A Direct PWM Loudspeaker Feedback System

by

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Submitted to the Department of Electrical Engineering and Computer Science

in Partial Fulfillment of the Requirements for the Degree of

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Abstract

A loudspeaker feedback system is built in this project. A class D switching amplifier and the loudspeaker are placed in a closed loop. The switching power amplifier is a simple pulse width modulator, but without the conventional output filter that is made up of expensive inductors and capacitors. The band-pass loudspeaker replaces the output filter. Feedback is achieved by an electret condenser microphone and a lag-pole compensator in the low frequency range of the audio spectrum. For a typical closed box type loudspeaker, the closed loop system frequency response is flat from 35Hz to 300Hz. The total harmonic distortion is also minimized to acceptable levels in the same frequency range. Various loudspeakers, drivers, and compensators are also studied for the completion of the final system.

Thesis Supervisor: James K. Roberge
Title: Professor of Electrical Engineering

This thesis would not have been possible without the help of many people. The names that follow are not necessarily listed in the order of importance.

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Finally I would like to dedicate this thesis to Dr. B. D. Wedlock and my parents, from whom I received financial support, to the Chiu family and to all my dear friends in America and abroad.

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Chapter 1

Introduction

1.1 Objective

Feedback is used extensively in audio electronics to reduce harmonic distortion and improve frequency response. However it has been little used in the loudspeaker. The loudspeaker is an important but weak part of the audio system. Its poor performance includes poor frequency response and large harmonic distortion. Its electromagnetic-mechano-acoustic nature makes it hard to utilize feedback compensation. Nevertheless, work has been done to include the loudspeaker in a feedback loop. Satisfactory results have been obtained for frequencies up to 500Hz [1] [2]. The reported results indicate a nearly flat frequency response and an average of about 5dB harmonic distortion reduction from about 30Hz to 150Hz [3]. However, such results come only by employing expensive and complicated sensors and compensators. There is a need to develop a simple low-cost feedback network that gives similar results to those of the previous complex feedback networks.

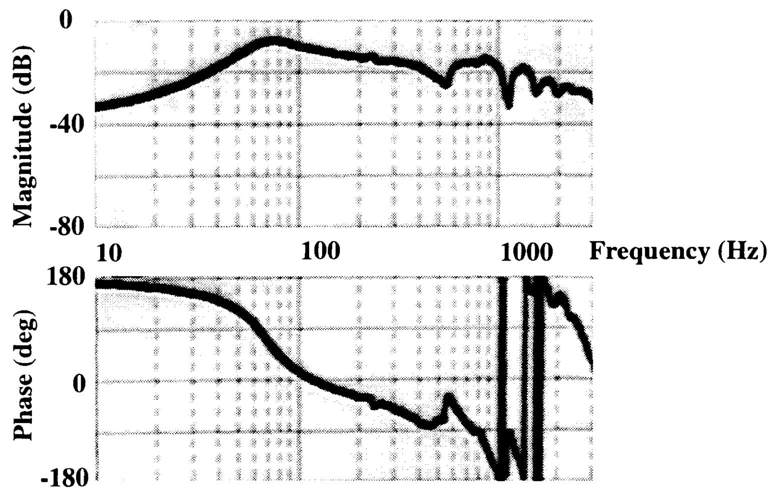
1.2 Historical Background

Feedback enables the modification of poor systems with uncertain characteristics to better systems with well-defined characteristics. As a result, harmonic distortion can be minimized while a flat frequency can be obtained. While most audio equipment takes advantage of feedback, the loudspeaker remains open-loop for the most part.

The loudspeaker is an electromagnetic-mechano-acoustic transducer. It takes in an electrical signal and converts it to sound by means of a moving voice coil and cone. The process involves electromagnetic-mechanical and mechano-acoustic conversions.

Because the system is not purely electronic, extracting the feedback signal can be challenging. Therefore the loudspeaker has been left open-loop.

Figure 1.1: Frequency Response of a Closed Box Loudspeaker



Note: The frequency response is measured using a dynamic signal analyzer (HP 3562A). The input to the amplifier-loudspeaker system is a sine sweep generated by the same instrument. The amplifier is a low distortion linear amplifier (LM12) (see Appendix B). The loudspeaker is a 10inch "woofer" (RS 40-1014A) mounted in a 50L closed box (see Appendix C). The output of the loudspeaker is sensed by an electret condenser microphone (RS 270-090B). The SPL is measured by a sound level meter (RS 33-2050). The measured SPL at 60Hz is 110dB at 1inch from the center of the cone.[1]

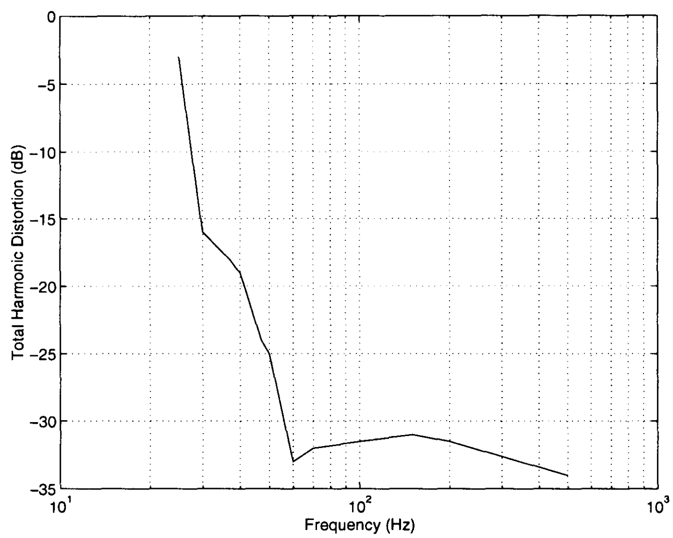
Figure 1.1 shows a measured frequency response of a loudspeaker used in this project. It is actually the open loop characteristics of a linear amplifier-loudspeaker-microphone system. Assuming that the amplifier and microphone contribute no magnitude deviation or phase shift to the Bode plots, Figure 1.1 is an accurate picture of the loudspeaker's frequency response alone. The peaks and valleys of the magnitude plot indicate that the loudspeaker has a very poor magnitude response. The magnitudes of the first peak and valley are nearly 20dB apart.

Using a distortion analyzer, we can also measure the total harmonic distortion (THD) of the loudspeaker. The results are plotted in Figure 1.2. Here we see that the harmonic distortion increases quickly as the frequency decreases near the lower end of the audio

1. The SPL reference is 0.0002μbar.

spectrum. The distortion can reach as high as -16dB, or 15.8% (at 30Hz).[1] [2] This distortion is significant because most of other audio equipment such as an average linear power amplifier introduces smaller than 1% harmonic distortion throughout the audio spectrum (20Hz-20kHz).

Figure 1.2: Total Harmonic Distortion of a Loudspeaker



Note: The distortion is measured by a distortion analyzer (HP 334A). The input to the amplifier-loudspeaker system is a sine wave from a signal generator (KH 2000). The amplifier is a low distortion linear amplifier (LM12) (see appendix B). The loudspeaker is a 10inch "woofer" (RS 40-1026A) mounted in a 50L closed box (see appendix C). The output of the loudspeaker is sensed by an electret condenser microphone (RS 270-090B). Constant SPL is obtained by manually adjusting the input to the system so that the RMS value of the microphone output signal is constant (at 300mVpp).[3] Using a sound level meter (RS 33-2050), the measured SPL at 60Hz is 110dB at 1inch from the center of the cone.

---

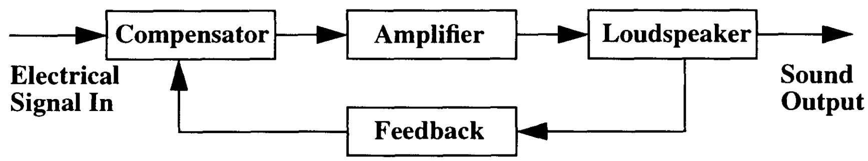
1. The distortion at 25Hz can be ignored since at that frequency the movement of the voice coil exceeds the maximum excursion limits of the loudspeaker. However the distortion at frequencies greater than 30Hz is valid.

2. Total harmonic distortion is the RMS ratio of the amplitudes of all harmonics (up to 3MHz, not including the fundamental) to those of the fundamental and harmonics. It is shown here in relative dB scale. The distortion analyzer (HP 334A) is an average-responding voltage-measuring device. Thus the relationship between dB and % is: -20dB = $20\log_{10}(10\%)$.

3. [Vpp] stands for "peak-peak Volt."

Work has been done in the past 40 years to improve the loudspeaker's performance utilizing feedback. The block diagram of a general loudspeaker feedback system is below.

Figure 1.3: Block Diagram of a General Loudspeaker Feedback System.



There are mainly two forms of loudspeaker feedback: "partial" and "motional." Partial feedback uses a nonintrusive method to extract the feedback signals from the loudspeaker's input terminals only. The signals can be extracted using a bridge circuit, a simple current sensing resistor, or other means [5] [6] [7] [8]. The signals can be the back electromotive force (EMF), load voltage and load current. It is believed that these signals are related to the sound output of the loudspeaker in a specific way. This type of feedback is usually inadequate since the actual output of the loudspeaker is not used. There is also disagreement about the linearity of the systems used to extract the signals [1]. For these reasons, partial feedback was not used in this project.

Another type of loudspeaker feedback is motional feedback. The term "motional feedback" is used to indicate any loudspeaker feedback network that includes in its feedback loop the motion of the speaker cone -- its displacement, velocity and/or acceleration [1] [2] [3]. It is believed that the cone motion, whether measured or derived, is related to the sound we hear in some known manner. Therefore the cone motion is a good representation of the sound -- the final output of the system.

To apply feedback, the feedback signal is combined with the input electrical signal to generate the error signal that is largely amplified as the new input of the loudspeaker. Feedback theory indicates that distortion can be reduced while a flat overall frequency response can be maintained.

1.3 Methods

Motional feedback has been used successfully in a variety of loudspeaker compensation systems. Although they fail at high frequencies for reasons we will see in Section 2.3, the compensation techniques are claimed to be applicable to all audio frequencies. Nevertheless their implementations are costly because the sensors and the compensation networks are either too delicate or too complex. This project examines a simple feedback method using a lag-pole network as the compensator and an electret condenser microphone as the sensor.

The microphone has been ruled out in most loudspeaker feedback systems. The main reason is the non-minimum phase shift associated with the time delay between the generation of the sound at the cone and the output of the converted signal from the microphone. The phase shift is caused by the propagation delays in the air medium and in the microphone. Although they avoided the use of the microphone and instead explored other sensors and feedback methods, researchers found that actual loudspeaker feedback can be achieved for frequencies up to only about 500Hz. At this frequency, the microphone's non-minimum phase problem is negligible. In fact, the phase problem is negligible for frequencies up to the kilohertz region, as we will see in Section 5.1. Since in this project we are concerned with low frequency (up to 300Hz) feedback compensation, we can safely use a microphone for feedback. In particular we will use an economical electret condenser microphone, which has a nearly flat frequency response and virtually no phase shift over the described frequency range.

Lastly the feedback system in this project features a "PWM driver" that resembles a class D switching amplifier. Previously many loudspeaker feedback systems used high cost linear amplifiers in the forward path. Sometimes the amplifier introduces unwanted phase shift that can make compensation difficult. For this project we will explore the use of a class D switching amplifier in place of a linear amplifier. A switching amplifier is low cost, small, highly efficient and provides high output power. However, most commercial class D switching amplifiers treat the loudspeaker as an external load of a fixed impedance (e.g. 8Ω). Thus the inclusion of complicated filters that are made up of high quality inductors and capacitors seems inevitable. In these switching amplifiers, feedback is usually applied to the output filter stage to regulate the load current or voltage delivered to the loudspeaker. This would be fine if the loudspeaker load were purely resistive. However as we will see in the next chapter, the loudspeaker is a complex device with an impedance considerably different than just the resistance of the voice coil. Furthermore the inductors and capacitors used in the filters are not just expensive, they are also problematic when the whole switching amplifier is used in the loudspeaker feedback system. Like the linear amplifier, the output filter of the switching amplifier introduces unwanted phase shift, which again complicates compensation. By realizing that the loudspeaker is a band-pass filter (see Section 2.2), we can eliminate the switching amplifier's output filter altogether and apply the switches (PWM driver) directly to the loudspeaker. The loudspeaker effectively becomes the output filter. Finally, simple feedback can be applied around the PWM driver and the loudspeaker to control the precise switching action so that the sound output is regulated.

1.4 Outline

The outline of this thesis is as follows: First, we will develop a simplified model of the loudspeaker based on its components and analogous electrical symbols. Then, we will look at different topologies of switching power converters and amplifiers and design the PWM drivers used for this project. Next we will discuss and also implement the loop compensation techniques based on the simplified loudspeaker model and the PWM drivers. We will then build and test the actual loudspeaker feedback system using an electret condenser microphone and a modified version of the compensator developed for the loudspeaker model. Finally, we will end with a discussion section and conclusion.

Chapter 2

Loudspeaker Model

In order to compensate the closed loop loudspeaker system, we must first understand the loudspeaker's open loop characteristics. In this section we will develop a relatively simple electrical model for the loudspeaker using the mobility analogy. We will also address the limitations and applications of this model in loudspeaker feedback compensation.

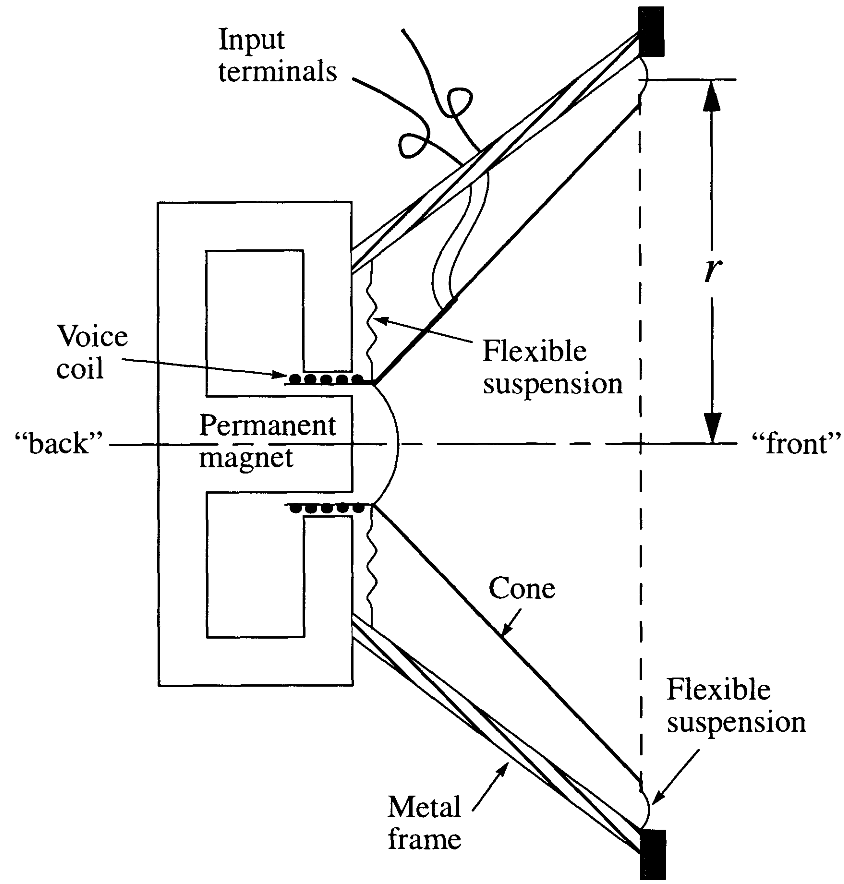
2.1 Elements of the Loudspeaker

The loudspeaker is a complicated device. It is an electromagnetic-mechano-acoustic transducer that converts electrical signals into sound. The type of loudspeaker that we are concerned with in this project is the common paper or polypropylene cone loudspeaker. Figure 2.1 shows a cross section of this type of loudspeaker. For reference, we shall call the side of the speaker where the cone opens up "the front," and the other side "the back." In such a loudspeaker, a strong magnet is mounted on the back on an open metal frame shaped much like a cone with most of the side panel cut away. On the rim of the metal frame hangs the speaker cone, or diaphragm. The materials used to suspend the cone around the rim and around the center are flexible so that the cone can move in only the axial direction: in and out (with respect to the front and back of the speaker). The center of the cone is connected to the voice coil that is in the magnetic gap of the permanent magnet. When a current is fed through the voice coil, the magnetic interaction in the air gap generates a force on the voice coil, as given by the electromagnetic property $F = Bli$.[1] As a result, the voice coil and the cone move away from the equilibrium point. The cone

1. Appendix A lists variables used in this thesis.

movement results in a change of air density and pressure locally, and sound waves propagate directly from the cone.

Figure 2.1: Cross Sectional View of a Loudspeaker



The front and back of the speaker can be acoustically shielded from each other if the speaker is mounted in a closed box or an infinitely large baffle. Caution should be taken to minimize the air leak between the two sides so that changes in air pressure on the two sides do not interfere. If the sound waves from two sides of the cone are permitted to interact with each other (such as in free air), the sound level would drop due to cancellation from the two sides. The cancellation is especially pronounced at low frequencies, where the waves are omnidirectional. The effect is analogous to trying to make a sound with only one hand moving. For practical reasons, a closed box instead of an infinite baffle is used to shield off the effect from the back of the loudspeaker. The box cannot be made too small because the enclosed air acts like a spring whose force increases as the enclosed volume decreases (see Appendix D). This "air suspension" changes the speaker system characteristics and should be accounted for in our loudspeaker model. We will only consider a closed box loudspeaker design in this project, as any other designs complicate the model by adding more acoustical elements to the system. Such complication is undesirable because it makes the closed loop compensation difficult and consequently degrade the overall system performance. We hope that the box's effect would not be too big and that the closed loop will be able to compensate for it.

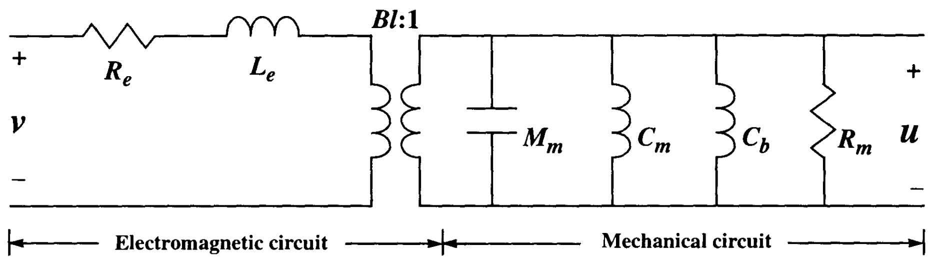
2.2 Loudspeaker Model

We have just seen the parts that make up the loudspeaker -- a coupler between these worlds: electrical, magnetic, mechanical, and acoustic. We would like to model the system elements by electrical components so that further analysis can be applied. The analogy we will use is the so called mobility analogy.

In the mobility analogy, variables such as current i, force F, and pressure p are considered "through" variables, while voltage v, velocity u and volume velocity v_v are called "across" variables [9]. The volume velocity is a property of the acoustic volume. It is a measure of how fast the volume changes and therefore the intensity of the sound at the output of the speaker. Volume velocity is analogous to velocity, which is a measure of how fast the displacement changes. Volume velocity is directional and has units [m^3/sec]. For the loudspeaker, the volume velocity is simply the product of the cone's velocity and area.

Using the mobility analogy, mechanical and acoustical resistances, masses and compliances (due to suspension) can be modeled by electrical components. Mechanical and acoustic resistances resemble electrical resistors. The moving mass of the cone or that of the air in front of the cone is equivalent to a capacitor, while the compliance of the cone suspension or that of the air is analogous to an inductor. With the above analogy, we are ready to construct the loudspeaker model. One valid model is shown in Figure 2.2. The equations for the variables are given in Appendix D [9] [10].

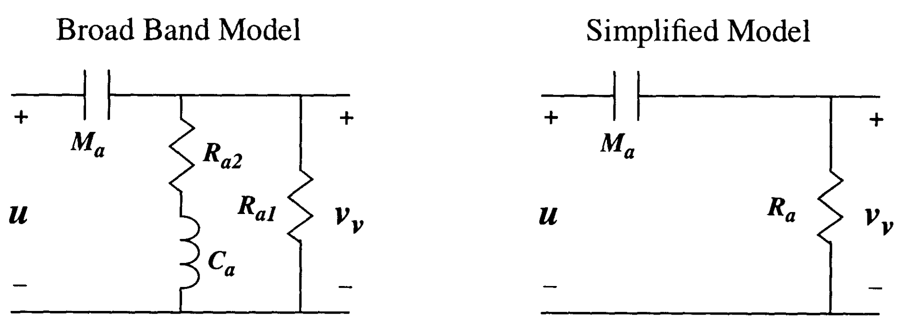
Figure 2.2: A Circuit Model for the Closed Box Loudspeaker (unloaded)



In this model, we have included the effect of the closed box, an extra compliance C_b, contributed by the spring-like air suspension inside the box. The reason that all the mechanical elements are in parallel is intuitive: the velocity is ideally identical over the entire cone structure. However, this assumption is oversimplified as we will see later in the discussion of the model's limitations. Nevertheless for low frequencies, the above model is adequate.

Next we will include the acoustic load. Baranek gives two models, one the broadband model and the other the simplified model for low frequencies [9].

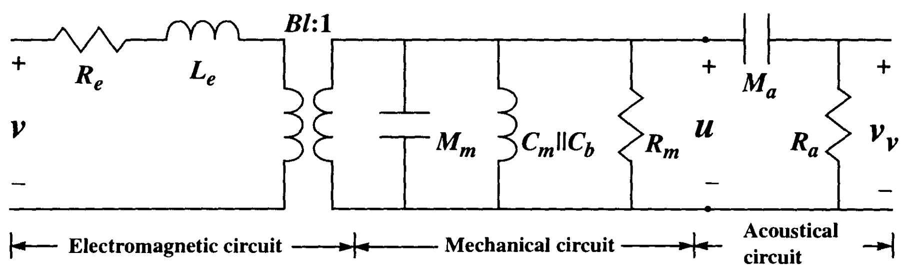
Figure 2.3: Circuit Models of the Acoustic Load



Though the simplified model is said to be accurate only up to $f < \frac{c}{4\pi r} = 275Hz$ for a 10inch ($r = 0.1$m) loudspeaker, in reality the error it introduces at even higher frequencies is small compared to that introduced by cone resonance, noise, and standing waves. For this reason, we can use it in most cases (unless, of course, the other errors are comparably small).

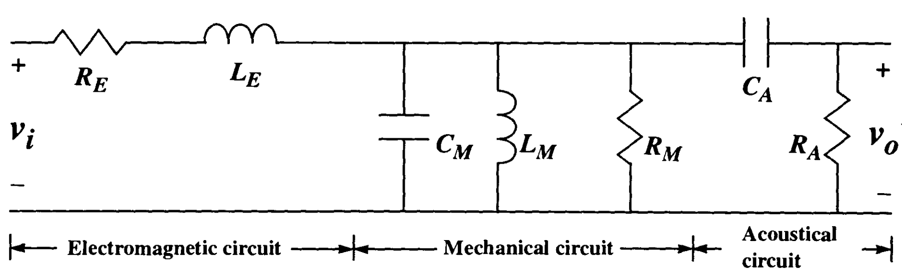
Thus we have the complete Loudspeaker model:

Figure 2.4: A Closed Box Loudspeaker Circuit Model with Acoustic Load



We now take out the transformer and let $v_i = v$, $v_o = Blv_v$, $R_E = R_e$, $L_E = L_e$, $C_M = \frac{M_m}{(Bl)^2}$, $L_M = (C_m \| C_b)(Bl)^2$, $R_M = R_m(Bl)^2$, $C_A = \frac{M_a}{(Bl)^2}$, $R_A = R_a(Bl)^2$. Then we have the following simplified model:

Figure 2.5: A Simplified Loudspeaker Circuit Model with Acoustic Load



For the 12inch loudspeaker used in this project, the corresponding values in the above circuit model are listed in Table 2.1.

From the above model, we wish to derive the system function, which will be useful for feedback compensation. From the system function, we can also extract poles and zeros of the system. Nyquist and Bode analyses can then be applied. Piece-wise estimation techniques have been used in determining the frequency response of the system [9] [10]. However, such techniques are not adequate for finding the estimated locations of the system poles and zeros. Though involved, finding the exact solution to the system is a start.

Table 2.1: Loudspeaker Model Parameters

<table>
<tr><td>$R_e = 5.6$</td><td>$R_E = 5.6$</td></tr>
<tr><td>$L_e = 4.5 \times 10^{-4}$</td><td>$L_E = 4.5 \times 10^{-4}$</td></tr>
<tr><td>$M_m = 0.0546$</td><td>$C_M = 1.23 \times 10^{-3}$</td></tr>
<tr><td>$C_m = 7.66 \times 10^{-4}$</td><td rowspan="2">$L_M = 4.77 \times 10^{-3}$</td></tr>
<tr><td>$C_b = 1.25 \times 10^{-4}$</td></tr>
<tr><td>R_m = large: 4.51</td><td>R_M = large: 200</td></tr>
<tr><td>$M_a = 6.87 \times 10^{-3}$</td><td>$C_A = 1.55 \times 10^{-4}$</td></tr>
<tr><td>$R_a = 0.0321$</td><td>$R_A = 1.42$</td></tr>
</table>

Note: The loudspeaker is a 12inch loudspeaker (RS 40-1026A). Values in the second column are derived from those in the first column (see Appendices A, C and D). MKS units are omitted. R_m is calculated from a nominal value of R_M, 200Ω [9].

Using Maple, we can find the exact system function for the simplified circuit model in Figure 2.5:[1]

$$H(s) = \frac{V_o}{V_i}(s) = \frac{Ks^2}{s^4 + a_1 s^3 + a_2 s^2 + a_3 s + a_4}, \text{ where} \tag{2.1}$$

$$K = \frac{1}{C_M L_E}, \tag{2.2}$$

$$a_1 = \frac{R_E}{L_E} + \frac{1}{R_A C_M} + \frac{1}{R_A C_A} + \frac{1}{R_M C_M}, \tag{2.3}$$

1. Maple is a registered trademark of Waterloo Maple Software.

$$a_2 = \frac{R_E}{L_E R_A C_M} + \frac{R_E}{L_E R_A C_A} + \frac{1}{L_M C_M} + \frac{1}{L_E C_M} + \frac{1}{R_A C_A R_M C_M} + \frac{R_E}{L_E R_M C_M}, \tag{2.4}$$

$$a_3 = \frac{1}{L_E R_A C_A C_M} + \frac{1}{R_A C_A L_M C_M} + \frac{R_E}{L_E R_A C_A R_M C_M} + \frac{1}{L_M C_M}, \tag{2.5}$$

$$a_4 = \frac{R_E}{L_E R_A C_A L_M C_M}. \tag{2.6}$$

Though it is precise, the above system function is of little use to us. One can simplify the system function further by eliminating/adding terms that are negligible for typical parameter values. For example, R_M is generally big compared to the other components that are in parallel. Therefore R_M can be eliminated. After deleting and adding negligible terms, the new system function becomes:

$$H(s) = \frac{V_o}{V_i}(s) \approx \frac{Ks^2}{s^4 + b_1 s^3 + b_2 s^2 + b_3 s + b_4}\text{, where} \tag{2.7}$$

$$K = \frac{1}{C_M L_E}, \tag{2.8}$$

$$b_1 = \frac{R_E}{L_E} + \frac{1}{R_A C_M} + \frac{1}{R_A C_A} + \frac{1}{R_E (C_M + C_A)}, \tag{2.9}$$

$$b_2 = \frac{R_E}{L_E R_A}\left(\frac{1}{C_M} + \frac{1}{C_A}\right) + \frac{1}{(C_M + C_A)}\left(\frac{1}{L_E} + \frac{1}{L_M}\right) + \frac{1}{R_A C_A R_E C_M}, \tag{2.10}$$

$$b_3 = \frac{1}{L_E R_A C_A C_M} + \frac{1}{R_A C_A L_M C_M} + \frac{R_E}{L_E L_M (C_M + C_A)}, \tag{2.11}$$

$$b_4 = \frac{R_E}{L_E R_A C_A L_M C_M}. \tag{2.12}$$

The above system function can be factored exactly to yield a simpler one:

$$H(s) = \frac{V_o}{V_i}(s) \approx \frac{Ks^2}{(s+a)\,(s+b)\left(s^2 + cs + d\right)}, \text{ where} \tag{2.13}$$

$$K = \frac{1}{C_M L_E}, \tag{2.14}$$

$$a = \frac{R_E}{L_E}, \tag{2.15}$$

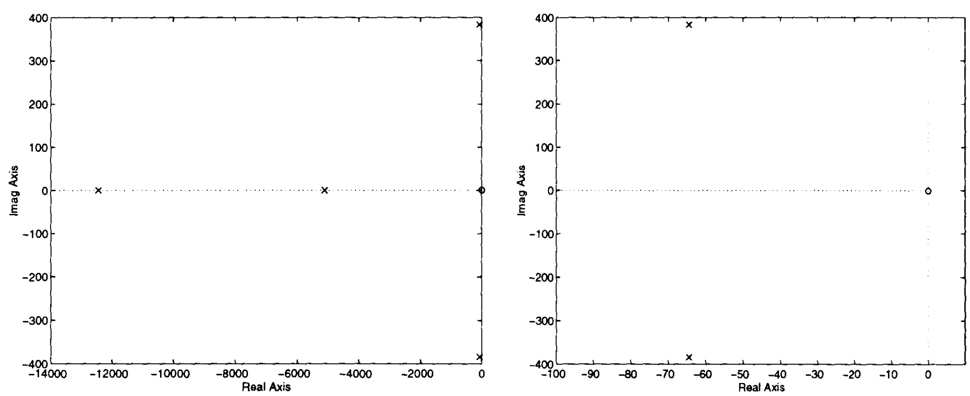
$$b = \frac{C_A + C_M}{R_A C_A C_M}, \tag{2.16}$$

$$c = \frac{1}{R_E (C_A + C_M)}, \tag{2.17}$$

$$d = \frac{1}{L_M (C_A + C_M)}. \tag{2.18}$$

The above system function represents the simplified model of the loudspeaker. It shows that the system has a double zero at the origin and 4 additional poles: two complex and two real. Figure 2.6 shows the pole-zero plot for the loudspeaker model with parameter values given in Table 2.1. We see that the system's complex poles are near the low frequency range of the audio band. Therefore the magnitude response of the system climbs steadily at a rate of 40dB/dec as the frequency increases up to the double pole (fundamental resonance) frequency. At the fundamental frequency, the magnitude usually peaks because of the closed box. However the magnitude turns nearly flat for mid frequencies, with a slight declining slope due to the effect of the two high frequency real poles. Then at high frequencies, because of the two closely located real poles, the magnitude gradually rolls off at 40dB/dec.

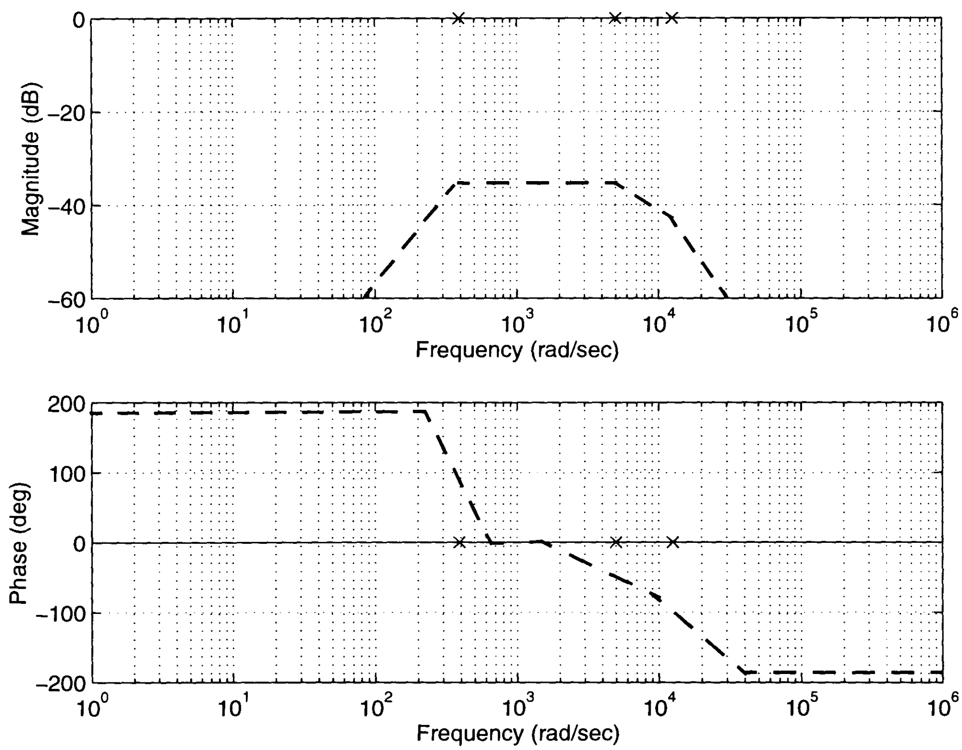
Figure 2.6: Pole-Zero Plot of the 12inch Loudspeaker Model System



Note: The plot on the left shows all system poles and zeros, particularly the high frequency real poles; the plot on the right is a magnified plot that shows the zeros at the origin and the low frequency complex poles. The numerical pole locations are: -5116, -12444, and -64±384i.

Equation 2.13 enables us to quickly plot the asymtotic Bode magnitude and phase plots once given the loudspeaker parameters (as shown in Table 2.1). Consequently closed loop compensation for the loudspeaker system using Bode analysis would be simplified and quick. Figure 2.7 shows the corresponding asymptotic Bode plots. As the figure indicates, the loudspeaker is a second order audio band pass filter. Thus the phase plot starts off at 180° and ends at -180°. It reaches 0° when the frequency is somewhere in between the fundamental resonance frequency and the two closely located real poles.

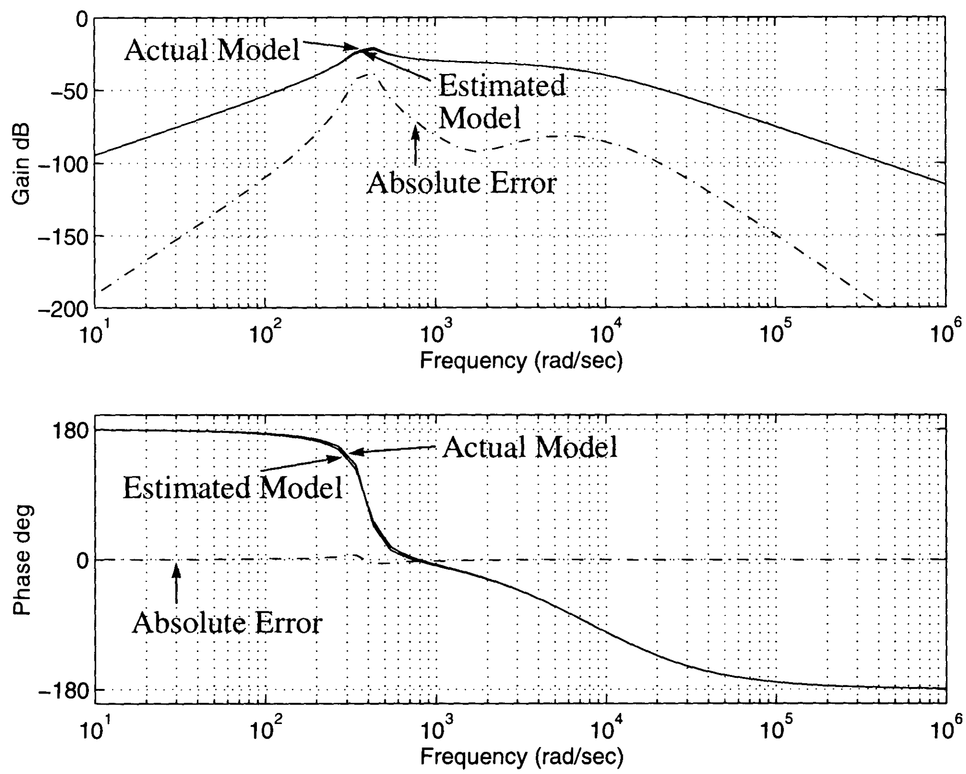
Figure 2.7: Asymptotic Bode Plots for the Simplified Loudspeaker Model



The estimated system function we have just developed is very accurate compared to the exact system function even for large variations in system parameters. We can see this from the Bode plots of the two system functions, and that of the absolute error -- the difference between the actual and estimated system functions. The plots are shown in Figure 2.8. As we can see, the two systems' plots nearly overlap. The phase error is virtually zero except for frequencies near the fundamental resonance frequency. The magnitude error function exhibits a similar behavior. It is 40dB or more below the actual or estimated magnitude plot for almost all frequencies. This means that the error is smaller than 1% of the actual system function for those frequencies. This value is negligible compared to other sources of errors. Even near the fundamental resonance frequency, the estimate is also very close to the actual model, with only about 10% error (-20dB). Furthermore, such an error is not a problem when we use feedback. The closed loop system can tolerate even larger magnitude deviation from a high resonance peak if the compensator is designed to account for sensitivity to system parameters. The system sensitivity requirement is not hard to meet, as we will see in Section 4.1.2.

Given the system parameters in Table 2.1, we can now quickly predict how the model behaves. We will then be able to compensate the system in a closed loop.

Figure 2.8: Bode Plots of the Actual and Estimated Speaker Models, and the Error



2.3 Limitations and Applications of the Model

The model presented in Section 2.2 have limitations. In reality the average loudspeaker does not behave as predicted by the above model at high frequencies. In the construction of the model, we have neglected the effects of the speaker cone resonance, standing waves, and the acoustic load at high frequencies.

Speaker cone resonance probably results in the worst kind of error. Ideally all parts of the cone move in a synchronous mode. This is true when the frequency is low in the audio band, where the corresponding wavelength is much greater (2π times) than the dimensions of the cone ($2r$): $\lambda = \frac{c}{f} > 4\pi r$. As the frequency increases, different parts of the cone start to resonate. Some parts move in while other parts move out. The patterns or modes are highly frequency dependent and unpredictable for different loudspeakers [9]. The net result is that the overall sound magnitude and phase are irregular at high frequencies (greater than 275Hz for the 10inch loudspeaker). We can see this effect in Figure 1.1. The magnitude peaks and valleys are undesirable in both open loop and closed loop systems. The additional phase shift at cone resonance frequencies also complicates the closed loop compensation. The irregular cone resonance imposes a practical limit on loudspeaker feedback systems. Given the reported results from previous work, the limit has been around 500Hz for paper cone loudspeakers [1] [2]. Unless there are mechanical changes to the loudspeaker design, the limit is hard to overcome. Such changes may include using lightweight rigid cones such as that made of aluminum, alloy or fiber glass, and even a new cone design. These topics are beyond the scope of this project and will not be dealt with here.

It appears that the model we developed is not useful because of the limitations imposed by the cone resonance and other effects. This is true for high frequencies where the corresponding wavelength is comparable to the dimensions of the cone. However the relativity of the cone dimensions and the wavelength can be changed. For example, a smaller cone, a lightweight rigid cone, or a cone with mechanically damped termination can be used [11].[1] If a loudspeaker with such a cone is used, the model would be valid up to a higher frequency limit. Though in this project the limit is around 300Hz for the loudspeakers, it can be higher for other loudspeaker systems. The model is therefore

1. If the cone is terminated (anywhere between the center and the edge inclusively) by a soft region, high frequency waves cannot reflect off the boundaries to create cone resonance. Furthermore sound waves travel faster in a rigid cone than in a soft cone. This implies that the effective wavelength is increased in a rigid cone because $\lambda = \frac{c}{f}$.

applicable to other loudspeaker systems at least up to the frequency limit given by the cone resonance and other effects.

In loudspeaker feedback compensation, using the model clearly has an advantage over directly measuring the open loop system characteristics. The model gives a quick picture of the system performance once the loudspeaker parameters are known. Since loudspeaker manufacturers usually provide specifications that contain enough related information, the parameters can be calculated easily (see Table 2.1 and Appendices C and D). This means that an experimental compensation network can be built by trial-and-error once the model is determined. No further open loop system characterization is necessary. The only caution when using the model is to make sure that the system crosses over before the frequency limit $f < \frac{c}{4\pi r}$.

Chapter 3

PWM Driver

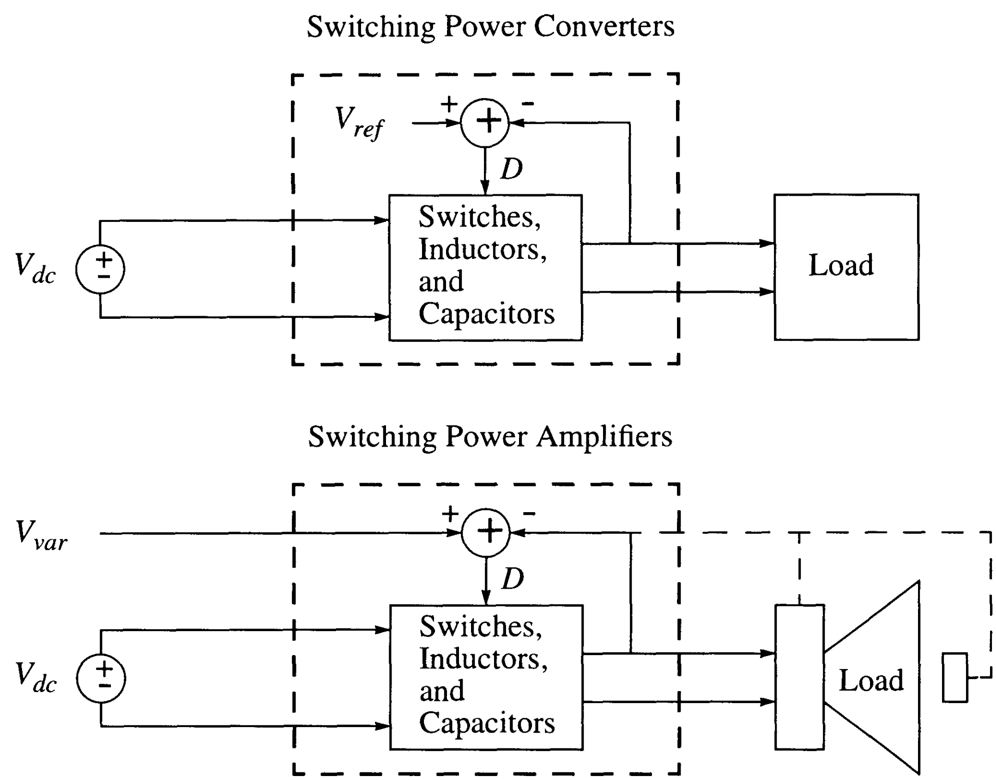
The amplifier used in this project is a class D switching type. However, it is distinct from regular switching amplifiers. This amplifier is without the output filter that is usually made up of inductors and capacitors in ordinary switching amplifiers. For brevity, we shall call it a "PWM driver." The reason for choosing a switching action driver is apparent given that the loudspeaker can be used as a filter. The idea is to exploit the advantages of a switching power converter and replace its costly output filter with the loudspeaker itself.

3.1 Switching Power Converter and Power Amplifier Topologies

Before presenting the PWM driver, let us look at the different topologies in switching power converters. With slight modifications any switching power converter can perform just like a switching amplifier. In this way, switching amplifiers evolve from switching power converters. The PWM driver used here evolves from one type of switching amplifier -- the buck (down) converter.

Figure 3.1 shows the block diagrams for switching power converters and power amplifiers. The core of the two diagrams is the same: a DC supply input, switches, inductors, capacitors and usually a feedback regulating network. In a switching power converter the output delivered to the load is internally set with or without a feedback loop. The voltage (or current) seen by the load is fixed by a reference voltage V_{ref}. If a feedback loop is used, the output is constant regardless of changes in the input supply voltage V_{dc}. The feedback loop operates in such a way that the switching duty ratio D responds to the changing V_{dc} so that the output stays at constant. Similarly in a switching power amplifier, the output is regulated if V_{var} (the reference voltage made available as in input) is fixed. Additionally, the feedback loop in a switching amplifier responds to a change in V_{var} proportionally. Furthermore, additional feedback (e.g. from the loudspeaker output) can be included in the feedback such that the final output is controlled.

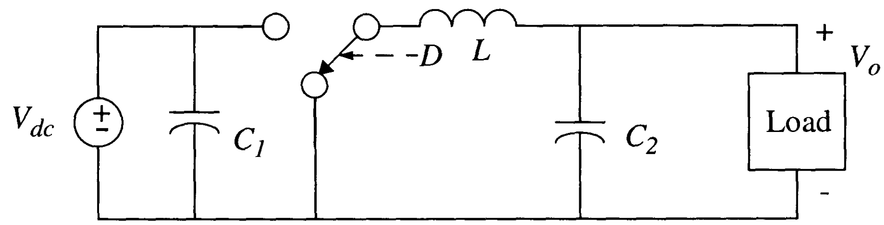
Figure 3.1: Block Diagrams of Switching Power Converters and Amplifiers



Switching power converters have been very popular in the power electronics field. Their small size, low cost, high efficiency and high power features make them the ideal choice in many DC and AC conversion applications. There are DC/DC, DC/AC, AC/DC, and AC/AC conversions. We are concerned with only DC/DC and DC/AC in this project.

DC/DC conversion can be achieved by three basic methods depending on the application: buck, boost, and buck-boost [12]. They are also called down, up, and up-down conversions respectively. In this project, we are only concerned with buck (down) converters. In a buck converter, a DC voltage input is converted to a lower DC output. Figure 3.2 shows the circuit topology of a buck converter. The inductor L, capacitor C_2 (and the load) form the output filter of the converter. The average output voltage V_o depends on the duty ratio D: $V_o = DV_{dc}$

Figure 3.2: Buck (Down) Converter

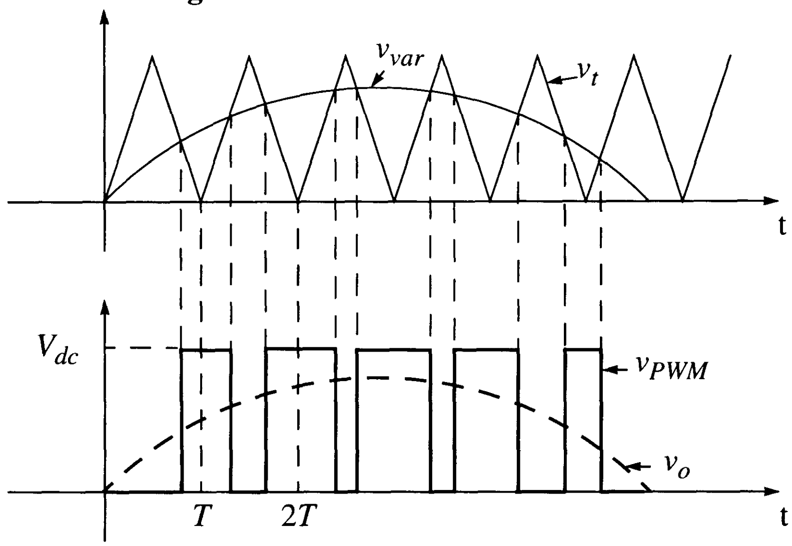


The buck converter emerges as the ideal choice for the switching amplifier in this project because the energy storage element, the inductor L, is on the output side of the switch. As we have seen in the previous chapter, there is an embedded inductor (the voice coil) in every electromagnetic loudspeaker. Using the voice coil alone as the energy storage element, we can eliminate the inductor at the output stage of the buck converter. The elimination of the inductor is not only cost efficient but also less complicated in terms of output feedback control. One fewer energy storage component (inductor or capacitor) means one fewer pole and/or zero in the system. A simpler system leads to a simpler compensation network. For the same reason, the filter capacitors (if there is any) can also be eliminated. All the filtering of switching ripple is performed by the band-pass like loudspeaker.

While DC/DC converters are made to output a fixed DC voltage, the DC/AC conversion is to output a variable voltage from a fixed DC source. DC/AC conversion is necessary in this project because the audio output is variable. One way of implementing DC/AC conversion stems from the buck converter. As we said before, one can vary the DC output of a DC/DC converter by changing V_{ref} and consequently the duty ratio D (see Figure 3.1). For most DC/DC switching power converters, D is regulated so that the output is fixed even though the input may change. The regulation is done by an internal feedback loop. In an actual buck converter the duty ratio is a function of both the reference voltage and the output. If the output drops because of a drop in the DC input or for any other reason, the duty ratio would be increased due to the negative feedback from the output. Consequently the output voltage increases to the desired fixed voltage. The increase in D counter-balances the drop in V_{dc} such that the output voltage ($V_o = DV_{dc}$) stays at a constant voltage fixed by V_{ref}.

In order to achieve DC/AC conversion, the internal reference voltage V_{ref} is replaced by an external variable signal V_{var}, so that the output voltage follows V_{var}. In other words, varying V_{var} causes V_o to change proportionally if the relationship between V_{var} and D is linear. The result is that the output is a pulse width modulated (PWM) waveform. In Figure 3.3, the PWM waveform (v_{PWM}) before the output filter, and the final output waveform (v_o) are plotted for a half-period sinusoidal V_{var} input. The switching frequency is fixed at $f_{sw} = \frac{1}{T}$ (100kHz<f_{sw}<500kHz), which is well above the audio frequency range. From the figure, we see that V_{var} changes the duty cycle (pulse width) in each fixed period as a result of changing D. Higher V_{var} contributes to a wider pulse, which averages to a higher output voltage after the output filter. Because $0<D<1$, $0<V_o<V_{dc}$, the DC/AC power converter becomes a switching power amplifier whose output voltage swing is $0<V_o<V_{dc}$.

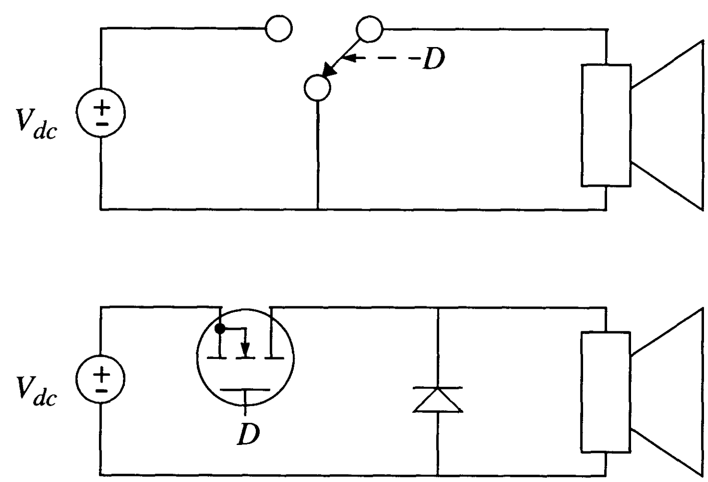
Figure 3.3: Pulse Width Modulation



3.2 Circuit Implementations of the PWM Driver

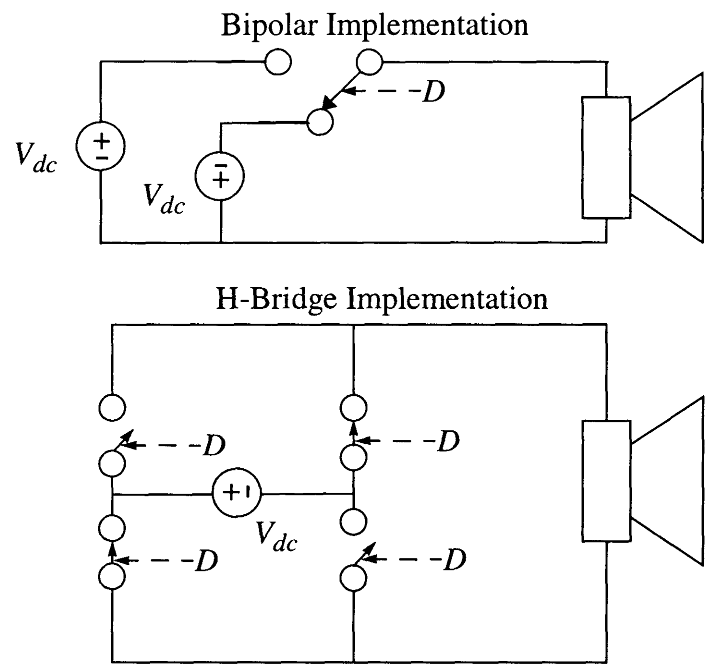
Now we proceed to construct the PWM driver. If we use the buck converter as shown by simply replacing the output filter with the loudspeaker, it is not possible to obtain negative output. Figure 3.4 shows the circuit topology and its practical circuit implementation. From section 3.1, we have seen that the output voltage is between the ground and V_{dc} (see Figure 3.3). Therefore, no negative voltage can be delivered to the loudspeaker, limiting the movements of the cone.

Figure 3.4: A Simple (Buck) PWM Driver-Loudspeaker System



To solve the above problem, we can use a bipolar or a bridge design for the PWM driver. The circuit topologies are given in Figure 3.5. The first design is bipolar. The input to the loudspeaker is switched between two voltage sources: $+V_{dc}$ and $-V_{dc}$. The voltage sources and the loudspeaker share a common ground.[1] This topology has few switches. Thus the duty ratio controller is relatively simple. Its expense is the bipolar voltage sources.

Figure 3.5: Bipolar and H-Bridge PWM Driver-Loudspeaker Systems



The second design is an H-bridge. The input terminals of the loudspeaker are simultaneously switched by two sets of switches, 180° out of phase. Only one set of diagonal switches are on at any time. As a result, the two terminals of the single DC voltage source are connected to the loudspeaker in a rotary fashion. There is only one voltage source but twice as many switches as the bipolar design. The switch drive

1. A common ground may become useful for current and voltage sensing at the speaker's input terminals. No additional circuitry is required (except a current sensing resistor). Otherwise an extra differential amplifier stage is necessary.

therefore becomes complicated. If the switches are implemented using only N-type MOSFET's, then four different switch drives are necessary, two being the complements of the other two.

For this project, both the bipolar and bridge designs were implemented, but the bridge circuit was used in the actual loudspeaker feedback system in the later stage of the project. To implement the switch drives, the "triangle intercept" method can be used [12]. Basically a fixed frequency triangle wave is compared to the varying input V_{var} (see Figure 3.3). The output of the comparator is a true PWM rectangular wave whose duty ratio D is linearly related to V_{var} even without the benefit of feedback. This is because the triangle wave is linear in its half period. True PWM is not necessary when negative feedback is applied, even though the fixed frequency waveform is not exactly triangular. The duty ratio D will be fixed by the feedback loop so that it gives an output that is linearly related to the input V_{var}. An exponential wave from the free running multivibrator can be used [13] [14]. However for low harmonic distortion, a true triangular wave is superior to the exponential wave. In the earlier stage of this project, an exponential wave was used. It was later replaced by a true triangular wave.

Figures 3.6 and 3.7 show the detailed schematics of the switch drives and the corresponding switch networks in the bipolar and bridge designs. One marked difference between the two schematics is that the circuitry for the bipolar design is powered by dual DC supplies while that for the bridge design is a single supply $+V_{dc}$. Another difference is the switch implementation. The switches of the bipolar design are discrete (NMOS IRFD110 and PMOS IRFD9110), whereas the ones in the bridge are monolithic (NMOS H-bridge LMD18200). Other features in the bipolar design include a speed up capacitor for the inverting npn stage, the nearly non-overlapping turn-on times for the switches, and push-pull buffers for the MOSFET gates. Dead times were not implemented for the bipolar circuit. However the faster falling edge of the comparator was exploited by the inverting npn stage. In the bridge design, a charge pump is an inherent feature for driving the high side N-type MOSFET's. For a switching frequency as high as 500kHz, the charge pump capacitors of about 0.01μF are suggested [4].

Figure 3.6: Bipolar PWM Driver Circuit

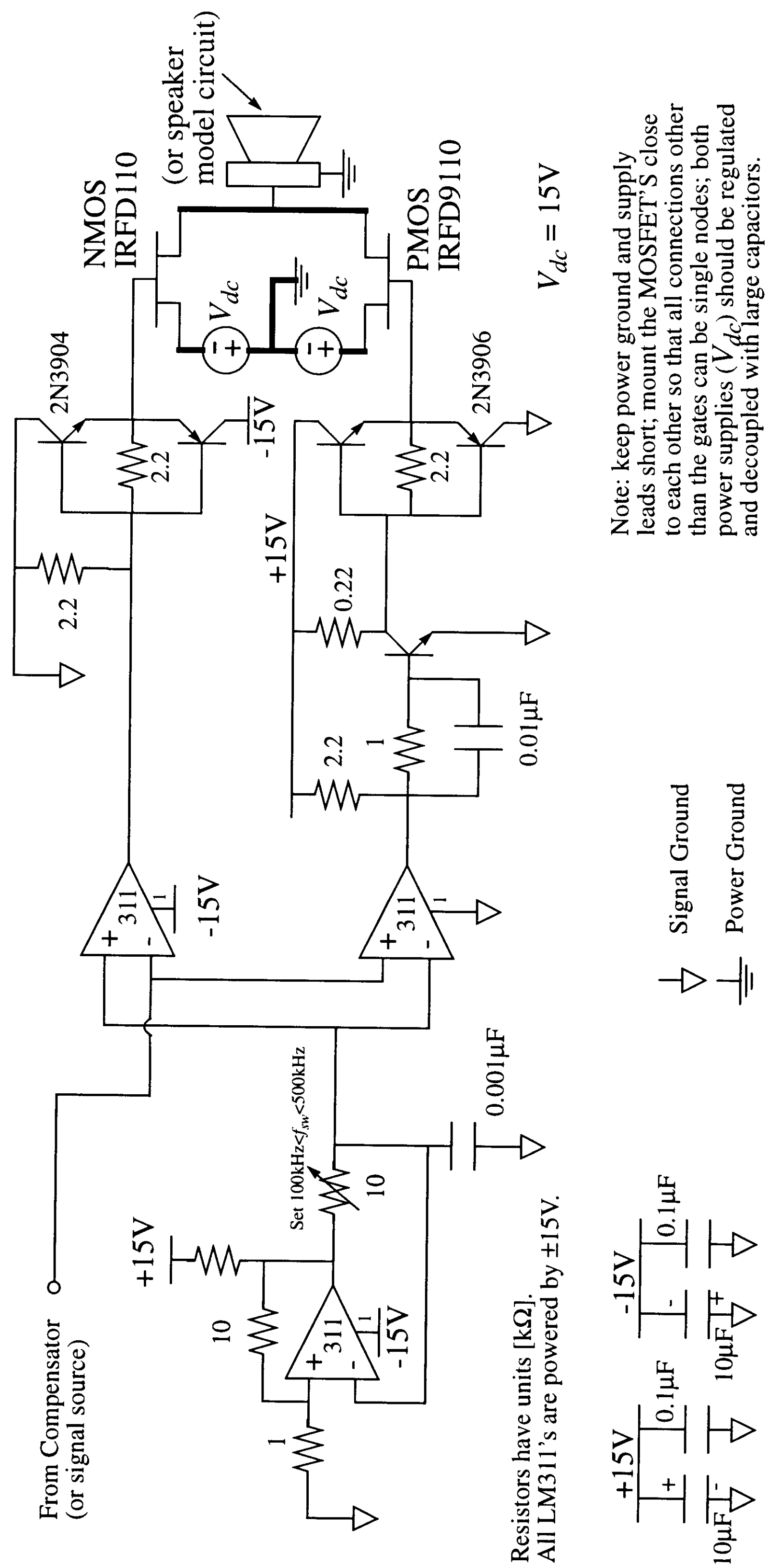
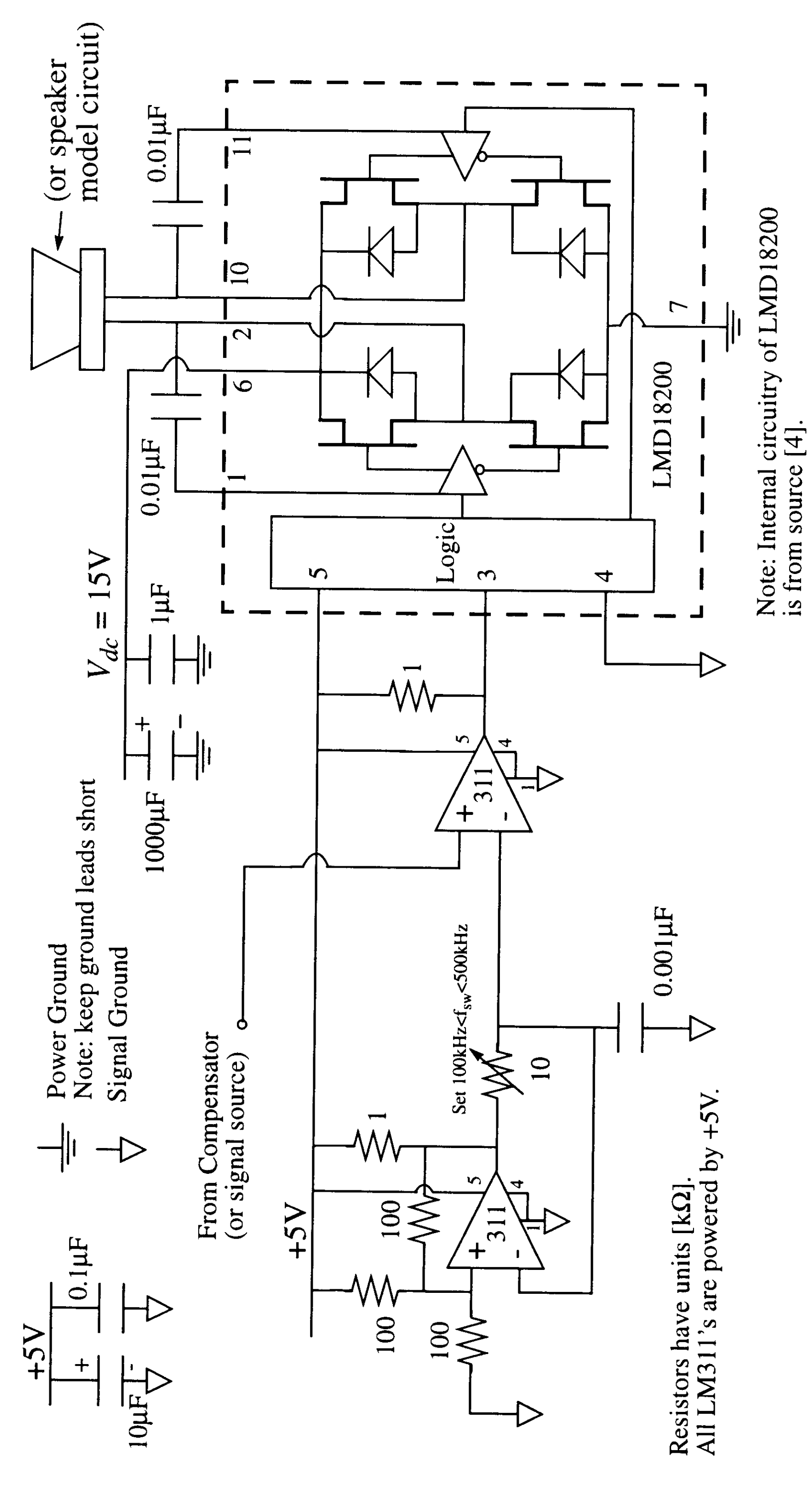


Figure 3.7: H-Bridge PWM Driver Circuit



Chapter 4

Loudspeaker Closed Loop Compensation

Now we have a fairly simple and accurate circuit model for the loudspeaker, and a working PWM driver/amplifier. Next we wish to find a compensator for the closed loop system that includes the PWM driver and the loudspeaker model. Then we will verify whether the compensator is adequate using a loudspeaker model circuit that is made up of discrete components. Though the model used here does not include the effects of cone resonance and other sources of errors we discussed in Chapter 2, the compensation of such a model has its significance in the event that a loudspeaker that fits the model (for all or part of the audio spectrum) is used. In this project, the compensation of the actual loudspeaker system near the low frequency crossover is certainly based on the compensation of the loudspeaker model, as we will see in the next chapter.

4.1 Theory of Loudspeaker Compensation

4.1.1 System Function and Compensation Network A

Figure 1.3 shows the block diagram of a general loudspeaker feedback system. The feedback signal is taken from the output of the speaker. It is then compared to the system's input. The resulted error is fed to a compensator, which basically has the effect of shaping the open loop amplifier-loudspeaker system characteristics. The compensated signal is then amplified by the amplifier. The output of the amplifier is finally fed to the loudspeaker. If the compensator is properly designed to suit the open loop amplifier-loudspeaker system characteristics, the loudspeaker output will track the system input with little distortion.

Since we know that the loudspeaker model circuitry driven by a linear amplifier is open loop stable, and that the PWM driver has no additional poles or zeros in the frequency range of interest, the PWM driver-loudspeaker system is also open loop stable and has no additional poles in the right half s-plane. Consequently, Nyquist diagrams or root loci are not necessary in determining system stability before applying Bode analysis. However, the actual loudspeaker system may be unstable due to cone resonance. Nyquist diagrams or root loci should be used to show stability first before applying Bode analysis. Furthermore when using Bode plots in closed loop compensation, the phase margin at crossover must be adequate to account for the negative non-minimum phase associated with any part of the system (e.g. the sensor's signal pick up and the amplifier's delays).

For the moment, let us assume that there is no time delay and that the loudspeaker can be modeled as shown in Figure 2.5. We also assume that the speaker's sound output signal is available to us, so that we can apply negative feedback easily. The open loop system function for the PWM driver-loudspeaker combination can be derived using incremental models:[1]

$$v_i = 2V_{dc}(d - 0.5) \text{ , where } 0<d<1; \tag{4.1}$$

$$v_i = V_i + \tilde{v}_i;\ d = D + \tilde{d};\ \text{and, } V_i = 2V_{dc}(D - 0.5)\ . \tag{4.2}$$

Substituting Equations 4.2 into Equation 4.1, we have

$$\tilde{v}_i = 2V_{dc}\tilde{d}. \tag{4.3}$$

The system function relating the duty ratio and the output of the PWM driver-loudspeaker system is therefore:

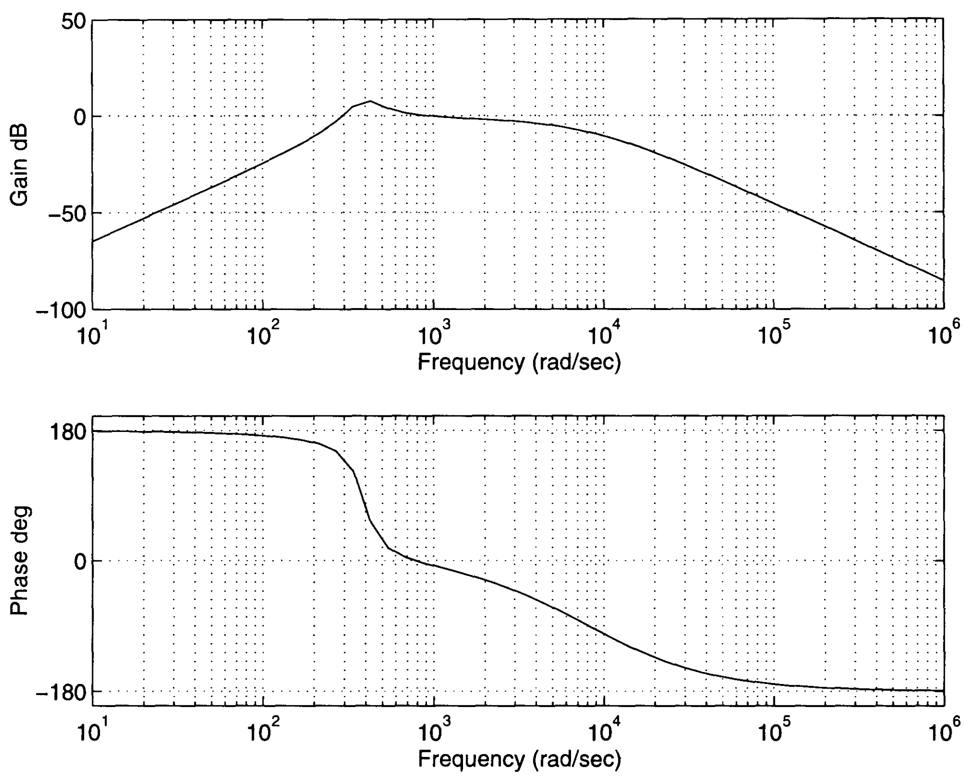
1. The notations used here are not IEEE standard. In the standard format, it is hard to denote the duty ratio incremental variable without a subscript. Therefore the tilde mark (~) is used for all incremental variables.

$$G(s) = \frac{\tilde{v}_o}{\tilde{d}} = 2V_{dc}\frac{\tilde{v}_o}{\tilde{v}_i} = 2V_{dc}H(s)\,, \qquad (4.4)$$

where H(s) is given by: (Equation 2.13)

$$H(s) = \frac{V_o}{V_i}(s) \approx \frac{Ks^2}{(s+a)\,(s+b)\left(s^2+cs+d\right)}.$$

Figure 4.1: Bode Plots of a PWM Driver-Loudspeaker System



From Equation 4.4, we see that the overall open loop system function *G(s)* is simply the loudspeaker model's open loop system function scaled by a constant, $2V_{dc}$. The Bode plots for *G(s)* are similar to that in Figure 2.8. For convenience, they are shown in Figure 4.1. If we require that the closed loop steady state error be minimized and the bandwidth be reasonably large (as close as possible to the frequency limit imposed by the cone resonance), then the compensator should include a double integrator and a lead network for extra phase at crossover. The compensator should have the form:

$$G_{cA}(s) = K_c \frac{\left(s + \frac{1}{\tau}\right)^2 \left(1 + \frac{s}{0.3\omega_c}\right)}{s^2 \left(1 + \frac{s}{3\omega_c}\right)}, \tag{4.5}$$

where ω_c is the desired crossover frequency and K_c is the compensator gain. τ is large so that at low frequencies $\frac{\left(s + \frac{1}{\tau}\right)^2}{s^2}$ becomes a double integrator but at high frequencies it is unity. The steady state error is thus:

$$e_{ss} = \lim_{s \to 0} \left(\frac{sV_i(s)}{1 + G_{cA}(s)} \right) = \frac{1}{1 + \frac{K_c K}{abd\tau^2}} \approx 0 \text{ if } \frac{K_c K}{abd\tau^2} \text{ is big.} \tag{4.6}$$

The closed loop crossover frequency is approximately ω_c. The phase margin at crossover is about 55° since the lead network adds a maximum amount of 55° to the otherwise 0° phase margin when the lead zero and pole are spaced one decade apart.

Note that the above compensator cannot achieve exactly zero steady state error for a step input because we cannot include more integrators in it. If we included another 1/s, the phase margin criterion at crossover would not be met and the system would be unstable. As a result of non-zero steady state error, the tracking error for a ramp input cannot be controlled to a bounded limit. Ideally the above compensator is very good, considering that there is almost no steady state error at low frequencies (0Hz to cone resonance frequency), where severe distortions are. In reality the above compensator is inappropriate. Here is why:

1. The average audio signal is hardly step-like. It is instead usually sinusoidal. Steady state accuracy is not really necessary.

2. Unless global DC feedback is used in the closed loop system, the op amps in the double integrator's circuit implementation eventually saturate. A non-zero offset (however small) at the input of a "perfect" integrator implementation is integrated indefinitely until the op amp saturates.
3. DC coupling is hard to implement in the loudspeaker feedback system. It requires that the sensor must be able to detect DC signal. An electret condenser microphone (or any microphone) has a slow roll off near low frequencies. Therefore a microphone cannot detect DC output. A broadband sensor, or one that has a flat frequency response up to at least the crossover frequency (i.e. a low pass characteristic) is required. Delicate and expensive motion sensors probably suffice.
4. DC coupling is also hard to implement for a single power supply.
5. Moreover, since the system's steady state error is not zero, any offset (at the input, from feedback, or in the op amps) may contribute significant DC offset at the output. A DC offset at the input to the loudspeaker is undesirable.
6. The system is susceptible to low frequency noise (1Hz - 10Hz). Any such noise picked up by the sensor will be considered as part of the input signal and amplified by the system. If a motional sensor is used to derive the feedback signal, low frequency vibrations from the environment (including the loudspeaker enclosure, the sensor's mounting structure and the floor) appear as noise.[1]

For the reasons listed above, the compensator given in Equation 4.5 is not suitable for this project. The next section presents an alternative.

<u>**4.1.2** Compensation Network B</u>

Since we are not concerned with steady state accuracy, we do not have to include the double integrator in the compensation network. This leads to the alternate Compensation Network B. For the compensation near the high frequency crossover ω_{c2}, the requirements are: [15]

1. Relative Stability: The loop transmission GcG must have enough phase margin at crossover: $Pm = 180° + \angle GcG > 45°$. This means that GcG magnitude plot should have adequate length of no more than -20dB/dec slope at or near crossover ω_{c2}.
2. Accuracy in Operating Range and System Sensitivity: In the mid frequencies, $|GcG|$ must be large in order to improve accuracy (reduce error), and to reduce sensitivity to disturbance and loudspeaker parameter variations. This requires that $|GcG|$ at some point before ω_{c2} should be steeper than at crossover.
3. Noise Rejection: The noise level at frequencies higher than ω_{c2} must be attenuated.

1. Noise applied to the sensor is different from a disturbance at the speaker cone. The latter is suppressed by feedback, while the former is amplified.

The slope of GcG should be more than -40dB/dec. This requirement can also reduce the effects of cone resonance peaks.

In addition to the above, we must make sure that there is enough phase margin at the low frequency crossover. The design of the compensation network near the low frequency crossover ω_{c1} is similar to that at the high frequency crossover ω_{c2}:

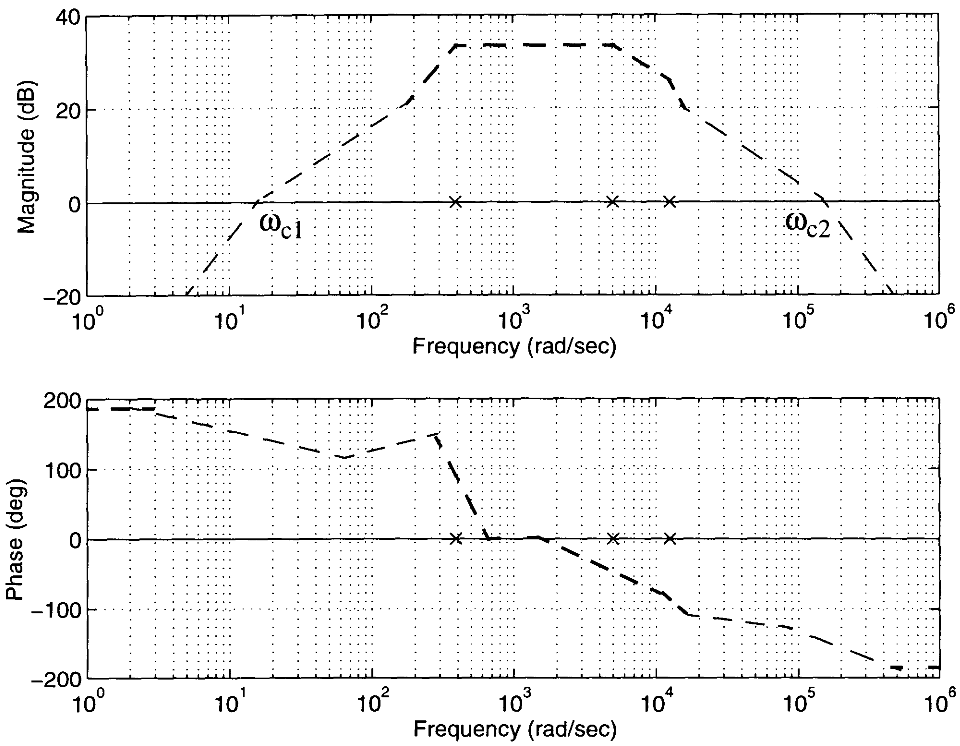
1. Relative Stability: At crossover, GcG must have enough phase margin, measured from 180° instead of from -180°: Pm = 180° - $\angle$GcG > 45°. This means that the Bode plot of GcG should have adequate length of no more than 20dB/dec slope at or near crossover ω_{c1}.
2. Accuracy in Operating Range and System Sensitivity: In the mid frequencies, |GcG| must be large to improve accuracy, and to reduce sensitivity to disturbances and loudspeaker parameter variations. This requires that the slope of |GcG| should be raised at some point after the low frequency crossover ω_{c1}.
3. Noise Rejection: The noise level must be reduced at frequencies lower than $\omega_{c1.}$ i.e. |GcG| should have a slope greater than 40dB/dec at those frequencies.

Given the above criterions, we proceed to design another compensator -- Network B. For the speaker model we have developed in Section 2.2, the driver-speaker system function is given by Equation 4.4. The corresponding Bode plots are shown in Figure 4.1. We see that the magnitude plot has a slope of +40dB/dec (or -40dB/dec) at low (or high) frequencies, while at the mid frequencies the system maintains a nearly flat magnitude response. For the requirements that we have just listed above, we need a lag compensation network near the low frequency crossover, and a lead near the high frequency crossover. The overall compensator should be of the form:

$$G_{cB}(s) = K_c \frac{(s + 10\omega_{c1})\left(\frac{s}{0.1\omega_{c2}} + 1\right)}{(s + \omega_{c1})\left(\frac{s}{\omega_{c2}} + 1\right)}. \qquad (4.7)$$

The asymtotic Bode plots of GcG is thus:

Figure 4.2: Asymptotic Bode Plots of the Compensated Loudspeaker Model System



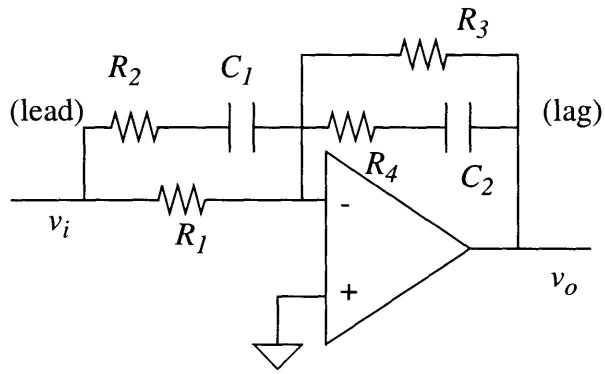
Compensation Network B solves most of the problems presented by the earlier version (Network A). For comparison, the features of Network B are listed below:

1. The system with Compensation Network B gives up steady state accuracy.
2. System B can be AC coupled. As long as signals whose frequencies are greater than the low crossover frequency can pass through the system, feedback is valid. DC offsets at the input and from the feedback are suppressed. Op amp offsets can be nulled by an offset adjust.
3. The sensor can be either broad band or band pass. Its frequency response only need to be wider than the GcG frequency range $\omega_{c1} < \omega < \omega_{c2}$.
4. System B is not susceptible to low or high frequency noise. However it is subject to mid frequency noise. If care is not taken, room noise picked up by the microphone can deteriorate the system's performance.
5. If needed to offset the unwanted phase due to time delay, cone resonance, standing waves, etc., the phase margin can be increased by crossing over early, but definitely beyond the fundamental resonance frequency.

4.2 Loudspeaker Model Closed Loop System and Its Performance

Assuming that the loudspeaker model in Equation 2.13 is valid, then the loop transmission of the PWM driver-loudspeaker system is given by Equation 4.4. A compensator of the type given in Equation 4.7 can be readily implemented for the closed loop driver-loudspeaker system. To realize the lag-lead network, we can use the following circuit:

Figure 4.3: Lag-Lead Network



The system function associated with the above circuit is:

$$\frac{V_o}{V_i}(s) = -\frac{R_3}{R_1}\frac{(R_4C_2s+1)}{[(R_3+R_4)C_2s+1]}\frac{[(R_1+R_2)C_1s+1]}{(R_2C_1s+1)}. \tag{4.8}$$

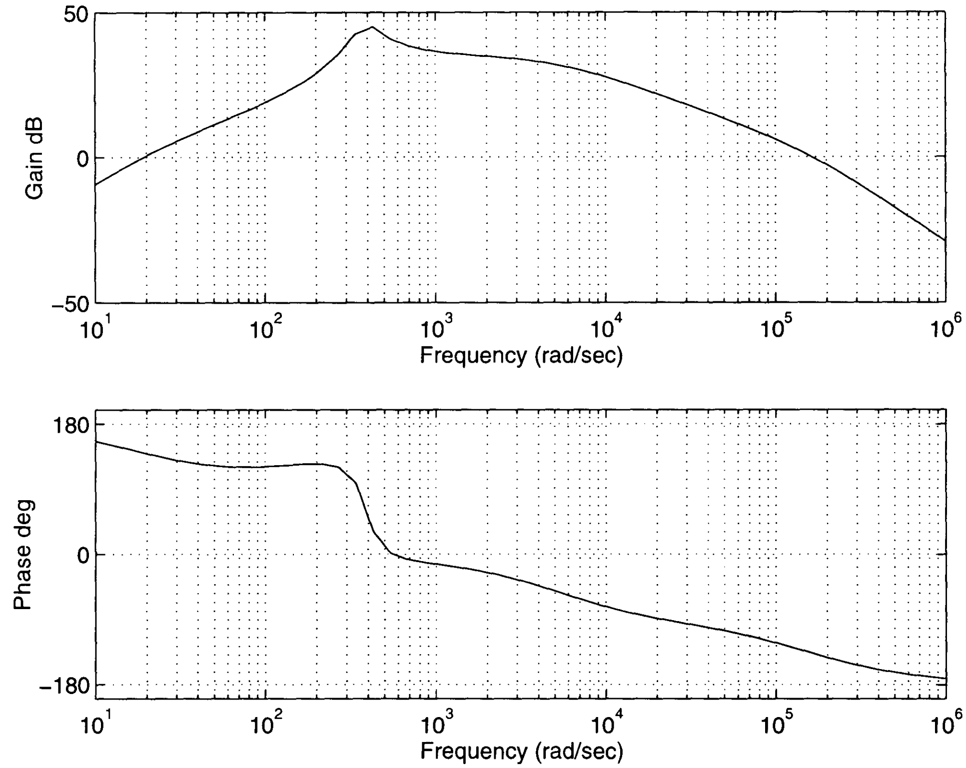
The desired compensator is:

$$G_c(s) = 2\times10^3\frac{(s+200)\left(\frac{s}{1.5\times10^4}+1\right)}{(s+20)\left(\frac{s}{1.5\times10^5}+1\right)}. \tag{4.9}$$

According to Matlab Bode plots,[1] the above compensator sets the system crossover frequencies to be ω_{c1}=20rad/sec and ω_{c2}=160krad/sec. Figure 4.4 shows the plots. Also evident from the Bode plots is that the phase margins at both crossovers are close to 45°.

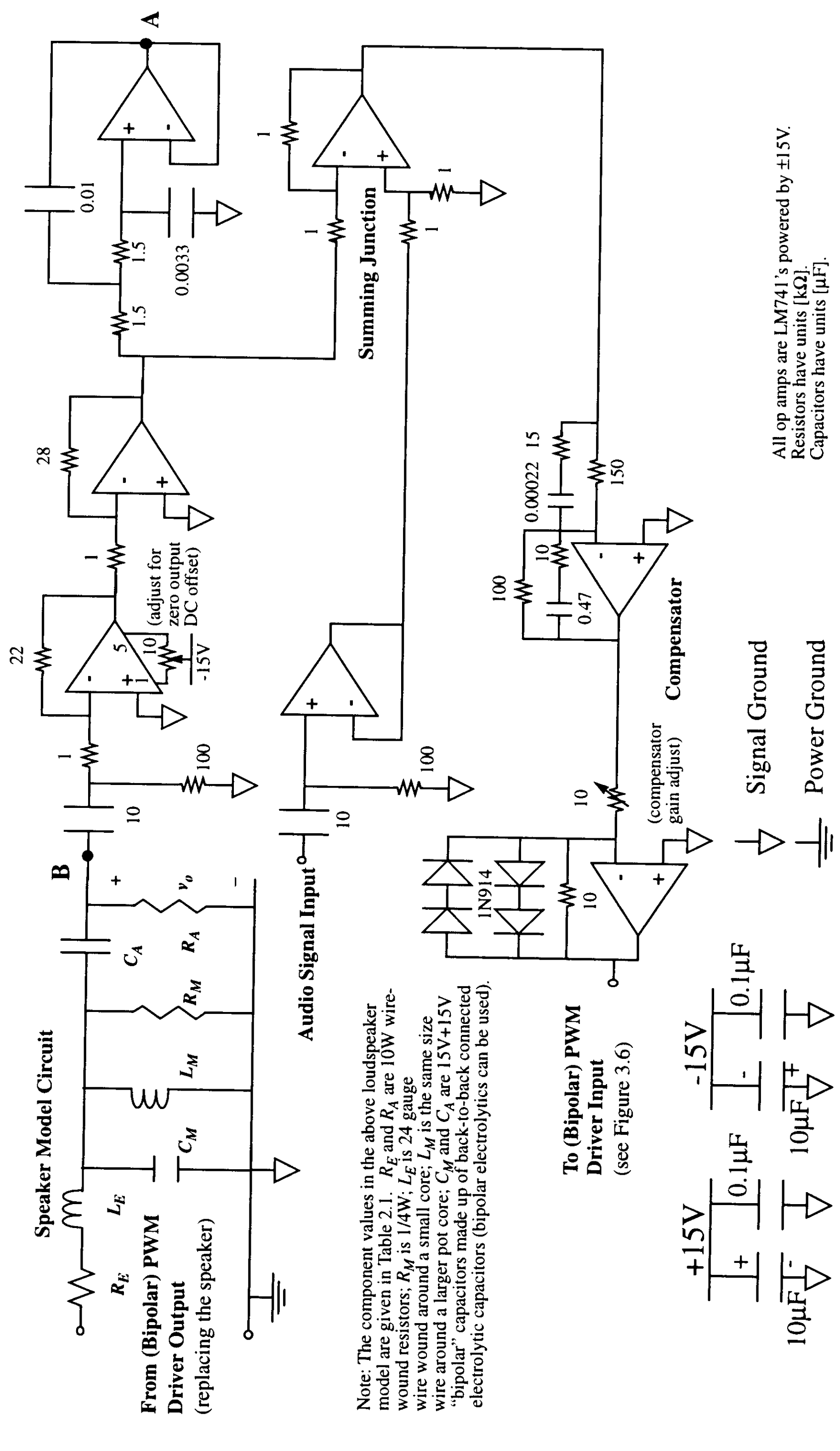
1. Matlab is a registered trademark of MathWorks, Inc.

Figure 4.4: Bode Plots of the Lag-Lead Compensated Loudspeaker Model System



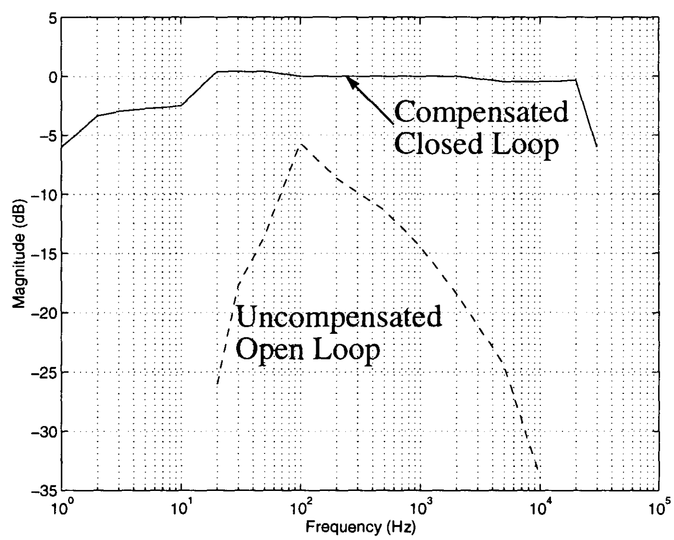
The corresponding values for the resistors and capacitors that make up the poles and zeros of the compensator in Equation 4.9 are shown in the circuit diagram of Figure 4.5. Loop gain is achieved by a couple of amplifier stages in the feedback path. More compensator gain is achieved by a variable gain inverter prior to the PWM driver. The "duty ratio limiting" diodes in the variable gain inverter are necessary to prevent latch up conditions. Upon start-up or if the input is out of the specified range ($-1.2V < v_i < 1.2V$), the PWM driver may latch in a state and the feedback loop malfunctions. Lastly both the input and feedback signal are AC coupled through an appropriate RC filter. Signals whose frequencies are as low as 0.2Hz can pass through. The loading of the RC network is negligible since $100k\Omega >> R_A$ ($<5\Omega$).

Figure 4.5: A Loudspeaker Model Closed Loop System



The compensator shown in Figure 4.5 has been built and tested in conjunction with the bipolar PWM driver shown in Figure 3.6. In Figure 4.6, the frequency response of the lag-lead compensated system is compared to that of the uncompensated open loop loudspeaker model system. The closed loop system output is taken from Point A, or through a separate amplifier from Point B for independent measurements. The two measuring schemes yield almost identical results. In the case of Point B, the output signal is small and should be amplified for maximum scope sensitivity. Furthermore, it contains high frequency noise at the switching frequency (100kHz). It should be filtered by a second order low pass filter to show meaningful results.[1]

Figure 4.6: Magnitude Plots of the Compensated and Uncompensated Loudspeaker Systems



Note: The compensated closed loop system is given in Figure 4.5. The uncompensated open loop system is the same circuit minus the summing junction and the compensator. Thus in the open loop system, the buffered input signal is directly connected to the input of the bipolar PWM driver (see Figure 3.6). The output of the open loop system is from point B. The input to the systems is a 100mVpp sinewave from a signal generator (KH 2000). The output is measured manually with an oscilloscope (Tektronix 2445).

1. The high frequency noise is beyond the audio range. It is merely the interference from the PWM driver, not part of the actual output of the loudspeaker or its model circuit. We shall disregard this noise.

As shown in Figure 4.6, the frequency response of the compensated system is flat from 20Hz to 20kHz. The bandwidth is from 3Hz to 25kHz. The results are almost as expected (see Figure 4.4).

Because a speaker model circuit is used instead of the actual loudspeaker, the systems have no harmonic or nonlinear distortion other than that introduced by the switching amplifier (because of the exponential wave and the timing of the switches). Therefore distortion analysis is not necessary. The distortion from the switching amplifier was later minimized in the final loudspeaker system using a true triangle wave and an improved switch drive (see Figure 5.11).

Chapter 5

A Closed Loop Loudspeaker System

Finally we are ready to implement the actual loudspeaker feedback system. In this chapter, the lag-lead compensator given in Chapter 4 will be modified to suit two box-type loudspeakers. An electret condenser microphone will be used as the sensor. The results are presented at the end of this chapter.

5.1 Measurement of Open Loop Driver-Loudspeaker Characteristics

The lag-lead compensator presented in Chapter 4 has been shown to give satisfactory results for the loudspeaker model given in Chapter 2. However, a loudspeaker that can be represented by the model over its entire operating range does not presently exist. Because of the limitations that we have discussed in Chapter 2, the loudspeaker model and its circuit equivalent are only valid for a narrow audio range: $f < \frac{c}{4\pi r}$, or up to about 275Hz for a 10inch speaker in a closed box. The cone resonance and other problems in all paper or polypropylene cone speakers are hard to solved by compensation, except the possibility of avoiding the problems altogether. This implies that the speaker system should cross over before the frequency limit, or 275Hz for the 10inch loudspeaker.

Previous work on closed loop loudspeaker systems has been claimed to be successful for frequencies up to around 500Hz. The main reason could well be the cone resonance problem. Some of these systems use complicated feedback systems including voltage/current and motional feedback techniques. Some also use complicated and expensive sensors such as a bridge, a capacitive distance meter, an accelerometer and a moving coil to sense the motion of the cone. The performance of these systems still falls under the cone resonance frequency. Since steady state accuracy is not of concern for audio, a microphone can be used as the sensor. In the following closed loop systems, we will use an economical electret condenser microphone (RS 270-090B) as the sensor.

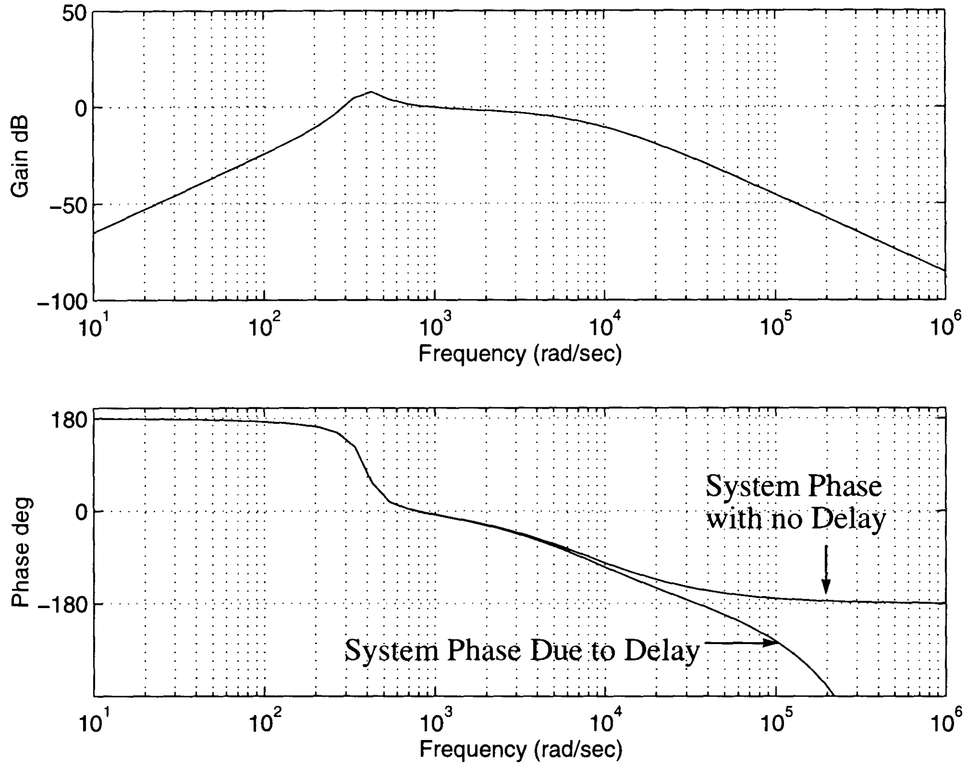
Before we measure the driver-loudspeaker system's open loop characteristics, we shall first validate the use of an electret condenser microphone. Such a microphone, according to the literature, can easily be designed to have a flat frequency response over a wide range (30Hz - 10kHz) [16]. It has virtually no phase shift until very near the upper limit of its operating range. Even at that limit, the phase shift is between 20° and 50° [16]. Since we are only concerned with low frequencies, we can use an electret condenser microphone for feedback.

Despite of its attractive features, the use of a microphone as a sensor has been avoided in most loudspeaker feedback systems. The main reason, as researchers have pointed out, is that the time delay associated with sound waves traveling from the cone to the microphone is non-minimum phase. The delay contributes no change in magnitude but significant phase shift at high frequencies:

$$\angle\left(e^{-j\omega T_d}\right) = -\omega T_d = -2\pi f\frac{\Delta}{c}, \tag{5.1}$$

where T_d is the propagation delay due to the sound traveling the distance Δ from the cone to the microphone, at the speed of c = 345m/sec.

Figure 5.1: Simulated Non-minimum Phase Effect Due to Propagation Delay



If we put the microphone close to the cone, say about 0.5cm apart, then the phase shift is minimal for low frequencies. Such a near field measurement of the sound field is valid in this project.[1] Using Equation 5.1 and Equation 4.4, the Matlab simulated phase plot of the open loop loudspeaker model and microphone system is given in Figure 5.1. Though it does not include the effect of the PWM driver's phase shift, the plot is accurate in that the PWM driver contributes little negative phase shift. From the plot, we see that the time delay contributes little negative phase below 10krad/sec. This suggests that as improved loudspeakers become available, a microphone can still be used as the sensor.

Most electret condenser microphones are also omnidirectional, meaning that sound waves can incident at any angle with little variation in the output. The omni-directivity

1. Far field measurement is subject to the unavoidable room acoustics just like in any system. We shall not use far field measurement as a performance indicator of our closed loop system.

feature can be beneficial or harmful. In terms of picking up the feedback signal, it is beneficial because the microphone is not sensitive to the sound's incidence angle. However the omni-directivity can be harmful because the microphone can pick up surrounding noise. As we have mentioned in Section 4.1.2, noise can deteriorate the system's closed loop performance. Fortunately the room noise problem can be solved by placing the microphone inside the closed box, right behind the cone. Since the back of the cone moves with the front of the cone exactly 180° out of phase for all frequencies, the feedback signal from the back microphone must be inverted before it is coupled to the negative feedback path. For this project, the microphone was placed in front of the cone for ease of installation. All measurements were done in a quiet environment.

The microphone used in this project is a low cost electret condenser microphone (Radio Shack No. 270-090B). This particular microphone has a built-in FET amplifier. The given frequency response is from 30Hz to 3kHz [17]. Because of the poor frequency response of this particular microphone, we will use a matching microphone in the measurements to offset its effect in the low frequency range. The sensitivity of the microphone is -63dB±3dB at 1V/μbar and a bias voltage of 4.5V. For a 115dB (at 0.0002μbar, 1cm distance) sound input, the equivalent output voltage is thus:

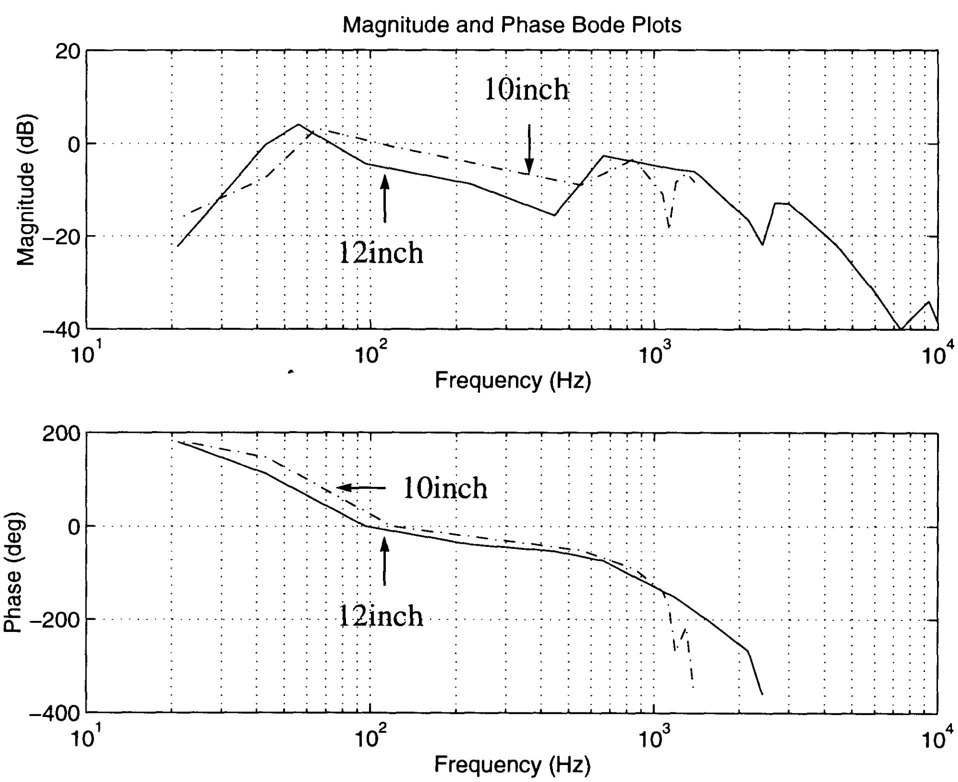
$$10^{\frac{20\log\left(10^{\frac{115}{20}}\times 0.0002\right)-63}{20}} = 0.08V. \quad (5.2)$$

This value is large enough for coupling to the feedback path without amplification.

Using the above microphone and a 100mVpp sinusoidal input, the open loop PWM driver-loudspeaker magnitude and phase characteristics are measured. The data essentially form the Bode plots of the loop transmission in the closed loop that includes the PWM driver, boxed loudspeaker and the microphone. The plots are shown in Figure

5.2. Two data sets have been collected: one for a 10inch speaker (RS 40-1014A) and the other for a 12inch speaker (RS 40-1026A). The speakers are mounted in $0.05m^3$ (50L) closed box enclosures. It is evident from the figures that both loudspeakers exhibit cone resonance. At frequencies lower than the fundamental resonance (the first peak due to the double complex poles) of the loudspeaker, the slope of the magnitude response is close to 40dB/dec and the phase is 180°. Then after the fundamental resonance peak, the magnitude slowly decreases due to the influence of its high frequency real poles. The phase plot crosses over 0° at some point after the fundamental resonance, but before cone resonance start to dominate at around 500Hz for both speakers. The phase quickly drops beyond -180° for frequencies higher than 1kHz.

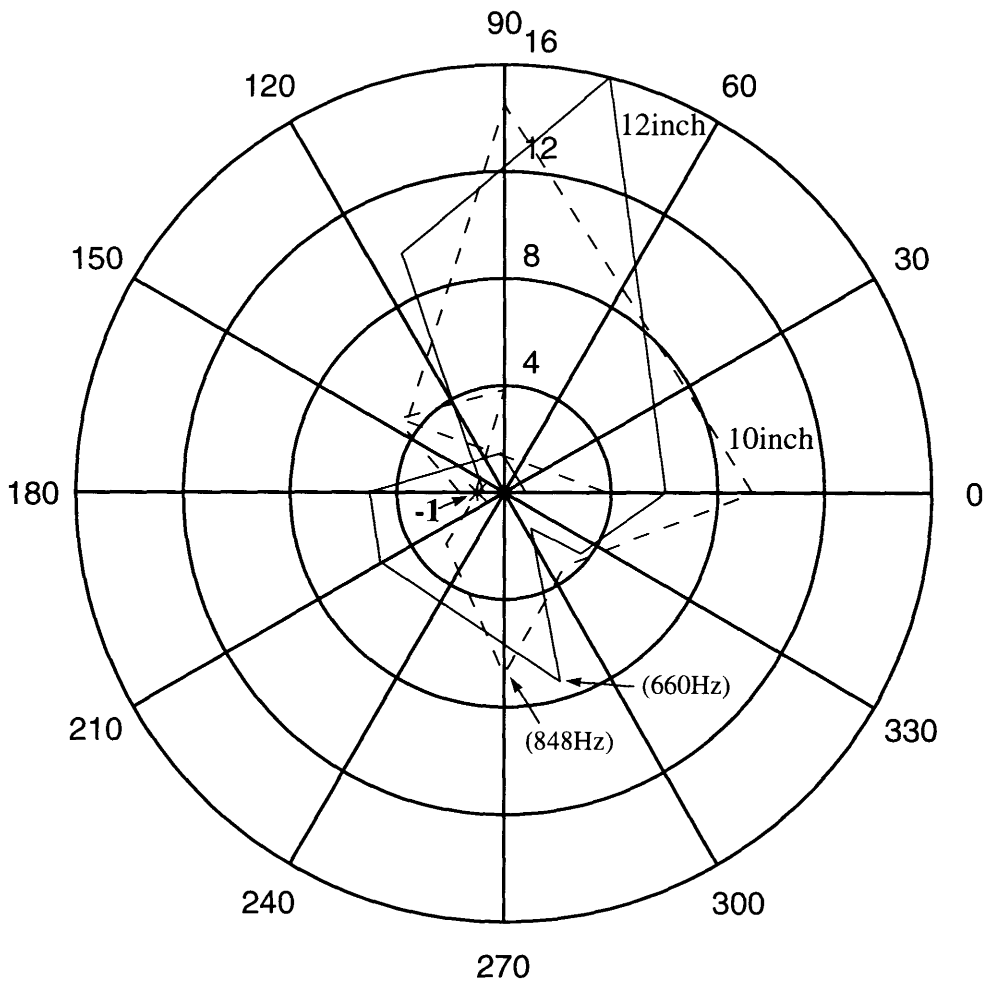
Figure 5.2: Measured Bode Plots of the Driver-Loudspeaker-Microphone Systems



Note: The PWM driver used is shown in Figure 3.7. Inputs to the systems are 100mVpp sine waves from a signal generator (KH 2000). The inputs are buffered and have a DC offset of 2.5V, the quiescent input voltage of the PWM driver. The loudspeakers (12inch RS 40-1026A and 10inch RS 40-1014A) are mounted in 50L closed boxes (see appendix C). The output is sensed by an electret condenser microphone (RS 270-090B), as shown in Figure 5.4.

If we use a simple gain of 10 compensator for the system hoping to improve frequency response over a wide operating range, we will see that the system is unstable as is evident from the Bode plots. The phase margin requirement cannot be met at the crossover frequencies if we simply raise the magnitude plot by 20dB. Alternatively, we can replot the data points on a magnitude-phase (Nyquist) plane and scale the magnitude by 10 times. From the Nyquist plots in Figure 5.3, we see that the systems are unstable because they enclose the -1 point.

Figure 5.3: Nyquist Plots of Simple Gain Compensator-Loudspeaker Systems



Note: The Nyquist plots are created by plotting the phase and scaled (10 times) magnitude data from Figure 5.2.

Since we know that the cone resonance hurts the phase margin significantly, we will compensate the system so that it crosses over before the cone resonance peaks. Furthermore, we must suppress the cone resonance peaks so that they are not above unity where the phase is close to -180°. Accomplishing this is equivalent to bending the Nyquist plot by reducing magnitude to avoid enclosing the -1 point, and making sure that the plot does not come back out to enclose the -1 point at higher frequencies.

To compensate for the low frequency crossover, a lag network (as a part of the lag-lead network used in Chapter 4) would suffice. It is repeated here for convenience:

$$G_{lag}(s) = K_1 \frac{(s + 10\omega_{c1})}{(s + \omega_{c1})}. \tag{5.3}$$

On the other hand, to compensate for the crossover before the cone resonance peaks, a lead network is inappropriate. Fortunately two alternatives of essentially the same principle are available:

1. Add one or two poles just before the first "valley" prior to the cone resonance, hoping to suppress the resonant peaks without contributing much phase shift to the crossover. i.e. ω_c must be before these poles.
2. Add one pole prior to crossover so that the gain at low frequencies can be raised in order to increase accuracy. The pole contributes a maximum -90° phase but can reduce the resonance peaks significantly. Then increase the system gain until the phase margin is just adequate or until the cone resonance peaks are just adequately below unity, whichever comes first. The crossover frequency ω_c must be after the pole and should be before the first "valley."

Option (1) is similar to (2) in that one or two poles suppresses the resonance peaks while contributing negative phase to crossover. However, option (2) is superior for it allows greater gain and thus higher accuracy at low frequencies. The difference between options (1) and (2) is that the second one adds the pole *before* crossover.

The alternatives listed above can be satisfied by one or two low pass filters:

$$G_{pole}(s) = \frac{K_2}{s+a}, \tag{5.4}$$

where $a > \omega_{c2}$ for option (1) and $\omega_0 < a < \omega_{c2}$ for option (2) (ω_0 is the fundamental resonance frequency, where the first peak is). The overall compensator is thus of the form:

$$G_{lag}G_{pole}(s) = K\frac{(s + 10\omega_{c1})}{(s+\omega_{c1})\ (s+a)}. \tag{5.5}$$

Because of the difficulty involved to simulate the actual loudspeaker in Matlab,[1] the constants in the above equation will be experimentally determined.

5.2 Circuit Implementations of the Closed Loop Loudspeaker System

Next we will realize the compensators given in the previous section. For practical reasons, we will use the single supply PWM driver. Consequently the compensators will also be single supply. We will reuse the 5V supply used for the PWM driver to bias the microphones. We also need a 7.5V reference for the op amps. It is implemented by two 1kΩ resistors. In addition, because the H-bridge input is TTL logic, the PWM driver input is $\frac{5}{2} \pm \frac{5}{3}$ V. The output from the compensator must be divided down 3 times. This can be easily done by a low resistance R-2R voltage divider. Note that we cannot easily use an op amp to implement the 1/3 factor because of the single supply limitation. Furthermore, we need an offset adjust in the feedback (or input) buffer op amp stage to null any op amp offsets and the static error due to the 7.5V reference. This prevents DC offsets in the loudspeaker. Though AC coupling can be used to isolate op amp DC offsets, it is not used in this project since additional circuitry in the forward path can complicate the system compensation.

---

1. Such a simulation requires a large number of poles and zeros. In the case of the actual loudspeaker, the unpredictable cone resonance makes the simulation almost impossible.

For the compensators given in the previous section, three circuits have been built. Each circuit has also been tested in a closed loop enclosing each one of the two loudspeakers.

5.2.1 Lag-2Pole Compensator

The first circuit (Figure 5.4) is the realization of the lag compensator $G_{lag}(s)$ and the two-pole compensator (1) given in the previous section. The system function is as follows:

$$G_{lag-2pole}(s) = K\frac{10\left(3.3\times 10^{-3}s+1\right)}{\left(33\times 10^{-3}s+1\right)\left(0.5\times 10^{-3}s+1\right)^{2}}, \tag{5.6}$$

where K is varied experimentally to give optimal results. The above compensator is designed to have a low system crossover frequency of around 15Hz and two poles at 320Hz, just before the first valley and the first cone resonance frequency.

In Figure 5.5, the measured frequency response of the compensated closed loop 12inch boxed speaker is compared to that of the uncompensated open loop (from Figure 5.2). Though there is some improvement in the closed loop system, the difference is small. Using the same compensator, the frequency response of the 10inch speaker system is given in Figure 5.6. Both Figures 5.5 and 5.6 show that the above compensation network does not improve performance substantially.

Figure 5.4: Lag-2Pole Compensator

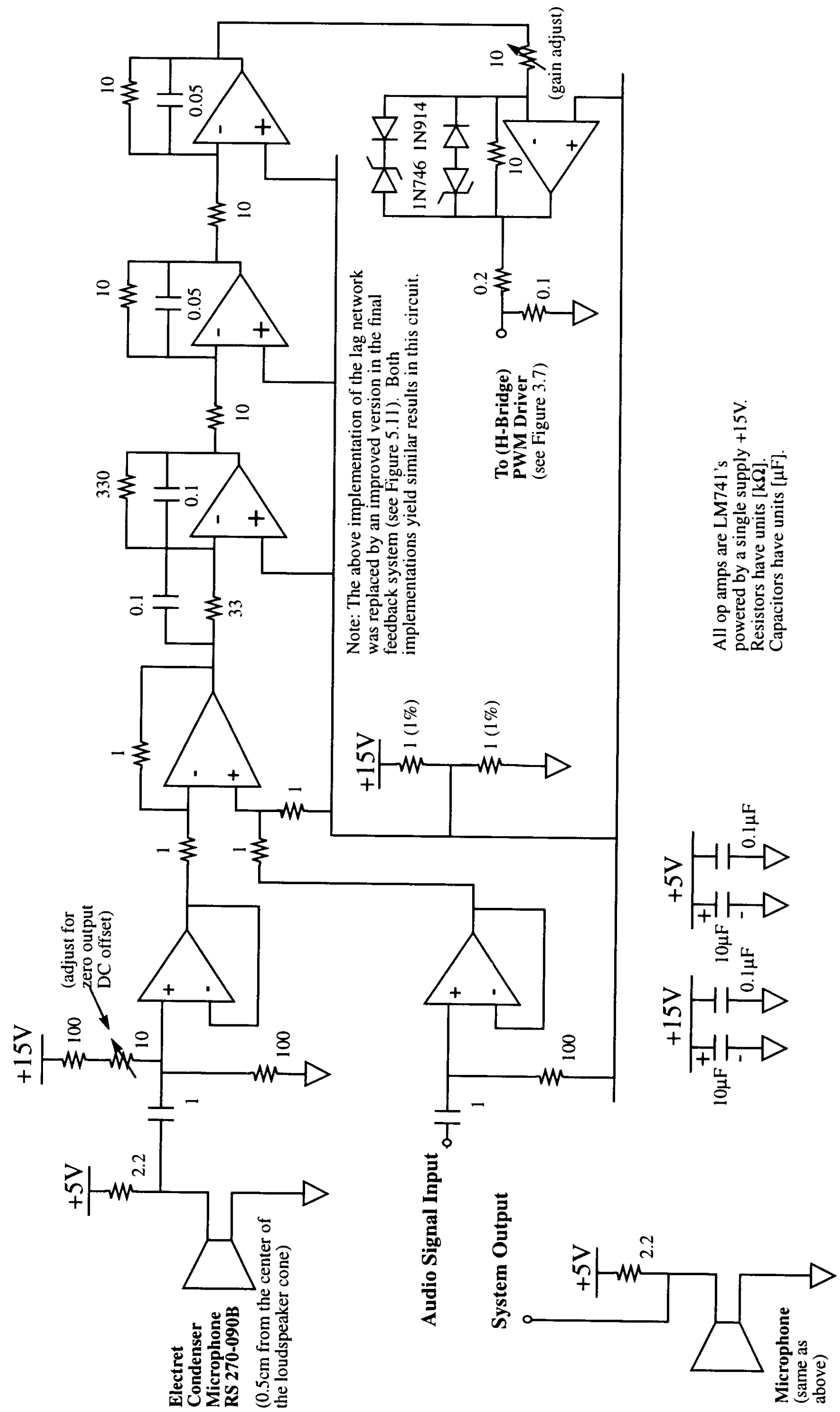
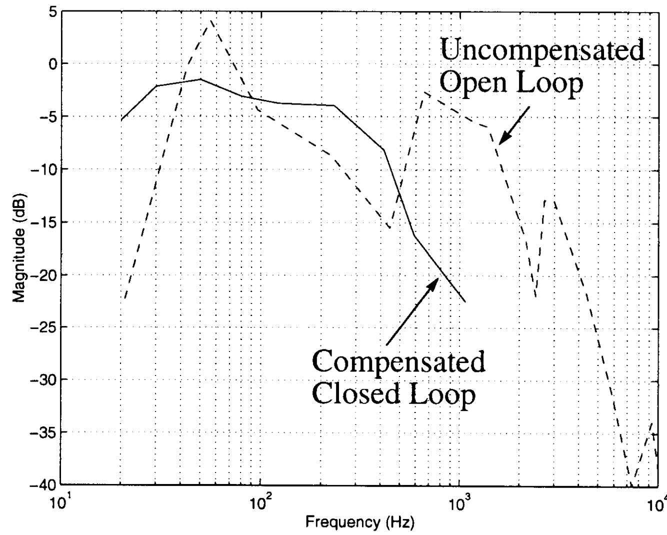
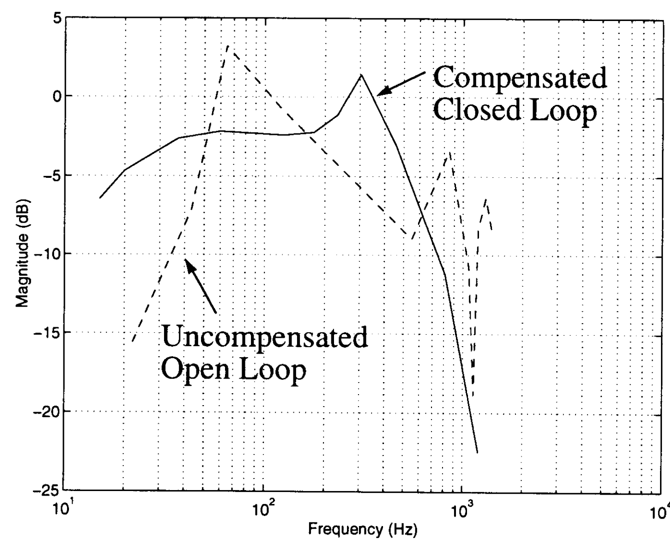


Figure 5.5: Frequency Response of the Closed Loop Lag-2Pole Compensator-Loudspeaker System (12inch)



Note: The measurement system setup is the same as that used for Figure 5.2. Input to the closed loop system is also a 100mVpp sinewave (from a signal generator KH 2000). The output is sensed by a matching microphone (RS 270-090B) as shown in Figure 5.4.

Figure 5.6: Frequency Response of the Closed Loop Lag-2Pole Compensator-Loudspeaker System (10inch)



Note: Same conditions as in Figure 5.5 apply.

5.2.2 Lag-Pole Compensator

The second circuit tested is the combination of $G_{lag}(s)$ and the single pole compensator (2) from the previous section. It is given in Figure 5.7. The corresponding system function for the compensator in this circuit is:

$$G_{lag-pole}(s) = K\frac{10\left(4.7\times 10^{-3}s+1\right)}{\left(47\times 10^{-3}s+1\right)\left(1.5\times 10^{-3}s+1\right)}. \tag{5.7}$$

This compensator is designed to have a low frequency crossover frequency of 10Hz, and a pole at around 100Hz -- shortly after the loudspeakers' fundamental resonance peak.

The measured closed loop characteristics of the 12inch and 10inch loudspeakers are given in Figures 5.8-5.10. They are compared to the open loop characteristics. The plots show that this compensator is superior to the earlier lag-2pole compensator for both loudspeakers. Again the 10inch system performs slightly better than the 12inch one does. This is partly due to the difference between the open loop characteristics of the two speakers. The 10inch speaker has a lower cone resonance peak so its gain can be raised slightly higher, thus improving accuracy in the operating range. For the compensated 10inch loudspeaker closed loop system, we see that it has a nearly flat frequency response from 30Hz to 500Hz (see Figure 5.9). The small peak at around 800Hz is caused by cone resonance and insufficient phase margin.

Because the compensator gain is maximized to increase accuracy in the operating range, frequency response and harmonic distortion at high frequencies can be worse. It is therefore necessary to band limit the input to this loudspeaker feedback system (and the next one). An electronic crossover network should precede the input so that only low frequency signals are passed to the loudspeaker feedback system.

Figure 5.7: Lag-Pole Compensator

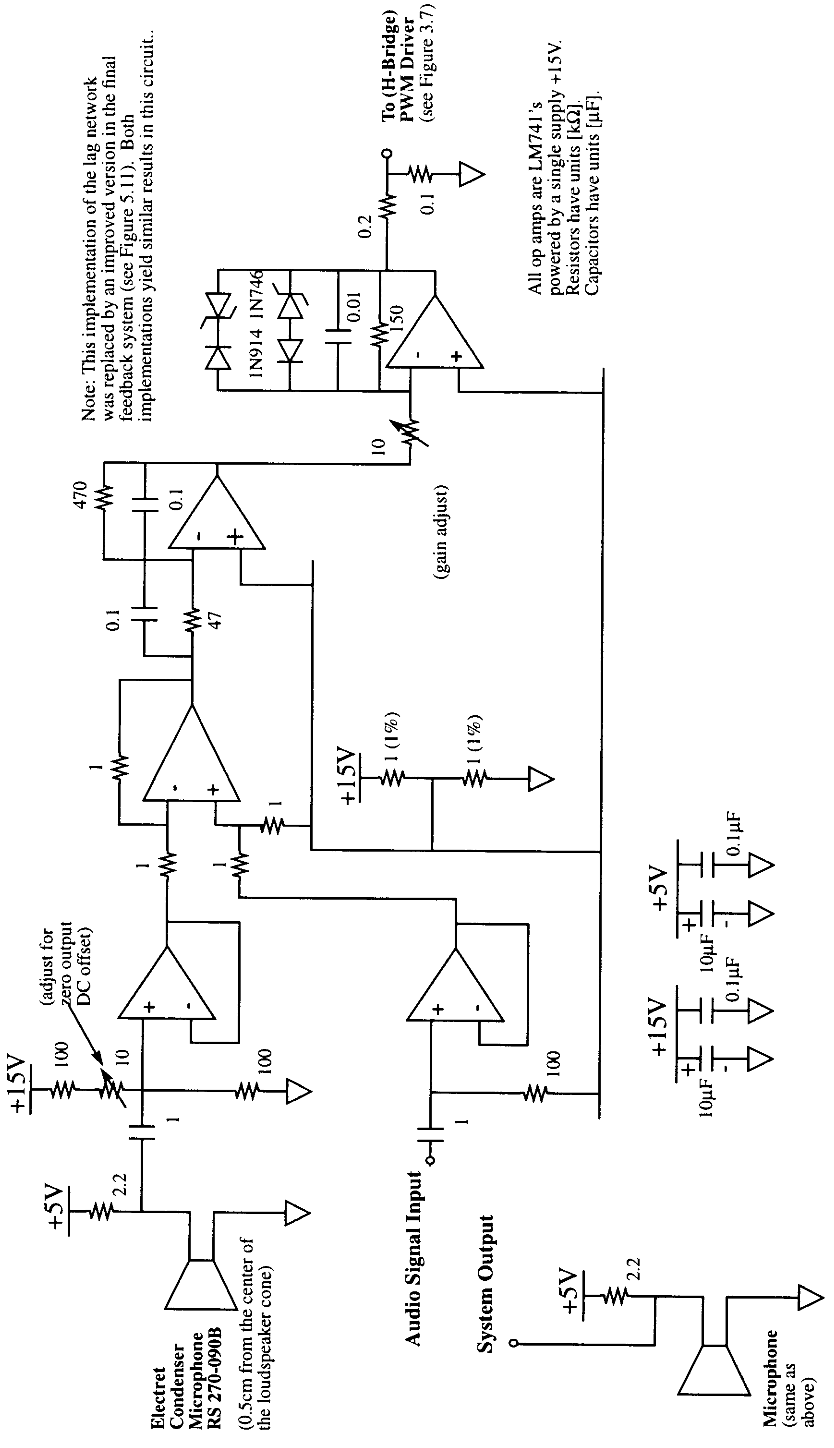
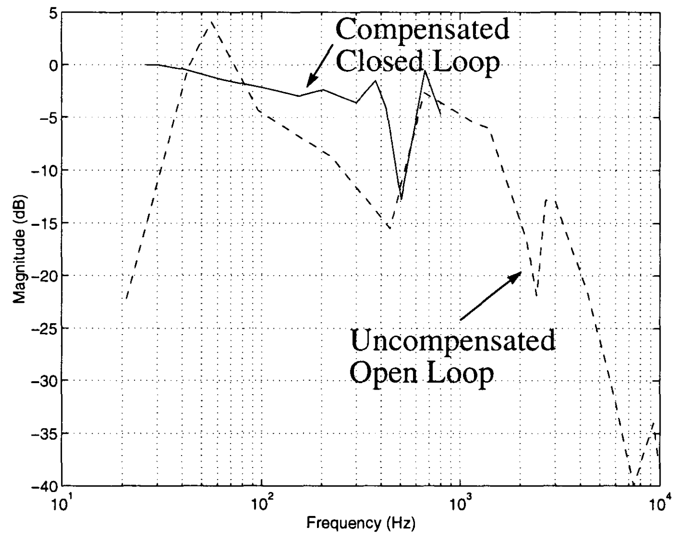
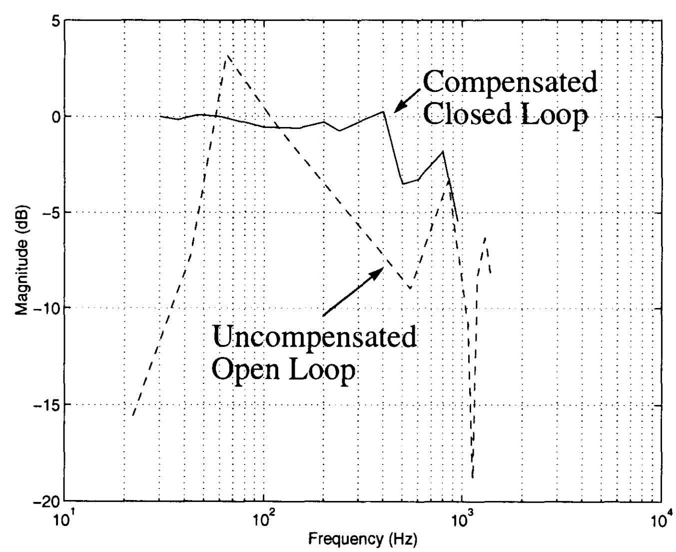


Figure 5.8: Frequency Response of the Closed Loop Lag-Pole Compensator-Loudspeaker System (12inch)



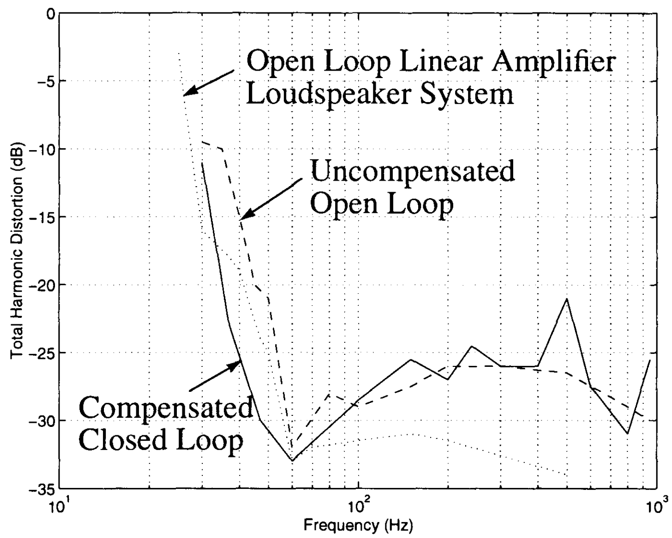
Note: The measurement system setup is the same as that used for Figure 5.2. Input to the closed loop system is also a 100mVpp sinewave (from a signal generator KH 2000). The output is sensed by a matching microphone (RS 270-090B) as shown in Figure 5.7.

Figure 5.9: Frequency Response of the Closed Loop Lag-Pole Compensator-Loudspeaker System (10inch)



Note: Same conditions as in Figure 5.8 apply.

Figure 5.10: Harmonic Distortion of the Closed Loop Lag-Pole Compensator-Loudspeaker System (10inch)



Note: The total harmonic distortion is measured by a distortion analyzer (HP 334A). Constant sound pressure level is obtained by manually adjusting the input to the systems such that the RMS value at the microphone output is 300mVpp. The measured SPL at 60Hz is 110dB at 1inch from the center of the cone.

In Figure 5.10, the harmonic distortion is compared among three systems: the open loop linear amplifier-loudspeaker system, the uncompensated open loop PWM driver-loudspeaker system and the compensated closed loop PWM driver-loudspeaker system. From the figure, we see that the closed loop reduces the harmonic distortion by about 2dB to 10dB from 35Hz to 80Hz.[1] However, it fails at frequencies greater than 100Hz. This is because at high frequencies the loop gain is no longer large enough to compensate for the distortion.

Figure 5.10 also indicates that the PWM driver introduces significant nonlinear distortion to the loudspeaker system, as compared to the linear amplifier. As we mentioned before, this nonlinearity is caused by the exponential waveform used in the pulse width modulation, and the timing of the switches in the switch drive. Even though

1. The low distortion level at around 60Hz for all three systems is a result of the loudspeaker system fundamental resonance. Only the fundamental at that resonance frequency is amplified by the system's acoustic structure, while all other harmonics are greatly attenuated.

the closed loop can reduce the harmonic distortion of the PWM driver-loudspeaker system, the improvement is not significant compared to the linear amplifier-loudspeaker system. Much of the improvement is offset by the extra nonlinearity of the PWM driver. It is therefore necessary to redesign the PWM driver so that it does not add significant nonlinearity to the loudspeaker system. This leads to the last circuit design.

5.2.3 Improved Lag-Pole Compensated Loudspeaker System

The final loudspeaker feedback system is basically a modified version of the above lag-pole system. Figure 5.11 shows the complete circuit diagram of the system. The switch drive in the PWM driver includes a true triangle wave oscillator. The triangle wave is obtained by integrating a schmitt trigger output in a positive feedback configuration [13]. In order to minimize the switching error in the comparator of the switch drive, the voltage swing of the oscillator is increased to 10Vpp and the frequency is decreased to around 26.5kHz.

Furthermore, a better lag network replaces the earlier one in the lag-pole compensator. This op amp implementation of a lag network is superior because it does not load the preceding circuitry through a capacitive path as in the earlier version. The circuit seems to be more stable with the new lag network. The preceding circuitry also includes an extra second order low pass filter. The damping ratio of the filter is about 0.7. The poles are at about 700Hz, where the loudspeakers' first cone resonance peaks are. The low pass filter does not add much negative phase shift at the crossover frequency, while suppressing the resonance peaks.

With the above modifications, the measured system characteristics are given in Figures 5.12-5.16.[1] Figure 5.12 shows that the closed loop 12inch loudspeaker system has a nearly flat frequency response from 35Hz to 200Hz. It is compared to the

1. Measurements at frequencies greater than 3kHz are corrupted by noise and are discarded.

uncompensated open loop frequency response in Figure 5.13. Figure 5.14 shows that the closed loop 10inch loudspeaker system is again better than the 12inch system, with a flat frequency response from 35Hz to 300Hz.[1] For comparison, the uncompensated open loop frequency response is also shown in Figure 5.15. Here we see that the uncompensated open loop frequency response of the PWM driver-loudspeaker system is slightly different from that of the linear amplifier-loudspeaker system given in Figure 1.1. Particularly at low frequencies, the magnitude plot of the former system seems to be more linear than that of the latter. The slope of the former plot at low frequencies is constant at 40dB/dec, as predicted by the model. This suggests that the op amp circuit actually changes the magnitude response of the loudspeaker system as given in Figure 1.1. However the magnitude deviation is small and mostly outside the audio range (20Hz - 20kHz).

The distortion plot in Figure 5.16 indicates that the modified PWM driver introduces little nonlinear distortion for almost all the frequencies of interest, as compared to the linear amplifier's distortion performance. The plot also shows that the closed loop is able to reduce the harmonic distortion by about 3dB to 10dB from 35Hz to 80Hz. For the same reason as in the earlier lag-pole compensator, the final feedback system fails to reduce the distortion at frequencies above 100Hz. From 100Hz to 300Hz, it seems that the loop even adds distortion. Nevertheless, the final closed loop loudspeaker system performs as expected at low frequencies (35Hz - 100Hz).

1. The small peak at around 250Hz is due to cabinet panel resonance that is also evident from the open loop Bode magnitude and phase plots of the loudspeakers. This resonance should not be considered as part of the system's performance.

Figure 5.11: A Loudspeaker Feedback System

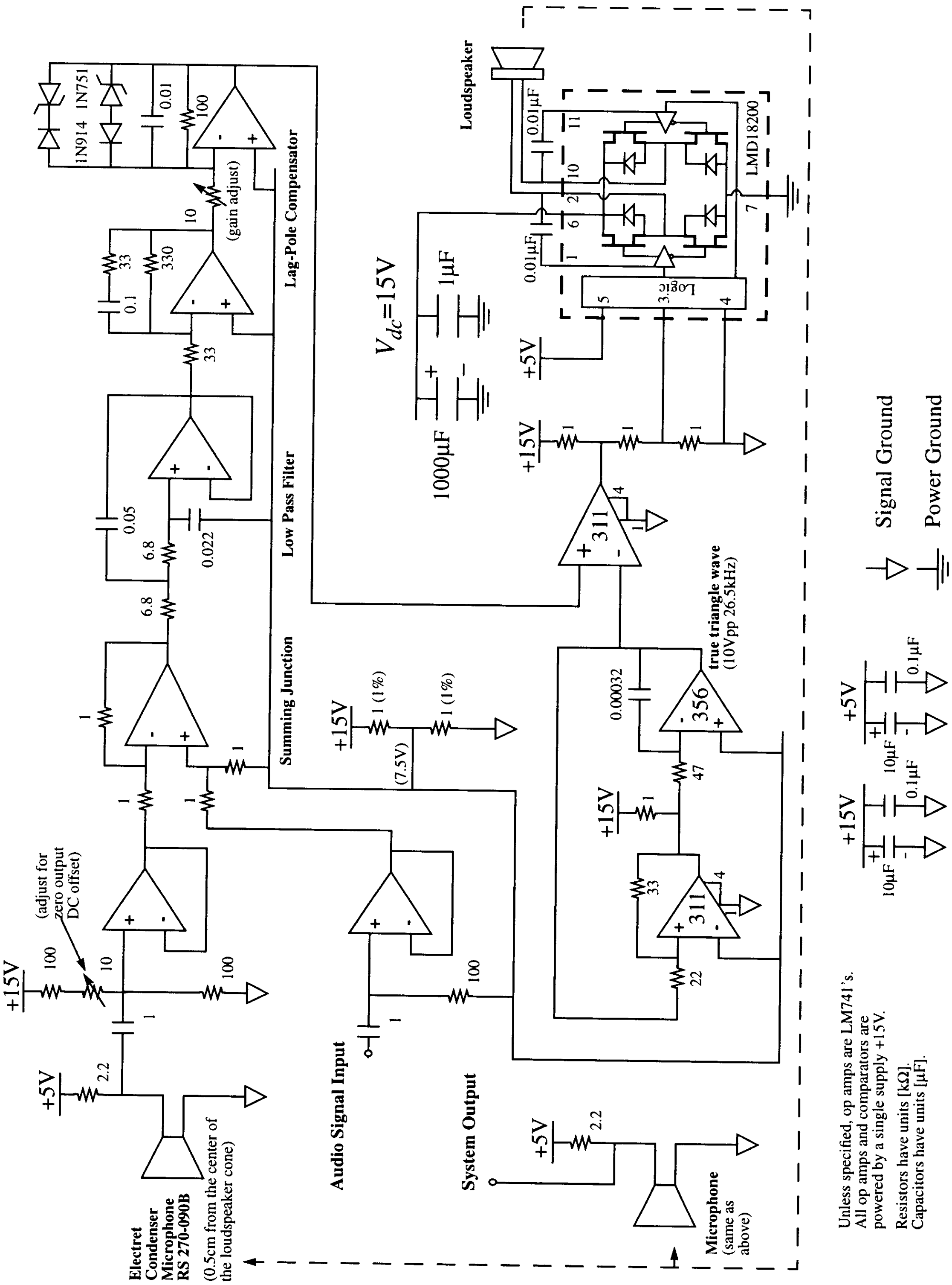
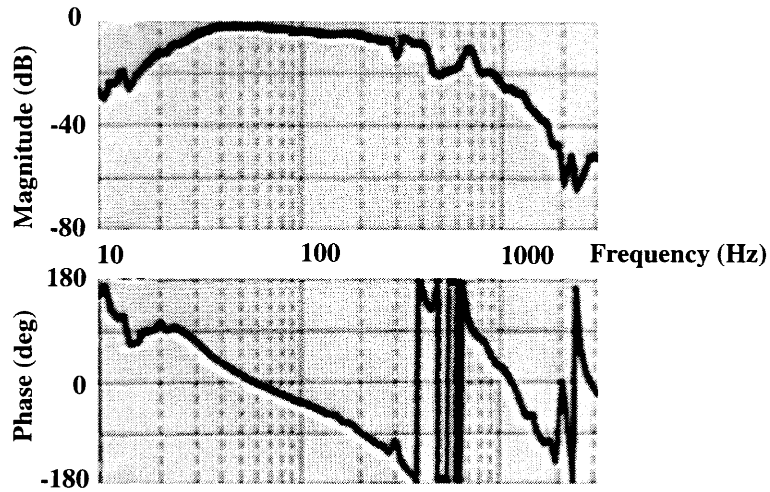
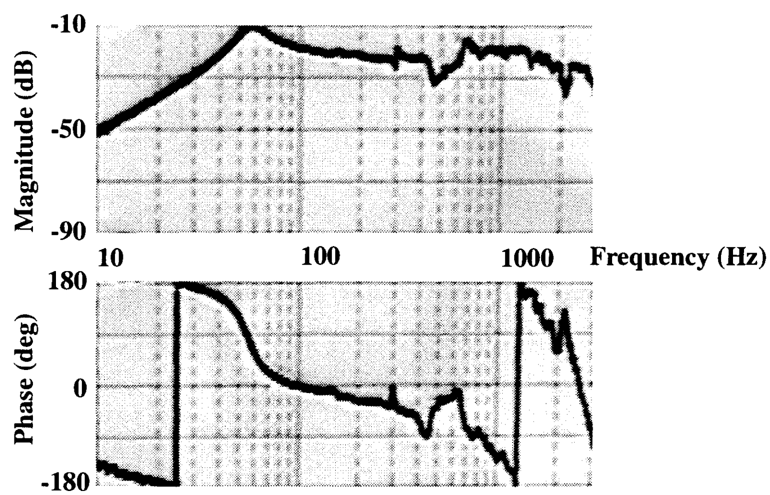


Figure 5.12: Frequency Response of the Loudspeaker Feedback System (12inch)



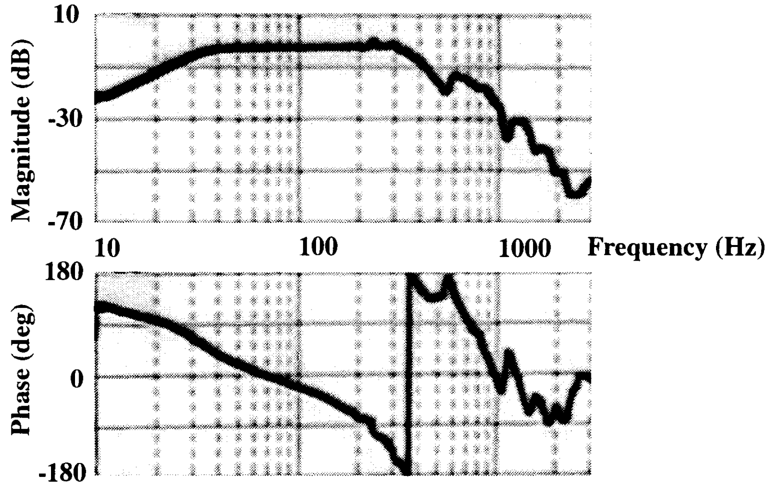
Note: The frequency response is measured using a dynamic signal analyzer (HP 3562A). The input is a sine sweep generated by the same instrument. The magnitude plot is automatically normalized to the peak magnitude. The measured sound pressure level at 60Hz is 110dB at 1inch from the center of the loudspeaker cone.

Figure 5.13: Frequency Response of the Open Loop System (12inch)



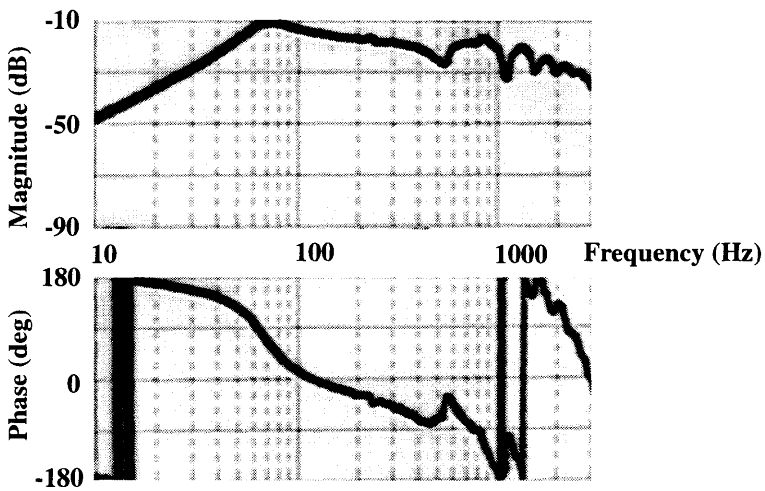
Note: The measurement setup is the same as that used for Figure 5.12. The open loop system is the same circuit given in Figure 5.11 minus the summing junction and the compensator prior to the switch drive. The measured sound pressure level at 60Hz is 110dB at 1inch from the center of the loudspeaker cone.

Figure 5.14: Frequency Response of a Loudspeaker Feedback System (10inch)



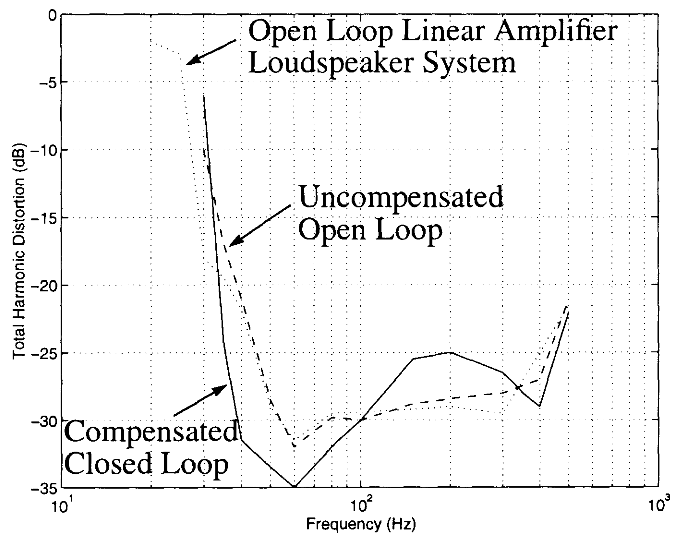
Note: The frequency response is measured using a dynamic signal analyzer (HP 3562A). The input is a sine sweep generated by the same instrument. The magnitude plot is automatically normalized to the peak magnitude. The measured sound pressure level at 60Hz is 110dB at 1inch from the center of the loudspeaker cone.

Figure 5.15: Frequency Response of the Open Loop System (10inch)



Note: The measurement setup is the same as that used for Figure 5.14. The open loop system is the same circuit given in Figure 5.11 minus the summing junction and the compensator prior to the switch drive. The measured sound pressure level at 60Hz is 110dB at 1inch from the center of the loudspeaker cone.

Figure 5.16: Harmonic Distortion of the Closed Loop 10inch Loudspeaker System



Note: The total harmonic distortion is measured by a distortion analyzer (HP 334A). Constant sound pressure level is obtained by manually adjusting the input to the systems such that the RMS value at the microphone output is 300mVpp. The measured SPL at 60Hz is 110dB at 1inch from the center of the loudspeaker cone.

Chapter 6

Discussion

The results given in Section 5.2 are satisfactory. The closed loop loudspeaker system shown in Figure 5.11 has been tested to give a nearly flat frequency response (30Hz - 300Hz) and reduced harmonic distortion at low frequencies (35Hz - 100Hz). Though the system built is a prototype, it is easy to see that the compensation technique is applicable to other loudspeaker systems. For example, a linear amplifier can be used rather than the PWM driver. Furthermore if better system components are used to replace the loudspeaker, microphone and the driver, then the closed loop system performance can be improved.

For the best performance in feedback applications, the loudspeaker should be reselected or redesigned to reduce cone resonance at mid and high audio frequencies. Even if the cone suspension (L_M) and weight (C_M) have to be altered, the cone resonance must be suppressed or pushed up to the high frequency range.

Since the PWM driver power supply voltages can be made arbitrarily large (up to the maximum limits of the op amps, comparators and switches), voltage scaling is possible in the circuits presented in this thesis. In particular, a large triangle wave is desired in order to reduce switching errors of the comparator in the PWM driver. Furthermore, the power amplification of the PWM driver can be increased by voltage scaling if necessary.

Lastly if a better microphone is used as the sensor, the closed loop system should perform better, especially at low frequencies. This is because a better microphone can detect low frequency sound more precisely. The microphone is the feedback element in the system. Therefore the performance of the microphone ultimately determines that of the closed loop loudspeaker system.

Chapter 7

Conclusions

A low-cost simple loudspeaker feedback system was built. The system consisted of a closed box loudspeaker, an electret condenser microphone and a simple "direct-drive" switching amplifier that is made up of mainly a switch drive and semiconductor switches. The final system is a modification of a closed loop system built for the ideal loudspeaker.

The ideal loudspeaker can be modeled by a simple equivalent electronic circuit. An estimated system transfer function can be quickly derived from given loudspeaker parameters. The estimated system function is accurate to within 1% error from the actual system function for all frequencies except very near the fundamental resonance of the loudspeaker, where the error is about 10%. Therefore the estimated system function can serve as a quick and accurate guide for system compensation.

The estimated system function shows that the ideal loudspeaker has two zeros at the origin and four additional poles -- two low frequency complex poles (at the fundamental resonance frequency, or around 20Hz - 150Hz), and two closely located real poles at high frequencies (at around 1kHz - 10kHz). Therefore this ideal loudspeaker exhibits a second order band-pass characteristic. The slope of the magnitude plot at low (and high) frequencies is 40dB/dec (-40dB/dec), while at mid frequencies it is nearly flat. The phase at low (and high) frequencies approaches 180° (-180°) while at mid frequencies it is close to 0°.

In addition, a simple direct PWM switching amplifier (PWM driver) was built. The semiconductor switches in the amplifier are applied directly to the loudspeaker. The loudspeaker therefore serves as the output filter for the switching amplifier, filtering out high frequency harmonics including that at the switching frequency (100kHz - 500kHz).

The switching action of the switches is done by fixed-frequency-variable-duty-ratio pulsed width modulation (PWM). Both dual-supply and single-supply versions of the PWM driver were built.

Using the above PWM driver, ideal loudspeaker model and a lag-lead compensator at the low and high frequency crossovers, the closed loop loudspeaker model system was built. The measured system performance was satisfactory. It shows that the system has a bandwidth close to that predicted by the crossover frequencies (3Hz<BW<25kHz). The system has a flat frequency response from 20Hz to 20kHz.

Though the ideal loudspeaker model system shows promising results, the real loudspeaker is difficult to compensate. Problems are caused by cone resonance and feedback signal sensing. The sensor used was an electret condenser microphone, which contributes little magnitude distortion and little phase shift over its (more-than-sufficient) operating frequency range. The non-minimum phase problem associated with the sound's propagation delay was minimized by placing the microphone next to the cone (0.5cm apart). However, the cone resonance problem cannot be solved by simple techniques. The cone resonance peaks and the associated large negative phase shift must be avoided.

Based on the closed loop loudspeaker model system, a modified compensator was used for the final loudspeaker system. The compensator is made of a simple lag network at around the low frequency crossover of the system, and a simple pole (low-pass) network before the high frequency crossover and the cone resonance frequencies. There is also a low pass filter included in the loop to further suppress cone resonance peaks. The closed loop PWM driver-loudspeaker-microphone system was built and tested. The measured performance of the compensated system is satisfactory. The system's frequency response is nearly flat at low frequencies (35Hz - 300Hz), and the total harmonic distortion is reduced by 3dB to 10dB from 35Hz to 100Hz.

The results also indicate that the compensation network performs better for a loudspeaker with a smaller cone. A smaller cone leads to less cone resonance. Unless the cone resonance problem is solved, the upper limit for this loudspeaker feedback system is at around 500Hz for a typical paper or polypropylene cone loudspeaker.

Appendix A

List of Variables

F	mechanical force, [N].
B	magnetic flux density in the air gap, [Weber/m^2].
l	length of wire in the voice coil, [m].
i	electric current, [A].
p	air pressure, [N/m^2].
v	velocity of the cone, [m/sec].
u	induced electromotive force, [V].
v_v	volume velocity, [m^3/sec].
r	radius of the cone, [m].
A	effective area of the loudspeaker cone, [m^2].
V	volume of the closed box, [m^3].
R_e, R_E	electrical resistance, [Ω].
R_m	mechanical responsiveness, [m/N-s].
R_M	mechanical responsiveness expressed as numerically equal electrical resistance, scaled by $(Bl)^2$, [Ω].
C	capacitance, [F].
L	inductance, [H].
L_e, L_E	inductance of the voice coil, [H].
M_m	mechanical mass of the cone, [kg].
C_M	mechanical mass expressed as numerically equal capacitance, scaled by $1/(Bl)^2$, [F].
C_m	mechanical compliance of cone suspension, [m/N].
C_b	compliance of the air in the closed box, [m/N].
L_M	total compliance of the cone and air suspension expressed as numerically equal electrical inductance, scaled by $(Bl)^2$, [H].
R_a, R_{a1}, R_{a2}	acoustic responsiveness, scaled by $1/A^2$, [m^5/N-s].
R_A	acoustic responsiveness expressed as numerically equal electrical resistance,

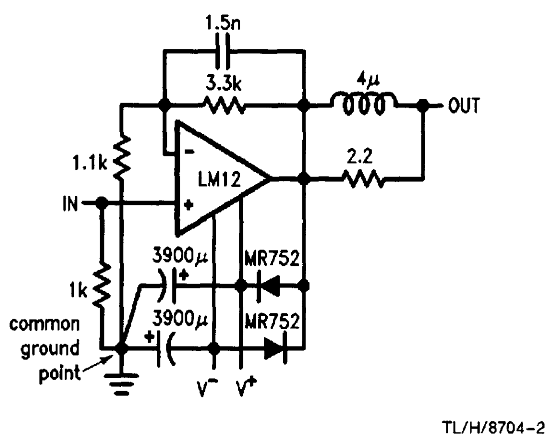
	scaled by $(Bl)^2$, [Ω].
M_a	acoustic mass of the air load, [kg].
M_{m+a}	total moving mass of the cone and air load, [kg].
C_A	acoustic mass expressed as numerically equal capacitance, scaled by $1/(Bl)^2$, [F].
C_a	acoustic compliance of the air in front of the cone, [m/N].
ρ_0	air density, 1.18kg/m^3.
c	speed of sound in ambient air, 345m/sec.
λ	wavelength of sound waves, [m].
f	frequency, [Hz].
v_i	system input, [V].
v_o	system output, [V].
D	duty ratio.
V_{ref}	reference voltage, [V].
V_{var}	variable voltage, [V].
V_{dc}	DC supply voltage input, [V].
V_o	output voltage, [V].
e_{ss}	steady state error.
G_c	compensator's system function.
G	open loop system function.
K	compensator gain.
ω_c	crossover frequency, [rad/sec].
ω_{c1}	low frequency crossover frequency, [rad/sec].
ω_{c2}	high frequency crossover frequency, [rad/sec].
ω_0	fundamental resonance frequency, [rad/sec].
f_0	fundamental resonance frequency, [Hz].
τ	time constant.
T_d	time delay, [s].
ω	frequency, [rad/sec].
Δ	distance, [m].

T	fixed period of the PWM waveform, [s].
f_{sw}	switching frequency, [Hz].
PWM	pulse width modulation.
BW	bandwidth.
SPL	sound pressure level, [dB].
THD	total harmonic distortion, [dB] or [%].
EMF	electromotive force, [V].
MOSFET	metal-oxide-silicon field effect transistor.
NMOS	N-type MOSFET.
PMOS	P-type MOSFET.
TTL	transistor-transistor logic.

Appendix B

Low Distortion Audio Amplifier

The circuit below is a linear amplifier that uses the power operational amplifier LM12 from National Semiconductor. The op amp is capable of driving ± 25V at 10A from ±30V supplies. It can deliver 80W of sinewave power into a 4Ω load with 0.01% distortion. Intermodulation distortion of the linear amplifier is 0.015% (60Hz/7kHz, 4:1). "Transient response and saturation recovery are clean, and the 9V/µs slew rate of the LM12 virtually eliminates transient intermodulation distortion." [4]



*Low distortion (0.01%) audio amplifier

Source: National Semiconductor

Appendix C

Loudspeaker Parameters

Listed are specifications for two loudspeakers used in this project. The speakers are polypropylene-cone replacement "woofers." One is 10inch (RS 40-1014A) and the other is 12inch (RS 40-1026A).

Parameters	Symbols	Units	10inch	12inch
Nominal Impedance		Ω	8	8
Frequency Response		Hz	30 -2500	25 - 3000
Free Air Resonance Frequency	f_0	Hz	30	25
Infinite Baffle Resonance Frequency		Hz	29	23
Effective Cone Area	A	m^2	0.0360	0.0523
Rated Power Input - Nominal		W	50	50
RMS Thermal Power Limit	P_{max}	W	100	100
Flux Density - Wire Length Product	Bl	Weber/m	8.27	6.66
Voice Coil Resistance	R_e, R_E	Ω	6.6	5.6
Voice Coil Inductance @1kHz	L_e, L_E	mH	1.13	0.45
Sound Pressure Level	SPL	dB/1W/1m	87±2	88±2
Electrical Q Factor			0.61	1.13
Mechanical Q Factor			4.6	3.13
Total Q Factor			0.54	0.83
Equivalent Acoustic Volume		L	181.2	297.19
Mechanical Compliance	C_m	μm/N	963.17	766
Mechanical Mass of Cone	M_m	g	31.14	54.59
Moving Mass of Cone and Air Load	M_{m+a}	g	35.19	61.46
Peak-to-Peak Linear Excursion		cm	2.25mm	2.90
Power Handling		W_{rms}	50	50

Source: Radio Shack

Appendix D

Additional Loudspeaker Model Parameters

The following equations are used to calculate the circuit parameters in the model in Chapter 2. Most of them are taken from sources [9] [10].

$$A = \pi r^2, \tag{D.1}$$

$$C_b = \frac{V}{\rho_0 c^2 A^2}, \tag{D.2}$$

$$R_m = \frac{R_M}{(Bl)^2}, \text{ where } R_M = 200\Omega \text{ nominally}, \tag{D.3}$$

$$M_a = M_{m+a} - M_m, \tag{D.4}$$

$$R_a = \frac{0.221}{r^2 \rho_0 c}, \tag{D.5}$$

$$R_{a1} = \frac{0.721}{r^2 \rho_0 c}, \tag{D.6}$$

$$R_{a2} = \frac{0.318}{r^2 \rho_0 c}, \tag{D.7}$$

$$C_a = \frac{0.6}{r \rho_0 c^2}. \tag{D.8}$$

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