Transform is given in Appendix C.)

Since the conductance is the transadmittance of two spatially coincident IDTs, G is derived from equation 2.3 as

$$G(\omega) = KF_1(-\beta)F_1(\beta) \tag{2.6}$$

However $F_1(\beta)$ is the Fourier transform of a real function (the source function) and therefore

$$F_1(-\beta) = F_1^*(\beta)$$

hence equation 2.6 becomes

$$G(\omega) = K|F_1(\beta)|^2 \tag{2.7}$$

Substituting equation 2.2 into equation 2.7 yields

$$G(\omega) \approx \frac{KA^2N^2L^2}{4}\left[\frac{\sin^2 a}{a^2} + \frac{\sin^2 b}{b^2}\right] \tag{2.8}$$

$$\text{where} \quad a = \frac{NL}{2}(\beta-\beta_o)$$

$$b = \frac{NL}{2}(\beta+\beta_o)$$

This is now in the form of equation 2.4 and, by using superposition and equation 2.5, an expression for the susceptance is found, viz:

$$B(\omega) \approx \frac{KA^2N^2L^2}{4}\left[\frac{\sin\ 2a - 2a}{2a^2} + \frac{\sin\ 2b - 2b}{2b^2}\right] \tag{2.9}$$

The approximations in this derivation arise from the evaluation of $|F_1(\beta)|^2$ and also G_o ($= \frac{KA^2N^2L^2}{4}$ in this case) which is not a constant as assumed but contains a $|\omega|$ term in the expression for K. However over the narrow bandwidths under consideration the approximations appear to be acceptable.

FORTRAN routine INPAD (Appendix A) is used to calculate both the input conductance and susceptance in the frequency domain according to equations 2.8 and 2.9. The input admittance is calculated for every IDT on the delay line and added to the transadmittances already calculated in XAD. Therefore, between XAD and INPAD, the total admittance (less static capacitance) of the delay line is calculated viz:

$$\text{total admittance (less capacitance)} = \sum_i \sum_j Y_{ij} \ .$$

2.3 STATIC CAPACITANCE

Apart from the transadmittances and input admittances which arise from the interaction of surface acoustic waves with the IDTs, each IDT exhibits an amount of capacitance due to electrostatic interaction between fingers in the IDT. The static capacitance may also be thought of as a part of the input admittance associated with the "cut-off" modes of propagation of the IDT.

For a uniform IDT of N fingers the static capacitance is:

$$C_s = \frac{1}{2}(\varepsilon_p + \varepsilon_o)(N-1)W$$

This capacitance is calculated for each IDT in FORTRAN routine STATCAP (Appendix A) and summed to give a total capacitance value for the delay line. This capacitance is not added directly to the total admittance calculation but is incorporated into the label antenna tuned circuit as described in section 2.8.

2.4 SYSTEM FUNCTIONS

It has been shown that routines XAD, INPAD and STATCAP are used to calculate the frequency domain admittance of the SAW one-port delay line. However expressions must be found for the frequency response of other parts of the interrogation system. In this way, a total system frequency response can be obtained. The following sections examine the main parts of the system and their frequency functions.

2.4.1 Transmitter Pulse $G_1(\omega)$

The transmitter pulse comprises 5 cycles of 200 megahertz sine wave (Figure 2.2). In the time domain,

$$g_1(t) = \mathrm{rect}\left(\frac{t}{T} - \frac{1}{2}\right) \sin(\omega_o t) \tag{2.10}$$

$$\text{where} \quad \mathrm{rect}(t) = 1 \quad \text{for} \quad -\frac{1}{2} \leqslant t \leqslant \frac{1}{2}$$

$$= 0 \quad \text{otherwise} .$$

An expression for the frequency function is readily obtained:

$$G_1(\omega) = j\left[\frac{\sin(\omega+\omega_o)\frac{T}{2}}{(\omega+\omega_o)} e^{-j(\omega+\omega_o)T/2} - \frac{\sin(\omega-\omega_o)\frac{T}{2}}{(\omega-\omega_o)} e^{-j(\omega-\omega_o)T/2}\right] \tag{2.11}$$

FORTRAN routine INPULS (Appendix A) calculates the frequency domain function for the transmitter pulse.

2.4.2 Transmit Antenna $G_2(\omega)$

The equivalent circuit used to model the transmit antenna is a series tuned circuit as shown in Figure 2.3.

The frequency response is

$$G_2(\omega) = \frac{j\omega C}{(1 - \omega^2 LC) + j\omega RC}$$

$$= \frac{j\,\frac{1}{Q}\left(\frac{\omega}{\omega_o}\right)}{1 - \left(\frac{\omega}{\omega_o}\right)^2 + j\,\frac{1}{Q}\left(\frac{\omega}{\omega_o}\right)} \tag{2.12}$$

Figure 2.2 Transmitter pulse

Figure 2.3 Transmit and receive antennae equivalent circuit

for $R = 1$. $G_2(\omega)$ is unity at $\omega = \omega_o$.

This function is implemented in FORTRAN routine TUNCCT Appendix A) with $\omega_o = 2\pi \times 200$ megahertz and $Q = 5$.

2.4.3 Receive Antenna $G_3(\omega)$

Since the same antenna is used for both transmit and receive functions in the interrogation system, the receive antenna is modelled exactly as for the transmit antenna.

$$G_3(\omega) = G_2(\omega) \tag{2.13}$$

2.4.4 Coupling to Label $G_4(\omega)$

The coupling between the transmit antenna and the label antenna is assumed to be purely by mutual inductance. The current in the transmit antenna induces a voltage in the label antenna with the induced voltage being determined by the mutual inductance.

$$G_4(\omega) = j\omega M_1 \tag{2.14}$$

2.4.5 Coupling from Label $G_5(\omega)$

The coupling between the label antenna and the receive antenna is also assumed to be by mutual inductance, as for $G_4(\omega)$.

$$G_5(\omega) = j\omega M_2 \tag{2.15}$$

When calculating $G_4(\omega)$ and $G_5(\omega)$ the mutual inductances, M_1 and M_2, are set to unity since the absolute amplitude of the response is unimportant in present investigations.

2.4.6 Label $G_6(\omega)$

The label comprises a coded SAW delay line connected to an antenna/tuned circuit which performs both receive and transmit functions. Figure 2.4 represents the model of the label. An equivalent model is shown in Figure 2.5 and is the model used by Burgess.

The current, I, in the label antenna due to an induced voltage, V, is

$$I = \frac{V}{Z_s} - \frac{V}{Z_s^2} \frac{Z_t}{(1 + YZ_t)} \tag{2.16}$$

Since YZ_t is significantly less than unity, the binomial expansion can be used to rewrite equation 2.16 as

$$G_6(\omega) = \frac{Z_s - Z_t}{Z_s} + \frac{Z_t}{Z_s^2}\left(YZ_t - Y^2Z_t^2 + Y^3Z_t^3 - \ldots\right) \tag{2.17}$$

The first term in equation 2.17 represents the response of the antenna/tuned circuit only. The second term represents the additional response due to the presence of the delay line. The first term will now be neglected since it contributes no useful information about the performance of the coded label system. It is, however, much larger in amplitude than the second term which contains

Figure 2.4 Circuit of label

Figure 2.5 Equivalent circuit of label

the code information (see Figure 2.6).

Therefore, the label's response is effectively

$$G_6(\omega) = \left(\frac{Z_t}{Z_s}\right)^2 \left(Y - Y^2 Z_t + Y^3 Z_t{}^2 - \ldots\right) \tag{2.18}$$

Upon inspection of the terms in equation 2.18, Y represents the response of the delay line and contains the code information. Other terms represent second and higher order responses as already outlined in section 1.5. Since only the second order response is of interest, the expression used in the frequency domain modelling is

$$G_6(\omega) = \left(\frac{Z_t}{Z_s}\right)^2 \left(Y - Y^2 Z_t\right) \tag{2.19}$$

2.4.7 Total System Response $G_s(\omega)$

In the frequency domain the total system frequency response is the product of the individual frequency responses which comprise the system, viz:

$$G_s(\omega) = G_1(\omega) G_2(\omega) G_3(\omega) G_4(\omega) G_5(\omega) G_6(\omega) \tag{2.20}$$

The total system frequency response is calculated in FORTRAN program MINISEC by using the routines already outlined in previous sections and some supporting calculations.

Figure 2.6 Total Label Response

2.5 BASEBAND EQUIVALENTS OF SYSTEM FUNCTIONS

Since all system frequency functions, and hence the total system function, are centred around the same frequency (in this case 200 megahertz), the total system response in the time domain is expected to take the form of a modulated 200 megahertz sinusoid. It is this modulation (both amplitude and phase) which contains the information pertinent to the system analysis (i.e. the code information). The carrier frequency is not particularly important except that it is known to be 200 megahertz.

It can be shown[8] that for a bandpass frequency function, $G(\omega)$, with a time domain equivalent of the form

$$g(t) = r(t)\cos(\omega_o t + \phi(t)) \tag{2.21}$$

where $r(t)$ and $\phi(t)$ represent amplitude and phase modulation respectively, then a baseband equivalent may be defined as

$$\hat{g}(t) = \frac{1}{2}r(t)e^{j\phi}$$

(In this case ϕ is constant, i.e. no phase modulation.)
In the frequency domain:

$$G(\omega) = \frac{1}{2}\, e^{j\phi}[R(\omega-\omega_o) + R(\omega+\omega_o)] \tag{2.23}$$

and $$\hat{G}(\omega) = \frac{1}{2}\, e^{j\phi}\, R(\omega) \tag{2.24}$$

where $R(\omega)$ is the Fourier Transform of $r(t)$.

In the frequency domain the baseband equivalent of the frequency function is simply the positive frequency part of the function translated to the origin (as shown in Figure 2.7).

The amplitude modulation (the envelope) is given by

$$r(t) = 2|\hat{g}(t)| \tag{2.25}$$

Other important characteristics of baseband equivalents are shown by considering an arbitrary bandpass frequency function, $G(\omega)$, which is the product of n bandpass functions, $G_i(\omega)$, for $i = 1$ to n.

$$G(\omega) = \prod_{i=1}^{n} G_i(\omega) \tag{2.26}$$

and hence in the time domain

$$g(t) = g_1(t)*g_2(t)* \ldots\ldots *g_n(t) \tag{2.27}$$

The baseband equivalents are similarly related, i.e.

$$\hat{G}(\omega) = \prod_{i=1}^{n} \hat{G}_i(\omega) \tag{2.28}$$

and $$\hat{g}(t) = \hat{g}_1(t)*\hat{g}_2(t)* \ldots\ldots *\hat{g}_n(t) \tag{2.29}$$

Equation 2.29 is particularly important since it is the basis of the time domain modelling approach described in Chapter 3.

However program MINISEC computes the total system frequency response as per equation 2.20 and then translates

Figure 2.7 Bandpass and baseband equivalent frequency functions

this function in frequency to form its baseband equivalent. A fast Fourier Transform is performed and the resultant complex time function comprises the envelope and phase information. The envelope is then extracted using equation 2.25.

The reason for working with baseband functions rather than bandpass functions is that the envelope is readily available. This has advantages when considering a practical implementation of the Fast Fourier Transform algorithm as explained in the following sections.

2.6 TIME DOMAIN ALIASING AND THE FAST FOURIER TRANSFORM

The problem of frequency domain aliasing due to an insufficient sampling rate of the time function is well known. This arises from interference between adjacent periodic spectra in the frequency domain. The solution to this problem is usually to pass the time function through an anti-aliasing filter before sampling to reduce its bandwidth. The effect of this is to narrow the individual periodic spectra and hence decrease interference. Another solution is to increase the time sampling rate which spreads the periodic spectra further apart in frequency.

Conversely consider the frequency function which corresponds to a time function of finite duration. If the frequency function is not sampled at close enough frequency intervals, the time function resulting from a Fast Fourier Transform will be distorted. This is due to overlapping

of individual periodic time functions comprising the time response. In effect, this is time domain aliasing.

The remedies for time domain aliasing parallel those for frequency domain aliasing. Either the frequency function must be convolved with a suitable frequency domain anti-aliasing filter (which would be computationally difficult and undesirable) or the frequency sampling rate must be increased. A suitable sampling rate must be chosen such that the period of the periodic time functions is greater than their duration in order to avoid overlapping.

At this point it is convenient to examine the case of the SAW label system. The duration of the time response is governed almost entirely by the SAW delay line admittance function, $Y(\omega)$. The time domain admittance, $y(t)$, has a duration of approximately 1.5 microseconds (the propagation time between the transmit IDT and the last IDT in the coding set for a 20-bit code). Furthermore, the second order response, which is due to a $Y^2(\omega)$ term, has a time function arising from $y(t)*y(t)$ which is of 3 microseconds duration.

Therefore in a 20-bit SAW label system the maximum duration of the time response (considering primary and second order responses only) is approximately 3 microseconds. For no time domain aliasing, this implies a maximum spacing between frequency samples of 333 kilohertz.

2.7 TIME DOMAIN RESOLUTION

Another constraint on the Fast Fourier Transform calculations is that the samples of the time function must be sufficiently close together so as to give an accurate representation of the time function. This is a constraint of time domain resolution.

It is known that the time domain response of the SAW label system is a modulated 200 megahertz carrier. In order to attain adequate resolution, it is necessary to have 10 samples (arbitrary) per cycle of carrier. Over the 3 microseconds duration of the time response this amounts to 6000 samples.

On the other hand, if only the envelope of the time response is required the number of samples may be reduced to 10 per bit period (i.e. 37.5 nanoseconds). Over 3 microseconds, the total number of samples would be 800. The advantage in using baseband equivalent functions is now apparent. The computational time required for calculating the envelope of the time response is considerably less than that for calculating the complete time response (due to the reduced number of samples).

For example, computing time for an N-point Fast Fourier Transform is proportional to $N \log_2 N$. If the complete time response is calculated using an 8192-point transform and the envelope using a 1024-point transform the computing times for the Fast Fourier Transform will

differ by a factor of 10.4. In addition, the computing times for the calculation of frequency domain samples will differ by a factor of 8, since there are 8 times the number of samples required for a complete time response.

For this reason program MINISEC calculates the envelope of the system time response using a 1024-point Fast Fourier Transform.

2.8 THE FREQUENCY DOMAIN COMPUTER MODEL

Program MINISEC is responsible for calculating the response of the total SAW label system based on frequency domain analysis.

The individual IDT weightings (i.e. the number of fingers for each IDT) and phasings (positive or negative connection) are input to the program. From this information the total transadmittance, input admittance and static capacitance for the SAW delay line are calculated using routines XAD, INPAD and STATCAP respectively. The label antenna values are calculated next. The inductance is calculated from the resonant frequency, the quality factor (Q=5) and a specified damping resistance (R = 50 ohms). To ensure resonance at 200 megahertz additional capacitance is added to the existing static capacitance of the delay line.

The label response is then calculated (based on equation 2.19) and multiplied by frequency functions for

a) the transmitter pulse,

b) the transmit antenna,

c) coupling from transmit antenna to label,

d) coupling from label to receive antenna,

and e) the receive antenna

to give the overall system frequency response (using routines INPULS and TUNCCT).

In preceding sections it has been seen that certain advantages arise from using the baseband equivalent response rather than the bandpass response. MINISEC translates the total system frequency response to its baseband equivalent and constructs a periodic frequency function necessary for the operation of the Fast Fourier Transform (see Figure 2.8). It should be noted that to construct this sampled interval accurately, the frequency response must be known over a wide bandwidth due to the overlapping of the periodic functions in the interval.

However, if the system is narrowband this overlapping is not significant and may be ignored. The resulting approximation requires the system frequency function to be calculated only over a bandwidth equal to the sampled interval. All values of the frequency response outside this interval are considered to be zero. This approximation is shown diagramatically in Figure 2.9 and is only useful if the system is sufficiently narrowband. Figure 2.10 shows that this is not an unreasonable approximation.

Figure 2.8 Periodic frequency function used for Fast Fourier Transform

Figure 2.9 Bandlimiting approximation used in calculations

Figure 2.10 System frequency response as calculated by MINISEC for the code of Figure 4.16

For example, the limits of the frequency interval in MINISEC are $200 \pm (512 \times 0.333)$ megahertz (i.e. 30 and 370 megahertz). At these points, the SAW delay line admittance (for 11-finger IDTs) is reduced by 39dB and the tuned circuits in the system reduce this further by 30dB and 39dB (30 and 370 megahertz respectively). Taking into account the 'ω' terms in the coupling (equations 2.14 and 2.15) and the transadmittance (equation 1.3), the total system response is reduced by 32dB and 118dB at 370 and 30 megahertz respectively. However the approximation becomes less valid as the system bandwidth increases which would occur if the number of fingers in any IDT was reduced.

Having obtained a suitable form of the baseband equivalent frequency function, the Fast Fourier Transform is used to transform to the time domain resulting in a complex baseband equivalent time response. The magnitude of this response is the envelope of the total system time response.

To simulate the integrations over the 25 nanosecond time windows used to generate the channel outputs of the receiver, MINISEC sums points in the complex baseband time function. The time windows are centred about the times at which pulses (corresponding to bits in the code) in the time response may occur. However due to the bandlimited system, these pulses are delayed by an amount of time (envelope delay). By inspection of plots produced by MINISEC this delay was found to be 43 nanoseconds. Allowing for this delay, the time windows were configured to sum 9 points

(window width = 9 × 2.93 = 26.4 nanoseconds) in the baseband time function around the expected pulse times. These numerical integrations simulate the outputs of the receiver channels from which the code bits are determined. MINISEC plots both the envelope and channel outputs as functions of time. Since the second order response is of particular interest, MINISEC also plots the envelope due to the second order response alone together with the corresponding channel outputs.

CHAPTER 3

TIME DOMAIN MODELLING

3.0 INTRODUCTION

In Chapter 2 the coded SAW label system was modelled by multiplying together a number of frequency responses representing parts of the system, to form a total system response in the frequency domain. The corresponding time response was derived by applying a Fast Fourier Transform to the frequency response. Theoretically this time response could also be obtained by convolving together the impulse responses of the sections of the system. However this approach is generally avoided due to the increase in computing time over the frequency domain and Fast Fourier Transform calculations.

Nevertheless, a time domain modelling approach using repeated convolutions has been developed, and by using several approximations, an efficient and fast computing algorithm for calculating the system time response has been realised. In particular, the algorithm gives an appreciation of the second order response which is not readily obtained from the frequency domain analysis.

3.1 SYSTEM TIME FUNCTIONS

In section 2.4 the frequency responses of the various parts which comprise the total system were examined. These bandpass functions were then multiplied together and the baseband equivalent was computed to give the total system response. The same result could have been achieved by taking the baseband equivalents of the system functions and then multiplying. This is because the baseband equivalent of a product of bandpass frequency functions is equal to the product of the baseband equivalents of the individual bandpass functions (see equation 2.28).

If the Fourier transforms of the baseband equivalents of the system frequency functions are taken, the result is the baseband equivalents of the system time functions. These time domain baseband equivalents contain the envelope and phase information of the system time functions as shown in section 2.5. More importantly, the baseband equivalent of the total system time response is the convolution of the baseband equivalent system time functions (see equation 2.29). Therefore to calculate the baseband equivalent (and hence the envelope) of the total system time response the nature of the baseband equivalents of the system time functions must be known. These time functions and their baseband equivalents are now examined.

3.1.1 Transmitter Pulse $\hat{g}_1(t)$

It has been shown previously that the transmitter pulse time function is

$$g_1(t) = \mathrm{rect}\left(\frac{t}{T} - \frac{1}{2}\right) \sin \omega_0 t \tag{3.1}$$

and the baseband equivalent is readily shown to be

$$\hat{g}_1(t) = -\frac{j}{2}\,\mathrm{rect}\left(\frac{t}{T} - \frac{1}{2}\right) \tag{3.2}$$

3.1.2 Transmit Antenna $\hat{g}_2(t)$

From the frequency function, $G_2(\omega)$, given in equation 2.12 the corresponding time equation is

$$g_2(t) = ae^{-bt} \cos(\omega_d t + \phi) \tag{3.3}$$

where
$$a = 2\omega_0/\sqrt{(4Q^2 - 1)}$$
$$b = \omega_0/2Q$$
$$\omega_d = \omega_0\sqrt{\left(1 - \frac{1}{4Q^2}\right)}$$
$$\phi = -\frac{\pi}{2} + \tan^{-1}\left(\frac{\omega_d}{b}\right)$$

and the baseband equivalent time function is

$$\hat{g}_2(t) = \frac{a}{2}\, e^{-bt}\, e^{j\phi} \cos(\omega_0 - \omega_d)t \tag{3.4}$$

However the duration of the exponential term is much less than the sinusoid whose value is close to unity

over the time period of interest. Hence equation 3.4 may be approximated by

$$\hat{g}_2(t) = \frac{a}{2} e^{-bt} e^{j\phi} \tag{3.5}$$

3.1.3 Receive Antenna $\hat{g}_3(t)$

The receive antenna is identical to the transmit antenna and hence

$$\hat{g}_3(t) = \hat{g}_2(t) \tag{3.6}$$

3.1.4 Coupling to Label $\hat{g}_4(t)$

Due to the frequency domain expression for $G_4(\omega)$ it is not possible to determine a meaningful expression for $g_4(t)$ and hence $\hat{g}_4(t)$. Rather than assign a time function for $\hat{g}_4(t)$, this part of the system is considered as a differential operator as suggested by the expression for $G_4(\omega)$. The consequences of this approach on the time domain system modelling calculations are discussed further in section 3.3.5.

3.1.5 Coupling from Label $\hat{g}_5(t)$

The approach adopted for $\hat{g}_4(t)$ is also used for $\hat{g}_5(t)$.

3.1.6 Label $\hat{g}_6(t)$

In section 2.4.6 the frequency domain function for the label was given by equation 2.19. In the time domain,

the function becomes

$$g_6(t) = \left(\frac{Z_t}{Z_s}\right)(t) * \left(\frac{Z_t}{Z_s}\right)(t) * \left(y(t) - y(t) * y(t) * Z_t(t)\right) \qquad (3.7)$$

A closer examination of equation 3.7 is needed before $g_6(t)$ may be determined. First consider the frequency function $G_7(\omega)$ where

$$G_7(\omega) = Z_t(\omega)/Z_s(\omega)$$

From Figure 2.5, $Z_t(\omega)$ is given by

$$Z_t = Z_s Z_c/(Z_s + Z_c) \qquad (3.8)$$

and therefore

$$G_7(\omega) = Z_c(\omega)/(Z_s(\omega) + Z_c(\omega))$$

or

$$G_7(\omega) = \frac{1}{1 - \left(\frac{\omega}{\omega_0}\right)^2 + j\left(\frac{\omega}{\omega_0}\right)\frac{1}{Q}} \qquad (3.9)$$

Note that $G_7(\omega)$ is similar to $G_2(\omega)$ and $G_3(\omega)$ and the time function shows a corresponding similarity.

$$g_7(t) = ae^{-bt}\cos(\omega_d t + \phi)$$

where $a = \omega_0/\sqrt{\left(1 - \frac{1}{4Q^2}\right)}$

$$b = \omega_0/2Q$$

$$\omega_d = \omega_0\sqrt{\left(1 - \frac{1}{4Q^2}\right)}$$

$$\phi = -\frac{\pi}{2}$$

Using the same approximation as in section 3.1.2, the baseband equivalent time function is

$$\hat{g}_7(t) = -j\,\frac{a}{2}\,e^{-bt} \tag{3.10}$$

Also the frequency function, $Z_t(\omega)$, is given from equation 3.8

$$Z_t(\omega) = \frac{R + j\omega L}{1 - \left(\frac{\omega}{\omega_0}\right)^2 + j\left(\frac{\omega}{\omega_0}\right)\frac{1}{Q}} \tag{3.11}$$

The time domain function, $z_t(t)$, also has the form

$$z_t(t) = ae^{-bt}\cos(\omega_d t + \phi)$$

$$\text{where} \quad b = \omega_0/2Q$$

$$\omega_d = \omega_0\sqrt{\left(1 - \frac{1}{4Q^2}\right)}$$

The expressions for the amplitude and phase constants (i.e. a and ϕ) are more involved than those for $g_2(t)$ and $g_7(t)$ but may be derived by superposition. However it is not necessary to evaluate these constants since amplitude and phase information is more easily derived from frequency domain functions as demonstrated in section 3.3.5.

It is important to note that the time functions $\hat{g}_2(t)$, $\hat{g}_3(t)$, $\hat{g}_7(t)$ and $\hat{z}_t(t)$ are all similar since they all display a tuned-circuit response. This similarity is exploited in section 3.3 when considering the different time responses which are combined to produce the total time response.

Finally the SAW delay line admittance, $Y(\omega)$, is examined. In the frequency domain,

$$Y(\omega) = \sum_{i=0}^{\infty} \sum_{j=0}^{\infty} Y_{ij}(\omega)$$

and in the time domain,

$$y(t) = \sum_{i=0}^{\infty} \sum_{j=0}^{\infty} y_{ij}(t)$$

The time domain admittance is the sum of both the time domain transadmittances and input admittances of all IDTs on the delay line. The transadmittance and input admittance both have distinctive time functions. For delay lines with identical IDTs, the transadmittance produces triangularly enveloped sinusoid pulses occurring at times determined by the spacing of IDTs on the substrate (see section 1.4). The input admittance produces half-triangular enveloped sinusoid pulses all occurring together near the time origin.

Thus the complete time domain delay line response comprises sinusoid pulses with either triangular or half-triangular envelopes. Hence $y(t)$ may be written as

$$y(t) = \frac{1}{2} b_0 y_i(t) + \sum_{k=1}^{\infty} b_k y_x(t-kT) \tag{3.12}$$

where $$y_x(t) = \frac{\omega_0 G_0}{\pi N} \Delta \left(\frac{\omega_0 t}{\pi N}\right) \cos \omega_0 t \tag{3.13}$$

$$y_i(t) = 2y_x(t) \quad \text{for } t \geqslant 0 \tag{3.14}$$

and $$\Delta(t) = 1 - |t| \quad \text{for } -1 \leqslant t \leqslant 1$$

$$= 0 \quad \text{otherwise}$$

and the b_k are coefficients yet to be determined but are dependent on the placement of IDTs on the substrate (see section 3.4.1).

The baseband equivalent of equation 3.12 is given by

$$\hat{y}(t) = \frac{1}{2} b_0 \hat{y}_i(t) + \hat{y}_x(t) * \sum_{k=1}^{\infty} b_k \delta(t-kT)(-1)^k \tag{3.15}$$

where $\hat{y}_i(t)$ and $\hat{y}_x(t)$ are the baseband equivalents of $y_i(t)$ and $y_x(t)$ of equations 3.13 and 3.14, and hence

$$\hat{y}_x(t) = \frac{\omega_0 G_0}{\pi N} \Delta \left(\frac{\omega_0}{\pi N}\right) \tag{3.16}$$

and $$\hat{y}_i(t) = 2\hat{y}_x(t) \quad \text{for } t \geqslant 0$$

The constant, G_0, is the conductance expression defined in section 2.2:

$$G_0 = \frac{-\pi^2 N^2 (\varepsilon_p + \varepsilon_0) |\omega_0| \Delta W}{13.75} \tag{3.17}$$

The $(-1)^k$ factor appearing in equation 3.15 is introduced when calculating baseband equivalents of delayed versions of $y_x(t)$. The transadmittance sequence of equation 3.12 is:

$$\sum_{k=1}^{\infty} b_k y_x(t-kT)$$

$$= y_x(t) * \sum_{k=1}^{\infty} b_k \delta(t-kT)$$

and in the frequency domain -

$$Y_x(\omega) \sum_{k=1}^{\infty} b_k e^{-jk\omega T}$$

translating to baseband -

$$Y_x(\omega+\omega_0) \sum_{k=1}^{\infty} b_k e^{-jk(\omega+\omega_0)T}$$

$$= \hat{Y}_x(\omega) \sum_{k=1}^{\infty} b_k e^{-jk\omega T} e^{-jk\omega_0 T}$$

and transforming back to the time domain -

$$\hat{y}_x(t) * \sum_{k=1}^{\infty} b_k e^{-jk\omega_0 T} \delta(t-kT) \qquad (3.18)$$

Examining the term $e^{-jk\omega_0 T}$ in equation 3.18, the product $\omega_0 T$ is evaluated as:

$$\omega_0 T = 15$$

(for $\omega_0 = 2\pi \times 200$ megahertz

and $T = 37.5$ nanoseconds)

and hence

$$e^{-jk\omega_0 T} = (-1)^k \qquad (3.19)$$

This term produces a phase reversal in alternate pulses, and is due to transducers being placed an integral number of half wavelengths apart on the substrate. (This effect follows from the discussion of transadmittance in Chapter 1, where a 'phase factor' associated with the transadmittance was a function of the distance between a pair of transducers).

3.1.7 Total System Response $\hat{g}_s(t)$

The baseband system time response, $\hat{g}_s(t)$, is the convolution of all individual baseband time responses comprising the system

$$\hat{g}_s(t) = \hat{g}_0(t) * [\hat{y}(t) - \hat{y}(t) * \hat{y}(t) * \hat{z}_t(t)] \tag{3.20}$$

where $$\hat{g}_0(t) = \hat{g}_1(t) * \hat{g}_2(t) * \hat{g}_3(t) * \hat{g}_4(t) * \hat{g}_5(t) * \hat{g}_7(t) * \hat{g}_7(t) \tag{3.21}$$

It can be seen that the total system response comprises a primary response, $\hat{y}(t)$, and a secondary response, $\hat{y}(t) * \hat{y}(t) * \hat{z}_t(t)$, convolved with a common time function $\hat{g}_0(t)$. Although the multiple convolutions involved in equations 3.20 and 3.21 would suggest a large amount of computation, an approximation can be used which reduces the computation considerably. This approximation is realised by a well known theorem in statistics; the Central Limit Theorem[9].

3.2 MULTIPLE CONVOLUTIONS AND THE CENTRAL LIMIT THEOREM

In statistics, a random variable is described by its characteristic function and its probability density function (p.d.f.) which is the inverse Fourier transform of the characteristic function. If a number of independent random variables are summed, the resultant random variable has a characteristic function which is the product of the individual characteristic functions. It follows that the overall p.d.f. is the convolution of individual p.d.f.s. Furthermore the mean and variance of the resultant random variable is equal to the sums of the means and variances respectively of the individual random variables. An analogy should be noted here between characteristic functions and frequency domain functions, and p.d.f.s and time domain functions.

The Central Limit Theorem states that a p.d.f. which is the convolution of a number of individual p.d.f.s tends to a Gaussian (or Normal) function as the number of individual p.d.f.s increases (regardless of the form of the p.d.f.s). Using the above analogy implies that the convolution of a number of time domain functions must approximate a Gaussian function. In order to completely specify a Gaussian function, the amplitude, mean and variance must be known. These may be calculated from the individual time functions since, for multiple convolutions, amplitudes (in this case, the areas under the functions) multiply together and the means and variances add.

Summarising, let there be m time functions:

$$h_i(t) \quad \text{for} \quad i = 1,2,\ldots m$$

Convolving these functions together gives:

$$h(t) = h_1(t) * h_2(t) * \ldots . * h_m(t)$$

The Central Limit Theorem states that h(t) is approximately Gaussian. The area under h(t) is H(0) where $H(\omega)$ is the Fourier transform of h(t). $H_i(\omega)$ are the Fourier transforms of $h_i(t)$ and

$$H(0) = H_1(0)\ H_2(0)\ \ldots\ldots\ H_m(0) \tag{3.22}$$

The means corresponding to each $h_i(t)$ are defined by

$$\mu_i = \frac{1}{H_i(0)} \int_{-\infty}^{\infty} t h_i(t)\, dt$$

and the mean of h(t) is given by

$$\mu = \sum_{i=1}^{m} \mu_i \tag{3.23}$$

The variances are defined by

$$\sigma_i^{\,2} = \frac{1}{H_i(0)} \int_{-\infty}^{\infty} (t-\mu_i)^2 h_i(t)\, dt$$

$$= \frac{1}{H_i(0)} \left[\int_{-\infty}^{\infty} t^2 h_i(t)\, dt - \mu_i^{\,2} \right]$$

and the variance of h(t) is

$$\sigma^2 = \sum_{i=1}^{m} \sigma_i^{\,2} \tag{3.24}$$

Since only H(0), μ and σ^2 are needed to specify h(t), all that must be known about the $h_i(t)$ functions are the statistics $H_i(0)$, μ_i and $\sigma_i^{\,2}$ for a good approximation to be determined for h(t). Thus a potentially complicated convolution may be simplified to a few arithmetic operations (to calculate H(0), μ and σ^2) providing that the statistics $H_i(0)$, μ_i and $\sigma_i^{\,2}$ are readily available.

The mean, μ, may be interpreted as the amount of delay introduced by the system of multiple convolutions and the variance, σ^2, as the "spread" of the resultant time function.

3.3 WAVEFORM STATISTICS (Figure 3.1)

Section 3.2 has shown that the convolution of many time functions produces an approximate Gaussian time function whose statistical properties may be determined from the statistics of its component time functions.

Fortunately, the baseband time functions which are convolved together to give the total system response are limited to four simple waveforms whose statistics are easily determined. These waveforms (and their statistics) are given below.

	AREA	MEAN	VARIANCE
Triangular	aT	0	$\frac{T^2}{6}$
Half-Triangular	$\frac{aT}{2}$	$\frac{T}{3}$	$\frac{T^2}{18}$
Exponential	$\frac{a}{b}$	$\frac{1}{b}$	$\frac{1}{b^2}$
Rectangular	aT	$\frac{T}{2}$	$\frac{T^2}{12}$

Figure 3.1 Waveform Statistics

3.3.1 Rectangular

The rectangular waveform arises from the transmitter pulse function, $\hat{g}_1(t)$.

$$\text{mean} = T/2$$

$$\text{variance} = T^2/12$$

where T = length of pulse.

For the transmitter pulse, T = 25 nanoseconds and hence:

mean = 12.5 nS variance = $(7.2\ \text{nS})^2$

3.3.2 Exponential

The exponential waveform is typical of all tuned circuit responses. These occur in the transmit/receive antenna as well as tuned circuits associated with the label (i.e. $\hat{g}_2(t), \hat{g}_3(t), \hat{g}_7(t), \hat{z}_t(t)$).

$$\text{mean} = 1/b$$

$$\text{variance} = 1/b^2$$

where b = exponential decay coefficient (nepers/second)

For a tuned circuit of known resonant frequency and Q then

$$b = \frac{\omega_0}{2Q} \qquad \text{(from section 3.1)}$$

Thus for $\omega_0 = 2\pi \times 200$ megahertz and a Q of 5,

$$b = 1.26 \times 10^8 \text{ nepers/second}$$

Hence mean = 8.0 nS and variance = $(8.0 \text{ nS})^2$.

3.3.3 Triangular

This waveform arises from the transadmittance between two identical IDTs. A series of these waveforms comprise the majority of the delay line admittance.

$$\text{mean} = 0$$

$$\text{variance} = T^2/24$$

where T = pulse width

For IDTs of 11 fingers, the transadmittance function has a duration of 11 cycles or 55 nanoseconds (for f_0 = 200 megahertz).

Thus variance = $(11.2 \text{ nS})^2$.

3.3.4 Half-Triangular

This waveform arises from the input admittance of an IDT. The waveform also contributes to the delay line admittance.

$$\text{mean} = T/3$$

$$\text{variance} = T^2/18$$

where T = pulse width

For an IDT of 11 fingers, the input admittance function has a duration of 27.5 nanoseconds.

Hence, mean = 9.2 nS and variance = $(6.5 \text{ nS})^2$.

3.3.5 Waveform Areas

The areas, $H_i(0)$, for each function are best determined from the bandpass frequency functions:

$$H_i(0) = \hat{G}_i(0) = G_i(\omega_0)$$

Thus the area under the baseband time function is equal to the bandpass frequency function evaluated at ω_0. The areas for functions used in the time domain computer model are given in Figures 3.2 and 3.3 together with the means and variances. Note that in general $H_i(0)$ are complex in which case the magnitude of $H_i(0)$ is the area and the phase of $H_i(0)$ is the carrier phase of the waveform.

3.3.6 Common System Function $\hat{g}_0(t)$

The common system function was defined in equation 3.21 and is

$$\hat{g}_0(t) = \hat{g}_1(t)*\hat{g}_2(t)*\hat{g}_3(t)*\hat{g}_4(t)*\hat{g}_5(t)*\hat{g}_7(t)*\hat{g}_7(t)$$

Using the Central Limit Theorem, $\hat{g}_0(t)$ is approximately Gaussian. The mean, variance and area of $\hat{g}_0(t)$ are calculated in accordance with equations 3.22, 3.23 and 3.24 respectively. Since no time functions were assigned to $\hat{g}_4(t)$ and $\hat{g}_5(t)$ these functions are considered to have zero mean and variance but an "area" equal to the frequency function evaluated at ω_0 as for other areas.

Function	Expression	Value
$\hat{g}_1(t)$		25×10^{-9}
$\hat{g}_2(t)$		1
$\hat{g}_3(t)$		1
$\hat{g}_4(t)$	$j\omega_0 M$	$j1.26 \times 10^9$
$\hat{g}_5(t)$	$j\omega_0 M$	$j1.26 \times 10^9$
$\hat{g}_7(t)$	$-jQ$	$-j5$
$\hat{g}_0(t)$		9.92×10^{11}
$\hat{y}_x(t)$	$\dfrac{-\pi^2 N^2(\varepsilon_p+\varepsilon_o)\lvert\omega_0\rvert\Delta W}{13.75}$	5.37×10^{-6}
$\hat{y}_i(t)$	"	5.37×10^{-6}
$\hat{z}_t(t)$	Q^2R (approx.)	1250

Figure 3.2 Areas for functions in the prototype 200 MHz system

Function	Area	Mean (nS)	Variance (nS^2)
$\hat{g}_1(t)$	25×10^{-9}	12.5	$(7.2)^2$
$\hat{g}_2(t)$	1	8.0	$(8.0)^2$
$\hat{g}_3(t)$	1	8.0	$(8.0)^2$
$\hat{g}_4(t)$	$j1.26 \times 10^9$	-	-
$\hat{g}_5(t)$	$j1.26 \times 10^9$	-	-
$\hat{g}_7(t)$	$-j5$	8.0	$(8.0)^2$
$\hat{g}_0(t)$	9.92×10^{11}	44.5	$(17.6)^2$
$\hat{y}_x(t)$	5.37×10^{-6}	0	$(11.2)^2$
$\hat{y}_i(t)$	5.37×10^{-6}	9.2	$(6.5)^2$
$\hat{z}_t(t)$	1250	8.0	$(8.0)^2$

Figure 3.3 Waveform statistics for functions in the prototype 200 MHz system

3.4 PRIMARY AND SECONDARY RESPONSES

The primary and secondary responses are due to terms appearing in equation 3.20, namely

$$\text{Primary} : \hat{p}(t) = \hat{y}(t)$$

$$\text{Secondary} : \hat{s}(t) = \hat{y}(t) * \hat{y}(t) * \hat{z}_t(t)$$

3.4.1 Primary Response

The primary response is defined by equation 3.15 as

$$\hat{p}(t) = \hat{y}(t) = \frac{1}{2} b_0 \hat{y}_i(t) + \hat{y}_x(t) * \sum_{k=1}^{\infty} b_k \delta(t-kT)(-1)^k \qquad (3.25)$$

where the b_k are coefficients which are, as yet, unknown. To determine these coefficients, consider the placement of identical IDTs on a substrate (as in Figure 1.12). The IDTs comprise the transmit IDT and other IDTs (the coding set) placed at integer multiples of D away from the transmit IDT (corresponding to multiples of T = D/v in the time domain).

The total source function at the surface of the substrate may be expressed as

$$f_t(x) = f(x) * \sum_{k=0}^{\infty} a_k \delta(x-kD) \qquad (3.26)$$

where f(x) = source function of a single IDT and $a_k = 1$ if an IDT exists at a distance of kD from the transmit IDT, and $a_k = 0$ if no IDT exists at that distance. The transmit IDT corresponds to k = 0 and hence $a_0 = 1$.

As the time and displacement domains are related by $x = vt$, equation 3.25 may be rewritten as

$$f_t(vt) = f(vt) * \sum_{k=0}^{\infty} a_k \ (t-kT)$$

From previous expressions for transadmittance and input admittance the delay line admittance may be written as

$$y(t) = 2K \ f_t(vt) * f_t(-vt) \qquad \text{for } t \geqslant 0$$

$$\text{where} \quad K = \frac{-|\omega| \Delta\varepsilon_o{}^2 W}{(\varepsilon_p + \varepsilon_o)}$$

and rewritten as

$$y(t) = K \ f(vt) * f(-vt) * \sum_{k=0}^{\infty} b_k \delta(t-kT) \qquad t \geqslant 0$$

and in the baseband

$$\hat{y}(t) = K \ \hat{f}(vt) * \hat{f}(-vt) * \sum_{k=0}^{\infty} b_k \delta(t-kT)(-1)^k \qquad t \geqslant 0 \qquad (3.27)$$

Thus the b_k are given by

$$b_k = 2 \sum_{i=0}^{\infty} a_i a_{i+k} \qquad k \geqslant 0 \qquad (3.28)$$

The sequence b_k is the autocorrelation of the sequence a_k for $k \geqslant 0$. In fact equation 3.27 is identical to equation 3.15.

3.4.2 Secondary Response

Using equation 3.15, the secondary response, $\hat{s}(t)$, is given by

$$\hat{s}(t) = \hat{p}(t)*\hat{p}(t)*\hat{z}_t(t)$$

$$= \hat{z}_t(t)*\Big[\frac{1}{2}\, b_0\hat{y}_i(t)*\hat{y}_x(t) * \sum_{k=0}^{\infty} b_k\delta(t-kT)(-1)^k$$

$$+ \hat{y}_x(t)*\hat{y}_x(t) * \sum_{k=1}^{\infty} c_k\delta(t-kT)(-1)^k$$

$$+ \frac{1}{4}\, b_0{}^2\hat{y}_i(t)*\hat{y}_i(t)\Big] \qquad (3.29)$$

and it may be shown that

$$c_k = \sum_{i=1}^{\infty} b_i b_{k-i} \qquad k \geqslant 1 \qquad (3.30)$$

3.5 TOTAL SYSTEM RESPONSE

The total system response (in baseband) is defined in equation 3.20 and may now be written as

$$\hat{g}_s(t) = \hat{g}_0(t) * (\hat{p}(t) - \hat{s}(t))$$

where $\hat{g}_0(t)$, $\hat{p}(t)$ and $\hat{s}(t)$ have been defined in equations 3.21, 3.25 and 3.29 respectively.

Substituting for $\hat{p}(t)$ and $\hat{s}(t)$ yields

$$\hat{g}_s(t) = [\tfrac{1}{2}b_o\hat{p}_o(t)$$

$$+ \hat{p}_1(t) * \sum b_k\delta(t-kT)(-1)^k$$

$$- \tfrac{1}{4}b_o{}^2\hat{s}_o(t)$$

$$- \hat{s}_1(t) * \sum\tfrac{1}{2}b_o b_k\delta(t-kT)(-1)^k$$

$$- \hat{s}_2(t) * \sum c_k\delta(t-kT)(-1)^k] \qquad (3.31)$$

where $\hat{p}_0(t) = \hat{g}_0(t)*\hat{y}_i(t)$ (3.32)

$$\hat{p}_1(t) = \hat{g}_0(t)*\hat{y}_x(t) \qquad (3.33)$$

$$\hat{s}_0(t) = \hat{g}_0(t)*\hat{y}_i(t)*\hat{y}_i(t)*\hat{z}_t(t) \qquad (3.34)$$

$$\hat{s}_1(t) = \hat{g}_0(t)*\hat{y}_i(x)*\hat{y}_x(t)*\hat{z}_t(t) \qquad (3.35)$$

$$\hat{s}_2(t) = \hat{g}_0(t)*\hat{y}_x(t)*\hat{y}_x(t)*\hat{z}_t(t) \qquad (3.36)$$

Thus the total system response is comprised of sequences of 5 different pulse waveforms (as defined in equations 3.32 to 3.36). According to the central limit theorem each of these waveforms is approximately Gaussian but with slightly different statistics. These waveforms are now examined separately.

3.5.1 Primary Zero Waveform $\hat{p}_0(t)$

This waveform is due to the input admittance of IDTs on the SAW delay line.

	Component Function	Area	Mean (nS)	Variance $(nS)^2$
	$\hat{g}_0(t)$	9.92×10^{11}	44.5	$(17.6)^2$
	$\hat{y}_i(t)$	5.37×10^{-6}	6.5	$(6.5)^2$
Total	$\hat{p}_0(t)$	5.33×10^{6}	51.0	$(18.2)^2$

3.5.2 Primary One Waveform $\hat{p}_1(t)$

This waveform forms the part of the primary response containing the required binary code (due to sequence b_k).

	Component Function	Area	Mean (nS)	Variance $(nS)^2$
	$\hat{g}_0(t)$	9.92×10^{11}	44.5	$(17.6)^2$
	$\hat{y}_x(t)$	5.37×10^{-6}	0	$(11.2)^2$
Total	$\hat{p}_1(t)$	5.33×10^{6}	44.5	$(20.9)^2$

3.5.3 Secondary Zero Waveform $\hat{s}_0(t)$

This waveform results from the auto-convolution of the IDT input admittance.

	Component Function	Area	Mean (nS)	Variance $(nS)^2$
	$\hat{g}_0(t)$	9.92×10^{11}	44.5	$(17.6)^2$
	$\hat{y}_i(t)$	5.37×10^{-6}	9.2	$(6.5)^2$
	$\hat{y}_i(t)$	5.37×10^{-6}	9.2	$(6.5)^2$
	$\hat{z}_t(t)$	1250	8	$(8)^2$
Total	$\hat{s}_0(t)$	3.58×10^4	70.9	$(21.4)^2$

3.5.4 Secondary One Waveform $\hat{s}_1(t)$

Similar to the primary one waveform but smaller in magnitude and delayed slightly due to $\hat{y}_i(t)$ and $\hat{z}_t(t)$.

	Component Function	Area	Mean (nS)	Variance $(nS)^2$
	$\hat{g}_0(t)$	9.92×10^{11}	44.5	$(17.6)^2$
	$\hat{y}_i(t)$	5.37×10^{-6}	9.2	$(6.5)^2$
	$\hat{y}_x(x)$	5.37×10^{-6}	0	$(11.2)^2$
	$\hat{z}_t(t)$	1250	8	$(8)^2$
Total	$\hat{s}_1(t)$	3.58×10^4	61.7	$(23.3)^2$

3.5.5 Secondary Two Waveform $\hat{s}_2(t)$

This waveform, convolved with the sequence c_k (see equation 3.31), is responsible for the main interference with the desired primary response.

	Component Function	Area	Mean (nS)	Variance $(nS)^2$
	$\hat{g}_0(t)$	9.92×10^{11}	44.5	$(17.6)^2$
	$\hat{y}_x(t)$	5.37×10^{-6}	0	$(11.2)^2$
	$\hat{y}_x(t)$	5.37×10^{-6}	0	$(11.2)^2$
	$\hat{z}_t(t)$	1250	8	$(8)^2$
Total	$\hat{s}_2(t)$	3.58×10^4	52.5	$(25.0)^2$

3.5.6 Combining Waveforms

Since it is now known that the five waveforms (i.e. $\hat{p}_0(t), \hat{p}_1(t), \hat{s}_0(t), \hat{s}_1(t), \hat{s}_2(t)$) are approximately Gaussian, only the coefficients b_k and c_k which determine the impulse sequences in equation 3.31 need to be found for $\hat{g}_s(t)$ to be calculated. From section 3.4.1, $a_k = 1$ if an IDT exists at the kth position on the delay line and $a_k = 0$ if not. It is also true that $a_k = -1$ if the IDT exists and connected in reversed electrical polarity. If all IDTs on the delay line are identical (although some may be connected in reversed polarity), then each a_k will have one of 3

values, viz. -1, 0 or +1. Therefore the transducer placements and phasings (positive or negative polarity) are described by the a_k coefficients. The coefficients b_k and c_k are derived from a_k by using equations 3.28 and 3.30.

When the b_k and c_k are known, all impulse sequences in equation 3.31 are defined and, after convolution with the appropriate waveform functions and summed, give the required total system response, $\hat{g}_s(t)$. It is seen from equation 3.31 that the important contributions to the total response come from the primary one, secondary one and secondary two responses. The primary zero and secondary zero responses occur close to the time origin and hence have no influence on the code which occurs later. For this reason the primary zero and secondary zero responses are ignored in the following analyses.

On closer examination, the impulse sequences responsible for the primary one, secondary one and secondary two responses all contribute to the total response at times when code bits occur (i.e. k = 21 to 40 inclusive for a 20-bit code system). The primary one response, due to the b_k sequence, contains the desired binary code (b_k for k = 21 to 40). The secondary one response is due to a sequence which is essentially the same as for the primary one response except that it is multiplied by a factor of $\frac{1}{2}b_0$. The secondary two response is due to the c_k sequence (the auto-convolution of the b_k sequence) and is capable of producing most

interference with the desired code. Although the waveform functions, $\hat{p}_1(t)$, $\hat{s}_1(t)$ and $\hat{s}_2(t)$, determine the shape of the pulses in the total response, it is the b_k and c_k sequences which are important when considering the composition and magnitude of the response.

From section 3.5, the magnitudes (areas) of the secondary waveforms are a factor of 149 less than the primary waveforms. This factor applies only for the prototype 200 MHz system with 11-finger IDTs and is a function of transducer characteristics and the label tuned circuit. For certain codes the magnitude of some impulses in the sequences corresponding to the secondary responses may approach or even exceed the factor of 149. At this stage the magnitude of the secondary responses are equal to or greater than the desired primary response. This situation should be avoided if severe corruption of the code is to be suppressed. Several strategies for minimising the magnitude of the secondary response are discussed in Chapter 5.

3.5.7 Non-identical transducers

In the previous section it was stated that the a_k coefficients describe the placement and phasing of transducers on a substrate. Implicit in this description and also in the determination of waveforms in previous sections is the assumption that all transducers are identical (11 fingers each).

The calculation of the system response using delay lines composed of non-identical (varying numbers of fingers) transducers is an extension of the case for identical numbers of fingers, the time domain trans-admittance function will not be triangular in form but, in general, will be trapezoidal depending on the resultant convolution of source functions. A common transadmittance function, $\hat{y}_x(t)$, cannot be used since the transadmittance of various transducer pairs will differ. The same argument is true for input admittances. Hence equations 3.31 to 3.36 do not strictly hold for delay lines with non-identical transducers. The shape and areas of admittance waveforms depend directly on the number of fingers in the transducers. The area under a baseband transadmittance function is proportional to the product of the numbers of fingers in the pair of transducers. Likewise the area of a baseband input admittance is proportional to the square of the number of fingers in the transducer.

The Central Limit Theorem states that any pulse waveform arising in the system response will be approximately Gaussian. It is also true that the shapes of these Gaussian waveforms are primarily dictated by the common system function, $\hat{g}_0(t)$, which is not dependent on delay line characteristics. Therefore the nature of the input admittance and transadmittance waveforms has little effect on pulse shape and positioning (i.e. on the 'variance' and 'mean' of the waveforms) but the magnitude of the pulses is highly dependent on the numbers of fingers in the

transducers and will affect the magnitude of pulses in the system response, but have only a small influence on pulse shapes and positioning. For this reason, the waveforms defined in equations 3.32 to 3.36 are assumed to be valid for delay lines using non-identical transducers. The system response will be given by equation 3.31. Coefficients b_k and c_k will be calculated according to equations 3.28 and 3.30, but the a_k coefficients will now contain information about the numbers of fingers in each transducer. The magnitude of the a_k coefficients will be given as

$$|a_k| = (\text{no. of fingers in } k^{th} \text{ IDT})/11 \ .$$

The sign of a_k will depend on the phasing of the IDT, as before. By assigning values in this way, the a_k coefficients are proportional to the areas of IDT source functions (in baseband). The admittances, which arise through convolutions of source functions, have areas which are products of areas of source functions. These products are contained in equations 3.28 and 3.30 which define appropriate values for b_k and c_k, thereby governing the magnitudes of pulses in the total response.

3.5.8 Phasing of transducers

In section 3.1.6 (and in particular, equation 3.15) it was demonstrated that alternate pulses in the code were antiphase for the prototype system. This was due to alternate transducer locations being placed an integral number of half wavelengths apart on the substrate. Thus,

although all transducers may be connected in the same electrical sense, some transducers produce pulses which are antiphase relative to others. This 'antiphase' effect manifests itself as a reversal of phase of the 'carrier' part of the pulses.

The effective phase of a transducer, relative to another transducer, may be varied by changing the position of the transducer on the substrate as described in previous sections. In the time domain, the corresponding time shift of a pulse has little effect on the envelope (for small shifts) but varies the carrier phase significantly. A phase shift of the carrier also results in a phase change in the 'area' of the corresponding baseband function as defined in Chapter 2. Thus it is possible to vary the phase of pulses by appropriate positioning of IDTs. This is the basis for phase encoded delay lines.

Phase encoding may be incorporated into the time domain model by assigning complex values to the a_k coefficients. The magnitude of a_k will depend on the number of fingers in an IDT (as before), while the phase of a_k will be the sum of the phase of the IDT relative to the transmit IDT and the phase due to the IDT's electrical connection. If this method is adopted, the $(-1)^k$ term of equation 3.15 would become redundant.

Considering the prototype 200 MHz system, the choice of 37.5 nanoseconds for the code bit spacing has inadvertently phase encoded (i.e. alternately antiphase)

the code bits. In general, the phase encoding of pulses (in the primary response) is not important in a system which only examines pulse amplitudes, such as the prototype system. However phasing of the pulses is very important when considering the secondary response, as will be demonstrated in Chapter 5.

3.6 INTERSYMBOL INTERFERENCE

Due to the bandlimited nature of the system, pulses which occur are spread over relatively large time periods. Therefore a pulse occurring at a given time (representing a bit in the code) may extend into time windows used for assessing the other code bits. In this way, adjacent pulses interfere with each other. From Figure 3.4 it is seen that the amount of interference depends on the relative location and spreading of the pulses. Since the channel outputs are proportional to the areas under the envelope of the received signal in corresponding time windows, the amount of pulse area overlapping into adjacent time windows is a measure of the interference.

For the purpose of the computer model, it is considered that a pulse only interferes with time windows (i.e. channels) immediately preceding and following its own window. Fractions of the pulse weight (area) are therefore distributed over the three contiguous channels (instead of one in the ideal situation of no intersymbol interference).

AREA RATIOS

primary 1	0.11	0.45	0.11
secondary 2	0.08	0.37	0.20
secondary 1	0.03	0.32	0.29

Figure 3.4 Intersymbol Interference for various waveforms

The areas from different pulses in a given time window add vectorially (including the "phase" of the area which contains the carrier phase information) to produce a nett channel output. It has already been stated in section 3.1.6 that the weighting of pulses occurring from adjacent code bits is antiphase due to bit spacing. This results in destructive interference between adjacent pulses. The ratios showing how pulse areas are distributed over adjacent time windows are also shown in Figure 3.4 for waveforms $\hat{p}_1(t)$, $\hat{s}_1(t)$ and $\hat{s}_2(t)$.

The three main components of the coded reply signal have waveforms characterised by $\hat{p}_1(t)$, $\hat{s}_1(t)$ and $\hat{s}_2(t)$ and sequences governed by the b_k and c_k in equation 3.31. Each of these three components contributes to the channel outputs.

Considering the primary one response, which comprises the primary one waveform $\hat{p}_1(t)$ convolved with a series of impulses weighted by coefficients b_k, a different method of representation can be used. If the weighted series of impulses are multiplied by the area of the waveform, $\hat{p}_1(t)$, and the resultant weighted impulses treated as data samples, then a channel output is derived from a given sample in addition to the preceding and following samples. In this way, intersymbol interference is modelled as a simple transversal digital filter. The input samples to the filter represent the pulses in the primary one response and the output of the filter represents the resultant channel outputs. The filter coefficients are simply the

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ratios shown in Figure 3.4. Likewise, digital filters can also be constructed to determine the channel outputs resulting from the secondary one and secondary two responses. The digital filter implementation is shown in Figure 3.5. Note that, due to the causality constraint on the filter, the outputs are delayed by one sample period in relation to the corresponding input samples.

3.7 COMPUTER IMPLEMENTATION

Based on concepts described in this chapter, FORTRAN program TIMDOM has been developed to interactively model a 200 MHz system with a capability of up to 30 code bits. TIMDOM requires information about the number of fingers and electrical polarity (i.e. the a_k coefficients) of each IDT in the coding set. The output gives the magnitudes of the receiver channel outputs, displayed numerically and in bar-graph form for easy interpretation.

TIMDOM assumes that the transmit IDT is an 11-finger IDT and that the spacing between the transmit IDT and the coding set conforms to Figure 1.12. All other system parameters have already been described in previous sections.

From the information about transducers and their placement on the substrate, TIMDOM sets up the a_k coefficients and adjusts the phases according to the $(-1)^k$ term of section 3.1.6. Coefficients b_k and c_k are then calculated and multiplied (weighted), as appropriate, by the areas of waveforms $\hat{p}_1(t)$, $\hat{s}_1(t)$ and $\hat{s}_2(t)$ (normalised

	a_2	a_1	a_0
Primary One	0.11	0.45	0.11
Secondary One	0.03	0.32	0.29
Secondary Two	0.08	0.37	0.20

Figure 3.5 Digital filter representation of intersymbol interference

to $\hat{p}_1(t)$) to produce sequences representing pulses in the primary one, secondary one and secondary two responses. These three sequences are passed through digital transversal filters to simulate intersymbol interference and then summed to produce the receiver channel outputs.

At present, TIMDOM does not have the capability of handling complex coefficients as required for a general phase encoded system. However, due to the relative simplicity of the program and its arithmetic operations, TIMDOM may easily be modified to cater for complex coefficients. At present, a 915 MHz phase encoded system is being developed in this Department and modifications to TIMDOM may be useful for modelling this system.

CHAPTER 4

RESULTS FROM THE PROTOTYPE SYSTEM AND MODELS

4.0 INTRODUCTION

This chapter covers the comparison of results obtained from the prototype 200 MHz system and results predicted by the two computer models, MINISEC (frequency domain) and TIMDOM (time domain). Unfortunately the results from the prototype system are limited since only three coded labels exist for this system. Furthermore, the codes which are available do not produce significant second order responses and hence cannot be used to verify the substantial second order responses predicted by the models for some codes.

4.1 PROTOTYPE SYSTEM CONFIGURATION

In addition to the main system blocks described in Chapter 1 which characterise the system response, the complete prototype system contains several other components.

Signals used for timing operations in the transmitter and receiver are produced by the timing pulse generator. The transmitter, receiver and timing pulse generator are incorporated into the interrogate unit (Figure 4.1). Overall system control is achieved by the use of a micro-

computer controller (and corresponding software) which both serves as an operator interface to the system and as a processor of channel outputs. Code extraction and determination of system signal levels are performed by the controller.

By communicating to the controller via a terminal, an operator may change various parameters of system operation. After the interrogation of a codel label by the system, the controller produces an output at the terminal comprising channel output levels, system signal levels (MSL, ZUL, ULL and UUL) and any error messages.

Figure 4.2 shows the hand held "wand" antenna which is used as both transmit and receive antennae in the prototype system.

4.1.1 Receiver Limitations

Although the prototype system was intended to have a capability of 20 code bits, only ten mixers have been incorporated into the receiver. The mixers and timing pulse generator have been configured such that the mixers correspond to the first ten bits in a 20-bit code. In addition to this limitation only 8 low frequency processors have been built into the receiver. Thus the prototype receiver is capable of producing outputs from only 8 channels.

However the 8 low frequency processors may be connected to the mixers in any permutation. The interconnections have been chosen such that complete receiver

Figure 4.1 The prototype system. From left, DECwriter terminal, controller and interrogate unit. Note wand antenna and label on top of controller.

Figure 4.2 Hand-held transmit/receive "wand" antenna

channels exist for code bits which are considered to be of most interest. In the prototype system, only channels 0, 1, 2, 3, 5, 7, 8 and 9 are operational (Figure 4.3).

4.1.2 Prototype Label Codes

Due to receiver limitations, the SAW delay lines were constructed to the general specifications required for a 20-bit code but only the first 10 code bits were considered for use.

The allocation of the 10 code bits was chosen to be -

1 0 0 X 0 Y 0 Z 0 1

where X, Y and Z denote either a '1' or '0', thus giving a range of 8 possible codes. Only 3 delay lines were produced and the codes are -

1 0 0 0 0 0 0 0 0 1

1 0 0 0 0 1 0 0 0 1

1 0 0 1 0 0 0 1 0 1

The delay lines with these codes were incorporated into labels for evaluation with the prototype system. Note that the codes were chosen such that 'ones' in the code did not correspond to non-existent receiver channels 4 and 6.

Figure 4.3 Receiver channel configuration for Prototype System

4.2 SUMMARY OF RESULTS

4.2.1 Prototype

System performance for the 3 coded labels has been investigated. The corresponding channel outputs (as output by the controller) appear in Figures 4.6, 4.10 and 4.14. To enable comparisons with envelopes predicted by MINISEC, oscilloscope traces were taken for the received reply signal (obtained from the output of the gated r.f. amplifier in the receiver). The traces are shown in Figures 4.4, 4.8 and 4.12. The reply for the code depicted in Figure 4.12 shows a reduced magnitude for one of the code bits. It is assumed that, since the delay line for that label was not hermetically sealed, corrosion of fingers for one of the transducers has occurred.

4.2.2 MINISEC

MINISEC has been used to model the performance of the system for the 3 coded labels. Since one of the labels appears damaged, the transducer corresponding to the reduced amplitude code bit has been modelled by assuming a 5-finger transducer (instead of the normal 11 fingers).

The envelopes of system responses are shown in Figures 4.4, 4.8 and 4.12. MINISEC also produces outputs corresponding to the receiver channel outputs (Figures 4.7, 4.11 and 4.15) and the envelopes of system second order responses (Figures 4.5, 4.9 and 4.13).

4.2.3 TIMDOM

System responses in the time domain have been modelled by TIMDOM. As for MINISEC, the damaged transducer of one label has been modelled as a 5-finger transducer. TIMDOM only predicts receiver channel outputs as shown in Figures 4.6, 4.10 and 4.14.

4.2.4 Damaged Label

It appears that damage to one of the prototype labels has occurred only recently. In earlier system testing by Dr. P.H. Cole, an output from the system corresponding to the label in question was obtained. This output showed no apparent damage. The prototype system output for this label and predictions by MINISEC and TIMDOM are shown in Figures 4.16, 4.17, 4.18 and 4.19.

Figure 4.4 System time response (MINISEC and prototype) for label code 1000000001

Figure 4.5 Secondary time response (MINISEC) for label code 1000000001

```
                    TABLE AND BAR GRAPH

     MSL = 3461 ZUL = 1581 ULL = 1906 UUL = 5205

                                   Z     U                          M
 0   3385  ************************************************************
 1   1305  **********************
 2   0185  ***
 3   0010
 4   0000
 5   0075  *
 6   0000
 7   0038
 8   1153  *******************
 9   3537  ***************************************************************


CHANNEL OUTPUTS

      .99  ********************
      .23  *****
     0.00
     0.00
     0.00
     0.00
     0.00
     0.00
     -.23  *****
     -.98  ********************
```

Figure 4.6 Channel outputs from prototype (top) and as predicted by TIMDOM for label code 1000000001

Figure 4.7 Simulated channel outputs (MINISEC) for label code 1000000001

Figure 4.8 System time response (MINISEC and prototype) for label code 1000010001

Figure 4.9 Secondary time response (MINISEC) for label code 1000010001

```
                    TABLE AND BAR GRAPH

     MSL = 3026 ZUL = 1382 ULL = 1666 UUL = 4550

                                        Z     U                          M
 0   3060  ************************************************************
 1   1019  ********************
 2   0151  **
 3   0054  *
 4   0000
 5   2969  ************************************************************
 6   0000
 7   0129  **
 8   1333  **************************
 9   2992  ************************************************************


CHANNEL OUTPUTS

      .99  ********************
      .23  *****
     0.00
     -.00
     -.24  *****
     -.99  ********************
     -.22  ****
     0.00
     -.23  *****
     -.97  ********************
```

Figure 4.10 Channel outputs from prototype (top) and as predicted by TIMDOM for label code 1000010001

Figure 4.11 Simulated channel outputs (MINISEC) for label code 1000010001

Figure 4.12 System time response (MINISEC and prototype) for label code 1001000101 (damaged label)

Figure 4.13 Secondary time response (MINISEC) for label code 1001000101 (damaged label)

TABLE AND BAR GRAPH

MSL = 3270 ZUL = 1494 ULL = 1801 UUL = 4917

```
                                  Z     U                          M
0   3247  ********************************************************
1   1069  *****************
2   0245  ****
3   3547  ************************************************************
4   0000
5   0081  *
6   0000
7   1683  ****************************
8   0580  *********
9   3293  ********************************************************
```

CHANNEL OUTPUTS

```
 .99  ********************
 .23  *****
-.24  *****
-.98  ********************
-.23  *****
-.00
-.12  **
-.45  *********
-.33  *******
-.97  ********************
```

Figure 4.14 Channel outputs from prototype (top) and as predicted by TIMDOM for label code 1001000101 (damaged label)

Figure 4.15 Simulated channel outputs (MINISEC) for label code 1001000101 (damaged label)

Figure 4.16 System time response (MINISEC) for label code 1001000101

Figure 4.17 Secondary time response (MINISEC) for label code 1001000101

```
                    TABLE AND BAR GRAPH

     MSL = 1474 ZUL = 0443 ULL = 1180 UUL = 1767

                              Z                                   U          M
 0   1474  ************************************************************
 1   0270  **********
 2   0121  ****
 3   1485  ************************************************************
 4   0000
 5   0000
 6   0090  ***
 7   1406  ************************************************************
 8   0322  ************
 9   1402  ************************************************************


CHANNEL OUTPUTS

      .99  ********************
      .22  ****
     -.24  *****
     -.99  ********************
     -.23  *****
     -.00
     -.25  *****
     -.98  ********************
     -.44  *********
     -.96  ********************
```

Figure 4.18 Channel outputs from prototype (top) and as predicted by TIMDOM for label code 1001000101

Figure 4.19 Simulated channel outputs (MINISEC) for label code 1001000101

4.3 DISCUSSION OF RESULTS

4.3.1 Second Order Responses

System responses in the prototype and both models show that second order responses are small for the codes examined. However slight reductions in pulse magnitudes for code bits are noticed. This is due to the fact that secondary response pulses have the same phase as the primary response pulses but are subtracted as shown in equation 3.20.

Results from the prototype system suggest that the second order response is larger than predicted by the computer models (comparing waveforms in Figures 4.4, 4.8 and 4.12). The larger secondary pulses in the prototype system are believed to occur from reflections due to transducer metallisation effects. These reflections are caused by discontinuities in wave propagation due to the metal fingers in transducers and are in addition to reflections caused by electro-acoustic regeneration. The computer models do not calculate responses due to metallisation effects. Further discussion on this effect is contained in section 5.2.

Since the codes used in the prototype system are very sparse (i.e. do not contain many 'ones'), most of the a_k coefficients will be zero. When correlation and convolution occur to produce the b_k and c_k coefficients, there is little possibility of any of these coefficients

reaching large magnitudes. Therefore for sparse codes the second order response is generally low in magnitude. For this reason the prototype labels do not suffer significantly from second order responses and hence are not good examples of codes which are corrupted by second order effects.

4.3.2 Waveforms

Previous discussion (Chapter 3) has shown that pulses in the system response should be approximately Gaussian. The 'variance' of a pulse can be interpreted as a measure of the 'spread' or 'width' of the pulse. For the purpose of this analysis, the pulse width can be defined as the width of the pulse at points which are 61% of the pulse peak. For a Gaussian waveform the width would be equal to twice the standard deviation (Standard deviation = $\sqrt{\text{variance}}$).

The widths of primary response pulses were measured for the prototype system and compared to values predicted by the computer models (assuming negligible second order responses). Results are -

i)	Prototype	:	42	nanoseconds
ii)	MINISEC	:	40	nanoseconds
iii)	TIMDOM	:	41.8	nanoseconds

Good agreement is noted between predicted and actual pulse widths. However it has been found that the transmitter pulse is not a rectangular sinusoid as assumed in

computer models. The actual transmitter pulse of the prototype system is shown in Figure 4.20.

It would be more realistic to model the transmitter pulse as having a triangular envelope of 50 nanoseconds duration. This would give the transmitter pulse a variance of $(10.2\ ns)^2$ rather than $(7.2\ ns)^2$ as previously assumed. The effect is to slightly increase the predicted pulse widths (e.g. from 41.8 to 42.4 nanoseconds for the time domain model).

Nevertheless a good general agreement of pulse widths is maintained. This emphasises the point that the actual shape of a waveform is not particularly critical since it is the waveform "statistics" which govern the pulse widths (and delays) in the total response.

The time domain model also predicts a delay (i.e. the mean) for pulse waveforms. The delay predicted for primary pulses by MINISEC has been measured from plots and is 43 nanoseconds. The time domain model predicts 44.5 nanoseconds. Unfortunately actual delays existing in the prototype system are difficult to determine due to variations between labels as discussed in the next section.

4.3.3 System Timing

The channel outputs rely heavily on correct receiver timing. The time windows corresponding to receiver channels (i.e. the gating times of receiver mixers) must be correctly positioned to ensure the correct channel outputs. The

Figure 4.20 Prototype system transmitter pulse

timing of the channels is constant with respect to each other (i.e. 37.5 nanoseconds apart and 25 nanoseconds wide) but may be varied relative to the transmitter pulse. This variation is needed to compensate for delays introduced by the system as already discussed. Figure 4.21 shows how variations in channel timing may affect the channel outputs and hence the perceived code.

Correct receiver timing can only be attained if code pulses occur repeatedly at their nominal times. Figure 4.22 shows the pulse corresponding to the first code bit for the 3 prototype labels relative to a time reference which is fixed with respect to the transmitter pulse. It can be seen that the time position of the pulses varies over almost a 20 nanosecond range. This inconsistency in the timing of the code bits is attributed to inaccuracies in the placement of coding set IDTs on the substrate. Since variations in channel outputs due to timing variations are comparable in magnitude to the effects of second order responses, it is considered that the prototype channel outputs are not a reliable indication of second order responses. For this reason, the waveforms of Figures 4.4, 4.8 and 4.12 have been a more useful comparison between the prototype and computer models. Nevertheless, prototype channel outputs are compared with those predicted by the computer models for completeness.

Nominally correct receiver timing

```
                    TABLE AND BAR GRAPH

MSL = 3478 ZUL = 1588 ULL = 1914 UUL = 5227

                                   Z    U                          M
0   3545  ************************************************************
1   1059  *****************
2   0145  **
3   0076  *
4   0000
5   3533  ************************************************************
6   0000
7   0132  **
8   1442  ************************
9   3408  ************************************************************
```

Incorrect receiver timing

```
                    TABLE AND BAR GRAPH

MSL = 4079 ZUL = 1864 ULL = 2246 UUL = 6134

                                   Z    U                          M
0   4064  ************************************************************
1   3808  ************************************************************
2   0672  *********
3   0160  **
4   0000
5   4064  ************************************************************
6   0000
7   0576  ********
8   2808  ****************************************
9   4095  ************************************************************
```

Figure 4.21 Effect of receiver timing on channel outputs

Figure 4.22 Relative timing of code pulses for prototype labels

CHAPTER 5

MINIMISATION OF SECOND ORDER RESPONSES

5.0 INTRODUCTION

In this chapter the second order responses are examined in more detail and methods for minimising the effects of second order responses and intersymbol interference are discussed. Some practical aspects of SAW delay line design and construction and the consequent effects on system response are also discussed.

5.1 GENERATION OF THE SECOND ORDER RESPONSES

From Chapter 3, the primary response is generated from the sequence of b_k coefficients which are generated by the auto-correlation of the a_k coefficients. The primary response comprises two parts -

i) the code which arises from the correlation of a_o (the transmit IDT) with the a coefficients for the coding sets IDTs.

ii) the auto-correlation of the a_k coefficients for the coding set IDTs, resulting in a response occurring prior to the code. This response is called the pre-code response.

Both the code and pre-code responses are readily identified in the figures of Chapter 4 and Figure 5.1.

The secondary responses are due to the auto-convolution of the primary response. The secondary one response, governed by the b_k sequence, arises from the convolution of the primary response and the first pulse of the pre-code (due to input admittances of IDTs). The secondary two response, governed by the c_k sequence, is produced by the auto-convolution of the primary response (less the input admittance pulse). The secondary two response is considered to generate the most interference with the code in the primary response since it is caused by convolutions and auto-convolutions of the pre-code and code responses. Such convolutions have the potential to produce large amplitude responses.

5.1.1 Worst Case Codes

The a_k, b_k and c_k coefficients are related by equations 3.28 and 3.30 -

$$b_k = 2 \sum_{i=0}^{\infty} a_i a_{i+k} \qquad k \geqslant 0$$

$$c_k = \sum_{i=1}^{\infty} b_i b_{k-i} \qquad k \geqslant 1$$

If all IDTs have the nominal eleven fingers then the magnitude of all a_k coefficients is unity. For the maximum second order responses (i.e. the b_k, and hence c_k,

coefficients are at a maximum) all a_k coefficients must be a unity in magnitude and have the same phase. That is -

$$a_k = 1 \qquad \text{for all } k$$

(It can also be shown that $a_k = (-1)^k$ also produces the same magnitudes for the b_k and c_k coefficients but different phases.) This definition of the a_k coefficients (and the corresponding IDT layout) will produce the worst possible case (i.e. largest magnitudes) for second order responses. However for a practical coded delay line not all the a_k can be unity since there are no IDTs between the transmit IDT and the coding set, and hence the a_k corresponding to these positions must be zero.

From equations 3.28 and 3.30 it can be shown that the worst case for second order responses in a practical delay line will be produced when the a_k coefficients corresponding to the transmit IDT and coding set are all unity. This implies a delay line with a code of all 'ones'.

Figure 5.1 shows the primary response (in the form of the b_k coefficients) for the worst case of an M-bit coded delay line. From this figure (or from equation 3.30) the values of c_k may be deduced. Second order responses which interfere with the code arise from the auto-convolution of the pre-code and the convolution of the pre-code with the code. (Note that the auto-convolution of the code will always occur after the code and hence produces no interference.)

Figure 5.1 b_k coefficients for a worst case code in an M-bit system

A sequence of d_k coefficients may be used to describe the auto-convolution of the pre-code, and e_k coefficients used for the convolution of the pre-code with the code. These coefficients may be determined from Figure 5.1 and are defined over the code period ($k = M+1$ to $2M$ for an M-bit delay line) as -

$$d_k = \frac{1}{3}\, x(x+1)(x+2) \tag{5.1}$$

where $x = 2(M-1) - k + 1$

for $M + 1 \leqslant k \leqslant 2M - 1$ (otherwise $d_k = 0$)

$$e_k = y(2M - y - 1) \tag{5.2}$$

where $y = k - M - 1$

for $M + 2 \leqslant k \leqslant 2M$ (otherwise $e_k = 0$)

Furthermore, over the code period -

$$c_k = d_k + e_k \qquad \text{for } M+1 \leqslant k \leqslant 2M \tag{5.3}$$

The maximum value of d_k occurs at $k = M + 1$ (i.e. the first code bit) and is

$$(d_k)_{max} = \frac{1}{3}\, M(M-1)(M-2) \tag{5.4}$$

The maximum value of e_k occurs at $k = 2M$ (i.e. the last code bit) and is -

$$(e_k)_{max} = M(M-1) \tag{5.5}$$

It is noticed that the maximum value of d_k is larger than the maximum value of e_k (for $M \geqslant 5$). Thus for most practical codes, where the number of code bits is greater than five, the maximum value of c_k is dictated by the maximum value of d_k.

While the c_k coefficients govern the secondary two response, the secondary one response is controlled by the b_k coefficients. Over the code period, the sequence corresponding to the secondary one response is the sequence of b_k coefficients multiplied by $\frac{1}{2} b_o$. For a worst case code -

$$\frac{1}{2} b_o = M + 1$$

and $\quad b_k = 2 \qquad$ for $\quad M + 1 \leqslant k \leqslant 2M$

Thus the secondary one response is governed by a sequence of coefficients whose values are equal to 2(M+1) over the code period.

In general, the secondary two response is the major contributor to the second order response. In particular, the d_k coefficients are responsible for the maximum values of the c_k sequence (and hence the second order response). Thus the majority of the second order response can be associated with the auto-correlation of the pre-code.

Figures 5.2 and 5.3 show the channel outputs predicted by TIMDOM for worst case codes in 5, 10, 15 and 20-bit systems.

```
NO. OF CODE BITS =? 5

IDT WEIGHTINGS =? 11 11 11 11 11

CHANNEL OUTPUTS

      .77  ***************
     -.51  **********
      .50  **********
     -.50  **********
      .73  ***************
```

```
NO. OF CODE BITS =? 10

IDT WEIGHTINGS =? 11 11 11 11 11 11 11 11 11 11

CHANNEL OUTPUTS

      .81  ****************
     -.50  **********
      .46  *********
     -.44  *********
      .42  ********
     -.41  ********
      .41  ********
     -.41  ********
      .41  ********
     -.65  *************
```

Figure 5.2 Channel outputs predicted by TIMDOM for 5 and 10-bit worst-case codes.

```
NO. OF CODE BITS =? 15

IDT WEIGHTINGS =? 11 11 11 11 11 11 11 11 11 11
                  11 11 11 11 11

CHANNEL OUTPUTS

      .42  ********
     -.18  ****
      .22  ****
     -.26  *****
      .28  ******
     -.29  ******
      .30  ******
     -.30  ******
      .30  ******
     -.30  ******
      .29  ******
     -.28  ******
      .28  ******
     -.27  *****
      .50  **********

NO. OF CODE BITS =? 20

IDT WEIGHTINGS =? 11 11 11 11 11 11 11 11 11 11
                  11 11 11 11 11 11 11 11 11 11

CHANNEL OUTPUTS

     1.77  ************************************
    -1.24  ***************************
     1.06  ***********************
     -.89  *****************
      .74  ***************
     -.61  ************
      .50  **********
     -.40  ********
      .32  ******
     -.25  *****
      .19  ****
     -.15  ***
      .12  **
     -.09  **
      .08  **
     -.07  *
      .06  *
     -.06  *
      .06  *
     -.30  ******
```

Figure 5.3 Channel outputs predicted by TIMDOM for 15 and 20-bit worst-case codes.

The second order responses should be most noticeable at the beginning and end of the code corresponding to the maximum values of d_k and c_k coefficients respectively. The reduction in channel outputs due to second order effects for the 5-bit system is not significant and the smaller magnitudes in all channels (especially all except the first and last) is due almost entirely to intersymbol interference.

Channel outputs for the 10 and 15-bit systems show a further reduction due to larger second order responses although the large magnitudes predicted by equations 3.28 and 3.30 are not evident. This is because intersymbol interference effectively cancels the effect of the secondary response pulses in the channel outputs. However for the 20-bit system, the second order responses become so large that, even after partial cancellation by intersymbol interference, they dominate the channel outputs. Note that the response is largest for the beginning of the code due to the d_k coefficients.

It was stated previously that the magnitudes of the secondary response pulses are the same for $a_k = (-1)^k$. However in this case the intersymbol interference between channels is additive and secondary pulses are not cancelled as for $a_k = 1$. This results in very large channel outputs. Thus two codes which produce the same magnitude pulses (although phased differently) can result in completely different channel outputs through the effect of intersymbol

interference. As a consequence, intersymbol interference must be taken into account when examining channel outputs for evidence of second order responses.

For further examples in this chapter, worst-case codes will be defined as $a_k = 1$ although $a_k = (-1)^k$ will produce larger channel outputs.

5.1.2 Primary to Secondary Ratio

The relative magnitudes of primary and secondary responses not only depend on the b_k, d_k and e_k coefficients but also on the relative magnitudes of primary and secondary pulse waveforms.

In section 3.5 it was shown that, for the prototype system, the magnitudes (areas) of the secondary waveforms were a factor of 149 less than the primary waveforms. This factor will now be defined for a general system.

Since the area of the common system function, $\hat{g}_o(t)$, is common to all waveforms, it does not affect the relative magnitude of primary and secondary waveforms. Also the area of $\hat{y}_x(t)$ is the same as the area of $\hat{y}_i(t)$. Therefore the ratio of primary waveform area to secondary waveform area is defined by -

$$PSR = \frac{[\text{Area of } \hat{y}_x(t)]}{[\text{Area of } \hat{y}_x(t)]^2[\text{Area of } \hat{z}_t(t)]}$$

where PSR = primary to secondary ratio.

But since the area of a baseband time function is equal to the bandpass frequency function evaluated at ω_o, then

$$PSR = \frac{Y_x(\omega_o)}{Y_x^{\,2}(\omega_o)\,Z_t(\omega_o)}$$

$$= \frac{1}{G_o\ Z_t(\omega_o)} \qquad (5.6)$$

since $Y_x(\omega_o) = G_o$ as defined in section 3.1.6.
For the prototype system:

$G_o = 5.37 \times 10^{-6}$ and $Z_t(\omega_o) \approx 1250$

and PSR = 149

The PSR is an important ratio since, together with sequences b_k and c_k, it determines the magnitude of the second order response with respect to the primary response. For example, if one of the c_k coefficients has a value equal to twice the PSR, then the corresponding pulse in the secondary two response would have the same magnitude as a pulse representing a code bit in the primary response.

5.2 METALLISATION EFFECTS

Apart from electro-acoustic regeneration, surface acoustic waves may be reflected from an IDT purely due to the presence of the metal fingers on the substrate surface. The metallisation effectively shorts out the electric field (associated with a surface acoustic wave) at the

surface of the substrate and alters the acoustic impedance of the substrate. Associated with the change in impedance is a change in wave velocity which is the measure of electro-acoustic coupling for SAW substrates. The change in velocity divided by the nominal SAW velocity gives the coupling coefficient, Δ.

A change in acoustic impedance also results in wave reflections at the discontinuities. Reflections of this type were described by Emtage[10] as the "back piezo-electric effect". Emtage derived an "intensity reflection coefficient, R" to quantify reflections from a short-circuited IDT. For narrowband operation, Emtage's expression can be simplified to -

$$R = \frac{A^2}{1 + A^2} \tag{5.7}$$

where $A = 2\pi s \Delta N$

s = factor dependent on the metallisation ratio, a (refer equation 1.1) for a = 0.5, s = 0.23 (refer graph[10])

N = no. of <u>finger pairs</u> in the IDT

Δ = coupling coefficient

In actual fact, R is a power reflection coefficient and an intensity reflection coefficient, r, is given by

$$r = j\sqrt{R} \tag{5.8}$$

Considering Figure 5.4, a voltage impulse is applied to IDT 1. Some time later a current pulse will appear in the terminals of IDT 1 due to reflections from IDT 2. If IDT 2 is terminated in an impedance, Z, then the magnitude of the current pulse in IDT 1 due to electro-acoustic reflection is -

$$A_{ea} = Y_x[(\omega_o)]^2 Z(\omega_o) \tag{5.9}$$

and for delay lines in the prototype system Z is approximately equal to Z_t. Hence, equation 5.9 is rewritten as -

$$A_{ea} = Y_x[(\omega_o)]^2 Z_t(\omega_o) \tag{5.10}$$

The magnitude of a current pulse in IDT 1 due to reflections from a short-circuited IDT 2 (i.e. reflections due to the back piezo-electric effect) is -

$$A_{bp} = r\, Y_x(\omega_o) \tag{5.11}$$

and the ratio A_{bp}/A_{ea} is -

$$\frac{A_{bp}}{A_{ea}} = r\ \mathrm{PSR} \tag{5.12}$$

For the prototype system

$$r = j\ 7.39 \times 10^{-3}$$

$$\mathrm{PSR} = 149$$

and therefore $A_{pb}/A_{ea} = j\ 1.1$.

Figure 5.4 Back-piezoelectric effect reflections

Thus reflections arising from the back piezo-electric effect are approximately the same magnitude as those arising from electro-acoustic regeneration (except for a 90° carrier phase shift). Note also that the ratio, A_{bp}/A_{ea}, increases with PSR. This is due to the fact that back piezo-electric effects are constant (for an IDT with constant number of fingers and metallisation ratio) and by decreasing the relative magnitude of the secondary response (increasing PSR), reflections due to the back piezo-electric effect become larger than those for electro-acoustic regeneration.

For the prototype system, back piezo-electric effects (reflections due to metallisation) are considered to be responsible for the larger-than-expected secondary response shown in Figures 4.4, 4.8 and 4.12.

Generally speaking, these reflections may be minimised by special transducer designs. Since methods for controlling this type of second order response are well documented[11,20] metallisation effects in transducers are not considered further in this thesis.

5.3 TRANSDUCER PHASING POLICIES

In section 5.1 it was shown that most of the second order response was due to convolutions involving the pre-code and code responses, with a major contribution arising from the former. If the magnitude of the pre-code can be reduced in some way, then the secondary response will be reduced accordingly.

The pre-code response arises from the auto-correlation of the a_k coefficients from the coding set IDTs. The same coefficients also determine the code (via the b_k coefficients). However the receiver of the interrogation system only measures the amplitude of code pulses and not the associated phases. Thus the phase of the a_k coefficients may be modified without changing the code perceived by the receiver (assuming negligible inter-symbol interference for the moment). Since the pre-code is determined by the auto-correlation of the coding set a_k coefficients, the phases of these a_k may be modified to minimise the pre-code while still retaining the required code.

Although it is possible to assign an arbitrary phase to any IDT, the phasing policies which have been investigated use only two phases (i.e. $a_k = \pm 1$). This bi-phase policy is easily implemented in practice by assigning electrical polarities for the connection of IDTs.

The largest second order responses occur in codes which contain all 'ones' (see section 5.1.1). Phasing policies are sought to minimise the pre-code (and hence second order responses) in these codes, with a view to extending the policy to other codes. Thus the a_k coefficients are all unity in magnitude and must be assigned phases to produce the smallest pre-code.

The phases of the a_k are adjusted to give a sequence

which has a "minimum" resultant auto-correlation. Maximum correlation results when the sequence is not shifted with respect to itself. This correlation peak is unaffected by any phasing policy. However by judicial choice of phasing policies, shifted versions of the sequence will result in small correlation products. Such policies are based on sequences exhibiting the desired auto-correlation properties. Sequences of this type are used widely in digital communication and spread-spectrum communication systems and are examined in following sections.

5.3.1 Barker Codes

Barker codes[12] were designed for synchronising digital transmission systems and have the property that auto-correlation sequence terms are either '0' or '-1' except for the correlation peak. Only 4 such codes are known to exist (for 3, 7, 11 and 13 bit code lengths) and are:

3 bits : + + -

7 bits : + + + - - + -

11 bits : + + + - - - + - - + -

13 bits : + + + + + - - + + - + - +

where '+' denotes '+1'

and '-' denotes '-1'.

As an example of the Barker code phasing policy, Figure 5.5 shows the results predicted by TIMDOM for a

```
NO. OF CODE BITS =? 13

IDT WEIGHTINGS =? 11 11 11 11 11 11 11 11 11 11
                  11 11 11

CHANNEL OUTPUTS

      .57  ***********
     -.32  ******
      .34  *******
     -.36  *******
      .36  *******
     -.37  *******
      .37  *******
     -.36  *******
      .35  *******
     -.35  *******
      .34  *******
     -.34  *******
      .57  ***********
```

```
NO. OF CODE BITS =? 13

IDT WEIGHTINGS =? 11 11 11 11 11 -11 -11 11 11 -11
                  11 -11 11

CHANNEL OUTPUTS

      .70  **************
     -.55  ***********
      .49  **********
     -.55  ***********
      .96  *******************
      .93  *******************
    -1.03  **********************
     -.96  *******************
      .97  *******************
     1.40  ****************************
     1.38  ****************************
     1.39  ****************************
     1.14  **********************
```

Figure 5.5 Channel outputs predicted by TIMDOM for a 13-bit worst case code (top) and using a 13-bit Barker code phasing policy (bottom).

worst-case 13-bit code and the same code phased using a 13-bit Barker code. Ideally all channel outputs should be the same, but significant improvements over the worst-case code have been achieved due to the reduction of the pre-code by the phasing policy. The residual variation in channel outputs can be shown to be mainly the result of intersymbol interference as will be examined in section 5.4.

Problems arise when the number of bits in a code is not equal to the length of a Barker code or if the code is not composed of all 'ones'. In these cases it is not clear which phasing policy is optimum but it is the author's opinion that good results can be obtained by multiplying a given code by a Barker code of the same or greater length. This has the effect of spreading the spectrum of the a_k sequence and thereby imparting the required auto-correlation properties. (The next section examines maximal sequences which are used extensively in spread spectrum system for the above purpose).

Thus a phasing policy may be used for codes which do not comprise all 'ones' by applying a phasing policy to a complete code or individually to sections of a code comprising many 'ones'. Figure 5.6 demonstrates this principle by phasing a 12-bit code in a 20-bit system. The Barker 13-bit code is used to phase the 12-bit code "locally" within the 20-bit structure. Note that a Barker code could not be used for a complete 20-bit code since the longest Barker code is 13 bits.

```
IDT WEIGHTINGS =? 11 0 11 11 11 0 11 11 0 11 0 11
                  0 0 0 0 0 0 0 0

CHANNEL OUTPUTS

      .97  ******************
      .47  *********
      .72  **************
     -.51  **********
      .71  **************
      .51  **********
      .71  **************
     -.71  **************
     -.50  **********
     -.92  ******************
     -.48  **********
     -.91  ******************
     -.22  ****
      .06  *
     -.00
      .02
     -.02
     -.02
     -.04  *
     -.03  *

IDT WEIGHTINGS =? 11 0 11 11 11 0 -11 11 0 -11 0 -11
                  0 0 0 0 0 0 0 0

CHANNEL OUTPUTS

      .97  ******************
      .46  *********
      .74  ***************
     -.50  **********
      .77  ***************
      .02
    -1.19  ************************
    -1.19  ************************
      .03  *
      .94  ******************
      .44  *********
      .93  ******************
      .20  ****
     -.03  *
      .02
      .05  *
      .04  *
      .05  *
      .00
     -.01
```

Figure 5.6 Channel outputs predicted by TIMDOM for a 12-bit code in a 20-bit system (top) and using a 13-bit Barker code 'local' phasing policy (bottom).

Unfortunately it is difficult to identify improvements due to the reduction of secondary responses in Figure 5.6 since the phasing policy alters channel outputs considerably due to intersymbol interference.

This problem of ISI was seen in Section 5.1.1 and will also be encountered in following sections. Phasing policies may reduce second order effects but will also significantly change the system response through the mechanism of ISI.

The changes in second order responses achieved by phasing policies may be partially or completely obscured by changes in the system response due to ISI. Therefore any investigation into the reduction of second order responses by the use of phasing policies must also take ISI into account when examining the system response.

Section 5.4 will propose methods for reducing the effect of ISI thereby allowing a better appreciation of the effects of the second order response on the system response.

5.3.2 Maximal Sequences

Maximal sequences[13] (or m-sequences) are defined to be the longest binary sequence that may be generated by a shift register of a given length (using feedback from various stages). M-sequences have the property that the correlation of a sequence with a repeated version of itself produces terms which are always '-1' except for the

correlation peaks. M-sequences only occur in lengths of $2^n - 1$ bits where n is the number of stages in the shift register required to produce the sequence.

Using an m-sequence as a phasing policy means that auto-correlation will occur for an m-sequence in isolation (i.e. not repeated). Auto-correlation products will not, in general, be '-1' as before but appear to be relatively small compared with the correlation peak. Examples of m-sequence are:

7-bit : + + + - - + -

15-bit : + + + + - - - + - - + + - + -

(Note that the 7-bit m-sequence is also a Barker code.)

There is an unlimited number of m-sequences available in varying lengths. M-sequences are used in the same manner as Barker codes. Figure 5.7 shows results predicted by TIMDOM for a 15-bit worst-case code and a 15-bit code modified using a 15-bit m-sequence as a phasing policy.

For codes with bit lengths which do not correspond to an m-sequence or which do not comprise all 'ones', procedures suggested for use with Barker codes should apply. Since m-sequences of a given length have low cross-correlation products with repeated m-sequences of the same length, sections of large sparse code may be phased "locally" with different m-sequences of the same length instead of phasing the complete code with a large length m-sequence.

```
NO. OF CODE BITS =? 15

IDT WEIGHTINGS =? 11 11 11 11 11 11 11 11 11 11
                  11 11 11 11 11

CHANNEL OUTPUTS

      .42  ********
     -.18  ****
      .22  ****
     -.26  *****
      .28  ******
     -.29  ******
      .30  ******
     -.30  ******
      .30  ******
     -.30  ******
      .29  ******
     -.28  ******
      .28  ******
     -.27  *****
      .50  **********


NO. OF CODE BITS =? 15

IDT WEIGHTINGS =? 11 11 11 11 -11 -11 -11 11 -11 -11
                  11 11 -11 11 -11

CHANNEL OUTPUTS

      .65  *************
     -.55  ***********
      .48  **********
     -.97  ******************
     -.94  ******************
      .49  **********
     -.94  ******************
    -1.40  ****************************
     -.90  ******************
     1.03  ********************
      .86  *****************
     -.93  ******************
    -1.32  **************************
    -1.33  **************************
    -1.11  *********************
```

Figure 5.7 Channel outputs predicted by TIMDOM for a 15-bit worst-case code (top) and using a 15-bit m-sequence phasing policy (bottom).

Although this proposal has not been pursued, it is considered that this approach should give acceptable results with sparse codes.

5.3.3 Phase Chirp

The phase chirp is a sequence in which the phase of the coefficients reverses with increasing periodicity. The sequence considered for use as a phasing policy is:

+ - - + + + - - - - + + + + + - . . . . etc.

The length of the sequence is not fixed and may be truncated as required. The reason for using the phase chirp as a phasing policy is the author's belief that such a sequence may display the required spectrum-spreading and hence auto-correlation properties.

Figure 5.8 shows an example of the phase chirp used for a 15-bit worst-case code.

Since the phase chirp is variable in length, it is easily applied to codes of any length. Although its auto-correlation properties may not be as good as Barker codes or m-sequences, the phase chirp is easy to apply in practice.

```
NO. OF CODE BITS =? 15

IDT WEIGHTINGS =? 11 11 11 11 11 11 11 11 11 11
                  11 11 11 11 11

CHANNEL OUTPUTS

      .42  ********
     -.18  ****
      .22  ****
     -.26  *****
      .28  ******
     -.29  ******
      .30  ******
     -.30  ******
      .30  ******
     -.30  ******
      .29  ******
     -.28  ******
      .28  ******
     -.27  *****
      .50  **********


NO. OF CODE BITS =? 15

IDT WEIGHTINGS =? 11 -11 -11 11 11 11 -11 -11 -11 -11
                  11 11 11 11 11

CHANNEL OUTPUTS

     1.23  *************************
     1.00  ********************
     -.88  ******************
    -1.02  ********************
      .43  *********
     -.88  ******************
     -.95  *******************
      .40  ********
     -.46  *********
      .94  *******************
      .88  ******************
     -.38  ********
      .46  *********
     -.53  ***********
      .81  ****************
```

Figure 5.8 Channel outputs predicted by TIMDOM for a 15-bit worst-case code (top) and using a phase chirp phasing policy (bottom).

5.4 INTERSYMBOL INTERFERENCE

In section 3.6 it was stated that the relationship between weighted sequences corresponding to pulses in the response and the channel outputs could be represented by transversal digital filters. It follows that if the channel outputs are applied to another digital filter (which is the inverse of the transversal filter) then the weighted sequence (corresponding to the primary response, say) could be recovered from which the code would be easily derived. Unfortunately this type of post-processing is not possible on the prototype system since the channel outputs contain no phase information (i.e. the outputs are unsigned).

Another approach is a pre-processing policy where the weightings of the IDTs on the delay line are modified to compensate for intersymbol interference (ISI). If the nominal coding set weightings are processed by an inverse ISI filter and the resultant transducer weightings used for delay line fabrication, then the channel outputs should be a direct measure of code bits and free from ISI. Modification of IDT weightings is achieved by varying the number of fingers in the IDTs as shown in section 3.5.7.

Assuming that secondary responses are not significant, the filter representing ISI may be described by its z-transform, F(z), defined as -

$$F(z) = az^{-2} + bz^{-1} + a \qquad (5.13)$$

where a and b are filter coefficients as per section 3.6.

The zeros of this filter occur at -

$$z = \frac{1}{-x \pm \sqrt{x^2 - 1}}$$

where $x = \frac{b}{2a}$

The inverse of this filter must have poles at these locations in the z-plane. Figure 5.9 shows the root locus of these pole locations versus the variable x. Since poles outside the unit circle in the z-plane indicate instability, the root locus shows that the inverse filter is marginally stable for $-1 \leqslant x \leqslant +1$ and unstable for other values of x. Since the value of x for the prototype system is approximately 2, the inverse ISI filter must be unstable. This instability was evident in initial attempts to pre-process the IDT weightings using a recursive form of the inverse filter.

However another filter configuration was found to approximate the inverse filter for values of $|x|$ in excess of unity. The approximation is defined by -

$$G(z) = -a^{-2} + bz^{-1} - a \qquad (5.14)$$

where a and b are the same coefficients as for F(z) in equation 5.13. This filter is inherently stable. Using this filter realisation it is possible to pre-process the code to provide a set of IDT weighting to minimise ISI. Pre-processing is implemented in FORTRAN program WEIGHT (Appendix D) which accepts the required binary code (including phasings) as input and produces a sequence of

Figure 5.9 Root locus of poles for an ideal inverse ISI filter

IDT weightings (number of fingers and electrical phase). WEIGHT allows for the alternate anti-phasing of code bits inherent in the delay line.

Figures 5.10 and 5.11 show the responses predicted by MINISEC and TIMDOM for the code of Figure 4.16 after modifying IDT weightings by WEIGHT. In this example, a noticeable reduction in ISI is evident.

The weighting policy is restricted to IDTs within the coding set. This does not allow for extra transducers at either end of the coding set which would be necessary for a strict implementation of weightings based on the inverse ISI filter. The absence of these two extra IDTs has the effect of increasing the magnitude of the first and last channels. This is demonstrated in Figure 5.11.

However for many codes, second order responses and ISI are both significant. A new weighting and phasing policy is required to reduce ISI and secondary responses together. A good compromise has been found by choosing the code bit phasings as determined by a suitable phasing policy and then processing this code using WEIGHT to give appropriate transducer weightings and phasings. This combined policy is demonstrated in Figures 5.12, 5.13 and 5.14 for a 20-bit code comprising all 'ones'. As a further test of this policy the same code is used with a 'zero' substituted for a 'one' in

in the 15th bit position (i.e. channel 14) and results are shown in Figures 5.15, 5.16 and 5.17. In both cases a phase chirp has been used as the phasing policy, followed by IDT weightings determined by WEIGHT. Figures 5.18, 5.19 and 5.20 show the results for the same code as Figure 5.15 but without phasing and weighting policies applied (i.e. the 20-bit worst-case code).

It should be noted that the magnitude of the response in Figures 5.18, 5.19 and 5.20 seems much smaller than would be expected from calculations in Section 5.1.1 for worst-case codes. This apparent discrepancy is due totally to the effect of intersymbol interference. The very large pulses predicted from Section 5.1.1 do exist but, due to ISI, the resultant magnitudes in the system response are reduced.

In this case adjacent pulses are anti-phase and ISI results in cancellation and hence a reduction in pulse magnitudes. If the IDTs were arranged such that adjacent pulses were in-phase, ISI would result in additive interference and a very large system response. Once again it is seen that the phasing of IDTs, combined with the mechanism of ISI, can greatly affect the system response.

It can be seen that this type of combined policy has removed the majority of secondary responses and

Figure 5.10 System time response (MINISEC) for label code 1001000101 compensated for ISI

Figure 5.11 Simulated channel outputs (MINISEC and TIMDOM) for label code 1001000101 compensated for ISI

Figure 5.12 Simulated channel outputs (MINISEC and TIMDOM) for a 20-bit code comprising all 'ones'

Figure 5.13 System time response (MINISEC) for the code of Figure 5.12

Figure 5.14 Secondary response (MINISEC) for the code of Figure 5.12

Figure 5.15 Simulated channel outputs (MINISEC and TIMDOM) for a 20-bit code of Figure 5.12 but with bit 14 = 0

Figure 5.16 System time response (MINISEC) for the code of Figure 5.15

Figure 5.17 Secondary response (MINISEC) for the code of Figure 5.15

Figure 5.18 Simulated channel outputs (MINISEC) for the 20-bit worst-case code

Figure 5.19 System time response (MINISEC) for the 20-bit worst-case code

Figure 5.20 Secondary response (MINISEC) for the 20-bit worst-case code

intersymbol interference. The secondary one response can never be removed using this policy as explained in section 5.6. Although intersymbol interference has been suppressed by weighting the transducers, a similar result could have been achieved by post-processing the receiver channel outputs (providing that phase information is available) or by pre-processing the transmitter pulse.

5.5 SHIFTING THE CODING SET

A large contribution to the secondary response comes from the auto-convolution of the pre-code (the d_k coefficients). This auto-convolution has a duration of twice the code duration and begins at the time origin. At present the later half of the pre-code auto-convolution overlaps into the code period giving rise to interference. If the code could be delayed in time, this overlap could be avoided and no interference would result from the auto-convolution of the pre-code. In practice this is achieved by increasing the separation between the transmit IDT and the coding set on the delay line. If the separation is increased to greater than twice the length of the coding set, interference from pre-code auto-convolution is avoided.

However secondary responses due to the convolution of the code with the pre-code (the e_k coefficients) will still remain and cannot be avoided since the position of this convolution always overlaps the code.

```
NO. OF CODE BITS =? 10

IDT WEIGHTINGS =? 11 11 11 11 11 11 11 11 11 11

CHANNEL OUTPUTS

      .81  ****************
     -.50  **********
      .46  *********
     -.44  *********
      .42  ********
     -.41  ********
      .41  ********
     -.41  ********
      .41  ********
     -.65  *************
```

```
NO. OF CODE BITS =? 20

IDT WEIGHTINGS =? 11 11 11 11 11 11 11 11 11 11
                   0  0  0  0  0  0  0  0  0  0

CHANNEL OUTPUTS

      .76  ****************
     -.47  *********
      .45  *********
     -.44  *********
      .43  *********
     -.42  ********
      .42  ********
     -.41  ********
      .41  ********
     -.65  *************
     -.34  *******
      .05  *
     -.04  *
      .02
     -.01
      .00
      .00
     -.01
      .01
     -.01
```

Figure 5.21 Channel outputs predicted by TIMDOM for a 10-bit worst-case code (top) and the same code sequence in a 20-bit system (bottom) to demonstrate the effect of shifted codes.

```
NO. OF CODE BITS =? 10

IDT WEIGHTINGS =? 11 -8 8 -8 8 -8 8 -8 8 -11

CHANNEL OUTPUTS

       .62  ************
       .65  *************
       .62  ************
       .66  *************
       .67  *************
       .68  **************
       .68  **************
       .68  **************
       .73  ***************
       .75  ***************


NO. OF CODE BITS =? 20

IDT WEIGHTINGS =? 11 -8 8 -8 8 -8 8 -8 8 -11
                   0 0 0 0 0 0 0 0 0 0

CHANNEL OUTPUTS

      1.14  ************************
      1.03  **********************
       .90  ******************
       .84  *****************
       .79  ****************
       .75  **************
       .72  **************
       .69  **************
       .74  ***************
       .75  ***************
     -.18  ****
     -.35  *******
     -.28  ******
     -.22  ****
     -.17  ***
     -.12  **
     -.08  **
     -.05  *
     -.02
     -.01
```

Figure 5.22 Channel outputs predicted by TIMDOM for codes of Figure 5.21 modified by WEIGHT to minimise intersymbol interference.

Figure 5.21 shows the TIMDOM predictions for a 10-bit worst-case code and the same code in a 20-bit system. (This is equivalent to a 10-bit system with the code delayed by an additional 10-bit periods.)

By shifting the coding set as suggested the maximum second order response will occur at the last bit in the code due to the maximum value of the e_k coefficients. If the coding set is not shifted the maximum second order response will occur at the first bit in the code due to the maximum value of the d_k' coefficients. Thus the suggested shift reduces the maximum possible second order response (i.e. for a worst-case code) by a factor of

$$\frac{(d_k)_{max}}{(e_k)_{max}} = \frac{1}{3}(M-2) \qquad (5.15)$$

This results in a six-fold reduction in the secondary response for a 20-bit worst-case code. In general, useful reductions in secondary responses can be achieved by shifting the coding set at the expense of a physically longer delay line.

Although Figure 5.21 does not appear to show any significant improvements associated with shifting the coding set, the actual reduction in secondary responses has been concealed by intersymbol interference. Figure 5.22 shows the same situation as Figure 5.21 except that transducer weightings have been modified (by WEIGHT) to compensate for the effects of intersymbol interference.

Figure 5.22 now clearly shows the reduction in secondary responses due to shifting the coding set. The shifted code displays increased magnitudes at the beginning of the code due to the removal of the d_k coefficients. However channel outputs are still reduced towards the end of the code since the effect of the e_k coefficients remains.

5.6 IDEAL PHASING

Various phasing policies have been proposed to minimise the pre-code and hence secondary responses. However there is no phasing policy which will reduce the first pulse of the primary response caused by input admittances. The convolution of this pulse with the primary response produces the secondary one response. Therefore phasing policies and shifting the coding set may reduce the effects of the secondary two response but will have no influence on the secondary one response.

Consider an "ideal" phasing policy (although its existence is doubted) in which the pre-code is made zero everywhere except for the input admittance pulse. In this case, the only secondary response is due to the secondary one response which increases linearly with the number of transducers on the delay line. From section 5.1.1 the impulses controlling the secondary one response will have a maximum value of 2(M+1) for an M-bit system. Thus, if M is large enough, the secondary one response will become significant. The magnitude of an impulse

corresponding to a code bit is two (i.e. $b_k = 2$ for code bits) and allowing for pulse spreading, the relative contribution to the corresponding channel output is $2 \times 0.45 = 0.9$. The magnitude of an impulse governing the secondary response at that code bit is $2(m+1)$ for m transducers on the delay line. The contribution of the secondary response pulse, allowing for pulse spreading, is $2(m+1) \times 0.32 \div PSR$. The secondary pulse will reduce the magnitude of the primary pulse and if an arbitrary level of 3dB is set for the maximum allowed reduction in the corresponding channel output then:

$$(m+1) \leqslant 0.42\ PSR$$

The worst case occurs when all transducers are present for an M-bit system in which case $m = M$ and

$$(M+1) \leqslant 0.42\ PSR \tag{5.16}$$

This expression imposes limits on the maximum number of transducers on a delay for a given primary-to-secondary ratio (PSR), and the minimum PSR for a given number of transducers (system code size, M). In the prototype system, PSR = 149 and hence the theoretical maximum number of code bits that could be satisfactorily used is 62.

Thus, even if an "ideal" phasing policy could be found, the secondary one response would not be affected and would impose constraints on the system.

5.7 LABEL CHARACTERISTICS

The primary-to-secondary ratio, PSR, of section 5.1.2 has a direct effect on the relative magnitude of the secondary responses. For a delay line with identical transducers:

$$PSR = \frac{1}{G_o Z_t(\omega_o)} \qquad \text{(from equation 5.6)}$$

where G_o = input conductance of a transducer at resonant frequency

and $Z_t(\omega_o)$ = effective source impedance of the local antenna seen by the delay line at resonant frequency.

To minimise the secondary response, the PSR must be as large as possible. This is achieved by decreasing G_o and/or $Z_t(\omega_o)$. From equation 3.17:

$$G_o = \frac{-\pi^2 N^2 (\varepsilon_p + \varepsilon_o) |\omega_o| \Delta W}{13.75}$$

Thus G_o may be decreased by decreasing -

i) ε_p and/or Δ. Values depend on type of SAW substrate.

ii) ω_o - system operating frequency.

iii) N and/or W - transducer geometry.

Unfortunately, G_o governs the absolute magnitude of the primary response and any reduction in G_o will also increase the effective insertion loss of the labels. Thus

G_o may only be decreased to a level consistent with satisfactory signal-to-noise performance of the system.

$Z_t(\omega_o)$ is the effective parallel damping resistance of the label antenna at resonance as seen by the delay line. Referring to Figure 5.23 some new quantities are defined:

i) G_r - parallel conductance of the antenna tuned circuit at resonance due to antenna radiation and losses.

ii) G_d - additional damping conductance required to adjust the total label quality factor, Q, to a given value.

iii) G_i - the total input conductance of the delay line at resonance.

iv) Q' - quality factor of antenna tuned circuit with damping from G_r only.

v) Q" - quality factor of antenna tuned circuit with damping from G_r and G_d.

vi) Q - quality factor of antenna tuned circuit with damping from G_r, G_d and G_i. Thus Q is the quality factor of the complete label.

vii) C - total parallel capacitance of the label comprising capacitances of the delay line and antenna.

Figure 5.23 Label quantities

Furthermore:

$$Q' = \frac{\omega_o C}{G_r} \tag{5.17}$$

$$Q'' = \frac{\omega_o C}{G_r + G_d} \tag{5.18}$$

$$Q = \frac{\omega_o C}{G_r + G_d + G_i} \tag{5.19}$$

$$Z_t(\omega_o) = \frac{1}{G_r + G_d} \tag{5.20}$$

From equation 5.20, the primary-to-secondary ratio can be written as:

$$PSR = \frac{G_r + G_d}{G_o} \tag{5.21}$$

Thus PSR may be increased by increasing $(G_r + G_d)$, but equation 5.19 predicts a corresponding decrease in label Q and hence increased label insertion loss. In general it is seen that any attempt to increase the PSR (i.e. decrease relative secondary responses) will also increase the label insertion loss (i.e. decrease the absolute primary response).

If it is desired to maintain a value of Q then equation 5.19 states that $(G_r + G_d + G_i)$ should be constant. If G_i is increased, then $(G_r + G_d)$ must decrease with a corresponding increase in PSR. Thus for a label of fixed Q, damping from the delay line must be kept to a minimum for low relative secondary responses.

Note that the above analysis holds for antennae that may be modelled by a single inductor (as for the loop antennae of the 200 MHz prototype labels). However other equivalent circuits may have to be used for other types of antennae (e.g. a dipole) and in these cases, the antennae impedance at resonance, $Z_t(\omega_o)$, as seen by the delay line must be kept as low as possible for satisfactory values of PSR. (Delay line capacitance is included in $Z_t(\omega_o)$). In conflict with this requirement is the need for the label Q to be maintained as high as possible (subject to bandwidth limitations) for acceptable label insertion loss.

In the previous analysis used in this thesis it has been assumed that the product Z_tY is much less than unity. Under these circumstances the binomial expansion used in Section 1.5 is appropriate and only second order responses are significant. This is the case in the 200 megahertz prototype system where the use of quartz substrates results in low values for the delay line admittance, Y.

However it is possible that Y could be considerably larger in label systems which employ stronger coupling substrates and/or larger transducers. For these cases the product Z_tY may not be significantly less than unity and third and higher order responses increase in magnitude. If the product is greater than unity the binomial expansion ceases to be valid and another method of mathematical analysis is required.

The binomial expansion nevertheless is a useful method of calculating second (and higher) order responses and some manipulation may be required to extend its validity to cases involving large delay line admittances. Consider an analysis where the total input admittances of transducers is not included in the delay line admittance but incorporated into the delay line load impedance, Z_t, to give a modified impedance, Z'_t. The resulting modified delay line admittance, Y', now comprises only the total of transadmittances.

i.e. $Z'_t = Z_t$ combined with transducer input admittances

$Y' =$ total of transadmittances.

Note that this manipulation has not changed the physical system being analysed but only re-arranged the mathematics. The modified expressions for Z_t and Y (i.e. Z'_t and Y') can be substituted into equation 1.11 giving rise to the product Z'_tY' (in place of the previous Z_tY). Once again, for the binomial expansion to be useful, the product Z'_tY' should be less than unity in magnitude. It is possible for Y' to be made small, especially in instances where its spectrum is spread by the use of phasing policies. Furthermore, Z'_t is less than Z_t due to the influence of the transducer input admittances. Thus the binomial expansion may still be valid by using Z'_t and Y' whereas Z_t and Y would have rendered it unusable.

Implications of this manipulation are:

- the primary zero, secondary zero and secondary one responses would no longer exist since the transducer input admittances are combined into the delay line load impedance, Z'_t. Only a primary one and a secondary two response would remain (suitably modified through the influence of Z'_t).

- the primary-to-secondary ratio is now defined as -

$$PSR' = \frac{1}{Z'_t(\omega_o)G_o} \tag{5.22}$$

Note that PSR' > PSR (equation 5.6) since

$$Z'_t(\omega_o) < Z_t(\omega_o).$$

Equation 5.22 can be thought of as an amended form of equation 5.6 for cases where the delay line admittance is not negligible compared with Z_t. Equation 5.22 can also be used to make some interesting predictions. It has been suggested that the delay line load impedance (i.e. the label antenna impedance, Z_t) should be matched to the total transducer input admittance of the delay line in the interests of better power transfer. In this case -

$$Z'_t(\omega_o) = \frac{1}{mG_o} \tag{5.23}$$

where m = number of transducers on the delay line. From equation 5.22 the primary-to-secondary ratio is -

$$PSR' = m$$

Thus the matching policy will result in poor values of PSR' for delay lines with small numbers of transducers. Although an increase in the number of transducers would improve PSR', the corresponding increase in the C_k coefficients would result in an overall increase in the relative magnitude of the second order response.

The previous few paragraphs are not proposed as a strict alternate method of calculation for cases where the delay line admittance is not negligible. For example, there is no guarantee that Y' will always be small as assumed (although phasing policies help to achieve this). However the above discussion is intended to show that methods which have been developed for the analysis of the 200 MHz prototype system (on which this thesis is based) may need to be modified for the analysis of other systems. In particular, the use of the binomial expansion is a convenient method, but is subject to constraints imposed by magnitudes of impedance and admittances.

For any system it can be stated that any reduction in the secondary response by varying admittances and impedances of the delay line and/or antenna can only be achieved by a corresponding reduction of the primary response (and hence a smaller retransmission of energy on the delay line) and a consequent increase in label insertion loss.

5.8 A 915 MEGAHERTZ SYSTEM

A coded SAW label system has been proposed for operation at 915 megahertz. The delay lines could be constructed on various substrates but quartz and lithium niobate ($LiNbO_3$) are particularly considered. The stronger piezo-electric coupling of $LiNbO_3$ has some advantages as will be discussed later. Transducers should comprise double-electrode fingers to suppress metallisation reflections and bulk waves. This transducer configuration will also allow third harmonic operation thereby permitting operation at high frequencies without the necessity of very small transducer geometries.

Both amplitude and phase encoding of the delay line are possible. If a 30-bit system is required, amplitude encoding would require 30 coding set transducers. Phase encoding can reduce the number of transducers by permitting more than one code bit per transducer. For example, if a transducer is assigned one of four phases (i.e. 90 degrees apart), two bits can be encoded into one transducer. Thus, for a 30-bit code, only 15 coding set transducers are required. Further advantages arise if the phase of each transducer is measured relative to the previous transducer in the coding set rather than relative to the transmit IDT. This means that positional accuracy must only be maintained over a distance equal to the spatial separation between contiguous coding set IDTs rather than the larger distances between the transmit IDT and coding set IDTs. This phasing

arrangement requires an extra transducer which is added to the beginning of the coding set as the initial phase reference, making a total of 16 transducers in the coding set for a 30-bit system.

This method of differential phase encoding may be achieved by shifting the position of a transducer relative to the previous coding set transducer. Another method is to place lengths of metallisation (phase encoding pads) between pairs of transducers in a set of equidistant coding set transducers. Since the metallisation alters the SAW propagation velocity, phase shifts between two transducers may be achieved by an appropriate length of the phase encoding pad. For example, at 915 megahertz phase encoding pad lengths are 930 and 42 μmetres for a 90 degree shift on quartz and $LiNbO_3$ respectively. For 270 degree phase shifts these lengths would triple. The spacing between transducers required to accommodate phase encoding pads may result in unacceptably long delay lines and reply pulse spacing. This problem is most noticeable in quartz due to its low piezo-electric coupling. For this reason stronger coupling substrates, such as $LiNbO_3$, are preferred when using phase encoding pads.

If a phase variation of ±20 degrees (say) in transducer phase is permitted for proper decoding, then the magnitude of a secondary pulse must not exceed 36% of the primary pulse magnitude (assuming the worst case of 90 degrees phase difference between the primary and secondary

pulses). This also assumes exact placement of transducers on the substrate since at 915 megahertz a phase shift of 20 degrees is equivalent to a displacement of 0.2 micrometres on quartz (and $LiNbO_3$). However all phase distortion is assumed to come from the presence of the secondary response and not inaccuracies in transducer placement.

The worst case for secondary responses occurs with all transducers having the same phase. This is equivalent to a worst-case code of Section 5.1.1. (In this case the secondary pulses are not 90 degrees apart from the primary pulses but the case in which this does occur is very similar to a worst-case code). For a worst-case code, the maximum secondary response pulse has a magnitude governed by -

$$(d_k)_{max} = \frac{1}{3} M(M-1)(M-2)$$

where M is the number of coding set transducers.

The relative magnitude of the corresponding primary pulse is given by -

$$b_k = 2$$

Therefore for no corruption of the code to occur -

$$(d_k)_{max} \div PSR \leqslant 0.36 \times 2$$

or

$$PSR \geqslant 0.46M(M-1)(M-2) \qquad (5.24)$$

and thus for a 915 MHz 30-bit phase encoded system (M = 16) -

$$PSR \geqslant 1545$$

Similarly, for a shifted coding set (as per Section 5.5) where the effect of the secondary response has been reduced, the magnitude of the maximum secondary pulse is governed by -

$$(e_k)_{max} = M(M-1)$$

and it may be shown that -

$$PSR \geqslant 1.37M(M-1) \qquad (5.25)$$

and also $\qquad PSR \geqslant 330 \qquad$ for $M = 16$

The expressions 5.24 and 5.25 set limits for the satisfactory operation of the system. For a given system code length the PSR must be above a minimum level for correct system operation. Conversely if the PSR is a given value there is a maximum limit on the number of code bits that can be used satisfactorily in the system.

For example, typical values of input conductance, G_o, for transducers operating at 915 MHz have been calculated (see Appendix E) for both quartz and lithium niobate substrates and are -

$$G_o = 5.47 \times 10^{-5} \quad \text{(quartz)}$$

and $\qquad G_o = 1.18 \times 10^{-2} \quad (LiNbO_3)$.

From the values of G_o, restrictions on the effective label antenna impedance, $Z_t(\omega_o)$, can be deduced from the

limitations imposed on PSR (using equations 5.24, 5.25 and 5.6) for a 30-bit 915 MHz phase encoded system.

For an unshifted coding set for which PSR $\geqslant$ 1545, equation 5.6 implies:

$$Z_t(\omega_o) \leqslant 11.8 \text{ ohms (quartz)}$$

$$\leqslant 0.055 \text{ ohms } (LiNbO_3)$$

For a shifted coding set (PSR $\geqslant$ 330):

$$Z_t(\omega_o) \leqslant 55.4 \text{ ohms (quartz)}$$

$$\leqslant 0.26 \text{ ohms } (LiNbO_3)$$

The label antenna must meet these impedance requirements for satisfactory system performance. Note that the impedance levels must be much lower for lithium niobate. This is necessary to reduce the larger primary response (and hence larger relative secondary response) which would occur from using a stronger coupling substrate. Section 5.7 has already stated that the relative magnitude of the secondary response can only be reduced with a corresponding decrease in primary response. A further implication is that label insertion loss and the relative magnitudes of primary and secondary responses are not independent. For this reason there seems to be no real advantage in using lithium niobate as a substrate in order to simply decrease label insertion loss since the relative level of the secondary response would increase. Stronger

coupling substrates are, however, useful for attaining given admittance levels with smaller transducers and for implementing phase encoding pads.

Note that the high values of PSR required for satisfactory operation of the 30-bit phase encoded system arise from the large secondary responses which would be encountered (especially in the worst-case and similar codes). In an amplitude encoded system these secondary responses could be greatly reduced by phasing policies; an option not available for phase encoded systems. Although an amplitude encoded system would require twice as many transducers as a phase encoded system (i.e. one transducer per bit versus one transducer per two bits respectively), the advantages inherent in the ability to use phasing policies for secondary response reduction make the amplitude encoded approach more attractive. Furthermore it would relax the limitations on antenna impedance, $Z_t(\omega_o)$.

5.9 DELAY LINE PRODUCTION CONSIDERATIONS

Previous sections have proposed transducer phasing and weighting policies to minimise second order responses and intersymbol interference. However some difficulties may be encountered in applying these policies to the

production of coded delay lines in volume quantities.

Methods for producing the desired metallisation patterns on substrates employ techniques common to semi-conductor microcircuit production. A major difference in requirements arises from the fact that microcircuit techniques are aimed at producing large numbers of identical devices whereas coded delay lines must, of necessity, differ from one another due to their coding.

Current methods used in the photofabrication stage of microcircuit production include -

- "step-and-repeat" camera exposure systems (e.g. wafer steppers) in which each device on a large substrate (wafer) is exposed individually.
- contact exposure of a complete wafer using a mask containing the patterns of many devices. This exposes all devices on the wafer in a single exposure. The masks may be produced either photographically or by electron-beam lithography.

Considering firstly the production of delay lines for amplitude encoded systems (e.g. the 200 MHz prototype system), exposure systems should allow for transducers with varying numbers of fingers (weighting) and electrical phase (polarity of connection to busbars). Each transducer in a delay line may be exposed individually using a set of reticles (each containing a different mask

of a transducer) covering all transducer weightings and phasings. The number of reticles may be reduced if phasing is implemented by accurately positioning transducers on the substrate. Since the accuracy of current commercial wafer steppers is approximately 0.5 μmetre, the phase of a transducer may vary by up to 10 degrees (at 200 MHz on quartz).

However the largest problem associated with the production of coded delay lines with wafer steppers is associated with throughput (the number of devices produced per unit time). For a 20-bit system there are 1048576 possible codes (and hence the same number of different delay lines) and a total of 10485760 coding set transducers (not including those introduced to reduce intersymbol interference through weighting policies). If each "move-and-expose" action takes one second (typical), production time for a complete set of delay lines would be 121 days using one wafer stepper. This figure does not include overheads such as exposing the busbars and transmit IDT which may be achieved with one extra exposure per device using one additional reticle. Neither has the time required for changing and aligning reticles been included. Since this time is much greater (a minute or more) than for a "move-and-expose" action, the throughput will almost totally depend on reticle changing and alignment times. For example, assuming a one-minute reticle change and alignment time and a reticle change for each transducer, production time

would increase by a factor of sixty (20 years for a full set of 20-bit coded delay lines or 1 delay line every 10 minutes). However by arranging for all transducers with the same weighting and phasing (on a given wafer) to be exposed as groups, reticle changes and hence production time may be reduced considerably.

An alternate method for producing weighted transducers is to expose each transducer on a "finger-by-finger" basis. Although this method requires only one reticle (with the mask for a single finger) the number of individual exposures would be increased and overall production time is greater than that for a multiple reticle method.

Clearly the weighting policy for transducers, and corresponding reticle changes, imposes a serious reduction in throughput. Therefore intersymbol interference should be controlled by increased bit spacing or compensation in the interrogation system rather than by transducer weighting policies. Also the effectiveness of phasing policies in minimising second order responses is limited by the positioning accuracy of the wafer stepper, particularly as the operating frequency of devices increase with a corresponding reduction in transducer dimensions.

For phase encoded lines all transducers have equal weightings, and phases may be determined by transducer positioning as already discussed. However due to accuracy

limitations in wafer steppers and the fact that phase encoded systems necessarily depend heavily on correct transducer phasing, phase encoding pads are considered to be a superior method. The positioning of the pads is not critical, as phase shift is dependent on pad dimensions.

Since only four phases are suggested for the 915 MHz system, three pad lengths corresponding to delays of one-quarter, one-half and three-quarter wavelengths are needed. Thus only three reticles are required to encode the delay lines (assuming that an extra reticle is used to produce the pattern containing all transducers and the busbars). Alternatively, one reticle containing the mask of a quarter-wavelength pad and a system of multiple exposures (up to 3 per IDT) to produce the range of phase shifts may be used. As for the amplitude encoded system, multiple exposures using small numbers of reticles is favoured from a throughput viewpoint (providing that reticle changing times are significantly greater than exposure times).

The alternative method of device photo-fabrication using contact exposure is now considered. This method results in the production of multiple devices on a wafer using a large mask and a single exposure. Transducer properties (weightings and phasings) are dependent only on the mask and by using electron-beam lithography in mask production good accuracy and definition is achievable.

However, depending on the number of devices per mask, large numbers of masks may be required to cover all possible codes. The most effective method of mask production for producing different devices is by a programmed electron-beam lithography system. Although this system is capable of producing the complete pattern for each device on a mask, a simpler procedure would be to produce multiple identical patterns of delay lines with all transducers present. Devices produced by such a mask may then be modified individually to cover the required range of codes as described below.

In an amplitude encoded system the codes may be imparted to the devices by selectively disconnecting transducers from the busbars by means of laser trimming. In such a system transducer weighting is not possible but phasing policies are implemented in the mask production. Consideration should be given to possible reflections from metallisation of disconnected transducer fingers although reflections of this type are believed to be small.

For phase encoded systems, coding is achieved by a second stage of exposure which creates the phase encoding pads on the delay lines. Thus masks corresponding to these exposures need only contain various pad combinations. For a 30-bit system (e.g. the 915 MHz system) there must be 2^{30} ($= 1.07 \times 10^9$) different phase encoding pad patterns to be incorporated into masks.

Assuming 200 (say) devices per mask implies that 5.4 million different masks must be produced to cover all possible codes. Due to the large number of masks, mask production time would be the limiting throughput factor.

Consider the case where 256 devices can be produced on a single wafer. The first stage mask (containing transducers and busbars) could also contain pads to encode the first eight bits in the codes. Thus the first stage exposure would produce devices with the first eight bits encoded. A second exposure stage is required to encode the remaining 22 bits (4.2 million codes) and, with 256 devices per mask, implies that a total of 16384 second stage masks are needed. Incorporation of the first 8 bits into the first stage exposure increases the throughput by a factor of almost 256.

It is now evident that production times for a set of coded delay lines not only depends on the number of code bits (and hence the total number of devices) but also on the choice of equipment and strategies used in production. Such strategies are required to resolve the conflict caused by the application of mass production techniques to produce large numbers of devices which must necessarily all be different.

CHAPTER 6

SUMMARY AND CONCLUSIONS

6.0 SUMMARY

Second order responses caused by electro-acoustic regeneration are an undesirable feature of coded one-port SAW delay lines. In certain circumstances second order responses become severe enough to interfere with the coded (primary) response. Rather than manufacture a range of delay lines to investigate second order responses in the prototype interrogation system, mathematical models have been developed to predict the system response. This approach has the advantage of being able to investigate the performance of delay lines with varied codes without the effort entailed in device manufacture.

Two mathematical models, and corresponding computer program implementations, have been developed. The first model simulates the 200 MHz prototype system in the frequency domain by deriving frequency responses for various sections of the system and multiplying these responses together to produce a frequency domain function for the total system response. Delay line characteristics are based on admittance expressions proposed by Burgess[4] (based on Engan[14]) and supported by other researchers.

The system time response (i.e. the time domain signal at the receive antenna) is derived from the system frequency response using sampled functions and a Fast Fourier Transform. The determination of system responses has been performed by using baseband equivalent functions which

- reduces the required computation time, and
- readily yields the envelope of the time response.

The computer program MINISEC produces plots of the envelopes of the total system time response and the secondary response. Plots of calculated receiver channel outputs are also produced. MINISEC is currently used as a batch program.

Multiplication of many frequency domain functions implies that the time response could be achieved by the convolution of corresponding functions in the time domain. The apparent complexity of this type of calculation is reduced by identifying that the time response comprises a number of weighted impulse sequences convolved with appropriate pulse waveforms. The impulse sequences (dependent on the b_k and c_k coefficients) are derived by simple correlation and convolution of an impulse sequence (the a_k coefficients) which depends only on transducer weightings and phasings.

On examination of the time functions within the system, it was found that all baseband equivalent time

functions possessed one of four basic waveshapes (rectangular, triangular, half-triangular and exponential). Furthermore, there are five pulse waveforms in the total system time response which are produced by convolutions of combinations of the four basic waveshapes. The Central Limit Theorem shows that all pulse waveforms are approximately Gaussian and that their salient characteristics (area, mean and variance) are readily derived from the "statistics" of the four basic waveshapes.

Of the five pulse waveforms only three (primary one, secondary one and secondary two) contribute to the system time response in the vicinity of the code as dictated by sequences involving the b_k and c_k coefficients.

The channel outputs are equivalent to the areas under the system time response envelope for corresponding time windows. Intersymbol (or inter-channel) interference due to the time spread of pulse waveforms is modelled by applying the impulse sequences governing the primary one, secondary one and secondary two responses to transversal digital filters whose outputs represent contributions to the channel outputs. The filter coefficients depend on the pulse waveform statistics. Outputs from the filters, when summed, give the receiver channel outputs.

Using these methods, computer program TIMDOM has been developed to model a 200 MHz system with a capacity of up to 30 bits. TIMDOM predicts the channel outputs

from information about coding set transducer weightings and phasings. Although TIMDOM does not calculate the envelope of the system time response as in MINISEC, TIMDOM is used interactively and is faster and easier to use than MINISEC.

A comparison of results from the prototype system and those predicted by the computer models show good agreement. However the comparisons are limited by the number of labels which are available (3 only). Furthermore, the labels which have been constructed have codes which do not produce large second order responses and therefore cannot be used with confidence to verify second order responses predicted for some codes by the computer models. Second order responses for the prototype system were larger than those predicted by the computer models and reflections from transducer metallisation are believed to be responsible. Reflections of this type are easily controlled by suitable transducer design.

The time domain model shows that the secondary responses are controlled by two impulse sequences; the b_k coefficients and, more significantly, the c_k coefficients which are derived from the a_k coefficients by correlation and convolution. By taking advantage of the auto-correlation properties of the a_k coefficients and auto-convolution of the b_k coefficients, methods have been proposed (i.e. transducer phasing policies and shifting the coding set) to minimise the magnitudes of the b_k and c_k coefficients and hence the secondary responses.

Such an insight into the mechanism for producing second order responses is not available from the frequency domain model. Also, by modelling intersymbol interference by digital filtering, transducer weighting policies (based on inverse filters) have been proposed to remove interference between channels. Examples have shown that combined phasing and weighting policies are successful. However phasing policies cannot reduce the secondary one response and this fact imposes an upper limit on the number of code bits that can be used with a system.

Based on investigations of the 200 MHz system with the time domain model, general requirements for delay lines and label antennae in a 915 MHz phase-encoded system are outlined. These requirements will affect the choice of substrate type, transducer design and label antenna for such a system. Finally some implications of transducer phasing and weighting policies on the manufacture of delay lines have been discusssed. It is deduced that current micro-circuit techniques may not be ideally suited to the mass production of coded SAW delay lines.

6.1 CONCLUSIONS

Using results from the prototype 200 MHz system and the two computer models, conclusions have been made by the author concerning the operation of the prototype system and more general systems (including a phase encoded system). These are outlined below:

- Second order responses caused by electro-acoustic regeneration are significant and may be large for some delay lines.

- The magnitude of second order responses (relative to the primary response) is determined by the code of the delay line and the value of the PSR (primary-to-secondary ratio) for the delay line.

- For amplitude encoded delay lines, transducer phasing policies (which spread the spectrum of the primary response) are used to reduce the secondary response by reducing the pre-code of the primary response.

- For both amplitude and phase encoded delay lines, shifting the coding set reduces the effect of the secondary responses (but not the magnitude).

- For both amplitude and phase encoded delay lines, reduction in the secondary responses by increasing the PSR will also reduce the primary response and increase the label insertion loss.

- Metallisation reflections are significant and should be controlled by the use of double-electrode fingers in the transducers.

- There is no advantage in using strong coupling substrates (e.g. $LiNbO_3$) except when using phase encoding pads. In this case strong coupling reduces the size of the pads for a required phase shift.

- For amplitude encoded systems, phasing policies cannot remove the secondary one response and this imposes a limit on the number of code bits which may be used with a delay line (dependent on the PSR).

- Similarly for phase encoded systems (where no phasing policies are possible), there is an upper limit on the number of transducers which may be used on a delay line (dependent on the PSR).

- Significant intersymbol interference exists in the 200 MHz system and transducer weighting policies (based on inverse digital filters) or other forms of compensation in the system may be used to overcome this problem.

6.2 FUTURE DEVELOPMENT

The phasing policies proposed in this thesis are capable of significantly reducing second order responses and are based on a reduction of the pre-code. This results in a consequent reduction in second order responses arising from the convolution of the pre-code with itself and with the code.

More comprehensive phasing policies should take the convolutions into account, thereby producing policies which may not necessarily reduce the magnitude of the pre-code but would result in small responses when the

pre-code is convolved with itself and the code. Having considered this possibility for some time, the author believes that the derivation of phasing policies based on the above principle would not be a trivial exercise.

For phase-encoded systems, where phasing policies are unable to be used, other possibilities may exist for reductions in second order responses. One possibility is a policy whereby transducers may only be placed at certain locations in the coding set. This would result in locations in the coding set where transducers should not be placed resulting in a larger spatial distribution of coding set transducers for a given number of code bits. Such policies have not been investigated but it is believed that some reduction in second order responses is achievable.

Finally, more investigation into equipment and strategies used in delay production is needed and it should be realised that transducer phasing and weighting policies will have substantial bearing on production procedures. Although based on a brief investigation in this thesis, it is the author's opinion that photo-fabrication techniques represent a major obstacle to be overcome before the economic mass production of coded SAW delay lines becomes a reality.

APPENDIX A

LISTINGS OF COMPUTER PROGRAMS FOR PROTOTYPE SYSTEM MODELLING

This appendix comprises programs and subroutines used in the frequency domain and time domain models of the 200 MHz coded label system.

Program MINISEC and associated subroutines (XAD, SF, INPULS, TUNCCT, STATCAP and INPAD) model the system in the frequency domain using the principles described in Chapter 2. The resultant plots are produced on a Calcomp 25-centimetre drum plotter.

Program TIMDOM models the system in the time domain as described in Chapter 3. TIMDOM is used interactively and predicts receiver channel outputs from information about transducers on the delay line. An example showing the use of TIMDOM is included.

Note - All code in the appendices has been written in the FORTRAN IV source language and all programs have been run on the University of Adelaide's Cyber 173 computer. With the exception of the Fast Fourier Transform routines (BFAST, TABLE and CON) and system plotting routines used in MINISEC, all programs and subroutines have been written by the author. BFAST, TABLE and CON are essentially the same as routines used by Burgess[4].

```
      PROGRAM MINISEC (INPUT,OUTPUT)
*
*  FREQUENCY DOMAIN MODEL FOR THE 200 MHZ
*   CODED SAW LABEL SYSTEM ( 20 CODE BITS )
*
      COMPLEX X(1024),Y(1024),ZS,H,ZC,ZT
      DIMENSION N(41),XI(1026),XR(1026),Z(1026)
      REAL LX
      CALL PLOT25
      CALL PAUPLOT(26HBLANK PAPER, BLACK INK PLS,26)
      CALL XLIMIT(300.)
*
*  TRANSDUCER WEIGHTINGS AND PHASINGS
*
      READ*,(N(I),I=1,41)
*
      PRINT21,(N(I),I=1,41)
   21 FORMAT(/,20(2X,I2))
*
      F0=2.E+8
      DATA XMTD,XMDR/1.,1./
*
      DO3 I=1,1024
    3 X(I)=(0.,0.)
*
*  CALCULATE TRANSADMITTANCES, INPUT ADMITTANCES AND STATIC
*  CAPACITANCES IN FREQUENCY DOMAIN
*
      CALL XAD(N,X)
      CALL INPAD(N,X)
      CALL STATCAP(N,X,CL,0.)
*
      DO11 I=1,1024
   11 Y(I)=X(I)
*
* CALCULATE LABEL ANTENNA CIRCUIT PARAMETERS
*
      PI=4.*ATAN(1.)
      Q=5.
      WO=2.*PI*FO
      RX=50.
      LX=Q*RX/WO
      C=1./WO/WO/LX
      CX=C-CL
      PRINT 20,CL,LX,RX,CX
   20 FORMAT(/,5X,4(E12.4,3X))
```

```
*
      DO10 M=1,2
*
* CALCULATE SYSTEM FREQUENCY RESPONSE AND COMBINE
* WITH ONE-PORT ADMITTANCE TO GIVE OVERALL RESPONSE
* IN FREQUENCY DOMAIN
*
*       M=1 :     SECONDARY RESPONSE ONLY
*       M=2 :     PRIMARY AND SECONDARY RESPONSES
*
      DF=1./3.E-6
      FL=F0-512.*DF
      DO6 I=1,1024
      X(I)=Y(I)
      Z(I)= FL+DF*FLOAT(I-1)
      WM=2.*PI*Z(I)+2.
* TAG ANTENNA SOURCE IMPEDANCE
      ZS=CMPLX(RX,WM*LX)
      ZC=CMPLX(0.,-1./WM/C)
      ZT=ZS*ZC/(ZS+ZC)
      X(I)=ZT*ZT/ZS/ZS*(FLOAT(M-1)*X(I)-ZT*X(I)*X(I))
* TRANSMITTER PULSE
      CALL INPULS(X(I),WM)
* TRANSMITTER ANTENNA
      CALL TUNCCT(X(I),WM)
* COUPLING TO LABEL
      X(I)=X(I)*CMPLX(0.,WM*XMTD)
* COUPLING FROM LABEL TO RECEIVE ANTENNA
      X(I)=X(I)*CMPLX(0.,WM*XMDR)
* RECEIVE ANTENNA
      CALL TUNCCT(X(I),WM)
    6 CONTINUE
*
*  FREQUENCY SHIFT TO BASEBAND AND SETUP FOR INVERSE F.F.T.
*
      DO7 I=1,512
      H=X(I+512)
      X(I+512)=X(I)*2.
    7 X(I)=H*2.
*
*       INVERSE F.F.T. TO TIME DOMAIN
*
      CALL BFAST(X,10,1.,DF)
*
* CALCULATE RECEIVER CHANNEL OUTPUTS
      DO15 I=1,1024
      Z(I)=DF*FLOAT(I-1)
   15 XI(I)=0.
      DO12 I=1,79
      J=FLOAT(I-1)*12.8+14.5
      JM4=J-4
      JP4=J+4
      H=(0.,0.)
      DO13 K=JM4,JP4
   13 H=H+X(K)
      A=REAL(H)
      B=AIMAG(H)
      D=SQRT(A*A+B*B)
      DO14 K=JM4,JP4
   14 XI(K)=D
   12 CONTINUE
```

```
*
*  PLOT CHANNEL OUTPUTS VS TIME
*
      CALL SCALE (Z,35.,1024,1)
      CALL SCALE (XI,18.,1024,1)
      CALL AXIS(0.,0.,4HTIME,-4,35.,0.,Z(1025),Z(1026),0)
      CALL AXIS(0.,0.,15HCHANNEL OUTPUTS,15,18.,90.,XI(1025),XI(1026),-1
     *)
      CALL LINE(Z,XI,1024,1,0,0)
      CALL PLOT(40.,0.,-3)
*
*  CALCULATE BASEBAND SIGNAL  (I.E. THE ENVELOPE)
      DO9 I=1,1024
      A=REAL(X(I))
      B=AIMAG(X(I))
    9 XR(I)=SQRT(A*A+B*B)
*
*
*  PLOT ENVELOPE VS TIME
*
      CALL SCALE(Z,35.,1024,1)
      CALL SCALE(XR,18.,1024,1)
      CALL AXIS(0.,0.,4HTIME,-4,35.,0.,Z(1025),Z(1026),0)
      CALL AXIS(0.,0.,5HREPLY,5,18.,90.,XR(1025),XR(1026),-1)
      CALL LINE(Z,XR,1024,1,0,0)
      CALL PLOT(40.,0.,-3)
   10 CONTINUE
      CALL SYMBOL(0.,10.,1.,7HBOX 142,90.,7)
      CALL PLOT(5.,0.,-3)
*
      END
```

```
      SUBROUTINE XAD (N,X)
*
*  XAD CALCULATES THE TOTAL TRANSADMITTANCE OF
*  THE CODED SAW DELAY LINE
*
      DIMENSION N(41)
      COMPLEX X(1024),SF,YI,YJ,ARG
      REAL NI,NJ
*
*    SYSTEM CONSTANTS
*
      DATA F0,AL,D/2.E+8,7.9E-6,1.185E-4/
      DATA V,DEL,E0,EP,W/3160.,9.3E-4,8.85E-12,3.921E-11,1.106E-3/
      PI=4.*ATAN(1.)
      DF=1./3.E-6
*
*  CALCULATE TRANSADMITTANCE FOR ALL POSSIBLE IDT PAIRS
*
      DO1 I=1,40
      IF(N(I).EQ.0) GO TO 1
      IP1=I+1
      DO2 J=IP1,41
      IF(N(J).EQ.0) GO TO 2
      NI=IABS(N(I))
      NJ=IABS(N(J))
      SI=NI/FLOAT(N(I))
      SJ=NJ/FLOAT(N(J))
      FL=F0-512.*DF
*
*  CALCULATE FOR 1024 FREQUENCY POINTS
*
      DO4 K=1,1024
      F=FL+DF*FLOAT(K-1)
      B=F*2.*PI/V
*
*  FOURIER TRANSFORMS OF SOURCE FUNCTIONS
*
      YI=SF(NI,B)
      YJ=SF(NJ,-B)
      A1=0.
      A2=B*D*FLOAT(I-J)
      ARG=CMPLX(A1,A2)
*
*  ADD TRANSADMITTANCE OF EACH IDT PAIR TO TOTAL
*
    4 X(K)=X(K)+2.*YI*YJ*SI*SJ*CEXP(ARG)
    2 CONTINUE
    1 CONTINUE
      DO3 K=1,1024
      WM=2.*PI*(FL+DF*FLOAT(K-1))
*
*  X = TOTAL TRANSADMITTANCE
*
    3 X(K)=X(K)*WM*DEL*PI*PI*W*(E0+EP)/3.437/AL/AL
      RETURN
      END
```

```
      SUBROUTINE INPAD (N,X)
*
*
*  INPAD CALCULATES THE INPUT ADMITTANCE OF AN
*  N-FINGER TRANSDUCER AND ADDS IT TO THE TOTAL
*  ADDMITTANCE, X.
      DIMENSION N(41)
      COMPLEX X(1024)
*
*  SYSTEM CONSTANTS
      DATA W,DEL,EO,EP/1.106E-3,9.3E-4,8.85E-12,3.921E-11/
      FO=2.E+8
      DF=1./3.E-6
      PI=4.*ATAN(1.)
      AL=7.9E-6
      V=3160.
*
      BO=2.*PI*FO/V
*  CALCULATE FOR ALL POSSIBLE TRANSDUCERS
      DO2 I=1,41
      IF(N(I).EQ.0) GO TO 2
      M=IABS(N(I))
      AN=FLOAT(M)
      FL=FO-512.*DF
      GO=W*2.*PI*FO*DEL*AN*AN*(EO+EP)*PI*PI/3.43//4.
      DO3 J=1,1024
      B=2.*PI*(FL+DF*FLOAT(J-1))/V+1.
      X1=AN*AL*(B-BO)/2.
      X2=AN*AL*(B+BO)/2.
      G1=SIN(X1)/X1
      G2=SIN(X2)/X2
      G=G1*G1+G2*G2
      B1=(SIN(2.*X1)-2.*X1)/2./X1/X1
      B2=(SIN(2.*X2)-2.*X2)/2./X2/X2
      B3=B1+B2
*
*  CONDUCTANCE = GO*G
*  SUSCEPTANCE = GO*B3
      X(J)=X(J)+GO*CMPLX(G,B3)
    3 CONTINUE
    2 CONTINUE
      RETURN
      END
```

```
      SUBROUTINE INPULS (X,WM)
*
*  INPULS CALCULATES THE FREQUENCY FUNCTION CORRESPONDING
*  TO THE TRANSMITTER PULSE AT A GIVEN FREQUENCY, WM.
*  THE FUNCTION IS MULTIPLIED WITH THE ADMITTANCE, X.
*
      COMPLEX X,A1,A2,B3,B4
      DT=2.5E-8
      PI=4.*ATAN(1.)
      FO=2.E+8
      WO=2.*PI*FO
      B1=(WM+WO)*DT/2.
      B2=(WM-WO)*DT/2.
      B3=CMPLX(0.,-B1)
      B4=CMPLX(0.,-B2)
      A1=SIN(B1)/(WM+WO)*CEXP(B3)
      A2=SIN(B2)/(WM-WO)*CEXP(B4)
      X=X*(A1-A2)*CMPLX(0.,1.)
      RETURN
      END
```

```
      SUBROUTINE TUNCCT (X,WM)
*
*  TUNCCT CALCULATES THE FREQUENCY RESPONSE OF A
*  GENERAL TUNED CIRCUIT WITH CENTRE FREQUENCY OF
*  200 MHZ AND Q OF 5 .
*
      COMPLEX X
      PI=4.*ATAN(1.)
      WO=2.*PI*2.E+8
      A=WM/WO
      X=X*CMPLX(0.,A/5.)/CMPLX(1.-A*A,A/5.)
      RETURN
      END
```

```
      SUBROUTINE STATCAP(N,X,C,D)
*
*  STATCAP CALCULATES THE STATIC CAPACITANCE OF A
*  N-FINGER TRANSDUCER AND HENCE THE TOTAL CAPACITANCE
*  OF THE DELAY LINE.
*
      DIMENSION N(41)
      COMPLEX X(1024)
*
*  SYSTEM CONSTANTS
      DATA E0,EP,W,F0/8.85E-12,3.921E-11,1.106E-3,2.E+8/
      DF=1./3.E-6
*
      S=0.
      DO2 I=1,41
      IF(N(I).EQ.0) GO TO 2
      NI=IABS(N(I))
      S=S+FLOAT(NI)-1.
    2 CONTINUE
      C=0.5*(E0+EP)*W*S
      PI=4.*ATAN(1.)
      FL=F0-512.*DF
*
* IF D IS NON-ZERO, ADD CAPACITANCE DIRECTLY
*  TO ADMITTANCE, X.
*
      IF(D.EQ.0.) RETURN
      DO3 I=1,1024
      WM=2.*PI*(FL+DF*FLOAT(I-1))
    3 X(I)=X(I)+CMPLX(0.,WM*C)
      RETURN
      END
```

```
      COMPLEX FUNCTION SF (AN,B)
*
*  SF CALCULATES THE FREQUENCY DOMAIN FORM ( I.E.
*  THE FOURIER TRANSFORM ) OF THE SOURCE FUNCTION
*  OF AN N-FINGER TRANSDUCER (AT A GIVEN FREQUENCY, B )
*
      COMPLEX B3
      PI=4.*ATAN(1.)
      AL=7.9E-6
      B0=PI/AL
      B1=0.
      B2=(AN-1.)*PI
      B3=CMPLX(B1,B2)
      SF=SIN(AN*AL*(B-B0)/2.)/(B-B0)*CEXP(B3)+SIN(AN*AL*(B+B0)/2.)/(B+B0
     $)*CEXP(-B3)
      RETURN
      END
```

```
      SUBROUTINE BFAST(Y,M,SIGN,DXY)
      COMPLEX A,U,W,T,Y
      DIMENSION A(1024),Y(2)
      COMMON /PERM/MM(1024)
      COMMON /TAB/TC(10),TS(10)
      DATA MSET/0/
*
      IF(MSET.EQ.M) GO TO 1
      CALL TABLE(M)
      CALL CON(M)
      MSET=M
    1 CONTINUE
*
      N=2**M
      DO7 I=1,N
      J=MM(I)
    7 A(I)=Y(J)
*
      DO20 L=1,M
      LE=2**L
      LE1=LE/2
      U=(1.,0.)
      W=CMPLX(TC(L),TS(L)*SIGN)
      DO20 J=1,LE1
      DO10 I=J,N,LE
      IP=I+LE1
      T=A(IP)*U
      A(IP)=A(I)-T
      A(I)=A(I)+T
   10 CONTINUE
      U=U*W
   20 CONTINUE
*
      DO30 I=1,N
      Y(I)=A(I)*DXY
   30 CONTINUE
      DXY=1./(DXY*N)
      RETURN
      END
```

```
      SUBROUTINE TABLE(M)
      COMMON /TAB/TC(10),TS(10)
      PI=4.*ATAN2(1.,1.)
      DO 1 I=1,M
      TC(I)=COS(PI/(2**(I-1)))
    1 TS(I)=SIN(PI/(2**(I-1)))
      RETURN
      END
```

```
      SUBROUTINE CON (M)
      INTEGER A,T
      COMMON /PERM/ A(1024)
      N=2**M
      NV2=N/2
      NM1=N-1
      J=1
      DO1 I=1,N
    1 A(I)=I
      DO7 I=1,NM1
      IF(I.GE.J) GO TO 5
      T=A(J)
      A(J)=A(I)
      A(I)=T
    5 K=NV2
    6 IF(K.GE.J) GO TO 7
      J=J-K
      K=K/2
      GO TO 6
    7 J=J+K
      RETURN
      END
```

```
      PROGRAM TIMDOM (INPUT,OUTPUT)
*  TIME DOMAIN MODEL FOR 200MHZ
*  CODED SAW LABEL SYSTEM
*
      DIMENSION A(62),B(62),S1(63),S2(63)
      DIMENSION C1(31),C2(31),F1(3),F2(3)
      DIMENSION C3(31),F3(3)

*   ISI FILTER COEFFICIENTS
      DATA F1/ 0.184,0.809, 0.447/
      DATA F2/ 0.072,0.710, 0.638/
      DATA F3/ 0.237,1., 0.237/

   98 PRINT10
   10 FORMAT(/,2X,*NO. OF CODE BITS =?  *)
      READ*,N
      IF(N.EQ.0) GO TO 99
      NP1=N+1
      DO1 I=1,NP1
    1 A(I)=0.
      A(1)=1
      NP2=N+2
      N=N*2+1
      PRINT11
   11 FORMAT(/,2X,*IDT WEIGHTINGS =?*)
      READ*,(A(I),I=NP2,N)
      SIGN=-1.
      DO7 I=NP2,N
      SIGN=-1.*SIGN
    7 A(I)=A(I)/11.*SIGN

*  CALCULATE PRIMARY RESPONSE
      DO2 J=1,N
      NMJP1=N-J+1
      B(J)=0.
      DO2 I=1,NMJP1
    2 B(J)=B(J)+A(I)*A(I+J-1)
      B1=B(1)/2.
      B(1)=0.
      B(N+1)=0.

*  CALCULATE SECONDARY RESPONSES
      DO3 J=1,N
      S1(J)=0.
      S2(J)=B(J)*B1
      DO3 I=1,J
    3 S1(J)=S1(J)+B(I)*B(J-I+1)
      S1(N+1)=0.
      DO4 I=2,N
    4 S1(N+1)=S1(N+1)+B(I)*B(N-I+2)
      S2(N+1)=0.
```

```
*  CALCULATE CHANNEL OUTPUTS INCLUDING ISI
      K=0
      DO5 J=NP2,N
      K=K+1
      C1(K)=0.
      C2(K)=0.
      C3(K)=0.
      DO5 I=1,3
      C1(K)=C1(K)+F1(I)*S1(J-I+2)
      C2(K)=C2(K)+F2(I)*S2(J-I+2)
    5 C3(K)=C3(K)+F3(I)*B(J-I+2)

      PRINT12
   12 FORMAT(/,2X,*CHANNEL OUTPUTS*,/)

      DO6 J=1,K
      X=C3(J)-C2(J)/149.-C1(J)/149.
      CALL BAR(X)
    6 CONTINUE
      GO TO 98
   99 CONTINUE
      END


*  PRODUCE BAR-GRAPH OF CHANNEL OUTPUTS
      SUBROUTINE BAR(X)
      Y=ABS(X)
      STAR=1H*
      NS=Y*20.+0.5
      IF(NS.EQ.0) GO TO 3
      IF(NS.GT.40) NS=40
      PRINT2,X,(STAR,I=1,NS)
    2 FORMAT(5X,F7.2,2X,40A1)
      RETURN
    3 PRINT2,X
      RETURN
      END
```

```
COMMAND- USE,RESPONSE,DMG,ASW,ASW

COMMAND- GET,TIMDOM

COMMAND- CONNECT,INPUT,OUTPUT

COMMAND- TIMDOM

 NO. OF CODE BITS =? 20

 IDT WEIGHTINGS =? 11 0 0 11 0 11 0 11 0 11
                    0 0 0 0 0 0 0 0 0 0

 CHANNEL OUTPUTS

        .99  ********************
        .22  ****
       -.27  *****
      -1.00  ********************
       -.47  *********
       -.96  ********************
       -.44  *********
       -.94  ********************
       -.43  *********
       -.93  ********************
       -.20  ****
        .04  *
       -.00
        .00
       -.01
       -.01
       -.02
       -.01
       -.01
       -.01

 NO. OF CODE BITS =? 0

      END  TIMDOM
     020500 MAXIMUM EXECUTION FL.
       .154 CP SECONDS EXECUTION TIME.
COMMAND-
```

APPENDIX B

LISTINGS OF COMPUTER PROGRAMS FOR DELAY LINE FABRICATION

This appendix lists the programs and subroutines written by the author in the early stages of his project to assist in the production of coded SAW delay lines for use in the prototype 200 MHz system.

Program TAPER20 and associated subroutines IDT, FINGER and ALIGN produce a plot of the metallisation pattern (250 times full size) of the delay line. The patterns were plotted on a Calcomp 75-centimetre drum plotter using special drafting film and photo-reduced to produce masks for actual device fabrication. Due to size limitations of the reduction camera, the plots were produced as two halves and alignment marks provided for correct pattern registration at the exposure stage of fabrication.

TAPER20 was subsequently modified to produce the plots on a XYNETICS flat-bed plotter to achieve greater accuracy and faster plotting. However, the method of producing masks from computer plots was abandoned before any plots were attempted on the XYNETICS plotter. Since the modifications to TAPER20 were minor, the program listings for use with the XYNETICS plotter have not been included.

Program CONTROL was written to produce a sequence of instructions (on paper tape) which are interpreted by the micro-computer controlled "step-and-repeat" camera exposure system at the University of New South Wales. The instructions control the movement and exposure action of the camera. Devices were produced by multiple exposures using only two masks -

- a long thin rectangle - used for producing the fingers of the transducers
- a large square - used for producing busbars and connection pads

The repertoire of controller instructions is included for information and comparison with the computer program listing.

```
      PROGRAM TAPER20(IDTDAT,TAPE1=IDTDAT,OUTPUT)
*
*****************************************************************
*
*  PROGRAM TAPER  INITIALISES THE PLOT75 FILE, INITIATES THE
*  PLOTTING OF THE DELAY LINE BUSES AND ALIGNMENT MARKS OF
*  EACH HALF OF THE PATTERN AND DETERMINES THE POSITION OF
*  INDIVIDUAL TRANSDUCERS ON THE PATTERN.
*  DATA WHICH IS READ FROM THE INPUT FILE CONTAINS THE
*  INFORMATION ABOUT TRANSDUCER WEIGHTINGS AND PHASINGS.
*  THIS INFORMATION COMPRISES 40 INTEGER VALUES CORRESPONDING
*  TO THE 40 POSSIBLE LOCATIONS OF TRANSDUCERS ON THE DELAY
*  LINE.
*  PROGRAM TAPER  PASSES ON THIS DATA TO S/R IDT FOR THE
*  PLOTTING OF INDIVIDUAL TRANSDUCERS.
*
*  THE PLOT PRODUCED BY THIS PROGRAM IS 250 TIMES FULL SIZE
*  AND IS TO BE USED FOR PHOTOREDUCTION TO PRODUCE THE
*  ACTUAL METALLISATION PATTERN OF THE ASW DELAY LINE.
*
*****************************************************************
*
      DIMENSION IW(40)
*
*  READ IN TRANSDUCER WEIGHTINGS AND PHASINGS
      READ(1,100) (IW(K),K=1,40)
  100 FORMAT(39(I3,/),I3)
*
*   INITIALISE PLOT FILE AND REQUEST NECESSARY PEN AND PAPER
      CALL PAUPLOT(17HDRAFTING FILM PLS,17)
      CALL PAUPLOT(17H0.4MM BLK PEN PLS,17)
      CALL PAUPLOT(18HAPPROX 1/2 HR PLOT,18)
      CALL PLOT75
*
      DATA FW,FL,PD,W,TSP/0.099,23.44,0.04,29.23,2.96/
*
*  PLOT BUSES FOR FIRST HALF OF THE PATTERN
      CALL FINGER(5.,52.62,67.5,56.57,0,PD)
      CALL FINGER(5.,23.39,67.5,19.44,0,PD)
*
*  PLOT FIRST 20 TRANSDUCERS ( RETRACING OUTLINES FOR
*  IMPROVED DEFINITION )
      DO2 J=1,2
      XC=5.+21.*FW/2.
      YS=38.
      FSP=FW
      IFILL=1
      IF(J.EQ.2) IFILL=0
      DO2 I=1,20
      IF(IW(I).EQ.0) GO TO 3
      NF=IABS(IW(I))
      IPH=ISIGN(1,IW(I))
      XS=XC-FW*FLOAT(2*NF-1)/2.
      CALL IDT(XS,YS,IPH,NF,FSP,FW,W,PD,FL,IFILL)
    3 XC=XC+TSP
    2 CONTINUE
*
```

```
*  PLOT ALIGNMENT MARKS ON FIRST HALF OF PATTERN
      CALL ALIGN(XC,62.57,4.)
      CALL ALIGN(XC,13.44,4.)
*
*  ADD IDENTIFICATION AND TERMINATE PLOT FILE
      CALL SYMBOL(15.,3.,2.,21HASW DELAY LINE PART 1,0.,21)
      CALL SYMBOL(20.,0.,2.,11HBOX 142 PLS,0.,11)
      CALL PLOT(74.,0.,3)
      CALL PLOT(74.,0.,-1)
*
*  INITIALISE A NEW PLOT FILE FOR THE SECOND HALF OF PATTERN
*  AND PROCEDE AS FOR FIRST HALF OF PATTERN
      CALL NEWPLOT
      CALL PAUPLOT(15HAPPROX 3HR PLOT,15)
      CALL PLOT75
      CALL FINGER(5.,52.62,67.5,56.57,0,PD)
      CALL FINGER(5.,23.39,67.5,19.44,0,PD)
      CALL ALIGN(7.+21.*FW/2.,62.57,4.)
      CALL ALIGN(7.+21.*FW/2.,13.44,4.)
      DO4 J=1,2
      XC=7.+21.*FW/2.
      YS=38.
      IFILL=1
      IF(J.EQ.2) IFILL=0
      DO4 I=21,40
      IF(IW(I).EQ.0) GO TO 5
      NF=IABS(IW(I))
      IPH=ISIGN(1,IW(I))
      XS=XC-FW*FLOAT(2*NF-1)/2.
      CALL IDT(XS,YS,IPH,NF,FSP,FW,W,PD,FL,IFILL)
    5 XC=XC+TSP
    4 CONTINUE
      CALL SYMBOL(15.,3.,2.,21HASW DELAY LINE PART 2,0.,21)
      CALL SYMBOL(20.,0.,2.,11HBOX 142 PLS,0.,11)
      CALL PLOT(74.,0.,3)
      CALL PLOT(74.,0.,-1)
      END
```

```
      SUBROUTINE IDT (XS,YS,IPH,NF,FSP,FW,W,PD,FL,IFILL)
*
*****************************************************************
*
*  SUBROUTINE IDT IS RESPONSIBLE FOR THE LAYOUT OF EACH
*  INTERDIGITAL TRANSDUCER.
*
*  PARAMETERS FOR THIS SUBROUTINE ARE -
*  XS - X COORDINATE AT WHICH TRANSDUCER BEGINS ON PLOT
*  YS - Y COORDINATE OF TRANSDUCER CENTRE ON PLOT
*  IPH - SETS ELECTRICAL PHASE OF TRANSDUCER,
*                                EITHER + OR - ONE
*  NF - NUMBER OF FINGERS IN THE TRANSDUCER
*  FSP - SPACING BETWEEN FINGERS
*  FW - WIDTH OF EACH FINGER
*  W - WIDTH BETWEEN DELAY LINE BUSES
*  PD - DIAMETER OF PEN USED ON PLOTTER
*  FL - LENGTH OF FINGERS IN THE TRANSDUCER
*
*  GIVEN DATA ABOUT THE LOCATION AT WHICH THE TRANSDUCER IS
*  TO BE PLOTTED, S/R IDT CALCULATES THE POSITION OF EACH
*  FINGER IN THE TRANSDUCER AND INITIATES PLOTTING BY
*  REPEATED CALLS TO S/R FINGER. EACH FINGER IS FILLED IN
*  BY SETTING THE CORRESPONDING PARAMETER IN S/R FINGER.
*
*****************************************************************
*
      J=-1
      IF(IPH.EQ.-1) J=-1*J
      J=-1*J
      N=0
      XA=XS
      CORR=W/2.+PD*5.
    2 YA=YS+CORR*FLOAT(J)
      YB=YA-(FL+PD*5.)*FLOAT(J)
      XB=XA+FW
      CALL FINGER (XA,YA,XB,YB,IFILL,PD)
      J=-1*J
      N=N+1
      IF(N.GE.NF) RETURN
      XA=XB+FSP
      GO TO 2
      END
```

```
      SUBROUTINE FINGER (X1,Y1,X2,Y2,IFILL,PD)
*
*****************************************************************
*
*  SUBROUTINE FINGER IS SIMPLY A ROUTINE WHICH DRAWS RECTANGLES
*  AND FILLS THEM IN IF DESIRED. S/R FINGER IS CALLED BY
*  S/R IDT TO DRAW THE INDIVIDUAL FINGERS OF EACH TRANSDUCER
*  AND BY PROGRAM LAYOUT TO DRAW THE LARGER DELAY LINE BUSES.
*
*  THE PARAMETERS ARE -
*  XA,YA,XB,YB - X AND Y COORDINATES OF DIAGONALLY OPPOSITE
*                CORNERS OF THE RECTANGLE TO BE PLOTTED
*  IFILL - IF THIS PARAMETER IS SET TO 1, THE RECTANGLE BEING
*          PLOTTED WILL ALSO BE FILLED IN
*  PD - DIAMETER OF THE PEN BEING USED ON THE PLOTTER
*
*  DUE TO THE PRECISION WHICH THIS PLOTTING REQUIRES,
*  S/R FINGER ALLOWS FOR THE PEN DIAMETER WHEN PLOTTING.
*
*****************************************************************
*
      PD2=PD/2.
      DX=X2-X1
      DY=Y2-Y1
      IF(DX.GT.0.) GO TO 2
      LX=-1
      XA=X1-PD2
      XB=X2-PD2
      GO TO 3
    2 XA=X1+PD2
      XB=X2-PD2
      LX=1
    3 IF(DY.GT.0.) GO TO 4
      YA=Y1-PD2
      YB=Y2+PD2
      LY=-1
      GO TO 5
    4 YA=Y1+PD2
      YB=Y2-PD2
      LY=1
    5 CALL PLOT(XA,YA,3)
      CALL PLOT(XA,YB,2)
      CALL PLOT(XB,YB,1)
      CALL PLOT(XB,YA,1)
      CALL PLOT(XA,YA,1)
      IF(IFILL.EQ.1) GO TO 6
      RETURN
    6 IF(ABS(DX).LT.ABS(DY)) GO TO 7
      L=0
      N=1
      Y=YA
      X=XB
      DEL=PD2*FLOAT(LY)
   10 Y=Y+DEL
      CALL PLOT(X,Y,1)
      IF(ABS(Y-YB).LT.PD2) RETURN
```

```
      IF(N.LT.0) GO TO 8
      X=XA
      GO TO 9
    8 X=XB
    9 N=-1*N
      GO TO 10
    7 L=0
      N=1
      X=XA
      Y=YB
      DEL=PD2*FLOAT(LX)
   11 X=X+DEL
      CALL PLOT(X,Y,1)
      IF(ABS(X-XB).LT.PD2) RETURN
      IF(N.LT.0) GO TO 12
      Y=YA
      GO TO 13
   12 Y=YB
   13 N=-1*N
      GO TO 11
      END
```

```
      SUBROUTINE ALIGN (X,Y,S)
*
*****************************************************************
*
*  SUBROUTINE ALIGN PROVIDES THE NECESSARY ALIGNMENT MARKS
*  ON THE PLOT. THESE MARKS ARE USED FOR ALIGNING THE TWO
*  HALVES OF THE DELAY LINE PATTERN DURING PHOTOFABRICATION
*  OF THE ASW DEVICE.
*
*****************************************************************
*
      D=S/2.
      CALL FINGER(X-D/5.,Y+D,X+D/5.,Y-D,0,0.)
      CALL FINGER(X-D,Y+D/5.,X+D,Y-D/5.,0,0.)
      RETURN
      END
```

Section of a plot produced by TAPER20 showing transducers, busbars and alignment marks.

```
      PROGRAM CONTROL (PHOTOPLOT,TAPE2=PHOTOPLOT)
      DIMENSION M(20)
      DATA M/11,0,-11,0,11,-4,-4,11,-2,-7,-8,0,10,3,11,-5,0,2,-6,11/
      WRITE (2,2)
    2 FORMAT(*I LOAD AND POSITION FINGER RETICULE *)
      WRITE(2,3)
    3 FORMAT(*P ZX ZY *)
      CALL XDUC(11)
      WRITE(2,4)
    4 FORMAT(*GX +5805 ZX *)
      DO5 I=1,20
      IF(M(I).EQ.0) GO TO 5
      NW=10*(2*M(I)-1)
      NXPOS=I*300-NW/2
      WRITE(2,6) NXPOS
    6 FORMAT(*RY GX +*,I6.6,* *)
      CALL XDUC(M(I))
    5 CONTINUE
      WRITE(2,7)
    7 FORMAT(*I LOAD AND POSITION BUS RETICULE *)
      WRITE(2,8)
    8 FORMAT(*P ZX MY -400 ZY MX -1700 E MX -2000 E MY +3360 MX +2000 E
     **)
      WRITE(2,9)
    9 FORMAT(*GX +5900 E *,4(*MX +2000 E *),*RY E *,4(*MX -2000 E *))
      WRITE(2,10)
   10 FORMAT(*GY -4000 GX +1700 E *,28(*MY +400 MX +20 E *),*MY +400 MX
     *-2000 *)
      CALL PAD
      WRITE(2,11)
      CALL PAD
      WRITE(2,12)
      CALL PAD
      WRITE(2,13)
      CALL PAD
      WRITE(2,14)
      CALL PAD
      WRITE(2,15)
      CALL PAD
      WRITE(2,16)
      WRITE(2,17)
   11 FORMAT(*GX -600 GY -10000 *)
   12 FORMAT(*GX -9980 GY -5600 *)
   13 FORMAT(*GX -9700 GY +3360 *)
   14 FORMAT(*GX +15900 GY +3360 *)
   15 FORMAT(*GX +15620 GY -5600 *)
   16 FORMAT(*I JOB COMPLETE *)
   17 FORMAT(*P *)
      END
```

```
      SUBROUTINE XDUC(N)
      M=IABS(N)
      I=0
      IF(N.GT.0) GO TO 1
      WRITE(2,2)
    2 FORMAT(*MY +83 E RY MX +20 *)
      I=1
    1 WRITE(2,3)
    3 FORMAT(*E *)
      I=I+1
      IF(I.EQ.M) RETURN
    6 WRITE(2,4)
    4 FORMAT(*MY +83 MX +20 E *)
      I=I+1
      IF(I.EQ.M) RETURN
      WRITE(2,5)
    5 FORMAT(*RY MX +20 E *)
      I=I+1
      IF(I.EQ.M) RETURN
      GO TO 6
      END
      SUBROUTINE PAD
      WRITE(2,2)
      WRITE(2,2)
      WRITE(2,2)
    2 FORMAT(*E *,14(*MY +400 MX +20 E *),*MY -5600 MX +1720 *)
      RETURN
      END
```

Command Repertoire for Photoplotter Control

Command	Meaning
M	Return to monitor without reset.
ZX	Reset the origin of X axis to current position.
ZY	Reset the origin of Y axis to current position.
RX	Return to the X axis origin.
RY	Return to the Y axis origin.
MX mm	Move in X direction by amount of 'mm' units.
MY mm	Move in Y direction by amount of 'mm' units.
E	Open shutter for preset exposure period.
I	Interpret remainder of line as instructions to operator (i.e. to be ignored by photoplotter).
P	Pause in software execution and switch off paper tape reader.
C	Resume software execution and switch on paper tape reader.
TL	Traverse to X movement lower limit.
TR	Traverse to X movement upper limit.
TD	Traverse to Y movement lower limit.
TU	Traverse to Y movement upper limit.
GX nnn	Move to absolute location 'nnn' for X axis.
GY nnn	Move to Absolute location 'nnn' for Y axis.
RL	Report current location co-ordinates.
ME	Enable manual control from front panel.
MD	Disable manual control from front panel.

APPENDIX C

INPUT SUSCEPTANCE AND THE HILBERT TRANSFORM

The input admittance, Y(x), of an interdigital transducer is -

$$Y(x) = G(x) + jB(x)$$

The input conductance, G(x), has the form -

$$G(x) = G_o \frac{\sin^2 x}{x^2}$$

where G_o is a constant

and x is a normalised frequency variable.

The input susceptance, B(x), is the Hilbert Transform of G(x) due to the causal nature of the input admittance in the time domain.

Now g(y) is defined (in the y-domain) as -

$$g(y) = \int_{-\infty}^{+\infty} G(x) e^{jyx} \, dx$$

From the definition of G(x), it can be shown that -

$$g(y) = \frac{1}{2} G \, \Delta\left(\frac{y}{2}\right)$$

$$\Delta(y) = 1 - |y| \qquad \text{for } |y| \leqslant 1$$

$$= 0 \qquad \text{for } |y| > 1$$

Also d(y) is defined as -

$$d(y) = g(y) \qquad \text{for} \quad y \leqslant 0$$

$$= 0 \qquad \text{for} \quad y < 0$$

and in the x-domain, D(x), is defined as

$$D(x) = \int_{-\infty}^{\infty} d(y) e^{-jyx} \, dy$$

From the definition of d(y), it can be shown that -

$$D(x) = \frac{1 - e^{-j2x}}{4x^2} + \frac{1}{j2x}$$

The y-domain form of the Hilbert Transform is -

$$h(y) = -j \text{ sgn } y$$

where

$$\text{sgn } y = 1 \; ; \quad y > 0$$

$$= -1 \; ; \quad y < 0$$

$$= 0 \; ; \quad y = 0$$

The y-domain susceptance must be -

$$b(y) = g(h)h(y)$$

$$= G_o[-jd(y) + jd(-y)]$$

and hence the x-domain susceptance is -

$$B(x) = G_o[-jD(x) + jD(-x)]$$

Substituting for D(x) gives

$$B(x) = \frac{jG_o}{2} \; \frac{e^{-j2x} - e^{+j2x} - 2x}{2x^2}$$

$$= G_o \; \frac{\sin 2x - 2x}{2x^2}$$

APPENDIX D

LISTING OF COMPUTER PROGRAM FOR TRANSDUCER WEIGHTING

This appendix contains the listing of FORTRAN program WEIGHT which is used to calculate transducer weightings (i.e. number of fingers in each transducer) for reducing intersymbol interference. The main algorithm is based on the inverse of the transversal filter representing intersymbol interference in the 200 MHz prototype system as described in Section 5.4.

WEIGHT is used interactively and an example is included.

```
      PROGRAM WEIGHT (INPUT,OUTPUT)
      DIMENSION KTAB(12),A(32),C(32),K(32)
      INTEGER PHI
      DATA KTAB/0,0,2,3,4,5,6,7,8,9,10,11/
      DATA A0,A1,A2/-0.24,1.,-0.24/
    7 PRINT1
    1 FORMAT(///,2X,*NO. OF CODE BITS =?*)
      READ*,N
      IF(N.EQ.0) GO TO 8
      PRINT2
    2 FORMAT(/,2X,*CODE =?*)
      READ*,(A(I),I=1,N)
      NP1=N+1
      A(NP1)=0.
      X=0.
      Y=0.
      Z=0.
      AMAX=0.
      DO4 I=1,NP1
      Z=Y
      Y=X
      X=A(I)
      C(I)=X*A0+Y*A1+Z*A2
      V=ABS(C(I))
      IF(V.GT.AMAX) AMAX=V
    4 CONTINUE
      PHI=1
      DO5 I=1,NP1
      ISIGN=1
      PHI=-1*PHI
      C(I)=C(I)/AMAX
      J=ABS(C(I))*11.+0.5
      IF(C(I).LT.0.) ISIGN=-1
    5 K(I)=KTAB(J+1)*ISIGN*PHI
      PRINT3
    3 FORMAT(5X,*IDT WEIGHTINGS*)
      PRINT6,(K(I),I=2,NP1)
    6 FORMAT(5X,10(I3,3X))
      GO TO 7
    8 CONTINUE
      END
```

```
COMMAND- USE,RESPONSE,DMG,ASW,ASW

COMMAND- GET,WEIGHT

COMMAND- CONNECT,INPUT,OUTPUT

COMMAND- WEIGHT

 NO. OF CODE BITS =? 10

 CODE =? 1 0 0 1 0 1 0 1 0 1

    IDT WEIGHTINGS
     11     3    -3   -11    -5   -11    -5   -11    -5   -11

 NO. OF CODE BITS =?0

      END  WEIGHT
     020000 MAXIMUM EXECUTION FL.
       .089 CP SECONDS EXECUTION TIME.
COMMAND-
```

APPENDIX E

ADMITTANCE CALCULATIONS

This appendix demonstrates the consistency between admittance expressions used by Engan[11] and those used in this thesis (based on Burgess and Engan). A sample calculation is included to illustrate the application of the admittance expressions.

According to Engan[11], the input conductance of a transducer is given by:

$$G_1 = \omega W(\varepsilon_p+\varepsilon_o)\Delta.2M^2R_G \qquad (E.1)$$

where ω = radian frequency

W = transducer aperture

Δ = coupling coefficient (= $\Delta v/v$)

ε_o = permittivity of free space

ε_p = effective permittivity of the substrate

$= (\varepsilon_{xx}\varepsilon_{zz} - \varepsilon_{xz}^2)^{\frac{1}{2}}$

M = number of electrical periods of the transducer

= number of wavelengths (at the transducer's fundamental frequency of operation) contained in the transducer's length

R_G = "normalised conductance" (Engan's nomenclature)

= the conductance of a transducer with unity aperture and one electrical period length. R_G is a function of transducer geometry and metallisation ratio (See Fig. 8 of Engan[11]). R_G is a measure of spatial harmonics of the transducer's electric field.

It has been shown (by Hartmann and Secrest[18]) that for a double-electrode finger transducer, the third spatial harmonic is the same magnitude as the fundamental. It follows that the value of R_G is the same when calculating admittances for either fundamental or third harmonic operation of a transducer with double-electrode fingers.

For a given transducer (where the number of fingers is constant), M will be the same for fundamental and third harmonic operation. Equation E.1 then implies that the difference between conductances for fundamental and third harmonic is dependent only on the 'ω' term. Thus for a given transducer with double-electrode fingers, the third harmonic conductance is three times greater than the fundamental conductance. This fact is stated by Engan[11] and supported by his calculations and measurements.

Equation E.1 can be compared with the conductance expression used in this thesis (equation 3.17) viz.:

$$G_o = \frac{-\pi^2 N^2 (\varepsilon_p + \varepsilon_o) |\omega_o| \Delta W}{13.75} \qquad (E.2)$$

for equations E.1 and E.2 to be consistent:

$$2M^2R_G = \frac{\pi^2N^2}{13.75}$$

but since $M = \frac{N}{2}$ = electrical periods of transducer, this implies that:

$$R_G = \frac{2\pi^2}{13.75} = 1.44$$

This value for R_G is confirmed by Engan's Fig. 8, and equations E.1 and E.2 are shown to be equivalent.

Sample Calculation

To exemplify the calculation and to support conductance values quoted in Section 5.8, consider a transducer with 25 (say) double-electrode fingers which is to operate at 915 megahertz. Operation at third harmonic is envisaged and hence the transducer's fundamental frequency is 305 megahertz. Transducers on both quartz and lithium niobate are considered.

- Transducer aperture is assumed to be 0.5 millimetres and hence $W = 5 \times 10^{-4}$
- There are 25 ÷ 2 = 12.5 electrical periods (fundamental wavelengths) in the transducer and thus M = 12.5
- For double-electrode fingers and metallisation ratio of 0.5 R_G = 1.4 from Engan's Fig. 8. (This value was also found from a comparison of

equations E.1 and E.2 above)

Operating frequency is 915 megahertz and thus $\omega = 2\pi \times 915 \times 10^6$.

Quartz

For quartz, $\varepsilon_p = 4.4\varepsilon_o$ and $\Delta = 9.1 \times 10^{-4}$.

From equation E.1, $G_1 = 5.47 \times 10^{-5}$.

$LiNbO_3$

For $LiNbO_3$, $\varepsilon_p = 46\varepsilon_o$ and $\Delta = 2.25 \times 10^{-2}$.

From equation E.2, $G_1 = 1.18 \times 10^{-2}$.

For fundamental operation, conductance values would be reduced by a factor of 3, viz.:

Quartz	:	G_1(305 MHz)	=	1.82×10^{-5}
LiNbO	:	G_1(305 MHz)	=	3.9×10^{-3}

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