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**A SAW BASED TRANSVERSAL FILTER
FOR ADAPTIVE TIME DOMAIN EQUALIZATION**

by

A. Neil Birkett, B. Eng.

**A thesis submitted to the Faculty of Graduate Studies and Research
in partial fulfilment of the requirements for the degree of
Master of Engineering**

**Department of Electronics
Carleton University
Ottawa, Ontario
March 1988**

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**CARLETON UNIVERSITY
FACULTY OF GRADUATE STUDIES AND RESEARCH**

**The undersigned recommend to the Faculty of Graduate Studies and
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FOR ADAPTIVE TIME DOMAIN EQUALIZATION"**

**submitted by A. Neil Birkett
in partial fulfillment of the requirements for
the degree of Master of Engineering.**

.....
Thesis Supervisor

.....
Chairman, Department of Electronics

**Carleton University
March 1988**

ABSTRACT

Surface Acoustic Wave (SAW) filter technology has reached the stage of maturity where inclusion into high spectral efficiency digital radio systems has not only become possible but also economically attractive. A research study was undertaken to investigate an application of a tapped SAW delay line in an adaptive time domain equalizer (ATDE) for use in future radio systems using multi-level quadrature amplitude modulation with symbol rates approaching the channel bandwidth.

This study outlines the design and construction of a seven tap SAW delay line and the accompanying electronics of a 70 MHz IF ATDE transversal filter. The study consists of a tap reflection analysis in the SAW delay line and the methods taken to suppress it and other spurious second-order effects.

The results show that tap reflections in the SAW device can be effectively suppressed proving that SAW delay lines can be used as an effective alternative to lumped element delay lines in the IF path. The measured performance of the SAW based ATDE shows that fading nulls 15 dB deep can be corrected for, demonstrating the feasibility of the proposed application .

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LIST OF SYMBOLS

<u>Symbol</u>	<u>Description</u>
a	In-phase delayed signal amplitude coefficient.
a'	New "a" after frequency translation.
a_c	Cavity width.
A	Fade level parameter.
A_n	Amplitude equivalent of nth finger overlap.
b	Quadrature-phase delayed signal amplitude coefficient.
b	Transducer mismatch ratio.
b'	new "b" after frequency translation.
B	Notch level parameter.
B	Bandwidth.
B_a	Acoustic susceptance.
B_d	Doppler spread of a multipath channel.
c_n	Tap coefficient of nth tap.
c_{In}	In-phase tap coefficient for nth tap.
c_{Qn}	Quadrature-phase tap coefficient for nth tap.
C_s	Interelectrode capacitance of one finger pair.
C_T	Static capacitance of a multi-finger IDT (parallel model).
C_T'	Static capacitance of a multi-finger IDT (series model).
d	distance between successive fingers.
d_T	Distance between transducers.
D	Peak distortion parameter.
E_1	Voltage at one port of a 3-port IDT model.
f	Frequency.
f_c	Carrier frequency.
f_{edge}	Frequency of ripple due to an edge reflection.
f_o	Centre frequency.
f_s	Symbol rate in Hertz.
f_{TTE}	Frequency of triple transit echo.
Δf	Acoustic 4dB bandwidth.
Δf_f	Spacing between samples in the frequency domain.
Δf_n	Matching network 3dB bandwidth.
Δf_z	0 dB bandwidth.
$(\Delta f)_c$	Coherence bandwidth of a multipath channel.
F	Function of the mass of a deposited metal.

LIST OF SYMBOLS (continued)

<u>Symbol</u>	<u>Description</u>
F_f	Total range in the frequency domain.
F_x	Function of the mass of a deposited oxide.
g	Conductance.
$g(t)$	Digital pulse shape in the time domain.
G	Additional loss shunt conductance.
G_a	Acoustic conductance.
G_{SNR}	Gain in tolerance to additive gaussian noise.
h	Height of a metal finger.
$h(t)$	Time domain response.
$H(f)$	Frequency domain response.
$H_c(f)$	Two-ray channel response.
$H(\omega)$	Frequency domain response, radian frequency.
$H'(\omega)$	Adjusted frequency domain response.
$i(t)$	I-axis transmitted waveform
I	Current.
I_k	kth information symbol for the In-phase axis.
I_{ok}	Output of equalizer at the kth sampling instant.
k	Magnitude of a binary signal.
k	wave number.
k^2	piezoelectric coupling constant.
K	SAW network bandwidth constant.
K	Number of taps in the discrete time channel model.
L_{ser}	Series matching inductor value.
L_p	Parallel matching inductor value.
L_{11}	Reflection loss.
L_{31}	Transmission loss.
L	Centre to centre distance between electrodes.
M_r	Number of steps in a reflection canceller.
M	Tap correction parameter.
N	Tap correction parameter.
n	Transformer turns ratio.
N_{sp}	Number of FFT sample points.
N_T	Total number of transducer fingers.
N_{opt}	Number of finger-pairs that maximizes the resonant bandwidth.
P_e	Probability of error.
q	Charge.
$q(t)$	Transmitted Q-axis waveform.

LIST OF SYMBOLS (continued)

<u>Symbol</u>	<u>Description</u>
Q	Quality factor.
Q_a	Acoustic quality factor.
Q'_a	Modified acoustic quality factor.
Q_k	k th information symbol for the Q axis.
Q_n	Network quality factor.
r	Reflection coefficient for a single step.
$r(t)$	Received signal waveform.
R	Vector distance between two points.
R	Additional series loss resistance.
R_a	Acoustic resistance.
R_e	Electrode resistance.
R_L	IDT load resistance.
R_o	Reference load.
R_p	Parallel tuning resistance.
R_{pp}	Peak to peak amplitude ripple.
R_s	Source resistance.
R_{ser}	Series tuning resistance.
R_T	Total IDT resistance.
$s(t)$	Signal waveform
SNR	Signal to noise ratio.
t	Time.
Δt_t	Spacing between samples in the time domain.
$(\Delta t)_c$	Coherence time of a multipath channel.
T	Symbol period.
T_m	Time spread of a multipath channel.
T_t	Total range in the time domain.
TTS	Triple transit suppression.
u_i	Noise signal from equalizer.
v	Velocity.
v_b	Bulk wave velocity.
v_o	Free surface SAW velocity.
v_r	Average surface wave velocity in a 50% metallization region.
Δv	Velocity difference between free and metallized surface.
V	Voltage.
$w(t)$	Response of a filter to a noise input.
w_k	Gaussian noise component for the k th symbol.
$x(t)$	Filter response to a pulse $g(t)$.
x_k	Filter response to the k th symbol.

LIST OF SYMBOLS (continued)

<u>Symbol</u>	<u>Description</u>
X_a	Acoustic reactance.
$y(t)$	Baseband filter output response.
$Y(z)$	Linear channel sampled model.
y_k	Time sampled coefficients of vector $Y(z)$.
Y	Admittance matrix.
Y_a	Acoustic admittance.
Z	Impedance.
z_k	Equalizer time sampled output for k th sample.
α	Filter roll-off parameter.
$\alpha_n(t)$	Attenuation factor.
β_i	Pole/zero parameter.
γ	EM feedthrough attenuation.
δ	Dirac delta function.
δ_{EM}	Ripple magnitude caused by EM feedthrough.
δ_s	Doppler shift.
δ_r	Small radian frequency increment.
ϵ_0/ϵ_p	Inverse relative dielectric constant.
λ	Wavelength in a piezoelectric substrate.
λ_c	Cut-off wavelength in a TE_{01} waveguide.
λ_0	Wavelength in a piezoelectric substrate at ω_0 .
η	Metallization ratio.
η^2	Equalizer noise variance.
σ^2	Noise variance of w_k .
θ	Radian angle.
$\theta_n(t)$	Delayed path phase.
θ	Phase difference.
θ_{nf}	Notch frequency parameter
ϕ	SAW potential
ρ	Substrate density
τ	Transmission delay
$\tau_n(t)$	Propagation delay for the n th path
τ_D	SAW travel time between IDT centres
τ_{EDGE}	SAW travel time between IDT centre and substrate edge
ω	Frequency, rads/sec.
ω_0	Radian centre frequency

CHAPTER 1

INTRODUCTION

1.1 Introduction to the Thesis Topic

The demand for large capacity digital radios in a nationwide network has accelerated the development of all related technologies. In order to satisfy the channel capacity requirements, multilevel quadrature amplitude modulation (QAM) systems have evolved. An efficiency of 7 b/s/Hz can be achieved using a 256 QAM system in conjunction with various techniques such as diversity and fading countermeasures in order to achieve an acceptable bit error rate (BER). In addition to space diversity techniques, correction of distortions introduced by selective fading is achieved by the use of an adaptive time domain equalizer (ATDE). Unfortunately, the complexity of QAM systems increases with the number of levels and hence there is a strong incentive to reduce system complexity and cost at every opportunity.

Adaptive time domain equalizers for high spectral efficiency digital radios may utilize baseband delay sections and multipliers in order to meet the stringent requirements on linearity and delay time. Unfortunately, this approach requires duplication of most of the baseband hardware for the in-phase (I) and quadrature (Q) channels, and, in multi-tap ATDEs can consume large areas of the available board space [Figure 1.1]. This problem can be solved if the ATDE can be implemented at an intermediate frequency (IF) where duplication of the delay sections and multipliers is not required [Figure 1.2]. This approach has conventionally been implemented with multiple lumped element delay sections which are

Figure 1.1(a) Digital radio receiver design with a baseband ATDE.

Figure 1.1(b) Conventional baseband implementation of ATDE requires duplication of much of the circuitry.

Figure 1.2(a) Digital radio receiver design with an IF ATDE.

Figure 1.2(b) Utilizing an IF ATDE reduces parts count and circuit complexity.

bulky, expensive, and take much time to be tuned by skilled personnel. A cost effective alternative, especially for large production runs, is a surface acoustic wave (SAW) delay line filter which is smaller than the IF lumped element equivalent, has fewer parts, and requires no tuning.

Until recently, inclusion of SAW devices into high spectral efficiency radios was not practical because of poor performance, and when integration into radio systems became possible they were generally restricted to functions such as roofing and band-reject filters. With the advent of high speed computers and special analysis software packages, it became possible to transform into the time domain a corrupted frequency response using Fast Fourier Transform (FFT) algorithms. This enabled the designer to pinpoint the exact time at which reflections and spurious excitations occurred, and also to determine the corresponding amplitude. A systematic approach to identifying and correcting these spurious signals has been documented in [10], [17], and [43] and has resulted in SAW filters which are of high enough quality to be used as pulse shaping filters in the IF signal path. Having achieved a solid understanding of the problems and solutions associated with designing and manufacturing such filters, the next logical step was to investigate the applications of such devices in areas other than pulse shaping, namely a tapped delay line for use in a SAW based ATDE. Figure 1.3 illustrates such a device.

A SAW filter is far from an ideal filter because there exists a number of second-order effects which can degrade the response. It is therefore necessary to compensate for these in the design of a high performance filter.

In Section 1.3 the thesis objective is defined and the sequence of work is outlined.

Figure 1.3 Photographs of a 70 MHz SAW filter delay line with seven tap outputs

1.2 Alternative Equalization Concepts

A number of papers describe the use of SAW devices in programmable transversal filters [16,28], however, these are generally limited to military applications of adaptive interference control and jamming countermeasures. Fujitsu [13] has developed a five-tap baseband transversal equalizer for their 64-QAM system and is currently developing an IF adaptive equalizer [Figure 1.4] which basically eliminates transmission path distortion using a resonator with variable f_0 and Q . The resonator is adjusted to have the reverse characteristics of the channel. Unfortunately, this type of IF equalizer can only correct for single fade notches and cannot equalize non-minimum phase distortion.

Linear and decision feedback (DFB) equalizers are two common forms of equalizers. The linear equalizer is discussed in detail in Chapter 2. The DFB equalizer consists of both a feedforward section and a feedback section. The input to the feedforward section is the received signal. The feedforward section is in this respect similar to the linear equalizer. The feedback system is used to remove that part of the interference from the present estimate caused by the previously detected symbols [8]. In terms of the loss in the SNR of the equalized signal, the DFB equalizer is superior to the linear equalizer. For the purposes of this thesis, only the linear equalizer is discussed.

Figure 1.4 Block diagram of a Fujitsu IF adaptive equalizer. The use of a variable Q resonator at the notch frequency equalizes the distorted input spectrum.

1.3 Thesis Objective

To date there has been no major effort in the development of SAW based equalizers because of the problems encountered in producing a high quality SAW device with a response free from second-order distortions. As mentioned in Section 1.1, it has only recently become possible to understand and compensate for second-order effects, allowing the radio designer to use this technology.

It is the objective of this thesis to demonstrate the feasibility of a 70 MHz IF transversal filter for adaptive time domain equalization utilizing a tapped SAW delay line as the delay element. The equalization is performed open loop by manually adjusting the multiplier coefficients. The design of the SAW delay line output taps is based on the results of an analysis which computes the frequency response degradation due to acoustic multi-reflections for various output tap finger topographies. The objective of the thesis also includes the development of new electronics to operate at an IF of 70 MHz as opposed to the conventional baseband approach. Since this work is exploratory in nature, any problems encountered in the process are to be analysed to determine the cause, and where possible, corrected.

A block diagram showing the hardware implementation of the SAW based ATDE is shown in Figure 1.5. Figure 1.6 outlines the work that has been done to meet the thesis objective. A distorted spectrum and an equalized spectrum obtained from the final circuit are illustrated in Figure 1.7.

Figure 1.6 Flowchart of the work presented in this thesis.

REF LEVEL	/DIV	MARKER 70 150 000.000Hz
-40.000dB	1.000dB	MAG(S21) -39.981dB
-34.250dB	1.000dB	MARKER 70 150 000.000Hz
		MAG(D3) -38.802dB

Figure 1.7 Spectrum showing a typical fade response and the response which has been corrected by the equalizer.

1.4 Thesis Outline

Chapter 2 outlines the theory behind adaptive equalization from the basic principles of matched filtering to the mathematics behind the zero-forcing algorithm.

Chapter 3 describes the basic operation of surface acoustic wave devices and presents terminology specifically related to SAW devices. The requirements for surface wave propagation are stated and a number of important second-order effects are introduced. An analysis of acoustic multi-reflections is given. Methods taken to compensate second-order effects are presented.

Chapter 4 describes the procedure used to design two different SAW delay lines. The design precautions used in the topographical layout of the output taps are presented. The design of the SAW delay line includes no compensation for the matching electronics. As a result the criteria for a low Q matching network are documented. Special packaging precautions are taken to reduce electromagnetic feedthrough.

Chapter 5 describes the design of the IF electronics developed for this thesis. The first section describes the SAW input and output buffer amplifier designs and associated printed circuit board layout considerations. The second section presents the designs of the remaining transversal filter elements, ie., the multipliers, summing amplifiers and quadrature hybrid. A special section on the ATDE alignment is included.

Chapter 6 presents the experimental results of the SAW delay line performance, the effect of the special techniques taken to implement the

ATDE, and the overall circuit performance. A description of the experimental test set-up is also included.

Finally, in Chapter 7, conclusions are drawn from the experimental results. Recommendations for future work are presented based on the experimental results and computer modelling.

Reflection modelling computer programs, SUPER-COMPACT amplifier models, and an overall circuit schematic are included as appendices.

CHAPTER 2

ADAPTIVE FILTER THEORY

2.1 Introduction

The classical approach to digital signal processing in optimal receivers involves the use of matched filter detection [3, 4] with filter characteristics based on expected or known channel distortions. Figure 2.1 illustrates such a receiver.

Figure 2.1 Optimal receiver with matched baseband filter detection.

Using this method in the decision process intrinsically assumes the use of an ideal channel with bandwidth B . As each pulse passes through the matched low-pass filter, the corresponding output $h(t)$ has a $\sin x/x$ shape (2.1), appropriately scaled and delayed by the pulse amplitude and position [Figure 2.2].

$$h(t) = \frac{\sin \pi t/T}{\pi t/T} \quad (2.1)$$

With pulses spaced at time intervals $T=1/(2B)$ seconds, the received waveform, when observed at the correct sampling instants, results in zero interference between adjacent pulses. In fact, any filter having frequency characteristics with odd symmetry about $f = 1/2T$ and $|H(f)|=0.5$ will result in no interference for symbol rates f_s equal to $2B$. The brickwall filter is just a special case of this.

Figure 2.2 Ideal brickwall filter results in zero intersymbol interference at the sampling instants (decision times). (a) Frequency response. (b) Time domain response. (c) Sampled input signal. (d) Corresponding pulse waveform output.

The zero interference condition between adjacent pulses requires exact synchronization between the transmitter and receiver and also requires that the transmission channel behave like an ideal low-pass filter. There are two basic problems associated with this. Firstly, the tails of $h(t)$ decay as $1/t$ and hence a mistiming error results in an infinite series of interference components at the decision times. Such a series is not absolutely summable and hence the sum of the resulting interference does not converge. Secondly, such a filter shape is not physically realizable. The use of a physically realizable filter causes each received pulse to be affected somewhat by the presence of adjacent pulses [19] as illustrated in Figure 2.3. This overlap between pulses gives rise to distortion, which is generally called intersymbol interference (ISI) [3,4,5].

Figure 2.3 Realizable filter function gives rise to ISI

If a simple band-limiting function is used, for example a single pole LPF, the ringing that results from a rectangular pulse excitation overlaps several symbol periods. The ringing of adjacent pulses can add or subtract to the desired pulse, increasing the likelihood of detecting an incorrect amplitude, and possibly causing an error in the demodulated data. To

minimize this problem, Nyquist filter shapes have been established [4] to minimize ISI by forcing the ringing to pass through zero at the exact moment the peak of the received pulse is sampled. A pulse shape that has found wide use in digital transmission on band-limited channels is one that has a raised-cosine spectral characteristic and pulse response $h(t)$ [3] as shown in Figure 2.4.

$$h(t) = \frac{\sin \pi t/T}{\pi t/T} \frac{\cos \alpha \pi t/T}{1-4\alpha^2 t^2/T^2} \quad (2.2)$$

where α is called the roll-off parameter and T is the symbol rate. The tails of this pulse decay as $1/t^3$, hence a mistiming error leads to a series of intersymbol interference components that converge. Use of SAW technology now allows these filters to be implemented at intermediate frequencies [17,19] as illustrated in Figure 2.5 .

In any real system, ISI is caused by both static and dynamic distortions. Static distortion results from non-ideal matched filtering and thermal noise. Dynamic distortion on the other hand results from frequency selective and multipath fading [7,18]. Since the designer has no foreknowledge of the exact channel distortion in a dynamic system, there is no way of exactly producing the optimal filter characteristics to counter it. The only recourse the designer has to mitigate the effects of dynamic distortion due to multipath fading is to include within the receiver an adjustable filter that can be trimmed to closely fit the required filter characteristics to compensate for any distortion scenario. Since the purpose of this adjustable filter is to compensate for imperfect channel transmission characteristics, these filters are called equalizers [3,5,8]. Since cost is proportional to the number of adjustable elements in an equalizer, the available parameters must be carefully divided between matched

(a)

(b)

Figure 2.4 Nyquist pulse shaping filter having a raised cosine spectrum. (a) Frequency response. (b) Pulse response causes ringing to pass through the zero point at the sampling instants.

(a)

(b)

Figure 2.5 SAW IF pulse shaping filters reduce the number of precise shaping structures by a factor of two. (a) Conventional baseband approach requires two pulse shaping filters. (b) Receiver design where pulse shaping is performed at IF requires one pulse shaping filter.

filtering (used to minimize the effects of additive noise) and equalization (used to minimize ISI due to fading). It may be possible to combine both of these requirements in an IF ATDE using a tapped SAW delay line with a Nyquist shaping response. This will reduce the number of components required for receiver operation even further since a separate pulse shaping filter prior to the equalizer would no longer be necessary.

Figure 2.6 The linear transversal filter is a popular equalizer design which employs a feedback algorithm to adjust the tap coefficients.

Because of its versatility and relative ease of implementation, the linear transversal filter is a popular equalizer design [7]. It lends itself to compensation of channels for which the time dispersion of pulses occurs over a few symbol periods. The linear transversal equalizer consists of a tapped delay line, as shown in Figure 2.6. The delay line is tapped at intervals of the symbol period T of the input pulse stream. The output of each tap is weighted by a variable gain factor and the weighted outputs are summed before sampling to produce the digitized output. A sophisticated

feedback system employing an equalizer algorithm is implemented to adaptively adjust the tap parameters for dynamic input signals. This is illustrated in Figure 2.7. An error signal is generated by the comparison of the received data and a known sequence which is sent during system start up.

Figure 2.7 The adaptive linear equalizer employs a sophisticated feedback network which generates an error signal to correct the distortion. The error signal controls the multipliers. The initial error parameters are established during system startup.

Significant research has been performed [5,8] on the criteria for optimizing the filter coefficients. Two criteria have found widespread use in optimizing the equalizer coefficients. One is the peak distortion criteria [3] which is simply the minimization of the worst-case ISI at the output of

the equalizer. The algorithm used to satisfy this criterion is called the zero-forcing algorithm. The second criterion used in optimizing the equalizer coefficients is called the mean-square-error (MSE) criterion. In the MSE criterion, the tap weight coefficients are adjusted to minimize the mean square value of the error between the k th information symbol and the estimate of that symbol at the output of the equalizer [3]. For the purposes of this thesis, only the zero-forcing algorithm will be considered.

The objective of this chapter is to illustrate the basic theory behind adaptive time domain equalization and show how tap coefficients may be calculated. Section 2.2 describes the theory behind signal distortion and ISI, and shows an equivalent representation of the fading channel by a Z-transform model. Section 2.3 explains the theory of linear equalization using the transversal filter. Sections 2.3.1 through 2.3.4 describe the zero-forcing algorithm, the additional tolerance to additive noise provided by the use of equalizers, the tap corrections required for an IF implementation, and the computer algorithm used to obtain the tap coefficients.

2.2 Signal Distortion and Intersymbol Interference

A digitally modulated signal transmitted over a limited bandwidth channel has the form

$$s(t) = \sum_{n=-\infty}^{\infty} I_n g(t - nT) \quad (2.3)$$

where $\{I_n\}$ represents the information bearing sequence of symbols and $g(t)$ is a pulse which is bandlimited. T is the symbol period. The output of the baseband receiving filter will have the form

$$y(t) = \sum_{n=-\infty}^{\infty} I_n x(t - nT) + w(t) \quad (2.4)$$

where $x(t)$ is a pulse representing the response of the receiving filter and channel to $g(t)$, and $w(t)$ is the response of the receiving filter to noise. Now, if $y(t)$ is sampled at times $t = kT + \tau$, $k=0,1,2$, etc. where τ is the transmission delay through the medium, the corresponding sampled response is

$$y_k = y(kT + \tau) = \sum_{n=-\infty}^{\infty} I_n x(kT + \tau - nT) + w(kT + \tau) \quad (2.5)$$

This expression may be re-written [3] as

$$y_k = x_0 I_k + \sum_{\substack{n=-\infty \\ n \neq k}}^{\infty} I_n x_{k-n} + w_k \quad (2.6)$$

The term I_k represents the desired information symbol at the k th sampling instant, w_k is the additive Gaussian noise variable at the k th sampling

instant and the term

$$\sum_{\substack{n=-\infty \\ n \neq k}}^{\infty} I_n x_{k-n} \quad (2.7)$$

represents the intersymbol interference.

As mentioned in Section 2.1, ISI can be introduced by both non-optimum filtering and channel distortions. This is in a sense misleading since any ISI in theory can be eliminated by a filter optimally matched to the channel distortion. However, in the case of fixed filters and time varying channel characteristics, this is not the case and hence an adaptive filter must be utilized to reduce ISI.

The characterization of signals in fading multipath channels has been well documented in a number of references [3-5,7,8]. However so as to be complete, a brief description of the points relevant to this thesis will be given.

If we consider the transmission of an unmodulated carrier at frequency f_c , then the received signal for the case of discrete multipath signals can be written as

$$\begin{aligned} r(t) &= \sum_n \alpha_n(t) e^{-j2\pi f_c \tau_n(t)} \\ &= \sum_n \alpha_n(t) e^{-j\theta_n(t)} \end{aligned} \quad (2.8)$$

where $\alpha_n(t)$ is the attenuation factor for the signal received on the n th path, $\tau_n(t)$ is the propagation delay for the n th path and $\theta_n(t) = 2\pi f_c \tau_n(t)$. The received signal consists of the sum of a number of time-variant phasors having amplitudes $\alpha_n(t)$ and phases $\theta_n(t)$. Large dynamic changes in the medium are required for $\alpha_n(t)$ to change sufficiently to cause a

significant change in the received signal. On the other hand, $\theta_n(t)$ will change by 2π radians whenever τ_n changes by $1/f_c$. Since $1/f_c$ is a small number, θ_n can change by 2π radians with relatively small motions of the medium. Since the delays $\tau_n(t)$ change in a random manner it means that for a large number of paths, $r(t)$ can be modelled as a complex-valued gaussian random process, in the time variable t [3].

The received signal $r(t)$ will experience signal fading which comes about as a result of the time variations in the phases $\{\theta_n(t)\}$. Hence, the vectors $\{\alpha_n e^{-j\theta_n}\}$ will at some times add destructively and at other times add constructively. A widely accepted model which describes the fading characteristics of multipath channels is the Rayleigh fading channel [8]. In this model, the magnitude of the envelope of the channel response to an applied impulse follows a Rayleigh distribution.

If an extremely short pulse is transmitted over a time-varying multipath channel, the received signal will experience a time delay τ and multipath spread of the channel, denoted by T_m . The reciprocal of the multipath spread is a measure of the coherence bandwidth $(\Delta f)_c$ of the channel.

$$(\Delta f)_c \approx 1/T_m \quad (2.9)$$

Thus two frequencies of separation greater than the coherence bandwidth will be affected differently by the channel. If $(\Delta f)_c$ is small compared to the bandwidth of the signal, the channel is said to be frequency-selective. The time variations in the channel result in an additional effect called the Doppler spread of the channel, denoted B_d . In addition, the spectral line will also experience a Doppler shift of δ_s Hz. The reciprocal of B_d is a

measure of the coherence time $(\Delta t)_c$ of the channel.

$$(\Delta t)_c \approx 1/B_d \quad (2.10)$$

Hence, if two signals arrive at the receiver greater than $(\Delta t)_c$ apart, then they are affected differently by the channel. These characteristics of the multipath channel determine the scattering function of the channel [Figure 2.8] which provides a measure of the average power output of the channel as a function of the time delay τ and doppler shift δ_s .

Figure 2.8 Scattering function of a medium range tropospheric scatter channel experiences both time and frequency spread, and small doppler shifts.

If a single multipath element is considered, the channel response for multipath fading on a line-of-sight path to a single antenna is often represented [52] by

$$H_c(f) = a [1 - b e^{-j2\pi(f-f_c)\tau}] \quad (2.11)$$

The depth, shape, and notch frequency of the fade are determined by the parameters a , b , and f_c respectively. A standard relative delay τ is 6.3 nanoseconds [52]. For convenience, the fade level parameter A , the relative notch depth parameter B , and the notch frequency parameter θ_{nf} are defined as,

$$(A)_{dB} = -20 \log a \quad (2.12)$$

$$(B)_{dB} = -20 \log (1 - b)$$

$$(\theta_{nf})_{rad} = 2\pi f_c \tau$$

These parameters are utilized in the algorithm of Section 2.3.4. Figure 2.9 illustrates the two-ray multipath model of (2.11).

Figure 2.9 Two-ray multipath model (a) Illustration of a single multipath signal. (b) Corresponding frequency response. (c) Transmitted pulse shape (d) Distorted pulse shape at the receiver.

The waveform of a transmitted QAM signal with carrier frequency f_c may be written [52] as

$$S(t) = i(t) \cos \omega_c t - q(t) \sin \omega_c t \quad (2.13)$$

where

$$\omega_c = 2 \pi f_c$$

$$i(t) = \sum_{n=-\infty}^{\infty} I_n g(t - nT)$$

$$q(t) = \sum_{n=-\infty}^{\infty} Q_n g(t - nT)$$

$\{I_n\}$ and $\{Q_n\}$ are mutually uncorrelated in-phase and quadrature-phase data sequences, T is the symbol period and $g(t)$ is the shape of the baseband pulse. Each element I_n and Q_n assumes with equal probability the values of $\pm 1, \pm 3, \dots, \pm(\sqrt{L}-1)$ for $L = 4, 16, 64, \dots$.

For the purposes of this discussion, only the single axis baseband case will be discussed. Since the transmitter sends discrete-time symbols at a rate of $1/T$, and the sampled output of the receiver is also a discrete-time signal with samples occurring at rate $1/T$, it follows that the cascade of the channel, sampler, transmit and receive filters may be represented (for a single axis) by an equivalent discrete-time channel model as shown in Figure 2.10. The sampled impulse response of this channel [8] is

$$Y(z) = y_0 + y_1 z^{-1} + y_2 z^{-2} + \dots + y_k z^{-k} + \eta_k \quad (2.14)$$

where $\{y_k\}$ is a set of tap coefficients of an equivalent discrete time

transversal filter and (η_k) is a gaussian noise sequence.

Figure 2.10 Equivalent discrete-time model of an intersymbol interference channel with additive noise.

If for the moment the noise component is ignored, equation (2.14) may be written as

$$\begin{aligned} Y(z) &= y_0 + y_1 z^{-1} + y_2 z^{-2} + \dots + y_k z^{-k} \\ &= y_0 (1 - \beta_1 z^{-1}) (1 - \beta_2 z^{-1}) \dots (1 - \beta_k z^{-1}) \end{aligned} \quad (2.15)$$

where $\{\beta_k\}$ represents the zero locations of the channel response. When all the zeros of this equation lie inside a unit circle, the channel is said to exhibit minimum phase [8]. Equation (2.11) exhibits a minimum phase state. The non-minimum phase state is obtained by replacing τ in (2.11) with $-\tau$. In the two-ray fade model of Figure 2.9, a non-minimum phase

state is analagous to having the delayed signal arrive at the receiver with an amplitude greater than the non-delayed signal.

The sampled impulse response of the channel can be represented by the vector $Y = [y_0 \ y_1 \ y_2 \ \dots \ y_k]$. The sampled impulse response of the corrupted channel is required in order to calculate the tap coefficients using the zero-forcing algorithm.

2.3 Linear Equalization Using the Transversal Filter

As mentioned in section 2.1, the linear feedforward transversal filter is a widely used equalizer design. It lends itself to the compensation of channels for which the time dispersion of pulse signals is only over a small number of symbol intervals [7]. In general, the equalizer can only correct for distortion which results from interfering signals occuring within the time span of the delay line. Hence for the $2N+1$ tap equalizer of Figure 2.6, interfering signals arriving at times beyond $\pm NT$ cannot be equalized. This is analagous to saying that for a $2N+1$ tap equalizer, the maximum number of ripples that the equalizer is capable of handling in the band $f = f_0 \pm 1/2T$ is less than or equal to N .

2.3.1 Zero-Forcing Algorithm

Figure 2.6 illustrates an equalizer with $2N+1$ taps, having corresponding tap weights $c_N, \dots, c_0, \dots, c_N$. The output z_k of the equalizer to a vector input $Y = [y_0 \ y_1 \ y_2 \ \dots \ y_k]$ and tap weights c_n can be expressed as

$$z_k = \sum_{n=-N}^N c_n y_{k-n} \quad (2.16)$$

$\{z_k\}$ is simply the convolution of the tap coefficients $\{c_n\}$ and the channel sampled impulse response $\{y_k\}$. The output of the equalizer at the k th sampling instant can be expressed in the form

$$I_{ok} = z_0 I_k + \sum_{n \neq k} I_n z_{k-n} \quad (2.17)$$

where the first term represents a scaled version of the desired symbol and the second term represents the intersymbol interference. The peak value of this interference is called the peak distortion [3] and can be defined as

$$D = \sum_{\substack{n=-\infty \\ n \neq 0}}^{\infty} |z_n| \quad (2.18)$$

Hence D is a function of the equalizer taps. For an equalizer having an infinite number of taps, it is possible to select tap weights such that $D = 0$ for all n except $n = 0$.

By taking the z transform of (2.16), the following is obtained

$$Z(z) = C(z)Y(z) = 1 \quad (2.19)$$

i.e.,

$$C(z) = 1/Y(z) . \quad (2.20)$$

hence, the equalizer with transfer function $C(z)$ is simply the inverse filter to the linear channel model $Y(z)$. In other words, complete elimination of the ISI requires the use of an equalizer with a response the inverse to $Y(z)$. This type of filter is called a zero-forcing filter.

Equation 2.16 yields $2N+1$ independent equations in terms of the tap coefficients c_n and limits us to $2N+1$ constraints. Hence we can impose the condition

$$z_k = \begin{cases} 1 & k = 0 \\ 0 & k = \pm 1, \pm 2, \dots, \pm N \end{cases} \quad (2.21)$$

Thus the equalizer can force the output to go to zero at N sample points to either side of the desired peak output. There is no guarantee that the output will be zero at intervals beyond $\pm N$, where $2N+1$ is the number of taps.

Combining (2.16) and (2.21) the $2N+1$ equations can be solved for the tap coefficients c_n as follows

$$\begin{bmatrix} y_0 & y_1 & \dots & y_N & \dots & y_{2N+1} & y_{2N} \\ y_1 & y_0 & \dots & y_{N+1} & \dots & y_{2N} & y_{2N+1} \\ \cdot & \cdot & & \cdot & & \cdot & \cdot \\ y_N & y_{N-1} & \dots & y_0 & \dots & y_{N+1} & y_N \\ \cdot & \cdot & & \cdot & & \cdot & \cdot \\ y_{2N-1} & y_{2N-2} & \dots & y_{N-1} & \dots & y_0 & y_{-1} \\ y_{2N} & y_{2N-1} & \dots & y_N & \dots & y_1 & y_0 \end{bmatrix} \begin{bmatrix} c_{-N} \\ c_{-N+1} \\ \cdot \\ c_0 \\ \cdot \\ c_{N-1} \\ c_N \end{bmatrix} = \begin{bmatrix} z_{-N} \\ z_{-N+1} \\ \cdot \\ z_0 \\ \cdot \\ z_{N-1} \\ z_N \end{bmatrix} \quad (2.22)$$

Although there are $2N+1$ adjustable parameters, there are $2N+K$ values in

$\{z_n\}$, the equalized response, where K represents the number of taps in the discrete time channel model of Figure 2.10. From (2.19), $z_0=1$ and $z_n=0$ for $n= \pm 1, \dots, \pm N$. However, for $n= \pm(N+1), \dots, \pm(N+K)$, z_n may not necessarily be zero. Hence, there is always some residual interference when the optimum coefficients are used. There also exists a noise penalty because the equalizer, in eliminating the ISI, enhances the additive noise. For example, if the channel contains a spectral null in its frequency response, the linear zero-forcing equalizer attempts to compensate for this by introducing an infinite gain at that frequency. This compensates the channel distortion at the expense of enhancing the additive noise [8].

2.3.2 Tolerance to Additive Noise

As mentioned in Section 2.3.1, for deep fading nulls there is a noise enhancement penalty for the linear feedforward equalizer. However, for more moderate fades and a given error rate in the received data, an equalizer can provide a reduction in the required signal to noise ratio. For example, consider the feedforward equalizer of Figure 2.11.

Figure 2.11 Noise signals on an $(m+1)$ tap linear feedforward transversal filter at the time instant $t = iT$.

The output noise signal u_i from the equalizer at time $t = iT$ is

$$u_i = \sum_{j=0}^m w_{i-j} c_j \quad (2.23)$$

The noise components $\{w_i\}$ are statistically independent Gaussian random variables with zero mean and variance σ^2 . Hence, u_i is also a Gaussian random variable with zero mean and variance η^2 defined by

$$\eta^2 = \sum_{j=0}^m \sigma^2 c_j^2 \quad (2.24)$$

The probability of an error occurring when a binary signal is sent over a time-invariant channel with additive white Gaussian noise can be shown [3,4,7] to be

$$P_e = Q\left(\frac{k}{\eta}\right) \quad (2.25)$$

where $Q(v) = \frac{1}{\sqrt{2\pi}} \int_v^{\infty} e^{-u^2/2} du$

is the complementary error function and k is the magnitude of the transmitted signal. The i th signal is equally likely to have either value $\pm k$.

The tolerance of any system to additive gaussian noise is taken to be the SNR required for some given error rate at the input to the receiver. This can be written as

$$\text{SNR}_{\text{dB}} = 20 \log (k/\sigma) \quad (2.26)$$

Normally for large values of SNR, the error rate is quite low, say 10^{-5} or 10^{-6} . The smaller the SNR for a particular error rate, the greater the tolerance of the system to additive noise. Thus the advantage in tolerance to noise of a given system over another is the reduction in the required SNR for a particular error rate.

For small tap weights, where the summation of the coefficients c_j^2 is less than 1 and $\eta^2 < \sigma^2$, the probability of error will be less than that predicted by (2.25). A more severe channel distortion will require a

larger set of tap coefficients in order to equalize the signal. In this case, $\eta^2 > \sigma^2$ and the probability of error will be slightly larger than the error probability of (2.25). This represents the noise enhancement penalty.

It should be noted however, that the error rate of the unequalized signal will be much worse than that of (2.25), which is the error rate for an undistorted, time-invariant channel. Hence the equalizer can be said to improve the tolerance of the receiver to noise. Even for equalizers of only a few taps, the improvement in the tolerance to additive noise can be several dB over the arrangement with no equalizer. This is a considerable improvement and is quite typical of the advantage that may be gained with a linear equalizer.

2.3.3 Tap Corrections for IF Implementation

The preceding sections calculate the tap coefficients from an algorithm based on a signal at baseband. The adaptive equalization in this thesis is being performed at IF, and as such, an additional correction in the taps must be made as follows.

Figure 2.12 shows the signals generated by the in-phase delay line of an IF equalizer. The k -th output of the demodulating mixer for the m -tap in-phase delay line has the form

$$I_{ok} = \cos \omega_c t \sum_{n=0}^m I_k(t-nT) c_{In} \cos [\omega(t-nT)] \quad (2.27)$$

where ω_c is the frequency used for the down conversion and T is the delay, equal to the symbol period, c_{In} are the tap weights for the in-phase components, and I_k is the in-phase baseband data signal.

Figure 2.12 Demodulation of the IF tap signal introduces an additional phase related magnitude term in the baseband signal.

The high frequency components are removed by the baseband filter. Assuming for the moment that $\omega = \omega_c$, the output signal is

$$I_{ok(\text{baseband})} = \sum_{n=0}^m I_k(t-nT) c_{In} \cos [n \omega_c T] \quad (2.28)$$

It can be seen that after the demodulation process, the desired baseband symbol $I_k(t-nT) c_{In}$ will be multiplied by an additional phase term of $\cos [n \omega_c T]$. Similarly, for the quadrature-phase delay line, the output of the demodulating mixer can be expressed as

$$Q_{ok(\text{baseband})} = \sum_{n=0}^m Q_k(t-nT) c_{Qn} \cos [n \omega_c T] \quad (2.29)$$

If the IF tap settings are used to implement the equalization at IF, then the baseband response will have slope in the frequency response. This happens when the bandpass centre frequency f_0 is not at an integer multiple of the symbol rate [Figure 2.13]. The slope that results can be represented as a cross coupling between the in-phase and quadrature components.

If a bandpass filter exhibits symmetry about some frequency $f = n/T$, where n can be any integer value, then there is no cross coupling. Transmission with symmetry about this frequency is required in order to avoid amplitude-to-phase modulation conversion on a bandpass signal [54]. It follows that if this is not the case, then some sort of a transformation network is required in order to implement a frequency shift in the tap response so that the peaks and notches occur at f_0 . Such a network is

shown in Figure 2.14 [54].

Figure 2.13 A bandpass signal not centred on the tap notch frequency will introduce cross coupling distortion, resulting in a finite attenuation and amplitude slope.

Figure 2.14 The network shown above generates tap weights which result in no cross coupling at IF.

The computation of the coefficients M and N can be obtained if we consider the single complex tap illustrated in Figure 2.15. The frequency response $H(\omega)$ of such a tap is,

$$H(\omega) = 1 + a e^{-j\omega\tau} + jb e^{-j\omega\tau} \quad (2.30)$$

where a and b are the in-phase and quadrature tap weightings and τ is the tap delay time, equal to an integer number of symbol periods T .

$$\tau = nT \quad n = 0, 1, 2, \text{ etc.} \quad (2.31)$$

The frequency response of (2.30) for $b = 0$ and $\tau = \pm T, \pm 2T$ and $\pm 3T$ is shown in Figure 2.16.

$$\tau = nT \text{ for } n = \pm 1, \pm 2, \pm 3; T = 44.3 \text{ ns}$$

Figure 2.15 The complex tap contains both in-phase and quadrature delay sections.

Figure 2.16 Frequency response of a complex tap for three values of delay τ .

The response $H'(\omega)$ of the tap which has been transformed by a frequency shift $\Delta\omega$ is,

$$\begin{aligned}
 H'(\omega) &= 1 + a e^{-j(\omega+\Delta\omega)\tau} + jb e^{-j(\omega+\Delta\omega)\tau} \\
 &= 1 + [a \cos(\Delta\omega\tau) + b \sin(\Delta\omega\tau)] e^{-j\omega\tau} \\
 &\quad + j [b \cos(\Delta\omega\tau) - a \sin(\Delta\omega\tau)] e^{-j\omega\tau}
 \end{aligned} \tag{2.32}$$

Equation (2.32) illustrates that for a complex tap at IF, a change in the tap weight "a" will cause a quadrature tap weight change.

In order to implement the transformation to prevent this cross coupling, the tap weights a and b can be replaced by a set of new tap weights a' and b' which are calculated using (2.33).

$$\begin{aligned}
 a' &= a M - b N \\
 b' &= b N + a M
 \end{aligned} \tag{2.33}$$

where $M = \cos(\Delta\omega\tau)$ and $N = \sin(\Delta\omega\tau)$.

Substitution of (2.33) into (2.32) results in a frequency response which has no cross coupling

$$H'(\omega) = 1 + a e^{-j\omega\tau} + jb e^{-j\omega\tau} \quad (2.34)$$

The algorithm used to compute the new tap weights to implement the required frequency translation is illustrated in Figure 2.17. This tap correction is used in the computer algorithm described in the next section.

$$\Delta\omega = 2\pi \times 2.245 \times 10^6$$

$$\tau = nT \text{ for } n = \pm 1, \pm 2, \pm 3; T = 44.3 \text{ ns}$$

Figure 2.17 Algorithm used to compute the IF tap weights from baseband tap weights. This algorithm prevents cross coupling between the I and Q axis.

2.3.4 Computer Algorithm

The flow chart of the computer algorithm used to compute the tap coefficients is shown in Figure 2.18.

A time domain response of an undistorted RD-6 digital radio pulse is retrieved from storage. This response is converted via an inverse DFT to the frequency domain.

Next a two ray multipath fade is set up from the values of A , B , and θ_{nf} as per (2.11). The frequency response of this fade is cascaded with the frequency response of the RD-6 pulse to obtain the frequency response of the corrupted pulse.

Subject to the designer, a frequency domain slope equalization may be applied to the corrupted response. In general however, this is ignored since slope equalization may be accomplished with an IF slope equalizer. If it is applied, the calculated tap coefficients are calculated based on the assumption that there is no frequency slope in the passband.

Next, the frequency response of the corrupted pulse is transformed back into the time domain via an IDFT. From this, the sampled impulse response $\{y_k\}$ of the received signal is obtained.

The zero-forcing algorithm of (2.22) is next applied via a matrix subroutine which calculates $2N$ tap coefficients based on $2N+1$ available equations. The I and Q baseband tap coefficients are then calculated.

Finally, the baseband taps are adjusted to correct for the phase due to the IF implementation as per (2.34). A branch decision at the end of the program offers four alternatives; (1) redo the time domain calculation (2) print the equalized pulse response values (3) re-try with new multipath distortion characteristics (4) exit.

Figure 2.18 Flowchart of the computer program used to compute tap coefficients based on the zero-forcing algorithm.

CHAPTER 3

THEORY OF SURFACE ACOUSTIC WAVE DEVICES

3.1 Principles of Operation

Surface acoustic waves are generated by an interdigital transducer (IDT) deposited on the surface of a piezoelectric substrate [Figure 3.1]. In a piezoelectric material, an electric polarization caused by a voltage difference between two IDT fingers will produce a mechanical deformation proportional to the applied potential. The mechanical deformation launches a surface wave which will propagate in both the forward and reverse directions. A complete SAW filter consists of one transmitting IDT and one receiving IDT. The transmitting IDT converts electrical energy to acoustic energy and the receiving IDT converts acoustic energy to electric energy.

Figure 3.1 Prototype of a SAW IDT on a piezoelectric substrate.

When the transducer is matched at the electrical port [Figure 3.2], half the power is transmitted towards the output transducer, and half of it is transmitted toward the acoustic load (usually absorber material). By reciprocity, the same applies to the output transducer, and hence there exists a minimum built-in loss of 6 dB for the complete filter (3dB per transducer).

Figure 3.2 Signal transmission in a filter with bi-directional IDTs

The concepts relating substrate material constants and the maximum bandwidth versus minimum insertion loss were originally investigated by R.Smith [14] in 1969. Since reference [14] was published, it has been verified experimentally [24] that the crossed-field model [Figure 3.3] is the best fit for LiNbO_3 substrates. The simplified equivalent electrical circuit consists of static capacitance C_T and an acoustic admittance $Y_a(\omega) = G_a(\omega) + jB_a(\omega)$. The real component of the acoustic admittance, $G_a(\omega)$, is the radiation conductance and it represents the electrical to mechanical energy transformation. The imaginary component $B_a(\omega)$ is the reactive component of the acoustic admittance. Both $G_a(\omega)$ and $B_a(\omega)$ are

frequency dependent terms. The static capacitance C_T is frequency independent. C_T depends on the number of fingers, the finger overlap, and the metallization ratio. The series equivalent circuit [Figure 3.3(c)] consists of a static capacitance C_T' and an acoustic impedance $Z_a(\omega) = R_a(\omega) + jX_a(\omega)$ where $R_a(\omega)$ and $X_a(\omega)$ are the radiation resistance and reactance respectively. R_a , X_a , and C_T' represent the series counterparts of G_a , X_a , and C_T , and are calculated via a parallel to series circuit transformation. For transducers with $G_a \ll \omega_0 C_T$, the transducer capacitance $C_T \approx C_T'$.

To a very good approximation it may be shown [14] that

$$G_a(\omega) = (G_a)_{\max} \left[\frac{\sin x}{x} \right]^2 \quad (3.1)$$

$$B_a(\omega) = (G_a)_{\max} \left[\frac{\sin 2x - 2x}{2x^2} \right] \quad (3.2)$$

where

$$x = N\pi (f/f_0 - 1)$$

$$(G_a)_{\max} = 8 k^2 f_0 N C_T \quad (3.3)$$

$k^2 = 5.4\%$ for 128° rotated LiNbO_3 ,

k^2 is the electromechanical coupling constant,

f_0 is the centre frequency,

N is the number of finger pairs of the IDT,

C_T is the total static interelectrode capacitance,

$C_T = N C_s$ where C_s is the interelectrode capacitance of one finger pair.

Figure 3.3 Crossed-field circuit model.(a) Transducer (b) Parallel equivalent circuit (c) Series equivalent circuit.

It can be shown [22] that when C_T is tuned with a matching inductor, there exists an optimum number of finger pairs that maximizes the resonant bandwidth of the transducer, given by

$$N_{opt} = [\pi/4 k^2]^{1/2} \quad (3.4)$$

The acoustic 4 dB bandwidth of an IDT is inversely proportional to the number of fingers [22].

$$\Delta f \approx f_0/N \quad (3.5)$$

Thus it can be seen that for wide bandwidths, k^2 should be as large as possible. Table 3.1 [1] shows some properties of commonly used substrates. It can be seen from Table 3.1 that 128°LiNbO_3 is the choice of material for wide bandwidth filters since it has a large value for k^2 . For better temperature stability, quartz is the preferred choice.

Table 3.1
Properties of some common piezoelectric substrates .

Material	Orientation		Velocity free surf. m/sec	k^2 %	ϵ_p/ϵ_0	Temp. Coef. of Delay ppm/°C	Atten.@ 1GHz dB/ μ sec
	Cut	Prop.					
LiNbO ₃	Y	Z	3488	4.5	50	94	1.07
LiNbO ₃	128° rot.		3992	5.4	-	75	-
Quartz	Y	Z	3159	0.23	4.52	-24	2.6
Quartz	ST	X	3158	0.25	4.55	0	3.1
LiTaO ₃	Y	Z	3230	0.66	47.9	35	1.14
LiTaO ₃	167° rot.		3394	1.5	47.9	64	-

As a SAW propagates on the surface of a piezoelectric substrate, successive wave crests may be tapped or generated by an IDT. The frequency at which maximum energy transfer occurs is called the synchronous frequency f_0 [2]. This is the frequency at which the period of two successive polarization reversals on the bus bars correspond exactly to the time taken for the crest of the SAW to travel to the next finger pair. Under these conditions, the generated SAWs add constructively for frequencies

$$f = (2n-1) v_r/2d \quad n = 1,2,3 \text{ etc.} \quad (3.6)$$

where v_r =average surface wave velocity on a metallized substrate
 d =distance between successive fingers.

As shown in Figure 3.4, the IDT may be modelled as a set of infinitely narrow sources located midway between two fingers [9]. The amplitude A assigned to each source is proportional to the overlap length of the IDT fingers, with positive or negative sign depending on the direction of the electric field [6].

Figure 3.4 The IDT can be modelled as a set of discrete sources with amplitudes proportional to the overlap length.

The impulse response of such an IDT will be a sum of delta functions

$$h(t) = \sum_n (-1)^n A_n \delta(t - t_n) \quad (3.7)$$

The frequency response is obtained by taking the Fourier transform of this expression

$$H(f) = \int_{-\infty}^{\infty} h(t) e^{-j 2\pi f t} dt = \sum_n (-1)^n A_n e^{-j 2\pi f t_n} \quad (3.8)$$

This represents the discrete fourier transform (DFT) of the input sequence

[21]. A similar conversion from the frequency to the time domain can be performed by an inverse discrete fourier transform (IDFT). A sequence of delta functions represents not only the sine wave at f_0 but also all the odd harmonics at frequencies $(2n+1)f_0$, for $n=0,1,2$, etc. Due to the half wavelength spacing between the fingers, the even harmonics are absent. The overall impulse response of the filter is a convolution [9] of the impulse responses of the input and receiving IDTs. The overall frequency response is obtained by performing a DFT on the resultant impulse response.

The analysis and design of transversal filters requires frequent transformations between time and frequency domains. This is usually done on a digital computer using Fast Fourier Transform (FFT) algorithms that require the information to be sampled in both domains [2]. This is covered in more detail in Section 4.1.

3.2 Second-Order Effects

Conventional surface wave filters have a higher insertion loss than their equivalent lumped circuit counterparts. It is therefore often desirable to minimize the insertion loss as much as possible. A SAW filter is far from ideal and the frequency response of an IDT can be corrupted by a number of unwanted second-order effects which tend to increase in importance when low insertion loss is required. For the design of high performance filters, it is therefore necessary to understand these effects fully and to adopt methods for an efficient reduction or compensation of them in the design process. Figure 3.5 illustrates a general model used to analyze the effects of second-order distortions [26]. A variation of this model is employed in Section 6.1.2 to compute the magnitudes of tap reflections. The most important second order effects are discussed in the following sections and methods for compensation are suggested.

Figure 3.5 General model used to analyze second-order effects in SAW filters.

3.2.1 Internal Reflections

The frequency response of an IDT can be significantly distorted by multiple acoustic reflections of the SAW from the electrodes, which modify its simple tapped delay line character. These reflections are often divided into two classes: mass/electrical loading (MEL) [15] and acoustic regeneration [24] as shown in Figure 3.6. The physics of reflections have been well documented in a number of reports [24,25,27], however, so as to be complete, a brief description will be presented below.

Mass loading is caused by the physical weight of the transducer material on the surface of the substrate. As the SAW propagates underneath this material, it must literally push it up and down. Electrical loading is caused by the metal electrodes on the surface of the piezoelectric crystal shorting out the tangential electric field. Electrical loading changes the elastic constants of the piezoelectric substrate and reduces the velocity of the propagating SAW. Electrical loading also causes delay distortion and other secondary perturbations in the frequency response [40]. The combined effect of mass/electrical loading causes the electrode and gap regions to have different wave impedances [25]. The change in impedance levels is analagous to an abrupt impedance discontinuity in a transmission line which causes reflections. MEL reflections in uniformly spaced transducers can be significantly reduced by the use of multi-finger electrodes which cause reflections to cancel at the center frequency.

Acoustic regeneration occurs when an incident SAW generates a voltage at the receiving IDT terminals, which in turn regenerates SAWs which propagate in both directions. Regenerated waves can be particularly large in bi-directional IDTs under matched conditions, but can be diminished by a intentionally mismatching the IDT with a low impedance. However, it should be noted that the magnitude of the regenerated wave

will be proportionally reduced at the expense of the insertion loss of the transducer. Another method of reducing regenerated SAWs is to employ uni-directional transducers [46] at the expense of a considerably more complex design and manufacturing process. Unidirectional transducers are also limited to narrow-band operation. Acoustic multi-reflections will be considered in more detail in the following sections since they are of importance in the multitap SAW design.

Figure 3.6 Two causes of acoustic multi-reflections in SAW devices (a) Mass/electrical loading (MEL) causes impedance discontinuities. (b) Acoustic regeneration due to a finite load impedance.

3.2.1.1 Regenerated Wave

Smith et al [14] have presented a paper analyzing the immitance, transfer, and scattering characteristics for IDTs on a piezoelectric substrate. The model used in the analysis consists of N periodic sections as shown in Figure 3.7(a). Each individual periodic section may be represented by the Mason equivalent circuit [Figure 3.7(b)] which has an immitance matrix S_n .

Figure 3.7 Mason equivalent circuit. (a) Equivalent N period representation of an IDT. (b) Each periodic section can be represented by the Mason equivalent model having an immitance matrix S_n .

Having formulated the equivalent circuit for one periodic section, the admittance matrix for the entire transducer can then be calculated. The transducer 3-port equation is written as

$$\begin{bmatrix} I_1 \\ I_2 \\ I_3 \end{bmatrix} = [Y] \begin{bmatrix} E_1 \\ E_2 \\ E_3 \end{bmatrix} \quad (3.9)$$

$$\text{Where } [Y] = K \begin{bmatrix} -1/N & 1/N & 4 \\ 1/N & -1/N & -4 \\ 4 & -4 & (-16N + \delta_r(\omega C_T/G_0)) \end{bmatrix} \quad (3.10)$$

for the crossed-field model. K is a constant, and δ_r is a small radian frequency increment. The increment δ_r is required since the matrix elements become infinite at synchronism otherwise.

We know that matching the output port causes a maximum reflection, however, we desire that reflections be at a minimum. Two cases which minimize the magnitude of regenerated SAWs are open circuit loading and a short circuit loading. These cases are commonly referred to as weakly coupled.

Case 1: By leaving port 3 open, we can set $I_3=0$ and hence solve for E_n and I_n in (3.9). The following is obtained.

$$\begin{bmatrix} I_1 \\ I_2 \\ 0 \end{bmatrix} = [Y] \begin{bmatrix} E_1 \\ E_2 \\ E_3 \end{bmatrix} = K \begin{bmatrix} -1/N & 1/N & 4 \\ 1/N & -1/N & -4 \\ 4 & -4 & (-16N + \delta_r(\omega C_T/G_o)) \end{bmatrix} \begin{bmatrix} E_1 \\ E_2 \\ E_3 \end{bmatrix} \quad (3.11)$$

Solving we obtain $E_1=E_2$ and $I_1=-I_2$. This represents total transmission and no reflection.

Case 2: By shorting port 3, we can set $E_3=0$. Setting up (3.9) we obtain

$$\begin{bmatrix} I_1 \\ I_2 \\ I_3 \end{bmatrix} = [Y] \begin{bmatrix} E_1 \\ E_2 \\ E_3 \end{bmatrix} = K \begin{bmatrix} -1/N & 1/N & 4 \\ 1/N & -1/N & -4 \\ 4 & -4 & (-16N + \delta_r(\omega C_T/G_o)) \end{bmatrix} \begin{bmatrix} E_1 \\ E_2 \\ 0 \end{bmatrix} \quad (3.12)$$

Solving we obtain $E_1=E_2$ and $I_1=-I_2$, and again conclude that according to the Mason equivalent model, a short on the output will result in no regenerated SAWs. Hence a weakly coupled transducer effectively

prevents regenerative effects and also allows the incident SAW to pass unperturbed under the IDT with no loss in energy.

Since presenting a zero impedance to the output transducer means that no signal energy is extracted, we must use a small finite load impedance in order to extract energy. However, at the same time we must be able to compute approximately the amount of regenerated signal. Referring to Figure 3.8 [1] we can see that the reflection loss and transmission loss are proportional to the transducer mismatch ratio $b = R_a/R_L$ where R_a is the acoustic radiation resistance of the IDT and R_L denotes the load resistance presented by a series equivalent circuit.

Figure 3.8 Plots of reflection loss L_{11} and transmission loss L_{31} as a function of transducer mismatch ratio b .

The reflection loss L_{11} (the amount the reflected signal is attenuated) is equal to

$$(L_{11})_{dB} = 10 \log [(1+b)^2] \quad (3.13)$$

and the transmission loss L_{31} is equal to

$$(L_{31})_{dB} = 10 \log [(1+b)^2/2b] \quad (3.14)$$

These equations illustrate that in order to achieve adequate reflection suppression and at the same time not have an excessive insertion loss, b should be greater than 20 but less than 600. These values of b correspond to $L_{11}=26$ dB, $L_{31}=10$ dB, and $L_{11}=56$ dB, $L_{31}=25$ dB. Since the measured radiation resistance of a ten finger output transducer is on the order of 20Ω , a load of approximately 1Ω (i.e. $b = 20$) will prevent excessive regeneration of SAWs.

3.2.1.2 MEL Reflections

The effect of MEL reflections is most clearly described if we initially assume that the transducer is very weakly coupled to the substrate so that we do not consider regenerative effects. The only source of acoustic multi-reflections for the weakly coupled case will be due to MEL effects.

Matthews [1] defines three cases where MEL reflections may occur [Figure 3.9]. For all cases, the reflection coefficient r is proportional to the ratio h/λ where h is the height of the step and λ is the wavelength of the SAW at frequency f . The reflection coefficient r is defined as the ratio of the reflected and incident voltage. For overlays and steps, the reflection coefficient goes to zero as the height h goes to zero. For conducting films, a finite reflection exists even for $h=0$. This is because a conducting film shorts out the electric field associated with the wave, changing the wave velocity and characteristic impedance.

Figure 3.9 Surface perturbations that cause surface wave reflections. (a) General case of a thin overlay. (b) Vertical step. (c) Conducting film on a piezoelectric substrate.

The reflection coefficient may be derived from the transmission line model shown in Figure 3.10. The mismatch may be embodied as a turns ratio n and is given by [1]

$$r = \frac{n^2 Z_1 - Z_0}{n^2 Z_1 + Z_0} \quad (3.15)$$

Z_0 and Z_1 are given by $Z_0 = \rho v_0$ and $Z_1 = \rho v_1$ where ρ is the substrate density and v_0 and v_1 are the wave velocities in the two areas shown.

Therefore, (3.15) may be re-written as

$$r = \frac{n^2 v_1/v_0 - 1}{n^2 v_1/v_0 + 1} \quad (3.16)$$

For a conducting film on a piezoelectric substrate, r is well represented by setting $n = 1$, hence $r \equiv \Delta v/2v_0$, where Δv is the velocity difference between the surface waves travelling on the metallized and unmetallized regions and v_0 is the free-surface wave velocity.

Figure 3.10 Transmission line model of reflections caused by an overlay.

The coupling coefficient k^2 of a piezoelectric substrate is defined as [1]

$$k^2 = 2 \frac{\Delta v}{v_0} \quad (3.17)$$

Hence the reflection coefficient for a metallized step may be calculated approximately as

$$r = (1/2) \Delta v / v_0 = k^2 / 4 \quad (3.18)$$

If mass loading is taken into account, the reflection coefficient has to be modified by the addition of a second term [1] since $\Delta v / v_0$ changes slightly.

$$\Delta v / v_0 = \frac{1}{2} k^2 \left(\frac{1}{1 + (\epsilon_0 / \epsilon_p) + (k^2 / 2)} \right) + F h (2\pi / \lambda) \quad (3.19)$$

where: ϵ_0 / ϵ_p is the inverse of the relative dielectric constant (≈ 0.02 for LiNbO_3)

h is the height of the metal

λ is the acoustic wavelength

F is a function of the mass of the metal (0.037 for LiNbO_3).

For 128°LiNbO_3 at 70 MHz with aluminum fingers $2000\text{\AA}$ thick and $\lambda/8$ wide, the reflection coefficient is approximately 1.34×10^{-2} (-37 dB). One can see from this calculation that mass loading has a relatively small effect on $\Delta v / v_0$.

Skeie and Engan [27] have developed a model to predict the

frequency response of filters corrupted by a number of second order effects. For MEL reflections, their model predicts a frequency response

$$|H'(\omega)| \approx |H(\omega)| \sqrt{R_L/R_0} \left(1 \pm \frac{|r| (\cos \theta - \cos N\theta)}{2 \sin \theta} \right) \quad (3.20)$$

where $H(\omega)$ is the uncorrupted frequency response

$H'(\omega)$ is the corrupted frequency response

R_L is the external load

R_0 is the reference load (usually 50Ω)

N is the number of taps

θ is the phase difference between f and f_0

$|r|$ is the reflection coefficient.

In transducers spaced by an integer number of acoustic half wavelengths, the following approximations may be used for narrow band operation

$$\sin \theta \approx \pm 2x/N$$

$$\cos \theta - \cos N\theta \approx \pm 2 \sin^2 x \quad \text{where } x = \frac{Nm\pi (f-f_0)}{2 f_0}$$

and m is the tap spacing in number of acoustic half wavelengths

Accordingly, one finds

$$|H'(\omega)| \approx |H(\omega)| \sqrt{R_L/R_0} \left(1 \pm \frac{|r| N \sin^2 x}{2 x} \right) \quad (3.21)$$

and the maximum ripple (proportional to N) is given by

$$d_{\max} = \pm 0.36 \ln N \quad (3.22)$$

The frequency response of a bandpass filter which is distorted by multi-reflections is compared with the undistorted filter in Figure 3.11.

Figure 3.11 (a) Distortion-free filter. (b) Filter with distortion caused by acoustic multi-reflections.

An efficient way of reducing reflections is to use double electrodes. The effect on the total amplitude may be seen by simply reducing the phase angle θ by a factor of 2. We obtain

$$|H'(\omega)| \approx |H(\omega)| \sqrt{R_L/R_0} \left(1 \pm \frac{\ln(\cos \theta/2 - \cos N\theta)}{2 \sin \theta/2} \right) \quad (3.23)$$

Again, for narrowband operation, the following approximations may be made

$$\cos \theta / 2 \approx 0$$

$$\sin \theta / 2 \approx 1$$

hence

$$|H'(\omega)| = |H(\omega)| \sqrt{R_L/R_0} \left(1 \pm \frac{|r| (-\cos N\theta)}{2} \right) \quad (3.24)$$

This shows that we get a uniform ripple of amplitude

$$d = \pm |r|/2 \quad (3.25)$$

with the frequency of ripple dependent on the number of taps. For 128° LiNbO_3 , assuming $|r| = -37$ dB, even this distortion amounts to approximately 0.12 dB _{peak-to-peak} and represents a fundamental limit to the ripple magnitude for split finger geometries. From the phase factor $N\theta$, the reflection may be regarded as due to a mismatch between the transducer area and the free surface area. This is difficult to compensate inside the IDT region, however as the SAW passes from the IDT to the free-surface region the effect may be reduced by weighting the last dummy tap or electrode as shown in Figure 3.12(a) [27] or by a reflection canceller array [47] [Figure 3.12(b)]. Reflection cancellers are discussed in greater detail in Section 3.2.4.

Figure 3.12 Acoustic reflections between a free surface and a transducer area may be reduced by employing special reflection canceller schemes.

A more rigorous analysis (Program REF3 in Appendix A-1.3) computes the reflections from every tap and sums them at the origin [Figure 3.13]. This allows one to see the effects of multiple taps on the frequency of ripple and also to examine the effects of metallization ratios other than 0.5. The metallization ratio η of an IDT is simply the ratio of the metallized to unmetallized surface area.

The results are shown in Figure 3.14. Notice that for metallization ratios less than $\eta=0.5$ the reflections are reduced in magnitude. However, as η is reduced, the electrode resistance of the electrodes increases dramatically and will cause an increase in the loss of the transducer.

Figure 3.15 shows the magnitude of reflections over frequency for a regular array of split-finger electrodes with $\eta=0.5$. This was computed by performing an FFT on the reflected signals in the time domain (Program FFT in Appendix A-1.5) and shows a $\sin x/x$ dependency with a maximum reflection at $2f_0$. Notice that when the number of electrodes is increased, the bandwidth of maximum reflection decreases and the magnitude of the

reflection at $2f_0$ approaches unity. More importantly, it can be seen that increasing the number of electrodes does not reduce the magnitude of MEL reflections. It only increases the frequency of the inband ripple.

Figure 3.13 Reflection algorithm used in REF3 program

Figure 3.14 Passband ripple based on the model of Figure 3.13 (Program PEF3) with various metallization ratios. Computed with a -37 dB reflection coefficient. (a) $\eta = 0.1$ (b) $\eta = 0.5$ (c) $\eta = 0.9$.

Figure 3.15 Reflection coefficient vs. frequency (Program FFT) for a regular array with a finite number of fingers assuming $\eta = 0.5$ and a single finger reflection coefficient of -37 dB. (a) Single finger (b) 4 fingers (c) 20 fingers.

3.2.2 Triple Transit Echo

As described in Section 3.1, a quarter of the incident power is reflected from a receiving IDT when it is conjugately matched. If both the transmitting and receiving IDTs are closely matched, then the signal is reflected twice [Figure 3.16] and arrives at the output IDT two transit delays after the main signal has reached the output. This is known as the triple transit echo (TTE). The delayed and attenuated TTE will beat with the main signal thus causing ripples in the magnitude and phase of the frequency response. The frequency of this ripple can be determined as the inverse of twice the transit time between transducers [6].

$$f_{\text{TTE}} = 1/2\tau_D \quad (3.26)$$

where τ_D is the time taken for a SAW to travel between IDT centres.

The amplitude level of the TTE is inversely proportional to the transducer mismatch ratio $b = R_a/R_L$. Equations (3.13) and (3.14) illustrate that when $b=1.0$ (ie. $R_L=R_a$) the insertion loss is at a minimum of 3dB but the reflection loss is only 6dB [Figure 3.8] and will result in a peak-to-peak passband ripple of approximately 4 dB. The relationship between the peak-to-peak ripple and triple transit suppression is [1]

$$(\text{TTS})_{\text{dB}} = 24 \text{ dB} - 20 \log(R_{\text{pp}})_{\text{dB}} \quad (3.27)$$

A small passband ripple means a large TTS and this means that a large mismatch ratio b is required for one or both of the transducers.

The measured value of the radiation resistance for a SAW031F

output tap is approximately 20Ω and hence for $b=20$, an input impedance on the pick-up amplifier on the order of 1Ω is required in order to obtain adequate reflection suppression, and consequently low TTE.

Figure 3.16 Triple transit signals are reflected off both transducers.

3.2.3 Substrate Edge Reflections

Reflections from the substrate edges will cause a ripple with a frequency equal to the inverse of twice the delay time between the centre of the IDT and the substrate edge.

$$f_{\text{edge}} = (2\tau_{\text{edge}})^{-1} \quad (3.28)$$

where τ_{edge} is the time taken for the SAW to traverse the distance between the tap centre and the substrate edge [Figure 3.17(a)]. Reflections may be generated for either the input or output taps. A very effective way [10] of reducing the reflections is by laying an absorptive material on the piezoelectric substrate, placed between the substrate edges and the transducers as shown in Figure 3.17(b).

Figure 3.17 (a) Substrate edge reflections. (b) Elimination of edge reflections by use of acoustic absorber material.

3.2.4 End-of-IDT Reflections

End-of-IDT reflections are caused by the mismatch in impedances between the transducer area and the free surface area as the SAW leaves the IDT region [Figure 3.18]. From the analysis of Section 3.2.1.2, it follows that the reflection coefficient at a metallized/free boundary is $20 \log |r| \approx -37$ dB. This reflection can be effectively reduced by employing an ingenious reflection cancellation scheme as shown in Figure 3.19 [47] .

Figure 3.18 A reflection occurs as the SAW leaves the metallized region and enters the free surface region.

Figure 3.19 Reflection cancellation scheme using successively shortened dummy electrodes.

The concept behind the reflection canceller is to spread the end of the IDT out such that the stepped termination of the IDT will not produce echos above -70dB. Dummy electrodes are used to equalize the velocity through the transducer region. In an apodized transducer, it is important to use dummy electrodes outside the coupling region so that the metallization density remains the same across the acoustic beam. Otherwise, serious distortions of the phase front may occur [27]. A continuation of these electrodes, each successively shorter than the previous one, would cause a finite reflection of [10]

$$\begin{aligned}
 V_r &= 20 \log (1/M_r) & (3.29) \\
 &= -40 \text{ dB} & \text{for } M_r = 100 \text{ steps}
 \end{aligned}$$

This, combined with a -37dB reflection coefficient gives a -77 dB echo.

3.2.5 Diffraction and Propagation Losses

The propagation of the surface wave from the input to the output transducer should ideally produce no delay dispersion. In the frequency domain this means a linear phase response and a constant amplitude response of 1.0. In practice however, there is some loss of signal because diffraction causes the beam to widen and hence some of it misses the output transducer. A more serious problem is that the amplitude and phase of the signal depends on the propagation distance so that it is different for various points on the transducer [Figure 3.20]. The usual assumption is that the received signal has a phase delay with respect to the input signal of kL , where $k = 2\pi/\lambda$ and L is the center-to-center distance between the electrodes. This is not exactly true since if any point X is taken on the transmitter and any point Y is taken on the receiver, then the phase delay between signal points X and Y is given by kR . The total signal output is obtained by summing over all points X and Y . Since $R \geq L$ the actual phase delay is always greater than kL . However, the absolute phase difference is not important, since only the relative phase difference between different electrode pairs will cause problems. This can be particularly troublesome in apodized transducers where the finger lengths vary, or in geometries where the apodization pattern follows a slope or V-line configuration. Diffraction effects alter the impulse response of the filter and reduce the out-of-band rejection. A high rejection level requires precision in the tap weight implementation. Also, for practical substrates, the anisotropy in the propagation velocity has to be evaluated as a function of θ and compensated for by adjusting the apodized IDT overlap distances. SAW diffraction compensation is covered in extensive detail in Ref. [6].

Figure 3.20 Diffraction effects in transducers.

In addition to diffraction losses, the SAW experiences an acoustic loss of approximately 0.1 dB per microsecond of delay between transducers. Acoustic loss is caused by non-ideal crystal properties. The loss value of 0.1 dB/ μ s. is an approximation derived for transducers operating below 100 MHz [1].

3.2.6 End Effects

End effects refer to the change in tap weight value due to fringing electrostatic fields not accounted for in the first-order analysis. Since electrostatic fields caused by a line charge decay quite slowly as a function of the distance, the charge on an electrode will depend quite strongly on the position, shapes, and polarity of the neighbouring electrodes. In a conventional transducer with equally spaced fingers, the charge distribution for the outermost electrodes deviates significantly from that of an infinite array. Examples of such charge distributions [Figure 3.21] show that for a small number of electrodes (i.e. the output IDT) the charge distribution can be far from ideal and must be accounted for in the design process [19,49]. The effect of the non-uniform charge distribution is to change the impulse response of the IDT.

Figure 3.21 Electrostatic charge distributions for various transducer configurations.

Another type of end effect, called the transverse end effect [48] is encountered in apodized transducers. The charge on an apodized tap is

assumed to be proportional to the overlap with the grounded finger. However, the electrostatic fields fringe out [Figure 3.22] beyond the length of the electrode making it seem longer than it actually is. In apodized transducers, this effect is proportionally larger for small electrodes with small overlaps.

Figure 3.22 Transverse end effects make the electrical overlap distance larger than the physical overlap distance.

A third type of end effect is called "end finger to ground excitation" [17]. This occurs at the ends of an IDT where charge coupling of the end dummy fingers to the case ground exceeds the overlap of the last finger. Figure 3.23 shows that when the charge is integrated over the end-most fingers, it will generally exceed the charge distribution established by the apodization. As a result, a spurious signal appears in the time domain corresponding to the position of these end fingers.

The V-line apodization topography [43,17] illustrated in Figure 3.24 has been successfully used to reduce ground coupling of electrodes

with small tap weights.

Figure 3.23 Driven finger to ground plane coupling causes end fingers to have charge distributions significantly larger than that established by apodization.

Figure 3.24 V-line apodization reduces ground plane effects to an acceptable level.

3.2.7 Spurious Losses

Two sources of spurious losses are electrode resistance, i.e. ohmic loss due to current flow through the electrodes, and the generation of bulk waves. Electrode resistance is fairly independent of frequency and is caused by the resistivity of the metal fingers. Naturally, reducing the metallization ratio η will increase the electrode resistance, however, this can be reduced by the use of thicker metal films for the electrodes.

Transducers excite not only Rayleigh waves but also bulk waves which propagate through the piezoelectric substrate rather than on the surface. These wave motions can propagate between transducers directly, or bounce off the bottom of the substrate. For YZ-LiNbO₃, bulk waves propagate steeper into the substrate as the frequency of excitation increases [Figure 3.25.] Bulk waves will propagate at angles θ for which constructive interference exists [9].

$$\theta = \sin^{-1} \left(\frac{2(n-1) v_b}{f_0 \lambda} \right) \quad n = 1, 2, 3 \text{ etc.} \quad (3.29)$$

where v_b is the bulk wave velocity.

The compressional waves travel at much higher velocities than the surface waves. Horizontal shear waves travel only 3% faster than do Rayleigh waves. To reduce or eliminate bulk wave coupling, careful choice of substrate material is required. YZ-LiNbO₃ is a popular choice of substrate material. It has a compressional bulk wave coupling at about twice the SAW frequency, but only about 20 dB below the main SAW coupling. Much of the bulk wave coupling may be scattered or absorbed by roughening the bottom of the substrate so that coherent reflection of the waves does not take place. However, roughening the backside of the substrate will not

prevent direct-coupled shear waves.

Horizontal shear waves often result in increased sidelobe levels on the high frequency side of the passband. Fortunately, 128° rotated, Y-cut X-propagating LiNbO_3 has a horizontal shear bulk wave coupling which is 60-70dB below the Rayleigh wave. It also has a slightly higher coupling coefficient k^2 of approximately 5.4%. YZ LiNbO_3 has a coupling coefficient of approximately 4.8%. This makes 128° LiNbO_3 to date the best material for wide-band filters.

Figure 3.25 Bulk waves in a piezoelectric substrate propagate into the substrate at steeper angles for increasing frequency.

3.2.8 Electromagnetic Feedthrough

Electromagnetic feedthrough in surface wave filters is defined as the energy that is electrically coupled or radiated from one transducer to another [1]. Since this signal travels at nearly the speed of light, it experiences zero delay when compared to the delay, τ_D of the filter [Figure 3.26]. In the frequency domain, EM feedthrough causes amplitude, phase, and delay ripple, as the EM signal beats with the main delayed signal. Another effect is poor rejection. If the device is improperly grounded, or if RF shielding between the input and output pins is poor, or if there is inductive or capacitive coupling in the SAW crystal or matching components, then the signal will jump from the input IDT to the output IDT without going through the filter.

Figure 3.26 The delay between signals in the time domain shows that the EM feedthrough experiences almost no delay compared to the acoustic signals.

With SAW filters capable of greater than 60 dB rejection, the effect of EM feedthrough can vary from a slight degradation to complete obliteration of the response, depending on its magnitude. A seriously degraded response due to EM feedthrough is shown in Figure 3.27.

Figure 3.27 SAW03L filter response. (a) Normal filter response (b) Response seriously degraded by EM feedthrough

EM feedthrough is most often a problem on printed circuit (PC) boards where isolation from input circuits to output circuits is often required. A number of techniques that can be used to hold EM feedthrough to an acceptable level are listed below [38].

- 1) Use double-sided PC board, preferably of epoxy resin or teflon.
- 2) Ground the SAW package by soldering it to the component side of the PC board.
- 3) Cut all leads as short as possible and reduce package elevation above the PC board to an absolute minimum.

- 4) Use plated-through holes or several feethrough jumpers around the SAW device to connect the bottom ground plane to the top one.
- 5) Place the matching components of the input network on one side of the board and the components of the output network on the other side of the PC board (if the SAW device has an output matching network). This technique prevents feedthrough due to lead and component radiation.
- 6) Reduction of feedthrough through the dielectric of the PC board can be accomplished by cutting a narrow strip through the dielectric between the input and output pins, inserting a thin metal shim between and soldering it to both the top and bottom ground planes.

Experimentation with various grounding topographies has shown that separate input and output ground planes with a physical shield placed between the input and output pins is one of the most effective methods for supressing EM feedthrough.

The most common method of electromagnetically feeding an interdigital array is to connect the two sum bars of the transducer to an unbalanced transmission line with gold bonding wires. In this arrangement, one side of the transducer is grounded to the metallic box housing the crystal. The resulting field distribution is shown in Figure 3.28(a). In this configuration, the package may be considered as a waveguide and the relative attenuation between any two points within the crystal cavity assuming a dominant TE_{10} mode can be determined using the following equation [1];

$$\gamma = \frac{54.5d_T}{\lambda_c} \{1 - (\lambda_c/\lambda)^2\} \quad (3.31)$$

where γ is the isolation between transducers in dB , λ the operating

wavelength in the cavity, λ_c the cavity cutoff wavelength, equal to twice the cavity width ($\lambda_c = 2a$), and d_T is the distance between transducers.

Providing sufficient isolation between closely spaced transducers can be a significant problem in the design of tapped delay lines with IDT spacings of fractions of a microsecond. A solution to this problem is to drive the IDT in a balanced fashion as shown in Figure 3.28 (b) such that there is no net normal electric field [1]. The use of a balanced feed on IDTs is very desirable, however, baluns are not very small and require additional volume to package. Also, most systems have unbalanced feeds for convenience and shielding and hence use of such a scheme will be determined by the system requirement.

Figure 3.28 EM field lines caused by different feed arrangements (a) Unbalanced electrical connection. (b) Balanced electrical connection.

For radios employing 25% QAM, an EM feedthrough of 60 dB down from the SAW filter output is required. Assuming a device insertion loss of

30 dB, this means that an overall isolation of 90 dB is required from the input to the output. The ripple generated by this amount of feedthrough is found from the following;

$$\delta_{EM} = 10^{(-60/20)} = 0.001$$

$$\text{ripple}_{dB} = 20 \log \left(\frac{1.0 + \delta_{EM}}{1.0 - \delta_{EM}} \right) = 0.02 \text{ dB}_{p-p} \quad (3.32)$$

At least 60 dB of EM isolation is required since EM feedthrough generally causes fast ripple in the filter passband. In a $(2N+1)$ tap equalizer, ripple with more than N peaks in the passband cannot be reduced for since it falls outside the range of the equalizer.

CHAPTER 4

SAW DELAY LINE DESIGN AND CONSTRUCTION

4.1 Design Analysis

Given a set of filter specifications, a set of real-valued tap weights must be calculated. Usually the required filter specifications contain one or more of the following:

- 1) centre frequency f_0 ;
- 2) 3 dB bandwidth ;
- 3) shape factor α ;
- 4) sidelobe level ;
- 5) insertion loss ;
- 6) in-band amplitude and phase ripple ;
- 7) insertion delay ;
- 8) temperature coefficient of delay/frequency drift.

Figure 4.1 Typical filter specification.

A typical filter specification is illustrated in Figure 4.1. The goal is to obtain a set of real and equally spaced (periodic) finger weights for both the input and output IDTs in order to satisfy these requirements. This is generally done by transforming the frequency domain samples into time domain samples using the inverse discrete fourier transform (IDFT), applying a window function and truncating. Some basic relationships for the DFT and IDFT are listed below.

F_f = total range in the frequency domain

Δf_f = spacing between samples in the frequency domain

T_t = total range in the time domain

Δt_t = spacing between time samples

$$T_t = (\Delta f)^{-1}$$

$$F_f = (\Delta t)^{-1}$$

N_{sp} = number of sample points in time and frequency domains

$$N_{sp} = \frac{F_f}{\Delta f_f} = \frac{T_t}{\Delta t_t} = F_f T_t = \frac{1}{\Delta f_f \Delta t_t} \quad (4.1)$$

Fast Fourier Transform (FFT) algorithms require that the number of sample points be a power of 2. Using split finger geometries (ie 4 fingers per wavelength) will result in 4 time samples per wavelength. Hence for $f_o = 70$ MHz, the time samples occur at $1/(4 \times 70 \text{ MHz})$ and $F_f = 280$ MHz. The overall length of the time domain response will be determined by the number of fingers required to satisfy the design requirements. However, for the type of filters presented in this thesis, the total number of fingers

does not generally exceed 600, hence a total of 1024 (2^{10}) samples is chosen for the FFT algorithm. This corresponds to $T_i = 1024/280\text{MHz} = 3.657 \mu\text{s}$, $\Delta t_i = 3.571 \text{ ns}$, and $\Delta f_i = 0.273 \text{ MHz}$. The filter time domain response will only last as long as $N_T \Delta t$ where N_T is the total number of transducer fingers including both the input and output transducer. It is within this constraint that the cascaded IDT impulse responses must be designed to obtain the required frequency characteristics.

The total transfer function $H(\omega)$ of the SAW filter is a product of the input IDT response $H_i(\omega)$, the output IDT response $H_o(\omega)$, the finger element factor $H_e(\omega)$ for split-fingers (analogous to the element factor in an antenna array), and the circuit factors $H_{c1}(\omega)$ and $H_{c2}(\omega)$ where the subscripts 1 and 2 relate to the input and output transducers.

$$H(\omega) = H_i(\omega) \times H_o(\omega) \times H_e(\omega) \times H_{c1}(\omega) \times H_{c2}(\omega) \quad (4.2)$$

If the circuit factors are designed to be wider than the acoustic response so that the amplitude response is not significantly affected, then $H_{c1}(\omega)$ and $H_{c2}(\omega)$ may be set to unity. A complete analysis will also include the second-order effects discussed in Chapter 3.

A straight forward and effective design technique for filters with tight tolerances makes use of window functions [31,50]. Several window functions such as the Hamming, Gaussian or Kaiser-Bessel [50] will give extremely low inband ripple content and good sidelobe suppression. However, in order to optimize certain parameters, a more sophisticated design method is required.

A quite simple design method is based on an iterative use of the

Fourier transform [46]. The required response $H(\omega)$ is first transformed into the time domain, truncated to a finite length D by a rectangular window function, and then transformed back into the frequency domain. Due to the truncation, this new frequency response will contain Gibbs oscillations [2] and will generally fall outside of the specified tolerances in some regions [Figure 4.2]. A new frequency specification is generated by multiplying the original specification by an error function. The error function is obtained by dividing the original specification by the truncated response. The procedure is then repeated, transforming to the time domain, truncating and returning to the frequency domain. Provided that D is large enough, the iteration of this process usually gives progressive improvement, yielding finally a frequency response meeting the filter specifications. The resulting finite-length impulse response will determine the apodized transducer geometry.

Figure 4.2 (a) Brickwall filter (b) Rectangular window function (c) Truncation of the time domain response causes Gibbs oscillations in the frequency domain.

The preceding discussion basically describes the approach taken in designing the filters used in this thesis. The number of fingers in the apodized IDT is arbitrarily determined by the designer in order to obtain the required ripple and sidelobe levels.

Two filter designs were considered as shown in Figure 4.3. The first design (SAW03F) has the following specifications for each tap output : a flat passband response (less than 0.05dB ripple) with a Nyquist bandwidth of 22.585 MHz, a filter roll off factor $\alpha = 12\%$, flat group delay, and a delay of 44.3 ns between taps. The second filter (SAW03L) has the following specifications for each tap output : a flat passband response (less than 0.05dB ripple) with a Nyquist bandwidth of 34.667 MHz, a filter roll off factor $\alpha = 25\%$, flat group delay, and a delay of 44.3 ns between taps. The SAW03L filter has a much wider bandwidth and larger α since it is basically intended to be used as a roofing filter and therefore does not require the narrow, sharply defined passband of a pulse shaping filter.

Figure 4.3 (a) SAW03F specification. (b) SAW03L specification.

4.2 Design Procedure

The output taps must all be identical hence, unapodized. The input IDT is apodized in order to generate the correct finger overlap weights as determined from the filter specifications. The design of the input and output IDTs are considered in the following sections.

4.2.1 Unapodized IDT Design

The first step in the design of the filter is to determine the number of finger pairs required in the unapodized transducer. From (3.4) the number of finger pairs that optimizes the resonant bandwidth for minimum insertion loss is calculated to be $N_{opt} = 3.8$ for 128° rotated Y cut LiNbO_3 . The 4dB bandwidth from (3.5) is $\Delta f \approx f_o/N = 26\%$. By using 2.5 finger pairs, a slightly larger acoustic bandwidth of 40% is obtained at the expense of a marginal increase in the insertion loss.

The unapodized IDT response carries uniform weighting for each finger pair and hence the frequency response will be the Fourier transform of a 2.5 finger pair impulse response [Figure 4.4]. The effect of using split-fingers is to introduce an element pattern $H_e(\omega)$ into the IDT frequency response. This causes an asymmetry in the unapodized response as shown by the dotted line in Figure 4.5. Since the edges of this response are below the specification (solid line), this discrepancy must be compensated for in the design of the apodized transducer.

Next the charge distribution of the output IDT is calculated [19]. The resulting non-uniform charge distribution becomes the modified impulse response of the IDT. The modified impulse response causes a slight up-slope in the frequency response.

The unapodized transducer pickup circuit has no matching elements since it is loaded with a low impedance non-reactive load. Hence, the circuit factor for the output transducer $H_{c2}(\omega)$ may be set to unity.

Figure 4.4 Equally weighted (unapodized) N finger pair transversal filter. (a) N finger pair IDT impulse response. (b) Split-finger element pattern. (c) Frequency response of an N split-finger pair IDT is comprised of a convolution of (a) and (b) in the time domain and a multiplication of (a) and (b) in the frequency domain.

Figure 4.5 The element pattern of a split finger transducer causes an assymetry in the frequency response of the unapodized IDT as shown by the dotted line. The filter specification is shown by the solid line.

4.2.2 Apodized IDT Design

The design specification for the apodized transducer is obtained by dividing the filter specification by the unapodized IDT response, as shown in Figure 4.6.

If the apodized IDT specification were used directly as an input to an FFT, the resultant time samples generated would be complex in nature due to the asymmetries in the amplitude and phase specifications. Since real-valued finger weights are to be implemented, we would be faced with the problem of trying to realize complex finger weights with real finger placements. To overcome this problem, a Hermitian [51] set of specifications is generated which results in equally spaced real-time samples. The reader is referred to [51] for an in-depth analysis and solution to this problem.

Figure 4.6 Apodized IDT specification (dotted line) vs. filter specification (solid line).

Once the Hermitian specifications have been generated, the transducer sample weights are synthesized via a 1024 point complex FFT. The resulting time samples are truncated at the designer's discretion.

There is a basic tradeoff at this point relating the number of fingers, the sidelobe level and the ripple content in the passband. For example, the ripple content and sidelobe levels of the wideband SAW03L delay line ($f_s=34.667$ MHz, $\alpha = 25\%$) for different numbers of fingers in the apodized IDT are shown below in Table 4.1.

Table 4.1

Computed values of in-band amplitude ripple and side lobe suppression for the SAW03L delay line for various numbers of split-fingers in the apodized IDT.

<u>No. AIDT</u>	<u>No. AIDT</u>	<u>Max. In-band Ripple</u>		<u>Sidelobe Level</u>
<u>Electrodes</u>	<u>Split-fingers</u>	$\pm$ (dB)	Δ (dB)	(dBc)
160	80	+0.124/-0.002	0.126	-42
200	100	+0.000/-0.073	0.073	-44.0
220	110	+0.010/-0.023	0.033	-43.7
260	130	+0.021/-0.003	0.024	-45.1
300	150	+0.004/-0.027	0.031	-46

The SAW03L device was chosen to have 150 split-fingers in the apodized IDT and the SAW03F device was chosen to have 208 split-fingers in order to achieve similar ripple and sidelobe suppression characteristics with a narrower bandwidth.

As with the design of the unapodized IDT, the split-finger element pattern $H_c(\omega)$ will modify the apodized IDT response. Also, the circuit factor $H_{c1}(\omega)$ for the apodized IDT is set to unity since the matching circuit is designed to minimize amplitude variations.

At this point the charge distribution model is applied to each sample and the charge is calculated via superposition, i.e., the individual

response of the apodized transducer is then calculated via an FFT.

Since the charge distribution model changes the sample weights, the resulting frequency response differs from the apodized IDT specification. To correct for these differences, an error function is generated and the synthesis and analysis is iterated three to five times [19].

On the final iteration, the apodized IDT design is saved and multiplied with the unapodized transducer response. The overall response is shown as the dashed line in Figure 4.7.

Figure 4.7 Analysis (dotted line) vs. specification (solid line) frequency response obtained from the cascade of the input and output IDTs.

A representation of the design process is illustrated in Figure 4.8 and a flow chart of the computer design routine is indicated in Figure 4.9 .

Figure 4.8 Basic steps in the design process of a SAW filter.

Figure 4.9 Flowchart showing computer design sequence for a SAW filter .

4.3 Output Tap Design

The function of the tapped SAW delay line is to provide seven identical output signals which are multiplied by the appropriate tap coefficients. The requirement that the frequency response of each output be the same is most easily achieved by making each output IDT topographically identical and as simple as possible. An unapodized design with a small number of fingers will provide the easiest solution and at the same time allow for independent adjustment of the delays between taps.

In a linear transversal equalizer the delay between adjacent taps is equal to the symbol period of the modulation scheme being used. For a symbol rate of f_s Hz, the delay between adjacent taps must be $1/f_s$. Assuming a symbol rate of 22.585 MHz, the required delay between adjacent taps will be 44.3 ns [Figure 4.10].

Figure 4.10 Tap spacing arrangement for 22.585 MHz symbol rate.

Optimal performance from the SAW device also requires that acoustic reflections and feedthrough from electromagnetic coupling be reduced to a minimum. EM coupling is primarily caused by non-ideal packaging and circuit board layout configurations. The problems associated with EM coupling are solved by the use of a glass carrier for the SAW device (described in detail in Section 6.1.3), and appropriate electronic printed circuit board layout techniques (Chapter 5).

As mentioned in Section 3.1, acoustic multi-reflections caused by MEL and regeneration effects [Figure 3.6] can distort the desired frequency response of the filter. MEL reflections can become unacceptable in designs where the metal electrodes are spaced by one-half wavelength. This geometry causes reflections to add in phase [27] causing undesirable ripples in the passband of a filter. MEL reflections are reduced by using split-finger geometries [2,10,22] which provide a high degree of cancellation of reflections from successive electrodes [Figure 4.11].

Figure 4.11 (a) $\lambda/8$ finger spacing reduces ripple caused by MEL effects. (b) $\lambda/4$ finger spacing causes reflections to add in phase.

Regeneration occurs when a primary SAW passes underneath a pickup transducer and induces a current which is delivered to an electric load. The voltage developed in the electric load regenerates a secondary SAW which propagates away from the pickup transducer in both forward and backward directions. By varying the load impedance, the magnitude of the load voltage can be reduced at the expense of the insertion loss of the transducer. By utilizing split-finger electrodes and a low impedance load, it is possible to reduce reflections due to MEL and regeneration to an acceptable level.

Obtaining a delay of 44.3 ns between adjacent taps would not present a problem to the topographical layout of the output IDT if we did not require metallization between the taps. However, in order to keep a tight tolerance on the propagation velocity (and hence delay) of the SAW between taps, it is necessary to maintain a metallization ratio of 50% using grounded dummy fingers. If it were possible, an integer number of $\lambda/8$ grounding bars could be placed between adjacent taps to make up a 44.3 ns delay. Unfortunately, for an f_0 of 70MHz, the standard $\lambda/8$ spacing of these grounding bars does not permit this and the only way to resolve the problem is to change the finger spacing of the output IDT, or the grounding bars, or both. Since changing the finger spacing of the output IDT effectively alters the frequency of operation, the only recourse is to change the spacing of the grounded dummy fingers [Fig. 4.12].

Figure 4.12 Dummy finger placement in order to achieve correct delay between taps. (a) Topography showing separately grounded dummy fingers. (b) Topography showing common grounding bar.

Section 4.3.1 describes the analysis of MEL reflections generated by a number of different output IDT designs and presents the reasons for choosing one design over the others. Section 4.3.2 describes the theory behind the regenerative wave phenomenon and the steps taken to reduce the effects.

4.3.1 Reflection Supression Analysis

As mentioned in Section 3.2, split finger geometries with finger widths of $\lambda/8$ provide a high degree of cancellation of MEL reflections from successive electrodes. The problem with having dummy fingers with finger widths slightly wider than $\lambda/8$ is that the reflections due to MEL no longer cancel completely and a small ripple will be present in the passband. As a result, we are faced with a choice of how to arrange the grounded dummy fingers so that the spacing from one tap to the next is equal to 3.0995λ , the spacing required to make up a delay equal to one symbol period. The choices available are illustrated in Figure 4.13. As already mentioned in Section 3.1, changing the finger spacing of both the grounded dummy fingers and the driven fingers [Figure 4.13 (e)] will change the frequency of operation and hence is not the preferred arrangement. However, so as to be complete, the frequency responses of all the configurations shown in Figure 4.13 are analysed to see the effect of MEL reflections.

The analysis of the MEL reflections illustrates [Figure 4.14] that for the configuration that includes no grounded dummy fingers, an undesirable ripple occurs in the passband. For the cases where 2, 4, and 6 grounded dummy fingers are placed between taps, the frequency responses are almost identical. The configuration consisting of two grounded dummy fingers between taps was chosen for the final design since it gives a response that has an acceptable ripple and minimizes the number of modifications required in the computer design routines.

The results illustrated in Figure 4.14 are obtained from program REF1 listed in Appendix A-1.1. REF1 computes MEL reflections based on a summation of reflections off the edges of the transducer fingers.

Figure 4.13 Five possible output tap finger spacing topographies showing the use of wider dummy fingers to make up for the extra distance the SAW travels during T (symbol period = 44.3ns). Each topography shown consists of twelve electrodes.

(a)

(b)

(c)

(d)

(e)

Figure 4.14 Computed frequency response illustrating ripple characteristics of MEI cases of Fig. 4.13 (a)-(e). Calculations are based on the addition of reflections from taps 2 to 7 on tap 1. (Refer to Program REF1)

4.3.2 Regenerated Wave Analysis

As described in [10], the insertion loss of an IDT can be reduced by electrically matching the IDT to the source or load impedance. However, a matched output transducer delivers half the energy of the incident SAW to the load and regenerates a secondary SAW which carries $1/4$ of the energy in a forward direction and $1/4$ of the energy backwards towards the input IDT. This constitutes in the case of successive tapped outputs a loss of 6 dB per tap. More importantly, the magnitude of the regenerated SAW from each tap will cause severe degradation of the frequency response in adjacent taps. Other secondary effects like triple transit echos will also degrade the output signal.

Since the objective of this work is to demonstrate the feasibility of using a tapped SAW delay line in an ATDE transversal filter, the basic requirements are that the frequency response from one tap to the next be identical and that the relative insertion loss from one tap to the next be small. Since matching the output transducers will generate a number of undesirable effects which will clearly violate this requirement, another method of extracting energy from the taps will have to be implemented.

Recall from Smith's paper [14] (Section 3.2.1.1) that reflections are at a minimum when the electrodes are weakly coupled to the substrate by either a short or open circuit on the electrical port. A short on the output will effectively prevent regenerative effects and also allows the incident SAW to pass unperturbed under the IDT with no loss in energy. Based on this premise, one can deduce that a low impedance load on the IDT output will produce almost no regenerated SAWs and at the same time prevent excessive attenuation of the incident SAW. The tradeoff however is that the output insertion loss will be rather high due to the large electrical mismatch and will require buffering to compensate for the loss in signal

strength. Hence it was proposed [23] that a low input impedance buffer be used to tap the signal energy from the output taps in order to simulate an AC ground and provide some degree of amplification for the attenuated signal.

As mentioned in Section 3.2.3 the reflection loss and transmission loss are proportional to the transducer mismatch ratio b . Since the measured value of R_a for a transducer with 2.5 split-finger pairs is approximately $20\ \Omega$, and the practical minimum value of input resistance for a common base amplifier is $1\text{--}2\Omega$, a value of $b \approx 20$ is achievable. For this value of b , (3.13) and (3.14) predict a transducer reflection loss L_{11} of 26 dB and a transducer transmission loss L_{31} of 10 dB.

4.4 Input Matching Network

In order to reduce system amplification requirements and prevent degradation in the signal to noise ratio it is desirable to minimize the insertion loss of the SAW device. Signal fidelity is achieved with a low spurious content and minimization of acoustic multi-reflections. Unfortunately in order to obtain a low spurious level, a sacrifice in the insertion loss must be made. A matching network is the means by which a satisfactory compromise can be achieved.

There are basically three different matching structures [37] available for a SAW IDT [Figure 4.15]. Conjugate matching with parallel or series tuned inductors provides an interface to external circuits by tuning out the interelectrode capacitance C_T for transducers with $Q_a \geq 1$. For transducers with $Q_a \ll 1$, it is customary to use magnitude matching. Magnitude matching in this case is as good as conjugate matching, since if $G_a \gg \omega C_T$ the capacitance does not really need to be tuned.

Figure 4.15 Three common IDT matching circuits (a) Shunt matching with parallel inductor (b) Series matching with series inductor (c) Magnitude matching. An added mismatch resistor R_{ser} or R_p may be introduced to lower peaking caused by the matching network.

4.4.1 Criteria for Network Design

The important parameters in the design of a matching network are the transducer insertion loss, triple transit suppression, and the maximum fractional bandwidth.

As mentioned in section 3.2.1.1, the reflection loss and transmission loss are proportional to the transducer mismatch ratio b , hence in order to achieve adequate reflection suppression and at the same time not have an excessive insertion loss, b should be chosen accordingly. The triple transit suppression (TTS) is approximately given by

$$(TTS)_{dB} \approx (L_{11 \text{ in}})_{dB} + (L_{11 \text{ out}})_{dB} \quad (4.3)$$

where $L_{11 \text{ in}}$ and $L_{11 \text{ out}}$ are the reflection loss for the input and output IDTs.

It can be seen from this expression that for conjugately matched input and output IDTs (i.e., 6 dB reflection loss each) the TTE is suppressed only 12 dB below the main signal. As mentioned previously, a matched input IDT is required to reduce the insertion loss and hence the overall TTS will be determined primarily by the reflection loss L_{11} of the output IDT. The overall TTS for a matched input IDT ($b = 1$) and the output IDT ($b = 20$) can be calculated approximately from (4.3) as $(TTS)_{dB} \approx 6\text{dB} + 26\text{dB} = 32\text{dB}$.

The next step is to determine the matching network parameters so that the maximum bandwidth is obtained for the minimum insertion loss. The following analysis illustrates the relationship between the transducer insertion loss and maximum fractional bandwidth as a function of the coupling constant k^2 .

The acoustic quality factor Q_a is defined as [1]

$$Q_a = \frac{\omega_o C_T}{G_a} = \frac{\pi}{4k^2} \frac{\Delta f}{f_o} \quad (4.4)$$

where Δf represents the acoustic 4 dB bandwidth. Although derived for an unapodized transducer, this formula is approximately correct for other types of transducer arrays. It is also necessary to account for other resistive losses in the electrodes, the matching network, and the conductance added to optimize the mismatch. These losses can be accounted for by adding a shunt conductance G or series resistance R to the circuit. If the matching network consists of a shunt or series inductor resonant with C_T and matching resistances, then the Q of the network is

$$Q_n = \frac{\omega_o C_T}{2(G+G_a)} = \frac{(\omega_o C_T)^{-1}}{2(R+R_a)} = \frac{f_o}{\Delta f_n} \quad (4.5)$$

where R and R_a are the counterparts of G and G_a for the series circuit, Δf_n denotes the 3 dB bandwidth of the network and the factor of 2 in the denominator arises because of the conjugate match.

The maximum fractional bandwidth $\{\Delta f/f_o\}'_{\max}$ obtainable without significantly altering the frequency response is obtained when the network bandwidth is at least twice the acoustic bandwidth [1], ie. $\Delta f_n = 2\Delta f$. Imposing this condition, the following is obtained.

$$\{\Delta f/f_o\}'_{\max} = 2 \left\{ \frac{k^2}{\pi} \left(\frac{G}{G_a} + 1 \right) \right\}^{1/2} = 2 \left\{ \frac{k^2}{\pi} \left(\frac{R}{R_a} + 1 \right) \right\}^{1/2} \quad (4.6)$$

The maximum intrinsic fractional bandwidth (i.e., when $G=0$) or

$R=0$) is related to the coupling coefficient by the following relationship

$$(\Delta f/f_o)_{\max} = 2 \left(\frac{k^2}{\pi} \right)^{1/2} \quad (4.7)$$

Figure 4.16 shows the maximum fractional bandwidths for a number of commonly used substrate materials.

When the desired acoustic frequency response of a filter is nearly flat, then a shunt conductance G_{sh} or series resistance R_{ser} must be chosen such that Δf_n is larger than $2\Delta f$ in order to preserve the amplitude and phase characteristics across the entire bandwidth. The higher the ratio of $\Delta f_n/\Delta f$ the higher the insertion loss but the less severe the band edge degradation. Matthews [1] has developed an important relation between the transducer insertion loss and the maximum fractional bandwidth.

$$IL = \left(\frac{K^2 \pi^2}{16 k^4} \right) (\Delta f/f_o)'_{\max}{}^4 \quad (4.8)$$

where $(\Delta f/f_o)'_{\max} \geq (\Delta f/f_o)_{\max}$ and $K = (\Delta f_n/\Delta f) \geq 2$. A composite family of curves [Figure 4.17] shows the relationship between the upper and lower band edge degradation, and actually is the result of two equations to determine both the upper and lower band responses. For large percentage bandwidths (greater than 10%), the amplitude degradation is higher on the lower band edge. The curves shown illustrate the response for a single transducer inductor-tuned matching network.

Figure 4.16 Curves that define the maximum fractional bandwidth for the minimum transducer insertion loss [1].

Figure 4.17 A composite family of curves that predicts the upper and lower band edge degradation due to a tuned matching network [1].

Generally the interelectrode capacitance of an IDT is much larger than the radiation resistance and hence, if interfaced directly with a purely resistive load, a frequency dependent mismatch is obtained and the transducer insertion loss is excessive. In this case it is desirable to introduce a series (or parallel) inductor to tune out C_T . The network Q_n of a series tuned transducer is

$$Q_n = 1/(2\pi f_o C_T' R_T) \quad (4.9)$$

$$\text{where } R_T = R_a + R_c + R_s + R_{ser}$$

$$\text{and } R_c = \text{series electrode resistance} = 4\rho W_{max}/(N\lambda_o)$$

where ρ is the metal resistivity and W_{max} is the maximum electrode overlap

$$R_a = \text{series radiation resistance} = G_a/(\omega_o C_T)^2$$

$$R_s = \text{generator source resistance}$$

$$R_{ser} = \text{added series mismatch resistor}$$

$$C_T' = \text{the series counterpart of } C_T. \text{ Note } C_T \approx C_T' \text{ for } G_a \gg \omega C_T.$$

The series resistance R_{ser} provides sufficient mismatch with the radiation resistance R_a . Of course, a shunt matching network can also be implemented, however, if the inductor is placed across the transducer, the impedance seen by the source at ω_o is $1/G_a(\omega_o)$. For the apodized transducers of SAW03F and SAW03L this is on the order of 800 Ω and is difficult to match. A series circuit allows one to match to the input IDT with a much lower impedance (10-20 Ω) and from the standpoint of the electronics, is the preferred choice.

4.4.2 Passive Matching Network

The following analysis shows how the input matching network values are selected.

The frequency at which the pulse shaping filter amplitude is nil occurs at frequencies of

$$f = f_0 \pm \frac{\Delta f_z (1+\alpha)}{2 (1-\alpha)} \quad (4.10)$$

where Δf_z is the 0 dB passband bandwidth. The bandwidth of the delay line must be greater than the bandwidth of the shaping filter in order not to perturb the shaping response. This works out to a bandwidth of approximately 26 MHz. The corresponding fractional bandwidth $(\Delta f/f_0)'_{\max}$ is approximately 37% which is greater than the maximum intrinsic fractional bandwidth $(\Delta f/f_0)$ of 26% for LiNbO_3 . From Figure 4.17 it can be seen that in order not to degrade the passband edges, K should be chosen to be 3. Based on this value of K , the insertion loss may be calculated from (4.8). For $K=2$, $k^2=5.4\%$ for 128°LiNbO_3 and $(\Delta f/f_0)'_{\max} = 0.37$, the transducer insertion loss is approximately $10 \log(36)$ or 15.5 dB. Since $K=\Delta f_n/\Delta f$, this implies that the network bandwidth be equal to three times the acoustic bandwidth.

The measured SAW03L input impedance is approximately $12-j115 \Omega$, corresponding to a C_T of 20pF.

For the series tuned transducer, the network $Q_n = f_0/\Delta f_n = 1/(2\pi f_0 C_T' R_T)$ where $R_T = R_a + R_e + R_s + R_{\text{ser}}$. The maximum acoustic bandwidth $\Delta f_{\max}$ is 37% for 128°LiNbO_3 and hence R_T is calculated to be

approximately 102Ω . $R_a + R_c + R_s = 17\Omega$ hence R_{ser} is 85Ω . A standard value of 82Ω is chosen for the mismatch tuning resistor.

The series inductor is calculated to resonate with the static capacitance C_T as follows;

$$\omega_o L_{ser} = \frac{\omega_o C_T}{(G_a)^2 + (\omega_o C_T)^2} \quad (4.11)$$

here G_a can be calculated at resonance from the the series equivalent circuit.

$$G_a = \frac{R_a}{R_a^2 + (\omega_o C_T)^{-2}} \quad (4.12)$$

From measured values of apodized transducer impedance, the value of inductor to be used for the SAW03L device is approximately $0.260\mu H$ and the value to be used for the SAW03F device is approximately $0.170\mu H$.

4.5 Packaging Structure

As mentioned in Section 3.2.8 control of EM feedthrough is of importance so that excessive fast ripple is not present in the passband. Proper packaging and grounding can effectively reduce most EM problems. These precautions are listed in Section 3.2.8.

In conventional two transducer SAW devices, the bonding wires connecting the IDTs and the package pins can be arranged to fan out away from each other [Figure 4.18(a)] so that there is very little EM coupling interaction between them. However, in a multi-tapped device, providing sufficient isolation between tap outputs can be a problem [Figure 4.18(b)]. The conventional solution to this type of problem is to use a balanced transformer to resonate the output so that the net electric field is zero, as described in Section 3.2.8. However, such a transformer has a self inductance which may be of sufficient value to resonate the small static capacitance of an output IDT. Since this clearly causes acoustic regeneration, this approach must be avoided. Regeneration may also exist with a large number of outputs, where the self inductance of long fan-out bond wires causes loading. Hence the criteria for the packaging structure are that it provide EM isolation and a low inductance path for the tapped outputs.

Experimental results (see section 6.1.2) show that the EM coupling is reduced as the angle between the input bond wire and the output bond wire goes to 180° . This suggested the use of a glass carrier [Figure 6.16] with aluminum fan-out lines deposited on a glass slide. The fan-out lines are made as wide as possible to reduce the characteristic impedance and the orientation is away from the input to reduce input-to-output EM coupling. The tapped delay line is mounted on the carrier which is then mounted in a 24 pin package as shown in Figure 6.17.

Figure 4.18 (a) Bond wires are nearly co-linear in a normal bonding configuration resulting in low EM coupling. (b) A multi-tapped device may experience enhanced EM coupling and SAW regeneration due to the reactance of long bond wires.

CHAPTER 5

ATDE ELECTRONICS

5.1 SAW Electronics

In order to compensate for the insertion loss of the output IDT arising from the low impedance load, it is desirable to amplify the signal and electrically match the input apodized IDT to obtain the minimum insertion loss possible. The input and output SAW buffer amplifiers are designed to perform this task.

There are a number of problems to be addressed using such a configuration: (1) The input admittance of the apodized IDT has an unusual resonance at the band edges [Figure 5.1(a)] and as such, obtaining an electrical match at these frequencies is very difficult, if not impossible to achieve; (2) The high Q of the LC tank circuit formed by the transducer capacitance and the resonant matching inductor will cause significant peaking [Figure 5.2(a)] in the frequency response ; (3) Due to the matched configuration of the input IDT, any SAWs reflected from the output IDTs will be re-reflected as triple transit signals.

Unfortunately, not much can be done to match the admittance resonance at the band edges and the only recourse the designer has to mitigate this effect is to make the passband of the delay section wider than the pulse shaping filter bandwidth, or use a low impedance driver with no inductive tuning. Using a wider bandwidth device results in matching distortions due to the admittance resonance to fall outside of the channel bandwidth [Figure 5.1(b)].

The distortion caused by the high Q of the LC tank circuit (formed

by the inductive matching and transducer capacitance) can be reduced by the addition of a damping resistance, or again, use of a low impedance driver with no inductive tuning. If inductive matching is used, the bandwidth of such a matching network must be much wider than the SAW filter bandwidth so that the overall frequency response is not significantly altered [Figure 5.2(b)]. This was discussed in Section 4.4.1 and the results are presented in Section 4.4.2.

The SAW buffer amplifier circuits presented in the following sections are designed to have extremely low amplitude variations in the pass band and low distortion. This is necessary in order to achieve optimum performance from the complete equalizer circuit.

Section 5.1.1 presents the design and experimental results of the input buffer amplifier circuit. Section 5.1.2 presents the design and experimental results of the output buffer amplifier. Subsection 5.1.3 lists some precautions that must be observed when manufacturing the PC board which contains the SAW buffer circuits.

SII SMITH
 REF 1 Y0.76 -j 74.00 ohms
 START 50.00000 MHz
 STOP 80.00000 MHz
 70.00000 MHz

(a)

(b)

Figure 5.1 (a) Input IDT has an admittance resonance at the band edges resulting in an electrical mismatch. (b) Use of a passband wider than the pulse shaping bandwidth minimizes mismatch effects.

S21 LOG MAG
 REF -10 dB
 5 dB/
 MRK 1 -10.29 dB 70.00000 MHz
 START 50.00000 MHz
 STOP 90.00000 MHz

(a)

(b)

Figure 5.2 (a) Inductive matching results in peaking of the frequency response (b) A damping resistor lowers the peaking.

5.1.1 Input Amplifier

The purpose of the input amplifier is to provide an input impedance match to a 75Ω system and also to provide the required output impedance to drive the apodized IDT. A two stage device [Figure 5.3] fulfills the requirements. Q1 and Q2 are arranged in a Darlington configuration to provide a high input impedance and a low output impedance. A 75Ω resistor (R1) provides the required input impedance. Resistors R6, R7, R8, and capacitors C5 and C6 provide the damping and compensation, respectively, in order to obtain a flat frequency response to within ± 0.1 dB and a flat group delay in the band 50-90 MHz.

Figure 5.3 Input amplifier design.

A SUPER-COMPACT (Copyright Compact Software Inc.) model was constructed [Figure 5.4] to fit the measured frequency response and output impedance. The model was then analysed for stability. According to the model, the amplifier is unconditionally stable driving a capacitive load of less than 50pf. Figure 5.5 shows the modelled and experimental frequency response. A photograph of the wideband frequency response is shown in

Figure 5.6. The group delay of the input amplifier is shown in Figure 5.7 and the modelled vs. experimental output impedance is shown in Figure 5.8

Figure 5.4 SUPER-COMPACT model of the input amplifier. The numbers 2-58 indicate the node numbers used in the SUPER-COMPACT network description. The reader is referenced to the program listing in Appendix A-2.1 for the parasitic element values.

The input driver was originally intended to directly drive the input IDT unmatched and as such, a low output impedance is preferred since it minimizes the effect of transducer capacitance and maximizes the current drive to the filter. However, as mentioned in Section 5.1, a matched input

is desirable in order to reduce the overall insertion loss and hence the low impedance is not absolutely required. It was decided however to leave the buffer design with a low output impedance in order to allow maximum flexibility during the testing and analysis stage of the thesis where the use of a low output impedance buffer might be required. The advantage of a low output impedance design is that a series resistor can be added to the buffer output in order to obtain any desired output impedance level.

Figure 5.5 A comparison of the experimental S21 frequency response of the input amplifier with the modelled response over the 50-90 MHz band.

Figure 5.6 Photograph of the measured S21 wideband frequency response of the input buffer amplifier over the band 0.5-1300 MHz.

S21 DELAY
 REF 0 nsec START 50.00000 MHz
 5 nsec / STOP 90.00000 MHz
 MRK 1 2.42 nsec. 70.00000 MHz

Figure 5.7 The measured S21 group delay response of the input amplifier over the 50-90 MHz band.

Figure 5.8 A comparison of the experimental output impedance of the input amplifier with the modelled response over the 50-90 MHz band.

5.1.2 Output Amplifier

As mentioned in Section 5.1 it is desirable to have a low input impedance buffer on the tap outputs in order to prevent acoustic regenerative effects from corrupting adjacent tap responses. The output buffer must also have a very flat frequency response and introduce very little delay distortion since it is placed directly in the IF path of the receiver. It must also be relatively simple, compact, and have a printed circuit board layout as narrow as possible in order to allow seven of them to be placed next to each other. Considering this, the use of inductive elements is discouraged. Since power consumption should also be kept to a minimum, a design goal is operation from ± 5 volt supply rails.

Originally, the use of a diode ring surface mount mixer was considered for the ATDE multiplier element. Tests revealed that the input impedance of a mixer used in a multiplying configuration changed by as much as $\pm 50\%$ depending on the attenuation level chosen. In order to provide sufficient isolation between the multiplier and the SAW output tap, a Darlington pair (Q2,Q3) was incorporated into the output buffer design shown in Figure 5.9. A common base input stage (Q1) provides a low input impedance. RC networks (R8,C5 and R11,C7) provide damping and compensation of the frequency response. A SUPER-COMPACT model was optimized [Figure 5.10] to fit both the experimental frequency response and input impedance. The model was then analysed for stability.

Results of the SUPER-COMPACT analysis indicate that the amplifier is unconditionally stable driving a load of 75Ω . Figures 5.11 through 5.13 show the modelled and measured results of the amplifier circuit.

Figure 5.9 Output amplifier design.

Figure 5.10 SUPER-COMPACT model of the output amplifier. The numbers 1-60 indicate the node numbers used in the SUPER-COMPACT network description. The reader is referenced to the program listing in Appendix A-2.2 for the parasitic element values.

(a)

(b)

Figure 5.11 A comparison of the experimental S21 frequency response of the output amplifier and the modelled response. (a) 0.1-200 MHz. (b) 0.5-1300 MHz.

Figure 5.12 Photograph showing the measured S21 frequency response of the output amplifier over 0.5-1300 MHz.

S21 DELAY
 REF 0 nsec START .1000000 MHz
 .5 nsec / STOP 200.0000 MHz
 MRK 1 .06 nsec. 70.06500 MHz

Figure 5.13 The measured S21 group delay response of the output amplifier over the 0.1-200 MHz band.

Figure 5.14 A comparison of the measured and modelled input impedance of the output amplifier over the 0.1-200 MHz band. A 75Ω resistor was placed on the amplifier input to obtain a match with the 75Ω measurement system. The measured results here indicate an input impedance of $2.1 + j11.3 \Omega$ when the matching resistor is ignored.

5.1.3 PCB Layout Considerations

Most of the precautions taken to eliminate spurious EM feedthrough problems associated with the SAW device are listed in Section 3.2.8: Separate grounds are used for the input and output IDTs, the case is grounded, and grounded shielding is placed between the pins of the device package. The complete ATDE contains much RF circuitry and therefore some additional precautions are necessary.

Due to the unavailability of surface mount multiplier elements, the multiplier section of the ATDE takes up a sizeable portion of the overall PC board area. For this reason it was decided to break the complete circuit down into three sub-circuits with interconnections made via small lengths of flexible coaxial cable [Figure 5.15]. The three sections are: (1) an input section containing the input amplifier and matching circuit, the SAW delay line and seven output amplifiers; (2) a multiplying section containing twelve multiplier circuits; (3) an output section containing the I-axis and Q-axis summing amplifiers, a 3 dB quadrature hybrid and twelve manually controlled voltage sources (op-amp followers) to provide the multiplier control voltages. All three boards are double sided copper clad. Most of the tracks have been put on the top side leaving the bottom as a continuous ground plane. For the input board, the special precaution of separating both the top and bottom ground planes is necessary since isolated input and output ground planes are required to keep EM feedthrough to an acceptable level.

Figure 5.15 Photograph of PC board assembly (a) Top view showing surface mount electronics.

Figure 5.15 (continued) PC board assembly (b) Bottom view showing interconnect cables and multipliers.

5.2 Adaptive Filter Electronics

The following sections describe the individual circuits designed to perform the multiplying, summing and quadrature addition functions. In addition to this, the last section contains a description of the alignment procedure required in order to ensure that the I and Q multipliers are orthogonal.

5.2.1 Multiplier

The requirements for the multiplier are: (1) that it be extremely linear and exhibit less than 1/2% non-linearity at signal levels in the vicinity of -15dBm; (2) that it be capable of an attenuation range of greater than 35 dB over the band 50-90 MHz in order to effectively implement the infinite attenuation criterion; (3) that the attenuation be controlled by a bipolar voltage with maximum attenuation for 0 volts and a minimum attenuation for either ± 5 volts; (4) that the relative insertion loss be the same for identical positive or negative control voltage magnitudes even though the phase of the output signal is reversed; (5) that the amplitude response be flat to within 1dB over the band 50-90 MHz.

The results of three different designs are presented in the following sections.

5.2.1.1 Diode Ring Mixer

The original idea was to use a surface mount mixer as a current controlled attenuator in order to implement the multiplying element [Figure 5.16]. Surface mount mixers are economically viable and their size makes them an attractive choice in an environment where PCB real estate is at a premium.

Figure 5.16 Diode ring mixer used as a bi-phase multiplier.

This design proved to have a very flat frequency response as illustrated in Figure 5.17. However, a number of undesirable characteristics were also observed. It was discovered that as the control current was changed to maximize or minimize the attenuation through the mixer, the input impedance changed from a magnitude of $59\ \Omega$ to $85\ \Omega$. This was not considered a serious problem because the output buffer design provided enough isolation so that the frequency response of the SAW filter would not be altered. However, the linearity of the mixer proved to be much worse than expected at intermediate attenuation levels. Figure 5.18 shows the mixer spectrum output in the passband for a two-tone input test.

Third order intermodulation products are at a level of -25 dBc. The resulting third-order intermodulation intercept point (IP_{o3}) is only -28 dBm. The use a high level mixer improved the 3rd-order intercept by about 10 dB ($IP_{o3} = -18$ dBm) at intermediate attenuation levels but the performance was still not acceptable. Since a maximum non-linearity of 1/2 % (intermodulation products approximately -60dBc) is required for reasonable performance in a 256 QAM system, mixers are clearly not a viable alternative.

Figure 5.17 Transmission (S21) frequency response of the diode ring mixer in the band 0.1 - 200 MHz for various values of bias current.

Two tone inputs @ -13 dBm each. $\text{IMR}_3 = 23 \text{ dB}$.

Medium attenuation setting

Figure 5.18 Output spectrum in the passband of the diode ring mixer (bias current at 0.4 mA) when excited by two equal level signals at 66 MHz and 74 MHz (each at -13dBm). Both third-order and fifth-order intermodulation products are produced.

5.2.1.2 Biased PIN Diodes

The next design which was investigated used biased PIN diodes as shown in Figure 5.19.

Figure 5.19 Schematic diagram of multiplier using biased PIN diodes.

The tapped signals are buffered through an emitter follower and then coupled through a transformer to produce a differential signal to a pair of PIN diodes. The multiplication is accomplished with biased PIN diodes acting as variable resistors so as to give both positive and negative tap weights. The control voltage will adjust the currents flowing through the PIN diodes D1 and D2 such that when V_c goes positive then the resistance of D1 increases while the resistance of D2 decreases. If V_c goes negative, the opposite is true and this causes a phase reversal in the output. If V_c is zero volts, there will be no output signal. PIN diodes have a very high turn-on voltage and thus produce very little distortion. Experimental results show that this design has a very high 3rd-order intermodulation

ratio that fully satisfies the linearity requirements [Figure 5.20] of the ATDE. However, the frequency response [Figure 5.21] is not very flat in the passband and exhibits a ripple of approximately 4dB peak to peak. Also, the frequency response changes with different attenuation levels and thus cannot be compensated for in the SAW filter or in the external electronics. Another disadvantage is that the design requires specially wound transformers and hand picked diodes in order to achieve a large attenuation range since mismatched diodes with different characteristics result in unequal cancellation of the out-of-phase signals.

Two tone inputs @ -13 dBm each. $IMR_3 = 49.9 \text{ dB}$.

Minimum attenuation setting

Figure 5.20 Output spectrum in the passband of the PIN diode multiplier (bias current at 15 mA) when excited by two equal level signals at 135 MHz and 145 MHz (each at -13dBm). Third-order intermodulation products which are 50 dB below the main signal are present in the output.

Figure 5.21 Photographs showing the frequency response of the biased PIN diode multiplier configuration. (a) Minimum attenuation has a 3 dB amplitude ripple. (b) Amplitude characteristics change by a few dB at 17 dB relative attenuation. (c) Maximum achievable attenuation.

5.2.1.3 MC12002 High Speed Multiplier

The MC12002 high speed multiplier is a double balanced analog mixer which includes an input amplifier feeding the mixer carrier port and a temperature compensated bias regulator. The input circuits for both the amplifier and mixer are differential amplifier circuits. The on-chip regulator provides all of the required biasing and hence minimizes the number of external components. The circuit schematic of the device is shown in Figure 5.22. The maximum differential voltages that can be applied to the LO input (pins 2 & 3) and the mixer input (pins 8 & 9) are 30 mV p-p and 300 mV p-p respectively.

Originally, the use of an MC1496 multiplier was considered for the multiplying element. However this device requires more external components for biasing since on-chip circuits are not provided. It also exhibits a poorer high frequency response than the MC12002. For these reasons, the MC12002 was the preferred choice.

The final design for the multiplier is shown in Figure 5.23.

Figure 5.22 MC12002 internal schematic.

Figure 5.23 Multiplier circuit employing the MC12002.

In order to achieve good linearity, the conventional LO and mixer ports have been interchanged. This is because the LO input has internal buffering which will cause clipping when large signal levels are presented, resulting in poor linearity. A bipolar control signal is required, therefore it is necessary to implement a special external biasing network [Figure 5.24(a)] such that the correct common mode voltages are applied at pins 2 and 3. Voltage superposition at the nodes 3 and 4 for various applied voltages allows the determination of both the differential and common mode voltages simultaneously. It is important to choose the correct resistor and node voltages so that the common mode voltages do not cut-off or saturate the internal multiplying transistors when either the minimum or maximum control voltage (± 5 volts) is applied.

In order to facilitate the selection of resistor values, a special spread sheet program was created on a Macintosh[®] personal computer to calculate all the node voltages and resistor currents for an arbitrary input

selection. The results are shown in Figure 5.24 (b). The resistors R_3 and R_4 and the node voltages V_{ref} and V_{bias} are chosen so that a differential voltage $DV = \pm 30$ mV is presented across pins 2 and 3 of the MC12002 when a control voltage $V_c = \pm 5$ V is applied. Also, $DV = 0$ V when $V_c = 0$ V.

(a)

MC12002

	2	3	4	5	6	7
1	VREF	2.4	V	V1	2.00696	V
2	VC	-5	V	V2	2.03854	V
3	VBIAS	5	V	DV	-0.0316	V
4	VINT	2.4	V			
5	RES3	62	Ω			
6	RES4	2446	Ω	I1	-1.4896	ma
7	RES5	1000	Ω	I2	0.58735	ma
8	RES6	1000	Ω	I3	32.3704	ma
9				I4	-0.1478	ma
10						
11	RES2	5095.8333	Ω	I5	-0.393	ma
12	RES1	4703.8462	Ω	I6	-0.3615	ma
13						

(b)

Figure 5.24 (a) Resistor network employed to transform floating differential voltage to a single-ended, ground referenced control voltage. (b) Results of Macintosh PC spreadsheet for $V_c = -5$ V.

In order to obtain a flat frequency response at all attenuation levels while still maintaining good linearity, it was found necessary to add a single stage transistor follower to provide isolation between the multiplier output and the summing amplifier. It should be noted that the input impedance of the mixer port is constant at about 700 ohms at all attenuation levels and hence the Darlington pair in the output buffer of section 5.1.2 is no longer required. This should reduce parts count in future designs. The frequency response for different attenuation levels is shown in Figure 5.25. The measurements of Figure 2.5 were obtained using an input power of -12 dBm. The worst case non-linearity is less than 0.5% [Figure 5.26]. The group delay deviation is less than 0.1 nsec over the 50-90 MHz band as shown in Figure 5.27. The required power consumption of the circuit is 0.35W (50ma @ +5V, 20ma @ -5V). This is an important factor to consider since twelve multipliers are required in the complete ATDE. The attenuation characteristics are listed below in Table 5.1.

Table 5.1

MC12002 multiplier circuit attenuation and gain slope for various control voltages.

Voltage (volts)	Gain at 70 MHz (dB)	Gain Slope (50 - 90 MHz) (dB)
±5	+0.3	0.00
±2	-4.6	-0.31
±1	-10.8	-0.50
±0.5	-16.1	-0.47
±0.2	-23.1	-0.09
0	-38.5	+6

(a)

(b)

Figure 5.25 (a) Frequency response of the MC12002 based multiplier circuit for various control voltages. (b) A smaller amplitude scale illustrates the relative amplitude slope. The above measurements exhibited a worst-case non-linearity of 0.5%. The input level is -12 dBm.

Figure 5.26 Measured results of the MC12002 multiplier non-linearity obtained from a microwave link analyzer. (a) Photograph showing 1/2% non-linearity at 0 dBm attenuation (worst case). (b) Photograph of the input signal (driving a 75 Ω load). The total power input is the sum of the powers contained in the high frequency and low frequency components.

S21 DELAY
 REF 2.1 nsec
 .5 nsec /

START 30.00000 MHz
 STOP 110.0000 MHz

Figure 5.27 Measured S21 group delay response of the MC12002 multiplier circuit over the band 30-100 MHz.

5.2.2 Summing Amplifier

The summing amplifier [Figure 5.28] employs a common base input stage to provide a low input impedance and an emitter follower output stage to provide buffering. The amplifier's measured frequency response is shown in Figure 5.29 and the group delay is shown in Figure 5.30. The input impedance [Figure 5.31] illustrates that the circuit has an input impedance of 77.6Ω which means that the summing node has an input impedance of approximately 2.6Ω . This means that for two signals fed into the summing node, the interaction between them will be attenuated by approximately 30 dB.

Figure 5.28 Summing amplifier circuit schematic

S21 LOG MAG
 REF -1.4 dB
 .1 dB/
 START 50.00000 MHz
 STOP 90.00000 MHz
 MRK 1 -1.42 dB 70.00000 MHz

(a)

X 0.5 1300 MHz

Y 10 dB div

(b)

Figure 5.29 The measured S21 transmission response of the summing amplifier (a) over the 50-90 MHz band; (b) over the 0.5-1300MHz band.

S21 DELAY

REF 1.15 nsec

START 40.00000 MHz

150

.5 nsec /

STOP 100.0000 MHz

MRK 1 1.3 nsec. 70.00000 MHz

Figure 5.30 The measured S21 group delay response of the summing amplifier over the band 40-100 MHz.

S11 SMITH

START 40.00000 MHz

REF 1 79.55 +j 2.31 ohms

STOP 100.00000 MHz

MRK 1 70.00000 MHz

Figure 5.31 The measured input impedance (S11) of the summing amplifier over the band 40-100 MHz indicates a close match to the 75Ω measurement system and almost zero reactance.

5.2.3 Quadrature Hybrid

The quadrature hybrid used is shown in Figure 5.32. For convenience, it is a design which is used in an existing down converter module. The hybrid has two quadrature inputs and two in-phase inputs, however only one of each is used. The input and output impedances are 75Ω . The measured phase and frequency responses are shown in Figure 5.33. The performance of the quadrature hybrid is summarized in Table 5.2.

Table 5.2

Measured performance of the quadrature hybrid.

Bandpass: 50 MHz-90 MHz

Insertion Loss: -3.5 dB at 70 MHz

Amplitude Ripple: +1 dB over bandpass, -90° input

-1.5 dB over bandpass, 0° input

Phase Ripple: 3° max. over the bandpass, both inputs

Figure 5.32 Quadrature hybrid circuit schematic

S21 PHASE

REF 79 °

45 ° /

START 55.00000 MHz

STOP 85.00000 MHz

MRK 1 78.79° 70.00000 MHz

In port 3
Out 78.79 degreesIn port 2
Out -12.08 degrees

(a)

S21 LOG MAG

REF 0 dB

1 dB/

START 50.00000 MHz

STOP 90.00000 MHz

MRK 1 -3.56 dB 70.00000 MHz

In 2

In 3

(b)

Figure 5.33 The measured S21 transmission response of the quadrature hybrid (a) phase response over 55-85 MHz; (b) amplitude response over 50-90 MHz.

5.3 ATDE Alignment

In order to ensure that the I and Q taps of the transversal equalizer are orthogonal, it is necessary to align the equalizer multipliers with respect to one another. One required alignment is the introduction of a delay Δt into the I_0 path which is equivalent to the delay introduced by the multipliers. This may be accomplished by the addition of a length of coaxial cable or an all-pass delay network [Figure 5.34] between the centre tap output buffer and the summing amplifier. An all-pass delay network will introduce a flat group delay into the signal path. The bandwidth of the delay network is inversely proportional to the delay time, but can easily be designed wide enough not to perturb the group delay response of the system for most practical values of delay. A length of coaxial cable was chosen to satisfy the I_0 delay requirement since it was the easiest method available to implement a flat delay with small attenuation. A 4.5 ns delay is required between the I_0 tap and the input to the I summing amplifier. A length of coaxial cable approximately 85 cm. long was inserted to satisfy the delay requirement.

In addition to this adjustment, it is necessary to ensure that the final I and Q outputs are orthogonal to each other. If the taps are not perfectly orthogonal, then the I tap multiplication will affect the Q tap response and vice versa as illustrated in Figure 5.35. The I and Q signals will not be orthogonal if slight differences exist in the path delays, or if the quadrature hybrid does not exactly sum the two signals in phase quadrature. This problem may be overcome by inserting a delay in the I or Q quadrature combiner input to compensate for this effect. A 14 cm. length of coaxial cable was inserted in the I path to satisfy this requirement. An overall circuit schematic in Appendix 3 gives the

additional lengths of coaxial cable inserted in order to correctly align the circuit.

In a closed loop system extremely accurate alignment of delays is not as important as in the open loop system presented in this thesis. One attractive feature of an equalizer in a closed loop operation is that small tap errors and residual RF cross talk between amplifiers and multiplier circuits will not affect system performance. This is because the equalizer will interpret these errors as ISI and adaptively correct itself.

Figure 5.34 An all pass delay network will introduce a flat group delay into the signal path. (a) Circuit schematic. (b) frequency response.

Figure 5-35 Alignment of the I and Q taps is necessary so that the I tap setting does not introduce any orthogonal multiplication and vice versa. In the diagram, ϕ is the phase alignment error introduced by $\omega_c \Delta\tau$ where $\Delta\tau$ is the small delay difference between equal multiplier paths.

CHAPTER 6

TESTING AND EXPERIMENTAL RESULTS

6.1 SAW Delay Line Experimental Results

The measurements presented in this section deal with the performance of the delay line and buffer electronics. The performance of the complete equalizer circuit is presented in Section 6.2.

The measurements are performed on the system shown in Figure 6.1. The test setup consists of an HP 3577A two-port network analyzer (5Hz-200MHz), an HP7470A plotter, an HP9000/216 console terminal and keyboard, and an HP89201M dual flexible disk drive unit. The ports of the network analyzer are terminated in 75 Ω . Experimental data may be uploaded to an IBM mainframe via a telephone modem connected to the computer subsystem.

Two different SAW filter designs were made. Essentially the only difference between the designs is the bandwidth, which in this case is primarily determined by the apodized input IDT. The output tap geometries are the same for both designs. The first design (SAW03F) has an α of 12%, a Nyquist bandwidth of 22.585MHz and 208 split-fingers in the apodized transducer in order to achieve the desired ripple and sidelobe response. The second design (SAW03L) is a wideband filter with a flat insertion loss passband of 26 MHz. This corresponds to a Nyquist bandwidth of 34.667 MHz and an α of 25%. It only requires 150 split-fingers in order to achieve the same ripple and sidelobe performance as SAW03F. A wide bandwidth device (discussed in Section 5.1) is required in order to mitigate the problem of matching to the admittance

resonance in an apodized transducer.

Figure 6.1 Block diagram of the test set-up. The HP 3577A Network Analyzer (75 Ω system) is under computer control.

A photograph of the SAW03L device is shown in Figure 6.2. Figure 6.3 shows the predicted frequency response versus the experimental results for the SAW03F and SAW03L designs.

Section 6.1.1 presents the preliminary group delay and insertion loss results obtained from SAW devices inserted in a prototype test jig. It also presents reflection measurements obtained using low impedance electronics on the tap outputs.

Section 6.1.2 deals with spurious tap reflections and presents a computer model used to establish the magnitudes of the EM induced and regenerated SAWs.

Section 6.1.3 presents the measurements of spurious signals caused by EM feedthrough and shows the improvement obtained by using a glass carrier with separate ground planes.

Section 6.1.4 deals with spurious MEL reflections.

Figure 6.2 Photograph of a finished 7-tap SAW delay line on a glass carrier. The apodized transducer is on the right.

(a)

(b)

Figure 6.3 Predicted frequency response of an ideal filter compared to the experimental result. (a) SAW03F (b) SAW03L.

6.1.1 Initial Group Delay and Insertion Loss Results

The first devices received were bonded in a 20 pin package as shown in Figure 6.4. A SAW03F device was inserted into a test jig [Figure 6.5] and tested for insertion loss and group delay. The results of this test are shown in Figure 6.6.

As can be seen from the frequency response of Figure 6.6, a spurious ripple is present in the passband at approximately 68 MHz. An investigation of the problem by computer modelling revealed that the spurious signals were being caused by three effects: EM coupling effects, regenerative effects caused by a finite load (i.e., lead inductance from the output tap to the buffer input), and electrode tap reflections. These are discussed in Sections 6.1.2 through 6.1.4.

Figure 6.4 Initial bonding pattern for the SAW03F device in a 20 pin package.

(a)

(b)

Figure 6.5 Prototype test jig and buffer electronics. (a) Test jig block diagram (b) Photograph showing prototype test jig and a packaged SAW delay line.

S21 LOG MAG
 REF -27 dB
 5 dB/
 START 50.00000 MHz
 STOP 90.00000 MHz
 MRK 1 -28.15 dB 70.00000 MHz

(a)

S21 LOG MAG
 REF -27 dB
 1 dB/
 START 50.00000 MHz
 STOP 90.00000 MHz
 MRK 1 -28.13 dB 70.00000 MHz

(b)

Insertion Gain
 at 70 MHz

Tap 1 -28.13 dB
 Tap 2 -28.56 dB
 Tap 3 -28.61 dB
 Tap 4 -28.66 dB
 Tap 5 -28.82 dB
 Tap 6 -28.95 dB
 Tap 7 -29.09 dB

Figure 6.6 SAW03F initial measured results. (a) S21 amplitude response for tap 1 and tap 7. (b) Insertion loss of taps 1 through 7 ((c),(d) continued on next page)

S21 DELAY
 REF 1223 nsec
 44.3 nsec /
 MARK 1 1223.38 nsec. 70.00000 MHz
 START 55.00000 MHz
 STOP 85.00000 MHz

S21 LOG MAG
 REF -35 dB
 10 dB
 START -100.0 nS
 STOP 1.9 μS

Figure 6.6 (continued) SAW03F initial measured results. (c) Group delay showing response for each tap exactly 44.3 ns apart (d) Tap 1 impulse response shows upper sidelobe degradation. The group delay measurement of (c) has been gated to remove a large ripple caused by EM feedthrough.

As mentioned in Chapters 3,4 and 5 a low input impedance amplifier will reduce regenerated SAWs being generated by the pickup taps. Figures 6.7 and 6.8 illustrate this statement. Figure 6.7 shows the effect of loading the output IDT of a single output SAW device with a small resistance. Notice that the regenerated signal is reduced by the use of a low impedance load and is attenuated to below the noise floor. A similar effect can be seen for a 7-tap device [Figure 6.8] loaded with a 150 Ω load or a 2 Ω load. A 10-15 dB improvement in the reflection can be seen for the 2 Ω load.

Figure 6.7 Effect of lowering the output IDT load resistance in a single input/single output IDT SAW filter. (a) S11 time domain measurement with 200 Ω load impedance on the output IDT. (b) Same as (a) except the load impedance is 0 Ω .

Figure 6.8 Time domain response illustrates spurious regenerated SAWs caused by a finite load impedance. SAW03F (a) S11 time domain measurement with 150 Ω load impedance (b) S11 with 2 Ω impedance presented by common base amplifier reduces the reflection by approximately 10 dB.

The results of Figures 6.7 and 6.8 verify that taps reflections due to electrical loading can be reduced by at least a factor of 10 using low input impedance amplifiers.

6.1.2 Tap Reflections

The first hypothesis concerning the peak in the spurious ripple at 68 MHz was that it was being caused by regeneration. The time domain measurement of Figure 6.6(d) shows that some sort of spurious signal is degrading the upper sidelobe response. In order to get a better understanding of the problem, the delay line was rebonded as illustrated in Figure 6.9(a). The resultant time domain response [Figure 6.9(b)] showed some sort of regenerated SAW arriving after the main signal output. A special feed arrangement (in tap 1 out tap 4) was necessary in order that this effect could be isolated in time.

(a)

(b)

Figure 6.9 Regenerated SAWs arrive after the main signal. (a) Bonding arrangement. (b) S21 time domain response showing both EM-induced and regenerated SAWs.

Under the assumption that these were in fact regenerated SAWs being caused by a finite load impedance, a general model was developed to quantify the magnitudes of the regenerated components. The model developed [Figure 6.10] includes both EM and regenerated effects. Refer to Appendix A-1.4 for the FORTRAN code for this model (Program name: CURVE). Using a curve fitting optimization routine, the coefficients of the model were determined to fit the experimental frequency response of SAW03L.

The results of CURVE [Figure 6.11] show that the fast ripple is caused by the regenerated SAWs and that the gradual rolloff is caused by EM coupling. The exact magnitudes of the computed coefficients are listed in Table 6.1. The results indicate that EM-induced SAWs are about an order of magnitude larger than the regenerated spurious SAWs.

Figure 6.10 Computer model (CURVE) used to quantify magnitudes of EM-induced and regenerated SAWs.

SPURIOUS REFLECTION ANALYSIS

Figure 6.11 Modelled vs. experimental results using the analysis model of Figure 6.10.

Table 6.1

Coefficient magnitudes obtained from CURVE optimization run illustrate that regenerated SAWs are an order of magnitude smaller than the EM induced components.

Coefficient Number i	EM Induced SAW Coefficients A_i	Regenerated SAW Coefficients B_i	EM Phase (degrees)
1	$+5.5 \times 10^{-2}$	-4.4×10^{-3}	55
2	-1.5×10^{-2}	-7.5×10^{-3}	55
3	$+1.8 \times 10^{-2}$	-1.0×10^{-2}	55
4	-3.65×10^{-3}	-7.8×10^{-3}	55
5	$+9.6 \times 10^{-3}$	-4.8×10^{-3}	55
6	-1.4×10^{-2}	-7.2×10^{-3}	55

6.1.3 EM Feedthrough Reduction Using a Glass Carrier

As mentioned in Section 3.2.8, EM feedthrough can potentially be a very serious problem if proper methods are not taken to reduce it to an acceptable level. When it is a problem, an easy solution is to use a balun to reduce the net EM field at the receiving IDT to zero. Unfortunately, the self inductance of a balun will resonate with the IDT and will cause undesirable regenerated SAWs. Hence another method must be devised to reduce EM feedthrough. Figure 6.12 shows a time domain response of SAW03F bonded as in Figure 6.4. The EM feedthrough suppression was worst for tap 1 and increased for taps 2 through 7. The EM feedthrough is only -43 dBc for tap 1 using this bonding arrangement and a feedthrough of less than -60 dBc is required. A time domain plot of EM-induced SAWs is shown in Figure 6.13.

Figure 6.12 S21 time domain response of SAW03F showing poor EM feedthrough rejection using the conventional bonding scheme of Figure 6.4

Figure 6.13 EM-induced SAWs arrive before the main signal (a) Bonding pattern. (b) S21 time domain response.

There are two possible factors responsible for generation of EM-induced SAWs: (1) EM coupling may be caused by the close proximity of the adjacent taps; (2) EM coupling may be caused by the close proximity of long bonding wires which act like miniature antennas inside the packaging structure. This would explain why the EM feedthrough suppression gets progressively better from tap 1 to tap 7. It was decided to investigate case (2) by rebonding a number of SAW03L devices to see the effect of moving the bond wires. The results illustrated in Figures 6.14 and 6.15 (before and after rebonding respectively) show that moving the input and output bond wires so that they are colinear reduces the EM feedthrough by approximately 18 dB. This experiment verified that a glass carrier was necessary to orient the output tap lines away from the input and to reduce EM coupling through the air from one bond wire to the next.

(a)

(b)

(c)

Figure 6.14 SAW03L EM feedthrough rejection for tap 1 (worst case). (a) Bonding pattern. (b) S21 transmission frequency response. (c) S21 time domain response showing poor EM feedthrough rejection of only 43 dB.

(a)

S21 LOG MAG
REF -33.25 dB
10 dB/
MPK 1 -75 dB 41.50000 MHz
START 40.00000 MHz
STOP 100.0000 MHz

(b)

S21 LOG MAG
REF -33.25 dB
10 dB/
MPK 1 -94.76 dB 0 nS (0 cm)
START -100.0 nS
STOP 0.9 uS

(c)

Figure 6.15 SAW03L EM feedthrough rejection for tap 1 (worst case). (a) New bonding pattern. (b) S21 transmission frequency response. (c) S21 time domain response showing dramatically improved EM feedthrough rejection of 60 dB.

Two different glass carrier designs illustrated in Figure 6.16 were fabricated. A rubylith pattern was cut and reduced, and then $2000\text{\AA}$ of aluminum were deposited on a glass slide using the reduced phototools. The entire assembly was then placed inside a 24 pin package. Initial tests showed that the carrier of Figure 6.16(a) had a slightly better EM feedthrough rejection for tap 1. It was also discovered that separating the input and output ground planes on the glass carrier [Figure 6.17] reduced the EM feedthrough by an additional 10 dB. Using this construction technique, the return RF currents for the input and output IDT are forced to exit via two different paths spaced as far apart as possible. The final results show that using a glass carrier with split ground planes inside a sealed package reduces the EM feedthrough for tap 1 (worst case) to less than -60 dBc [Figure 6.18].

Figure 6.19 shows the group delay and insertion loss measurements for a tapped delay line utilizing the improvements listed above.

Figure 6.16 Glass carrier mask patterns. (a) Fanout is symmetrical about the centre tap. (b) Fanout is sloped away from the input IDT.

Figure 6.17 Better EM suppression was obtained with a split ground plane carrier illustrated above.

Figure 6.18 SAW03L measured S21 time domain response shows an improved EM feedthrough suppression using a split ground glass carrier. As can be seen here for tap 1, better than 60 dB of EM feedthrough suppression is achieved.

3 OF/DE 3

Microl

S21 LOG MAG

REF -29 dB

1 dB/

MRK 1 -29.58 dB 70.00000 MHz

START 50.00000 MHz

STOP 90.00000 MHz

175

(a)

S21 DELAY

REF 1015 nsec

44.3 nsec /

MRK 1 1014.32 nsec. 70.00000 MHz

START 50.00000 MHz

STOP 90.00000 MHz

(b)

Figure 6.19 Measurements of the SAW03L tapped delay line utilizing split ground plane glass carrier of Fig. 6.17 show improved performance. (a) S21 insertion loss over the 50-90 MHz band for taps 1 through 7.(b) S21 group delay response over the 50-90 MHz band. This measurement was not gated since the EM feedthrough was sufficiently suppressed.

In addition to the split ground plane carrier mentioned above, the installation of a series tuning inductor will result in a reduced EM feedthrough at high frequencies. However, the real advantage is that the passband insertion loss is reduced with a series inductor, hence effectively reducing the EM feedthrough by the same amount. Measured results of a SAW03F device in the prototype jig of Figure 6.5 are listed in Table 6.2.

Table 6.2

Additional EM feedthrough rejection is achieved with a series tuning inductor.

	<u>No inductor</u>	<u>With Series tuning Inductor</u>
Passband Insertion Gain	-34dB	-29dB
Feedthrough 0-40 MHz	-75dB	-80dB
Feedthrough at 100 MHz	-57dB	-58dB
Feedthrough at 200 MHz	-38dB	-50dB

It can be seen that with a series tuning inductor, a 5 dB improvement in EM feedthrough is obtained.

The insertion loss of the device can also be reduced by decreasing the tuning resistance in series with the matching inductor. Unfortunately, as mentioned in Section 4.4, this arrangement will cause peaking in the passband. Measured results of a SAW03F device in the prototype jig of Figure 6.5 are listed in Table 6.3

Table 6.3

Reducing the series tuning resistor reduces the insertion loss but increases the peaking in the passband.

R_s (Ω)	Insertion Loss (dB)	Passband Peaking (60-80 MHz) (dB)
10	-24.5	± 0.95
24	-28.6	± 0.75
82	-37.4	± 0.42
510	-52.8	± 0.30

The effect of a damping resistance on the insertion loss is shown in Figure 6.20.

REF LEVEL /DIV MARKER 70 000 000.000Hz
-24.500dB 2.000dB MAG(S21) -24.442dB

Figure 6.20 Reducing the series damping resistance decreases the relative insertion loss but causes peaking in the passband.

The analysis of Section 4.4.2 shows that a series resistor of 82 Ω will give the optimum insertion loss/passband distortion design.

6.1.4 Electrode MEL Reflections

Program CURVE successfully demonstrated that the frequency response of the first tap in the tapped SAW delay line can be accurately modelled by including spurious signals generated by EM induced and regenerated SAWs. However, after successfully reducing the inband feedthrough to -60 dB below the carrier, a small spurious ripple was still present. Based on this conclusion two more devices were bonded as illustrated in Figures 6.21 and 6.22 under the assumption that bonds to ground on all unused taps would effectively prevent any regenerated SAWs and at the same time reduce any EM coupling to a minimum. This would allow an uncontaminated response to be observed. Measurements shown in Figure 6.21 illustrate however that a residual spurious ripple of approximately 0.3 dB p-p is still present even when the EM feedthrough is gated out. The results shown in Figure 6.22 indicate that regenerated SAWs are no longer contaminating the time domain response and are therefore at a sufficiently low level.

Since the EM feedthrough in such a bonding scheme is far below the magnitude required to produce the residual ripple observed in the passband and is gated out anyways, and regeneration effects can be discounted because of the bonding, then, by the process of elimination electrode multi-reflections caused by MEL effects are responsible for the spurious ripple. Program REF2 (Appendix A-1.2) was constructed to model the effects of MEL reflections only. REF2 was designed to fit the gated frequency response of Figure 6.21. The gate function is used to prevent REF2 optimizing to any EM ripple. REF2 shows that when EM induced and regenerated SAWs are reduced to a minimum, the residual ripple can be modelled by MEL effects alone. This represents the fundamental limit to the amplitude ripple as discussed previously in

Section 3.2.1.2.

(a)

(b)

Figure 6.21 Bonding pattern of SAW03L device used to analyze magnitude of MEI reflections. (a) Bonding pattern. (b) Residual magnitude ripple caused by MEI effects.

(a)

(b)

Figure 6.22 Time domain response of SAW03F device shows very little sidelobe contamination by regenerated SAWs. (a) Bonding pattern. (b) S21 time domain response showing a clean impulse response.

Figure 6.23 Optimization of the MEL reflection coefficients in program REF2 illustrates that the frequency response of Fig. 6.21(b) can be modelled by MEL electrode interactions. Precautions have been taken to reduce EM-induced and regenerated SAWs to a minimum.

6.2 ATDE Experimental Results

The measurements reported in this section deal with the performance of the complete equalizer circuit. Measurements are performed on the system shown in Figure 6.24. The test set-up consists of the equipment of Figure 6.1 plus a channel simulator which is shown in Figure 6.25. The channel simulator simulates a fading channel response.

Section 6.2.1 deals with the performance of the equalizer when it is manually aligned with a selected set of tap weights. It presents the experimental results and compares the performance with the computed ideal tap response.

Section 6.2.2 deals with the equalization of a spectrum distorted by a simulated fade. The equalization of the distorted spectrum is performed by manually adjusting the equalizer tap weights to the values obtained by the computer algorithm of Section 2.3.4. The objective is to use the equalizer to remove the distortion and restore the signal to its undistorted shape. The maximum level of distortion that the equalizer can handle is then determined.

Figure 6.24 Block diagram of the test set-up used to measure the performance of the SAW based transversal equalizer.

(a)

REF LEVEL 10.000dB /DIV 10.000dB MARKER 70 065 000.000Hz
MAG (S21) 3.571dB

(b)

Figure 6.25 (a) Block diagram of the channel simulator. A 70 MHz IF signal is upconverted to 6 GHz where the fade is applied. A down converter module translates the distorted signal back to IF. (b) Typical response showing both a non-distorted passband and a deep fade.

6.2.1 Equalizer Tap Response

This section presents the results of the ATDE tap performance. Experimental results are uploaded to an IBM mainframe computer and compared with an ideal response. The computed response is obtained by cascading the SAW delay line response with an equalizer model that has a set of complex tap weights. The SAW delay line response is obtained using the algorithm presented in Section 4.2. The model used to compute the ideal response is illustrated in Figure 6.26.

Figure 6.26 Model used to determine the theoretical equalizer response for a selected set of complex tap weights.

Figure 6.27 shows the I-tap responses compared to theory. Only the feedforward tap responses are shown since the response for the feedback taps are identical, except for an additional delay of three symbol periods (approximately 132.9 nsec.). The Q-tap responses are also similar to the I-tap responses except that the nulls and peaks occur at frequencies half-way between the peaks and nulls of the I-tap responses. As shown in the photograph of Figure 5.15, the module interconnect cable lengths have been manually adjusted so that the I and Q taps are perfectly orthogonal.

The cable lengths are shown in the circuit schematic in Appendix 5. Figure 6.28 shows the computed and experimental response for an arbitrary set of complex coefficients.

As can be seen from Figures 6.27 and 6.28, the performance of the equalizer behaves as predicted. The noise floor of the ATDE circuit is higher than that predicted by the model and is due to (1) the summing of the noise powers in all the taps and (2) a limit to the isolation afforded by the SAW delay line. The summation of seven taps in the I channel and six in the Q channel will raise the noise floor approximately 13 times above the noise floor of a single tap. The noise floor of the twelve multipliers and the I_0 path are all summed in the final output, raising the noise floor approximately 11 dB above that of a single output.

(a)

(b)

-----Predicted

————Measured

Figure 6.27 Theoretical versus experimental performance of the SAW based transversal equalizer (a) $L_3 = +1$. (b) $L_3 = -1$.

(a)

(b)

-----Predicted

————Measured

Figure 6.27 (continued) Theoretical versus experimental performance of the SAW based transversal equalizer. (c) $I_2 = +1$. (d) $I_2 = -1$.

(a)

(b)

-----Predicted

————Measured

Figure 6.27 (continued) Theoretical versus experimental performance of the SAW based transversal equalizer. (a) $L_1 = +1$. (b) $L_1 = -1$.

(a)

(b)

-----Predicted

————Measured

Figure 6.28 Theoretical versus experimental performance of the SAW based transversal equalizer for an arbitrary set of complex tap weights. (a) $I_{+1}=1$, $Q_{+1}=1$, all other taps set to 0. (b) $I_{+2}=1$, $Q_{+3}=1$, all other taps set to zero.

6.2.2 Equalization of the Data Spectrum

This section presents the results of the ATDE circuit under fading conditions. A simulated fade is applied to the equalizer and then the tap weights are manually set to the coefficients computed using the zero-forcing algorithm of Section 2.3.4.

One of the first tests performed was the equalization of the channel response under non-fading conditions. This was performed by manually adjusting the tap coefficients to remove any non-ideal amplitude variations inherent in the SAW filter. The purpose of this test is to demonstrate that the spurious ripple that occurs at 68 MHz (discussed in Section 6.1) can be corrected to some extent. From the time domain measurements of section 6.1, most of the energy of this spurious signal appears at a time which is within the equalizer's correcting range of $\pm 3T$. The results of this test are illustrated in Figure 6.29, and prove that the SAW filter response can be equalized to remove both benign amplitude peaking due to inductive matching circuits and the spurious ripple at 68 MHz.

The equalization of a distorted data spectrum is performed in an open loop system where the adjustment of the tap weights is performed manually. It is important in an open loop system that the taps be aligned very accurately otherwise an amplitude error will exist in the equalized response. In a closed loop system, small errors in the tap responses are corrected adaptively and do not degrade the performance of the system. This is because the equalizer will interpret the tap error as ISI and adaptively correct or "right" itself. However, in an open loop system, this mode of performance is not possible and hence the alignment of the tap responses is critical to the performance evaluation. Sufficient precautions have been taken to ensure that the alignment is correct and this has been verified in section 6.2.1.

The results showing the equalization of two arbitrarily distorted spectrums are shown in Figures 6.30 and 6.31. Both the unequalized and equalized responses are plotted for comparison. A standard delay value of 6.3 ns was selected for the delayed path. The phase and amplitude of the delayed signal are adjusted to obtain two different distortion scenarios: (1) the fade notch is centred on the passband [Figure 6.30]; (2) the fade notch is offset from the passband centre so that a frequency dependent slope is introduced [Figure 6.31].

Figure 6.29 The equalizer can be pre-stressed to remove passband amplitude variations caused by matching circuits and second-order effects. Equalization is performed by manually adjusting the tap coefficients so that an ideal non-fading response (i.e., flat) is obtained.

Figure 6.30 Equalized response of a distorted spectrum for a fade notch at 70 MHz.

Figure 6.31 Equalized response of a distorted spectrum for a fade notch at 60 MHz.

Results of a series of tests indicate that the equalizer is capable of correcting notches with fade depths of up to 15 dB. At this fade depth, the amplitude ripple at the equalizer output is approximately 1 dB_{peak-to-peak}. For less severe fades, the equalized amplitude ripple is reduced. The results of a number of different tests showing the maximum equalized fade depths and notch frequencies versus the equalized spectrum response are tabulated in Table 6.4.

Table 6.4

Tabulated results showing the performance of the equalizer circuit under a variety of fading conditions.

Fading Parameters			Equalizer results
Notch Depth	Notch Position	Notch Delay	Passband ripple 50-90MHz
(dB)	(MHz)	(ns)	(dB peak-to-peak)
-1	68	6.3	0.45
4	70	6.3	0.43
4	76	6.3	0.56
5	70	6.3	0.15
8	70	6.3	0.72
12	70	6.3	0.85
15	70	6.3	1.04
4.5	64	6.3	0.65
7.5	64	6.3	0.55

CHAPTER 7

CONCLUSIONS AND RECOMMENDATIONS

7.1 Conclusions

This thesis contains a novel design approach to the problem of implementing a 70 MHz IF adaptive equalizer based on the use of SAW technology, and an experimental verification of the feasibility of such an approach.

The experimental results indicate that fading nulls of up to 15 dB can be effectively compensated for using this approach, to yield equalized amplitude variations of less than 1 dB peak-to-peak in the 50-90 MHz band. The in-band amplitude ripple decreases in magnitude for shallower fading notches.

It has been experimentally verified that a number of potentially serious second-order effects in the SAW delay line, such as acoustic multi-reflections, EM feedthrough, and tap interactions can be controlled or reduced using special construction techniques based on computer analysis of the problem. Two seven-tap SAW delay line designs were fabricated for use in the IF equalizer based on a computer analysis showing that tap reflections can be minimized by using different topographies. It has been shown that the use of grounded dummy fingers between taps is required to minimize electrode reflections. The dummy fingers must have widths slightly larger than the standard split finger width of $\lambda/8$ in order to compensate for symbol periods which are not integer divisors of the operating frequency.

It has also been shown that the SAW delay line response can be

equalized to remove benign amplitude peaking due to inductive matching circuits by pre-stressing the tap coefficients. Residual spurious ripple occurring within the equalizer bandwidth can also be reduced by pre-stressing the tap coefficients.

In addition, a number of circuits that operate with flat frequency characteristics up to 150 MHz have been developed in order to implement the circuit. These include a bi-phase IF multiplier circuit that exhibits good linearity and frequency characteristics, an input buffer amplifier, low input impedance output buffers, a summing amplifier, and a reactive matching circuit for the delay line.

7.2 Recommendations for Future Work

This thesis has illustrated a number of important results which verify the feasibility of using SAW devices in adaptive IF equalizers. A number of suggestions will be given here which should be considered by those wishing to pursue the development of SAW based equalizers for future radio products utilizing multilevel QAM digital modulation techniques. Given that tap reflections can be controlled to an acceptable level, one of the great advantages of utilizing SAW delay lines in IF ATDEs is that the number of output taps can be increased substantially with little increase in SAW device complexity or size. It also reduces the hardware requirements when compared to the baseband design. Hence, when considering equalization for systems beyond the 256 QAM level, SAW technology becomes an attractive alternative.

Suggestions for future work presented in this thesis include: (1) the integration of the SAW electronics to permit placement of input and output circuits directly adjacent to the SAW delay line; (2) hybrid integration of the components required to implement the twelve multiplier

circuits to reduce board space; and (3) inclusion of the pulse shaping into the SAW delay line to eliminate the use of a separate pulse shaping filter, thus further reducing the cost and complexity of the receiver.

7.2.1 Integration of SAW Electronics

Since it has been verified that the ripples in the passband are being caused in part by regenerated SAWs, it would be advantageous to incorporate into the design a miniaturized integrated version of the output buffer which could be positioned to minimize the electrical loading. If the miniature amplifiers are butted directly up to the output taps, the lead lengths from the output taps to the low impedance inputs would be the minimum possible length. Reduction in lead lengths reduces the inductive impedance of the bond wire and this is desirable since the magnitude of regenerated SAWs is directly proportional to load impedance as seen by the IDT. Since diode ring mixers are not being implemented as the multiplying element, the isolating effect of the Darlington stage is not required and a single transistor is probably all that is required per tap in the miniaturized version. Another advantage of this approach is that stray EM coupling effects due to long bond wires would also be reduced, minimizing the risk of spurious EM-induced SAWs.

If integrated output buffers are used, it would make sense to incorporate the input buffer into the design as well, since it has about as many components as one of the output buffers. This approach has already been taken in the development of SAW circuits used for signal processing and includes applications such as correlators [36], adaptive interference suppression [16], hybrid SAW/CCD signal processors [28], and clock recovery and data retiming circuits [44]. Reference [44] illustrates a packaging technique where a SAW filter is placed on one side of a ceramic

DIP and the die containing the integrated interface buffers are placed on the other side. For large production runs, the obvious advantages to using this approach are increased circuit reliability and overall reduction of the assembly costs.

7.2.2 Multipliers

The multiplier circuitry consumed a large proportion (approximately 35%) of the overall board space required to implement the ATDE circuit. A great saving in PC board space would be achieved if the multiplier circuits could be implemented using surface mount electronics or hybrid packages which would contain the MC12002 die wirebonded to the other required elements. The limiting performance factor of the MC12002 multiplier is its maximum attenuation (40 dB). Given that the cost is justified by system requirements, it would be desirable to integrate a new multiplier design which has greater than 40 dB maximum attenuation. As mentioned in the previous section, the advantages to using integration for large production runs include increased circuit reliability and reduced assembly costs.

7.2.3 Inclusion of Pulse Shaping into the ATDE

If the use of an integrated SAW/electronics package is contemplated, it would also be desirable to eliminate the use of a matching inductor since it is bulky and behaves unpredictably at VHF. If this is done, then the problem encountered in matching to the admittance resonance in the apodized SAW IDT at the band edges is no longer of concern. If the apodized SAW is driven in this manner with a low output impedance integrated buffer, then it is not only possible but also extremely

desirable from the system integration level to incorporate pulse shaping into the SAW delay line. This one simple step would make redundant the use of the IF pulse shaping filter shown in Figure 1.2 since it would be incorporated into the ATDE. The advantages of this approach include a simpler receiver design, a reduced insertion loss in the IF path (hence a greater tolerance to noise since only one SAW device would be required), and reduced assembly costs.

If separate pulse shaping and ATDE circuits are contemplated, then it would be advantageous to make the delay line bandwidth as wide as possible. This would mean that the number of apodized IDT fingers would be small. There are a number of advantages to be gained taking this approach: (1) The static capacitance C_T of the apodized IDT would be small enough that resistor magnitude matching could be used. For small C_T , the insertion loss using magnitude matching would be the same as an inductor matching network, however, phase linearity problems associated with inductive tuning would be eliminated; (2) A wide-band device would allow the grounded dummy fingers of the output taps to be a standard $\lambda/8$ width. The slight offset in the centre frequency of operation would not affect the circuit operation and the SAW design process would be simplified.

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APPENDIX 1**REFLECTION MODELLING COMPUTER PROGRAMS**

This appendix contains listings of the programs used to model electrode reflections in SAW IDTs.

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A-1.1 REF1	205
A-1.2 REF2	209
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A-1.4 CURVE	220
A-1.5 FFT	225

```

C PROGRAM REF1
C*****
C
C THIS PROGRAM COMPUTES THE REFLECTIONS GENERATED BY FINGERS
C THAT ARE NOT FRACTION LENGTHS OF A WAVELENGTH.
C
C N1=NUMBER OF WIDER FINGERS
C N2=NUMBER OF LAMBDA/8 FINGERS
C WD=WIDTH OF LAMBDA/8 FINGERS
C CONST=MULTIPLYING CONSTANT FOR WIDE FINGERS
C REFLEC=COMPLEX SUM OF REFLECTIONS
C FREQ=FREQUENCY IN MHZ
C POL=+/-1 SIGN CHANGE CONSTANT
C NTAPS=NUMBER OF TAPS OVER WHICH REFLECTIONS ARE SUMMED
C AMP=AMPLITUDE OF REFLECTIONS RELATIVE TO MAIN SIGNAL
C
C
C WRITTEN N. BIRKETT OCT 23, 1987
C*****
      IMPLICIT REAL*8(A-H,O-Z)
      REAL*8 D(24),VALDB(101),AMPDB(6),AMP(6),FREQ(101)
      COMPLEX*16 REFLEC,SUM
      COMMON/DATA1/VALDB,FREQ

      PI=DABS(DARCOS(-1.0D0))
      N1=6
      DO 900 M=1,1
      NTAPS=7
      N2=12-N1
      CONST=1.06117
C SET UP REFLECTION COEFFICIENT FOR TAPS (AMPDB)
      AMPDB(1)=-37.0D0
      AMPDB(2)=-37.0D0
      AMPDB(3)=-37.0D0
      AMPDB(4)=-37.0D0
      AMPDB(5)=-37.0D0
      AMPDB(6)=-37.0D0
      DO 10 I=1,6
      AMP(I)=10.0D0**(AMPDB(I)/20.0D0)
10 CONTINUE
      WD=1.0D0/8.0D0
      IF (N1.EQ.0) CONST=1.7952324D0
      IF (N1.EQ.2) CONST=1.1590529D0
      IF (N1.EQ.4) CONST=1.0883663D0
      IF (N1.EQ.6) CONST=1.0611706D0
      IF (N1.GE.0) GOTO 20
      N1=0
      CONST=1.0D0
      WD=1.0331424D0*WD
20 CONTINUE
C SET UP INITIAL PARAMETERS
      N2=12-N1
      WDG=WD*CONST
      D(1)=WDG
C

```

```

REF00010
REF00020
REF00030
REF00040
REF00050
REF00060
REF00070
REF00080
REF00090
REF00100
REF00110
REF00120
REF00130
REF00140
REF00150
REF00160
REF00170
REF00180
REF00190
REF00200
REF00210
REF00220
REF00230
REF00240
REF00250
REF00260
REF00270
REF00280
REF00290
REF00300
REF00310
REF00320
REF00330
REF00340
REF00350
REF00360
REF00370
REF00380
REF00390
REF00400
REF00410
REF00420
REF00430
REF00440
REF00450
REF00460
REF00470
REF00480
REF00490
REF00500
REF00510
REF00520
REF00530
REF00540
REF00550

```

C FIRST SET OF WIDE FINGERS	REF00560
DO 100 I=1,N1	REF00570
D(I+1)=D(I)+WDG	REF00580
100 CONTINUE	REF00590
I=I-1	REF00600
IF (N1.EQ.0) I=0	REF00610
C SET OF NARROW FINGERS	REF00620
DO 200 J=(I+1),(I+(2*N2-1))	REF00630
D(J+1)=D(J)+WD	REF00640
200 CONTINUE	REF00650
J=J-1	REF00660
C SECOND SET OF WIDE FINGERS	REF00670
DO 300 K=(J+1),(J+N1)	REF00680
D(K+1)=D(K)+WDG	REF00690
300 CONTINUE	REF00700
C OUTPUT VALUES	REF00710
WRITE(6,1001)N1,N2,WD,CONST,WDG,AMPDB	REF00720
WRITE(9,1001)N1,N2,WD,CONST,WDG,AMPDB	REF00730
C PRINTING DISTANCES	REF00740
DO 400 I=1,24	REF00750
WRITE(6,1002)I,D(I)	REF00760
C WRITE(9,1002)I,D(I)	REF00770
400 CONTINUE	REF00780
DO 450 I=1,101	REF00790
FREQ(I)=60.000+DFLOAT(I-1)*0.200	REF00800
450 CONTINUE	REF00810
C 500 LOOP TO CALCULATE REFLECTIONS FOR DIFFERENT FREQ	REF00820
DO 500 K=1,101	REF00830
REFLEC=DCMPLX(0.000,0.000)	REF00840
A=(-4.000*PI*FREQ(K)/70.000)	REF00850
C CALCULATION OF REFLECTIONS FROM 24 EDGES	REF00860
DO 600 I=1,24	REF00870
POL=DCOS(DFLOAT(I+1)*PI)	REF00880
REFLEC=REFLEC+DCMPLX(POL*DCOS(A*D(I)),POL*DSIN(A*D(I)))	REF00890
600 CONTINUE	REF00900
C SUMMING REFLECTIONS FROM (NTAPS-1)	REF00910
SUM=DCMPLX(0.000,0.000)	REF00920
DO 700 I=1,(NTAPS-1)	REF00930
F1=DFLOAT(I-1)	REF00940
B=AMP(I)	REF00950
SUM=SUM+REFLEC*DCMPLX(B*DCOS(A*F1*D(24)),B*DSIN(A*F1*D(24)))	REF00960
700 CONTINUE	REF00970
C CALCULATE MAGNITUDE FOR THIS FREQUENCY	REF00980
SUM=DCMPLX(1.000,0.000)+SUM	REF00990
VALDB(K)=20.000*DLOG10(CDABS(SUM))	REF01000
C	REF01010
500 CONTINUE	REF01020
C	REF01030
C PRINT FREQ AND MAGNITUDE	REF01040
C	REF01050
DO 800 I=1,101	REF01060
C WRITE(6,1003)FREQ(I),VALDB(I)	REF01070
C WRITE(9,1003)FREQ(I),VALDB(I)	REF01080
800 CONTINUE	REF01090
C	REF01100

C PLOT RESPONSE	REF01110
CALL FRQPLT	REF01120
C	REF01130
N1=-2+M*2	REF01140
C	REF01150
900 CONTINUE	REF01160
C	REF01170
CALL DONEPL	REF01180
C	REF01190
1001 FORMAT(' N1=',I3,' N2=',I3,' WD=',F7.3,' CONST=',F9.6,' WDG=',	REF01200
+F6.3,/,', AMPDB=',6(F7.3,' DB '),//)	REF01210
1002 FORMAT(' EDGE ',I3,' DISTANCE=',F9.4)	REF01220
1003 FORMAT(' FREQ=',F5.1,' MIIZ AMP=',F9.3,' DB')	REF01230
2000 FORMAT(I3)	REF01240
C	REF01250
C	REF01260
C	REF01270
STOP	REF01280
END	REF01290
C	REF01300
C	REF01310
C	REF01320
C	REF01330
C	REF01340
C	REF01350
C	REF01360
C	REF01370
C	REF01380
C	REF01390
C	REF01400
C	REF01410
C	REF01420
C	REF01430
C	REF01440
C	REF01450
C	REF01460
C	REF01470
C	REF01480
C	REF01490
C	REF01500
C	REF01510
C	REF01520
C	REF01530
C	REF01540
C	REF01550
C	REF01560
C	REF01570
C	REF01580
C	REF01590
C	REF01600
C	REF01610
C	REF01620
C	REF01630
C	REF01640
C	REF01650

C*****FRQPLT*****	REF01660
SUBROUTINE FRQPLT	REF01670
C PLOT FREQUENCY RESPONSE	REF01680
C*****N.BIRKETT OCT87*****	REF01690
IMPLICIT REAL*8(A-H,O-Z)	REF01700
REAL*4 XM(101),Y1(101),XMIN,XMAX,XSTP,YMN,YMX,YST	REF01710
REAL*8 FREQ(101),VALDB(101)	REF01720
COMMON/DATA1/VALDB,FREQ	REF01730
C	REF01740
CALL COMPRS	REF01750
CALL PAGE(11.0,8.5)	REF01760
CALL AREA2D(5.8125,4.6875)	REF01770
CALL FRAME	REF01780
CALL GRACE(0.0)	REF01790
CALL NOCHECK	REF01800
CALL XNAME('FREQUENCY(MHZ)\$',100)	REF01810
XMIN=60.0	REF01820
XMAX=80.0	REF01830
XSTP=(XMAX-XMIN)/10.0	REF01840
C	REF01850
DO 100 I=1,101	REF01860
XM(I)=FREQ(I)	REF01870
Y1(I)=VALDB(I)	REF01880
100 CONTINUE	REF01890
C	REF01900
NPTS=101	REF01910
C	REF01920
CALL YAXANG(0.0)	REF01930
CALL YNAME('MAGNITUDE(DB)\$',100)	REF01940
YMX=+02.0	REF01950
YMN=-02.0	REF01960
YST=(YMX-YMN)/10.0DO	REF01970
CALL GRAF(XMIN,XSTP,XMAX,YMN,YST,YMX)	REF01980
CALL CURVE(XM,Y1,NPTS,0)	REF01990
CALL DOT	REF02000
CALL GRID(1,1)	REF02010
CALL RESET('DOT')	REF02020
CALL ENDPL(0)	REF02030
RETURN	REF02040
END	REF02050
	REF02060
	REF02070
	REF02080
	REF02090
	REF02100

```
C*****REF00010
C  PROGRAM REF2REF00020
CREF00030
C  THIS PROGRAM IS A SPURIOUS FREQUENCY CURVE FITTING PROGRAM USINGREF00040
C  ISML ROUTINE ZXMINREF00050
C  MEL REFLECTIONS EFFECTS ARE MODELLEDREF00060
C*****N.BIRKETT NOV87REF00070
      IMPLICIT REAL*8 (A-H,O-Z)REF00080
      EXTERNAL FUNCTREF00090
      REAL*8 X(7),H(28),G(7),W(21),F,EXPDB(101),CALCDB(101),FREQ(101),REF00100
      +DIST(24)REF00110
      INTEGER N,NTAPSREF00120
      COMMON/DATA1/EXPDB,CALCDB,FREQ,DIST,NTAPSREF00130
CREF00140
      N=7REF00150
      NSIG=1REF00160
      MAXFN=500REF00170
      IOPT=0REF00180
      N1=2REF00190
      NTAPS=7REF00200
CREF00210
      CALL INPUTREF00220
      CALL EDGE(N1)REF00230
CREF00240
C  MEL REFLECTION COEFFICIENTSREF00250
      X(1)=+0.0024D0REF00260
      X(2)=+0.0100D0REF00270
      X(3)=+0.0089D0REF00280
      X(4)=+0.0100D0REF00290
      X(5)=+0.0079D0REF00300
      X(6)=+0.0059D0REF00310
      X(7)=+0.0023D0REF00320
C  CALL OPTIMIZING PROGRAMREF00330
      CALL ZXMIN(FUNCT,N,NSIG,MAXFN,IOPT,X,H,G,F,W,IER)REF00340
CREF00350
      DO 100 I=1,101REF00360
C      WRITE(6,1000)FREQ(I),EXPDB(I),CALCDB(I)REF00370
C      WRITE(10,1000)FREQ(I),EXPDB(I),CALCDB(I)REF00380
100  CONTINUEREF00390
CREF00400
      WRITE(6,1001)X,FREF00410
      WRITE(10,1001)X,FREF00420
CREF00430
      CALL FRQPLTREF00440
CREF00450
CREF00460
1000 FORMAT(' FREQ= ',F5.1,' MHZ    EXPDB= ',F9.3,' CALCDB= ',F9.3)REF00470
1001 FORMAT(' X1,7= ',4(D10.2),/, '    ',3(D10.2),/, ' F= ',F7.3)REF00480
CREF00490
      CALL DONEPLREF00500
CREF00510
      STOPREF00520
      ENDREF00530
CREF00540
CREF00550
```



```
C*****REF00560
C                                             REF00570
C      SUBROUTINE INPUT                      REF00580
C                                             REF00590
C      THIS READS IN DATA FROM A FILE AND REFERENCES TO THE CENTRE POINT REF00600
C                                             REF00610
C*****N.BIRKETT NOV87 REF00620
C      IMPLICIT REAL*8(A-H,O-Z)             REF00630
C      REAL*8 EXPDB(101),EXPVAL(101),FREQ(101),CALCDB(101),DIST(24) REF00640
C      INTEGER NTAPS                        REF00650
C      COMMON/DATA1/EXPDB,CALCDB,FREQ,DIST,NTAPS REF00660
C                                             REF00670
C                                             REF00680
C      READ(09,*)STRFRQ                     REF00690
C      READ(09,*)STPFRQ                     REF00700
C      READ(09,*)NPTS                       REF00710
C      READ(09,*)PWR                        REF00720
C                                             REF00730
C      DF=(STPFRQ-STRFRQ)/(DFLOAT(NPTS-1)) REF00740
C                                             REF00750
C      DO 100 I=1,NPTS                      REF00760
C      READ(09,*)EXPRES,EXPIM               REF00770
C      EXPVAL(I)=20.0D0*DLOG10(CDABS(DCMPLX(EXPRES,EXPIM))) REF00780
C      FREQ(I)=STRFRQ+DFLOAT(I-1)*DF        REF00790
100  CONTINUE                              REF00800
C                                             REF00810
C      DO 200 I=1,NPTS                      REF00820
C      EXPDB(I)=EXPVAL(I)-EXPVAL(60)        REF00830
200  CONTINUE                              REF00840
C                                             REF00850
C      RETURN                               REF00860
C      END                                  REF00870
C                                             REF00880
C                                             REF00890
C                                             REF00900
C                                             REF00910
C                                             REF00920
C                                             REF00930
C                                             REF00940
C                                             REF00950
C                                             REF00960
C                                             REF00970
C                                             REF00980
C                                             REF00990
C                                             REF01000
C                                             REF01010
C                                             REF01020
C                                             REF01030
C                                             REF01040
C                                             REF01050
C                                             REF01060
C                                             REF01070
C                                             REF01080
C                                             REF01090
C                                             REF01100
```

C	*****	REF01110
	SUBROUTINE EDGE(N1)	REF01120
C		REF01130
C	COMPUTES EDGE DISTANCES FOR A SET OF 12 FINGERS	REF01140
C	*****N.BIRKETT NOV87	REF01150
	IMPLICIT REAL*8(A-H,O-Z)	REF01160
	REAL*8 DIST(24),EXPDB(101),CALCDB(101),FREQ(101)	REF01170
	COMPLEX*16 REFLEC,SUM	REF01180
	INTEGER NTAPS,N1	REF01190
	COMMON/DATA1/EXPDB,CALCDB,FREQ,DIST,NTAPS	REF01200
C		REF01210
	P1=DABS(DARCOS(-1.0D0))	REF01220
	WD=1.0D0/8.0D0	REF01230
	IF (N1.EQ.0) CONST=1.7952324D0	REF01240
	IF (N1.EQ.2) CONST=1.1590529D0	REF01250
	IF (N1.EQ.4) CONST=1.0883663D0	REF01260
	IF (N1.EQ.6) CONST=1.0611706D0	REF01270
	IF (N1.GE.0) GOTO 20	REF01280
	N1=0	REF01290
	N2=12	REF01300
	CONST=1.0D0	REF01310
	WD=1.0331424D0*WD	REF01320
20	CONTINUE	REF01330
	N2=12-N1	REF01340
C	CALCULATION OF EDGE DISTANCES	REF01350
	WDG=WD*CONST	REF01360
	DIST(1)=WDG	REF01370
C	FIRST SET OF WIDE FINGERS	REF01380
	DO 100 I=1,N1	REF01390
	DIST(I+1)=DIST(I)+WDG	REF01400
100	CONTINUE	REF01410
	I=I-1	REF01420
C		REF01430
	IF (N1.EQ.0) I=0	REF01440
C	SET OF NARROW FINGERS	REF01450
	DO 200 J=(I+1),(I+(2*N2-1))	REF01460
	DIST(J+1)=DIST(J)+WD	REF01470
200	CONTINUE	REF01480
	J=J-1	REF01490
C	SECOND SET OF WIDE FINGERS	REF01500
	DO 300 K=(J+1),(J+N1)	REF01510
	DIST(K+1)=DIST(K)+WDG	REF01520
300	CONTINUE	REF01530
	WRITE(6,1001)N1,N2,WD,CONST,WDG	REF01540
	WRITE(10,1001)N1,N2,WD,CONST,WDG	REF01550
C	PRINTING DISTANCES	REF01560
	DO 400 I=1,24	REF01570
C	WRITE(6,1002)I,DIST(I)	REF01580
C	WRITE(10,1002)I,DIST(I)	REF01590
400	CONTINUE	REF01600
	1001 FORMAT(' N1=',I3,' N2=',I3,' WD=',F7.3,' CONST=',F9.6,' WDG=',	REF01610
	+F6.3,//)	REF01620
	1002 FORMAT(' EDGE ',I3,' DISTANCE= ',F9.4)	REF01630
	RETURN	REF01640
	END	REF01650

```
(*****REF01660
C                                     REF01670
      SUBROUTINE FUNCT(N,X,F)                                     REF01680
C                                     REF01690
C THIS SUBROUTINE COMPUTES THE ERROR BETWEEN THE EXPERIMENTAL FREQUENCYREF01700
C RESPONSE AND THE RESPONSE OBTAINED BY THE POLYNOMIAL COMPUTATIONS REF01710
C                                     REF01720
(*****N.BIRKETT NOV87 REF01730
      IMPLICIT REAL*8 (A-H,O-Z)                                   REF01740
      INTEGER N,NTAPS                                             REF01750
      REAL*8 X(N),F,FREQ(101),EXPDB(101),CALCDB(101),DIST(24) REF01760
      COMMON/DATA1/EXPDB,CALCDB,FREQ,DIST,NTAPS                  REF01770
C                                     REF01780
      PI=DABS(DARCOS(-1.0D0))                                     REF01790
      F=0.0D0                                                     REF01800
      DO 100 I=1,101                                             REF01810
      W=1.0D6*FREQ(I)*2*PI                                       REF01820
C                                     REF01830
      CALL REFLEC(X,I)                                           REF01840
C                                     REF01850
      U=10.0D0**(CALCDB(I)/20.0D0)                               REF01860
      EXPAMP=10.0D0**(EXPDB(I)/20.0D0)                           REF01870
      F=F+DABS(U-EXPAMP)                                         REF01880
C                                     REF01890
100  CONTINUE                                                    REF01900
C                                     REF01910
C                                     REF01920
      RETURN                                                     REF01930
      END                                                         REF01940
C                                     REF01950
C                                     REF01960
C                                     REF01970
C                                     REF01980
C                                     REF01990
C                                     REF02000
C                                     REF02010
C                                     REF02020
C                                     REF02030
C                                     REF02040
C                                     REF02050
C                                     REF02060
C                                     REF02070
C                                     REF02080
C                                     REF02090
C                                     REF02100
C                                     REF02110
C                                     REF02120
C                                     REF02130
C                                     REF02140
C                                     REF02150
C                                     REF02160
C                                     REF02170
C                                     REF02180
C                                     REF02190
C                                     REF02200
```

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C*****REF02210
C                                     REF02220
      SUBROUTINE REFLEC(AMP,K)         REF02230
C                                     REF02240
C   CALCULATION OF REFLECTIONS OFF EDGES REF02250
C                                     REF02260
C*****REF02270
C                                     REF02280
      IMPLICIT REAL*8 (A-H,O-Z)       REF02290
      COMPLEX*16 REF,SUM               REF02300
      REAL*8 EXPDB(101),CALCDB(101),FREQ(101),DIST(24),AMP(7) REF02310
      INTEGER NTAPS,K                 REF02320
      COMMON/DATA1/EXPDB,CALCDB,FREQ,DIST,NTAPS REF02330
C                                     REF02340
      PI=DABS(DARCOS(-1.0D0))          REF02350
C   CALCULATE REFLECTIONS FOR K'TH FREQ REF02360
      REF=DCMPLX(0.0D0,0.0D0)          REF02370
      SHIFT=1.50D0                     REF02380
      A=(-4.0D0*PI*(FREQ(K)+SHIFT)/70.0D0) REF02390
C   CALCULATION OF REFLECTIONS FROM 24 EDGES REF02400
      DO 600 I=1,24                    REF02410
        POL=DCOS(DFLOAT(I+1)*PI)       REF02420
        REF=REF+DCMPLX(POL*DCOS(A*DIST(I)),POL*DSIN(A*DIST(I))) REF02430
600   CONTINUE                         REF02440
C   SUMMING REFLECTIONS FROM NTAPS     REF02450
      SUM=DCMPLX(0.0D0,0.0D0)          REF02460
      DO 700 I=1,NTAPS                 REF02470
        FI=DFLOAT(I-1)                 REF02480
        B=AMP(I)                       REF02490
        SUM=SUM+REF*DCMPLX(B*DCOS(A*FI*DIST(24)),B*DSIN(A*FI*DIST(24))) REF02500
700   CONTINUE                         REF02510
C   CALCULATE MAGNITUDE FOR THIS FREQUENCY REF02520
      SUM=DCMPLX(1.0D0,0.0D0)+SUM      REF02530
      CALCDB(K)=20.0D0*DLOG10(DABS(SUM)) REF02540
C                                     REF02550
C                                     REF02560
      RETURN                           REF02570
      END                              REF02580
C                                     REF02590
C                                     REF02600
C                                     REF02610
C                                     REF02620
C                                     REF02630
C                                     REF02640
C                                     REF02650
C                                     REF02660
C                                     REF02670
C                                     REF02680
C                                     REF02690
C                                     REF02700
C                                     REF02710
C                                     REF02720
C                                     REF02730
C                                     REF02740
C                                     REF02750
```

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C*****FRQPLT*****REF02760
      SUBROUTINE FRQPLT                                REF02770
C          FREQUENCY PLOTS                            REF02780
C*****N.BIRKETT NOV87REF02790
      IMPLICIT REAL*8(A-H,O-Z)                        REF02800
      REAL*4 XM(101),Y1(101),Y2(101),XMIN,XMAX,XSTP, YMN, YMX, YST, REF02810
      +IPAK(100)                                       REF02820
      REAL*8 FREQ(101),EXPDB(101),CALCDB(101),DIST(24) REF02830
      INTEGER NTAPS                                    REF02840
      COMMON/DATA1/EXPDB,CALCDB,FREQ,DIST,NTAPS        REF02850
C                                                    REF02860
      CALL COMPRS                                     REF02870
      CALL PAGE(11.0,8.5)                             REF02880
      CALL AREA2D(5.8125,4.6875)                       REF02890
      CALL FRAME                                       REF02900
      CALL GRACE(0.0)                                  REF02910
      CALL NOCHECK                                    REF02920
C                                                    REF02930
      CALL HEADIN('SPURIOUS REFLECTION ANALYSIS$',28,1.,1) REF02940
      CALL LINESP(3.0)                                REF02950
      CALL HEIGHT(0.10)                               REF02960
      CALL LINES('EXPERIMENT$',IPAK,1)                 REF02970
      CALL LINES('MODEL$',IPAK,2)                     REF02980
      CALL RESET('HEIGHT')                             REF02990
      CALL MYLEGN('$ ',1)                              REF03000
      CALL XNAME('FREQUENCY(MHZ)$ ',100)               REF03010
C                                                    REF03020
      XMIN=60.0                                         REF03030
      XMAX=80.0                                         REF03040
      XSTP=(XMAX-XMIN)/10.0                            REF03050
C                                                    REF03060
      DO 100 I=1,101                                   REF03070
      XM(I)=FREQ(I)                                     REF03080
      Y1(I)=EXPDB(I)                                   REF03090
      Y2(I)=CALCDB(I)                                  REF03100
100  CONTINUE                                          REF03110
      NPTS=101                                          REF03120
      CALL YAXANG(0.0)                                  REF03130
      CALL YNAME('MAGNITUDE(DB)$ ',100)                REF03140
C                                                    REF03150
      YMX=+01.0                                         REF03160
      YMN=-01.0                                         REF03170
      YST=(YMX-YMN)/10.000                             REF03180
C                                                    REF03190
      CALL GRAF(XMIN,XSTP,XMAX,YMN,YST,YMX)           REF03200
      CALL LEGLIN                                       REF03210
      CALL CURVE(XM,Y1,NPTS,0)                         REF03220
      CALL DASH                                         REF03230
      CALL LEGLIN                                       REF03240
      CALL CURVE(XM,Y2,NPTS,0)                         REF03250
      CALL DOT                                          REF03260
      CALL GRID(1,1)                                   REF03270
      CALL RESET ('DOT')                               REF03280
      CALL HEIGHT(0.10)                                REF03290
      CALL LEGEND(IPAK,2,4.0,0.2)                     REF03300

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```
      CALL RESET('HEIGHT')
C
C      CALL ENDPL(0)
      RETURN
      END
C
```

```
REF03310
REF03320
REF03330
REF03340
REF03350
REF03360
REF03370
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C*****REF00010
C  PROGRAM REF3                                REF00020
C                                              REF00030
C  THIS PROGRAM COMPUTES THE FREQUENCY RESPONSE OF A FINGER ARRAY WITH REF00040
C  ARBITRARY SPACING BETWEEN FINGERS (DETERMINED BY CONST) REF00050
C  MEL REFLECTIONS EFFECTS ARE MODELLED REF00060
C  NOTE:4 FINGERS PER WAVELENGTH ARE USED REF00070
C  CONST=METALIZATION RATIO REF00080
C                                              REF00090
C  WRITTEN N. BIRKETT DEC 87 REF00100
C*****REF00110
C      IMPLICIT REAL*8 (A-H,O-Z) REF00120
C      REAL*8 X(6),CALCDB(101),FREQ(101),D(24) REF00130
C      INTEGER N,NTAPS REF00140
C      COMMON/DATA1/CALCDB,FREQ,L,NTAPS REF00150
C                                              REF00160
C      NTAPS=7 REF00170
C      CONST=0.9D0 REF00180
C                                              REF00190
C  MEL REFLECTION COEFFICIENTS REF00200
C      X(1)=0.0141D0 REF00210
C      X(2)=0.0141D0 REF00220
C      X(3)=0.0141D0 REF00230
C      X(4)=0.0141D0 REF00240
C      X(5)=0.0141D0 REF00250
C      X(6)=0.0141D0 REF00260
C                                              REF00270
C      DF=20.0/100.0D0 REF00280
C      DO 50 I=1,101 REF00290
C      FREQ(I)=60.0D0+DFLOAT(I-1)*DF REF00300
50  CONTINUE REF00310
C                                              REF00320
C      CALL EDGE(CONST) REF00330
C                                              REF00340
C  CALCULATE REFLECTION FOR EACH FREQUENCY REF00350
C      DO 100 I=1,101 REF00360
C      CALL REFLEC(X,I) REF00370
C      WRITE(6,1000)FREQ(I),CALCDB(I) REF00380
C      WRITE(10,1000)FREQ(I),CALCDB(I) REF00390
100  CONTINUE REF00400
C                                              REF00410
C      CALL FRQPLT REF00420
C      CALL DONEPL REF00430
C                                              REF00440
C                                              REF00450
1000 FORMAT(' FREQ= ',F5.1,' MHZ  CALCDB= ',F9.3) REF00460
C                                              REF00470
C      STOP REF00480
C      END REF00490
C                                              REF00500
C                                              REF00510
C                                              REF00520
C                                              REF00530
C                                              REF00540
C                                              REF00550
```

```
C*****REF00560
C                                             REF00570
C      SUBROUTINE EDGE(CONST)                REF00580
C                                             REF00590
C      COMPUTES EDGE DISTANCES FOR A SET OF 12 FINGERS REF00600
C                                             REF00610
C*****N.BIRKETT DEC87REF00620
C      IMPLICIT REAL*8(A-H,O-Z)              REF00630
C      REAL*8 D(24),CALCDB(101),FREQ(101),CONST REF00640
C      INTEGER NTAPS                          REF00650
C      COMMON/DATA1/CALCDB,FREQ,D,NTAPS      REF00660
C                                             REF00670
C      DO 100 I=1,12                          REF00680
C        J=2*I                                REF00690
C        K=J-1                                REF00700
C        D(J)=DFLOAT(I)*0.25D0               REF00710
C        D(K)=D(J)-CONST*0.25D0              REF00720
C 100  CONTINUE                               REF00730
C                                             REF00740
C      DO 200 I=1,24                          REF00750
C        WRITE(6,1002)I,D(I)                  REF00760
C        WRITE(9,1002)I,D(I)                  REF00770
C 200  CONTINUE                               REF00780
C                                             REF00790
C 1002 FORMAT(' EDGE ',13,' DISTANCE= ',F9.4) REF00800
C                                             REF00810
C      RETURN                                REF00820
C      END                                    REF00830
C                                             REF00840
C                                             REF00850
C                                             REF00860
C                                             REF00870
C                                             REF00880
C                                             REF00890
C                                             REF00900
C                                             REF00910
C                                             REF00920
C                                             REF00930
C                                             REF00940
C                                             REF00950
C                                             REF00960
C                                             REF00970
C                                             REF00980
C                                             REF00990
C                                             REF01000
C                                             REF01010
C                                             REF01020
C                                             REF01030
C                                             REF01040
C                                             REF01050
C                                             REF01060
C                                             REF01070
C                                             REF01080
C                                             REF01090
C                                             REF01100
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(*)*****REF01110
C                                     REF01120
      SUBROUTINE REFLEC(AMP,K)                                     REF01130
C                                     REF01140
C   CALCULATION OF REFLECTIONS OFF EDGES                           REF01150
C                                     REF01160
C*****N.BIRKETT DEC87REF01170
C                                     REF01180
      IMPLICIT REAL*8 (A-H,O-Z)                                     REF01190
      COMPLEX*16 REF,SUM                                           REF01200
      REAL*8  CALCDB(101),FREQ(101),D(24),AMP(6)                   REF01210
      INTEGER NTAPS,K                                              REF01220
      COMMON/DATA1/CALCDB,FREQ,D,NTAPS                             REF01230
C                                     REF01240
      PI=DABS(DARCOS(-1.0D0))                                       REF01250
C   CALCULATE REFLECTIONS FOR K'TH FREQ                             REF01260
      REF=DCMLPX(0.0D0,0.0D0)                                       REF01270
      A=(-4.0D0*PI*FREQ(K)/70.0D0)                                   REF01280
C   CALCULATION OF REFLECTIONS FROM 24 EDGES                         REF01290
      DO 600 I=1,24                                                  REF01300
        POL=DCOS(DFLOAT(I+1)*PI)                                     REF01310
        REF=REF+DCMLPX(POL*DCOS(A*D(I)),POL*DSIN(A*D(I)))          REF01320
600  CONTINUE                                                       REF01330
C   SUMMING REFLECTIONS FROM (NTAPS-1)                             REF01340
      SUM=DCMLPX(0.0D0,0.0D0)                                       REF01350
      DO 700 I=1,(NTAPS-1)                                           REF01360
        FI=DFLOAT(I-1)                                              REF01370
        B=AMP(I)                                                     REF01380
        SUM=SUM+REF*DCMLPX(B*DCOS(A*FI*D(24)),B*DSIN(A*FI*D(24))) REF01390
700  CONTINUE                                                       REF01400
C   CALCULATE MAGNITUDE FOR THIS FREQUENCY                         REF01410
      SUM=DCMLPX(1.0D0,0.0D0)+SUM                                    REF01420
      CALCDB(K)=20.0D0*DLOG10(DABS(SUM)+1.0D-9)                     REF01430
C                                     REF01440
C                                     REF01450
      RETURN                                                         REF01460
      END                                                            REF01470
C                                     REF01480
C                                     REF01490
C                                     REF01500
C                                     REF01510
C                                     REF01520
C                                     REF01530
C                                     REF01540
C                                     REF01550
C                                     REF01560
C                                     REF01570
C                                     REF01580
C                                     REF01590
C                                     REF01600
C                                     REF01610
C                                     REF01620
C                                     REF01630
C                                     REF01640
C                                     REF01650

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C*****FRQPLT*****REF01660
SUBROUTINE FRQPLTREF01670
C          FREQUENCY PLOTSREF01680
C*****N. BIRKETT NOV8REF01690
IMPLICIT REAL*8(A-H,O-Z)REF01700
REAL*4 XM(101),Y1(101),Y2(101),XMIN,XMAX,XSTP,YMN,YMX,YST,REF01710
+IPAK(100)REF01720
REAL*8 FREQ(101),CALCDB(101),D(24)REF01730
INTEGER NTAPSREF01740
COMMON/DATA1/CALCDB,FREQ,D,NTAPSREF01750
CREF01760
CALL COMPRSREF01770
CALL PAGE(11.0,8.5)REF01780
CALL AREA2D(5.8125,4.6875)REF01790
CALL FRAMEREF01800
CALL GRACE(0.0)REF01810
CALL NOCHEKREF01820
CALL HEADIN('SPURIOUS REFLECTION ANALYSIS$',28,1.,1)REF01830
CALL LINE$P(3.0)REF01840
CALL HEIGHT(0.10)REF01850
CALL LINE$('EXPERIMENT$',IPAK,1)REF01860
CALL LINE$('MODEL$',IPAK,2)REF01870
CALL RESET('HEIGHT')REF01880
CALL MYLEGN('$',1)REF01890
CALL XNAME('FREQUENCY(MHZ)$',100)REF01900
XMIN=60.0REF01910
XMAX=80.0REF01920
XSTP=(XMAX-XMIN)/10.0REF01930
DO 100 I=1,101REF01940
XM(I)=FREQ(I)REF01950
Y1(I)=CALCDB(I)REF01960
Y2(I)=CALCDB(I)REF01970
100 CONTINUEREF01980
NPTS=101REF01990
CALL YAXANG(0.0)REF02000
CALL YNAME('MAGNITUDE(DB)$',100)REF02010
YMX=+01.0REF02020
YMN=-01.0REF02030
YST=(YMX-YMN)/10.000REF02040
CALL GRAF(XMIN,XSTP,XMAX,YMN,YST,YMX)REF02050
CALL LEGLINREF02060
CALL CURVE(XM,Y1,NPTS,0)REF02070
CALL DASHREF02080
CALL LEGLINREF02090
C CALL CURVE(XM,Y2,NPTS,0)REF02100
CALL DOTREF02110
CALL GRID(1,1)REF02120
CALL RESET('DOT')REF02130
CALL HEIGHT(0.10)REF02140
C CALL LEGEND(IPAK,2,4.0,0.2)REF02150
CALL RESET('HEIGHT')REF02160
CREF02170
CALL ENDPL(0)REF02180
RETURNREF02190
ENDREF02200

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C *****	CUR00010
C PROGRAM CURVE	CUR00020
C	CUR00030
C THIS PROGRAM IS A SPURIOUS FREQUENCY CURVE FITTING PROGRAM USING	CUR00040
C ISML ROUTINE ZXMIN	CUR00050
C BOTH REGENERATIVE AND EM COUPLING EFFECTS ARE MODELLED	CUR00060
C *****N.BIRKETT DEC87	CUR00070
IMPLICIT REAL*8 (A-H,O-Z)	CUR00080
EXTERNAL FUNCT	CUR00090
REAL*8 X(13),H(91),G(13),W(39),F,EXPDB(101),CALCDB(101),FREQ(101)	CUR00100
INTEGER N	CUR00110
COMMON/DATA1/EXPDB,CALCDB,FREQ	CUR00120
C	CUR00130
C	CUR00140
C	CUR00150
N=13	CUR00160
NSIG=1	CUR00170
MAXFN=500	CUR00180
IOPT=0	CUR00190
C	CUR00200
C X(1) TO X(6) DETERMINE THE AMPLITUDE OF THE EM GENERATED SAWS	CUR00210
C X(7) TO X(12) DETERMINE THE AMPLITUDE OF THE REGENERATED SAWS	CUR00220
C X(13) DETERMINES THE ARBITRARY PHASE OF THE EM FEEDTHROUGH	CUR00230
C	CUR00240
CALL INPUT	CUR00250
C	CUR00260
C	CUR00270
DO 200 J=1,3	CUR00280
C	CUR00290
C EM INDUCED AMPLITUDE COEFFICIENTS	CUR00300
X(1)=0.0150D0	CUR00310
X(2)=-0.0020D0	CUR00320
X(3)=0.0072D0	CUR00330
X(4)=-0.0000D0	CUR00340
X(5)=0.0000D0	CUR00350
X(6)=-0.0049D0	CUR00360
C REGENERATED SAW AMPLITUDE COEFFICIENTS	CUR00370
X(7)=-0.0033D0	CUR00380
X(8)=-0.0034D0	CUR00390
X(9)=-0.0000D0	CUR00400
X(10)=-0.0060D0	CUR00410
X(11)=-0.0000D0	CUR00420
X(12)=-0.0043D0	CUR00430
C	CUR00440
C X(13) IS THE EM PHASE IN DEGREES	CUR00450
C	CUR00460
X(13)=10.0D0+DFLOAT(J-1)*5.0D0	CUR00470
C	CUR00480
C OPTIMIZE DATA TO FIT EXPERIMENTAL DATA	CUR00490
CALL ZXMIN(FUNCT,N,NSIG,MAXFN,IOPT,X,H,G,F,W,IER)	CUR00500
C	CUR00510
C	CUR00520
C	CUR00530
C	CUR00540
C	CUR00550

C WRITE DATA INTO A FILE	CUR00560
DO 100 I=1,101	CUR00570
C WRITE(6,1000)FREQ(I),EXPDB(I),CALCDB(I)	CUR00580
C WRITE(10,1000)FREQ(I),EXPDB(I),CALCDB(I)	CUR00590
100 CONTINUE	CUR00600
C	CUR00610
C WRITE COEFFICIENTS AND ERROR FUNCTION F	CUR00620
WRITE(6,1001)X,F	CUR00630
WRITE(10,1001)X,F	CUR00640
C	CUR00650
C PLOT THE EXPERIMENTAL AND OPTIMIZED DATA	CUR00660
CALL FRQPLT	CUR00670
C	CUR00680
200 CONTINUE	CUR00690
C	CUR00700
1000 FORMAT(' FREQ= ',F5.1,' MHZ EXPDB= ',F9.3,' CALCDB= ',F9.3)	CUR00710
1001 FORMAT(' X1,6 ',6(D10.2),/, ' X7,12 ',6(D10.2),/, ' PHASE=',D10.2	CUR00720
+,/, ' F= ',F7.2)	CUR00730
C	CUR00740
CALL DONEPL	CUR00750
C	CUR00760
STOP	CUR00770
END	CUR00780
C	CUR00790
C	CUR00800
C	CUR00810
C	CUR00820
C	CUR00830
C	CUR00840
C	CUR00850
C	CUR00860
C	CUR00870
C	CUR00880
C	CUR00890
C	CUR00900
C	CUR00910
C	CUR00920
C	CUR00930
C	CUR00940
C	CUR00950
C	CUR00960
C	CUR00970
C	CUR00980
C	CUR00990
C	CUR01000
C	CUR01010
C	CUR01020
C	CUR01030
C	CUR01040
C	CUR01050
C	CUR01060
C	CUR01070
C	CUR01080
C	CUR01090
C	CUR01100

```
C CUR01110
C*****CUR01120
C CUR01130
C SUBROUTINE INPUT CUR01140
C CUR01150
C THIS READS IN DATA FROM A FILE AND REFERENCES TO THE CENTRE POINT CUR01160
C CUR01170
C*****N.BIRKETT DEC87CUR01180
C IMPLICIT REAL*8(A-H,O-Z) CUR01190
C REAL*8 EXPDB(101),EXPVAL(101),FREQ(101),CALCDB(101) CUR01200
C COMMON/DATA1/EXPDB,CALCDB,FREQ CUR01210
C CUR01220
C CUR01230
C READ(09,*)STRFRQ CUR01240
C READ(09,*)STPFRQ CUR01250
C READ(09,*)NPTS CUR01260
C READ(09,*)PWR CUR01270
C CUR01280
C DF=(STPFRQ-STRFRQ)/(DFLOAT(NPTS-1)) CUR01290
C CUR01300
C DO 100 I=1,NPTS CUR01310
C READ(09,*)EXPRE,EXPIM CUR01320
C EXPVAL(I)=20.0D0*DLOG10(CDABS(DCMPLX(EXPRE,EXPIM))) CUR01330
C FREQ(I)=STRFRQ+DFLOAT(I-1)*DF CUR01340
100 CONTINUE CUR01350
C CUR01360
C DO 200 I=1,NPTS CUR01370
C EXPDB(I)=EXPVAL(I)-EXPVAL(51) CUR01380
200 CONTINUE CUR01390
C CUR01400
C RETURN CUR01410
C END CUR01420
C CUR01430
C CUR01440
C CUR01450
C CUR01460
C CUR01470
C CUR01480
C CUR01490
C CUR01500
C CUR01510
C CUR01520
C CUR01530
C CUR01540
C CUR01550
C CUR01560
C CUR01570
C CUR01580
C CUR01590
C CUR01600
C CUR01610
C CUR01620
C CUR01630
C CUR01640
C CUR01650
```

```

C*****CUR01660
C                                          CUR01670
      SUBROUTINE FUNCT(N,X,F)                CUR01680
C                                          CUR01690
C THIS SUBROUTINE COMPUTES THE ERROR BETWEEN THE ACTUAL FREQUENCY    CUR01700
C RESPONSE AND THE RESPONSE OBTAINED BY THE POLYNOMIAL COMPUTATIONS    CUR01710
C                                          CUR01720
C*****N.BIRKETT DEC 87CUR01730
      IMPLICIT REAL*8 (A-H,O-Z)              CUR01740
      INTEGER N                               CUR01750
      COMPLEX*16 V,PHASE,ONE,EMSUM,REFSUM     CUR01760
      REAL*8 X(N),F,FREQ(101),EXPDB(101),CALCDB(101),EXPVAL,T(6)      CUR01770
      COMMON/DATA1/EXPDB,CALCDB,FREQ         CUR01780
C                                          CUR01790
C      PI=DABS(DARCOS(-1.0D0))                CUR01800
      DELAY=1.0D0/22.585D6                    CUR01810
      DO 50 I=1,6                             CUR01820
      T(I)=DFLOAT(I)*DELAY                     CUR01830
50    CONTINUE                                CUR01840
C                                          CUR01850
C      F=0.0D0                                CUR01860
      PHI=X(13)                                CUR01870
      PHASE=DCMPLX(DCOS(-PHI*PI/180.0D0),DSIN(-PHI*PI/180.0D0))    CUR01880
      ONE=DCMPLX(1.0D0,0.0D0)                 CUR01890
C                                          CUR01900
C      DO 100 I=1,101                          CUR01910
      W=1.0D6*FREQ(I)*2*PI                     CUR01920
C                                          CUR01930
C      EMSUM=DCMPLX(0.0D0,0.0D0)               CUR01940
      DO 80 J=1,6                             CUR01950
      V=DCMPLX(X(J),0.0D0)*DCMPLX(DCOS(-W*T(J)),DSIN(-W*T(J)))    CUR01960
      EMSUM=EMSUM+V                           CUR01970
80    CONTINUE                                CUR01980
C                                          CUR01990
C      REFSUM=DCMPLX(0.0D0,0.0D0)              CUR02000
      DO 90 J=1,6                             CUR02010
      TT=2.0D0*T(J)                            CUR02020
      V=DCMPLX(X(J+6),0.0D0)*DCMPLX(DCOS(-W*TT),DSIN(-W*TT))    CUR02030
      REFSUM=REFSUM+V                         CUR02040
90    CONTINUE                                CUR02050
C                                          CUR02060
C      U=CDABS(ONE+PHASE*EMSUM+REFSUM)          CUR02070
      EXPAMP=10.0D0**((EXPDB(I)/20.0D0))      CUR02080
      F=F+DABS(U-EXPAMP)                      CUR02090
      CALCDB(I)=20.0D0*DLOG10(U+1.0D-9)      CUR02100
C                                          CUR02110
100   CONTINUE                                CUR02120
C                                          CUR02130
C                                          CUR02140
      RETURN                                  CUR02150
      END                                    CUR02160
C                                          CUR02170
C                                          CUR02180
C                                          CUR02190
C                                          CUR02200

```

```
C*****FRQPLT*****CUR02210
      SUBROUTINE FRQPLT                                CUR02220
C  THIS ROUTINE PLOTS THE DATA                        CUR02230
C*****N.BIRKETT' DEC87 CUR02240
      IMPLICIT REAL*8(A-H,O-Z)                        CUR02250
      REAL*4 XM(101),Y1(101),Y2(101),XMIN,XMAX,XSTP,YMN,YMX,YST, CUR02260
      +IPAK(100)                                       CUR02270
      REAL*8 FREQ(101),EXPDB(101),CALCDB(101)          CUR02280
      COMMON/DATA1/EXPDB,CALCDB,FREQ                  CUR02290
C                                                    CUR02300
      CALL COMPRS                                       CUR02310
      CALL PAGE(11.0,8.5)                               CUR02320
      CALL AREA2D(5.8125,4.6875)                       CUR02330
      CALL FRAME                                       CUR02340
      CALL GRACE(0.0)                                   CUR02350
      CALL NOCHECK                                     CUR02360
      CALL HEADIN('SPURIOUS REFLECTION ANALYSIS$',28,1.,1) CUR02370
      CALL LINESP(3.0)                                  CUR02380
      CALL HEIGHT(0.10)                                CUR02390
      CALL LINES('EXPERIMENT$',IPAK,1)                 CUR02400
      CALL LINES('MODEL$',IPAK,2)                     CUR02410
      CALL RESET('HEIGHT')                             CUR02420
      CALL MYLEGN('$',1)                                CUR02430
      CALL XNAME('FREQUENCY(MHZ)$',100)                CUR02440
      XMIN=60.0                                         CUR02450
      XMAX=80.0                                         CUR02460
      XSTP=(XMAX-XMIN)/10.0                             CUR02470
      DO 100 I=1,101                                   CUR02480
      XM(I)=FREQ(I)                                     CUR02490
      Y1(I)=EXPDB(I)                                    CUR02500
      Y2(I)=CALCDB(I)                                  CUR02510
100  CONTINUE                                          CUR02520
C                                                    CUR02530
      NPTS=101                                          CUR02540
      CALL YAXANG(0.0)                                  CUR02550
      CALL YNAME('MAGNITUDE(DB)$',100)                 CUR02560
      YMX=+01.0                                         CUR02570
      YMN=-01.0                                         CUR02580
      YST=(YMX-YMN)/10.000                             CUR02590
      CALL GRAF(XMIN,XSTP,XMAX,YMN,YST,YMX)            CUR02600
      CALL LEGLIN                                       CUR02610
      CALL CURVE(XM,Y1,NPTS,0)                         CUR02620
      CALL DASH                                         CUR02630
      CALL LEGLIN                                       CUR02640
      CALL CURVE(XM,Y2,NPTS,0)                         CUR02650
      CALL DOT                                          CUR02660
      CALL GRID(1,1)                                    CUR02670
      CALL RESET('DOT')                                CUR02680
      CALL HEIGHT(0.10)                                CUR02690
      CALL LEGEND(IPAK,2,4.0,0.2)                     CUR02700
      CALL RESET('HEIGHT')                             CUR02710
      CALL ENDPL(0)                                     CUR02720
      RETURN                                           CUR02730
      END                                              CUR02740
```

```

C*****
C FFT * FFT00010
C FFT PROGRAM TO COMPUTE MAGNITUDE OF REFLECTIONS BASED ON * FFT00020
C MAGNITUDE AND NUMBER OF REFLECTION TIME SAMPLES. * FFT00030
C WRITTEN N. BIRKETT NOV 27, 1987 * FFT00040
C*****N.BIRKETT NOV87 * FFT00050
      IMPLICIT REAL*8(A-H,O-Z) FFT00060
      REAL*8 VALDB(1024),FREQ(1024),VR(1024) FFT00070
      COMPLEX*16 VIN(1024),VOUT(1024) FFT00080
      COMMON/DATA1/VALDB,FREQ FFT00090
      PI=DABS(DARCOS(-1.0D0)) FFT00100
      REFDB=-37.0D0 FFT00110
C
      DO 500 K=1,5 FFT00120
      IF(K.EQ.1) NFI=1 FFT00130
      IF(K.EQ.2) NFI=2 FFT00140
      IF(K.EQ.3) NFI=4 FFT00150
      IF(K.EQ.4) NFI=8 FFT00160
      IF(K.EQ.5) NFI=20 FFT00170
      NF=10 FFT00180
      NFFT=1024 FFT00190
      FO=70.0D0 FFT00200
      DF=4.0D0*FO/DFLOAT(NFFT) FFT00210
      R=10.0D0**(REFDB/20.0D0) FFT00220
C
      DO 100 I=1,NFFT FFT00230
      FREQ(I)=DFLOAT(I)*DF FFT00240
      VIN(I)=DCMPLX(0.0D0,0.0D0) FFT00250
100 CONTINUE FFT00260
C
      AB=(1.0D0-R)*(1.0D0+R) FFT00270
      COEF=0.0D0 FFT00280
      VR(1)=R FFT00290
C
      DO 200 I=2,2*NFI FFT00300
      VR(I)=R*(-AB)**(I-1)+COEF*(R**3)*(-AB)**(I-2) FFT00310
C
      COEF=COEF+DFLOAT(I-1) FFT00320
200 CONTINUE FFT00330
      DO 250 I=1,2*NFI FFT00340
      VIN(I)=DCMPLX(VR(I),0.0D0) FFT00350
250 CONTINUE FFT00360
C
      CALL TINFRQ(VIN,VOUT,NF,NFFT) FFT00370
      DO 300 I=1,NFFT FFT00380
      VALDB(I)=20.0D0*DLOG10(DABS(VOUT(I))) FFT00390
C
      WRITE(11,1001)I,FREQ(I),VALDB(I) FFT00400
300 CONTINUE FFT00410
C
      CALL FRQPLT FFT00420
500 CONTINUE FFT00430
      CALL DONEPL FFT00440
1000 FORMAT(' EDGE ',14,2X,D16.9) FFT00450
1001 FORMAT(14,2X,2D12.3,' DB') FFT00460
      STOP FFT00470
      END FFT00480

```



```

C*****TIMFRQ*****AUG 1985FFT00560
C                                FFT00570
      SUBROUTINE TIMFRQ(VINP,VOUT,MFT,NFT)                                FFT00580
C                                TIME TO FREQ DFT TRANSFORM                FFT00590
C*****MSSUTHERS*****SFFT00600
      IMPLICIT REAL*8(A-H,O-Z)                                FFT00610
      DIMENSION IWK(14)                                FFT00620
      REAL*8 VALDB(1024),FREQ(1024)                                FFT00630
      COMPLEX*16 VINP(NFT),VOUT(NFT),SAVE(8192)                                FFT00640
      COMMON/DATA1/VALDB,FREQ                                FFT00650
C                                FFT00660
      DO 101 I=1,NFT                                FFT00670
      SAVE(I)=DCONJG(VINP(I))                                FFT00680
      VINP(I)=DCONJG(VINP(I))                                FFT00690
101  CONTINUE                                FFT00700
C                                FFT00710
      CALL FFT2C(VINP,MFT,IWK)                                FFT00720
C                                FFT00730
      DO 102 I=1,NFT                                FFT00740
      VOUT(I)=DCONJG(VINP(I))                                FFT00750
      VINP(I)=SAVE(I)                                FFT00760
102  CONTINUE                                FFT00770
C                                FFT00780
C                                FFT00790
      WRITE(10,901)                                FFT00800
C                                FFT00810
      DO 108 I=1,50                                FFT00820
      WRITE(10,900) I,FREQ(I),VINP(I),VOUT(I)                                FFT00830
C108 CONTINUE                                FFT00840
      900 FORMAT(I5,1X,F6.1,2X,4D12.3)                                FFT00850
      901 FORMAT(7X,'FREQ',9X,'VIN ',8X,'VOUT',/)                                FFT00860
      RETURN                                FFT00870
      END                                FFT00880
C                                FFT00890
C                                FFT00900
C                                FFT00910
C                                FFT00920
C                                FFT00930
C                                FFT00940
C                                FFT00950
C                                FFT00960
C                                FFT00970
C                                FFT00980
C                                FFT00990
C                                FFT01000
C                                FFT01010
C                                FFT01020
C                                FFT01030
C                                FFT01040
C                                FFT01050
C                                FFT01060
C                                FFT01070
C                                FFT01080
C                                FFT01090
C                                FFT01100

```

C*****FRQPLT*****	FFTO1110
SUBROUTINE FRQPLT	FFTO1120
C FREQUENCY PLOTS	FFTO1130
C*****N.BIRKETT NOV8	FFTO1140
IMPLICIT REAL*8(A-H,O-Z)	FFTO1150
REAL*4 XM(1024),Y1(1024),XMIN,XMAX,XSTP,YMN,YMX,YST	FFTO1160
REAL*8 FREQ(1024),VALDB(1024)	FFTO1170
COMMON/DATA1/VALDB,FREQ	FFTO1180
C	FFTO1190
CALL COMPRS	FFTO1200
CALL PAGE(11.0,8.5)	FFTO1210
CALL AREA2D(5.8125,4.6875)	FFTO1220
CALL FRAME	FFTO1230
CALL GRACE(0.0)	FFTO1240
CALL NOCHEK	FFTO1250
C	FFTO1260
CALL XNAME('FREQUENCY(MHZ)\$',100)	FFTO1270
XMIN=0.0	FFTO1280
XMAX=280.0	FFTO1290
XSTP=(XMAX-XMIN)/10.0	FFTO1300
C	FFTO1310
DO 100 I=1,1024	FFTO1320
XM(I)=FREQ(I)	FFTO1330
Y1(I)=VALDB(I)	FFTO1340
100 CONTINUE	FFTO1350
C	FFTO1360
NPTS=1024	FFTO1370
C	FFTO1380
CALL YAXANG(0.0)	FFTO1390
CALL YNAME('MAGNITUDE OF REFLECTION(DB)\$',100)	FFTO1400
YMX=+10.0	FFTO1410
YMN=-90.0	FFTO1420
YST=(YMX-YMN)/10.0DO	FFTO1430
CALL GRAF(XMIN,XSTP,XMAX,YMN,YST,YMX)	FFTO1440
CALL CURVE(XM,Y1,NPTS,0)	FFTO1450
CALL DOT	FFTO1460
CALL GRID(1,1)	FFTO1470
CALL RESET('DOT')	FFTO1480
CALL ENDPL(0)	FFTO1490
RETURN	FFTO1500
END	FFTO1510
	FFTO1520

APPENDIX 2**BUFFER AMPLIFIER SUPER-COMPACT MODELS**

This appendix contains listings of the programs used to model the SAW input and output buffer amplifiers.

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A-2.1 IPBUF	229
A-2.2 OPBUF	232

SUPER-COMPACT Version 1.7 + 001 04/09/84

10:52:47 01/13/88

* IPBUF MODEL

*****TRANSISTOR ELEMENTS*****

BLK

TWO 1 2 3 Q7

A7:3POR 1 2 3

END

BLK

TWO 1 2 3 Q10

A10:3POR 1 2 3

END

BLK

TWO 1 2 3 Q20

A20:3POR 1 2 3

END

BLK

TWO 1 2 3 Q30

A30:3POR 1 2 3

END

*****CIRCUIT ELEMENTS*****

R1:750H

R2:330H

R3:330H

R4:4300H

R5:200H

R6:5600H

R7:360H

R8:330H

R9:1800H

R10:0.10H

R11:8200H

R12:820H

C1:10NF

C2:5.6PF

C3:10NF

*****PARASITIC ELEMENTS*****

LP1:1PH

LP2:1PH

LP3:1PH

LP4:1PH

LP5:1PH

LP6:11NH

CP1:1.0PF

CP2:1.0PF

CP3:1.0PF

CP4:1.0PF

CP5:1.0PF

CP6:10.0PF

W1:0.12CM

*****POWER SUPPLY RAILS*****

BLK

CAP 12 0 C=10NF

CAP 18 0 C=10NF

```

CAP 54 0 C=10NF
CAP 56 0 C=10NF
CAP 34 0 C=20PF
CAP 42 0 C=20PF
TRL 12 34 W=0.10IN P=1.0CM SUB
TPI 34 32 W=W1 P=0.3CM SUB
TRL 32 54 W=0.10IN P=0.5CM SUB
TRL 18 42 W=0.10IN P=0.5CM SUB
TRL 42 55 W=W1 P=0.2CM SUB
TRL 55 40 W=0.10IN P=1.0CM SUB
TRL 40 56 W=W1 P=0.3CM SUB
TRL 56 58 W=0.10IN P=0.5CM SUB
SLC 54 0 L=100UH C=10NF
SLC 58 0 L=100UH C=10NF
*****INPUT MATCHING*****
TRL 2 4 W=W1 P=1.0CM SUB
RES 4 0 R=R1
RES 4 6 R=R2
*****INPUT STAGE Q1*****
IND 12 10 L=LP1
RES 10 8 R=R3
A10 6 8 14
RES 14 16 R=R4
IND 16 18 L=LP2
CAP 6 0 C=CP1
CAP 8 0 C=CP2
CAP 14 0 C=CP3
*****FOLLOWER STAGE Q2*****
RES 14 20 R=R5
RES 20 22 R=R6
CAP 22 0 C=C1
CAP 20 0 C=C2
RES 20 24 R=R7
IND 24 26 L=LP3
A20 26 28 36
RES 32 30 R=R8
IND 30 28 L=LP4
RES 36 38 R=R9
IND 38 40 L=LP5
CAP 26 0 C=CP4
CAP 28 0 C=CP5
CAP 36 0 C=CP6
*****OUTPUT MATCHING*****
CAP 36 44 C=C3
RES 44 46 R=R10
TRL 46 48 W=0.2CM P=1.5CM SUB
IND 48 50 L=LP6
RES 44 52 R=R11
RES 52 0 R=R12
*
IREF:2POR 2 50
END
*
*****FREQUENCY DATA*****
FREQ

```

```
1MHZ
STEP 10MHZ 200MHZ 5MHZ
STEP 300MHZ 1300MHZ 100MHZ
END
*
*****OUTPUT DATA*****
OUT
PRI IPBUF S Z1=75 0 Z2=75 0
PRI IPBUF Z Z1=75 0 Z2=75 0
END
*
*****OPTIMIZATION DATA*****
OPT
IPBUF Z1=75 0 Z2=75 0 F=40MHZ 100MHZ MS21=-6.40dB
END
*
*****TRANSISTOR DATA*****
DATA
Q7: 7MA FILE=NEC85637
Q10: 10MA FILE=NEC85633
Q20: 20MA FILE=NEC85633
Q30: 30MA FILE=NEC85633
SUB: MS H=63MIL ER=4.5 MET1=CU 0.125MIL TAND=0.0001
END
```

SUPER-COMPACT Version 1.7 + 001 04/09/84 10:47:51 01/13/88

* OPBUF MODEL

*****TRANSISTOR ELEMENTS*****

BLK

TWO 1 2 3 Q7

A7:3POR 1 2 3

END

BLK

TWO 1 2 3 Q10

A10:3POR 1 2 3

END

BLK

TWO 1 2 3 Q20

A20:3POR 1 2 3

END

BLK

TWO 1 2 3 Q30

A30:3POR 1 2 3

END

*****CIRCUIT ELEMENTS*****

R1:750H

R2:1500H

R3:120H

R4:81.30H

R5:320H

R6:5100H

R7:320H

R8:2400H

R9:750H

R10:330H

R11:330H

C2:8.2NF

*****PARASITIC ELEMENTS*****

LP1:8NH

LP2:13NH

LP3:1NH

LP4:1NH

LP5:1NH

LP6:1NH

LP11:1.25NH

LP12:10NH

LP13:7NH

LP14:1NH

CP1:0.1PF

CP2:0.1PF

CP3:0.001PF

CP4:0.001PF

CP5:0.001PF

CP6:0.5PF

CP7:0.5PF

W1:0.12CM

*****CIRCUIT DEFINITION*****

BLK

```

CAP 36 0 C=10NF
CAP 60 0 C=10NF
CAP 30 0 C=10NF
CAP 34 0 C=10NF
CAP 38 0 C=20PF
CAP 32 0 C=20PF
TRL 36 38 W=W1 P=0.2CM SUB
TRL 38 40 W=W1 P=0.8CM SUB
TRL 30 32 W=W1 P=1.5CM SUB
TRL 32 34 W=W1 P=1.0CM SUB
TRL 40 60 W=W1 P=1.0CM SUB
SLC 60 0 L=100UH C=8.2NF
SLC 34 0 L=100UH C=8.2NF
*****INPUT MATCHING*****
IND 1 2 L=LP1
RES 2 4 R=R1
TRL 4 6 W=0.100IN P=4.5CM SUB
*****COMMON BASE STAGE Q1*****
RES 6 30 R=R2
SRL 8 0 R=R3 L=LP14
RES 10 5 R=R4
CAP 10 0 C=CP7
IND 5 36 L=LP11
A30 8 10 6
CAP 6 0 C=CP3
*****DARLINGTON STAGE Q2*****
RES 10 12 R=R5
A10 12 13 14
CAP 14 0 C=CP1
CAP 13 0 C=CP4
RES 14 15 R=R6
IND 15 32 L=LP3
IND 13 37 L=LP5
RES 37 38 R=R10
*****DARLINGTON STAGE Q3*****
RES 14 51 R=R7
IND 51 16 L=LP13
RES 16 11 R=2200H
CAP 11 50 C=3.3PF
IND 50 0 L=LP12
A30 16 17 18
CAP 18 0 C=CP2
CAP 17 0 C=CP5
RES 18 19 R=R8
IND 19 34 L=LP4
IND 17 39 L=LP6
RES 39 40 R=R11
*****OUTPUT MATCHING*****
CAP 18 20 C=C2
RES 20 22 R=R9
CAP 20 0 C=CP6
IND 22 24 L=LP2
*
OPBUF:2POR 1 24
END

```



```
*
*****FREQUENCY DATA*****
FREQ
1MHZ
STEP 10MHZ 200MHZ 5MHZ
STEP 300MHZ 1300MHZ 100MHZ
END
*
*****OUTPUT DATA*****
OUT
PRI OPBUF S Z1=75 0 Z2=75 0
PRI OPBUF Z Z1=75 0 Z2=75 0
END
*
*****OPTIMIZATION DATA*****
OPT
OPBUF Z1=75 0 Z2=75 0 F=40MHZ 100MHZ MS21=-6.40DB
END
*
*****TRANSISTOR DATA*****
DATA
Q7: 7MA FILE=NEC85637
Q10: 10MA FILE=NEC85633
Q20: 20MA FILE=NEC85633
Q30: 30MA FILE=NEC85633
SUB: MS H=63MIL ER=4.5 MET1=CU 0.125MIL TAND=0.0001
END
```

FILE: NEC85633 DATA * BELL-NORTHERN RESEARCH

*

*NEC85633 S PARAMETER DATA OBTAINED FROM MANUFACTURER DATA SHEETS

*

7MA 100MHZ 2000MHZ 6 50 1

TRANSISTOR S PARAMETERS FOR VCE=10V, IC=7MA

*

100MHZ	0.77	-43	16.55	145	0.01	71	0.88	-18
200MHZ	0.57	-74	12.28	123	0.04	58	0.70	-28
500MHZ	0.36	-126	6.20	95	0.08	61	0.49	-34
1000MHZ	0.29	-163	3.38	76	0.12	64	0.44	-41
1500MHZ	0.27	171	2.36	62	0.18	66	0.45	-52
2000MHZ	0.30	155	1.89	49	0.23	62	0.45	-64

*

10MA 100MHZ 2000MHZ 6 50 1

TRANSISTOR S PARAMETERS FOR VCE=10V, IC=10MA

*

100MHZ	0.69	-50	20.27	139	0.01	74	0.83	-22
200MHZ	0.48	-82	14.10	117	0.03	59	0.64	-31
500MHZ	0.30	-134	6.67	93	0.07	66	0.44	-33
1000MHZ	0.25	-168	3.59	75	0.13	66	0.41	-41
1500MHZ	0.23	168	2.50	62	0.19	66	0.41	-52
2000MHZ	0.26	152	2.00	50	0.25	62	0.42	-64

*

20MA 100MHZ 2000MHZ 6 50 1

TRANSISTOR S PARAMETERS FOR VCE=10V, IC=20MA

*

100MHZ	0.56	-64	26.42	129	0.01	69	0.72	-27
200MHZ	0.35	-96	16.30	109	0.02	64	0.53	-32
500MHZ	0.24	-147	7.22	89	0.07	71	0.38	-31
1000MHZ	0.21	-175	3.81	73	0.13	69	0.37	-40
1500MHZ	0.20	162	2.64	62	0.20	67	0.38	-53
2000MHZ	0.23	149	2.10	49	0.26	61	0.37	-65

*

30MA 100MHZ 2000MHZ 6 50 1

TRANSISTOR S PARAMETERS FOR VCE=10V, IC=30MA

*

100MHZ	0.49	-72	28.53	124	0.01	68	0.67	-27
200MHZ	0.30	-104	16.78	105	0.02	66	0.49	-30
500MHZ	0.23	-154	7.25	88	0.07	73	0.37	-29
1000MHZ	0.21	-179	3.83	72	0.13	70	0.36	-39
1500MHZ	0.20	160	2.64	61	0.20	67	0.37	-52
2000MHZ	0.23	147	2.11	49	0.26	61	0.37	-65

*

APPENDIX 3

7-TAP ATDE CIRCUIT SCHEMATIC

This appendix contains a circuit schematic of the complete equalizer. The reader is referenced to the appropriate sections in this thesis for a more detailed description of the constituent parts.

7 TAP SAW BASED A/D CONVERTER SCHEMATIC
 DRAWN: NEIL BIRKETT JAN 1988
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