An Adaptive Digital Dynamic Range Controller

by

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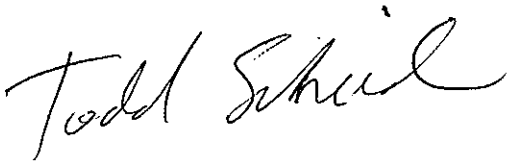

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Abstract

High fidelity digital audio sources are capable of reproducing a much wider dynamic range than most conventional consumer media (eg. AM/FM radio and audio cassettes). The research presented here addresses the problem of matching this wide dynamic range to that of a device (or channel) with lower dynamic range capabilities using a Dynamic Range Controller (DRC). Currently available digital signal processing hardware allows the implementation of entirely Digital DRC's (DDRC's) that interface directly to digital sources and eliminate unnecessary data (analog $\leftrightarrow$ digital) conversions.

The DDRC design presented in this thesis uses an adaptive level measurement scheme and an adaptive recovery time to improve performance. The *perceived* distortion introduced by rapid gain reductions (attack) is lessened by allowing attacks only at the zero crossing preceding a transient. A single-channel version of the Adaptive DDRC has been implemented for real-time operation on a DSP56000 evaluation board.

Tests showed that the Adaptive DDRC has insignificant total harmonic distortion. Intermodulation distortion measurements compare favourably with a previous DDRC design [11] that was reported as having good subjective performance. The results of our listening tests show great promise for the Adaptive DDRC. Listeners rated the average sound quality of an Adaptive DDRC configuration higher than a conventional design (with peak level gain control). However, since other Adaptive DDRC configurations (i.e. different parameter sets) did not perform as well, further testing is required to optimize the Adaptive DDRC parameter set.

Acknowledgements

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Chapter 1

Introduction

1.1 General

With the advent of digital audio equipment (like Compact Disc and Digital Audio Tape), the quality of audio fidelity has increased dramatically. In particular, these digital sources allow the reproduction of wider dynamic range (the ratio of the softest to the loudest sound level) than the majority of previously available analog media. This wide dynamic range enhances the realism of audio programs, but it can cause over-load or over-modulation in some situations.

Typically, the dynamic range of a high fidelity digital source is greater than 90 dB. While this is still less than the dynamic range of a live performance (>100 dB), it still exceeds the dynamic range capabilities of many conventional recording and reproduction methods (Figure 1.1). Since these media are still in use, the problem of matching the dynamic range of the source to that of the media must be addressed. This problem is particularly relevant for radio broadcasting where the robust nature of digital sources is desirable but where their dynamic range can cause over-modulation and a visit from the Department of Communications!

Dynamic Range Controllers (DRC's) have many applications. Typically they are used

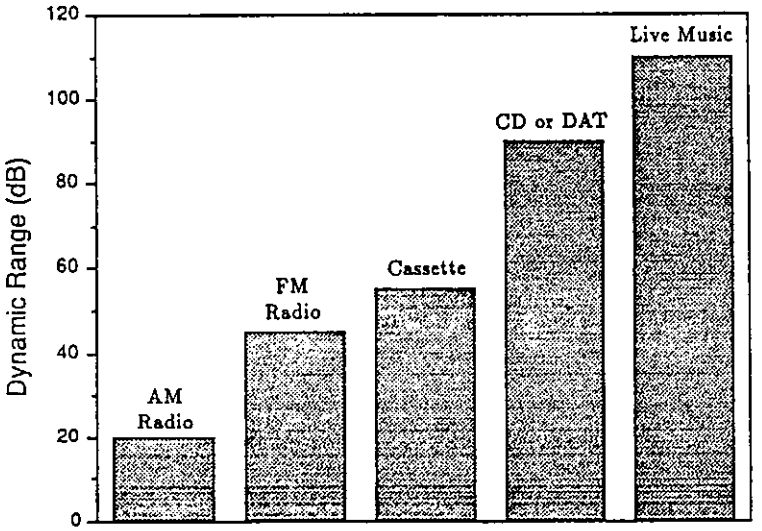


Figure 1.1: Dynamic Range of Various Media

for dynamic range control of signals (music and speech) for recording and broadcasting. They also find application in hearing aids where they are used to compress the dynamic range of a listening environment into the residual dynamic range of a hearing impaired person. Some noise reduction systems [2],[4],[19] use compression and a complementary expansion to increase dynamic range and reduce the media noise for magnetic tape and similar media. By making the amount of compression a function of the background noise, a DRC can also be used to match listening levels to a listener's environment in noisy areas (e.g. car audio).

Dynamic range control is accomplished via a Dynamic Range Controller (DRC). The gain of a DRC varies with time, as a function of the input signal level. A DRC can produce a reduction in dynamic range (know as compression) or an increase in dynamic range (know as expansion). Many DRC's also perform limiting, which is just severe compression. A DRC intentionally amplitude distorts an input signal to reduce the dynamic range while introducing minimum *perceived* distortion. Thus, DRC evaluation is subjective, although objective measurements are used in the initial stages of design.

Extensive work has been done into the design and implementation of DRC's in the past. Recently, some Digital Dynamic Range Controllers (DDRC's), DRC's that utilize

digital signal processing (DSP) methods, have been developed. Using DSP provides flexibility that is not attainable via analog methods. If a DDRC provides a digital input port (e.g. an AES/EBU interface), unnecessary analog-to-digital (A/D) and digital-to-analog (D/A) conversions can also be eliminated. By using at least 24 bit values for intermediate processing very high fidelity can be maintained.

A survey of previous work indicated that older DDRC designs had not exploited the full power of DSP, such as adaptation. All of the designs used (expensive) custom hardware. With the powerful single-chip DSP processors available, it is now possible to implement an adaptive DDRC using *off-the-shelf* signal processing hardware.

1.2 Overview

This thesis covers the design and implementation of a single-channel adaptive DDRC. The design uses adaptive recovery time and level measurement to obtain improved performance. To reduce distortion, sudden gain reductions (known as attack) occur only at zero crossings of the output signal. The design provides wide frequency response (40Hz to 20kHz) and uses all of the capabilities of the Motorola DSP56000 to achieve the best sound fidelity possible.

The goal of this research is to design and implement (in real-time) a flexible single-channel DDRC on a single DSP processor. The possiblies for real-time adaptation of the recovery time and signal level measurement are also explored. Because of the large number of user adjustable parameters, only simple tests are conducted to confirm operation and compare the performance of non-adaptive and adaptive DDRC's. No attempt is made to arrive at the "best" set of parameters since this is dependent on user requirements.

The thesis is divided into six chapters. Chapter two introduces the terminology necessary to understand the design of the adaptive DDRC. It also covers basic topologies and presents a brief survey of previous research into DDRC's. The high-level design of

the Adaptive DDRC (ADDRC) is presented in Chapter three. Heuristic adaptation rules and the rationale for them are presented here. Chapter four covers the real-time implementation of the ADDRC on the Motorola DSP56000 Evaluation Module. The trade-offs and compromises made for the real-time implementation are examined. The results of subjective and objective tests are presented in Chapter five. Chapter six summarizes the test results, presents conclusions and suggests further work to improve the ADDRC design.

Chapter 2

Background

2.1 General

The simplest form of dynamic range control is a volume (or level) control that is adjusted (over time) by a human operator to maintain a sound level that is not "buried" in channel (or media) noise and does not introduce objectionable distortion as a result of clipping. However, a human operator is very slow acting and would be likely to disregard short, isolated transients. A better approach would be to design a device (a DRC) that automatically measures the signal and adjusts its gain to maintain a signal that is above the noise floor and below clipping. To achieve good performance, the DRC should perceptually fool the listener into thinking that the processed signal is almost the same as the original.

In operation, a DRC measures the input and/or output signal to obtain a set of parameters used to determine the gain applied to the input signal. It is desirable that the parameters have some psychoacoustic significance. However, knowledge of psychoacoustics is so limited that any parameter set will be a rough approximation to the ideal set of perceptually relevant parameters.

The channel (or media) characteristics are well understood. The channel has a maximum level above which clipping will occur and a minimum level below which the signal will be inaudible over noise. The dynamic range of the channel is the ratio of the clipping level to the minimum audible level (often called the noise floor). The difference between the dynamic range of the channel and the dynamic range of the signal (computed in a manner similar the the channel dynamic range) is the difference in amplification for high and low level signals. This is equivalent to the DRC output for periodic input waveforms (i.e. steady state). These characteristics are specified by the **static parameters.**

The temporal characteristics of the gain must also be considered. The gain must change over time—otherwise the DRC will be nothing more than a volume control. The onset and duration of gain changes are specified by the **dynamic characteristics.** The dynamic characteristics are related to the modulation distortion, effective compression ratio and noise masking capabilities of the DRC.

When specifying the characteristics of a DRC, the application must also be considered. DRC's are used for a number of basic tasks:

1. maintain output peaks below a specified level (peak limiting)
2. reduce output dynamic range (compression)
3. increase output dynamic range (expansion)
4. eliminate all signal below a specified level (noise gate)
5. any combination of the above

These applications place restrictions on the performance requirements that must be considered when selecting a topology and specifying static and dynamic characteristics. For example, if a DRC is to be used for peak limiting, the peak level must always be kept below the maximum output level. Clearly, it is desirable to design a DRC that can perform all of the above functions.

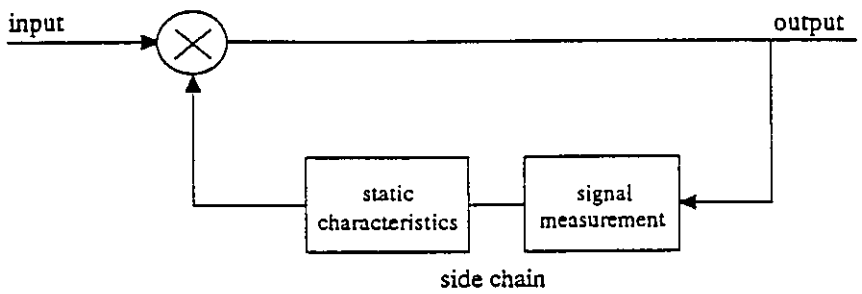


Figure 2.1: Negative Feedback DRC

2.2 Topologies

The first DRC's were constructed using negative feedback control loops or side chains (Figure 2.1). Because negative feedback linearizes and stabilizes the feedback loop, the exact shape of the side chain characteristics are unimportant. However, the side chain propagation delay allows a transient to reach the output before the gain is reduced. Thus, it is inevitable that overshoots will reach the output. This is a serious problem for peak limiter designs.

A feedforward system with a delay in the forward signal path can eliminate this overshoot problem (Figure 2.2). The delay in the signal path accounts for the side chain propagation delay. Thus, the gain can act to suppress a transient before if reaches the output. This also improves the subjective performance [13]. This design has the disadvantage that the linearity and stability of the side chain must be precisely controlled. This presents a challenge for analog implementations, but not for DSP methods.

Feedforward and feedback topologies may be combined to yield a structure that does not require variable gain elements with accurately specified characteristics (Figure 2.3). With this design, the gain control elements must be matched. This is still a difficult

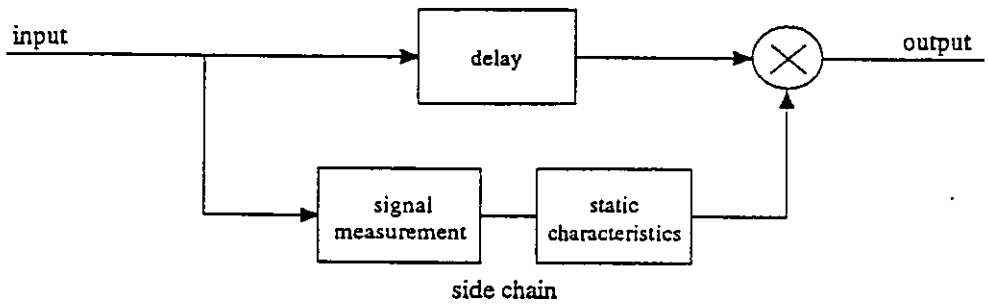


Figure 2.2: Feedforward DRC

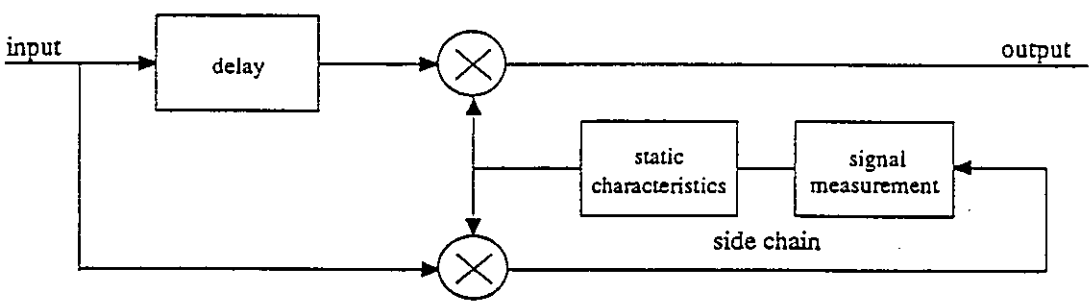


Figure 2.3: Feedback DRC with Duplicate Gain Stage

task but it is easier to accomplish than accurately specifying a single characteristic. This technique is used successfully in modern high-performance compressors and limiters.

Both the feedforward and combined methods require a delay in the signal path. In analog designs, it is costly and difficult to generate a high-fidelity delay (this is typically done using CCD delay lines). Using DSP, accurate control over the side chain characteristics is easy to achieve if high precision fixed point arithmetic is used (24 bits). Also, it is easy to generate a delay.

Olivera [14] showed that feedforward designs are superior to feedback designs because they can achieve infinite compression (i.e. limiting) with finite side chain gain while a feedback design requires infinite gain. For large compression ratios, a feedback design operates at almost open-loop, a situation that can lead to instability. As well, as the amount of compression is changed in a feedback system, the dynamic characteristics are altered.

2.3 Static Characteristics

The static characteristics specify the steady-state performance of the DRC. They describe the "instantaneous" input level versus output level (in dB) relationship and do not consider the temporal variations of the DRC gain. To match the input signal dynamic range to that of the channel (or media), the difference in the dynamic range between the channel and the program must equal the difference in amplification for high and low-level signals.

This difference specifies the range of the static characteristics, but it provides no information about the distribution of the gain. The gain distribution is an important consideration because it will affect the output program quality. For example, if we were to compress a signal (reduce its dynamic range) we might (1) apply a gain of one to peak signals and amplify the low-level signals or (2) attenuate peaks and amplify low-level

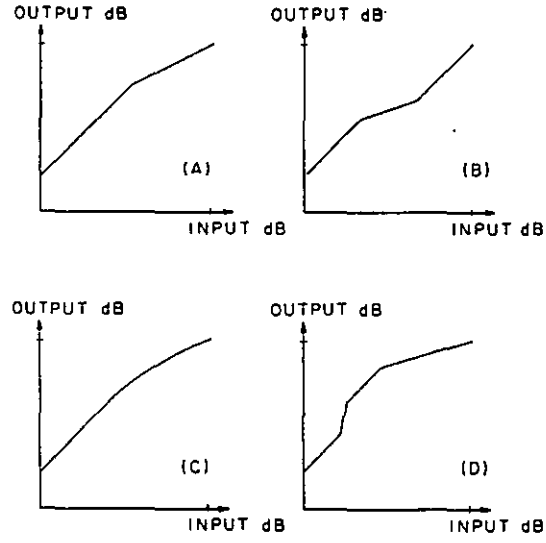


Figure 2.4: Possible Static Characteristics

signals somewhat. We may concentrate the gain variation or spread it across the entire input range. Because the gain distribution is user and input signal dependent, there is no optimum solution to this problem. We can specify any number of characteristics that provide the correct total dynamic range compression (Figure 2.4 [3]).

Static characteristics are typically specified using ratios and thresholds:

1. **Ratio** (R): The ratio between changes in level (measured in dB) at the input and output of the DRC. R is the slope of the static characteristics (output versus input).
2. **Threshold**: The input signal level where the static characteristics change slope (i.e. a new Ratio).

The definitions can be used to formalize the notions of compression, expansion and limiting:

- **Compression:** $0 < R < 1$
- **Expansion:** $R > 1$
- **Limiting:** Ideally, $R = 0$ but typical designs use $0 < R < 0.01$.

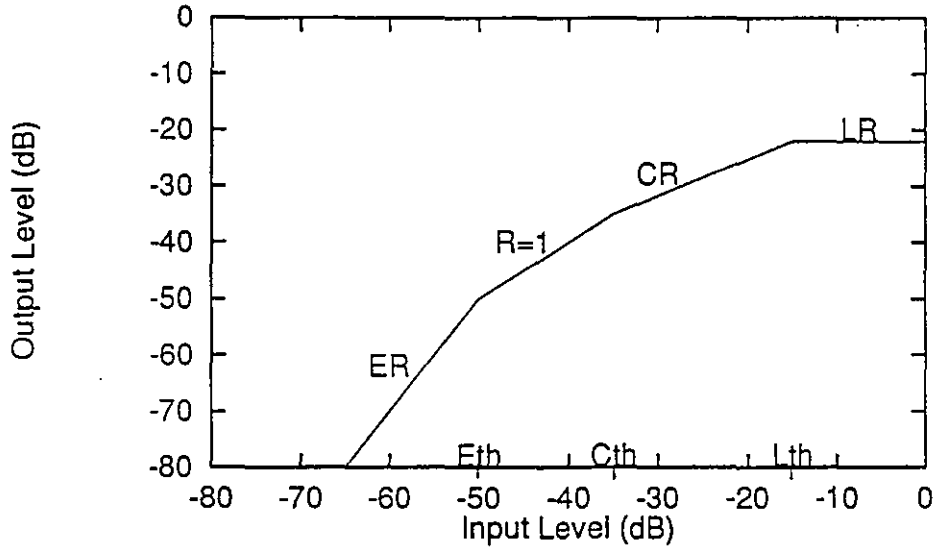


Figure 2.5: Four Region Static Characteristics

The properties of level independence and low-level expansion have been shown to be desirable attributes of static characteristics [12],[18],[3].

1. **Level Independence:** Static characteristics where R is constant (i.e. straightline on a dB scale) between thresholds. That is, R is not a function of the input level.

2. **Low-level Expansion:** Whenever a signal is compressed, the signal-to-noise ratio (SNR) is reduced by the amount that the signal is compressed. Low-level expansion is used to restore the SNR to its original value by suppressing (low-level) signals below an expansion threshold. This threshold is typically set just above the noise floor.

In principle, a large number of regions (each with a ratio and threshold) can be used as a piece-wise linear approximation to any desired characteristic. However, many designs incorporate these feature into static characteristics with four regions (Figure 2.5). These characteristics are simple to implement and they have been used successfully in many compressor/limiter designs in the past [3],[12] [18]. Four region characteristics have three thresholds:

- **Expansion Threshold** (E_{th}): Below this threshold expansion is used to maintain the SNR.
- **Compression Threshold** (C_{th}): Above this threshold (and below the limit threshold) compression is used the reduce the output signal dynamic range.
- **Limit Threshold** (L_{th}): Above this threshold (and below the maximum input level) limiting keeps the output from exceeding the maximum input to the channel.

There is also a **no-action** region between the expansion and compression thresholds. In this region the signal is passed with a gain of one. The ratios within all other regions are within the ranges listed previously. Some designs also use characteristics with "soft-knees" [14]. That is, the ratios change asymptotically from one region to another with smooth transitions as opposed to sharp breaks.

Let X be the input level in decibels (dB) and Y be the output level in dB.

$$X = 20\log_{10}(\frac{V_{in}}{V_{max}})$$

$$Y = 20\log_{10}(\frac{V_{out}}{V_{max}})$$

The input and output are both scaled by the maximum input level, which is equal to $32767/32768 \simeq 1.0$ for 16 bit fractional twos complement arithmetic. Within each region, the input-output relationship is

$$Y = RX$$

where R is the ratio. Since $V_{max} \simeq 1.0$, this can be rewritten as

$$V_{out} = V_{in}^{R}$$

We see that equal level differences on the input are mapped to equal level differences on the output that are R times smaller.

The DRC output signal is computed as $V_{out} = V_{in} \times G$ where G is the DRC gain. The gains within each region are as shown below, where R_E is the expansion ratio; R_C is the compression ratio and R_L is the limit ratio:

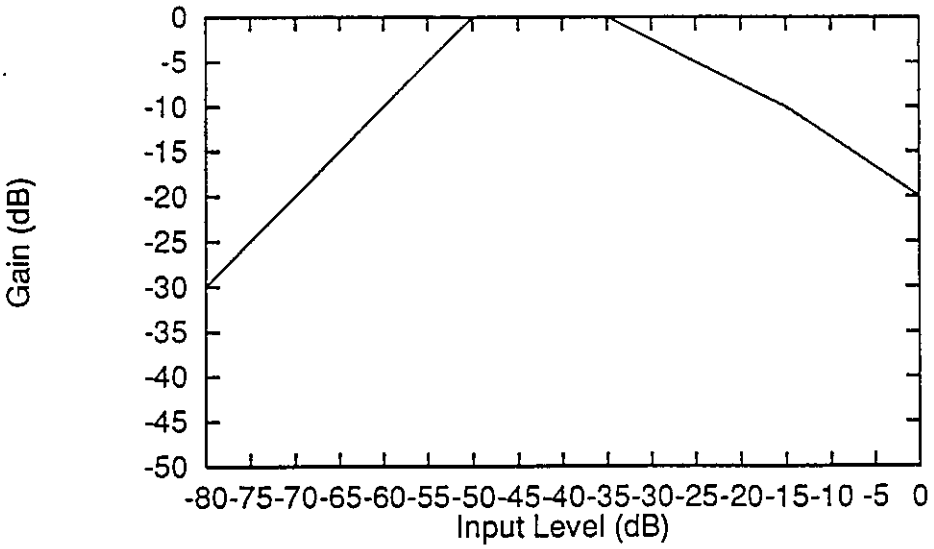


Figure 2.6: Gain for Four Region Static Characteristics

- **Expansion:** $G_{dB} = (R_E - 1)(X - E_{th}) = ES(X - E_{th})$
- **Compression:** $G_{dB} = (R_C - 1)(X - C_{th}) = CS(X - C_{th})$
- **No-action:** $G_{dB} = 0$
- **Limiting:**

$$\begin{aligned} G_{dB} &= (R_L - 1)(X - L_{th}) + (R_C - 1)(L_{th} - C_{th}) \\ &= LS(X - L_{th}) + CS(L_{th} - C_{th}) \end{aligned}$$

where $ES = R_E - 1$ is the expansion slope. The limit slope and compression slopes are defined in a similar manner with their respective thresholds. The gain expressions for the expansion and compression regions are similar in form. The gain expression for the limiting region contains an additional term that accounts for the gain applied by the compression before the limit threshold. Table 2.1 shows typical static characteristics for a DRC. Figure 2.5 shows the input-output relationships for these characteristics. Figure 2.6 shows the input-gain relationship for the same characteristics.

Region	Ratio (R)	Threshold
Expansion	2	$E_{th} = -50dB$
No-action	1	—
Compression	0.5	$C_{th} = -35dB$
Limiting	0.01	$L_{th} = -15dB$

Table 2.1: Typical Static Characteristics

2.4 Dynamic Characteristics

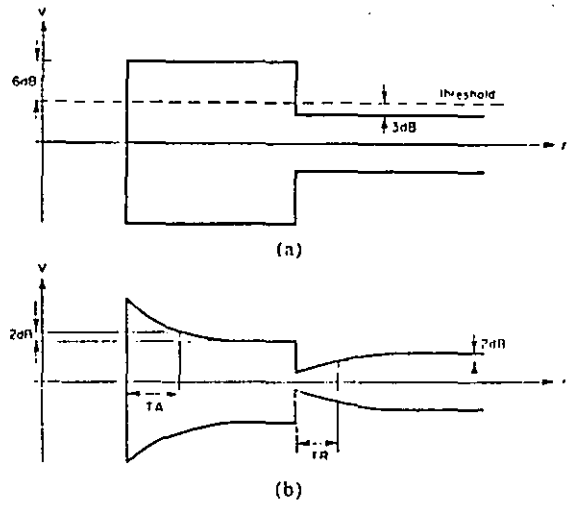
As indicated previously, the dynamic characteristics describe the temporal variations of the DRC gain. When specifying the dynamic characteristics, there are three basic questions (originally suggested by Blesser [3]) that must be answered:

1. When should the gain change?
2. How quickly should the gain increase?
3. How quickly should the gain decrease?

Blesser also presents three rules to aid in the design of the dynamic characteristics:

1. The gain should not change during the duration of a single note to ensure that the output is a faithful reproduction of the input.
2. Transients should not dominate the program (i.e. the gain control)
3. The subjective balance between different musical notes should appear the same even if the actual balance has changed.

Because the gain is computed instantaneously by the static characteristics, the dynamic characteristics are always included in the measurement section. All designs use



Specification of attack and recovery times. (a) Envelope of input signal. (b) Envelope of output signal.

Figure 2.7: Attack and Recovery Times

either peak detection, average signal level or RMS signal level for gain control. Each of these methods is characterized by an attack time and a recovery time. A hold time is used in some designs.

2.4.1 Attack Time

Attack times have been subject to a number of definitions on the past [12]. For our work, the attack time (T_a) is defined as the time taken for 63.2% of the total change in gain when the input exceeds a DRC threshold (Figure 2.7 [12]). This definition is consistent with the conventional notion of a time constant.

Most previous designs have used a first-order exponential attack time that is typically less than 10ms for compression/limiting of musical signals. This value is a compromise. A short T_a makes the output look more like the input while a long T_a accentuates the initial part of a transient. Short attack times (less than 10ms) are generally not perceived; however, the gain tends to be determined by the peaks in the program. Since the ear responds to loudness and not peaks, signals with high-level peaks are not necessarily loud. Under some conditions, this can produce "a very *flat* sound with an apparently negative compression ratio" [3]. This problem is typically solved by using an attack time

of approximately 5ms. Extensive subjective tests conducted by Wagenaars et al. [20] agree with this value.

A recent peak limiter design [11] has taken a different approach. In this design the gain attacks at zero crossings. "This makes it possible to obtain an instantaneous attack without distortion of the waveform peak." The reports on the subjective performance of this device are very promising.

2.4.2 Recovery Time

As with attack times, recovery times have been defined in a number of ways in the past [12]. To be consistent with our definition of attack time, the recovery time (T_r) is defined as the time taken for the DRC gain to achieve 63.2% of the total change in gain when the input falls below a threshold (Figure 2.7). This definition is consistent with the notion of a first-order time constant. Most recovery characteristics are first-order exponential.

Typically, DRC's use a long T_r to reduce the distortion introduced by the gain amplitude modulating the signal. However, with a short T_a and a long T_r, a single high-level transient can create a "hole" in the program by blanking out the musical program following it. Also, a very long T_r will change the gain so slowly that the compression/limiting will be more like volume control. A short T_r can also lead to an exaggeration of speech and breath noises and modulation of otherwise constant amplitude background noise. These phenomena are often called "gain pumping," "breathing" and "swishing" [11]. Thus, if a fixed T_r is used, a compromise between modulation distortion, the creation of holes in the program, effective signal compression and gain pumping must be made.

Compromise T_r's on the order of 200-250ms are typically used. A study by Wagenaars et al. [20] found that a T_r of approximately 200ms[1] produced good subjective results.

[1] This design[18] used a 3^{rd}-order recovery stage with $T_{r3} = 62.5ms$ to "achieve a kind of hold effect." An equivalent 1^{st}-order recovery time is $T_{r1} = 3.25T_{r3} \simeq 203ms$

In a classic paper on the design of an adaptive compressor/limiter [3], Blesser argues that "a single recovery time does not take into account the psychophysical characteristics of hearing." He considered the dynamic range of a musical program in two components:

- **Short-term dynamic range** - The ratio of maximum to minimum level over about a second. It corresponds to the "fullness" or "muddiness" of a piece of music.
- **Long-term dynamic range** - The ratio of maximum to minimum level over a thirty second interval. It corresponds to the "mood" of a piece.

If there is a large discrepancy between the ratio of the short- and long-term dynamic range of the input to that of the output, the music does not sound natural. When a DRC compresses or limits a piece of music, the long-term dynamic range is unaffected by T_r; it is determined by the static compression ratio adjustment. The short-term dynamic range depends on T_r.

Consider a piece of music that is heavily compressed using long recovery time. The gain changes very slowly after a high peak and for normal programs the gain is constant. Thus, there is no change in the dynamic range (i.e. we have implemented a volume control). The gain must vary to produce effective compression. Blesser shows that the effective compression ratio (as a percentage, 100% corresponding to maximum compression) can be written as

$$CR_e = 1 + \frac{(CR_s - 1)T}{kT_r}$$

for $T < kT_r$, where T is the time interval between peaks; CR_s is the static compression ratio; CR_e is the effective compression ratio and k is some constant. The relationship shows that the recovery time function has only one degree of freedom—one parameter determines all of its properties. Clearly this is undesirable.

To overcome this difficulty, Blesser split T_r into two parts, a short-term T_r and a long-term T_r. The short-term T_r controls the recovery of a percentage of the gain change,

P_s, caused by the last peak. The long-term T_r controls the recovery of the remaining percentage, $1 - P_s$. He also includes a **hold time** (T_h) that keeps the gain constant (for $T_h = 100ms$) until most of the musical event that caused the gain change has passed. If a new peak arrives during the hold time, the recovery is again postponed. The hold time allows the compression section to decide if it really "wants" to recover.

The essence of good dynamic range compression is the proper adjustment of P_s. Empirically, Blesser found that making P_s a piecewise linear function of T (the time between peaks) produced good results. He also allowed the peak-to-average loudness ratio of the program to control T_r so that a program's original dynamic range determined how much compression was used. It is not revealed how this is accomplished in the actual design.

Blesser also includes an adaptive T_r in the limiting section of his DRC design. He sets another threshold 2dB below the limit threshold (L_{th}). The frequency with which peaks enter this window (i.e. between L_{th} and L_{th}-2dB) is used to adjust T_r. For a single peak, the DRC reduces the gain and then immediately returns it to its previous value. For repetitive peaks, the DRC reduces the gain for the first peak and then observes that succeeding peaks lie within the window. T_r then increases and the gain remains constant. T_r varies from 150ms to more than 30s in limiting.

Mapes-Riordan and Leach [11] developed a digital peak limiter that recovers in fixed amounts at zero crossings of the signal. They explain that this "makes the recovery time inversely proportional to frequency, thus minimizing the problems associated with a fixed recovery time." Thus, distortion of low-frequency signals caused by too short a recovery time is minimized while program "holes" and "dropouts" caused by short duration peaks are minimized. They report good subjective results with this design.

McNally [12] proposes a scheme for adapting T_r (Figure 2.8). T_r is adapted based on the peak and RMS level of the input signal so that a short T_r is used for isolated peaks but with signals of higher average level, the T_r is extended to avoid excessive limiting or

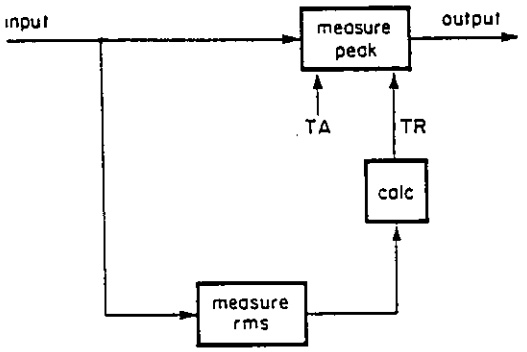


Figure 2.8: Block Diagram for McNally's Autorecovery Method

compression. He terms this method "autorecovery."

2.5 Measurement Methods

In the past, DRC designs have used three methods of signal measurement for gain control: peak detection, average signal level or RMS signal level. Each of these methods has perceptual consequences and an effect on the attack and recovery times of the DRC gain.

2.5.1 Peak Detection

Peak level detection provides a simple method of achieving a short T_a and a long T_r. It ensures that the output peak level never exceeds the maximum input level of the channel. Most peak measurement systems are constructed as shown in Figure 2.9. In this design, R_a controls T_a and R_r controls T_r. Typically, $R_r >> R_a$ so $T_a \simeq R_a C$ and $T_r \simeq R_r C$.

A digital implementation of a peak detector is shown in Figure 2.10. Here, TA controls T_a and TR controls T_r. The non-linear element that receives the difference of the full-wave rectified input ($x[n]$) and the delayed output ($y[n]$) simulates an ideal diode using

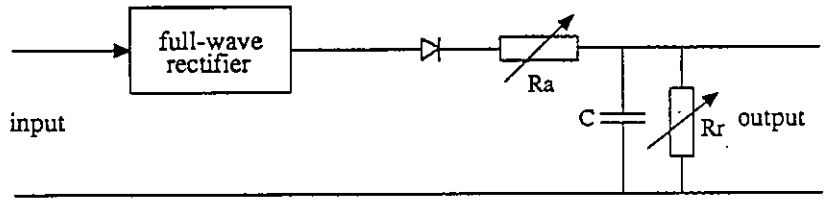


Figure 2.9: Block Diagram of Peak Detector

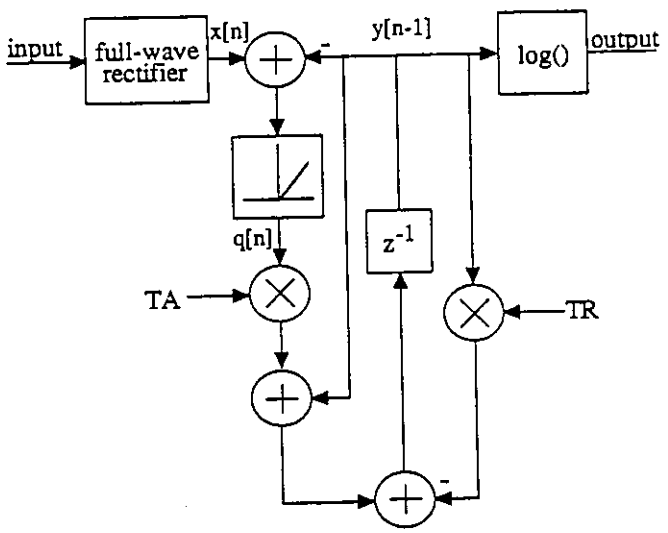


Figure 2.10: Block Diagram of Digital Peak Detector

$$\begin{aligned}
&\text{if } x[n] - y[n] > 0 \\
&\qquad q[n] = x[n] - y[n] \\
&\text{else} \\
&\qquad q[n] = 0
\end{aligned}$$

In their design of a digital peak limiter, Mapes-Riordan and Leach [11] use the peaks between zero crossings and compute a gain to reduce a peak below the maximum input to the channel if the peak exceeds the limit threshold.

Using peak level for gain control has the disadvantage that the ear does not respond to the peak level of a signal. The ears response resembles a RMS or average level measurement [5]. Thus, peak level control causes gain changes to be perceived as being unrelated to program content.

2.5.2 Average and RMS Level

Some designs have used average or RMS levels for gain control. Generally this improves the perceptual performance of a DRC because the gain control is better related to the perceived signal level [1]. However, to approximate the ear's response, averaging times that are much longer than the attack times required to suppress transients are used. Thus, transients exceeding the maximum channel input may appear on the output of the DRC.

These methods also have the disadvantage that $T_r = T_a$ for most simple implementations. Figure 2.11 shows a digital implementation of a RMS level measurement. To implement an average measurement, the squaring operation on the input is replaced with a full-wave rectifier (i.e. absolute value).

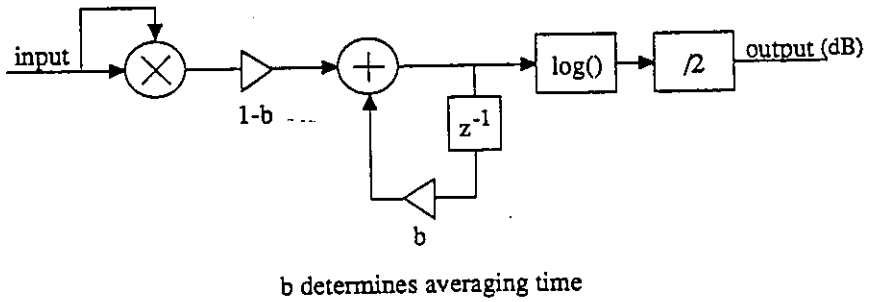


Figure 2.11: Block Diagram of RMS Level Detector

2.6 Summary

Previous research has been done in the design and implementation of DRC's. This chapter presented general background on a number of designs and introduced the terminology required to understand the operation of a DRC.

Most recent DDRC designs have used custom hardware. Typically, they have been re-implementations of traditional designs. These designs have made a number of compromises that will be addressed in our improved DDRC design that is presented in the next chapter.

Chapter 3

Design of an Improved DDRC

3.1 General

Previous DRC designs have been implemented via analog or digital methods. We selected DSP as the implementation method because it provides powerful processing options like adaptation that are difficult to implement using other methods. Also, DSP-based designs are immune to temperature variations; they do not depend on component tolerances and are generally more stable than analog implementations.

A feedforward topology was selected because it offers a number of advantages over other options (Section 2.2). The side chain linearity requirement of this topology does not pose any serious problems in a DSP-based implementation.

The four region static characteristics described in Section 2.3 will be used. These characteristics are reasonably easy to implement and they have been used successfully in many compressor/limiter designs in the past [12],[3],[18].

Because musical signals are non-stationary, it is not possible to select a set of fixed parameters (e.g. T_r and the time constants for level measurements) that will provide good performance for all input signals (or even different portions of the same input signal).

With fixed parameters, a compromise must always be made. To improve performance, our design is adaptive.

The input signal and peak level are buffered so adaptation can be based on the future (with respect to the output) statistics of the input signal. The level measurement and T_r adapt based on input signal statistics. Novel attack characteristics are also used to improve performance. Figure 3.1 shows a block diagram of the adaptive DDRC. The remainder of this chapter describes the high-level design of the adaptive T_r section, the adaptive level measurement and the attack portion of the adaptive DDRC design.

3.1.1 Statistics Selection

A set of input signal statistics to control the adaptation must be selected. Ideally, a set of perceptually relevant, orthogonal input signal statistics should be used. This is impossible to derive except for very simple inputs so an approximation to the ideal set of statistics must be used. In a real-time implementation, the statistics must be computed rapidly. Thus, a set of easy to compute statistics that have some perceptual bearing on the signal or that are quantities we must control will govern the adaptation of T_r and the level measurement.

After some background experimentation and consideration of the issues involved in real-time computation, the statistics listed below were selected for use in the adaptive DDRC. Further explanation of the rationale behind the selection of these statistics is presented in later sections of this chapter.

1. **Peak Level:** The peak level between zero crossings. This is a quantity that we wish to control. Zero crossings provide a convenient block size for the determination of peak level.

2. **Average Level:** This is the local level of the signal. It is related to the perceived signal level. An averaging time of 100ms will be used.

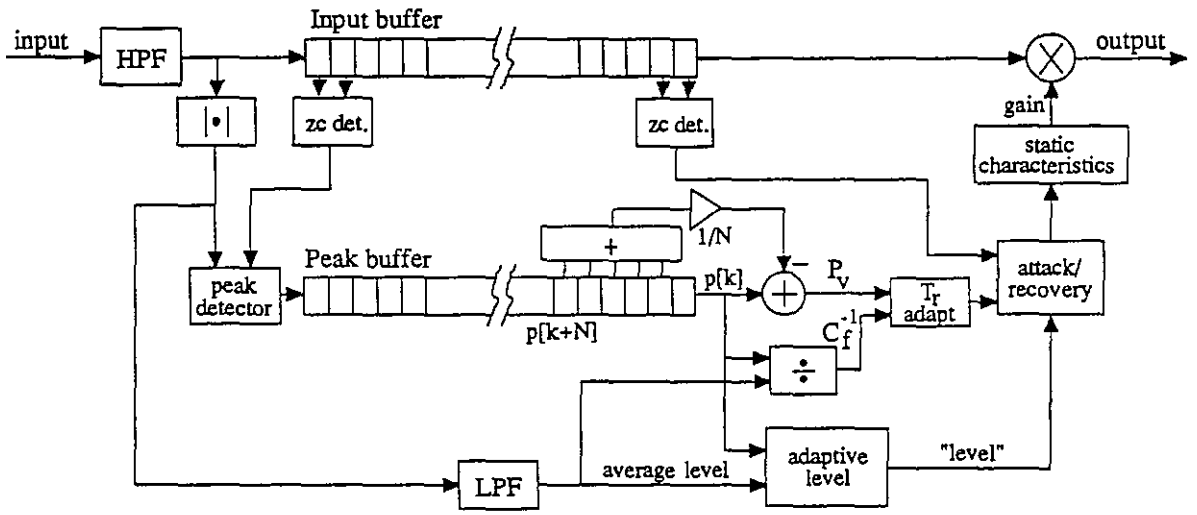


Figure 3.1: Block Diagram of the Adaptive DDRC

3. **Crest Factor** (C_f): This is the ratio of the peak level to the RMS level of the signal. Our design will approximate this using,

$$C_f^{-1} = \frac{\bar{x}}{p[k]}$$

where $p[k]$ is the present peak level between zero crossings and $\bar{x}$ is the average input level. This quantity is easier to compute in real-time and it is always fractional since $p[k] \geq \bar{x}$. This statistic provides information about the "peakiness" of the original signal (i.e. the signal's short-term dynamic range).

4. **Peak Variation** (P_v): This value is computed as,

$$P_v = p[k] - \frac{1}{N} \sum_{n=1}^{n=N} p[k+n]$$

where $p[k]$ is the present peak level. It provides a rough measure of the *isolation* of a peak level by indicating how the current peak level compares to the future average peak level (Figure 3.2).

The peak variation is computed using forward averages. The input samples and peak level are buffered so that statistics based on the future signal properties (with respect to the output) can be computed.

Although the RMS signal level is better related to the perceived signal level [5], the DDRC will use the average signal level because it is easier to compute and does not suffer the underflow problems associated with squaring fractional input samples[1]. It is possible to compute all of these statistics in real-time using relatively simple computations. These measurements will be used to adapt T_r and the level measurement.

3.1.2 Design Method

To control the adaptation of the DDRC, adaptation rules (or equations) will be developed. The adaptation must be computed in real-time, so it is imperative that the rules be simple.

[1]The DSP56000 uses fractional twos-complement arithmetic

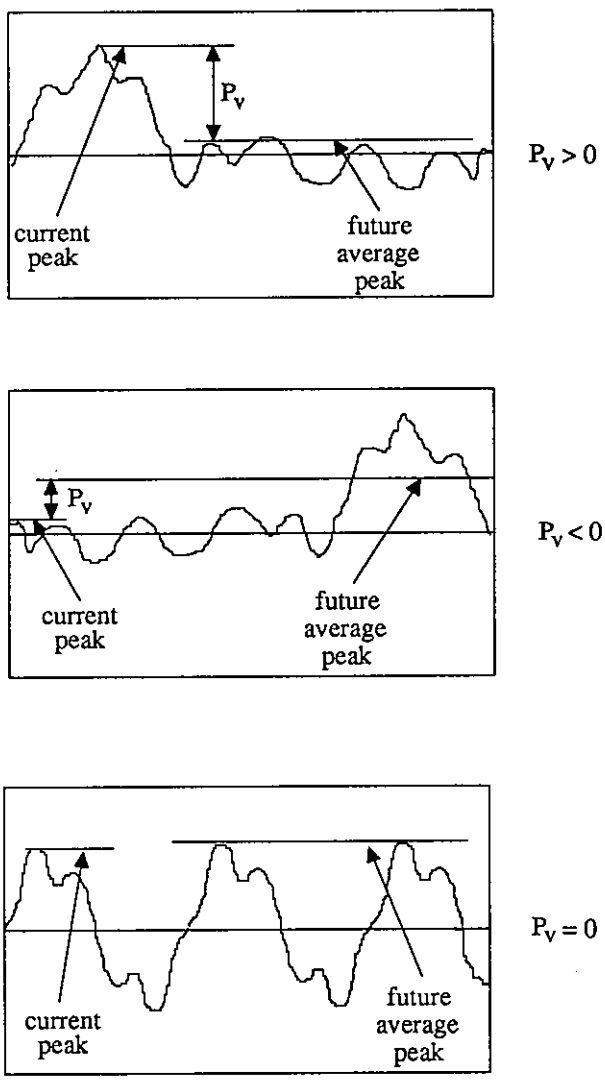


Figure 3.2: Calculation of Peak Variation

It is very difficult to specify the required performance mathematically—this makes it impossible to use conventional adaptation methods like least mean square algorithm [17]. Thus, heuristic methods will be employed.

Two approaches for the development of rules were explored. We first attempted to classify all input signals based on the set of statistics presented above. Each statistic was split into three ranges. If we assume (pretend) that the statistics presented above are orthogonal, this is equivalent to partitioning a four dimensional measurement "space" and specifying a set of parameters (to control the adaptation) for each region. Although, this method provides an organized methodical approach to the adaptation design, it requires the consideration of too many classes. Thus, this method was not used.

A second approach proved more successful. This method can be split into five steps:

1. Specify the problems to be solved by the adaptation.
2. Consider how adjustment of the available parameters can be used to compensate for these situations.
3. Determine the input statistics that indicate these problems.
4. Relate these statistics to the parameter adjustments required to solve the problems.
5. Develop simple (heuristic) rules based on the relationships derived in step 4.

This approach is used to develop the adaptation rules for T_r and the adaptive level.

3.2 Recovery Time Adaptation

Non-adaptive DRC's have the best performance when they use short T_a's and long T_r's. A long T_r ($> 250ms$) is required for low modulation distortion and to maintain constant short-term dynamic range [3]. However, this set of parameters allows isolated transients

to cause "holes" in the musical program by rapidly reducing the gain and then recovering slowly. If a short T_r is used, repetitive high-level peaks will cause excessive distortion by applying a rapidly changing gain to the output signal (i.e. modulation distortion). Most DRC's use a compromise T_r of 200 to 250ms to balance these difficulties.

If T_r is adaptive, a compromise will not be necessary and improved performance will result. The goals of adapting T_r are summarized below:

1. Maintain the relative (short-term) dynamic range of the output signal whenever possible by using a long T_r (e.g. $T_r = 200ms$).
2. Use a short T_r following isolated high-level peaks to eliminate "holes" in the musical program.
3. For repetitive peaks, use a long T_r (in effect, a hold time) to avoid excessive modulation distortion caused by frequent gain changes.

The adaptation algorithm will use these rules to adapt T_r based on the input statistics. Like most DRC's our design will use a first-order exponential recovery.

T_r should be a function of the "peakiness" (relative peak size) of the input signal and the "isolation" of these peak levels. C_f is a measure of the relative peak size of a signal (in relation to the average level). P_v is a measure of the isolation of a peak level. The input signal properties reflected in each of these statistics and their relation to T_r is considered below.

- **Crest Factor (C_f)**: A large C_f indicates a signal with large peaks relative to its average level. For this signal a short T_r will avoid "holes" in the program. A small C_f indicates a signal with low "peakiness". For this class of signal, a long T_r will reduce modulation distortion and will not introduce "holes" in the program.
- **Peak Variation (P_v)**: A positive P_v indicates that the future average peak level is less that the present peak level. For the situation, a short T_r is required because

the present peak is isolated. A negative P_v indicates that the average peak level is increasing; this calls for a long T_r to reduce gain modulation (in effect a hold time). A $P_v = 0$ indicates no change in the present a future average peak level. This requires a long T_r.

From this discussion, we see that T_r is inversely related to both C_f and P_v.

3.2.1 Adaptation Equation

To design the rules for the control of T_r, the priority of these statistics must be determined. P_v is not important if the signal has a low peak level. C_f is a more important statistic in the adaptation of T_r because it provides information about the relative size of the signal peaks. If an input signal has a high C_f, additional information is required to adapt T_r. If the peak is isolated, T_r should be short to avoid causing a "hole" in to program. Conversely, if the peak is not isolated, the gain should recover slowly (in effect a hold time) to reduce gain modulation. P_v provides the information about the isolation of the peak level. Because the DDRC will be operating from low-noise digital sources, excessive background noise will not be a problem and the noise masking properties of the recovery time are not a consideration.

Typically, the peak and RMS levels of musical signals differ by 10dB [7] (i.e. $C_f \simeq 3.2$). For low C_f signals, a fixed T_r will be used because the signal has low "peakiness"—a fixed T_r can be used without causing holes in the program. For higher C_f signals (i.e $C_f > 3.2$), T_r is adapted based on C_f and P_v. A linear adaptation equation is used to simplify real-time computation. By using C_f^{-1} in the adaptation, the inverse relationship between T_r and C_f can be obtained. C_f^{-1} is always fractional—an important consideration for implementation on the DSP56000 which uses fractional arithmetic. The adaptation rule used in the design is

if $C_f < 3.2$

T_r is constant

else

$$T_r \propto k_1(C_f^{-1}) - k_2(Pv) + c$$

In this equation, $k_1, k_2 \geq 0$. A negative sign in front of $k_2 P_v$ provides the inverse relationship between T_r and P_v as described above. The constant c controls the starting point of the adaptation while k_1 and k_2 control the range of the adaptation. The values used for these parameters are derived in Chapter 4. The range of the T_r adaptation will typically be $50ms \leq T_r \leq 300ms$ or so. This range will be refined through listening tests.

3.3 Level Adaptation

Previous DDRC designs have used either peak or RMS signal level for gain control [12],[18],[11]. As discussed in Chapter 2, peak detection is a simple-to-compute measure that ensures the channel will never be overdriven. However, the peak signal level has a weak relation to the perceived signal level. This can lead to undesirable results when compressing/limiting signals [3].

The average or RMS signal level is better measure of the perceived level of the signal. However, these measures do not respond quickly to transients because of their inherent averaging. Thus, the peak output signal level may exceed the maximum input to the channel (the *saturation level*) for isolated transients.

In our design, the peak and average signal levels will be adaptively combined so that, whenever possible, the average signal level is used for gain control. Using the average level will provide improved perceptual performance because the gain will be better related to the perceived signal level. Peak level or a linear combination of peak and average level will be used when necessary to ensure that the peak level does not exceed the saturation level. In summary, the goals of this design are:

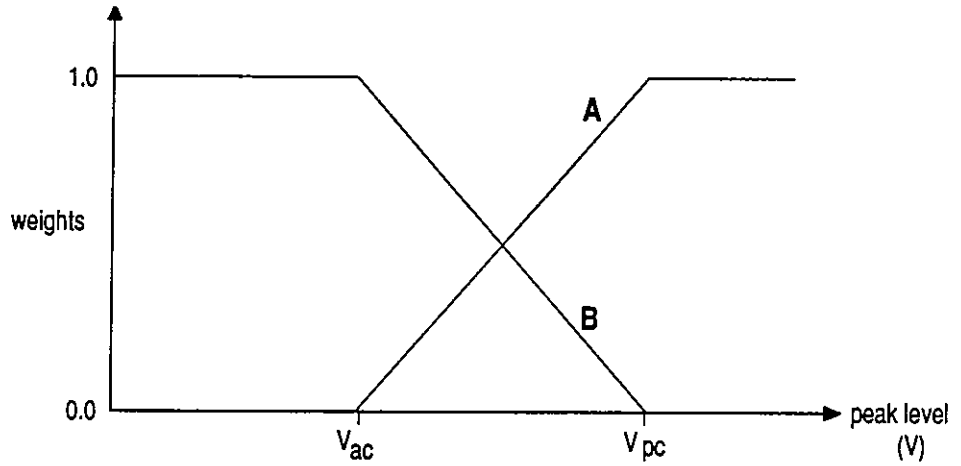


Figure 3.3: Adaptive Level Weight Adjustment

- Whenever possible, use a signal measure for gain control that reflects the perceived level of the signal.
- Suppress output transients that will cause the peak output level to exceed the channel saturation level.
- Provide a smooth measure of the signal level to the gain section.

3.3.1 Adaptation Equation

The first design goal can be satisfied if the average level is used to control the gain (whenever possible). That is, if the output will not exceed the saturation level, the average level is used for gain control. If transients will cause the output to saturate the channel, the peak level will be used for gain control. This satisfies the second goal.

To realize these characteristics, two thresholds, the *peak control threshold* (V_{pc}) and the *average control threshold* (V_{ac}) are set as shown in Figure 3.3. For signals with low peak level (between zero crossings), the average level is used for gain control. For signals with high peak level, the peak level is used for gain control. Between these thresholds, a

linear combination of the average and peak levels is used as the signal "level"

$$level = Ap[k] + B\bar{x}$$

Typically, we will set $V_{pc} = L_{th}$ so that the peak level is used for limiting. Because the average level of a signal ($\bar{x}$) is always less than the peak level, the thresholds must be spaced apart by a reasonable amount (e.g. 10 to 15dB) to ensure that the transition from average to peak control (and vice-versa) occurs smoothly (i.e. there is not an abrupt change in level). Thus, $V_{ac} \simeq V_{pc} - 15dB$.

This design is also very flexible. The weights (A, B) and thresholds (V_{pc}, V_{ac}) can be adjusted to realize most conventional DRC designs.

3.4 Attack Design

Most conventional DRC designs attack at any point on the input signal. No effort is made to perform rapid gain reductions at points where they introduce lower distortion. Our design attacks instantaneously at the zero crossing preceding a transient in an effort to reduce the distortion introduced by rapid gain reductions.

Mapes-Riordan and Leach [11] presented a digital peak limiter design with instantaneous attack at the zero crossing preceding a transient. Their design also recovers in "fixed dB amounts" at zero crossings. They conducted tests for intermodulation distortion (using the sum of sinusoids with incommensurate periods as an input) and found that their limiter (with typical settings) introduced less total harmonic distortion (THD) than a peak clipper. The limiter output spectra for their limiter consisted only of difference frequency components while the clipper spectra contained harmonic and difference frequency components. They reported good subjective results for their design.

Our design strategy pursues a similar course to that of Mapes-Riordan and Leach. Blesser [3] reports that an instantaneous attack time will not produce good subjective

results. However, our design will only have instantaneous attack when the DRC switches into limiting and the peak level is used for gain control. Otherwise, our design will have an attack time that is not instantaneous because a linear combination of the average signal level and the peak level or the average level is used for gain control.

We conducted preliminary tests to determine whether instantaneous gain reductions at zero crossings introduced less distortion than instantaneous gain reductions at any other point on a signal. To simplify the tests, gain reductions at peaks were chosen for the comparison because they offered the greatest variation in performance from gain reductions at zero crossings. Sinusoidal input signals were used for all tests because they have clearly defined peaks and the location of the gain reductions will not vary significantly.

3.4.1 THD and Spectral Tests

Two objective tests were conducted to assess the effects of gain changes at peaks and zero crossings.

Total Harmonic Distortion: THD was used to measure the distortion performance. Although this measure does not indicate the perceptual performance of any method, it does provide an "objective yardstick" to roughly gauge the performance of each method. THD is computed as

$$THD = \frac{\sum_{k=0}^{N-1} (A(\omega_k) - B(\omega_k))^2}{\sum_{k=0}^{N-1} A^2(\omega_k)} \times 100\% \tag{3.1}$$

where $A(\omega_k)$ are the components of the magnitude spectra for the reference signal (no gain change) and $B(\omega_k)$ are the components of the magnitude spectra for the gain-reduced signal.

Spectral Test: The spectra for gain reductions of 3,6 and 9 dB for each gain reduction method were plotted. The frequency domain characteristics of each method was exam-

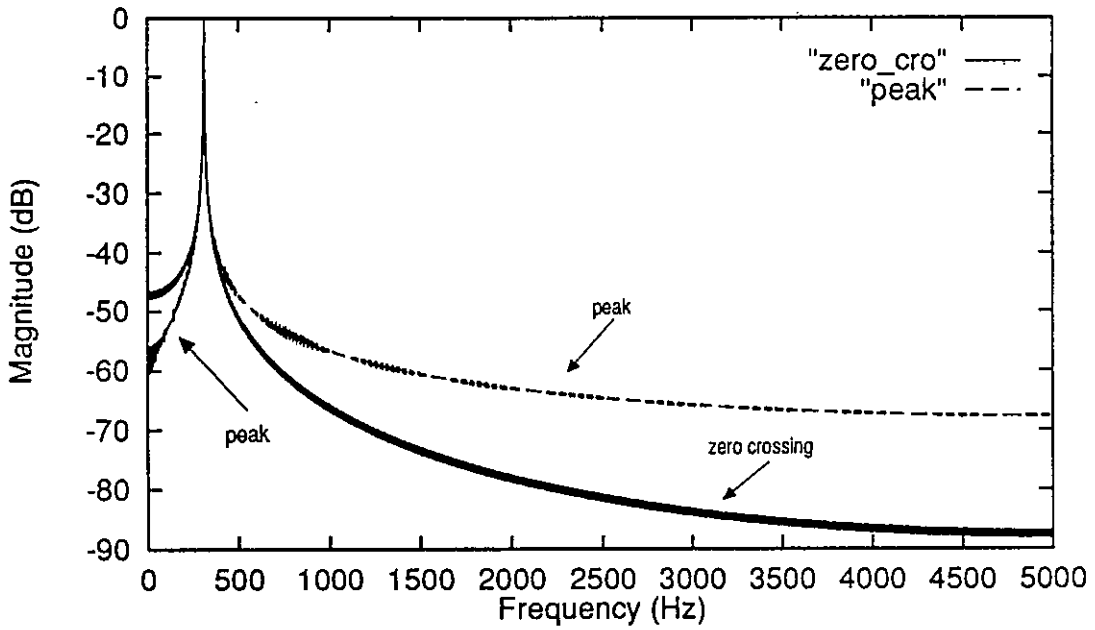


Figure 3.4: Spectra of Low-frequency Test Signal

ined. Because the spectra are all similar in nature, only the 9 dB gain reduction graphs are shown here.

To obtain the frequency domain results a 4096-point sinusoid with a single gain reduction at the peak of zero crossing closest to the 2048^{th} sample was Hamming windowed and a 4096-point FFT was taken. A sampling rate of 10kHz was used.

Single gain reductions of 3, 6 and 9dB were applied to each signal. These gain changes are typical of the gain reductions that may be employed in a DDRC. Two test signal frequencies (312.5Hz and 1562.5Hz) were selected. The FFT provides the best results for these frequencies because the signal fits into the window with an integral number of half-periods. A Hamming window is used to taper the ends of the data sequence and reduce any DC offset introduced by non-integral numbers of full periods.

Figures 3.4 and 3.5 show the spectra for the 9dB gain reductions (f=312.5 and 1562.5 Hz). Results for the other gain reductions are similar. Table 3.1 shows the THD for these tests.

As expected both of these methods have identical THD and the THD is proportional

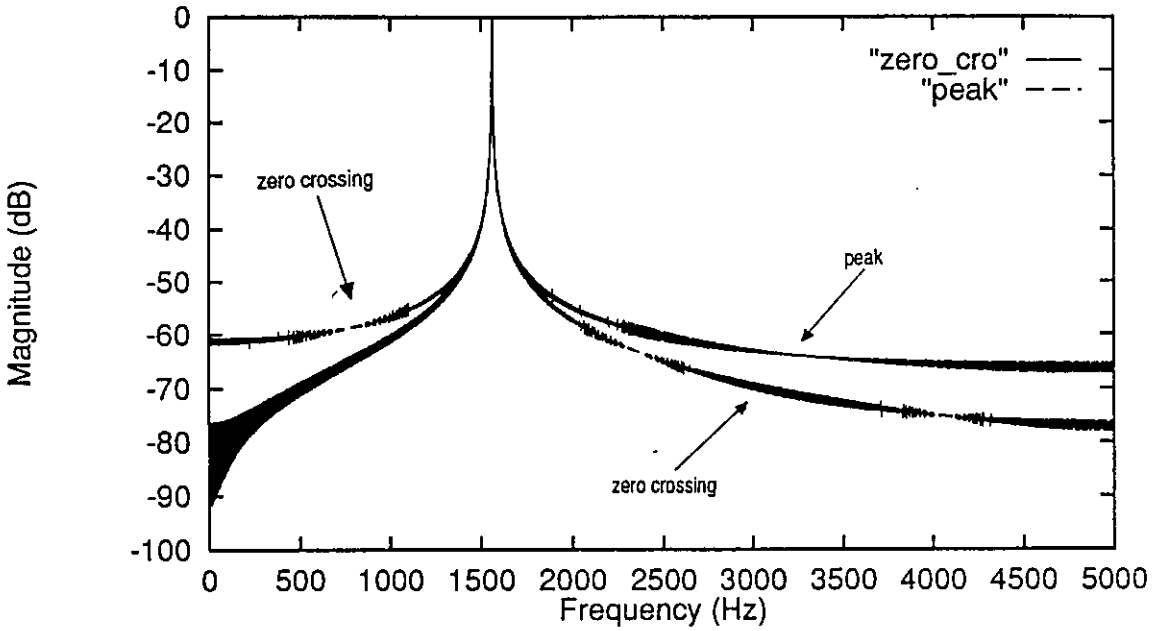


Figure 3.5: Spectra of High-frequency Test Signal

Gain Red. (dB)	Peak		Zero Crossing	
	312.5 Hz	1562.5 Hz	312.5 Hz	1562.5 Hz
-3.0	1.21%	1.21%	1.21%	1.21%
-6.0	4.83%	4.83%	4.84%	4.83%
-9.0	10.52%	10.55%	10.57%	10.56%

Table 3.1: THD for Peak and Zero Crossing Gain Reductions

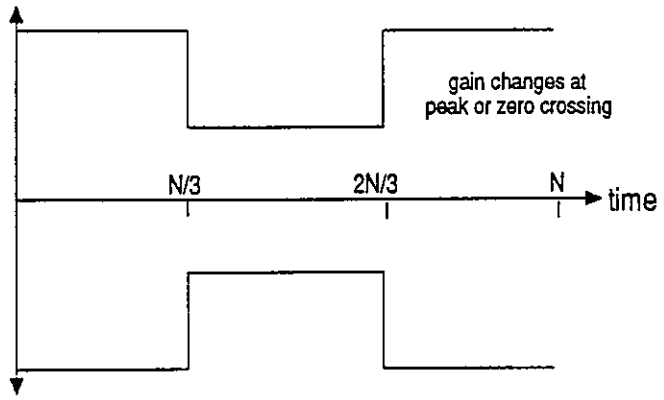


Figure 3.6: Envelope of Test Signal Segment

to the size of the gain reduction. In all cases, peak gain reduction introduces more high frequency harmonics while zero crossing gain reduction introduces more low frequency harmonics. The slight differences in THD are caused by the time differences in the location of the gain change. As the input frequency approaches the Nyquist rate, the samples where peak gain and zero crossing gain reduction occur converge. This is reflected in the spectra—they move together as the frequency increases.

3.4.2 Listening Tests

For the listening tests, a 4096 sample signal with the envelope shown in Figure 3.6 was generated. A number of these signals were concatenated together to form a test signal with a large number of periodic gain changes at peaks or zero crossings.

All listening tests were conducted using only a 9dB gain reduction. The test set-up is shown in Figure 3.7. A pure tone with no gain reductions was used as a reference in all listening tests.

Informal listening tests showed that the gain reductions are easily heard. For the low frequency tests (f=312.5), all listeners commented that the peak gain changes sounded more "clicky" than the zero crossing gain changes. This is a result of the high frequency content that is evident in the spectral plots. The low frequency distortion introduced by

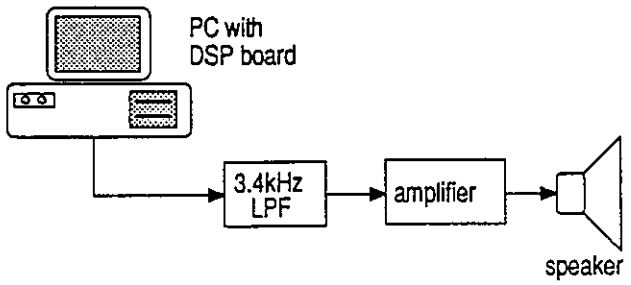


Figure 3.7: Listening Test Set-up

zero crossing gain changes did not seem to have any perceptual effect. It is possible that the strong fundamental masks the lower frequency distortion components.

For the high frequency tests (f=1562.5), the gain reductions could be clearly heard. However, there was not an audible difference between the gain-reduced signals. The 3.4kHz reconstruction filter does not allow all the high-frequency distortion introduced by peak gain changes to be heard. This should make it easier to hear the low frequency distortion introduced by zero crossing gain reduction. Yet, the listeners reported no significant differences.

From these results, we can see that zero crossing gain reduction is superior to peak gain reduction because it introduces less high frequency distortion which gives better subjective performance. This distortion is frequency dependent— as the input frequency approaches the Nyquist rate, the samples where peak and zero crossing gain reduction occur converge.

3.4.3 Attack Implementation

Based on the results of the tests conducted above, our design attacks at the zero crossing preceding an input signal peak. This method introduces less perceivable distortion.

In operation, the DDRC detects zero crossings of the input signal and stores the peaks

between zero crossings in the peak level buffer (Figure 3.1). A zero crossing detector at the output is used to keep the output of the peak level synchronized with the input buffer. The output zero crossing information is also provided to the attack/recovery block. If there is a zero crossing and the input level from the adaptive level computation exceeds the previous output level from the attack/recovery block, the gain will attack. Otherwise the gain will recover.

3.5 Summary

Three problems of conventional designs: recovery time, level measurement and attack characteristics have been addressed in our improved DDRC design. This chapter explained the high-level design of the features in our adaptive DDRC that solve these problems.

To obtain low modulation distortion without introducing "holes" in the program, the adaptive DDRC adapts the recovery time based the input signal crest factor and the isolation of peak levels. Rapid gain reductions (attack) occur instantaneously at zero crossings preceding a transient. This introduces less distortion than conventional methods. The DDRC gain is controlled by an adaptive signal level that uses the peak level for limiting and the average signal level for compression and expansion. This provides improved perceptual performance because the average signal level is better related to the perceived level than the peak signal level.

Chapter 4

Implementation

4.1 General

Having completed the high-level design of the adaptive DDRC, we now turn to the real-time implementation. The DDRC will be implemented on a Motorola DSP56000 Application Development System (DSP56000ADS). This system consists of a single DSP56000 with $8k \times 24$ bit of RAM and the necessary support circuitry. To simplify testing, a Motorola supplied A/D and D/A sub-system (DSP56ADC16) will be used. A direct digital input to the DSP56000 from a digital source will not be used.

The DSP56000 is a powerful DSP chip that uses 24 bit values internally to reduce round-off error. It also features two 56 bit accumulators and high-speed serial A/D and D/A input/output ports. These features make the DSP56000 ideally suited for audio signal processing. Although it is possible to use other number representations (e.g. floating point), the DSP56000 architecture is optimized for fixed-point fractional twos complement arithmetic. Thus, it is much faster to use fractional arithmetic (and scaling, if necessary). The adaptive DDRC will use 16 bit input and output samples.

Because the adaptive DDRC algorithm execution time is variable, an interrupt driven implementation is used. The DSP56000 uses a base clock rate of 20.48MHz ($T_{clock} \simeq$

49ns). Most simple DSP56000 instructions execute in two clock cycles (approximately 98ns). The DDRC will operate at a sampling rate (= interrupt rate) of 44.1kHz—the standard sampling rate for Compact Disk players. Thus, the entire adaptive DDRC algorithm must execute between two interrupts in 20.48MHz/44.1kHz = 464 clock cycles or less.

The DDRC design can be split into input and output sections (Figure 3.1). The input section consists of a high-pass filter, a zero crossing detector, a peak level detector/buffer, an average level filter and an input buffer. The output section operates on delayed input and peak level samples. It performs the level adaptation, T_r adaptation, attack/recovery decision and computes the static characteristics. Figure 4.1 shows an overview flow chart of the adaptive DDRC algorithm.

The remainder of this chapter covers the "block-by-block" implementation of the adaptive DDRC. Complete code listings are contained in Appendix A. A memory map is in Appendix B.

4.2 Input High-pass Filter

The input highpass filter (HPF) removes any DC component of the input signal. Stikvoort [18] included a filter is his design so input DC offsets would not be modulated by the gain control signal and cause "plops" in the output signal. Because the offset voltage and signal level are indistinguishable, a DC offset may also corrupt the average level measurement.

Mapes-Riordan and Leach [11] included an input HPF in their digital peak limiter to restrict the lowest input signal frequency so the input signal buffer would not overflow between zero crossings. In our design, the input signal buffer must store the input samples between zero crossings, but we assume that the input is bandlimited to greater than 20Hz (with a possible DC offset). Thus, the HPF will only be required to block DC. This

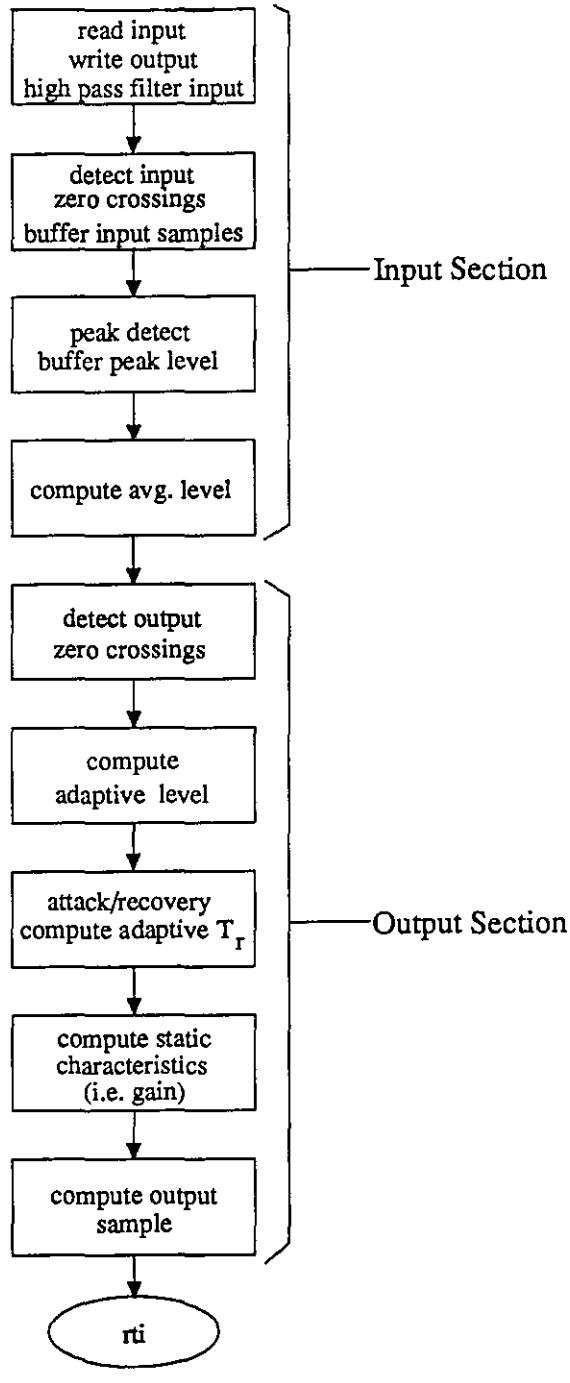


Figure 4.1: Overview Flow Chart of Adaptive DDRC Algorithm

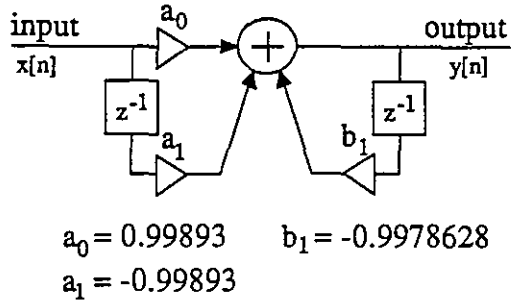


Figure 4.2: First-order Input High-pass Filter

assumption allows the use of a low-order filter which saves computation time.

To minimize the computation time, our design will use a first-order HPF. The filter (Figure 4.2) has a -3dB cutoff frequency of 15Hz. The simulated frequency response is shown in Figure 4.3. Notice that the filter has infinite rejection at DC because $a_0 = -a_1$. The real-time implementation uses parallel updated pointers and orders coefficients in memory to speed execution (see gain.asm listing in Appendix A).

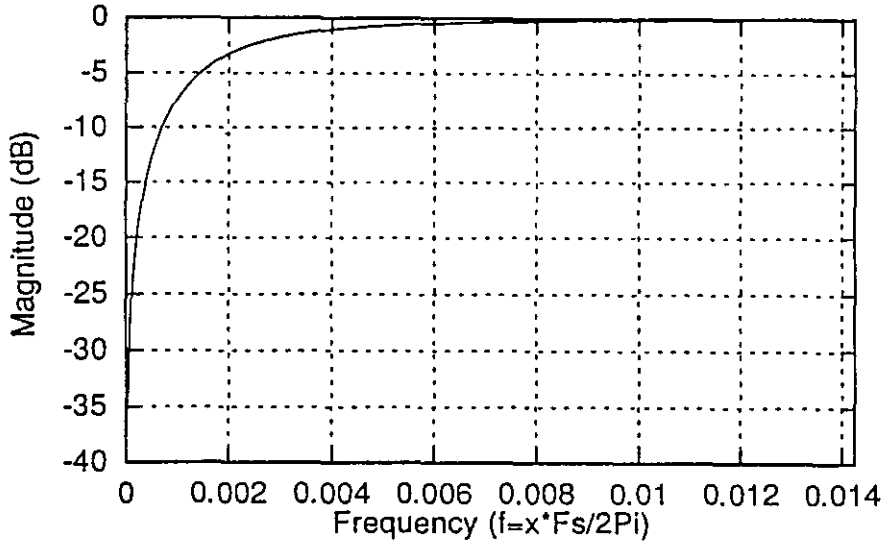


Figure 4.3: Frequency Response for First-order HPF

4.3 Zero Crossing Detector and Buffering

The input section uses two circular buffers to hold the input samples and the peak signal levels detected between input signal zero crossings (zc's). A modulo addressing mode and parallel pointer updating are used to provide fast input and peak level buffering.

4.3.1 Zero Crossing Detector

To detect zc's in real-time, an algorithm that uses multiplications and as few comparisons as possible was developed. Multiplications execute much faster than comparisons on the DSP56000. The zc detector uses the current and previous input samples (z[n] and z[n-1]) to determine if the present input sample is a zc:

```
if z[n]z[n − 1] ≤ 0 then
        input is a zero crossing
else
        input is not a zero crossing
```

For $z[n] = 0$ this algorithm detects two zc's and gives a peak value of zero. This may cause the peak value to be less than the average level for a signal segment (i.e. the samples between two zc's) —the DDRC algorithm is designed so that this condition will not cause any problems. A previous zc detection algorithm that used three input samples to detect zc's did not suffer from this fault. However, it executed much more slowly than the two-sample method presented above.

4.3.2 Buffers

The circular peak and input buffer sizes are determined by the maximum (tolerable) delay through the DDRC and the number of peak levels (detected between zc's) used in the P_v calculation. For the input signal buffer of size M, the output sample is $M - 1$ samples

behind or one sample ahead of the current input sample. Whenever there is an input zc, the current peak input level is stored and an input peak pointer is incremented. A similar zc detector at the output end of the input sample buffer increments the output peak pointer. The input buffer is filled with small positive values at initialization, so the difference between the input and output peak pointers always equals the number of zc's in the input buffer.

The input buffer must store the samples between two zc's of the lowest input frequency to the DDRC. For a sampling rate of $f_s = 44.1kHz$ and a minimum input frequency of $f_{min} = 20Hz$, the buffer must be

$$M = \left\lfloor \frac{f_s}{2f_{min}} \right\rfloor = 1103$$

samples long. The minimum size of the peak buffer is also M. If the highest input frequency is $f_{max} = f_s/2$ (i.e. the Nyquist rate), there will be an input zero crossing every sample. Thus, the number of zc's is equal to the number of input samples.

Recall that the P_v calculation uses N future peak values—each detected between a zc on the input. The input buffer must hold at least $N + 1$ zc's (i.e. N future zc's and the present zc) of f_{min}. Thus, the input buffer must be extended to at least

$$M = (N+1)\left\lfloor \frac{f_s}{2f_{min}} \right\rfloor = (N+1)(1103)$$

samples long. The DDRC currently uses $N = 5$ (see Section 4.4). This means that the delay through the DDRC is

$$\begin{aligned} T_{pd} &= (5+1)(1103)\left(\frac{1}{f_s}\right) \\ &\simeq 150ms \end{aligned}$$

This is a tolerable delay. The delay time is fixed so synchronization is not a problem and in most applications users will never hear the unprocessed input to the DDRC. For a delay of this length, the average measurement must also be delayed so that is does not act too far ahead of the input signal.

Presently, the DDRC uses a fixed delay of $1125(1/44100.0) \simeq 25.5ms$. This is constrained by the memory available on the DSP56000ADS (i.e. money!), the restrictions of the ADS memory configurations [9] and the starting addresses that may be used for modulo addressing [15]. If more high speed memory were purchased, this would not be a limitation. This means that P_v calculations for some input signals (those with less than $(N+1)$ zc's in the input buffer) will be incorrect because the future peak values will be wrong. For this short delay, the average level does not require buffering. We conducted some informal listening tests and found that this limitation on the P_v calculation did not seriously compromise the DDRC's performance.

4.4 Level Measurement

Two signal measurements, the peak level and the average level, are computed by the input section of the DDRC. These levels are adaptively combined as explained in Section 3.3.1. The peak level is used primarily for peak limiting; the average level is used for compression/expansion.

4.4.1 Peak Detector

The peak detector measures the peak level between zc's and stores it in the peak buffer. Zc's are detected as described above. They provide a convenient block size for the determination of peak level. Peak detection between zc's is also convenient because the DDRC attacks instantaneously at zero crossings. The peak detector has an (effective) instantaneous attack because it immediately updates the peak level.

At the output of the peak buffer, the peak output pointer is the address of the peak level between the present and next zc. This "future" peak level allows the DDRC (when the gain is controlled by the peak level) to reduce the gain at the zc *before* a transient. The output peak level remains constant between zero crossings. Figure 4.4 shows how

the contents of the peak and input buffers are related. The input buffer contains high pass filtered samples from the source and the peak buffer contains peak levels between zc's of the input signal.

4.4.2 Average Level

The average level measurement is used as an approximation to the perceived signal level. A first-order IIR filter is used because it is computationally efficient and has no overshoot. An IIR structure provides sufficient averaging in a low order design (Figure 4.5).

Good averaging characteristics require that the time constant of the filter (τ) be much longer than the period of the lowest input frequency. The minimum input frequency ($f_{min} = 20Hz$) has a period of 50ms. Thus, for good averaging we require $\tau >> 50ms$. Cabot [5] reports that "the ear responds more to the power of the signal averaged over some moderate time period (typically one-quarter of a second) than to the average level." This implies that we ought to use $\tau = 250$ms. However, for a value or τ in this range, the peak and average levels may differ by so much that the transition from average to peak (via level adaptation) will not occur smoothly. To achieve a balance, we will use $\tau = 100ms$. This will provide reasonable averaging and should respond to program level changes rapidly so that the transition from average to peak level control will occur smoothly. If required, this value can be refined through listening tests.

McNally [12] showed that the time constant of this filter (Figure 4.5) can be written as

$$\tau = \frac{T_s\sqrt{1-a}}{a}$$

(T_s is the sampling period) providing

$$\tau > 2T_s$$

That is, $\tau > 2/44100 \simeq 45\mu$s—a condition that will be met. Solving this equation for a

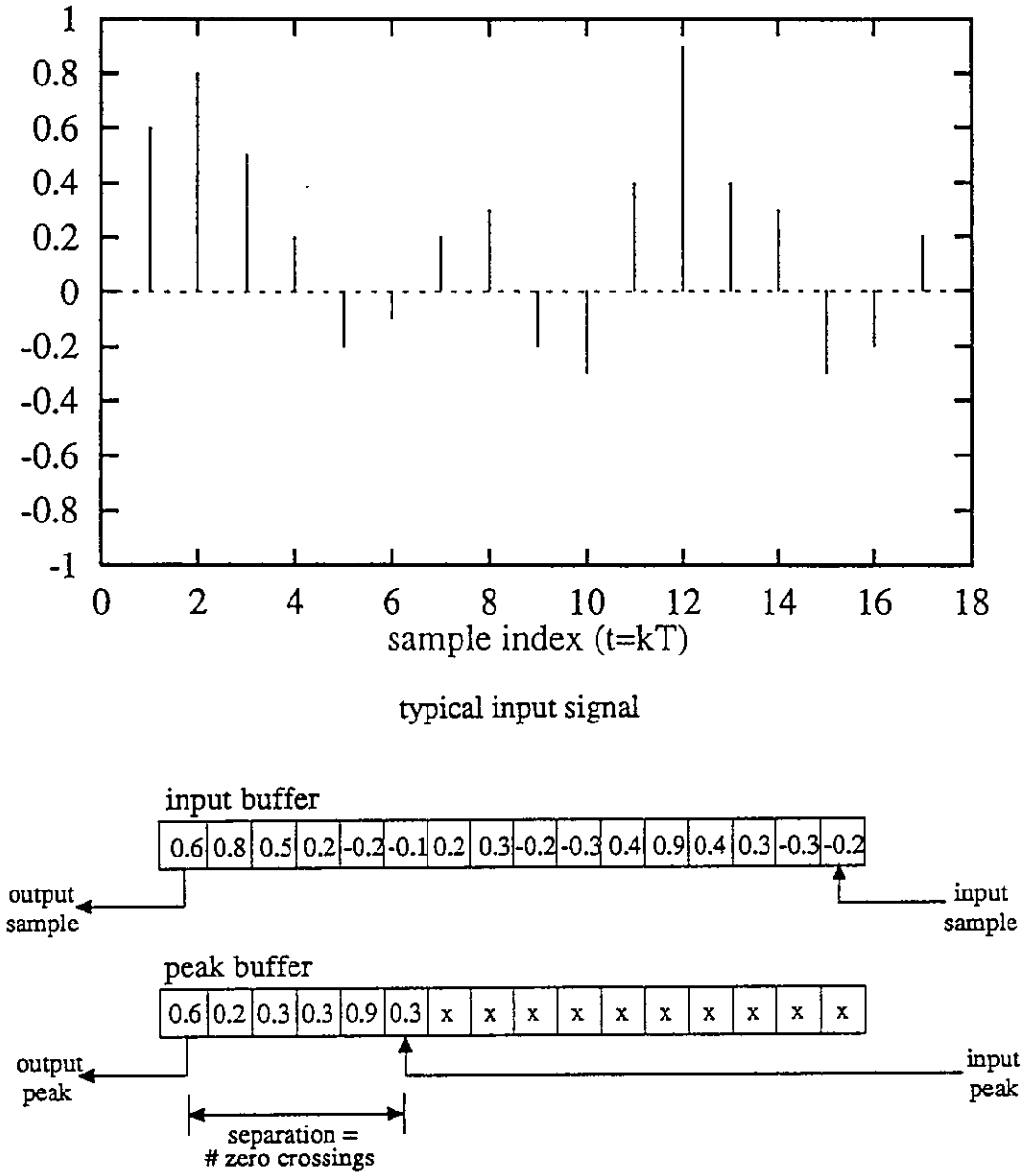


Figure 4.4: Input and Peak Buffer Operation

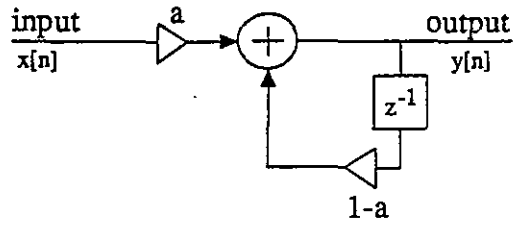


Figure 4.5: Averaging Filter

results in

$$a = \frac{1}{2}\left(\frac{T_s}{\tau}\right)^2 \left[-1 + \left(1 + 4\left(\frac{\tau}{T_s}\right)^2\right)^{1/2}\right]$$

For $T_s = 1/44100$ and $\tau = 0.100$ we find $a = 2.2673316 \times 10^{-4}$. This value of a is used in the adaptive DDRC implementation. a is written into RAM so the value may be easily changed to meet different requirements.

4.5 Level Adaptation

As described previously, the adaptive level section combines the peak and average level measurements so (1) the peak output level is always below the saturation level of the channel and (2) the gain signal is better related to the perceived signal level (whenever possible). Figure 4.6 shows (in block diagram form) how this scheme (described in Section 3.3.1) is implemented in the adaptive DDRC.

Recall, that below the average control threshold (V'_{ac}), the average signal level is used for gain control. Above the peak control threshold (V'_{pc}), the peak level (detected between zc's) is used for gain control. Between these two thresholds, a linear combination of the

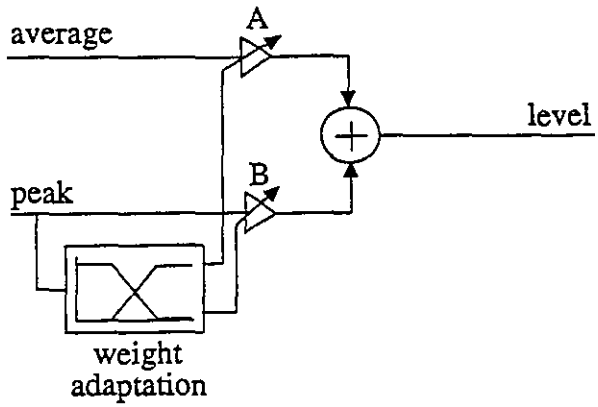


Figure 4.6: Block Diagram of Adaptive Level Scheme

peak and average level control the gain. The sum of A and B is always less than or equal to 1.0—this ensures that the adaptive level output is always fractional. The equations for the adaptation are

- $p[k] < V'_{ac} \Rightarrow A = 1 - B = 0, B = 1$
- $V'_{ac} \leq p[k] \leq V'_{pc} \Rightarrow A = 1 - B, B = (1/\Delta)(p[k] - V'_{pc})$
- $p[k] > V'_{pc} \Rightarrow A = 1 - B, B = 0$

where $\Delta = V'_{ac} - V'_{pc}$, $V'_{ac} = 10^{V_{ac}/20}$ and $V'_{pc} = 10^{V_{pc}/20}$ (V_{ac} and V_{pc} are both specified in dB).

To implement these equations on the DSP56000, all values must be fractional. Both V'_{ac} and V'_{pc} will be fractional since $V_{pc}, V_{ac} \leq$ 0dB. $1/\Delta$ will be non-fractional, so it must be scaled. A maximum value of Δ must be found so a scaling factor (2^j) can be determined to make $\Delta/2^j \leq 1.0$. If we choose $\epsilon = V_{pc} - V_{rc} >$ 10dB and $V_{pc} > -$50dB a scaling factor of $2^9 = 512$ can be used (Appendix C). These restrictions on ϵ and V_{pc} are required to ensure that there is not an unnecessary loss of accuracy in the adaptive level calculations. This scaling factor gives results that are accurate to 14 bits. This is adequate for these computations.

To increase the speed of these computations, pre-computed (and scaled) values of $1/\Delta$ and V_{pc}/Δ are used. Also, since we always have $A = 1 - B$, only B must be computed. Whenever possible, variables used in sequence are ordered in memory and fetched via parallel updated pointers. The entire level adaptation algorithm is implemented as shown if Figure 4.7.

The values of V'_{pc} and V'_{ac} are user adjustable since they are written into RAM. Typically, V_{pc} will be set equal to the limit threshold (L_{th}) so that the DDRC will use the peak level for limiting. V_{ac} will be set 10-15dB below V_{pc} to ensure a smooth transition from average to peak level gain control.

4.6 Dynamic Characteristics

As described in Section 3.2.1, the recovery time (T_r) of the adaptive DDRC is controlled by the crest factor (C_f) and peak variation (P_v) of the input signal. T_r is adapted to achieve low modulation distortion and avoid causing "holes" in the musical program. T_r will be in the range $50ms \leq T_r \leq 200ms$. Our design uses a first order exponential recovery characteristic. Gain attack occurs instantaneously at the zero crossing preceding a transient.

4.6.1 Recovery Time Adaptation

The adaptive DDRC design uses a first order recovery filter (Figure 4.8). By considering the unit sample response, it can be shown that the relationship between T_r and a (the filter parameter) is

$$a = exp\left(-\frac{1}{T_r/T_s + 1}\right)$$

where T_s is the sample period. (Note: This could have also been solved using the relationship derived by McNally [12] as in Section 4.4.2.) Figure 4.9 shows a graph of this

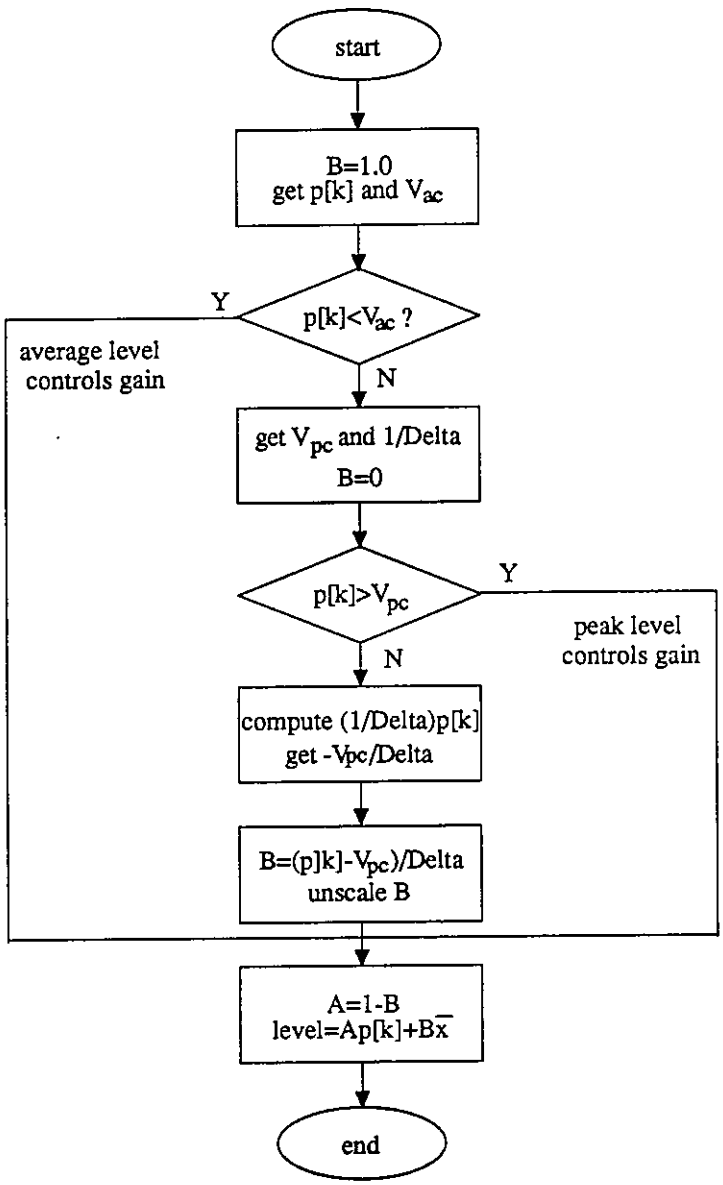


Figure 4.7: Flow Chart of Adaptive Level Implementation

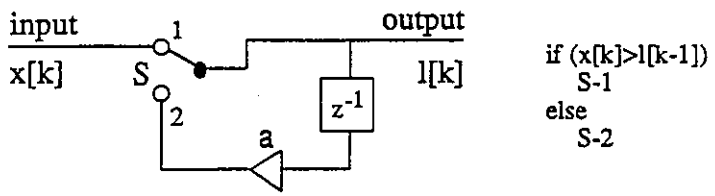


Figure 4.8: First Order Recovery Filter

relationship. Notice that there is a very small range of a over which T_r can can adapt within the range $50ms < T_r < 200ms$. Also, notice that increasing a implies a longer T_r.

As described earlier, the filter parameter (a) is adapted using

$$a = k_1 C_f - k_2 P_v + c$$

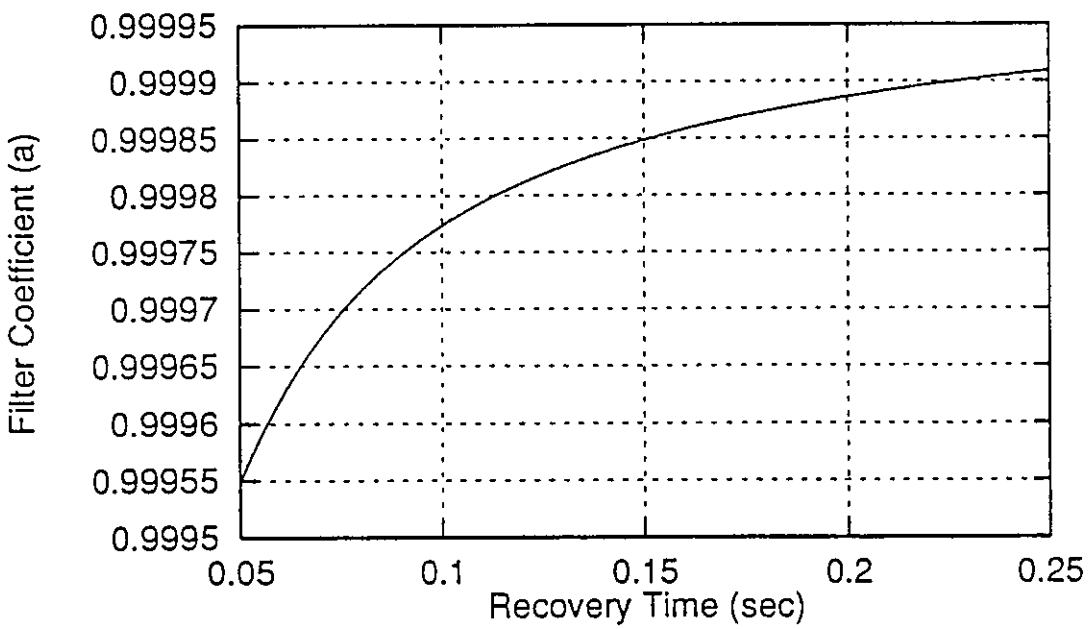


Figure 4.9: Filter Parameter (a) versus T_r

P_v is computed using

$$P_v = p[k] - \frac{1}{5}\sum_{n=1}^{n=5} p[k+n]$$

where $p[k]$ is the present peak value between zero crossings. The value $N = 5$ is a compromise. We want to measure the average future peak level so that the DDRC will adapt based on a future trend—not just one or two future peak levels. However, N must not be too large or isolated future peak levels will not affect the average at all. $N = 5$ provides a good compromise and allows the P_v calculation to execute in a reasonable number of clock cycles. The multiply and accumulate operation (*mac*) of the DSP56000 is used in conjunction with parallel pointer updates to "speed up" the P_v computation.

For the real-time implementation, we found that C_f^{-1} took too long to calculate because it required a division. C_f^{-1} was replaced with a quantity we named *crest difference* (C_d). C_d provides similar information to C_f^{-1}, but it can be computed much faster:

$$\begin{aligned} C_f^{-1} &= \frac{\bar{x}}{p[k]} = 1 + \frac{\bar{x} - p[k]}{p[k]} \\ C_d &= 1 + \bar{x} - p[k] \end{aligned}$$

We also found (in later experiments) that the peak level between zc's was not always greater than the average signal level—this should have been apparent earlier! Also, the peak level between zc's provides no information about the previous peak level used for gain control. This information is important when deciding how to adapt T_r. For these reasons, the peak level between zc's (p[k]) was replaced by the recovered output level ($l[k]$) in the C_d calculation. Now we have,

$$C_d = 1 - l[k] + \bar{x}$$

$l[k]$ is always greater than or equal to $\bar{x}$ and it is always fractional. Thus, C_d is always fractional.

Parameter Selection

The parameters for the adaptation equation

$$a = k_1 C_d - k_2 P_v + c$$

must be chosen to keep a within the desired range of recovery times. Let the threshold that determines whether T_r is adaptive or constant be Q. Then, T_r is constant for $C_d > Q$ and adaptive for $C_d < Q$.

Q is set so that T_r will adapt when the DDRC goes into peak limiting. Thus, T_r should adapt when the peak level is greater than the peak control threshold (i.e. $p[k] > V_{pc}$). As described previously, we set $V_{pc} = L_{th} = -15dB$. Typically, the peak level of a signal is 10-15dB greater than its RMS level [7]. If we consider that average and RMS level are roughly equivalent, the RMS level can be used in the C_d calculation in place of the average signal level. Thus, for a typical signal we expect

$$\begin{aligned} 1 - 10^{\frac{L_{th}}{20}} + 10^{\frac{L_{th}-15}{20}} \leq \quad & C_d \quad \leq 1 - 10^{\frac{L_{th}}{20}} + 10^{\frac{L_{th}-10}{20}} \\ 0.854 \leq \quad & C_d \quad \leq 0.878 \end{aligned}$$

when $p[k] > V_{pc}$. For our design we selected $Q = 0.86$. The adaptation equation is only used for $0 \leq Q \leq 0.86$. For $0.86 < Q < 1.0$, a constant value of a=0.999886641 gives a constant $T_r = 200ms$.

The values for the other adaptation parameters are determined by considering the conditions that will lead to the maximum and minimum values of T_r. When T_r is adaptive

$$a = k_1 C_d - k_2 P_v + c$$

where $0 \leq C_d \leq 0.86$, $-1.0 \leq P_v \leq 1.0$.

T_r should be a minimum ($\Rightarrow a$ is minimum) when $C_d = 0$ (i.e. $l[k] >> \bar{x}$) and $P_v = 1.0$ (i.e. the current peak is isolated). T_r should be a maximum ($\Rightarrow a$ is maximum) when

$C_d = Q = 0.86$ (i.e. T_r has just become adaptive) and $P_v = -1.0$ (i.e. future peaks are large). Thus,

$$\begin{aligned} min(a) &= k_1(0) - k_2(1.0) + c \\ &= -k_2 + c \qquad (4.1) \\ max(a) &= k_1(0.86) - k_2(-1.0) + c \qquad (4.2) \end{aligned}$$

where $min(a) = 0.999546794$ ($T_r = 50$ms) and $max(a) = 0.999886641$ ($T_r = 200$ms). Solving equation 4.1 for c and substituting it into equation 4.2 we find

$$max(a) - min(a) = 0.86k_1 + 2k_2$$

We have three variables and only two equations, so one parameter must be chosen. If we select $k_1 = k_2$, P_v will have twice as much influence on the adaptation as C_d does since it has roughly twice the range of C_d. Making this substitution (i.e. $k_1 = k_2 = k$), results in

$$k = \frac{max(a) - min(a)}{2.86} = 1.18827 \times 10^{-4}$$

These values of $k_1 = k$ and $k_2 = k$ will serve a starting points and can be refined via listening tests. Substituting this value of k_2 into equation 4.1, results in

$$c = min(a) + k_2 = 0.999665621$$

4.6.2 Attack/Recovery Implementation

The first order recovery filter used in the adaptive DDRC is a peak hold type. That is, the output ($l[k]$) is controlled such that the larger of the adaptive level (x[k]) or the recovered output level $al[k-1]$ is written to the output. Since the peak level is updated at zc's, gain attack occurs instantaneously at zc's. Figure 4.10 shows a block diagram of the attack/recovery section (including T_r adaptation) design used in the real-time implementation of the adaptive DDRC.

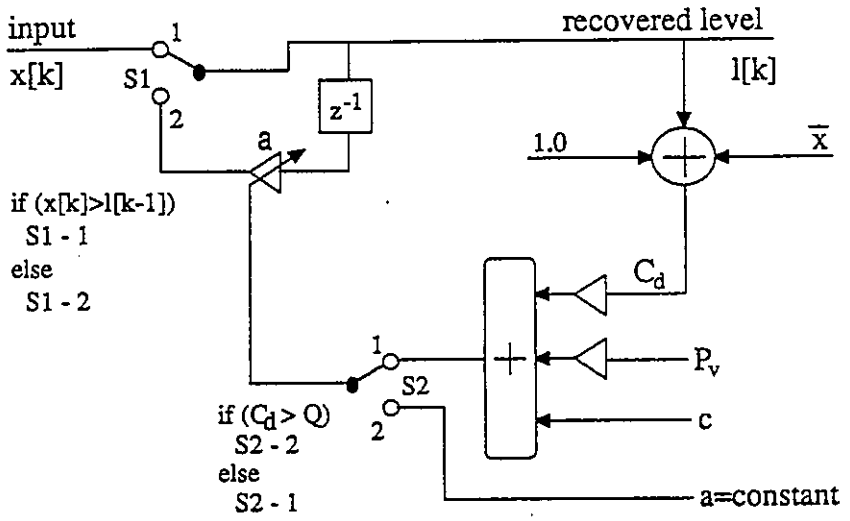


Figure 4.10: Attack/Recovery Section of DDRC

4.7 Static Characteristics

As discussed previously (Section 2.3), the static characteristics specify the steady state performance of the DDRC. They describe the instantaneous relationship between the input level and the DDRC gain (both in dB). The adaptive DDRC uses four region level independent static characteristics.

There are three basic considerations when implementing static characteristics:

1. **Speed:** The static characteristics should be computed as rapidly as possible.
2. **Accuracy:** The gain value should be as accurate as possible.
3. **Storage:** The computations should use as little memory as possible.

The computation of the static characteristics can be split into four steps: (1) determine the input region (i.e. expansion, no-action, compression or limiting) (2) compute the input level in dB (3) perform the gain calculation and (4) convert the gain from dB to a linear

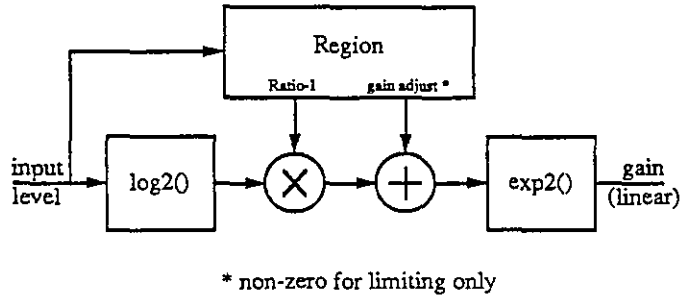


Figure 4.11: Computation of Static Characteristics

value. Using the equations presented in Section 2.3, these computations are implemented as shown in Figure 4.11.

The computation of the static characteristics is now reduced to real-time calculation of logarithms and exponents. Since the these functions must be computed on a binary processor (the DSP56000), base two logarithms and exponents can be used to reduce the computation time. To increase the accuracy provided by low-order approximations, a previous design used *binary floating point* (BFP) arithmetic [12].

BFP arithmetic uses an approximation between 0.5 and 1.0 (i.e. 2^{-1} and 2^{0}) along with normalization/denormalization to compute a base two logarithm or exponent. The logarithm of any number x can be written as

$$\log_2 x = \log_2 m + E$$

where E is the *exponent* and m is the *mantissa.* Assuming x: $0 \leq x \leq 1.0$ is a binary number[1], E is the number of left shifts necessary to set the most significant bit (MSB) of x. This procedure is known as *normalization.* The normalized value is m, $0.5 \leq m \leq 1.0 - 2^{-23}$. $\log_2(m)$ is found using an approximation.

The computation of the 2^x is similar in nature. If $y = \log_2(m) + E$ where m and E are defined as above and $-1.0 \leq (f \equiv \log_2 m) \leq 0$ (since $0.5 \leq m \leq 1.0$) and E is a negative

[1]The DSP56000 uses 24 bit twos complement fractional values so $0 \leq x \leq 1 - 2^{-23}$

integer, then

$$\begin{aligned} x &= 2^y \\ &= (2^f)(2^E) \end{aligned}$$

2^f is found via an approximation. 2^E is simply a right shift for a binary number. The right-shifting operation *denormalizes* 2^f.

BFP arithmetic provides better accuracy for low order approximations because a smaller range is approximated. For the computation of $\log_2(x)$ this approach greatly increases the accuracy because of the large slope as $x \rightarrow 0$. The disadvantage of this approach is the normalization/denormalization operation. Two clock cycles per bit are required to implement bit-shifting in software.

The accuracy required for the gain calculations is difficult to specify. The input samples (from the source) are 16-bits. This provides an upper limit on the accuracy. The lower limit is determined by subjective analysis. McNally [12] states that coarse errors in the static characteristics may be subjectively acceptable "if gain changes are sufficiently small and occur infrequently." However, if short attack and recovery times are used, rapid gain changes may generate *zipper noise*—a subjectively displeasing effect caused by a fast changing, severely quantized gain value.

McNally used a 256×8 bit lookup table for the mantissa in his $log_2(m)$ calculations. BFP arithmetic was used and the exponent was limited to $0 \leq E \leq 7$ to save computation time. To achieve good subjective results, he employed an adaptive filtering scheme to smooth "cogs" in the gain signal and reduce zipper noise. We will consider 8 bit accuracy for the mantissa as our lower limit. Our design will aim for an accuracy between 8 and 16 bits in the mantissa computation. If necessary, McNally's adaptive smoothing scheme can be used to reduce zipper noise.

Two basic methods, lookup tables and polynomial approximations, were examined for the real-time calculation of base two logarithms and exponents. In all approximations

considered, the peak error is minimized. Two basic lookup table types are examined in this section, *uniform* tables and *non-uniform* tables. Although interpolated lookup tables are not presented here, they were also under consideration for real-time implementation. They were not used because they required too much computation time. Chebyshev polynomial approximations were also examined.

4.7.1 Uniform Lookup Tables

Lookup tables sequentially store the value of a function at a number of sample points across the (BFP) input range. Lookup table approximation error can be attributed to two sources:

1. **Quantization Error:** error resulting from the finite precision of the samples stored in the table (i.e. The table uses fixed point entries).

2. **Interpolation Error:** error caused by the finite number of table entries.

These two sources of error are dependent: enough precision must be used so that adjacent table entries are distinct. Otherwise the effective table size will be reduced by overlapping entries. The DSP56000 uses 24 bit fractional twos complement arithmetic so the total error is bounded by the worse case interpolation error for any look-up table.

A lookup table can be considered as a quantizer. The input is split into ranges separated by decision levels, $y[k]$. In each range, a representation level, $x[k]$, is selected to represent all values within the range. The value $q[k] = Q[f(x[k])]$[2] is stored in the look-up table. For our application, the input 24 bit binary value. This value can be considered continuous when compared to the interpolation error introduced for the table sizes under consideration.

[2] $Q[\bullet]$ represents the quantization operation.

Table Size	$log_2(x)$ Error	2^x Error
256	2.815×10^{-3}	1.353×10^{-3}
512	1.408×10^{-3}	6.767×10^{-4}
1024	7.043×10^{-4}	3.384×10^{-4}
2048	3.523×10^{-4}	1.692×10^{-4}

Table 4.1: Peak Error for BFP Look-up Tables

To generate an output value (z), for an input (x) the *decision rule*

$$z = Q[f(x[k])] = q[k]$$

iff $y[k-1] \leq x \leq y[k]; 0 \leq k \leq N$ is used. Considering lookup tables as quantizers provides a flexible analysis framework because no restrictions are placed on the location of representation or decision levels.

For a uniform BFP lookup table, $x[k]$ and $y[k]$ are uniformly spaced. The input value is a fractional fixed-point mantissa ($0.5 \leq m \leq 1.0$). Since $\log_2(x)$ is a monotonic non-decreasing function for $0 < x \leq 1.0$, the worst case peak error occurs in the interval where the slope is greatest. Thus, the error for the $\log_2(x)$ look-up table is

$$\begin{aligned} \epsilon &= |\log_2 y[0] - \log_2 \left(y[0] \left(1 - \frac{1}{2N} \right) + \frac{y[N]}{2N} \right)| \\ \epsilon &= |-1 - \log_2 \left(0.5 \left(1 - \frac{1}{2N} \right) + \frac{1}{2N} \right)| \end{aligned}$$

Table 4.1 shows the peak error for a $\log_2(x)$ BFP lookup table.

By a similar analysis, the peak error for a 2^x BFP look-up table is

$$\epsilon = |2^{y[N]} - 2^{x[N]}|$$

where $y[N] = 0$ and $x[N] = y[0]/2N = \log_2(0.5)/2N$. Thus,

$$\epsilon = |1 - 2^{-1/2N}| \tag{4.3}$$

Table 4.1 shows the error for some typical table sizes.

For both tables, the BFP number representation causes the error to repeat in powers of two. Within each interval (i.e. power of two) the error increases as the input moves from the low slope to high slope end of the interval. For both table types, doubling the table size halves the peak error (i.e. the accuracy increases by one bit). The error is highly non-linear across and within decision intervals. For the 2^x lookup table, quantization error makes a significant contribution to the error at the lower extreme of the input range. However, the error does not exceed the bound set by Equation 4.3.

4.7.2 Non-uniform Lookup Tables

Nonuniform tables have nonuniformly spaced decision and/or representation levels. These tables are designed to some optimization criterion. We will not consider tables with non-uniform decision and representation levels because they are difficult to address in real-time. However, tables with only non-uniform representation levels will be examined because they are addressed as uniform tables. Since peak error was selected as the error metric— tables with non-uniform representation levels that minimize peak error will be examined.

Both $log_2(x)$ and 2^x are monotonic non-decreasing functions. Thus, the maximum peak error within any decision interval may be minimized by selecting

$$x[k] = f^{-1}\left(\frac{f(y[k]) + f(y[k-1])}{2}\right)$$

For both 2^x and $log_2(x)$ look-up tables, the worst case error occurs where the slope of the function is the greatest. Thus, for $log_2(x)$ the worst case error is

$$\begin{aligned}
\epsilon_1 &= |f(y[0]) - f(x[1])| \\
&= |f(y[0]) - \frac{f(y[k]) + f(y[k-1])}{2}| \\
&= |\frac{f(y[0]) - f(x[1])}{2}|
\end{aligned}$$

Table Size	$log_2(x)$ Error	2^x Error
256	2.8123×10^{-3}	1.3520×10^{-3}
512	1.4075×10^{-3}	6.7644×10^{-4}
1024	7.0410×10^{-4}	3.3833×10^{-4}
2048	3.5213×10^{-4}	1.6920×10^{-4}

Table 4.2: Error for Non-Uniform Look-up Tables for $log_2(x)$ and 2^x

Notice also that $\epsilon_1 = |f(y[1]) - f(x[1])|$ since the peak error is *equalized* across the interval. For 2^x, a similar analysis show that the maximum peak error is

$$\epsilon_N = |\frac{f(y[N]) - f(y[N-1])}{2}|$$

Again we may also write $\epsilon_N = |f(y[1]) - f(x[1])|$ because the peak error is the same at each end of the interval. Table 4.2 shows the maximum peak error for BFP $log_2(x)$ and 2^x tables.

For a non-uniform $log_2(x)$ BFP table, there is not a significant improvement in the error when compared to the same size uniform table. The non-uniform 2^x BFP table shows only a very small reduction (approximately 1.5%) in error compared to the uniform table. If tables are selected for the final design a non-uniform (representation level) table should be used because it provides slightly lower peak error than a uniform (representation level) table. Both of these tables are simple to address.

4.7.3 Polynomial Approximations

There are many methods available for approximating an arbitrary function with a polynomial. However, a majority of these methods require that $n + 1$ samples be used to obtain an n^{th}-order approximation. This form of a solution matches exactly at the n sample points, but may deviate far from a *best-fit* solution at points between the samples.

This behavior is especially evident for high-order approximations.

For the approximation of $log_2(x)$ and 2^x we want a method that can approximate a large number of samples using a low-order polynomial and minimumize the peak error. This can be accomplished via a *minimax* approximation— an approximation that minimizes the maximum error. This can be shown to be equivalent to the *Chebyshev Criterion*:

> Of all polynomials, $p_n(x)$ of degree n, having a leading coefficient equal to 1, the polynomial
>
> $$\frac{T_n(x)}{2^{n-1}}$$
>
> has the smallest least upper bound for its absolute value in the interval $-1 \leq x \leq 1$ (where $T_n(x)$ is a Chebyshev polynomial) [8].

Note that for cases where the approximation will be for $x \geq 0$, it is necessary to work with the shifted Chebyshev polynomials.

$$T_n^*(x) = T_n(2x - 1)$$

which are defined over the range $0 \leq x \leq 1.0$. Chebyshev polynomials can be also be *stretched* to provide approximations over arbitrary ranges greater than $0 \leq x \leq 1.0$ or $-1 \leq x \leq 1$. Normally the calculation of Chebyshev polynomial coefficients is a complicated procedure [8] that involves the evaluation of a number of Fast Fourier Transforms. However, a powerful symbolic algebra package (MAPLE[3]) was used to simplify this process.

MAPLE was used to find the minimum-order Chebyshev approximation to each function for a specified accuracy. Approximations over the BFP input ranges: $0.5 \leq x \leq 1.0$ for $log_2(x)$ and $-1.0 \leq y \leq 0$ for 2^x were found. Table 4.3 shows the peak error for 2^{nd}-, 3^{rd}- and 4^{th}-order Chebyshev approximations. The coefficients for these approximations

[3] A program developed at the University of Waterloo.

Order	$log_2(x)$	2^x
2	5.5830×10^{-3}	1.2628×10^{-3}
3	7.2527×10^{-4}	5.4737×10^{-5}
4	1.0019×10^{-4}	1.8895×10^{-6}

Table 4.3: Peak Error for BFP Chebyshev Polynomial Approximations

are tabulated in a Appendix D. To give an idea of the nature of these curve fits, Figure 4.12 shows plots of the BFP approximations to $log_2(x)$.

The 3^{rd}- and 4^{th}-order BFP approximations give acceptable results—for $log_2(x)$ the 3^{rd}-order approximation gives 10.5-bits accuracy while the 4^{th}-order approximation gives 13-bits accuracy. Notice that the approximations are not exactly equiripple because $log_2(x)$ is non-linear.

For 2^x, the 3^{rd}-order approximation provides 14-bits accuracy and the 4^{th}-order curve fit gives 19-bits of accuracy. The 2^x approximations are closer to equiripple because of

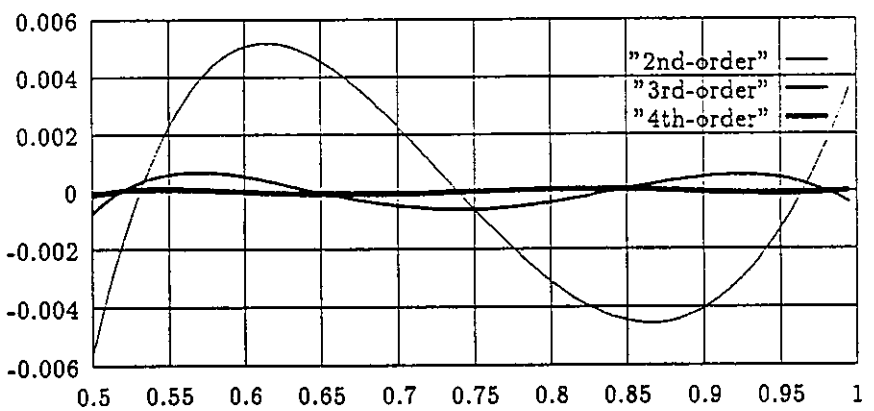


Figure 4.12: Error for Chebyshev Approximation over BFP Range for $log_2(x)$

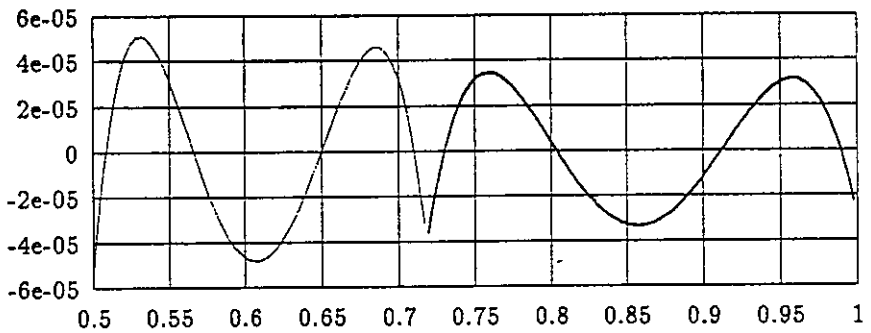


Figure 4.13: Error for Two Chebyshev Approximations to $log_2(x)$ for BFP range

the lower variation in slope over the input range. These results are excellent.

To improve the results for the 3^{rd}-order approximation to $log_2(x)$, two separate approximations can be applied across the input range. Figure 4.13 shows a plot of the error for such an approximation. The "break point" in the approximation was chosen to provide roughly equiripple error with 3^{rd}-order approximations for each segment of the input range. The coefficients for these approximations are tabulated in Appendix D. This approximation provides a peak error of 5.2125×10^{-5} or approximately 14-bits accuracy. An approximation of this type can be implemented with only a small increase in computation time—we only need to determine which range the input lies and specify a pointer to the desired polynomial coefficients.

Clearly, the 3^{rd}-order BFP Chebyshev approximation to 2^x is acceptable. If 10-bits accuracy can be tolerated then a 3^{rd}-order approximation for $log_2(x)$ could also be used. If this does not provide sufficient accuracy then two 3^{rd}-order approximations since they provide better accuracy than the 4^{th}-order approximation for a roughly equal increase in computation time.

It should be noted that for a fixed point implementation on the DSP56000 the coefficients will be quantized and there will be round-off error in the multiplications, particularly for high powers of x. This may cause slightly poorer results for the polynomial approximations.

4.7.4 Real-time Implementation

Approximation Method

If the approximation methods are compared for accuracy, we see that they all provide at least 10 bits of accuracy. The 3^{rd}-order Chebyshev approximation for 2^x provides 14-bits accuracy. The same range of accuracy ($\simeq$ 14-bits) can be achieved for $log_2(x)$ if two 3^{rd}-order approximations are applied across the BFP input range.

The look-up tables require much more memory than the polynomial approximations. Because it is convenient to use 24-bit entries (DSP56000 word size), look-up tables will use substantial portion of the $6k \times 24$ bit memory available on the (stock) DSP56000ADS. Chebyshev polynomial approximations require very little storage, just $(k+1)$ 24 bit values for a k^{th}-order approximation.

Table 4.4 shows the estimated number of clock cycles to compute the approximation given a normalized input value. (De)Normalization is not considered since it is used by all approximations. All possible "tricks" are used to get the best performance from each method.

Clearly, the 3^{rd}-order BFP Chebyshev polynomial approximation provides that lowest computation time for the desired accuracy. It also uses the least memory. This method was selected for real-time implementation.

BFP Approximation Type	clock cycles	
	best case	worst case
interpolated table	54 (N=256)	66 (N=2048)
non-uniform table	24 (N=256)	30 (N=2048)
3^{rd}-order Chebyshev	20	26 (two approx.)

Table 4.4: Computation Time for Selected Approximation Methods

Normalization and Denormalization

The normalization and denormalization of input/output values are two distinctly operations: normalization can be generalized as a search operation. We want to find the largest bit that is set in a binary number—that is, we have no prior knowledge of which bit is set. Denormalization is a right shifting an amount specified by the exponent. This operation can be quickly done using a small look-up table and multiplication where the exponent provides the address of the required shift constant.

Numerous search methods may be applied to normalize a number. Two simple algorithms, *exhaustive search* and *binary search* were investigated. Theoretically, binary search should execute much faster than exhaustive search: $N/log(N)$ faster for a search through N elements [16]. However, in practice the additional overhead required for set-up reduces the practical advantage of a binary search over an exhaustive search.

An exhaustive search is easily implemented via bit-shifting. To simplify the implementation, the DSP56000 instruction set provides the *norm* instruction [15]. A conditional jump instruction (*jnn*) that redirects program execution depending on the state of the normalization flag in the SR and a low-overhead loop instruction (*rep*) are also provided.

Two implementations of the exhaustive search were tested: a *norm—jnn* loop and a *rep #23—norm* loop. Table 4.5 shows the best case, worst case and average (estimated) execution times for these implementations.

A binary search uses the divide-and-conquer paradigm. At each iteration the location of the desired element is resolved to half of the previous possible locations. Tests on the DSP56000 showed that the set-up time for each execution of the search was significant. Thus, a combined binary/exhaustive search process was used. A single iteration of the binary search is done to "narrow down" the location of the MSB that is set to within 12-bits. Then, an exhaustive search is applied to find the MSB that is set within these 12-bits. Figure 4.14 shows a flow chart of the algorithm use to implement this search technique. Two methods, a *norm—jnn* loop and a *ref #11—norm* loop, were applied to the exhaustive search portion of the algorithm. The execution times for these algorithms are shown in Table 4.5.

Name	Type	Execution Times (clk cyc)		
		best	worst	average
norm1	*norm/jnn* loop	8	140	74
norm2	*rep#23/norm* loop	58	58	58
norm3	binary-*norn/jnn* loop	34	88	61
norm4	binary-*rep#11/norm* loop	48	54	54

Table 4.5: Execution Times for Search Algorithms on DSP56000

We see that norm4 provides the best performance. It also has the advantage of offering roughly the same execution time for the best and worst cases. This simplifies the analysis of the overall execution time of the DDRC. A complete implementation of the static characteristics is shown in Figure 4.14.

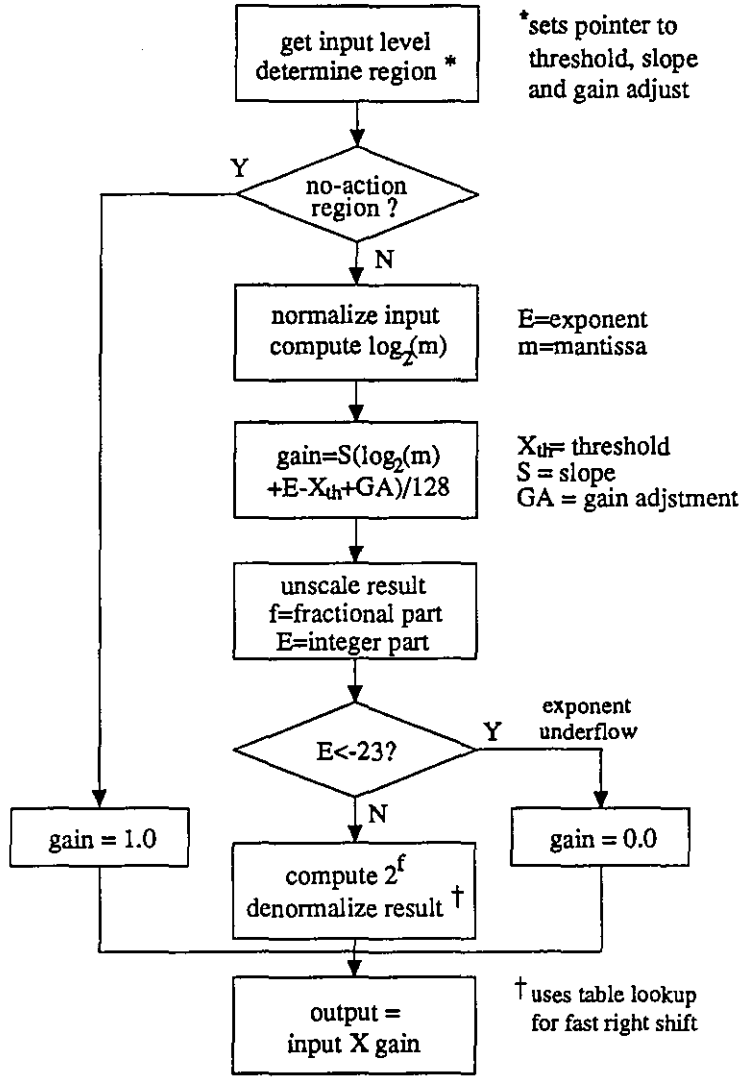


Figure 4.14: Flow Chart of Static Characteristics Implementation

4.8 Summary

The adaptive DDRC was implemented in real-time on a Motorola DSP56000ADS. It operates at the standard (Compact Disk) sampling rate of 44.1kHz with 16 bit input and output samples. The design is interrupt driven—the entire DDRC algorithm must execute between two interrupts in $1/44100 = 22.67\mu s$ or 464 clock cycles of the DSP56000.

The design uses a 150ms delay to adapt based on future statistics (with respect to the output) of the input signal. Our present design in constrained to 25ms delay because of memory and DSP56000 architecture limitations. For this short delay, the P_v calculation will be in error for very low frequency input signals. This problem is easily solved by purchasing more memory.

Numerous parameters for the design of the input/output section components were derived in this section. A time constant of 100ms was selected for the average level filter. This value is a compromise between the need for good averaging and a reasonable response time to program level changes. The static characteristics are computed using Chebyshev polynomial approximations and BFP arithmetic. To "speed-up" normalization, a combined binary/exhaustive search algorithm was used.

Chapter 5

Testing

5.1 General

This chapter describes the testing and evaluation of the adaptive DDRC. In previous chapters, the variables used to adapt the dynamic characteristics (Table 5.1) and the static characteristics (Table 5.2) of the DDRC have been presented. The values used for the static parameters are typical of those that would be used for a compressor/limiter. The adaptive parameter values were developed theoretically and refined through preliminary (informal) listening tests. Unless specified otherwise, the parameter values shown in these tables will be used for all tests conducted in this chapter.

The tests are split into two basic types: objective and subjective. Objective tests are used to confirm that the DDRC operates as designed. The majority of these tests are conducted using the DSP56000 simulator (SIM56000). This program exactly simulates a DSP56000 and runs the same compiled and linked *.lod* files as the evaluation board. The simulator can read and write a number of input/output files and provide an exact count of the clock cycles required for execution of the algorithm. At any point, simulations can be stopped and restarted. However, the simulator has the disadvantage of slow operation (eg. it takes 2.25 hours to process 186ms of audio); this limits the length of tests that

Section	Parameter	Function	Typ. Value
Level Measurement	a	average level filter parameter	(τ =100ms) 2.2673316×10^{-4}
	N	number of samples for P_v calculation	5
Recovery Adaptation	k_1	C_d weight	1.18827×10^{-4}
	k_2	P_v weight	1.18827×10^{-4}
	c	adaptive recovery starting point	0.99966562
	a	fixed T_r filter parameter	0.99988664 (T_r=200ms)
	Q	adaptive T_r threshold	0.86
Level Adaptation	V_{pc}	avg. control threshold	V_{pc} – 10dB
	V_{pc}	peak control threshold	-15dB (=L_{th})

Table 5.1: Parameters for DDRC Adaptation Control

Parameter	Function	Typ. Value
L_{th}	limit threshold	-15dB
C_{th}	compression threshold	-35dB
E_{th}	expansion threshold	-50dB
R_L	limit ratio	1/100
R_C	compression ratio	1/3
R_E	expansion ratio	2/1

Table 5.2: Parameters for Static Characteristics

can be conducted on the simulator.

Subjective or listening tests were conducted to determine the sonic performance of the adaptive DDRC relative to (almost) conventional designs. Through specific combinations of parameters, the adaptive DDRC can be configured to perform like conventional (non-adaptive) designs. The subjective tests will also show whether any subjectively displeasing artifacts are introduced by the adaptation. In these tests, no attempt will be made to arrive at the optimum set of adaptation parameters since this evaluation is beyond the scope of the present research.

5.1.1 Test Set-up

Simulator: Input files containing fractional samples (i.e. $-1 < x[n] < 1$) are generated with the desired waveforms using programs written in C. These are "ideal" signals containing no noise and having precisely controlled characteristics. The simulator (SIM56000) executes the DDRC algorithm and reads these input files. Output files of specified variables (e.g. the recovered signal level) are generated.

Development System: The test set up for real-time tests is shown in Figure 5.1. An adjustable gain mixer (Appendix E) was constructed to mix the left/right outputs of the CD player and provide a balanced input to the A/D converter. For a majority of the tests, a test CD (Denon Audio Technical CD #38C39-7147) was used. This CD provides standard test signals for distortion measurements. Sinusoids at various input levels (including 0dB peak level) are also provided. These signals proved invaluable for setting the input level to the A/D (i.e. the gain of the two channel mixer). It is imperative that the input level to the A/D be properly adjusted so that the DDRC receives the maximum input dynamic range.

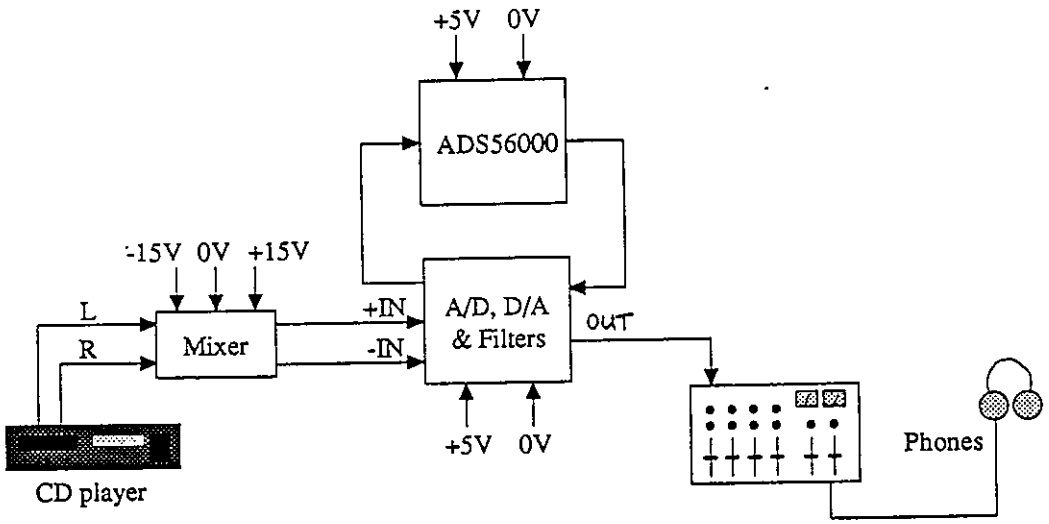


Figure 5.1: Real-time Test Setup

5.2 Objective Tests

5.2.1 Execution Time

As discussed previously, for the adaptive DDRC to operate in real-time, the algorithm must execute within one sample period (T_s) (where $T_s \simeq 22.67us$). The ADS56000 operates at a base clock rate of 20.48MHz. Thus, there are 464 clock cycles available to execute the entire algorithm. Table 5.3 shows the number of clock cycles required for each component of the algorithm and the percentage for each component expressed relative to the maximum execution time of 464 cycles. All features of the DSP56000 (i.e. parallel pointer updating, parallel data fetching and pointer addressing) are used to speed execution.

From this table, we see that the worst case path through the algorithm requires more time than is available. The worst case path through the algorithm will be followed rarely (possibly never). Some quick listening tests confirmed that the DDRC operated properly. There was no evidence of blocks of samples being skipped because the algorithm did not

Section	Component	# clock cycles	% total time
Input	interrupts and I/O	16	3.4
	HPF/setup pointers	26	5.6
	peak/zc detect	50	10.8
	avg. level	12	2.6
Output	zc detect	18	3.9
	adaptive level	56	12.1
	attack/recovery	48	10.3
	adaptive recovery	66	14.2
	region	28	6.0
	normalize	54	11.6
	$log_2()$	16	3.4
	compute gain	64	13.8
	$exp2()$	16	3.4
	denormalize	12	2.6
	output sample	4	0.9
Totals		486	104.7

Table 5.3: Execution Times for DDRC Algorithm Components

complete between interrupts. We could leave the algorithm as is. However, to ensure a reliable design, the HPF filter was eliminated from the input stage. This reduces the worst case execution time to 464 cycles. As discussed previously, this may have some adverse affects on the sonic quality of the DDRC if the input has a DC offset.

5.2.2 Static Characteristics

Tests on the static characteristics were conducted to determine the accuracy of the gain computation. This computation includes the normalization of the input value, the computation of polynomial approximations, calculation of the static characteristics and the denormalization of the result.

Initially, tests were conducted in real-time on the ADS56000 using the test setup shown in Figure 5.1 except a 1kHz input from a signal generator was used in place of the CD player signal source. A test program that reads an input sample and writes it to the output (*intioevb.asm*) was run of the ADS56000. The input level was set so that the output was just below (digital) clipping. At this point, the input level to the A/D was 1.596Vrms (measured differentially across the A/D input with a floating DVM). The output level was 1.042Vrms. These levels are the 0dB reference voltages for the input and output.

To test the static characteristics the adaptive DDRC program was run. The input level was varied and the output level was recorded. The adaptive DDRC was configured so that the adaptive level section was disabled and the peak input level was used for level control. The results of this test are tabulated in Table 5.4 and plotted (along with the ideal static characteristics) in Figure 5.2. These results show that for our test setup, the measured and theoretical static characteristics (over the range we were able to measure) differ by no more than 2dB. Our test setup had severe noise problems, we could not measure below -40dB input level because the signal was "buried" in noise (i.e. the test setup has a noise floor of approximately -45dB). Clearly, the noise introduced into our

Input Level (dB)	Output Level(dB)
0.0	-28.18
-1.02	-28.18
-1.50	-28.18
-2.80	-28.18
-3.78	-28.18
-5.88	-28.40
-8.88	-28/40
-17.25	-29.38
-22.74	-31.70
-25.68	-32.76
-29.90	-34.42
-36.48	-38.42
-40.00	-40.18

Table 5.4: Measured Static Characteristics

test setup by the surrounding equipment (computers, power supplies etc.) makes it very difficult to assess the accuracy of the static characteristics.

To obtain more accurate results, tests were conducted using the DSP56000 simulator. The adaptive DDRC was configured as described above and 1024 input levels from approximately -75dB to 0dB were applied to the input (512 uniformly spaced samples from -75dB to -40dB and 512 uniformly spaced samples from -40dB to 0dB). The results of this test are plotted in Figure 5.3. Figure 5.4 shows the error in these gain computations compared to floating point theoretical calculations.

The error plot (Figure 5.4) shows that the static characteristics of the adaptive DDRC differ by less than 0.008dB from the theoretical static characteristics over the usable input

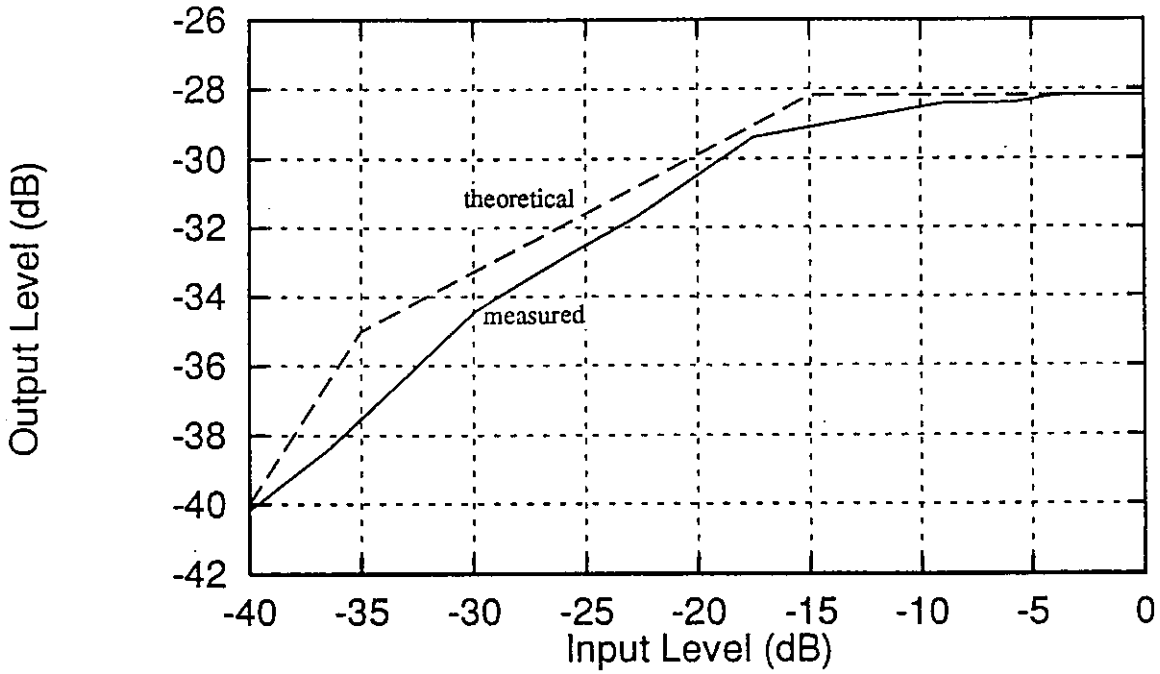


Figure 5.2: Measured Static Characteristics

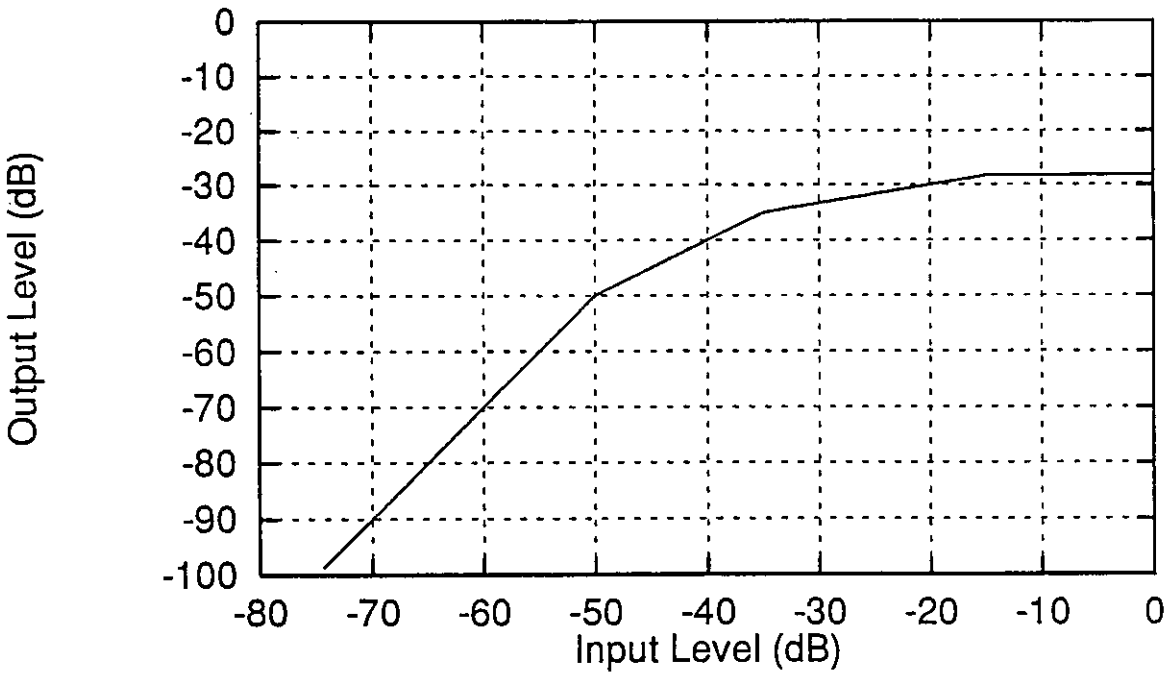


Figure 5.3: Static Characteristics Measured Using Simulator

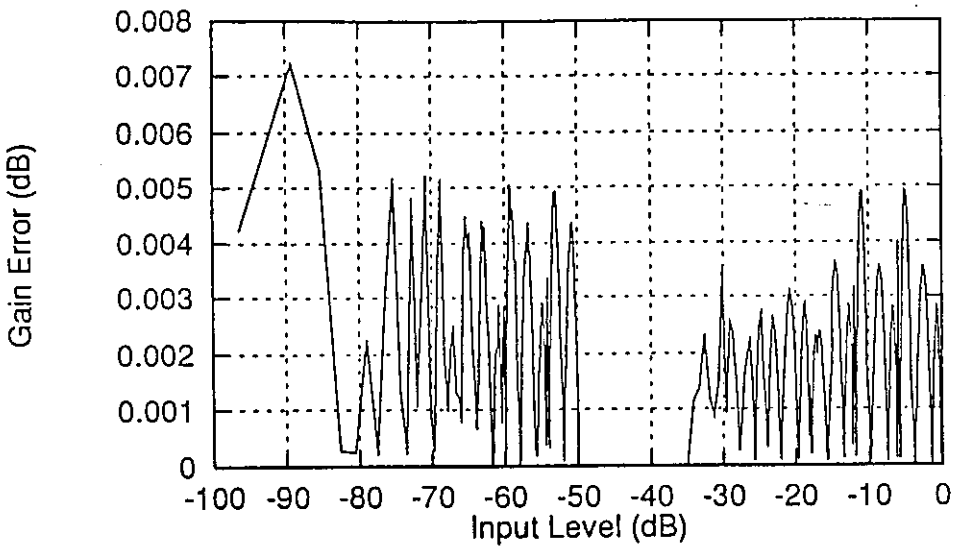


Figure 5.4: Error for Static Characteristics

range (i.e. before the output level underflows 16 bits). Clearly this is accurate enough since no one can perceive an error this small. The error plot shows increasing error as the input level falls below -80dB. This is a result of exponent underflow (i.e. $E < -23$) in the gain calculation. In the no action region of the characteristics, the only error is caused by the use of fixed point arithmetic since the gain is set to $1 - 2^{-23}$. In all other regions of the static characteristics (limiting, expansion and compression), the error is a combination of the polynomial approximation error, fixed point arithmetic error and error introduced by scaling.

5.2.3 Level Measurement

To check the operation of the average level measurement, the step response of the first-order averaging filter was measured. A unit step that changed from 0V to 1.0V at the first sample was used as the input. The output voltage at the 4091^{st} sample (for a 4096-point

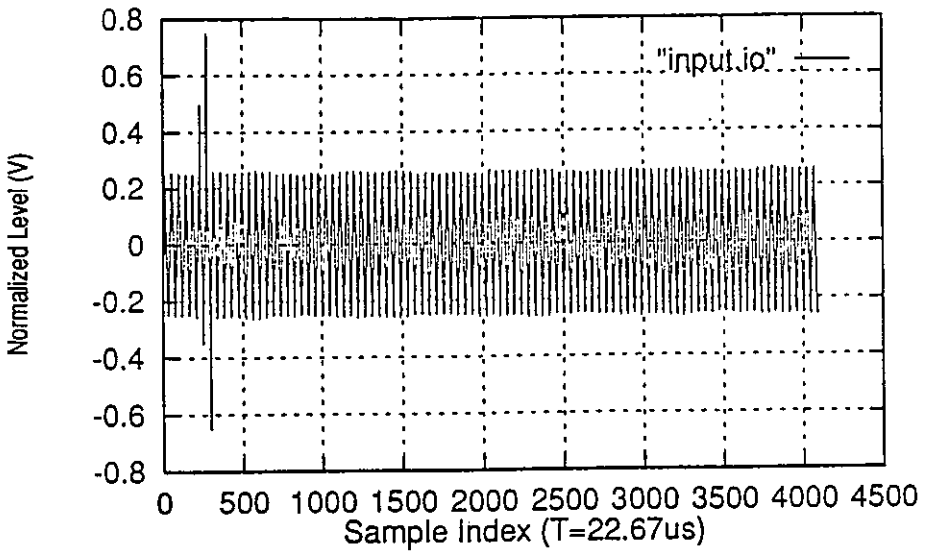


Figure 5.5: Input Waveform

input step) was $V_{out}[4091] = 0.6043706$. Thus, the time constant of the filter is

$$\tau = \frac{-4091(T_s)}{ln(1 - 0.6043706)} \simeq 100ms$$

This is as designed.

The peak level detector was tested by applying a 1kHz, 4096-point, 0.25V peak level sinusoid with some modified peak levels (Figure 5.5) to the input of the adaptive DDRC. Figure 5.6 shows the output peak level for a portion of the input waveform where the peak input levels have been altered. Notice that all peak levels are greater than V_{pc}. (Recall that $V_{pc} = -15dB$ or 0.17783 volts.) This causes the input signal peak level to be used for gain control.

Clearly, the peak detector operates as desired: the peak level is updated at the zero crossing preceding the peak. During attack, the peak level is constant between zero crossings (so that peaks are properly suppressed). Recovery begins once the present peak level falls below the recovered signal level.

A simple test was also conducted to confirm the operation of the adaptive level mea-

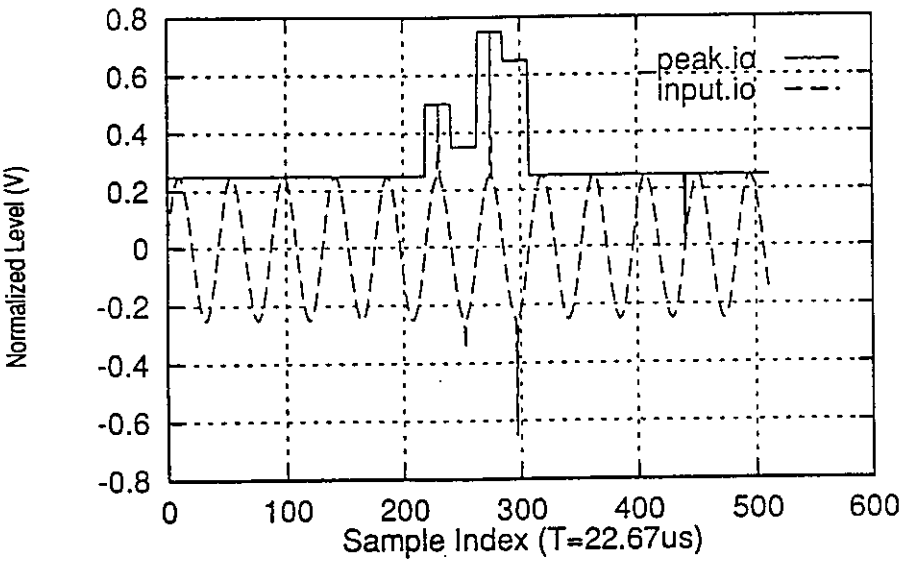


Figure 5.6: Peak Detector Output

surement. Using the simulator, the characteristics shown in Figure 3.3 were reproduced within the errors introduced by scaling and fixed point arithmetic (Section 4.5).

5.2.4 Dynamic Characteristics

Tests were also conducted to confirm the operation of the attack and (adaptive) recovery times. Typically, these tests are conducted using sinusoidal tone bursts with level changes at zero crossings [6]. However, a waveform of this type does not demonstrate the instantaneous attack capabilities of the adaptive DDRC (i.e. the gain changes at the zero crossings before the transient).

The dynamic characteristics of the adaptive DDRC operate in three different modes. Let the peak level between the present output zero crossing be p[k]. Recall that τ is the time constant of the filter used to measure the average signal level. Also, let $T_{\tau f}$ be the recovery time of the adaptive recovery filter. When the adaptive signal level is controlled by the average level (i.e. $p[k] < V_{pc}$), we have $T_a = \tau$ and $T_r = f(\tau, T_{\tau f})$. When the adaptive signal level is controlled by the peak level between zero crossings (i.e.

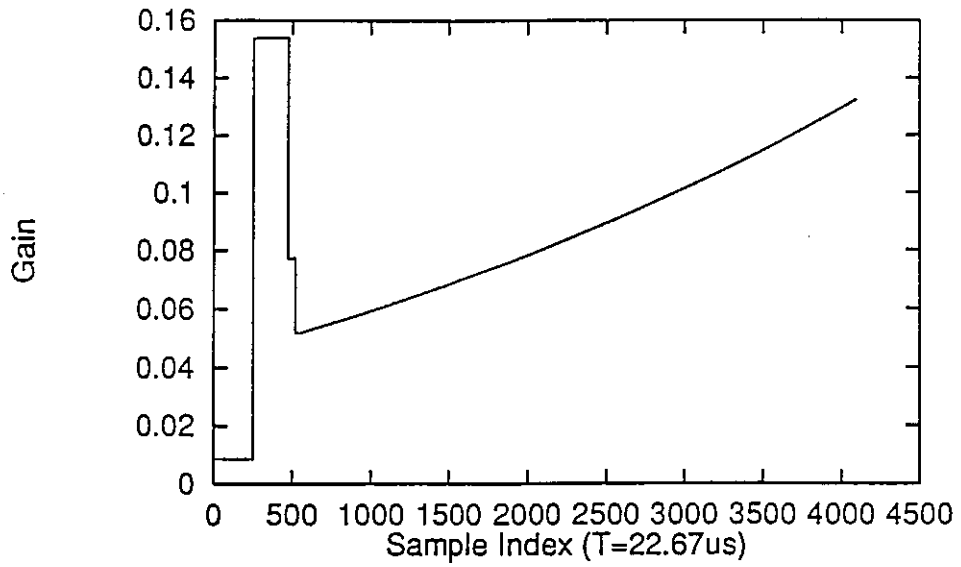


Figure 5.7: DDRC Gain Response to Isolated Peaks

$p[k] > V_{pc}$), T_a is instantaneous and the gain attacks at the zero crossing preceding the transient. For this case, $T_r = T_{rf}$. If $V_{pc} \leq p[k] \leq V_{pc}$ the adaptive signal level is a linear combination of $p[k]$ and the average signal level. Then, T_a and T_r will lie between the two extremes presented above.

We want to test the instantaneous attack time and the adaptive recovery time. Thus, the input signal must be such that $p[k] > V_{pc}$ for all k. For these tests, the sinusoidal signal presented above (Figure 5.5) was used. This signal has isolated transients that will demonstrate the unique attack and recovery characteristics of the adaptive DDRC. To obtain a reasonable number of non-zero output samples for a 4096-point input signal, the input and peak buffers were set to 256 samples. This size ensures that the P_v calculation will always be done using "valid" peaks for a 1kHz input waveform (Section 4.3.2).

Figure 5.7 shows the DDRC gain response to the input signal. When the peak level is used to control the gain, the attack is instantaneous and it precedes the transient. The gain is held constant between zero crossings during attack so all peaks between the zero crossings are properly suppressed.

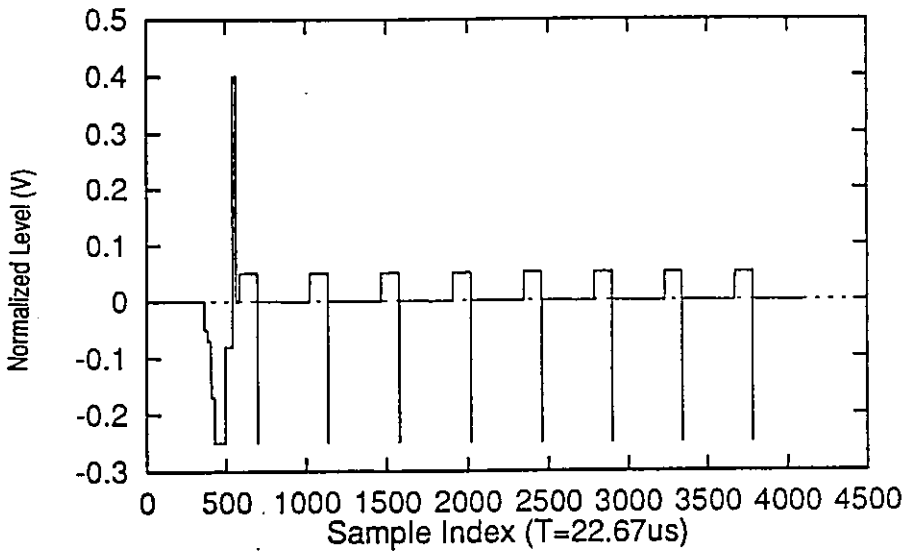


Figure 5.8: Peak Variation Measurement

To test the adaptive recovery operation, the peak variation (P_v) and crest difference (C_d) measurements were examined. Again, the input signal shown in Figure 5.5 was used. Figure 5.8 shows the P_v for this input signal. P_v is constant when the DDRC attacks since it is only computed during recovery. The initial 277 samples of the P_v are zero (256 while the peak buffer empties and 21 while the DDRC attacks). As expected, when the future average peak level is large, $P_v < 0$. For a smaller future average peak level, $P_v > 0$. For a constant peak level (i.e. a periodic signal), $P_v \simeq 0$.

The P_v for this input waveform (Figure 5.5) also demonstrates a disadvantage of the simple two-sample zero crossing detection method. For the periodic part of the input signal, we expect $P_v = 0$. However, the sinusoidal input signal has periodically occurring 0V samples. This causes the peak detector to store a peak level of zero. When this peak value of zero is time shifted through the P_v calculation, $P_v = 0.25 - 1.0/5 = 0.05$ when the peak value of zero is in the future. When the present peak level is zero, $P_v = 0 - 1.25/5 = -0.25$. For a single 0V sample, this transient lasts for only $6(T_s) = 136.02\mu s$ which is not long enough to be audible. Thus, this small anomaly will not have any affect on the

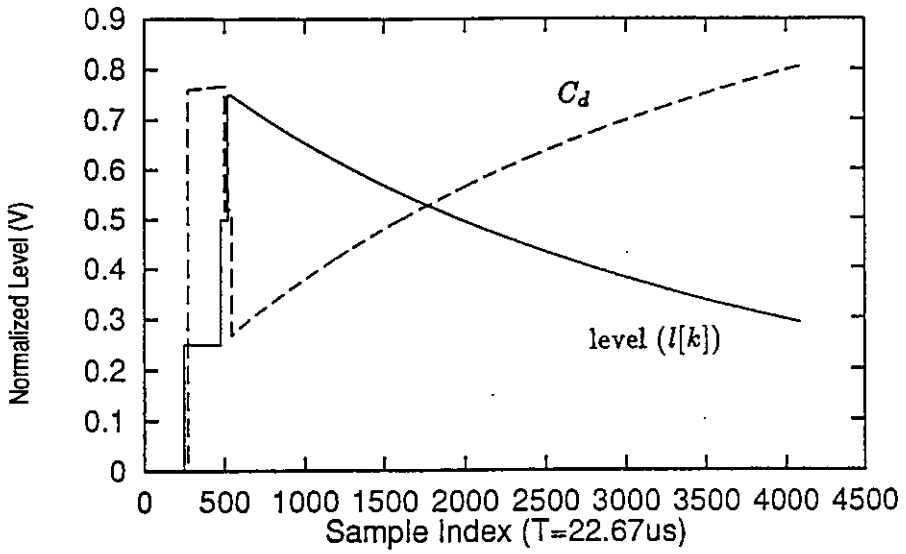


Figure 5.9: Crest Difference Measurement

sonic performance of the DDRC.

C_d for this input waveform is shown in Figure 5.9. Compared to the recovered signal level (Figure 5.9), the average signal level is small since the simulation is only 90ms long. Thus, $C_d \simeq 1 - l[k]$. We see that where peaks are large, C_d is large ($C_d = 0.75$). Where the peaks are small, C_d is smaller. Since $C_d < 0.86$, T_r is always adaptive.

T_r for this same input waveform (Figure 5.5) is shown in Figure 5.10. T_r was found by post-processing the recovery filter coefficient. The first 277 samples are not shown because they are constant at the initialized value of zero. Clearly, T_r lies within specified range of $50ms < T_r < 200ms$. Also, when the input signal (Figure 5.5) has a high peak level (relative to the average level), T_r is short ($\simeq$ 80ms). T_r increases as the gain recovers and the signal level ($l[k]$) decays towards the average signal level. Because the level ($l[k]$) will take a minimum of 50ms to recover, T_r does not adapt too quickly. The anomaly in the P_v calculation (discussed above) does not have any serious affect on the recovery time computation.

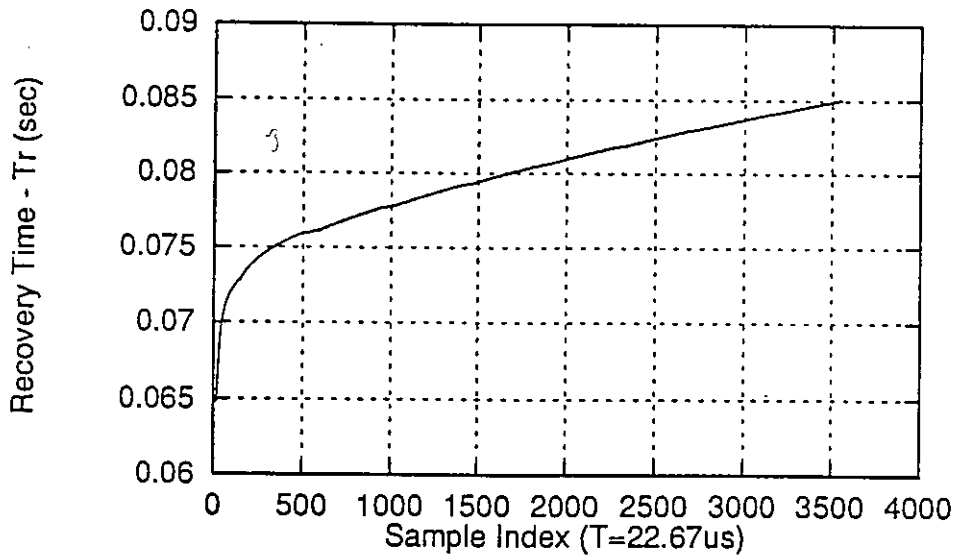


Figure 5.10: Recovery Time (T_r)

5.2.5 Distortion Measurements

To perform dynamic range control, the adaptive DDRC uses time-varying non-linear processing. These operations will generate harmonic and intermodulation distortion products. Because, the DDRC performs all calculations in the digital domain, it is possible for these components to alias into the passband of the corresponding lowpass filter. (In an analog system, these components are removed by low pass filtering.) This problem is most severe in the level measurement section where the input is the full bandwidth audio signal [12]. To assess the effects of the distortion on the DDRC performance a harmonic distortion and two intermodulation distortion tests were conducted.

In the past many methods have been developed for the characterization of distortion [6]. Typically, these measures cannot gauge the subjective effects of distortion. However, a distortion analysis based on simple input signals can provide a means of comparison with other devices. Also, by examining the input and output spectra, insight into the nature of the device characteristics can be gained. Also, it can provide a basis for optimizing the parameters of the device under test.

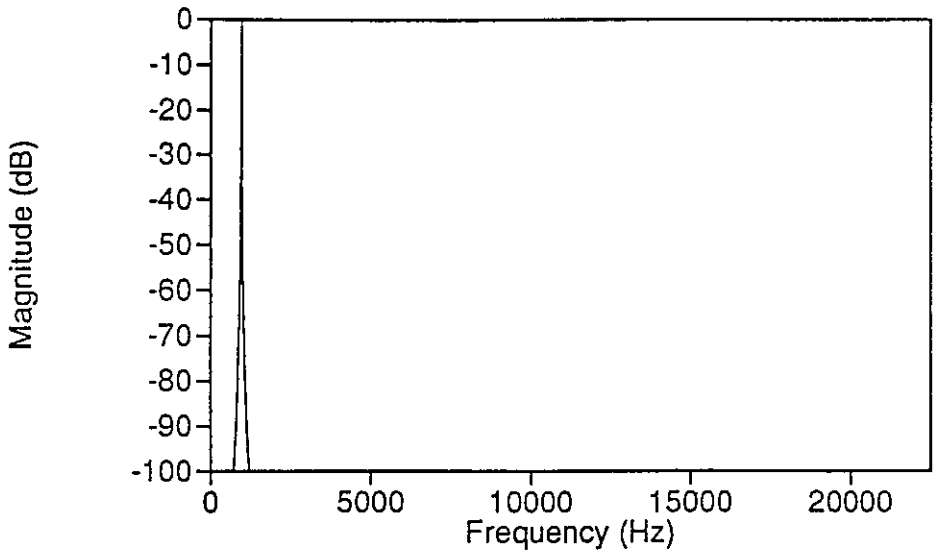


Figure 5.11: Input Spectrum for THD Test

To determine the distortion of the adaptive DDRC, a total harmonic distortion (THD) test and two intermodulation distortion (IMD) tests were conducted. Our original plan was to use test signals provided on the Denon Test CD and perform the tests in real-time. However, our test setup had a noise floor of approximately -45dB. This made it difficult to make accurate measurements. Thus, simulations were used for both tests.

For both tests, peak controlled gain was used. This method of gain control introduces the most distortion since the peak level (peaks between zero crossings) is not as smooth as the average signal level. Also, the output reaches steady state much faster when the peak level is used for gain control— this allows shorter duration simulations. Both simulations were run for 8192 samples. Spectral plots were made using an FFT of the last 4096 samples to ensure that transients (eg. the buffers emptying at initialization) did not affect the results.

For the THD test, a 1.0V peak (i.e. full-scale input), 1kHz sinusoid was used. Figures 5.11 and 5.12 show the input and output spectra for the THD test. From these results, the THD was computed to be (using Equation 3.1) 0.000031%.

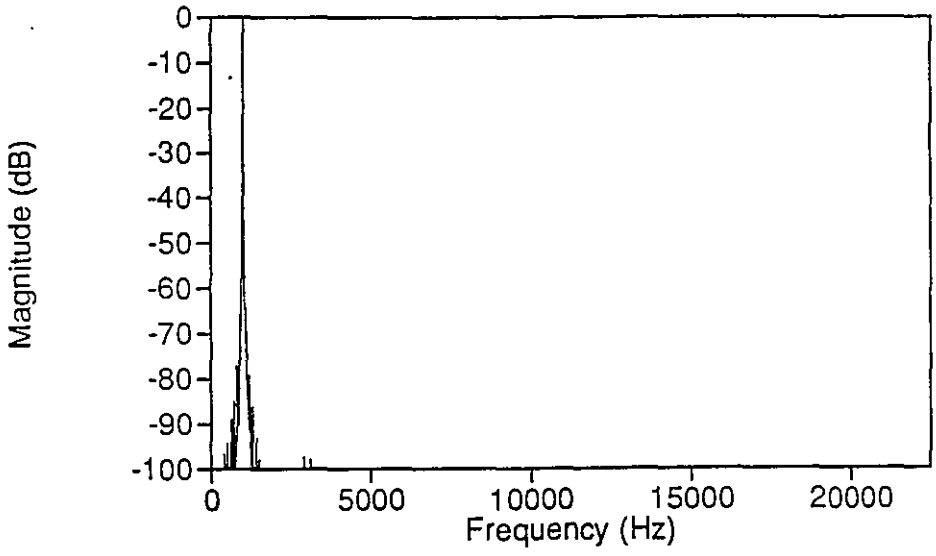


Figure 5.12: Output Spectrum for THD Test

For the IMD test, a standard IEC test signal

$$x(t) = 0.80sin(2\pi f_1 t) + 0.20sin(2\pi f_2 t)$$

where $f_1 = 250Hz$ and $f_2 = 8020Hz$ was used. Figures 5.13 and 5.14 show the input and output spectra for these tests. The IMD was computed (using Equation 3.1) to be 0.000525%.

Both of these tests result in distortion levels that are inaudible. For the THD test, this is expected. The adaptive DDRC should have no steady state distortion for a single sinusoidal input because the device simply computes a gain based on the constant peak level and applies it to the input signal. Since the peak level is constant, the gain is constant and the output is simply the input at a lower level. The small amount of distortion that is present is a result of the inaccuracies in the gain computations and the use of fixed point arithmetic and scaling in the algorithm.

The IMD test results in a higher, but still inaudible distortion level. Again, this low distortion results from the almost constant peak level of the input signal. The small

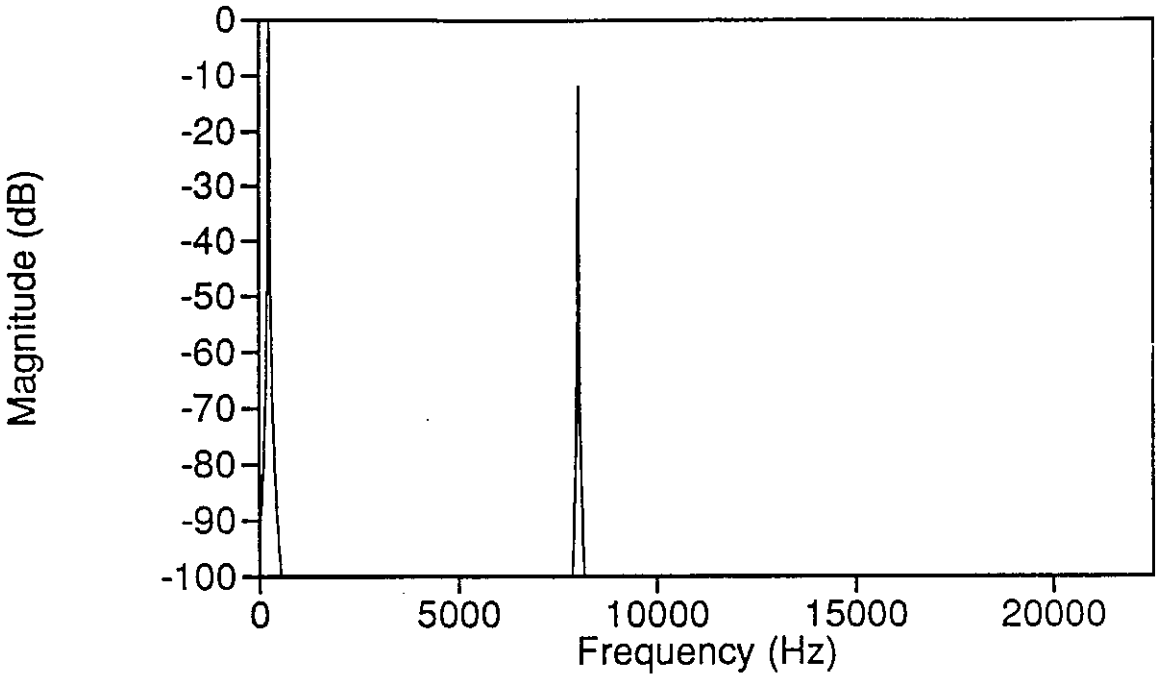


Figure 5.13: Input Spectrum for IMD Test

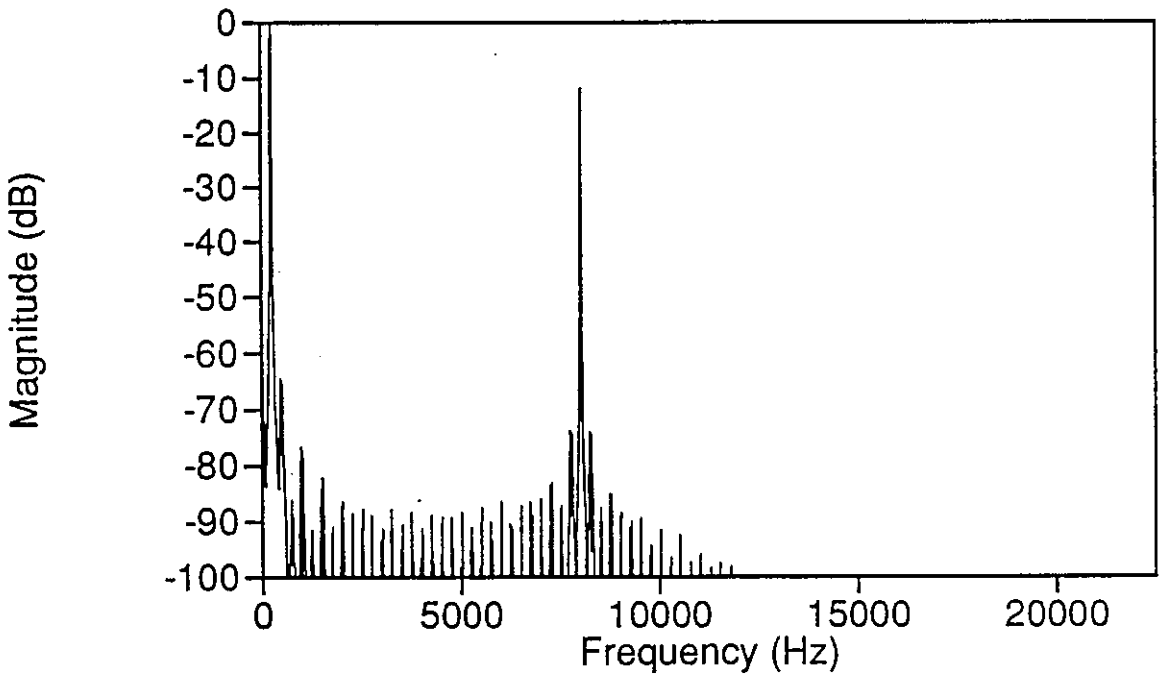


Figure 5.14: Output Spectrum for IMD Test

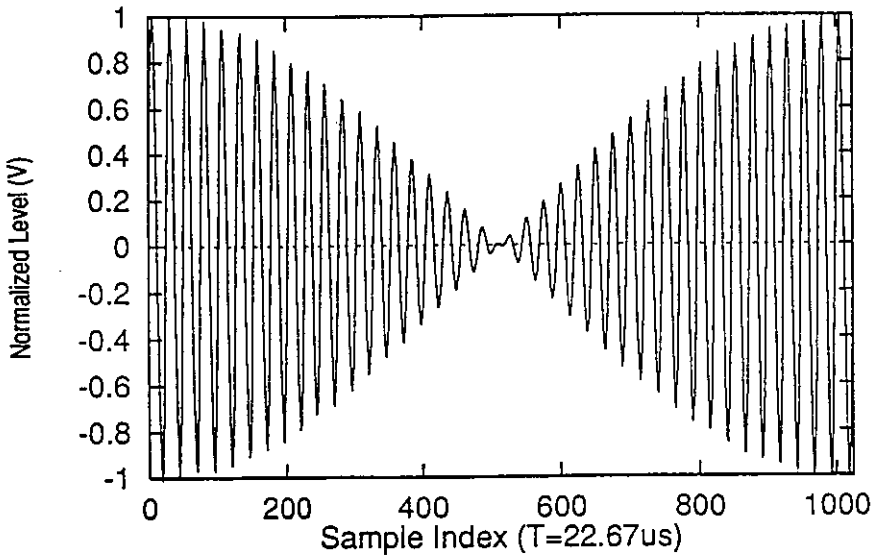


Figure 5.15: Input Signal for Second IMD Test

amount of distortion can be attributed to the same sources that are responsible for the harmonic distortion.

To obtain results that would better reflect the adaptive DDRC performance to "real world" (musical) input signals, we conducted an additional IMD test. The input signal (Figure 5.15) used was one employed by Mapes-Riordan and Leach [11] in their tests on a digital peak limiter. Their signal was 1024 points long. We concatenated eight such signals (8192 samples long) and applied this as an input to the adaptive DDRC. Figures 5.16 and 5.17 show the input and output spectra for the last 4096 samples of each signal. For these signals, we obtained a distortion level of 0.438%. If all 8192 input and output samples are used in the distortion calculations (i.e. the initialization and gain attack are included in the analysis), a distortion of 0.450% is obtained. The spectra for the 8192 point analysis was very similar to the 4096 point spectra. This test signal results in higher distortion levels than the previous IMD test because it has a larger variation in peak level.

These distortion levels may be audible. For their tests on a digital peak limiter

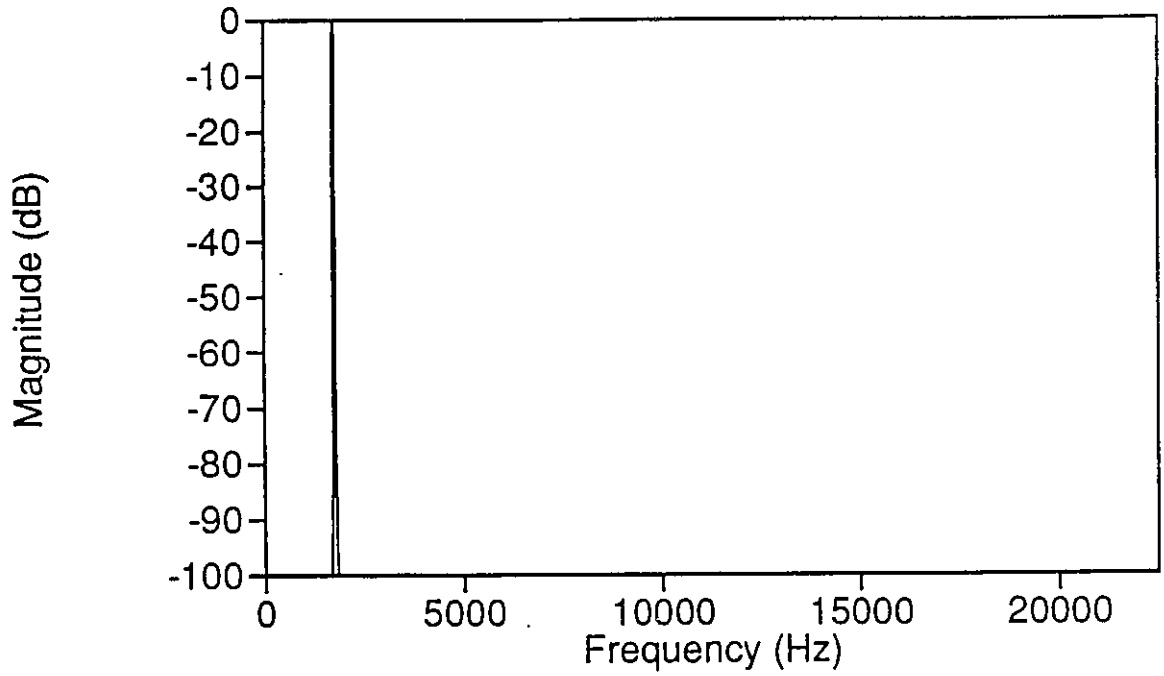


Figure 5.16: Input Spectrum for Second IMD Test

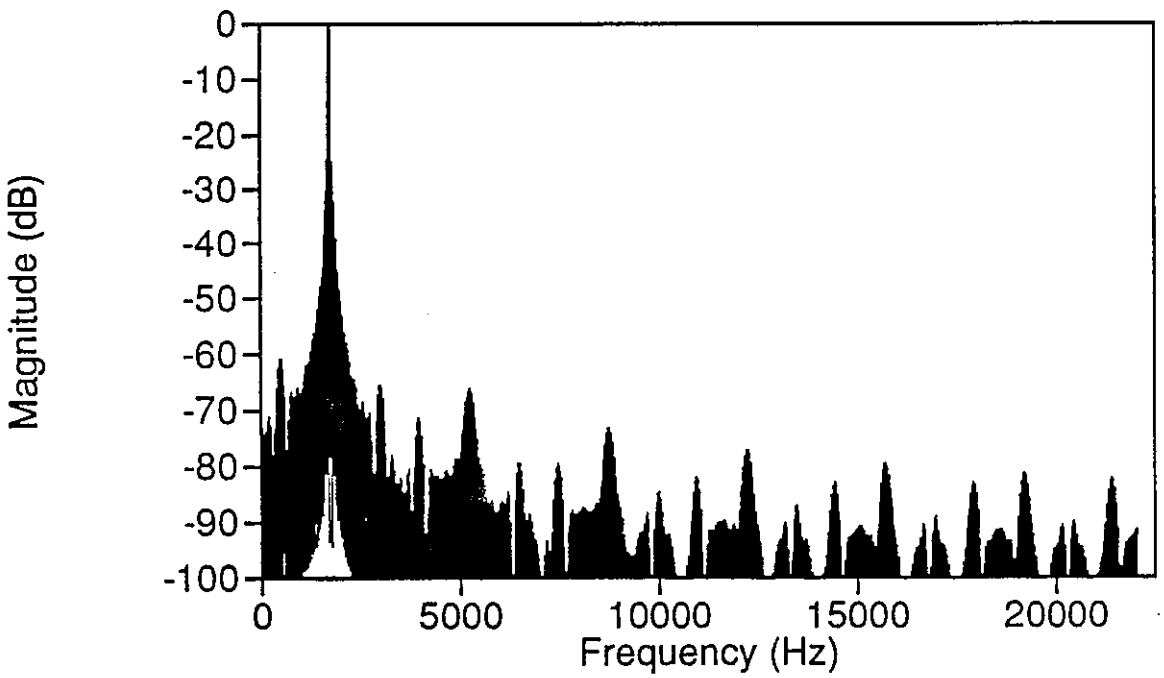


Figure 5.17: Output Spectrum for Second IMD Test

with gain attack and recovery at zero crossings, Mapes-Riordan and Leach [11] report a distortion of 5.4% with T_r=0. For long recovery times (on the order of those used in our adaptive DDRC) and a 6dB limiting level, they report distortions of "less than 1%." These results compare favourably with our results. They also indicate that good subjective results are obtained with these long recovery times.

5.3 Subjective Tests

Signal processing may sometimes generate annoying distortion or artifacts that will not be apparent in the objective tests. To uncover these problems, listening tests are usually performed— they are the definitive measurement of performance.

An important requirement of these experiments is one of presenting a totality of stimuli with maximal randomness of order to protect against the possible effects of presentation order and with sufficient redundancy for averaging out variability in subjects decisions ([10] - Appendix F). However, the large number of tests that are required to average out variability make listening tests very time consuming.

For our evaluation of the adaptive DDRC performance, we did not conduct exhaustive testing; instead, seven subjects were used to get an idea of how the performance of the adaptive DDRC compares to conventional DDRC designs. Comparative listening tests were used exclusively: listeners compared processed signals to unprocessed signals and rated their relative quality on a five point opinion scale (Table 5.5).

Preliminary listening tests showed that there was "clicking" present in the attack portion of some processed signals. This artifact was most evident when the average signal level was used for gain control. Clicking was present for both the adaptive and conventional designs. Through a number of tests, we determined that this artifact was not a result of input buffer overflow or (input) dc level modulation [18]. We concluded this artifact was *zipper noise* [12] caused by the combination of attack at zc's and the rise

Score	Quality	Comments
5	Excellent	sounds the same as the original
4	Good	compressed, but still of high quality
3	Fair	compressed with occasional annoying distortion
2	Poor	compressed with annoying distortion
1	Unsatisfactory	sound quality is unacceptable

Table 5.5: Five-point Opinion Scale

time of the average level measurement.

When the average signal level is used for gain control, an increase in level results in a exponentially increasing average level measurement. This results in a sequence of gain attacks at zc's. This is perceived as a sequence of clicks—zipper noise! McNally's adaptive gain smoothing scheme [12] could be used to eliminate the zipper noise. However, our algorithm currently uses 100% of the available processing time. Thus a faster processor or an improved DDRC algorithm (faster) would be required to allow this addition. Because this artifact has a known cause and cure, listeners were asked to ignore it in their evaluation of the Adaptive DDRC's sound quality.

Three high-quality test signals (compact disk "DDD" recordings) were selected for use in the listening tests. These selections represent typical wide dynamic range listening material.

- (A) Larry Coryell
  Scheherazade
  1. The Sea and Sindbad's Ship
  ©1983 Nippon Phonogram Co., Ltd., Tokyo
  (solo guitar)

- (B) Tears for Fears

The Seeds of Love
Badman's Song
©1989 Phonogram Ltd., London
(piano, electric guitar and heavy drums)

- (C) W.A. Mozart
  Violin Concerto No. 5 in A Major (KV 219)
  1. Allegro Aperto
  ©1988 Enigma Classics
  (string ensemble and solo violin)

Seven processing methods that were developed in the preliminary round of testing were used. (Note: Unless specified otherwise, the adaptation parameters in Table 5.1 are used. Recall that τ is the time constant of the average level filter).

1. Adaptive DDRC (τ =30ms)
2. Adaptive DDRC (τ =100ms)
3. Peak Level Controlled Gain
4. Average Level Controlled Gain
5. Adaptive Level and Fixed T_τ=200ms
6. Adaptive DDRC (Extended T_τ range 50ms to 250ms)
7. No Processing

Each signal was processed using all of the above methods to give a total of 21 test signals. Between 45s and 60s of each unprocessed/processed signal pair were recorded in random order on a high-quality cassette. The use of a cassette slightly degrades the unprocessed examples. However, all testing is relative so this minor degradation is not of

any consequence. In our opinion, the recorded versions of the processed signals suffered very little degradation.

Listeners were given the following list of terms to use when commenting on performance they felt was degraded:

- **holes:** the program is momentarily blanked out following a transient
- **muddy:** the program is noticeably modulated by its level (IM distortion)
- **dense:** the output program is heavily compressed. The original balance of the program components is lost or degraded.
- **strident:** the timbre of the program is altered. Some instruments sound "rough" or too "bright."
- **fuzzy:** harmonic distortion introduces additional harmonics

Listeners were allowed to use their own descriptive terms provided they explained them.

Seven listeners were used to obtain the results shown in Table 5.6. To obtain the mean opinion score results, all scores for each processing option were averaged. The raw data for this test is contained in Appendix F.

As expected, listeners judged the unprocessed signal to be of the highest (relative) quality. Signals compressed using the average signal level were judged to be of the next highest quality. This was also expected because average level gain control is better related to the perceived signal level. Thus, the signal is not as heavily compressed. The adaptive DDRC with $\tau = 30ms$ was ranked third in quality by the listeners. The subjects preferred this processing to the peak gain controlled example and either of the other adaptive examples (extended T_r and τ=100ms). The fixed T_r example was judged to be the worst quality.

These results indicate that the adaptive DDRC has the potential for better performance than conventional designs that use only the peak signal level for gain control. The

Test	Processing	Score	Rank
1	Adapt τ=30ms	4.00	3
2	Adapt τ=100ms	3.69	6
3	Peak Control	3.83	4
4	Average Control	4.38	2
5	Adapt (fixed T_r)	3.50	7
6	Adapt (extend T_r)	3.81	5
7	None	4.95	1

Table 5.6: Mean Opinion Scores

fact that two of the adaptive DDRC examples ranked lower than the peak gain control test show that further studies are required to obtain the "best" possible parameter set for the adaptive DDRC. Clearly, the adaptive DDRC achieves its intended goal of improved subjective performance. With further tests to achieve a better parameter set, the subjective evaluation of the Adaptive DDRC may almost equal that of the average level controlled gain example.

The comments made by listeners also provide some insight into the quality of the processed signals. Listeners reported that $\simeq$29% of the peak controlled gain processed signals contained holes. However, the other processed signals were reported as having roughly equivalent "hole counts" ($\simeq$14%). This means that either listeners were not reporting all the holes (some listeners did not indicate hearing any) or that the T_r adaptation algorithm (parameters) require refinement. When signals were rated as being low quality, they were usually reported as being strident or muddy. This results from poor control of the recovery time. These results also indicate that further tests are required to improve the adaptation parameter set.

5.4 Summary

The results for the subjective tests indicate that the Adaptive DDRC operates as designed. The Adaptive DDRC has very low total harmonic distortion. The intermodulation distortion results compare favourably with a previous design that was reported to have good subjective performance [11]. Concern expressed about the two sample zero crossing detection method used in the design was unwarranted. Simulations show that this has a negligible effect on performance.

Listening tests were conducted using seven listeners with seven processing methods and three different musical input signals. One Adaptive DDRC example was ranked third in average quality— behind the unprocessed signal and the average level controlled gain example. However, two adaptive DDRC examples were ranked below the peak controlled gain example. Clearly, the adaptive DDRC shows promise for improved performance. However, extensive listening tests are required to arrive at the "best" parameter set.

Chapter 6

Conclusions and Recommendations

This thesis describes the design and real-time implementation of an Adaptive DDRC. Three features of this design offer improved performance over previous DDRC designs.

1. **Attack at Zero Crossings:** The Adaptive DDRC attacks at the zero crossing preceding a transient. This reduces the *perceived* distortion introduced by gain attacks.

2. **Adaptive Recovery Time:** The recovery time is adjusted based on two input signal statistics (peak variation and crest difference) to avoid introducing "holes" into the output program, yet achieve low modulation distortion. The recovery time adaptation characteristics are controlled by user adjustable parameters.

3. **Adaptive Level Measurement:** For peak levels below the average control threshold, the average level controls the gain. When the peak level exceeds the peak control threshold (= limit threshold) the gain is controlled by the peak signal level. Between these thresholds, a linear combination of the peak and average signal levels

is used to provide a smooth transition from peak to average-control (and vice-versa). This arrangement gives improved perceptual performance because the average signal level (which is better related to the perceived signal level than the peak signal level) is used for gain control whenever possible.

Our test results show that the Adaptive DDRC performs well. The total harmonic distortion is insignificant. Intermodulation distortion measurements are comparable to those for another DDRC design [11] that is reported to have good subjective performance. Listening tests using seven listeners, three test signals and seven processings methods were conducted. In these tests, one configuration of the Adaptive DDRC received better mean opinion scores than conventional designs (peak level gain control, non-adaptive recovery time). Although these tests cannot be considered exhaustive, they do indicate that the Adaptive DDRC has promise.

Some versions of the Adaptive DDRC (different parameter sets) were rated below a conventional design (peak controlled gain). This indicates a need for more extensive testing to obtain the "best" set of adaptation parameters. With this "best" parameter set, our design offers the control of a peak limiter (i.e. the peak level never exceeds the limit threshold) with perceptual performance almost on par with a DDRC that uses average level gain control.

The present Adaptive DDRC implementation uses all of the available processing time (at a sampling rate of 44.1kHz). To achieve this execution time, an input HPF was eliminated. This degrades performance since input dc offsets can corrupt level measurements and cause "plops" in the output signal [18].

Under some conditions, the output signal contains an annoying artifact known as *zipper noise*. This artifact is caused by successive attacks at zero crossings when the average level is used for gain control. In his DDRC design, McNally [12] implemented an adaptive smoothing scheme to remedy this problem. The lack of processing time precluded the implementation of this smoothing scheme. A new implementation on a

faster processor should be investigated so listening tests with the HPF and adaptive smoothing schemes can be conducted.

Bibliography

[1] M. G. Duncan amd D. Rosenberg and G. W. Hoffman. Design of a universal compandor for the elimination of audible noise in tape, disc and broadcast systems. *J. Audio Engineering Society*, 23(8):610–621, October 1975.

[2] D. E. Blackmer. A wide dynamic range noise reduction system. *dB Magazine*, pages 54–56, Aug-Sept 1972.

[3] B. A. Blesser. Audio dynamic range compression for minimum perceived distortion. *IEEE Transactions on Audio and Electroacoustics*, AU-17(1):22–32, Mar 1969.

[4] R. S. Burwen. Design of a noise eliminator system. *J. Audio Engineering Society*, 19(11):906–911, December 1971.

[5] R. C. Cabot. Audio measurements. *J. Audio Engineering Society*, 35(6), June 1987.

[6] R. C. Cabot. Audio tests and measurements. K. Blair Benson, editor, *Audio Engineering Handbook*, chapter 16. McGraw-Hill, 1988.

[7] H. A. Chinn, D. K. Garnnett, and R. M. Morris. A new standard volume level indicator and reference level. *Bell System Technical Journal*, 19(1):94–137, January 1940.

[8] R. W. Hamming. *Numerical Methods for Scientists and Engineers.* Dover Publications Inc., second edition, 1986.

[9] Motorola Inc. *DSP56000ADS - Application Development System Reference Manual.* Motorola Inc., version 2 edition, 1989.

[10] N.S. Jayant and P. Noll. *Digital Coding of Waveforms.* Prentice-Hall, 1984.

[11] D. Mapes-Riordan and W. M. Leach. The design of a digital signal peak limiter for audio signal processing. *J. Audio Engineering Society*, 36(7/8):562–574, Jul/Aug 1988.

[12] G. W. McNally. Dynamic range control of digital audio signals. *J. Audio Engineering Society*, 32(5):316–327, May 1984.

[13] G. W. McNally and T. A. Moore. A modular signal processor for digital filtering and dynamic range control of high quality audio signals. In *ICCASP '81 Proceedings (Atlanta, Georgia)*, pages 590–594, 1981.

[14] A. J. Olivera. A feedforward side-chain limiter/compressor/de-esser with improved flexibility. *J. Audio Engineering Society*, 37(4):226–240, April 1989.

[15] Motorola Technical Operations. *DSP56000/DSP56001 Digital Signal Processor User's Manual.* Motorola Inc., rev 2 edition, 1990.

[16] R. Sedgewick. *Algorithms.* Addison-Wesley Publishing Company, 1983.

[17] S. D. Stearns. Fundamentals of adaptive signal processing. J. S. Lim and A. V. Oppenheim, editors, *Advanced Topics in Signal Processing*, chapter 5. Prentice-Hall, 1988.

[18] E. F. Stikvoort. Digital dynamic range controller for audio. *J. Audio Engineering Society*, 34(1/2):3–9, Jan/Feb 1986.

[19] D. B. Talbot. A satellite communications, broadcast-quality amplitude compander. *J. Audio Engineering Society*, 29(10):690–698, October 1981.

[20] W. M. Wagenaars, A. J. M. Houtsma, and R. A. J. M. van Lieshout. Subjective evaluation of dynamic compression in music. *J. Audio Engineering Society*, 34(1/2):10–17, Jan/Feb 1986.

Appendix A

Code Listings

C:\TODD\56K\JUL10\MAKEFILE 1

```
#-------------------------------------------------
#
#                Makefile for ddrc.lod
#
#       revisions: created Feb 6/91 - T. Schneider
#-------------------------------------------------

ddrc.lod: main.lnk gain.lnk coeff.lnk rshftab.lnk static.lnk
        lnk56000 -Bddrc.lod -M gain coeff rshftab static main

# main program
#
main.lnk: ioequ.asm input.asm main.asm
        asm56000 -B -Lmaint main

# ddrc algorithm [interrupt handler]
#
gain.lnk: gain.asm fshft.asm input.asm ioequ.asm
        asm56000 -B -Lgaint gain

# polynomial approximation coefficients
#
coeff.lnk: coeff.asm
        asm56000 -B -Lcoefft coeff

# lookup table for fast right shifts (denormalization)
#
rshftab.lnk: rshftab.asm
        asm56000 -B -Lrshftabt rshftab

# static compression parameters
#
static.lnk: static.asm
        asm56000 -B -Lstatict static
```

C:\TODD\56K\JUL10\MAIN.ASM 1

```
;**** main.asm ******************************************************************
;*
;*      Main module of adaptive DDRC program.
;*
;*      This program executes an endless loop and services interrupts
;*      from the SSI. The SSI interrupt service routine is the DDRC
;*      algorithm. This module is based on the program intioevb.asm
;*      by C. Thompson.
;*
;*********************************************************************************
        opt cex,cre,w
        page 160,66,4,4,5
        section ddrc

        include 'ioequ'

        xref    bufsiz

        org     x:
        xref    in_buf
        xref    pk_buf

        org     y:
        xref    gn_buf

;=======================================
;=      Start of Main Program
;=======================================
        org     p:$40

;---------------------------------------
;-      Mask all Interrupts
;---------------------------------------
        ori     #$03,MR

;---------------------------------------
;-      Initialize SSI and BCR
;---------------------------------------
        movep   #0,x:M_BCR      ;zero wait states to all memory
        movep   #$3000,x:M_IPR  ;SSI Rx interrupt priority

;---------------------------------------
;- Set SSI for external continuous
;- synchronous clock, normal mode.
;---------------------------------------
        movep   #$4000,x:M_CRA  ;set SSI word_length = 16

        movep   #$B200,x:M_CRB  ;word long frame sync, RX interrupts enab.,
                                ;external clock and frame sync

        movep   #$01FF,x:M_PCC  ;turn on SSI port

;---------------------------------------
;- Set up ADS board in case of force
```

C:\TODD\56K\JUL10\MAIN.ASM 2

```
;- break instead of force reset
;---------------------------------------
        movec   #1,sp           ;init stack pointer
        movec   #0,sr           ;clear loop flag/interrupt mask bits

;---------------------------------------
;-  Warm restart and 'one-time'
;-  initialization for DDRC
;---------------------------------------
_init   move    #logcof,r4              ;point to polynomial coefficients
        move    #(8-1),m4               ;mod 8 addressing

        move    #in_buf,r3              ;circular input buffer
        move    #(bufsiz-1),m3          ;mod(bufsiz) addressing
        move    #-1,n3

        ;* input buffer is initialized with small +ve values
        ;* it initially contains no zero crossings
        move    #pk_buf,r2
        move    #(bufsiz-1),m2          ;for INPUT section with
        move    #-1,n2                  ;mod(bufsiz) addressing

        move    #pk_buf,r6
        move    #(bufsiz-1),m6          ;for OUTPUT section with
        move    #1,n6                   ;mod(bufsiz) addressing

        ;* set r7 for mod(bufsiz) addressing
        move    #(bufsiz-1),m7
        move    #-1,n7

        move    #-1,n5
;---------------------------------------
;- fill input buffer with small +ve
;- samples.
;---------------------------------------
        move    #>$000001,a             ; a <- 2^-23

        rep #(bufsiz)                   ; fill input buffer with 2^-23
        move    a,x:(r3)+

        clr     a                       ; clear initial output value

;---------------------------------------
;- zero fill peak buffer.
;---------------------------------------
        rep #(bufsiz)                   ; zero fill peak buffer
        move    a,x:(r2)+

;---------------------------------------
;-      Unmask all Interrupts
;---------------------------------------
        andi    #$FC,MR

;---------------------------------------
```

C:\TODD\56K\JUL10\MAIN.ASM 3

```
;-      Loop & wait for Rx Interrupt
;---------------------------------------
self    jmp     self

;!!!    jmp     rdwrite

;---------------------------------------
;-  I/O Interrupt Handler
;---------------------------------------
        org     p:
        xref    rdwrite

;---------------------------------------
;- Interrupt exception handlers for
;- SSI RX-TX overrun/underrun
;---------------------------------------

; clear the exception flag for the SSI transmitter and return
txcept  movep   x:M_SR,a0               ; read status
        movep   a1,x:M_TX               ; send something
        nop                             ; place a BREAKPOINT here
        rti                             ; for debugging exceptions

; clear the exception flag for the SSI receiver and return
rxcept  movep   x:M_SR,a0               ; read status
        movep   x:M_RX,a1               ; receive something
        nop                             ; place a BREAKPOINT here
        rti                             ; for debugging exceptions


;---------------------------------------
;- Install interrupt handlers
;- in vector table
;---------------------------------------
        org     p:I_SSIRD       ; SSI interrupt vector-receive
        jsr     rdwrite

        org     p:I_SSIRDE      ; SSI exception vector-receive
        jsr     rxcept

        org     p:I_SSITDE      ; SSI exception vector-transmit
        jsr     txcept

        endsec
        end
```

C:\TODD\56K\JUL10\GAIN.ASM 1

```
;**** gain.asm ***************************************************************
;*
;*      Adaptive DDRC algorithm
;*      This module is an interrupt service routine. In response to a
;*      SSI Rx interrupt, it reads an input sample, writes an output
;*      sample (computed last time the algorithm was executed) and
;*      calculates the output sample for the next SSI Rx interrupt.
;*
;*****************************************************************************
gain    ident 0,1
        opt cex,cre,w
        page 140,66,4,4,5
        section gain
;-----------------------------------------------
;-      Include files for the DDRC routine
;-----------------------------------------------
        include 'ioequ'         ; std. io equates

        include 'input'         ; constants for input section etc.

        include 'fshft'         ; "fast" shift macros

        org y:
        xref logcof             ; polynomial coefficients [coeff.asm]

        xref unflow             ; fast right shift table [rshft.asm]
        xref rshtop
        xref rshbot

        xref limit              ; static parameters [static.asm]

        org x:
        xref lthresh            ; adaptive level thresholds [static.asm]

        org y:$047E             ; !!! temporary storage for testing only !!!
temp3   dc      $000000
temp4   dc      $000000

;=======================================
;= Start of DDRC routine
;=======================================
        org     p:$80
        xdef    rdwrite                 ; address installed in vector table
;---------------------------------------
;- Real-time I/O:
;- Read A/D data from the SSI RX register
;- and write data to the SSI TX register.
;- Both registers operate synchronously so
;- TX empty flag need not be checked. TX is
;- always empty due to reception of new A/D word.
;---------------------------------------
rdwrite     movep   x:M_RX,x0           ;A/D input-->x0
            movep   a,x:M_TX            ;write output-->M_TX
```

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```
;=========================================
;= Input Section of Routine
;=========================================
;*************************************************************************
;*
;*      Highpass filter the input signal. (1st order filter)
;*
;*      Entry:  x0 - input sample
;*              r3 - pointer to input buffer
;*              n3 - offset to previous input sample (-1)
;*
;*      Exit:   x0 - filtered output sample
;*              r5 - pointer to peak value
;*              r1 - pointer to avg. level filter coefficients
;*
;*************************************************************************
;*      move    #hpcoff,r5                      ; pointer to input params (y)
       move    #peak,r5
;*      move    #hpstate,r1                     ; pointer to input params (x)
       move    #iflтcof,r1

;*      move    y:(r5)+,y0                      ; y0 <- a0
;*      mpy     x0,y0,a                         ; a <- a0x[n]

;*      move    x:(r1),x1       y:(r5)+,y1      ; x1 <- x[n-1], y1 <- a1
;*      mac     x1,y1,a         x0,x:(r1)+      ; a <- a0x[n] + a1x[n-1]
                                                ; x[n-1] <- x[n]
;*      move    x:(r3+n3),x0                    ; x0 <- z[n-1]
;*      move    y:(r5)+,y0                      ; y0 <- -b1
;*      mac     x0,y0,a                         ; a <- a0x[n] + a1x[n-1]
                                                ;   - b1z[n-1]
;*      move    a,x0

;*************************************************************************
;*
;*      Find the peak level between zero crossings and store it
;*      in the peak level buffer.
;*
;*      Entry:  x0 - filtered input sample (z[n])
;*              r2 - pointer to peak buffer
;*              r3 - pointer to input buffer
;*              r5 - pointer to peak value (c_peak)
;*              n3 - offset to previous input sample (-1)
;*
;*      Exit:   y0 - input level squared (z[n]^2)
;*              r5 - pointer to peak level
;*
;*      Reg'r:  a,b,x0,x1,y1 - temporary storage
;*
;*************************************************************************
_pkdet  clr     b
        move    x:(r3+n3),x1    b,y1            ;x1<-z[n-1], y1<-0
        move    x0,b                            ; b <- z[n]
        abs     b       x0,x:(r3)+              ; in_buf <- z[n]
```

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```
        mpy     x0,x1,a         b,y0            ;a<-z[n]*z[n-1], y0<-|z[n]|
        cmp     y1,a            y:(r5),x1       ;IF (z[n]*z[n-1]>0), x1<-c_peak
        jgt     _notzc                          ;{ goto _notzc }

_iszc   move    x1,x:(r2)+      y0,y:(r5)       ;pk_buf<-c_peak, c_peak<-|z[n]|
        jmp     _avgl

_notzc  cmp     x1,b                            ;IF (c_peak > |z[n]|)
        jmi     _avgl                           ;{ goto _avgl }
                                                ;ELSE
        move    y0,y:(r5)                       ;{ c_peak <- |z[n]| }

;************************************************************************
;*
;*      Compute the average signal level (Output is q[n]).
;*
;*      Entry:  y0 - abs(input sample)
;*              r5 - pointer to previous st level
;*              r1 - pointer to filter coefficients (ifltcof)
;*
;*      Exit:   r5 - points to avg. signal level
;*
;************************************************************************/
_avgl   move    y:(r5)+,y1                      ; update r5 (y1 <- junk)

        move    x:(r1)+,x0                      ;x0<-tav1
        move    y:(r5),y1                       ;y1<-q[n-1]
        mpy     x0,y0,a  x:(r1)+,x1             ;a<-tav1*|z[n]|, x1<-omtav1
        mac     x1,y1,a                         ;a<-tav1*z[n]+omtav1*q[n-1]
        move    a,y:(r5)                        ; store st_avg

;=========================================
;=  Output Section of Routine
;=========================================
;************************************************************************
;*
;*      Find zero crossings at the output
;*
;*      Entry:  r3 - pointer to input buffer (@ z[n-N+1])
;*              r6 - output peak pointer (@ last peak)
;*
;*      Exit:   r6 - pointer to peak level between present & next zc
;*
;*      Reg'r:  x0,x1,a - temporary storage
;*
;************************************************************************
_outzc  move    x:(r3)+,x0              ; x0 <- z[n-N+1] (current output sample)
        move    x:(r3)-,x1              ; x1 <- z[n-N+2]

        mpy     x0,x1,a                 ; a <- z[n-N+1]*z[n-N+2]

        tst     a                       ; IF (z[n-N+1]*z[n-N+2]>0)
        jgt     _level                  ; goto _level
                                        ; ELSE
```

```
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        move    x:(r6)+,a               ; update output peak pointer (r6)
                                        ; (a <- junk )
;*************************************************************************
;*
;*      Compute adaptive level
;*
;*      Entry:  r5 - points to loc'n past stl (y-memory)
;*              r6 - points to current peak value (x-memory)
;*
;*      Exit:   b  - contains the adaptive level
;*
;*************************************************************************
_level  move    y:(r5)+,x0                      ; x0 <- s (short-term level)
        move    x:(r6),x1                       ; x1 <- peak (adapt via peak)
        move    y:(r5)+,a                       ; a <- 1.0
        move    y:(r5)+,b                       ; b <- Tac
        cmp     x1,b                            ; IF (1 < Tac)
        jpl     _dobee                          ; goto _dobee

        clr     a       y:(r5)+,b               ; a <- 0, b <- Tpc
        cmp     x1,b    y:(r5)+,y0              ; IF (1 > Tpc), y0 <- 1/d*scale
        jmi     _dobee                          ; goto _dobee

        mpy     x1,y0,a y:(r5)+,b
        add     a,b                             ; b <- (1 - Tpc)/(d*scale)
        move    b,y1
        MSHL    y1,x1,(9),b                     ; unscale result
        move    b0,a                            ; a <- (1 - Tpc)/d

        ;* preserve x0 = average signal level
_dobee  move    #one,b                          ; b <- 1.0
        sub     a,b     x:(r6),y0               ; b <- 1 - A, y0 <- p
        move    a,x1                            ; x1 <- A, y1 <- B
        move    b,y1                            ; y1 <- B
        mpy     x0,x1,b                         ; b <- s*A
        mac     y0,y1,b                         ; b <- s*A + p*B

;!!!    move    x0,b                            ; !!! testing !!! avg level
;!!!    move    x:(r6),b                        ; !!! testing !!! peak level

;*************************************************************************
;*
;*      Make attack/recovery decision, attack if (x[n]-y[n-1])>0
;*
;*      Entry:  b   - adaptive signal level, x[n]
;*              r3  - pointer to input buffer (@z[n-N+2])
;*              r1  - pointer to previous recovered output sample, y[n-1] .
;*              r6  - pointer to peak buffer (output section)
;*
;*      Exit:   a   - "recovered" or "attacked" level, y[n]
;*
;*************************************************************************
_attack move    x:(r1),y0                       ; y0 <- y[n-1]
        cmp     y0,b                            ; if (x[n] - y[n-1])<0
```

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```
        jmi     _recov                  ; goto _recov

;* Attack ... y[n-1] <- x[n]
        move    b,a                     ; a <- x[n]
        jmp     _arend

;*************************************************************************
;*
;*      Compute adaptive recovery filter coefficient.
;*
;*      Entry:  r5 - points to loc'n containing 1.0
;*              r6 - points to peak level
;*
;*      Exit:   x0 - contains recovery filter coefficient
;*
;*************************************************************************
        ;* compute Cd - relative peak measure
_recov  move    #radapt,r5              ; pointer to coefficients
        move    x:(r1),b                ; b <- prev recovered level (y[n-1])
        move    y:(r5)+,a               ; a <- 1.0
        sub     b,a                     ; a <- 1 - p
        move    y:stlev,b               ; b <- s
        add     b,a                     ; a <- 1 - p + s = Cd

        ;* compute peak variation (Pv)
        ;* r7 set for mod(bufsiz) addressing
        lua     (r6)+n6,r7              ; temp ptr to peak buffer (n6=1)
        clr     b       a,x0            ; b <- 0, x0 <- Cd (store Cd)
        move    x:(r7)+,x1              ; x1 <- p[k+1] (ie. future peak)
        move    y:(r5)+,y1              ; y1 <- 1/N

        rep     #N                      ; compute avg of future peak levels
        mac     x1,y1,b x:(r7)+,x1      ; b <- b + (1/N)sum(p[n+k])

        move    x:(r6),a                ; a <- p[n]
        sub     b,a                     ; a <- p[n] - (1/N)sum(p[n+k])

        ;* To adapt or not to adapt? What is the condition.
        ;* x0 contains Cd
        move    y:(r5)+,x1              ; x1 <- fixed a
        move    y:(r5)+,b               ; b <- Q
        cmp     x0,b                    ; IF (Cd > Q)
        jmi     _rlev                   ; goto _rlev (ie. a=fixa)
;!!!    jmp     _rlev                   ; !!! testing !!! no adaptation

        ;* compute adaptive recovery coefficient
        ;* a contains Pv
        move    a,x1    y:(r5)+,y1      ; x1 <- Pv, y1 <- k1
        mpy     x0,y1,a y:(r5)+,y1      ; a <- k1*Cd, y1 <- -k2
        mac     x1,y1,a y:(r5)+,y1      ; a <- k1*Cd - k2*Pv, y1 <- c
        add     y1,a                    ; a <- k1*Cd - k2*Pv + c
        move    a,x1                    ; x1 <- adaptive recov. coefficient

;*************************************************************************
```

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```
;*
;*      Recover if (x[n]-y[n-1])<0
;*
;*      Entry:  x1 - contains the filter coefficient, a
;*              y0 - contains the prev. recovered level, y[n-1]
;*              r1 - points to the prev. recovered level, y[n-1]
;*
;*      Exit:   a - contains recovered signal level (l[k])
;*
;*************************************************************************
_rlev   mpy     x1,y0,a                 ; a <- y[n-1]a = y[n]

        ;* store output for next time 'thru
_arend  move    a,x:(r1)                ; y[n-1] <- y[n]

;*************************************************************************
;*
;*      Determine the input region
;*
;*      Entry: a - input value.
;*
;*      Exit:  r1 - pointers to static parameters (+7).
;*             A - gain for no-action region (jmp to _gend only)
;*
;*      Reg'r:  x1 - temporary srotage of linear (ms) thresholds.
;*              n1 - offset for static parameter structures.
;*
;*************************************************************************
_region move    #limit,r1               ;point to static parameters
        move    #sstep,n1               ;step value for static data

        nop                             ;!!! PIPELINE DELAY !!!
        move    y:(r1)+n1,x1            ;get Lth and update r1

        cmp     x1,a    y:(r1)+n1,x1    ;get Eth and update r1
        jpl     _norm                   ;IF input>Lth THEN limit; goto _norm

        cmp     x1,a    y:(r1)+n1,x1    ;get Cth and update r1
        jmi     _norm                   ;IF input<Eth THEN expand; goto _norm

        cmp     x1,a    y:(r1)+n1,x1    ;get dummy thresh. and update r1
        jpl     _norm                   ;IF input>Cth THEN compr; goto _norm

;-------------------------------------
;- input level is in no-action region
;-------------------------------------
        move    #one,a                  ;no-action region has gain of 1.0
        jmp     _gend                   ;goto end of gain computation

;*************************************************************************
;*
;*      Normalize an input value supplied in x0 and correct the pointer
;*      to the static data.
;*
```

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```
;*      Entry:  a - input value
;*              r1 - static data pointer (+7)
;*
;*      Exit:   B  - normalized input value
;*              r0 - integer part or exponent
;*              r1 - points to static data (gain) for input region.
;*
;*      Reg'r:  x1 - left shift constant
;*              x0 - temporary input storage
;*              n1 - pointer adjustment value
;*
;*************************************************************************
_norm   move    #7,n1                   ;correction value for pointer
        move    a,x0                    ;x0<--input (set-up for compare)
        lua     (r1)-n1,r1              ;adj. static data pointer

        move    #0,r0                   ;initialize number of shifts
        tfr     a,b     #>k1,x1         ;set-up for norm./get left shift const.
        cmp     x1,a                    ;IF (k1>inval)
        jpl     _doshft                 ;THEN
        mpy     x0,x1,a #>-shft,r0      ;left shift input by shft & set up r0
        move    a0,b1                   ;set up for normalization

_doshft rep     #(shft-1)
        norm    r0,b                    ;normalize B
                                        ;exponent is in r0!

;*************************************************************************
;*
;*      Compute log2() of value supplied in B using a third order
;*      polynomial approximation (a1*x^3+a2*x^2+a3*x+a4)/8.
;*
;*      Entry:  B - normalized input value (i.e. 0.5 <= B < 1.0 )
;*              r4 - points to scaled polynomial coefficients
;*                   (initialized in main.asm - circular)
;*
;*      Exit:   A  - log2(B)/8
;*
;*      Reg'r:  x1 - contains x**3
;*              x0 - contains x
;*              y1 - contains x**2
;*              y0 - temp. storage for coefficients
;*
;*************************************************************************
_log2   move    b,x0                    ; put normalized input into x0
        mpyr    x0,x0,a                 ;compute x**2
        move    a,y1                    ;store x**2 in y1
        mpyr    y1,x0,a y:(r4)+,y0      ;compute x**3 - get a1
        move    a,x1                    ;store x**3 in x1

        mpyr    y0,x1,a y:(r4)+,y0      ;compute a1x**3 - get a2
        mac     y0,y1,a y:(r4)+,y0      ;compute a2x**2 and sum - get a3
        mac     y0,x0,a y:(r4)+,y0      ;compute a3*x and sum - get a4
        add     y0,a                    ;add a4 to result
```

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```
;*****************************************************************************
;*
;*      Compute the gain in log2() "dB".
;*
;*      Entry:  r0 - exponent (integer)
;*              A  - log2(m)/16 (fractional)
;*              r1 - points to array of static parameters
;*
;*      Exit:   x0 - fractional part of gain (16-bits accuracy)
;*              n0 - integer part of gain (table offset)
;*
;*      Reg'r:  x1 - temp storage
;*              x0 - temp storage
;*              y1 - temp storage
;*              y0 - temp storage
;*              A - temp storage
;*              B - temp storage
;*
;*      Comment Notation: E - integer exponent (two's complement)
;*                        m - fractional mantissia
;*                        S - gain slope
;*                        Xth - log2() threshold for input region
;*                        Adj - gain adjustment (non-zero for limiting only)
;*
;*      Modified:       Apr 28/91 - changed scaling to fix error and get
;*                                  extra bit of accuracy.
;*****************************************************************************
;*  Evaluate E*S
_gain   move    r0,x1                   ; x1 <- E
        MSHR    x1,y1,(scale+2),b       ; scale down by 2^5
                                        ; b0 <- E/32
        move    b0,x1                   ; x1 <- E/32
        move    y:(r1)+,y1              ; y1 <- S/4
        mpy     x1,y1,b    a,x1         ; b <- (E/32)*(S/4)
                                        ; x1 <- log2(m)/8
;*  Evaluate log2(m)*S
        mpyr    x1,y1,a                 ; a <- (log2(m)/8)*(S/4)
        asr     a
        asr     a                       ; a <- (log2(m)*S)/128

;*  Sum exponent, mantissia and unscale result
        add     b,a     y:(r1)+,x1      ; a <- S*(E + log2(m))/128
                                        ; x1 <- Xth/16
        mpy     x1,y1,b y:(r1),x1       ; b <- (S/4)*(Xth/16)
                                        ; x1 <- Adj/128
        asr     b                       ; b <- S*Xth/128

        sub     b,a                     ; a <- S(log2(m)+E-Xth)/128
        add     x1,a                    ; add gain correction

        clr     b       a1,x0           ; b <- 0, x0 <- result/128
        MSHL    x0,y0,(7+1),a           ; unscale result
                                        ; (extra shift to recover int. part)
        move    a0,b1                   ; b <- fract_part
```

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```
;*  Check integer part of exponent for underflow
        move    #m24,x1                 ; x1 <- -23
        move    #zero,a0                ; zero fractional part for compare

        move    #unflow,n0              ; use shift coefficient of zero
        cmp     x1,a                    ; if (int_part - -23) < 0
        jmi     _dnorm                  ; goto _dnorm

;*  Recover integer portion of result
;*      asl     b                       ; combined with above MSHL
        move    a1,r0                   ; store (int_part-1) in r0

        move    #$800000,x1             ; set-up to adj. sign of fract. part
                                        ; pipeline delay
        lua     (r0)+,n0                ; store corrected int_part
                                        ; (used for de-normalization)

;*  Recover fractional part of result ( b1 contains fract_part )
        asr     b                       ; get bits into correct places
        or      x1,b                    ; set MSB (all results are -ve)
        move    b1,x0                   ; set-up for exp2() calculation

;************************************************************************
;*
;*      Compute a third order polynomial approximation to 2^x for a
;*      fractional input x.
;*
;*      Entry:  x0 - fractional part of gain (db)
;*              r4 - points to polynomial coefficients
;*
;*      Exit:   A  - 2^(x)
;*
;*      Reg'r:  x1 - contains x**3
;*              x0 - contains x
;*              y1 - contains x**2
;*              y0 - temp. storage for coefficients
;*
;************************************************************************
_exp2   mpyr    x0,x0,a                 ;compute x**2
        move    a,y1                    ;store x**2 in y1
        mpyr    y1,x0,a y:(r4)+,y0      ;compute x**3 - get a1
        move    a,x1                    ;store x**3 in x1

        mpyr    y0,x1,a y:(r4)+,y0      ;compute a1x**3 - get a2
        mac     y0,y1,a y:(r4)+,y0      ;compute a2x**2 and sum - get a3
        mac     y0,x0,a y:(r4)+,y0      ;compute a3*x and sum - get a4
        add     y0,a                    ;add a4

;************************************************************************
;*
;*      Do right shift as specified by integer part of log2().
;*      (i.e. un-normalize result)
;*
;*      Entry:  a - 2^x for fractional part of input
```

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```
;*
;*      Exit:   a - right shifted result
;*
;*      Reg'r   x0 - temp storage for shift constant
;*              y0 - temp storage of input value
;*              r0 - table index
;*              n0 - integer part part result (= table offset)
;*
;*************************************************************************
_dnorm  move    #rshtop,r0              ;get table address
        move    a,y0                    ;y0<--2^(fract(gain))
        move    y:(r0+n0),x0            ;get shift constant from table
        mpy     x0,y0,a                 ;shift right
                                        ;result is in A
_gend   move    a,y1                    ; y1 <- gain

;----------------------------------------
;-  Apply Gain to Output Sample
;----------------------------------------
        move    x:(r3),x1               ; x1 <- z[n-N+1]
        mpyr    x1,y1,a                 ; output = a <- z[n-N+1]*gain
        asl     a

;========================================
;=      End of DDRC Algorithm
;========================================

        rti

        endsec
        end
```

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```
;*************************************************************************
;*
;*      User Adjustable Parameters for Static Characteristics.
;*
;*************************************************************************
static  ident   0,2
        title   'Static Parameters'
        opt     cex,cre,w
        page    160,66,4,4,5

        section static
;---------------------------------------
;- Thresholds in dB
;---------------------------------------
LTH     equ     -15.0
CTH     equ     -35.0
ETH     equ     -50.0

;---------------------------------------
;- Ratios are output/input in dB
;---------------------------------------
LRatio  equ     1.0/100.0
CRatio  equ     1.0/3.0
ERatio  equ     2.0/1.0

;*************************************************************************
;*  Internal Variables (DO NOT ALTER!!!)
;*************************************************************************
;---------------------------------------
;- Gain slopes
;---------------------------------------
LS      equ     LRatio-1
CS      equ     CRatio-1
ES      equ     ERatio-1

;---------------------------------------
;- Thresholds converted to log2()
;---------------------------------------
LTH2    equ     LTH/(20*@L10(2.0))
CTH2    equ     CTH/(20*@L10(2.0))
ETH2    equ     ETH/(20*@L10(2.0))

;- Gain adjustment for limiting (gain already applied by compression)
GA      equ     CS*(LTH2-CTH2)

;-------------------------------------------------------------------------
;- Static data.
;-      linear threshold: linear threshold (mean square values).
;-      gain slope: S/4
;-      log2(threshold): dB thresholds converted to log2()
;-------------------------------------------------------------------------
        org     y:$04A1
        xdef    limit
limit   dc      @pow(10.0,LTH/20.0)             ;linear threshold
```

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```
        dc      LS/4.0                          ;(gain slope)/16
        dc      LTH2/16                         ;log2(threshold)/16
;xxx changed GA/256 to GA/128 May 22/91 TS
        dc      GA/128                          ;gain adjust. for limiting
expand  dc      @pow(10.0,ETH/20.0)
        dc      ES/4.0
        dc      ETH2/16
        dc      0.0
compr   dc      @pow(10.0,CTH/20.0)
        dc      CS/4.0
        dc      CTH2/16
        dc      0.0
noact   dc      0.0
        dc      0.0
        dc      0.0
        dc      0.0

        endsec
        end
```

C:\TODD\56K\JUL10\INPUT.ASM 1

```
;*** input.asm ***************************************************************
;*
;*      User Adjustable Paramters
;*
;*****************************************************************************
times   equ     4095            ; # of times for gain loop !!! testing only !!!

;-----------------------------------------------
;- Average Level Filter Parameters
;-----------------------------------------------
tav1    equ     0.000226733     ; avg. level coefficient (tau=100ms)
;*tav1  equ     0.00007558      ; avg. level coefficient (tau=30ms)
omtav1  equ     1-tav1          ; coefficient for st_level filter

;-----------------------------------------------
;- Adaptive Level Thresholds
;-
;- TPC = peak control threshold > -50 dB
;- TAC = average control threshold (TPC - TAC > 10dB)
;-----------------------------------------------
TPC     equ     -15
TAC     equ     -25

;-----------------------------------------------
;- Adaptive Recovery Parameters
;-
;- fixa = fixed recovery time (low Cd signals)
;- Q = (crest difference threshold)^-1
;- k1 = crest difference coefficient
;- k2 = peak varitation coefficient
;- c = adaptation equation constant
;- N = # of future samples for Pv calculation
;-----------------------------------------------
;*fixa  equ     0.999886641             ; fixed Tr = 200 ms
;*k1    equ     0.000118827             ; max Tr = 200 ms
;*k2    equ     0.000118827             ; min Tr = 50 ms
;*c     equ     0.999665621

fixa    equ     0.999909309             ; fixed Tr = 250 ms
k1      equ     0.000285002             ; max Tr = 250 ms
k2      equ     0.000285002             ; min Tr = 25 ms
c       equ     0.999379205

Q       equ     0.86
N       equ     5
pvc     equ     1.0/(1.0*N)             ; 1/N ... coefficient for Pv samples

;*****************************************************************************
;*      INTERNAL PARAMETERS - DO NOT ALTER !!!
;*****************************************************************************
avmem   equ     $04B1
        xdef    bufsiz
bufsiz  equ     1150            ; size of input and peak buffers
```

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```
;-----------------------------------------------
;- adaptive level parameters
;-----------------------------------------------
Tac     equ     @pow(10.0,TAC/20.0)     ; linear adaptive level thresholds
Tpc     equ     @pow(10.0,TPC/20.0)
dlt     equ     (Tac-Tpc)               ; delta
dscale  equ     512.0                   ; scaling coeff. for adaptive level

;-----------------------------------------------
;- Storage Locations for Testing
;-----------------------------------------------
        org     x:avmem
inval   dc      $000000                 ;input value
temp1   dc      $000000                 ;temp storage !!! testing only !!!
temp2   dc      $000000

        org     y:$0000
stor1   dc      $000000
stor2   dc      $000000
stor3   dc      $000000
stor4   dc      $000000
stor5   dc      $000000
stor6   dc      $000000

;-----------------------------------------------
;- Input and Gain buffers
;-----------------------------------------------
        org     x:$0                    ;input signal buffer
        xdef    in_buf
in_buf  dsm     bufsiz

        org     x:$800                  ;peak signal buffer
        xdef    pk_buf
pk_buf  dsm     bufsiz

;-----------------------------------------------
;- State variable for input HPF
;- Coefficients for IIR averaging filters
;-----------------------------------------------
        org     x:$4c0                  ; HP state variables
hpstate         dc      $000000         ; x[n-1] (output)
ifltcof         dc      tav1            ; ST Avg. filter coefficients
                dc      omtav1
recovprev       dc      $000000         ; recov. filter - delayed output

;-----------------------------------------------
;- Coefficients for input HPF
;- Peak detection stuff
;- Adaptive level stuff
;-----------------------------------------------
        org     y:$4c0
;*hpcoff        dc      0.99893         ; a0 (HP filter coefficients)
;*              dc      -0.99893        ; a1
;*              dc      0.9978628       ; -b1
```

C:\TODD\56K\JUL10\INPUT.ASM 3

```
hpcoff          dc      $7fffff         ; no filter
                dc      0.0
                dc      0.0
peak            dc      $000000         ; peak between zero-crossings
stlev           dc      $000000         ; short-term average signal level
ladapt          dc      $7fffff         ; approx. 1.0
                dc      Tac             ; st average control threshold
                dc      Tpc             ; peak control threshold
                dc      1/(dscale*dlt)  ; scaled delta^-1
                dc      (-1.0*Tpc)/(dscale*dlt) ; scaled -Tpc/delta
radapt          dc      $7fffff         ; approx. 1.0
                dc      pvc             ; coefficient for Pv samples (1/N)
                dc      fixa            ; fixed recovery coefficient
                dc      Q               ; adaptive recovery threshold
                dc      k1              ; crest factor coefficient
                dc      -1.0*k2         ; Pv coefficient
                dc      c               ; constant for adaptation equation

;*****************************************************************************
;*
;*      Output Section Constants
;*
;*****************************************************************************
;-----------------------------------------------
;- Constants for determination of the region
;-----------------------------------------------
sstep   equ     4                       ;step size 'thru static data table
one     equ     $7fffff                 ;approx. 1.0

;-----------------------------------------------
;- Constants for log2() computation
;-----------------------------------------------
shft    equ     12                      ;bits to shift for binary search
k1      equ     @cvi(@pow(2,shft-1))    ;test value & shift constant

;-----------------------------------------------
;- Constants for gain computation
;-----------------------------------------------
scale   equ     4                       ;log2(scale_factor)
m24     equ     $FFFFE8                 ;used for underflow check
zero    equ     $000000                 ;used to zero fractional part
```

C:\TODD\56K\JUL10\RSHFTAB.ASM 1

```
;************************************************
;*
;*      Lookup table for FAST right shifts
;*
;************************************************
rshftab ident   0,3
        title   'Fast Right Shift Lookup Table'
        opt     cex,cre,w
        page    160,66,4,4,5

        section shift_table
        org     y:$0488

        xdef    unflow
        xdef    rshbot
        xdef    rshtop

unflow  dc      $000000                         ;underflow table
rshbot  dc      @pow(2,-23)                     ;2^-23
        dc      @pow(2,-22)
        dc      @pow(2,-21)
        dc      @pow(2,-20)
        dc      @pow(2,-19)
        dc      @pow(2,-18)
        dc      @pow(2,-17)
        dc      @pow(2,-16)
        dc      @pow(2,-15)
        dc      @pow(2,-14)
        dc      @pow(2,-13)
        dc      @pow(2,-12)
        dc      @pow(2,-11)
        dc      @pow(2,-10)
        dc      @pow(2,-9)
        dc      @pow(2,-8)
        dc      @pow(2,-7)
        dc      @pow(2,-6)
        dc      @pow(2,-5)
        dc      @pow(2,-4)
        dc      @pow(2,-3)
        dc      @pow(2,-2)
        dc      @pow(2,-1)                      ;2^-1
rshtop  dc      $7fffff                         ;approx. 2^0
        endsec
        end
```

C:\TODD\56K\JUL10\FSHFT.ASM 1

```
;* FAST SHIFT MACROS ***********************************************************
;* Macro definitions for generating fast right and left shift constants,
;* KR and KL, and performing the right and left shifts.
;*
;*      Let     s = the source register
;*              m = the multiplier register
;*              n = the number of bits to be shifted
;*              acc = the destination accumulator
;*
;* where s,m can be one of X0,X1,Y0,Y1 and acc can be A or B
;*
;* Taken from: "Fractional and Integer Arithmetic Using the DSP56000
;*              Family of General-Purpose DSP's" pg 9
;*
;*******************************************************************************

MSHR    macro   s,m,n,acc               ;four input variables
        move    #@pow(2,-n),m           ;load the multiplier register
        mpy     s,m,acc                 ;shift right n bits
        endm

MSHL    macro   s,m,n,acc               ;four input variables
        move    #>@cvi(@pow(2,n-1)),m   ;load the mult. reg.
        mpy     s,m,acc                 ;shift left n bits
        endm
```

C:\TODD\56K\JUL10\COEFF.ASM 1

```
;***********************************************
;*      Coefficients for 3rd-order polynomial
;*      approximationd to log2() and exp2()
;***********************************************
coeff   ident   0,4
        title   'Polynomial Coefficients for log2 and exp2'
        opt     cex,cre,w
        page    132,66,4,4,5

        section coefficients
;***********************************************
;*      Coefficients for 3rd order polynomial
;*      approximation to log2().
;***********************************************
        org     y:$0480
        xdef    logcof
logcof  dc      1.24356684/8.0          ;a1
        dc      -4.15703264/8.0         ;a2
        dc      6.05895642/8.0          ;a3
        dc      -3.14494063/8.0         ;a4

;***********************************************
;*      Coefficients for 3rd order polynomial
;*      approximation to 2^x
;***********************************************
        xdef    expcof
expcof  dc      0.03954285              ;a1
        dc      0.23088672              ;a2
        dc      0.69134016              ;a3
        dc      0.99994464              ;a4

        endsec
        end
```

C:\TODD\56K\JUL10\IOEQU.ASM 1

```
;*****************************************************************************
;*
;*      Equates for DSP56000 I/O registers and ports
;*
;*****************************************************************************

ioequ   ident   1,0

;---------------------------------------
;-
;-      Equates for I/O Port Programing
;-
;---------------------------------------

;       Register Addresses

M_BCR   equ     $FFFE           ;Port A Bus Control Register
M_PCC   equ     $FFE1           ;Port C Control Register

;---------------------------------------
;-
;-      Equates for SSI
;-
;---------------------------------------

;       Register Addresses

M_RX    equ     $FFEF           ;Serial Rx Data Register
M_TX    equ     $FFEF           ;Serial Tx Data Register
M_CRA   equ     $FFEC           ;SSI Control Register A
M_CRB   equ     $FFED           ;SSI Control Register B
M_SR    equ     $FFEE           ;SSI Status Register

;---------------------------------------
;-
;-      Equates for Exception Processing
;-
;---------------------------------------

;       Register Addresses

M_IPR   equ     $FFFF           ;Interrupt Priority Register

;*****************************************************************************
;*
;*      Equates for DSP56000 interrupts
;*
;*****************************************************************************

I_RESET         equ     $0000           ;H/W Reset
I_STACK         equ     $0002           ;Stack Error
I_TRACE         equ     $0004           ;Trace
I_SWI           equ     $0006           ;SWI
I_IRQA          equ     $0008           ;_IRQA
```

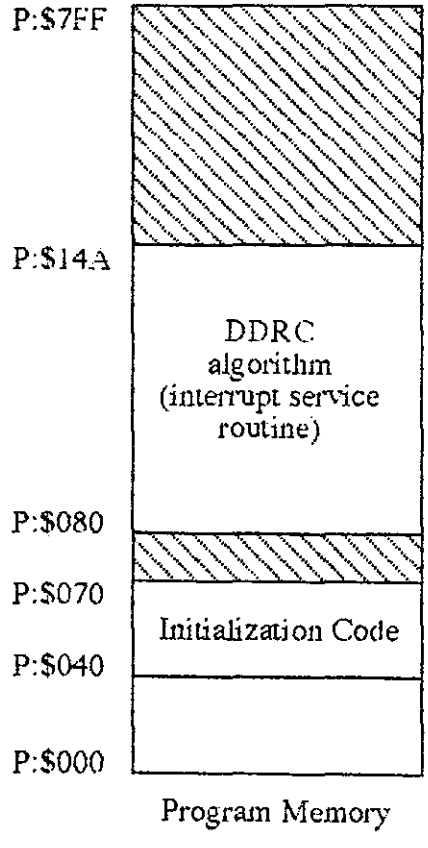
C:\TODD\56K\JUL10\IOEQU.ASM 2

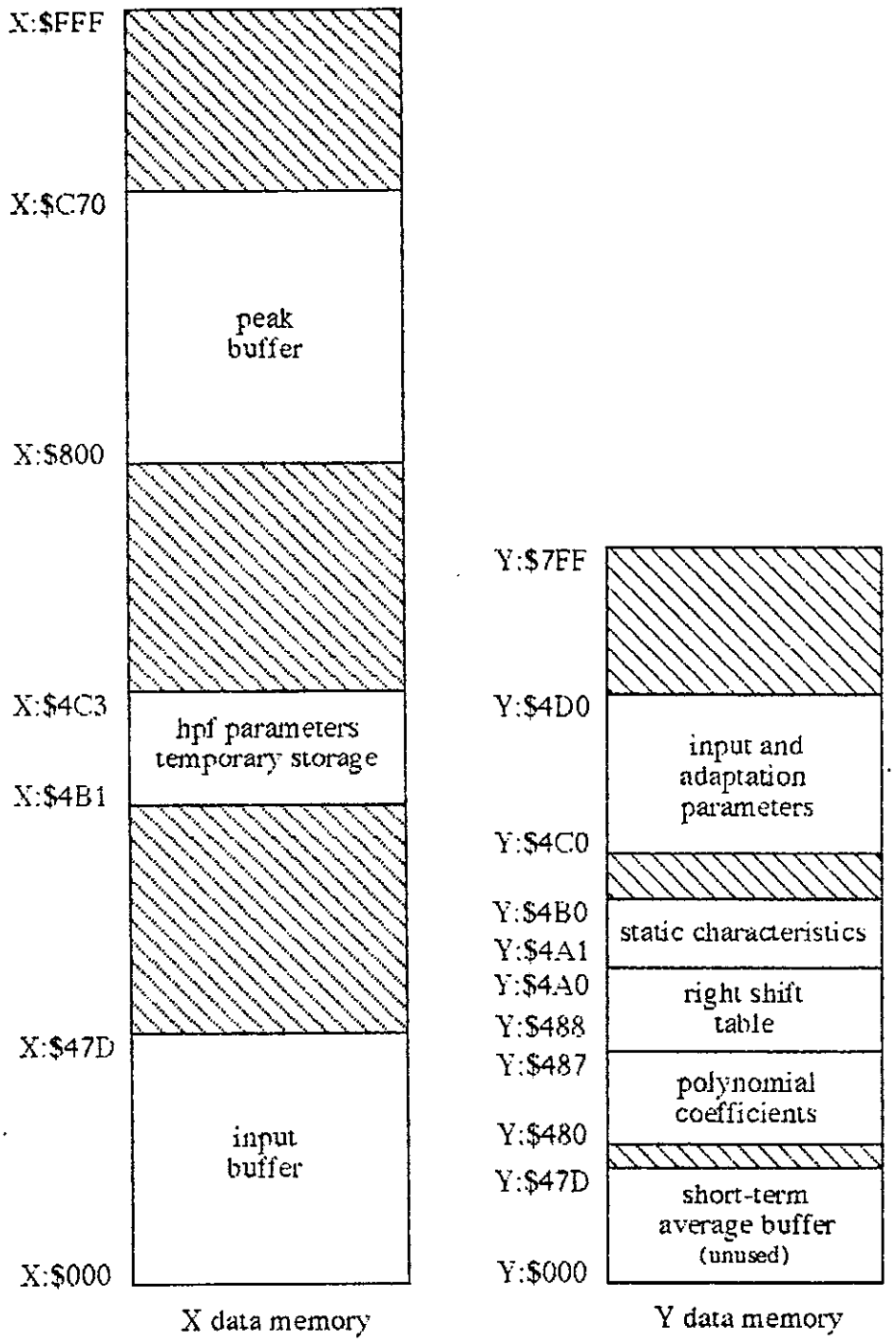
```
I_IRQB          equ     $0010           ;_IRQB

I_SSIRD         equ     $000C           ;SSI Rx Data
I_SSIRDE        equ     $000E           ;SSI Rx Data with exception
I_SSITD         equ     $0010           ;SSI Tx Data
I_SSITDE        equ     $0012           ;SSI Tx Data with exception
```

Appendix B

Memory Map




X:$FFF
X:$C70
peak buffer
X:$800
X:$4C3
hpf parameters temporary storage
X:$4B1
X:$47D
input buffer
X:$000
X data memory
Y:$7FF
Y:$4D0
input and adaptation parameters
Y:$4C0
Y:$4B0
static characteristics
Y:$4A1
Y:$4A0
right shift table
Y:$488
Y:$487
polynomial coefficients
Y:$480
Y:$47D
short-term average buffer (unused)
Y:$000
Y data memory

Appendix C

Level Adaptation Scaling Factor

In this appendix, the restrictions on the locations of the peak control threshold (V_{pc}) and the average control threshold (V_{ac}) are developed. The maximum value of the slope of the linear combination portion of the characteristic depends on the seperation of the two thresholds. The scaling factor required for these calculations must also be found. It is desirable that this scaling factor be as small as possible.

Maximize

$$\left|\frac{1}{\Delta}\right| = \left|\frac{1}{V'_{ac} - V'_{pc}}\right|$$

with respect to V'_{ac} and V'_{pc}. This can be accomplished by minimizing

$$\begin{aligned} |\Delta| &= \left|V'_{ac} - V'_{pc}\right| \\ &= \left|10^{V_{ac}} - 10^{V_{pc}}\right|^{1/20} \end{aligned}$$

Let $V_{ac} = x$ and $V_{pc} = y$. We must always have $y > x$ and $x, y > -100dB$ (i.e. the minimum input level). This can be expressed as

$$\begin{aligned} -100 < \quad & y \quad < 0 \\ -100 < \quad & x \quad < y - \epsilon \end{aligned}$$

where $\epsilon = y - x > 0$. Now we may write

$$|\Delta| = 10^{y/10}\left|10^{-\epsilon/20} - 1\right|$$

Now $|\Delta|$ may be minimized by minimizing $10^{y/20}$ and $\left|10^{-\epsilon/20} - 1\right|$. $10^{y/20}$ is a minimum for y = -100.

$$10^{-100/20} = 10^{-5}$$

$\left|10^{-\epsilon/20} - 1\right|$ is a minimum for the smallest value of ϵ. Notice that $\epsilon = 0 \Rightarrow$ minimum value is 0. After much trial and error, we set min(y) = -50dB and min(ϵ) = 10dB. This gives

$$\begin{aligned} max(|\Delta|) &= \frac{1}{|min(\Delta)|} \\ &= 462.5 \end{aligned}$$

This maximum Δ must be scaled down by 2^9 so that it is fractional. The calculations will be accurate to 23 - 9 = 14 bits (only positive twos complement values are used for the thresholds). This implies that the lower limit of the average control threshold is

$$20\log_{10}(2^{-14}) \simeq -84dB.$$

In summary, the restrictions on the thresholds are

$$V_{pc} > -50dB$$

$$\epsilon > 10dB$$

$$V_{ac} > -84dB$$

$$V_{ac} = V_{pc} - \epsilon$$

Appendix D

Polynomial Approximation Coefficients

The coefficients for polynomial approximations over the BFP input range $0.5 \leq x \leq 1.0$ are tabulated below. The coefficients are all tabulated as

$$f(x) = \sum_{i=0}^{i=N} a_i x^i$$

where N is the order of the approximation.

	$\log_2(x)$				
Order	a_0	a_1	a_2	a_3	a_4
2	-2.66403001	4.01872957	-1.35900725	—	—
3	-3.14494063	6.05895642	-4.15703264	1.24356684	—
4	-3.50561466	8.09923378	-8.39760912	5.08408893	-1.28017403

Table D.1: BFP Chebyshev Polynomial Coefficients

	2^x				
Order	a_0	a_1	a_2	a_3	a_4
2	0.99870893	0.66909732	0.17157244	—	—
3	0.99994464	0.69134017	0.23088672	0.03954285	—
4	0.99999809	0.69305067	0.23943921	0.05322683	0.00684199

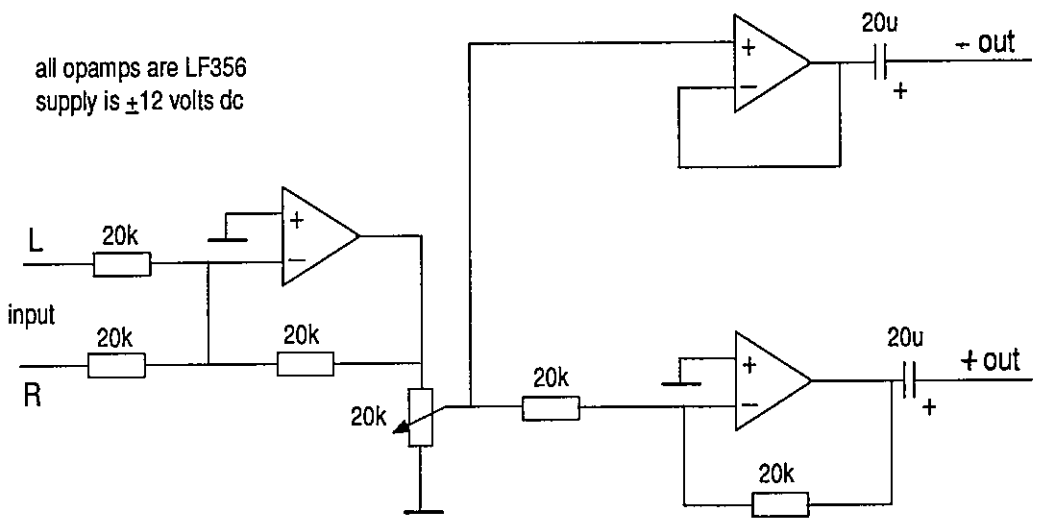
Table D.2: BFP Chebyshev Polynomial Coefficients

$\log_2(x)$ for Split Ranges				
Range	a_0	a_1	a_2	a_3
$0.5 \leq x < 0.71875$	-3.383159930	7.199881431	-5.955824132	2.177818986
$0.71875 \leq x < 1.0$	-2.883159931	5.093505086	-2.983556684	0.7732426917

Table D.3: BFP Chebyshev Polynomial Coefficients (two polynomials)

Appendix E

Mixer Schematic



Appendix F

Listening Test Results

This appendix contains the raw listening test results. In the table below, each listener is identified by his/her initials. The tests are indicated by a number/letter combination where the letter indicates the musical signal used and the number indicates the processing. The numbers and letters for each signal and processing option are listed in Section 5.3.

Test Number	Subject							Trial Averages
	JVH	PC	PEH	ACD	RLB	JPH	AV	
1A	5	4	3	4	4	4.5	3	3.93
1B	4	3	5	5	3	4	4	4.00
1C	4	4	5	3	4	4.5	4	4.07
2A	4	4	4	4	4	4.5	3	3.93
2B	3	3	4	5	3.5	4	3	3.64
2C	3	4	3	4	3.5	4	3	3.50
3A	5	4	4	4	4	4.5	4	4.21
3B	3	3	5	4	4	4	4	3.86
3C	4	2	3	4	4	4	3	3.43
4A	5	4	5	4	4	4.5	3	4.21
4B	5	4	5	5	5	5	5	4.86
4C	4	4	4	4	4	4.5	4	4.07
5A	4	3	4	3	3.5	4.5	3	3.57
5B	3	2	5	3	3	3.5	4	3.36
5C	4	2	3	4	4	4	4	3.57
6A	4	3	4	4	4	4.5	3	3.79
6B	4	3	4	5	3	4	5	4.00
6C	3	4	4	3	4	4.5	3	3.64
7A	5	5	5	5	5	5	5	5.00
7B	5	5	4	5	5	5	5	4.86
7C	5	5	5	5	5	5	5	5.00

Table F.1: Listening Test Results