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Amplifier Design in a Motional Feedback Audio System

Bachelor Graduation Project - EE3L11

Thesis

Delft University of Technology

Project Goup G

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Challenge the future

Abstract

The quality of audio amplifiers used in Hi-Fi sound installations is largely determined by the distortion of the output signal. This report proposes a method of improving the distortion reduction in audio amplifiers by using a topology with active feedback. The characteristics of the standard non-inverting amplifier are extended to this new topology and the stability problems caused by the active feedback are resolved with frequency compensation. The amplifier discussed in this report is developed for a motional feedback application, therefore a full audio system including cross-over filters, a subtracting amplifier and a displacement sensor is designed. Lastly a PCB is designed to integrate the different circuits into an end product.

Preface

This report is written as part of the Bachelor Graduation Project (BAP) for the Electrical Engineering bachelor at the TU Delft. The project group consists of six electrical engineering bachelor students, who are divided into subgroups of two and are each devoted to a subsystem of the total project.

The assignment was provided by Dr. Ir. G.J.M. Janssen out of interest in the digital implementation of the motional feedback system. Janssen provided further guidance during the project, for which we would like to thank him.

Other than Janssen we would like to thank the employees of the Tellegen Hall at the TU Delft for their help and the budget they provided. We also want to thank Ioan Lager for the coordination of the project. Lastly we want to thank the members of the jury for reviewing our work.

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Definition of Parameters

Parameter	Definition
α	Attenuation coefficient
a_{rms}	The RMS value of the acceleration of the speaker cone
A	Open loop voltage gain
A_1	Open loop voltage gain of the high speed operation amplifier
A_2	Open loop voltage gain of the audio amplifier
A_ϕ	The speaker cone area
β	Feedback gain
β_1	Feedback gain of the nested feedback amplifier
β_2	Local feedback gain of the audio amplifier in the nested feedback amplifier
B	Magnetic flux density
C_m	The Miller input capacitance
C_{m0}	The Miller output capacitance
d	Distance from the speaker to the receiver
D	Attenuation
D_{out}	Output distortion
D_a	Distortion of the audio amplifier
ΔA	Factorial improvement in open loop voltage gain
f	Frequency
f_0	Cutoff frequency
f_m	Cutoff frequency of the dominant pole compensation network
f_p	The frequency placement of the pole introduced by lead compensation
f_z	The frequency placement of the zero introduced by lead compensation
g	Subtracting amplifier gain
G	Closed loop gain
$H(s)$	Transfer function in the Laplace domain
$H(s)_{HP4}$	Transfer function of a 4th order high-pass filter in the Laplace domain
$H(s)_{LP4}$	Transfer function of a 4th order low-pass filter in the Laplace domain
$H(s)_{LR4}$	Transfer function of a 4th order Linkwitz-Riley cross-over in the Laplace domain
I	Current through the voice coil
l	Length of the wire in the voice coil
L	Loop gain
m	Mass of the piston
ϕ	Phase shift of the loop gain
Φ	Phase margin
Q	Quality factor
ρ	The density of air

Parameter	Definition
S	Sensitivity of the amplifier
s	Laplacian variable
t	Time in seconds
θ	Phase of the system at $G = 1$
$\hat{v}$	Amplitude of the velocity of the speaker cone
v_{rms}	The RMS value of the velocity of the speaker cone
V_{audio}	Maximum input voltage amplitude of the audio signal
V_{mfb}	Maximum input voltage amplitude of the motional feedback signal
$V_{in,A}$	Maximum input voltage amplitude of the amplifier
V_{ios}	Offset voltage originating from the typical input bias offset current of an operation amplifier
V_{off}	Typical offset voltage of an operation amplifier
$V_{out1,A}$	The maximum output voltage amplitude of the high speed operational amplifier
$V_{out2,A}$	The maximum output voltage amplitude of the amplifier
$V_{out2,rms}$	The maximum RMS output voltage of the amplifier
V_{-amp}	Voltage at the minus terminal of the op amp
V_{out}	Voltage at the output of the op amp
V_{in}	Input voltage
ω	Frequency
ω_0	Cutoff frequency

Abbreviations

Abbreviation	**Definition**
AC	Alternating current
DC	Direct current
MEMS	Microelectromechanical system
MFB	Motional feedback
N	Noise
Op Amp	Operational amplifier
PCB	Printed circuit board
RC	Resistor-capacitor
RMS	Root mean square
SNR	Singal to noise ratio
THD	Total harmonic distortion

Chapter 1

Introduction

1.1 Background[1]

1.1.1 History of motional feedback

In the year 1968 Philips released a paper on motional feedback (MFB),[1] a technique where electro-acoustic feedback is used to reduce the linear and non-linear distortion of a speaker system. Philips and various other manufacturers sold audio systems with the motional feedback system for about a decade. Unfortunately the costs of the analogue equipment necessary for the implementation did not make the system worthwhile in the long run and it disappeared from the market.

Recent development in the costs and capabilities of digital devices have provided an opportunity for bringing back the motional feedback system by designing a digital implementation. The advantages of the modern techniques can make motional feedback an affordable extension to modern day audio systems.

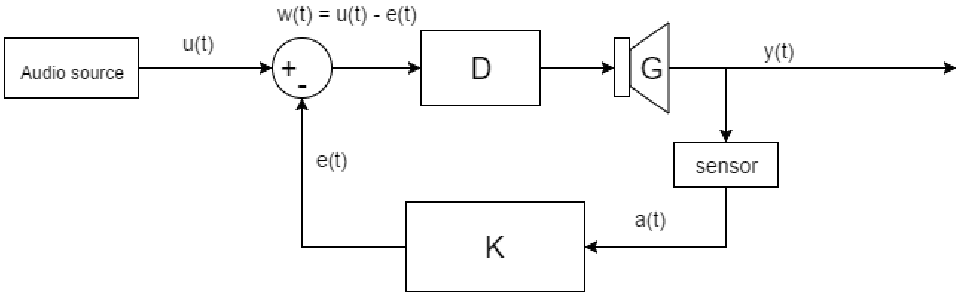
1.1.2 Principle of digital motional feedback

When playing sound over a loudspeaker, an electrical waveform resembling the sound is converted into air pressure waves which are sensed by human ears and perceived as sound by the brain. Because a loudspeaker is a non-ideal, physical system, this conversion from voltage to air pressure distorts the sound by for example damping out low frequencies (linear distortion) and adding harmonics that weren't present in the original waveform (non-linear distortion). Ideally, a loudspeaker would output the exact sound it receives at it's input, so finding a way to decrease this distortion is of interest.

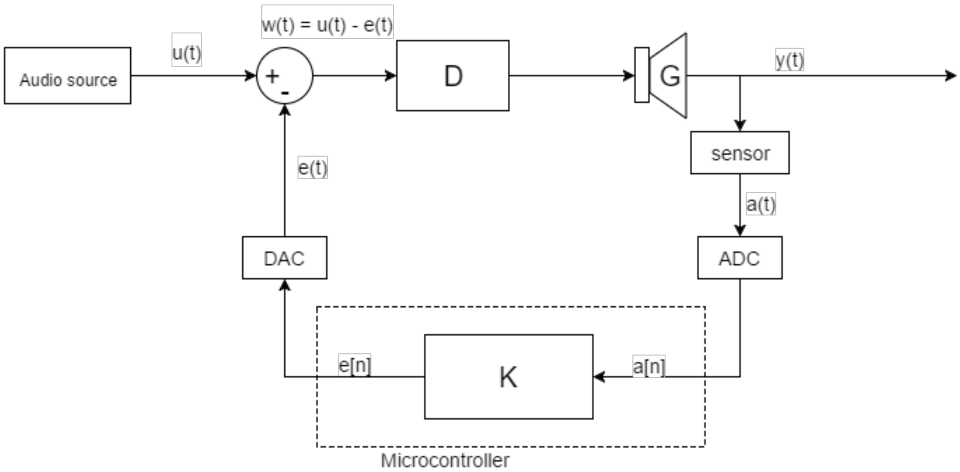
One way to do so is by using a motional feedback system, where the output of the speaker is measured by a sensor and used to calculate the error in the input signal. This error signal is then fed back to the input (fig. 1.1a) to correct it. To calculate this error signal from the measured output, an LTI filter called the *controller* is designed using Control System theory so that it corrects the input of the loudspeaker, thereby altering the output sound to more closely resemble the input waveform.

Digital MFB implements this controller on a digital platform. This requires sampling of the loudspeaker's measured output signal and reconstruction of the controller's output signal to an analogue voltage for comparison with the input signal (fig. 1.1b).

[1]Because it discusses the project in general, this section occurs in the theses of all three subgroups (see section 1.2).



(a) Controller implemented as analogue system.



(b) Controller implemented digitally.

Figure 1.1: Layout of a motional feedback system around a loudspeaker system with speaker G and amplifier D. Shown are the controller K, audio input signal $u(t)$, speaker output pressure $y(t)$, conus acceleration $a(t)$ and $a[n]$ and the error signal calculated by the controller $e(t)$ and $e[n]$.

1.2 Division into subsystems[2]

The design of the digital motional feedback device is split into three parts for the purpose of this project. The interfaces between these parts are defined as part of the Programme of Requirements (see chapter 2).

Digital signal processing
The first part is devoted to the digital signal processing, which is highlighted green in figure 1.2. The main tasks of this design are:

- Digital-Analogue conversion
- Analogue-Digital conversion
- Providing a configurable platform for the controller

The performance of this part is highly dependant on the delay in all respective stages.

The design of this subsystem is documented in a different thesis. See [2] for details.

[2]Because it discusses the project in general, this section occurs in the theses of all three subgroups.

Analogue components
The second part is the design of the analogue system components which are highlighted red in figure 1.2. The aim of this part is designing the motional feedback system for high fidelity applications. Fittingly the amplifier constructed uses active feedback to reduce output distortion. All tasks belonging to this group are:

- Subtracting the controller's error signal from the audio
- Cross-over filtering of the audio
- Amplifying the audio signal with minimal distortion
- Characterising the speaker
- Measuring the displacement of the loudspeaker cone

This thesis will focus on the analogue components subsystem. For details on the design of the other two subsystems, please refer to [2] and [3].

Control and system identification
The final group designs the controller highlighted in blue in figure 1.2 and implements system identification. The control group is tasked with implementing the motional feedback that makes the system closed loop stable and reduces distortion. All tasks of this group are:

- Designing the motional feedback controller
- Implementing system identification that identifies the second order characteristic of the speaker the motional feedback is applied to

Because the system contains system identification it can be applied to all enclosed speakers. This means no re-design is required for installing the motional feedback module on a different second order speaker system.

The design of this subsystem is documented in a different thesis. See [3] for details.

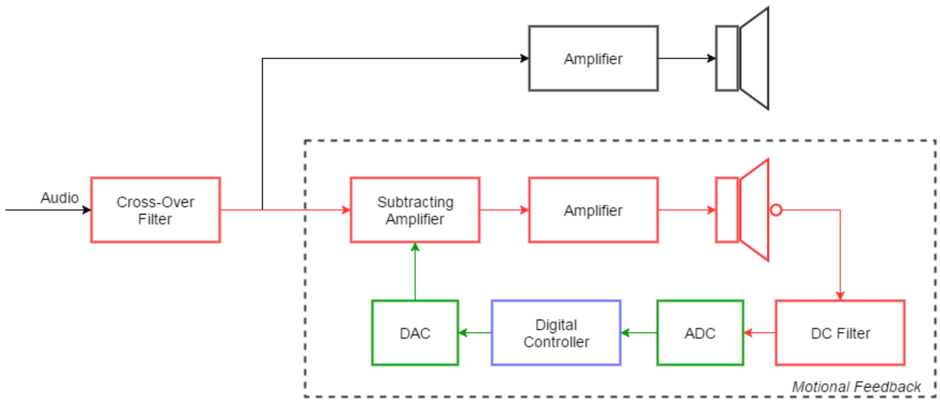


Figure 1.2: The motional feedback system designed by the three groups. The green parts are designed by the digital group. The parts highlighted in red are related to the amplifier/filter group. The blue part is the controller implemented by the control group. The MFB system is shown here in a two-way loudspeaker setup, where the tweeter (on top) does not use motional feedback.

1.3 State-of-the-art Analysis

Amplifiers used for audio applications are generally designed using a single feedback loop [5] as is shown in figure 1.3.

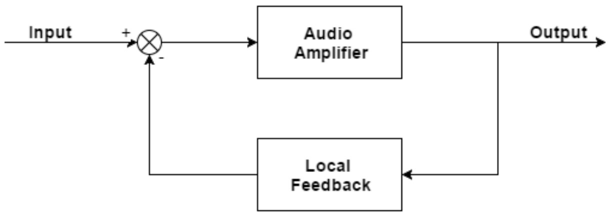


Figure 1.3: The general control scheme used for implementing audio amplifiers.

The feedback in this circuit reduces the distortion produced by the audio amplifier significantly, however the distortion in the output is not negligible. To improve the distortion reduction an additional amplifier is placed before the non-inverting amplifier and a secondary feedback path is added. The topology, called nested feedback, contains both local and overall feedback. Because the added amplifier is included in the feedback of the audio amplifier, the amplifier is also called an active feedback amplifier. The control scheme of this topology is shown in figure 1.4.

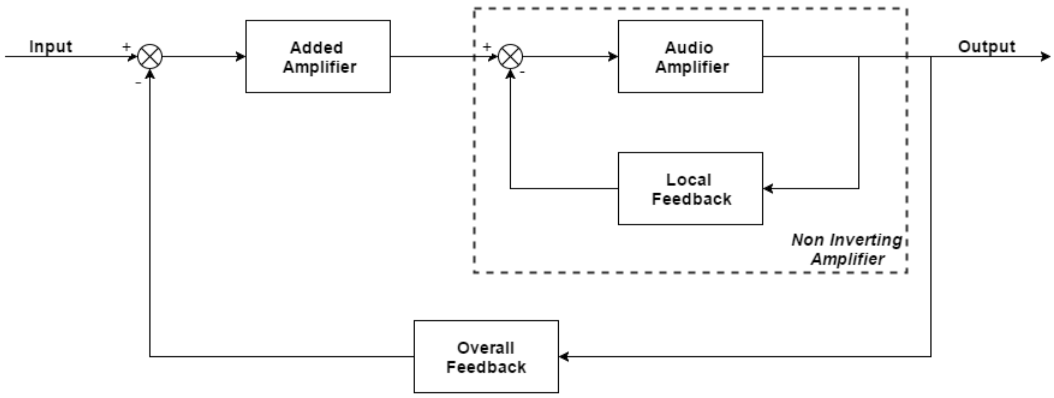


Figure 1.4: The control scheme of the nested feedback topology.

In [6] the application of local feedback in an audio amplifier and the results that follow are discussed. The author designs the resulting amplifier with bipolar transistors for multiple local feedback loops. In this report the control scheme depicted in figure 1.4 is implemented using op amps. A single amplifier and feedback loop is added to the non-inverting amplifier.

The choice of op amps with regards to their frequency behaviour is important for the stability of circuit, as phase shift accumulation of the devices can cause oscillations [5]. Selecting the right op amps eliminates the problems caused by their phase shift, however it does not resolve all stability problems in the topology. Therefore the frequency behaviour will need adjustment.

Single drivers can not have an adequate sound pressure level at a bandwidth of 20 Hz to 20 kHz on their own. Multiple drivers are needed, which requires the use of cross-over filters. There is one rule for the design of these filters: *The cross-over of a loudspeaker must be inaudible.* [4]. This implies that the power response of the two drivers must be similar in the cross-over region.

1.4 Chapter organisation

The instability in the active feedback topology is due to the increase in loop gain. The remainder of this report will address the stability issues without losing the improvement in loop gain. To that end, the non-inverting amplifier is analysed on its distortion and stability characteristics. The results of the non-inverting amplifier analysis are then extended to the new topology and are used as a basis for the design of the circuit. Frequency compensation is then adequately applied to stabilise the new topology and lastly simulation results are provided.

Additionally the other analog components in the motional feedback system are constructed. Firstly the cross-over filters are discussed and designed. Then the device implementing the subtraction of input audio and input feedback is designed. Lastly various methods of measuring the piston displacement are discussed based on the physics involved. The best method and sensor type is derived based on both a physical and practical motivation.

Chapter 2

Program of Requirements

This section covers the requirements of the system based on the state of the art analysis and limitations of the project itself. First the requirements of the total system are listed. These outline the overall purpose, the time and money restrictions, the input and output, and the criteria of the system transfer. Next the derived criteria for the subsystem designed in this thesis are listed.

1. **Total System**
   - 1.1. **Overall**
     - 1.1.1. *The feedback must be implemented digitally.*
     - 1.1.2. *The system must be able to log its input audio stream and controller parameters.*
     - 1.1.3. *The system must be adjustable to the speaker.*
     - 1.1.4. *The system must work on second order speaker systems.*
   - 1.2. **Input Output**
     - 1.2.1. *The maximum peak-to-peak input voltage of the audio signal is* $4\,V$.
     - 1.2.2. *The maximum continuous output power is* $50\,W$.
     - 1.2.3. *The DC output may not exceed* $50\,mV$.
   - 1.3. **System Transfer**
     - 1.3.1. *The gain of the system must be* $28\,dB \pm 1\,dB$.
     - 1.3.2. *The THD is restricted by the mask shown in appendix A.*
     - 1.3.3. *The motional feedback is applied in the bandwidth of* $10\,Hz - 1\,kHz$.
     - 1.3.4. *The total system outputs signals in the bandwidth of* $10\,Hz - 20\,kHz$.
     - 1.3.5. *The SNR is at least* $80\,dB$ *at the output inside the specified bandwidth of the motional feedback.*
   - 1.4. **Development**
     - 1.4.1. *The development time is 9 weeks.*
     - 1.4.2. *The development costs must be below €300.*

2. **Amplifier / Filter Specific**

2.1 **Input Specifications**

2.11 *The maximum peak-to-peak input voltage of the audio signal is* $4\,V$.

2.12 *The maximum peak-to-peak input voltage of the motional feedback is* $4\,V$.

2.13 *The input impedance should be higher than* $1\ k\Omega$.

2.14 *The bandwidth of the input signal is* $10\,Hz - 20\,kHz$.

2.2 **System Specifications**

2.21 *The system amplifies voltage by a factor of* $28\ dB$.

2.22 *The magnitude of the transfer for the amplifier of bass frequencies is* $28\ dB$ *from* $10\ Hz$ *to* $1\ kHz$, *with a maximal error of* $\pm 1\ dB$.

2.23 *The phase of the amplifier transfer is* $0°$ *from* $10\ Hz$ *to* $1\ kHz$, *with a maximal error of* $\pm 22.5°$.

2.24 *The amplifier is stable with a phase margin of at least* $45°$.

2.25 *The source voltages of the amplifier and filters are less than* $50\,V$

2.26 *The amplifier in nested configuration must have a minimal improvement in distortion reduction of* $20\ dB$ *compared to an amplifier with equal gain and the same op amp in a non-inverting configuration.*

2.3 **Output Specifications**

2.31 *The maximal continuous output power is* $50\,W$.

2.32 *The output bandwidth is* $10\,Hz - 1\,kHz$.

2.33 *The phase shift of the cross-over filter is linear at the bandwidth of* $10\,Hz - 1\,kHz$.

2.34 *Attenuation of the filter at* $1.5\ kHz$ *is at least* $15\ dB$.

2.35 *The SNR is at least* $80\ dB$ *at the output inside the specified bandwidth of the motional feedback.*

2.36 *The DC output may not exceed* $50\,mV$.

2.4 **Mechanical Specifications**

2.41 *The weight of the sensor is less than* $50\ g$

2.42 *The circuit with the amplifier, filters and subtracting op amp weighs less than* $1\ kg$.

2.43 *The maximal dimensions of the end product in the previous requirement are* $15x15x5\ cm$.

Chapter 3

Design Process

3.1 Non-inverting Amplifier Analysis

Audio amplifiers are generally constructed using a single op amp. Figure 3.1 shows an op amp in a non-inverting configuration, dimensioned for the specified gain of 28 dB.

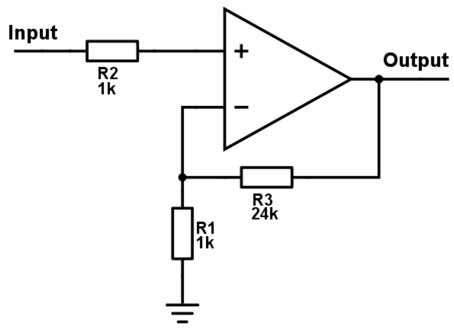


Figure 3.1: A non-inverting op amp amplifier with a gain of 28 dB as specified by specification 2.21

The control scheme of this amplifier is depicted below in figure 3.2.

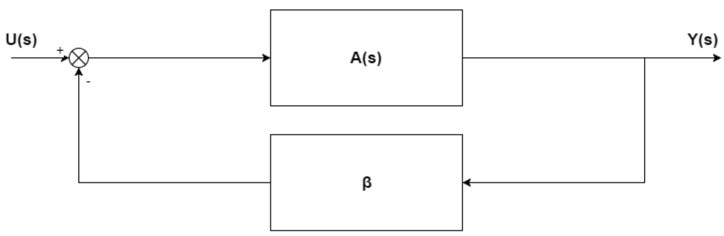


Figure 3.2: The classic feedback control scheme

$A(s)$ represents the transfer function of the amplifier and is called the open loop gain. The Laplace variable s denotes the dependency of this variable on frequency, but is omitted further on, to keep the equations compact and readable. β is the feedback gain. The loop gain in this scheme is defined as:

$$L = A\beta \tag{3.1}$$

The closed loop gain, which defines the relation between output and input, is given by:

$$G = \frac{Y}{U} = \frac{A}{1 + A\beta} \tag{3.2}$$

Generally the loop gain $A\beta$ is large and allows G to be approximated as:

$$G = \frac{1}{\beta} \tag{3.3}$$

Which shows that the closed loop gain, for large amplifier gains, is determined by the feedback network.

3.1.1 Stability Analysis

The stability of the system shown above is examined by analysing the loop gain. When the loop gain is not attenuated and shifted 180° in phase, the negative feedback is translated into positive feedback and the systems transfer becomes unstable [5]. A mathematical translation of these criteria is given by equation 3.7.

$$|A\beta| \geq 1 \qquad \textit{and} \qquad |\phi| > 180^\circ \tag{3.4}$$

Where ϕ denotes the phase shift of the loop gain. This criterion can be translated into the frequency domain using the phase margin. The phase margin is defined as:

$$\Phi = 180^\circ - |\theta| \tag{3.5}$$

In this equation θ denotes ϕ at $L = 1(0\,dB)$. Equation 3.4 is translated to:

$$\Phi > 0 \tag{3.6}$$

This makes the system stable in theory, however a smaller phase margin results in higher overshoot and negatively impacts the behaviour of the circuit. Traditionally control systems are designed to have a phase margin of at least 45° [9]. The system thus needs to satisfy equation 3.7 (as stated in specification 2.24).

$$\Phi > 45 \tag{3.7}$$

3.1.2 Distortion Analysis

By differentiating the closed loop gain to the transfer A, the sensitivity of the system can be derived [6, pp. 266]:

$$\frac{\partial G}{\partial A} = \frac{1}{(1 + A\beta)^2} \tag{3.8}$$

The relation between small changes in the amplifier and the resulting change in the output is given by equation 3.9.

$$\frac{\delta G}{G} = \frac{1}{1 + A\beta} \frac{\delta A}{A} \tag{3.9}$$

This equation shows the output is varying with a factor on changes in the amplifier. This factor is called the sensitivity S and in this system equals:

$$S = \frac{1}{1 + A\beta} = \frac{1}{1 + L} \tag{3.10}$$

Non-linearities originating from the op amp are reduced by this sensitivity as well[6, pp. 266]. Using this result the distortion in the output may be calculated:

$$D_{out} = \frac{D_a}{S} = \frac{D_a}{1+L} \tag{3.11}$$

D_a represents the distortion that is produced by the op amp. In order to reduce the distortion in the output, the sensitivity should be increased. This is accomplished by increasing the loop gain. Since the loop gain is large equation 3.11 can be approximated to:

$$D_{out} \approx \frac{D_a}{L} \tag{3.12}$$

This equation holds for a single op amp amplifier.

The loop gain of the non-inverting amplifier using the LM3886 as op amp (see section 3.2.3) is simulated in figure 3.3. Equation 3.12 directly relates the loop gain of a single op amp amplifier to the distortion reduction. The distortion reduction of the amplifier is therefore minimally 50 dB in the specified output region. The design discussed in this report aims on reducing the distortion measured at the output. Specification 2.26 specifies the expected improvement is at least 20 dB, which means the designed circuit will need a distortion reduction of at least 70 dB in the specified output region. This concludes the analysis of the non-inverting amplifier

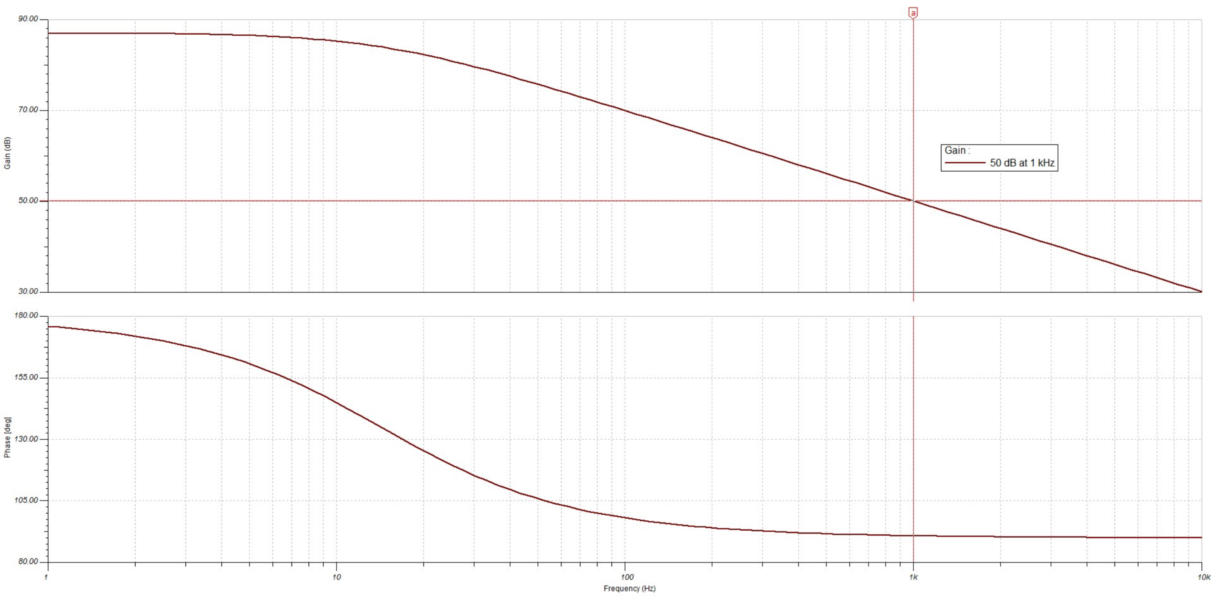


Figure 3.3: A simulation of the loop gain for the amplifier in the non-inverting configuration, as depicted in figure 3.1

3.2 Nested Feedback

As the loop gain is given by $L = A\beta$, it can be increased in two ways. Increasing the feedback gain β increases the loop gain, but does reduce the closed loop gain, which is inversely proportional to β. This solution can only be applied if the closed loop gain can be altered, which is not applicable in this design (*Specification 2.21*).

A more potent solution is found in increasing the open loop gain A. Since the closed loop gain is approximately independent of A, the loop gain can be increased without changing the closed loop gain. The proposed way of increasing A is shown in figure 3.4. The following sections will determine the characteristics of this topology and will find a solution to the problems that arise mainly in its stability.

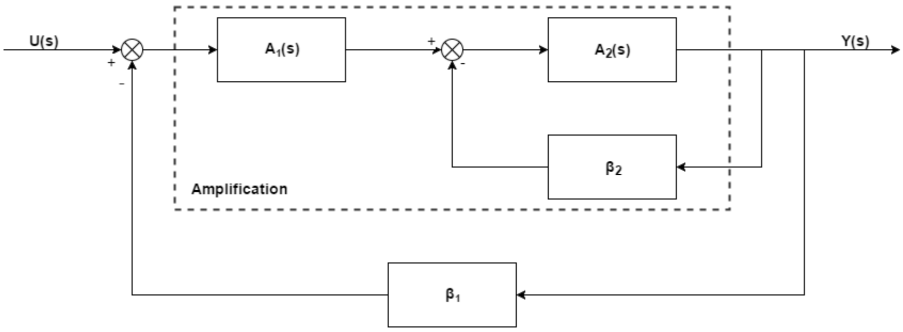


Figure 3.4: The proposed control scheme, where the open loop gain is increased compared to figure 3.2

3.2.1 Nested Feedback Distortion Analysis

The parameter A in this nested feedback configuration can be determined using the expression for the closed loop gain of A_2 and β_2 (equation 3.2):

$$A = \frac{A_1 A_2}{1 + A_2\beta_2} \tag{3.13}$$

The closed loop gain is then given by:

$$G = \frac{A_1 A_2}{1 + A_2\beta_2 + A_1 A_2} \tag{3.14}$$

The sensitivity of the output on both op amps is derived similarly to the case of the single op amp amplifier. For the primary op amp the sensitivity is given by equation 3.16.

$$\frac{\partial G}{\partial A_1} = \frac{A_2(1 + A_2\beta_2)}{(1 + A_2\beta_2 + A_1 A_2\beta_1)^2} \tag{3.15}$$

$$\frac{\delta G}{G} = \frac{1 + A_2\beta_2}{1 + A_2\beta_2 + A_1 A_2\beta_1} \frac{\delta A_1}{A_1} \tag{3.16}$$

Generally the open loop gains A_1 and A_2 are far larger than 1 and feedback gains β_1 and β_2 are smaller than 1. Equation 3.16 thus may be approximated to:

$$\frac{\delta G}{G} \approx \frac{A_2\beta_2}{A_1 A_2\beta_1} \frac{\delta A_1}{A_1} \tag{3.17}$$

Similarly for the secondary op amp the sensitivity is given by equation 3.19.

$$\frac{\partial G}{\partial A_2} = \frac{A_1}{(1 + A_2\beta_2 + A_1A_2\beta_1)^2} \tag{3.18}$$

$$\frac{\delta G}{G} = \frac{1}{1 + A_2\beta_2 + A_1A_2\beta_1}\frac{\delta A_2}{A_2} \tag{3.19}$$

Making the same assumptions as before 3.19 may be simplified to:

$$\frac{\delta G}{G} \approx \frac{1}{A_1A_2\beta_1}\frac{\delta A_2}{A_2} \tag{3.20}$$

The total expression for the distortion of the output can be found by summing both sensitivities in equations 3.17 and 3.20:

$$\frac{\delta G}{G} \approx \frac{\frac{\delta A_2}{A_2} + A_2\beta_2\frac{\delta A_1}{A_1}}{A_1A_2\beta_1} \tag{3.21}$$

Equation 3.21 gives the mathematical prove that the distortion of both op amps is further suppressed by the nested feedback amplifier. This is visible in the extra gain factor A_1 in the denominator. When the distortion given by $\frac{\delta G}{\delta A_1}$ is small, the nested feedback amplifier output is not more distorted then the non-inverting amplifier output. It is noted here that low distortion in the added amplifier is required to improve the distortion suppression. The secondary op amp drives the output and therefore needs to be an amplifier with good driving capabilities. These types of op amps have relatively large distortion specifications in the range of 0.02% The primary op amp does not need good driving capabilities and can be a high precision op amp. High precision op amps have low distortion specifications (typically in the range of 0.00002%). This low distortion characteristic allows for additional distortion reduction in the topology.

3.2.2 Closed Loop Gain

Figure 3.4 shows the circuit topology of the nested feedback amplifier.

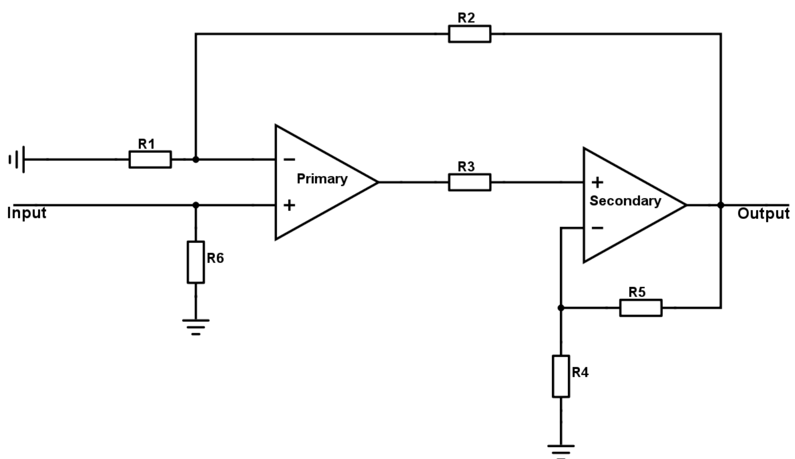


Figure 3.5: The topology of a nested feedback amplifier

Since the feedback in the control scheme has not been changed, the closed loop gain of the amplifier is still determined by the feedback gain β_1. For the specified gain of 28 dB (*Specification 2.21*), the following expression applies:

$$G = \frac{R_1 + R_2}{R_1} = 25 \tag{3.22}$$

$$R_2 = 24 \cdot R_1 \tag{3.23}$$

R_1 needs to be as large as possible to allow for a low cut-off frequency of the DC filter, given that large bipolar capacitors are not readily available (see section 3.2.4). Polar capacitors are larger but do not guarantee reliability, since the voltage in the branch may become negative. R_2 is paired with a capacitor to implement lead compensation. Because of the order of magnitude of this compensation network, R_2 is limited to about 100 $k\Omega$ (see section 3.2.3). Setting R_2 to the value of 100 $k\Omega$ and substituting in equation 3.23 yields a value of 4.2 $k\Omega$ for R_1. With these values the closed loop gain meets specification 2.21.

3.2.3 Nested Feedback Stability Analysis

Since the loop gain of the amplifier has been increased, the 0 dB point of the loop gain has moved to the right. As a result the phase margin has decreased, making the amplifier less stable. Furthermore the phase shift of the two op amps in the loop add up, thereby also decreasing the stability. These sources of instability indicate that an amplifier configured in a nested feedback configuration is not inherently stable and adjustment to ensure stability must be made.

Operational Amplifier Selection

Each pole of an op amp causes a phase shift of 90°. If the secondary poles of the two amps are spaced closely together, the phase shift will occur simultaneously. The resulting 180° of phase shift makes the system unstable. This situation can be prevented by selecting two op amps which secondary poles are spaced far apart (a factor 5 is generally sufficient [5]).

The secondary op amp delivers the power to the load and therefore needs to have good driving capabilities. For this op amp a standard audio amplifier can be used (which would have been used in the single op amp amplifier). The op amp selected for this design is the LM3886[1]. The LM3886 can deliver 50 W output power continuously which meets specification 2.31. It has low noise and distortion specifications for an audio op amp ($THD+N = 0.03\%$). Its gain bandwidth product is typically 8 MHz. Figure 3.6 shows the magnitude-phase characteristics of the LM3886. This figure shows that the op amp is only stable for gains larger than 19 dB (9x). This is the gain where the secondary pole is passed.

The primary op amp has to be selected based on the following criteria:

Gain-Bandwidth Product The gain-bandwidth product needs to be a factor 5 separated from that of the secondary op amp and therefore needs to be at least 40 MHz, to prevent simultaneous phase shift accumulation.

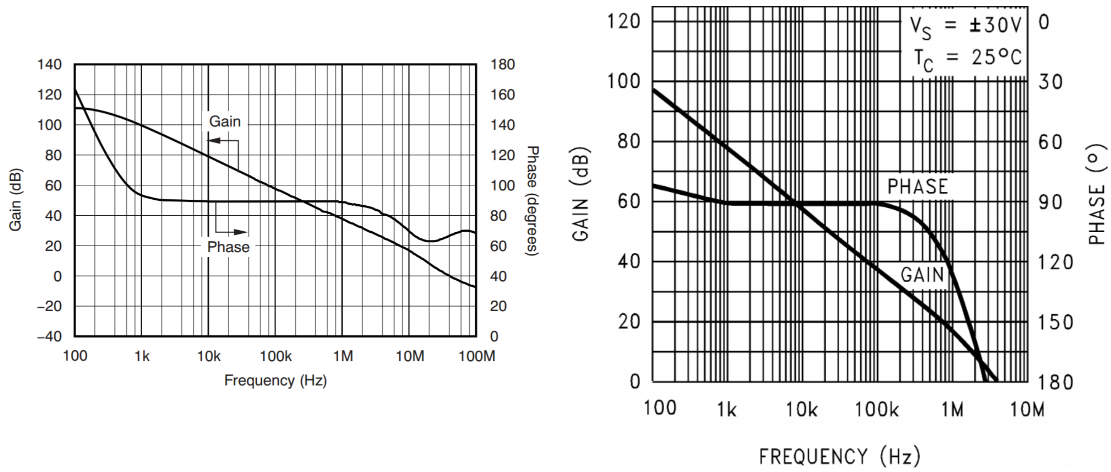
Low Distortion As discussed before in section 3.2.1, the primary op amp can be a high precision op amp and contributes significantly to the output distortion. The typical THD of this op amp should therefore be as low as possible.

Open Loop Voltage Gain Equation 3.21 shows that the distortion reduction of the circuit is proportional to the open loop voltage gain of the primary op amp. A high open loop voltage gain is desirable.

[1] http://www.ti.com/product/LM3886

The op amp selected for this application is the OPA1611[2]. The gain bandwidth product of this op amp is typically 40 MHz, which meets the given restriction. Furthermore it has a (THD + N) of 0.000015% and an open loop voltage gain of 130 dB.

Appendix B shows the most relevant specifications of all the op amps selected for this project. The magnitude-phase characteristics of both op amps are shown in figure 3.6.



(a) The magnitude-phase characteristic of the OPA1611 **(b)** The magnitude-phase characteristic of the LM3886

Figure 3.6: The magnitude-phase characteristics of both op amps. The secondary poles are spaced far apart.

Secondary Gain
The secondary gain must exceed 19 dB to stabilise the local closed loop. Equation 3.23 is rewritten for a gain of 19 dB as:

$$R_5 = 8 \cdot R_4 \tag{3.24}$$

Setting R_4 to 1 $k\Omega$ yields a resistance value of 8 $k\Omega$ for R_5.

Lead Compensation
Although selecting the right op amps is essential to stability, it does not guarantee that the amplifier is stable. As with single op amp amplifiers, the stability can be improved by applying frequency compensation. The first method applied is lead compensation [7, pp. 18-21]. This method compensates a pole present in the loop gain by placing a capacitor C_2 parallel to resistor R_2 in the feedback β_1 as is visualised in figure 3.7.
This parallel combination of C_2 and R_2 places a zero at:

$$f_z = \frac{1}{2\pi C_2 R_2} \tag{3.25}$$

It also introduces a pole combined with R_1 that is located at:

$$f_p = \frac{1}{2\pi C_2 (R_1 \parallel R_2)} \tag{3.26}$$

[2] http://www.ti.com/product/OPA1611

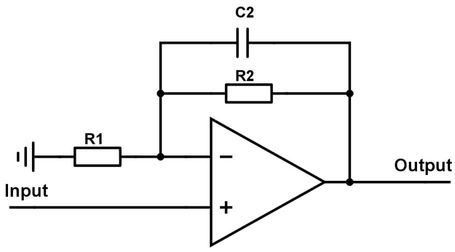


Figure 3.7: An amplifier where C_2 is added to implement lead compensation.

Lead compensation has effectively moved the compensated pole to a higher frequency. The loop gain without compensation is shown in figure 3.8. The system has a pole at a frequency of 878 kHz, which can be compensated by a capacitor of:

$$C_2 = \frac{1}{2\pi \cdot 100 \cdot 10^3 \cdot 878 \cdot 10^3} = 1.8\ pF \tag{3.27}$$

A simulation of the circuit with lead compensation is shown in figure 3.8. The pole at 878 kHz has been moved to a frequency of 4.86 MHz, thereby improving the phase behaviour of the circuit.

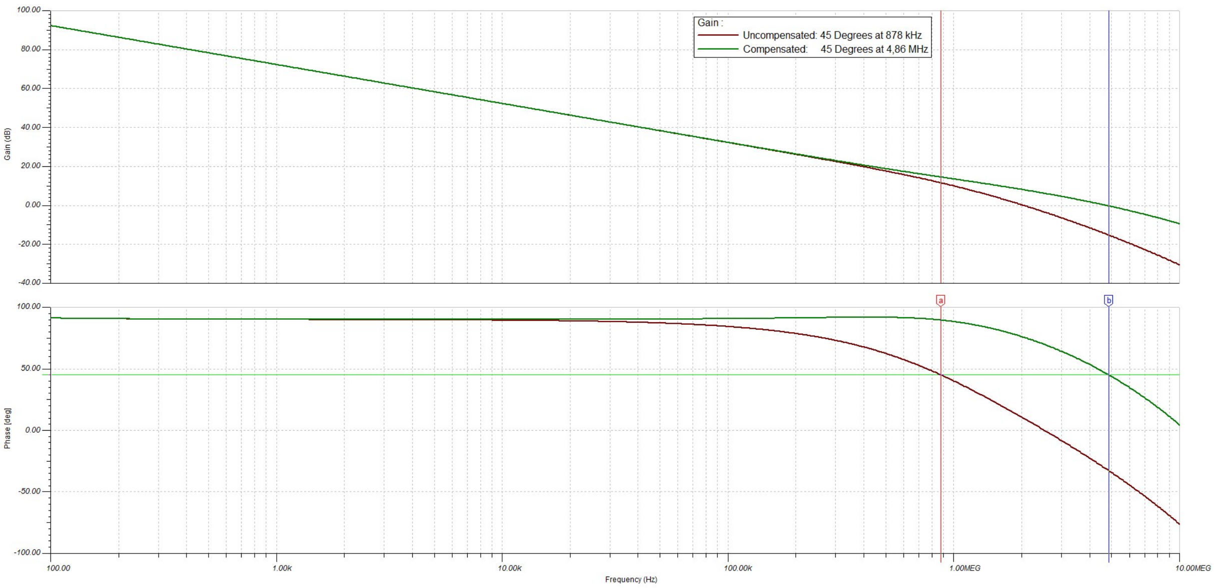


Figure 3.8: A simulation of the loop gain of the uncompensated and lead compensated circuits. The lead compensation improves the phase behaviour of the system.

Dominant Pole Compensation

Dominant pole compensation [7, pp. 15-17] is a type of frequency compensation that provides the possibility of trading loop gain for stability. It places a pole before the dominant pole of the primary op amp. The cutoff frequency of the gain effectively moves left towards the introduced pole, which increases the stability and decreases the loop gain. This method of compensation is invaluable since loop gain and stability are the two main specifications of the circuit design. The precise specifications for these quantities are:

Stability Specification 2.24 states the minimum phase margin at 45°.

Loop Gain As discussed in section 3.2 the loop gain must be as large as possible, since the distortion reduction will be improved.

Dominant pole compensation is generally done by placing a capacitor between the output or input of the amplifier and ground and is also called pole splitting, because the dominant pole of the op amp is moved to a lower frequency and the secondary pole is moved to a higher frequency [8]. However in order to place the pole low in the frequency spectrum a large capacitor value would be required. These capacitors are not readily available. A solution to this problem is provided by the miller effect. The miller effect states that a single capacitor C_1 in the feedback of an amplifier is equivalent with a capacitor $C_m = C_1(1+A)$ between the negative input of the op amp and ground and a capacitor $C_{m0} \approx C_1$ between the output of the op amp and the ground [8]. This is illustrated in figure 3.9.

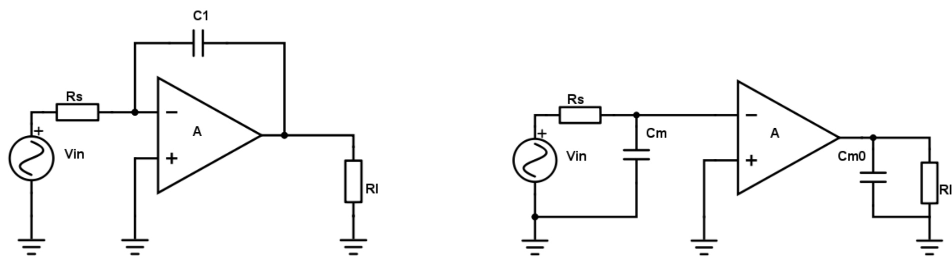


Figure 3.9: Two equivalent circuits. The miller effect greatly increases the capacitance at the input due to the capacitance between the input and output terminals of the amplifier.

The capacitance C_m is multiplied by the gain of the amplifier and because of that is much larger than its physical origin, capacitance C_1. The miller capacitance together with resistance R_1 in figure 3.5 forms a first order low-pass filter that places a pole at a frequency of:

$$f_m = \frac{1}{2\pi R_1 C_m} \tag{3.28}$$

Since the capacitor C_{m0} is small and forms a low-pass filter with the output resistance of the primary op amp, which is also small, its effects are much smaller than those of capacitor C_m and can be neglected.

The dominant pole compensation can decrease the loop gain in favour of stability. The phase margin should be increased until specification 2.24 is met. Figure 3.10 shows a simulation of the loop gain for various values of capacitance C_1.

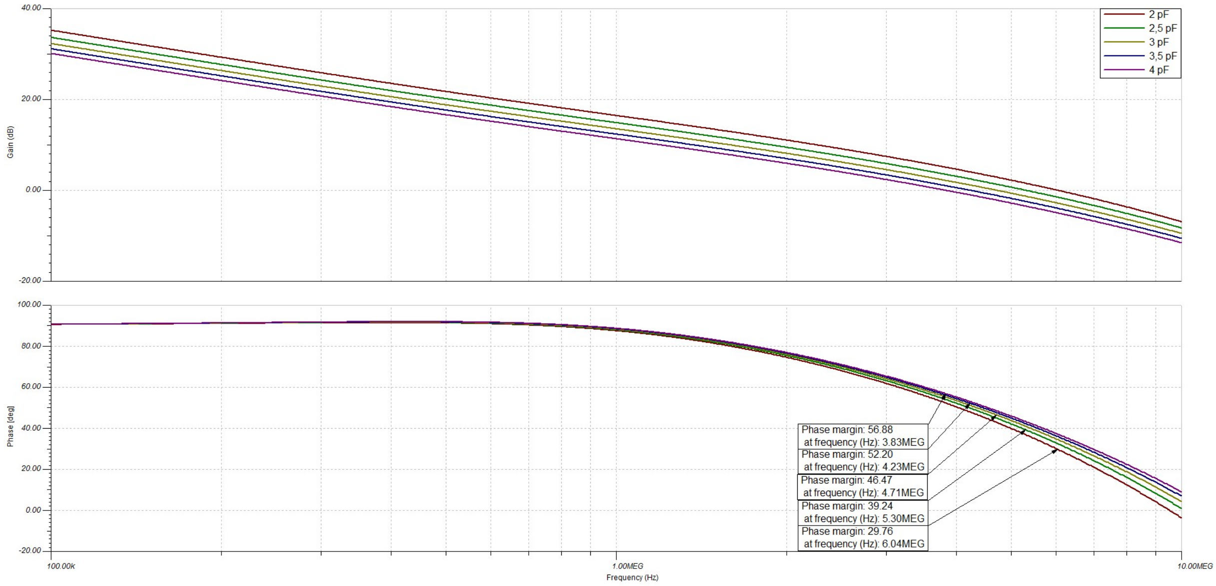


Figure 3.10: A simulation of the loop gain with dominant pole compensation for different values of C_1.

A value of 3 pF yields a phase margin of 46° which meets the specified stability criterion.

3.2.4 DC Behaviour

The bias current is the parasitic current that enters the op amp terminals at all times. This current produces an offset voltage at the input terminals due to the input resistances. The effect of this current can be removed by placing a resistor at the positive input terminal of the op amp that matches the parallel combination of the feedback network at the negative input terminal, so that the resulting voltage on both terminals is equal. This principle is applied to both op amps in the circuit shown in figure 3.5. R_3 needs to match the equivalent resistance of the feedback network, which is:

$$R_3 = R_4 || R_5 = 890\ \Omega \tag{3.29}$$

For the primary op amp, the same effect is accomplished by resistor R_6. The DC filters, discussed further on will block DC signals from the input and in the circuit branch of R_1. The bias current can only flow through resistors R_2 and R_6. R_6 needs to be matched to R_2 and therefore has a value of 100 $k\Omega$.

The two resistor introduced mitigate the effects of the bias current, but do not eliminate the effect of the bias offset current or the offset voltage. These two effects will be present in the final circuit.

The closed loop gain of the system can be approximated by:

$$G = \frac{1}{\beta_1} \tag{3.30}$$

Placing a DC filter in the circuit branch with R_1 ensures unit gain for DC signals, since β_1 is set to 1. Placing another DC filter at the input, the output DC voltage becomes equal to the parasitic effects described above on the primary op amp. The bias offset current of the OPA1611 is typically ± 25 nA. This current runs through a 100 $k\Omega$ resistor and thus causes an offset voltage at the input terminals of:

$$V_{ios} = 100 \cdot 10^3 \cdot 25 \cdot 10^{-9} = 2.5\ mV \tag{3.31}$$

The typical offset voltage of the OPA1611 is specified as $\pm 100\,\mu V$. The maximal total offset voltage on the input will be:

$$V_{off} = 2.5 + 0.1 = 2.6\,mV \tag{3.32}$$

This will also be the DC output voltage and thereby meets specification 2.36.

The filters discussed above are implemented as a first order high pass filter with a low cutoff frequency. The highest bipolar capacitor values are in the order of $10\,\mu F$. Combined with resistances of $4.2\,k\Omega$ and $100\,k\Omega$ these result in cutoff frequency of respectively $3.8\,Hz$ and $0.16\,Hz$. The cutoff frequency of $3.8\,Hz$ is not removed far from the lowest output frequency of $10\,Hz$ and causes a phase shift of nearly 21° which falls within the error specified in specification 2.23.

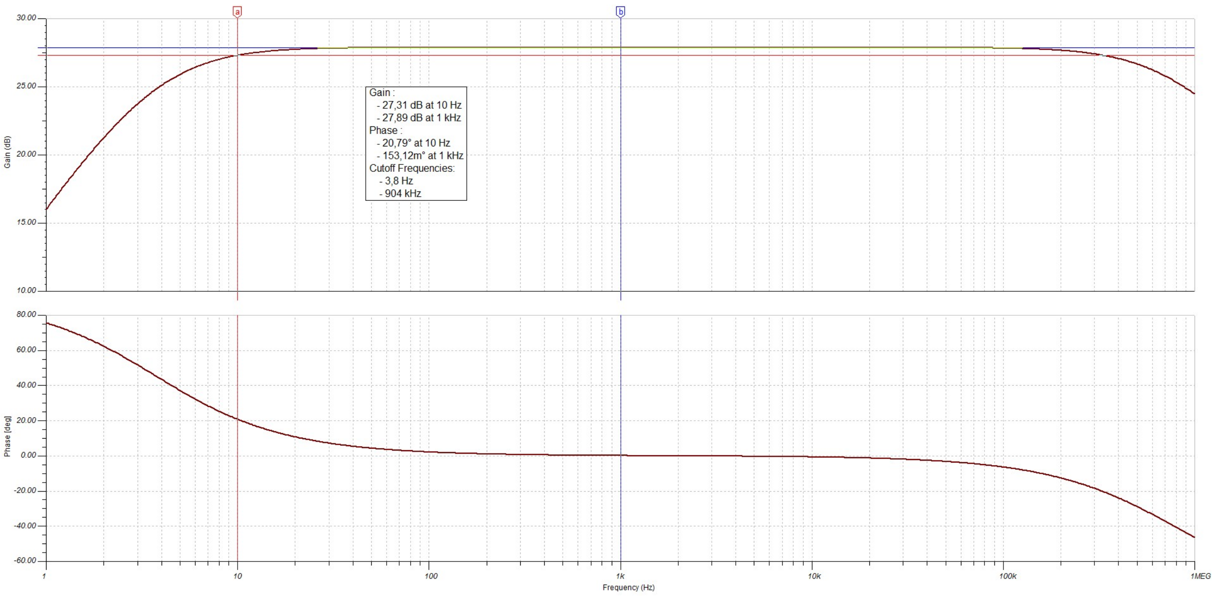


Figure 3.11: A simulation of the closed loop transfer after adding the DC filters to the circuit. The simulation meets specifications 2.22 and 2.23.

3.2.5 Source Voltages

The maximum continuous output power of the LM3886 is $50\,W$ for an impedance of $8\,\Omega$. This means the maximum rms output voltage of the circuit is:

$$V_{out2,rms} = \sqrt{50 \cdot 8} = 20\,V \tag{3.33}$$

The amplitude of the output voltage will then be:

$$V_{out2,A} = 20 \cdot \sqrt{2} = 28\,V \tag{3.34}$$

Incorporating a small margin, the source voltage of the secondary op amp is set to $\pm 30\,V$.

The closed loop gain of the secondary amplification is $9x$. Which means the voltage on the output terminals of the primary op amp must be 9 times smaller than the output voltage of the circuit, which is:

$$V_{out1,A} = \frac{28}{9} = 3.1\,V \tag{3.35}$$

The op amps used in the filters discussed in section 3.3 use a source voltage of 9 V. This voltage is within the source voltage range of the OPA1611. For convenience this voltage is also used for the primary op amp.

3.2.6 Final Circuit

The final circuit designed in the previous section is shown in figure 3.12.

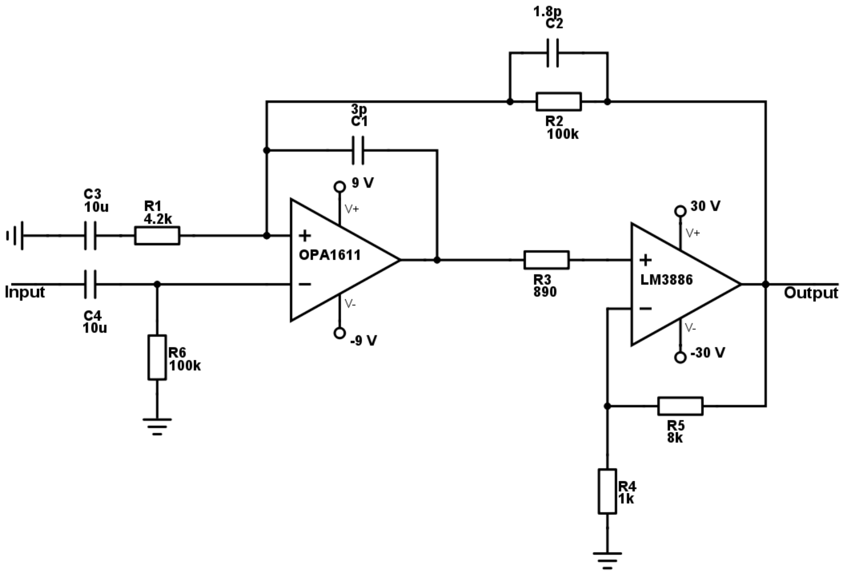


Figure 3.12: The fully dimensioned nested feedback amplifier

3.2.7 Simulation

TINA Spice software[3] is used to simulate the behaviour of the circuit depicted in figure 3.12. The closed loop simulation circuit is depicted in figure 3.13. The result of the transfer simulation was shown in figure 3.11. The TINA Spice software is capable of plotting the SNR as a function of the bandwidth. The resulting simulation is shown in figure 3.14. Specification 2.35 states that the SNR must be at least 80 dB within the bandwidth of the motional feedback. The simulation meets this specification.

[3] http://www.ti.com/tool/tina-ti

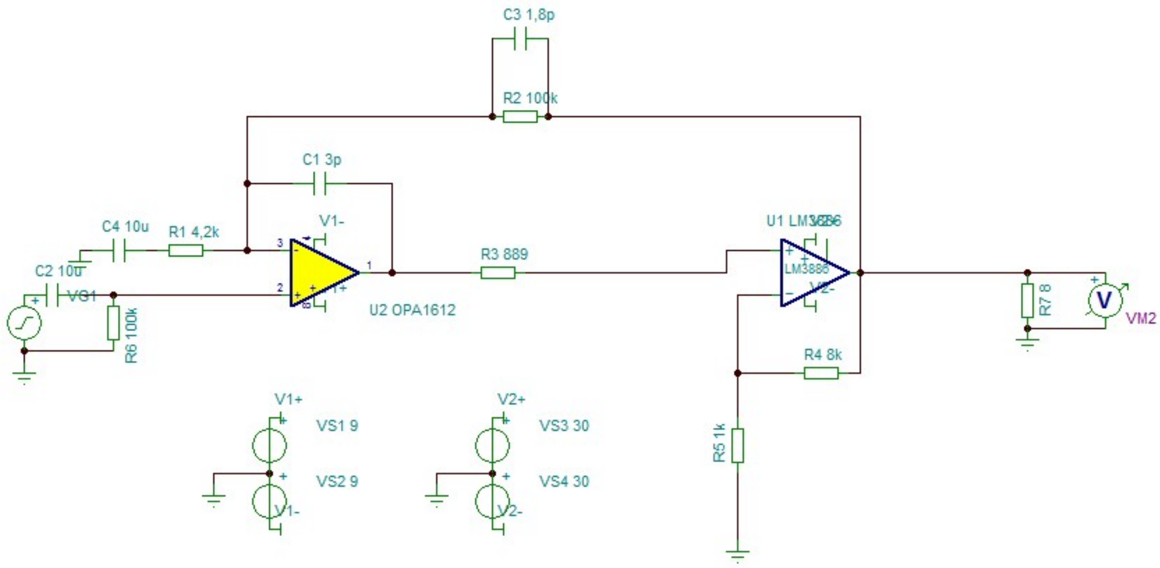


Figure 3.13: The closed loop simulation circuit.

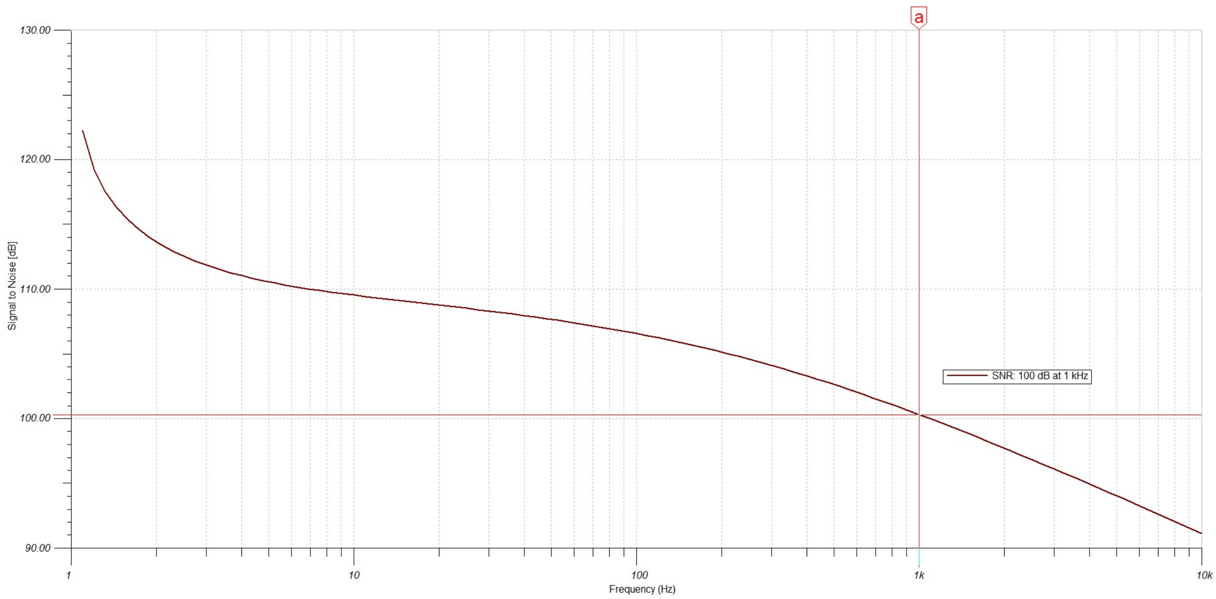


Figure 3.14: The SNR of the nested feedback amplifier.

3.3 4th Order Cross-over Filter

Loudspeaker drivers are not made to operate over the entire audible frequency spectrum. Therefore, a speaker usually consist of more then one driver. Typically, a speaker produces frequencies from 20 Hz to 20.000 Hz. A large speaker cone is designed to emit low frequencies while a smaller cone emits higher frequencies. Cross-over filters separate the spectrum into a matched bandwidth for each driver. Mainly, there are two types of these filters, passive and active.

3.3.1 Passive Filters

Passive filter implementations are the simplest implementation of a given transfer function. Because the components are passive, no power supply is needed. Passive components are not restricted to frequency limitations as much as the active filters. The passive components can handle larger voltages and currents than active components. Also the noise performance of these networks is in general sufficient, only thermal noise is produced. Passive filters do not use any power supply, so they can not produce gain in a circuit.

3.3.2 Active Filters

Active filters use amplifying elements, with passive elements in the feedback loop. Active elements, like an op amp, have a high input impedance and a low output impedance. Active filters are easier to design, also because of the lack of inductors. Op amps are restricted in bandwidth, so higher frequency filters are more difficult to design. Amplifying circuits produce noise. A solution is to choose for low-noise amplifiers. Low-noise amplifiers produce less noise than regular op amps. [12]

3.3.3 Cross-over Design

The design of the cross-over filter is simpler when using an active network. The required power supply is not a limitation, because the amplifier of the woofer needs power. Also, the bandwidth of op amps is large enough for the audible frequencies. Because the input signal is smaller than 4 volt peak-to-peak, there is no need to use high voltage passive components. So, the cross-over will be an active circuit. The noise produced by the circuit will be reduced by a low-noise amplifier.

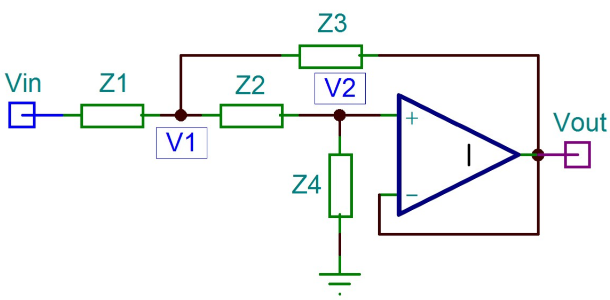


Figure 3.15: The Sallen-Key Topology

A simple second order active filter, is a Sallen-Key filter, which is shown in figure 3.15. With this topology both a low-pass and a high-pass filter can be designed. A combination of a low-pass and a high-pass filter gives a cross-over filter. The + and - inputs of the operational amplifier have to match. Because of the negative feedback, this results into:

$$V_2 = V_{-amp} = V_{out} \tag{3.36}$$

Applying Kirchhoff's current and voltages laws, gives:

$$\frac{V_{in} - V_1}{Z_1} = \frac{V_1 - V_{out}}{Z_3} + \frac{V_1 - V_{out}}{Z_2} \tag{3.37}$$

$$\frac{V_1 - V_2}{Z_2} = \frac{V_2}{Z_4} \tag{3.38}$$

The last equation can be written as:

$$V_1 = V_{out}\left(1 + \frac{Z_2}{Z_4}\right) \tag{3.39}$$

Combining equation 3.37 and 3.39 gives:

$$\frac{V_{in} - V_{out}(1 + \frac{Z_2}{Z_4})}{Z_1} = \frac{V_{out}(1 + \frac{Z_2}{Z_4}) - V_{out}}{Z_3} + \frac{V_{out}(1 + \frac{Z_2}{Z_4}) - V_{out}}{Z_2} \tag{3.40}$$

This equation can be written as:

$$\frac{V_{out}}{V_{in}} = \frac{Z_3 Z_4}{Z_1 Z_3 + Z_2(Z_1 + Z_3) + Z_3 Z_4} \tag{3.41}$$

For a second order low-pass filter, the transfer function in the Laplace domain is given:

$$H(s)_{LP2} = \frac{\omega_0^2}{s^2 + 2\alpha s + \omega_0^2} \tag{3.42}$$

With:

$$s = j\omega \tag{3.43}$$

To create a low-pass filter, Z_1 and Z_2 are resistors and Z_3 and Z_4 are capacitors. The transfer function 3.41 can be transformed to the Laplace domain:

$$H(s)_{LP2} = \frac{\frac{1}{s^2 C_3 C_4}}{\frac{R_1}{sC_3} + R_2(R_1 + \frac{1}{sC_3}) + \frac{1}{s^2 C_3 C_4}} = \frac{\frac{1}{R_1 R_2 C_3 C_4}}{s^2 + s\frac{1}{C_3}(\frac{R_1 + R_2}{R_1 R_2}) + \frac{1}{R_1 R_2 C_3 C_4}} \tag{3.44}$$

Thus:

$$\omega_0 = 2\pi f_0 = \frac{1}{\sqrt{R_1 R_2 C_3 C_4}} \tag{3.45}$$

$$2\alpha = \frac{1}{C_3}\left(\frac{R_1 + R_2}{R_1 R_2}\right) \tag{3.46}$$

$$Q = \frac{\omega_0}{2\alpha} = \frac{\frac{1}{\sqrt{R_1 R_2 C_3 C_4}}}{\frac{1}{C_3}(\frac{R_1 + R_2}{R_1 R_2})} \tag{3.47}$$

Vice versa, for an high-pass filter, Z_1 and Z_2 are capacitors and Z_3 and Z_4 are resistors. The transfer function of a second order high-pass filter is given by:

$$H(s)_{HP2} = \frac{s^2\omega_0^2}{s^2+2\alpha s+\omega_0^2} = \frac{s^2\frac{1}{R_1R_2C_3C_4}}{s^2+s\frac{C_3C_4}{R_2C_3C_4}+\frac{1}{R_1R_2C_3C_4}} \tag{3.48}$$

There are some issues with a single Sallen-Key filter topology. The problem with these second order filters is the roll-off, which is only 40 dB per decade or 12 dB per octave. The accelerometer in the woofer has a resonance peak at 1.5 kHz (see section 3.5). So the amplification at 1.5 kHz has to be low enough to keep the resonance peak under control (specification 2.34). These second order filters are easily realised with only passive components. Furthermore, a steeper roll-off is useful, in the order of 80 dB per decade or 24 dB per octave. An advantage of the Sallen-Key filter is the property to cascade single filters to get a higher order filter network. This work is done by S.H. Linkwitz. The cascade of n Sallan-Key filter gives a n^{th} order filter. Each second order filter has a 180 phase shift, because of the two capacitors per op amp.

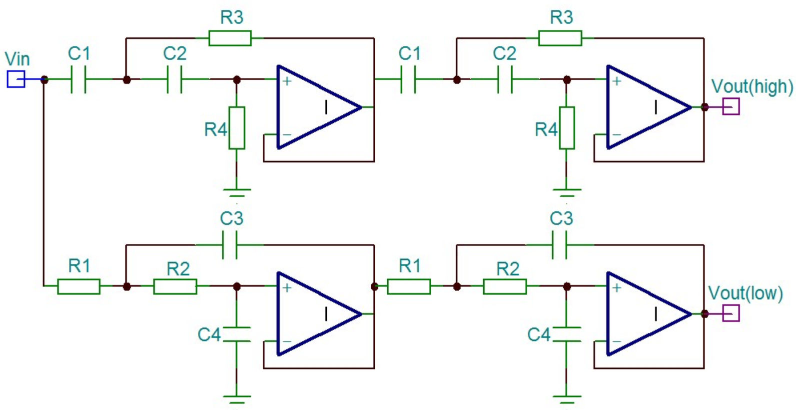


Figure 3.16: Linkwitz-Riley cross-over Topology [10]

The transfer function of these fourth order low-pass filter is the multiplication of two second order transfer functions:

$$H(s)_{LP4} = \frac{\omega_0^2}{s^2+2\alpha s+\omega_0^2}\cdot\frac{\omega_0^2}{s^2+2\alpha s+\omega_0^2} = \frac{\omega_0^4}{(s^2+2\alpha s+\omega_0^2)^2} \tag{3.49}$$

For a fourth order high-pass filter, the transfer function is:

$$H(s)_{LP4} = \frac{s^2\omega_0^2}{s^2+2\alpha s+\omega_0^2}\cdot\frac{s^2\omega_0^2}{s^2+2\alpha s+\omega_0^2} = \frac{s^4\omega_0^4}{(s^2+2\alpha s+\omega_0^2)^2} \tag{3.50}$$

The Linkwitz Riley cross-over is shown in figure 3.16. At this stage, the filters need to be dimensioned. The op amps have to match with the resistor and capacitor values.

Operational Amplifier Selection

The primary op amp has to be selected based on the following criteria:

Low Distortion The op amps should not add significant distortion to the circuit, because the good distortion characteristics of the topology would decrease.

Low Voltage Noise To meet specification 2.35 of a SNR of 80 dB, the noise produced by the op amps has to be as low as possible.

Output voltage offset Because of the DC-filter if front of the audio amplifier, it is allowed to have a low output voltage offset. Low enough, to prevent clipping of the op amp.

Number of Channels This cross-over filter uses four op amps. It is cheaper to use multiple channel op amp instead of four single channel op amps.

The op amp selected for this application is the LM833NG[4]. This two channel op amp has a low distortion of 0.002 % and a low noise voltage of $4,5\ \frac{nV}{\sqrt{Hz}}$.

Resistors and Capacitors Selection

The resistors and capacitors has to be selected based on the following criteria:

Cross-over Point The cross-over point is limited by the bandwidth of the output. (specification 2.32)

Flat Transfer Function To get a proper acoustic flat transfer function, the addition of the two filter transfer functions has to be flat.

Phase It is desired to have linear phase from 10 Hz to 1 kHz. (specification 2.33)

Tolerance The tolerance of resistors must be 1 %, and 10 % for the capacitors, due to the sensitivity of the circuit.

The bandwidth of the motional feedback is limited to 1 kHz (*specification 1.3.3.*). This is the bandwidth where the bass driver will be used. The cross-over point thus also needs to be placed at 1 kHz. The capacitor values have to be between the 2.2 nF and 470 nF, to avoid loading of the op amp. The resistor values have to be between the 2.2 $k\Omega$ and 22 $k\Omega$, to minimize noise. Lower values can cause problems with the input impedance of the op amp. Higher values can cause current problems. The op amp will not work when the currents are too low.

A flat transfer function indicates that the addition of the two filters has to be 1:

$$H(s)_{LR4} = H(s)_{LP4} + H(s)_{HP4} = \frac{\omega_0^4}{(s^2+2\alpha s+\omega_0^2)^2} + \frac{s^4\omega_0^4}{(s^2+2\alpha s+\omega_0^2)^2} = 1 \tag{3.51}$$

To get a flat transfer function at the cross-over point, the value of R4 has to be twice as the value of R3 at the high-pass. The value of C3 has to be twice of the value of C4 at the low-pass. The values of R1 and R2 have to be equal and lastly the values of C1 and C2 have to be equal. [10] The cutoff frequency becomes:

$$f_0 = \frac{1}{2\pi\sqrt{R_1 R_1 2C_4 C_4}} = \frac{1}{2\sqrt{2}\pi R_1 C_4} \tag{3.52}$$

First, the values of capacitors are chosen. C4 becomes a 10 nF capacitor, which is between the 2.2 nF and 470 nF. R1 can be calculated with equation 3.52:

$$R_1 = R_2 = \frac{1}{2\sqrt{2}\pi f_0 C_4} = \frac{1}{2\sqrt{2}\pi \cdot 1000 \cdot 10 \cdot 10^{-9}} = 11.3\ k\Omega \tag{3.53}$$

Which is between the the $2.2k\ \Omega$ and $22k\ \Omega$. C3 becomes 22 nF, because 20 nF is not available. At the high-pass filter: C1 and C2 are 10 nF, R3 is 11.3 $k\Omega$ and R4 is twice R3, so 23.2 $k\Omega$, because $22,6k\Omega$ is not available. The circuit is shown in figure 3.17.

[4] http://www.farnell.com/datasheets/81893.pdf

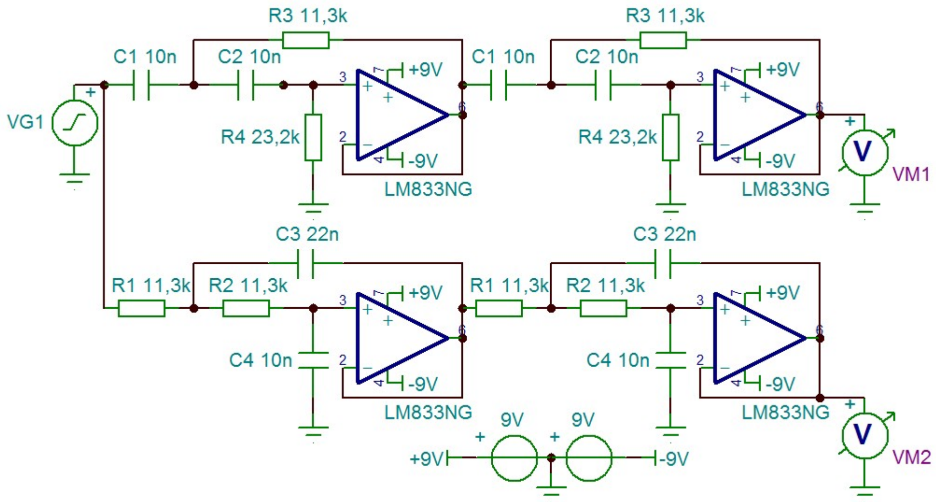


Figure 3.17: Linkwitz-Riley cross-over Circuit

At least, the quality factor of each stage becomes:

$$Q_{LR4} = \frac{\frac{1}{\sqrt{R_1 R_2 C_3 C_4}}}{\frac{1}{C_3}\left(\frac{R_1+R_2}{R_1 R_2}\right)} = \frac{\frac{1}{\sqrt{11.3\cdot 10^3\cdot 11.3\cdot 10^3\cdot 10\cdot 10^{-9}\cdot 22\cdot 10^{-9}}}}{\frac{1}{10\cdot 10^{-9}}\left(\frac{11.3\cdot 10^3+11.3\cdot 10^3}{11.3\cdot 10^3\cdot 11.3\cdot 10^3}\right)} = 0.34 \tag{3.54}$$

Simulation

The TINA Spice software includes an AC analysis simulation to test the circuit. First, the AC analysis gives the the transfer function of both filters. This is shown in figure 3.18. An addition of both filters gives the total transfer function. The total transfer is flat as specified, with a small peak at 656.4 Hz of 387.6 mdB. This is caused by the value of the resistors and capacitors, which are not available. It is not necessary to remove this peak, since it is audible. The amplification at 1.5 kHz is -16.6 dB, which meets specification 2.34.
Both filters have a phase shift of 360°, due to the 4 capacitors per filter. Linkwitz mentioned in the paper that the delay is not audible [10]. The phase of these filter are equal till 20 kHz. After the 20 kHz, the phase of the low-pass filter drops. This is caused by the non ideal characteristics of the op amp. [11] The phase till 1 kHz is linear, which meets the specifications. The delay analysis is shown in figure 3.20. There is a peak of 586.1 us at 652.8 Hz in the low-pass filter, which also is not audible.
At last, a DC analysis is required. DC voltage is not desired, because it can destroy the woofer.
There is an issue with this cross-over filter topology.

Bias Current Every op amp has a bias current. This bias current creates an offset at the input terminals of the op amp. With the low-pass Sallen-Key filter this parasitic current causes problems. The bias current can not flow from the minus terminal through C3 or C4. Therefore, the op amp will not operate correctly and will clip to the positive supply voltage. To correct this, a resistor (R5) is placed parallel to C3. A large resistor will cause a large offset voltage at the output of the op amp. A small resistor will adapt the filter properties.

The effect of resistor R5 can be simulated, which is shown in figure 3.21. A low value of resistor R5 attenuates the input signal. So a high value resistor is suitable, which causes an offset voltage at the output of the op amp. The

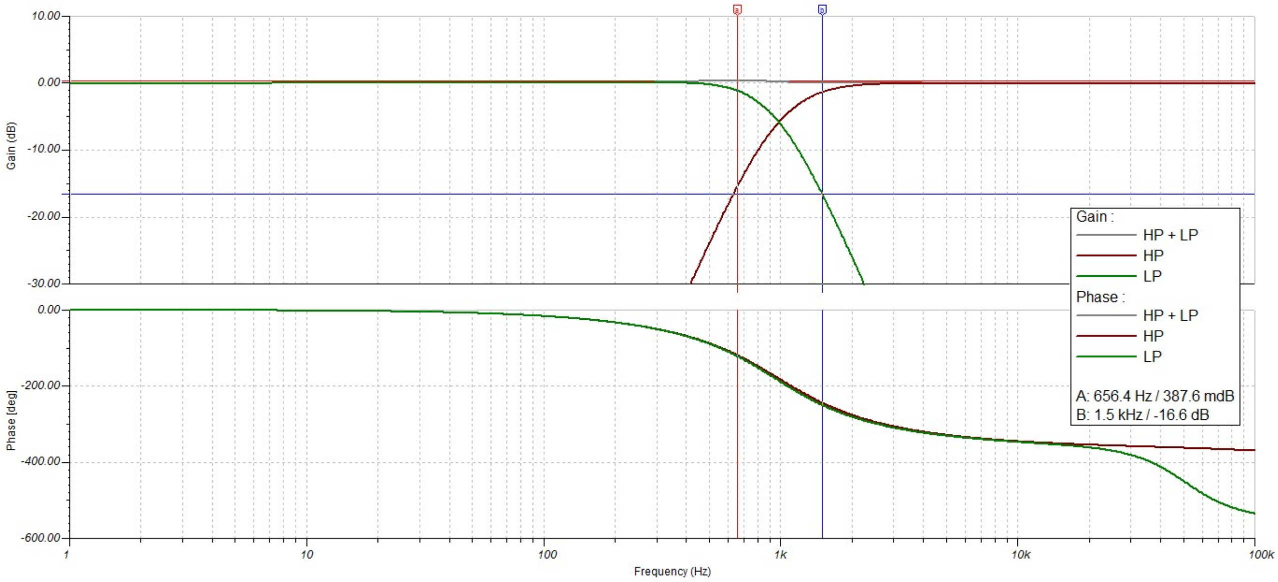


Figure 3.18: AC Analysis of the Linkwitz-Riley Cross-over Circuit

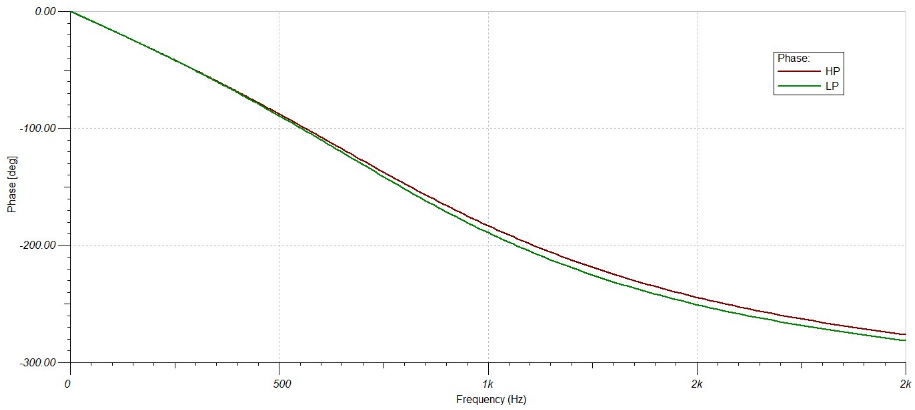


Figure 3.19: Phase Analysis of the Linkwitz-Riley cross-over Circuit

typical bias current of an LM833 op amp is 300 nA. This can create an offset voltage, if the current flow through resistor R5, of:

$$V = I \cdot R = 300 \cdot 10^{-9} \cdot 1 \cdot 10^{6} = 0.3\,V \tag{3.55}$$

3.3.4 Final Circuit

The final circuit, with the last adjustments, is shown in figure 3.22. The AC analysis of this circuit is shown in figure 3.23. The small peak is moved to 638.8 Hz and decreased to 325.3 mdB, which still is not audible. The attenuation at 1.5 kHz is lowered to 16.7 dB. The noise-bandwidth performance is shown in figure 3.24. The lowest signal-to-noise ratio is 115.8 dB. The total signal to noise ratio of the total system must be at least 80 dB

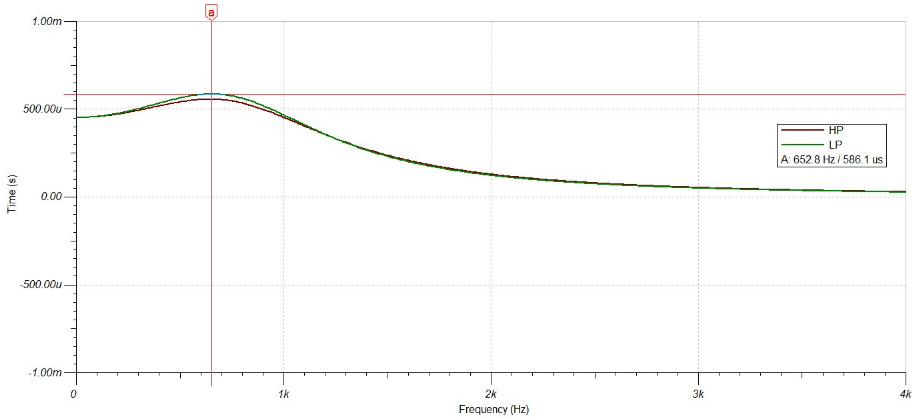


Figure 3.20: Delay Analysis of the Linkwitz-Riley cross-over Circuit

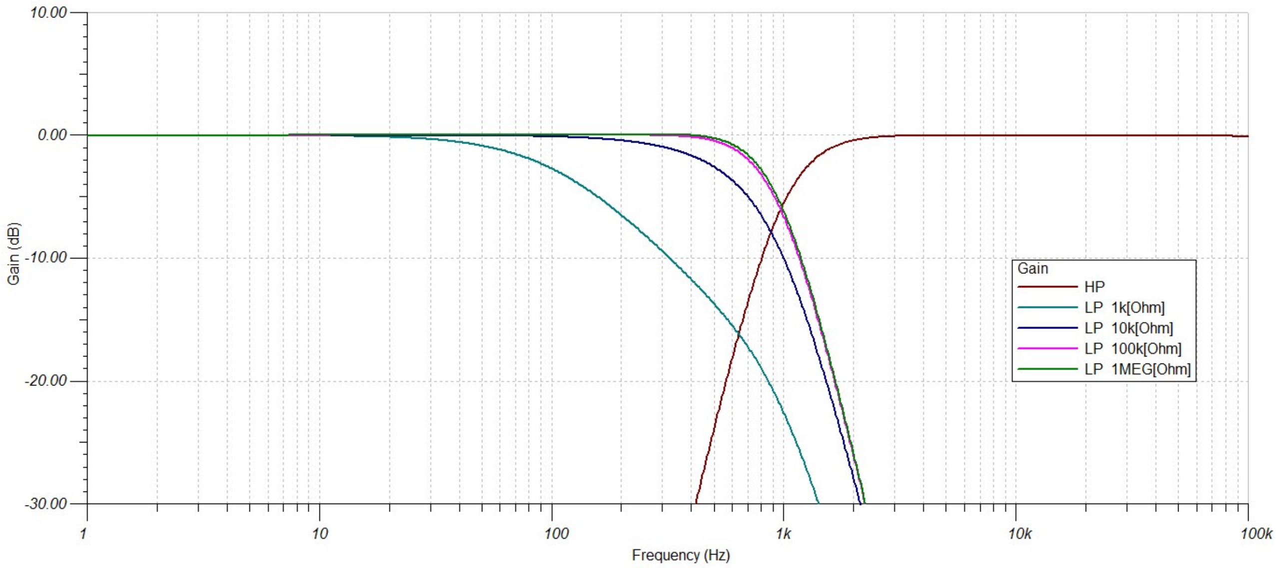


Figure 3.21: Effect of adding Resistor R5

according to specification 2.35. So this circuit is not a limitation to the specification.

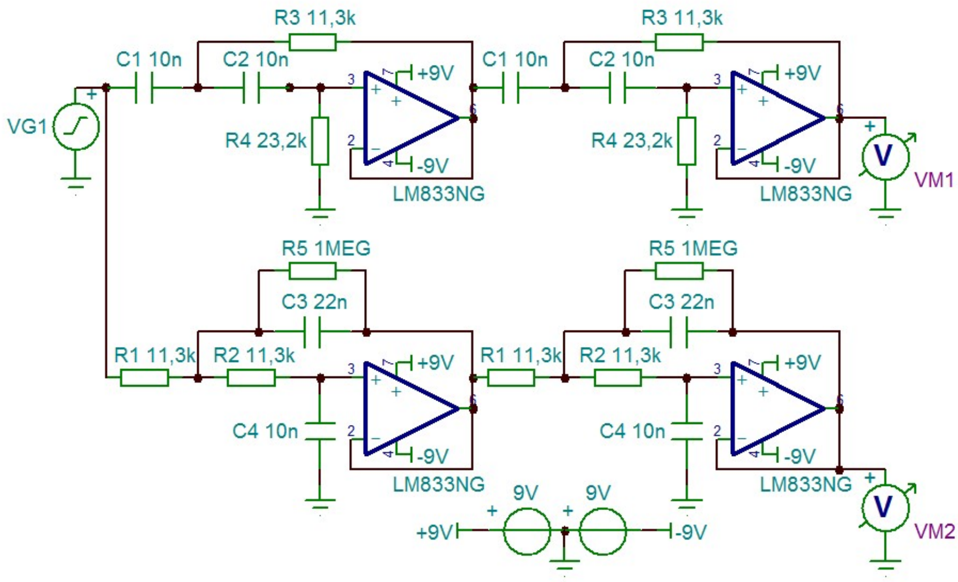


Figure 3.22: Final Linkwitz-Riley Cross-over Circuit

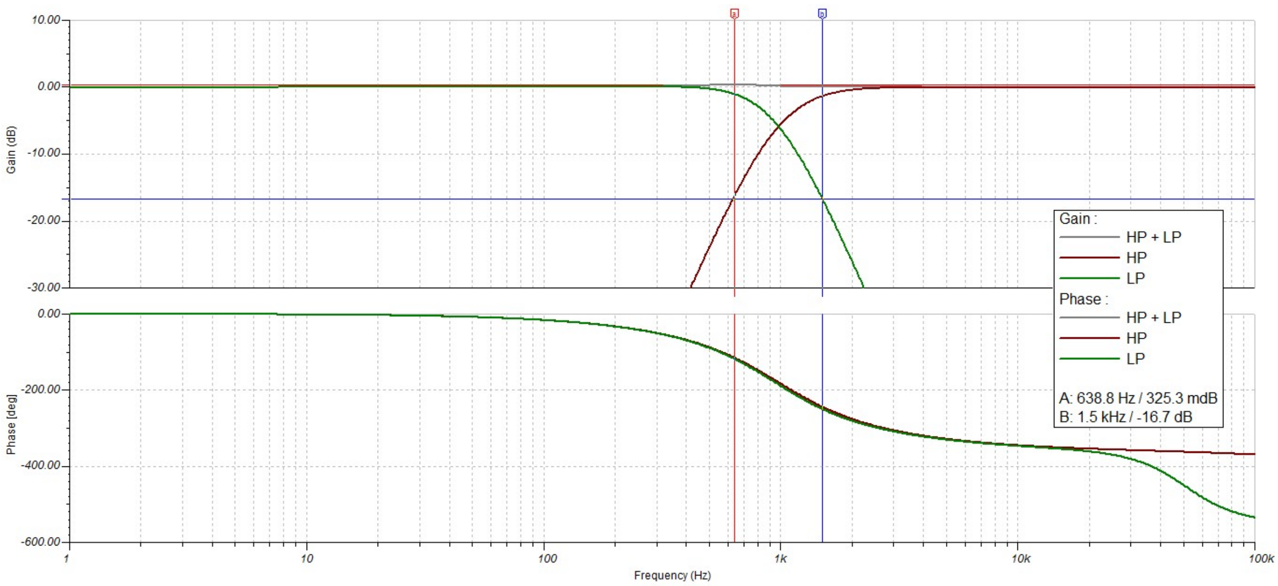


Figure 3.23: AC analysis of the Final Linkwitz-Riley cross-over Circuit

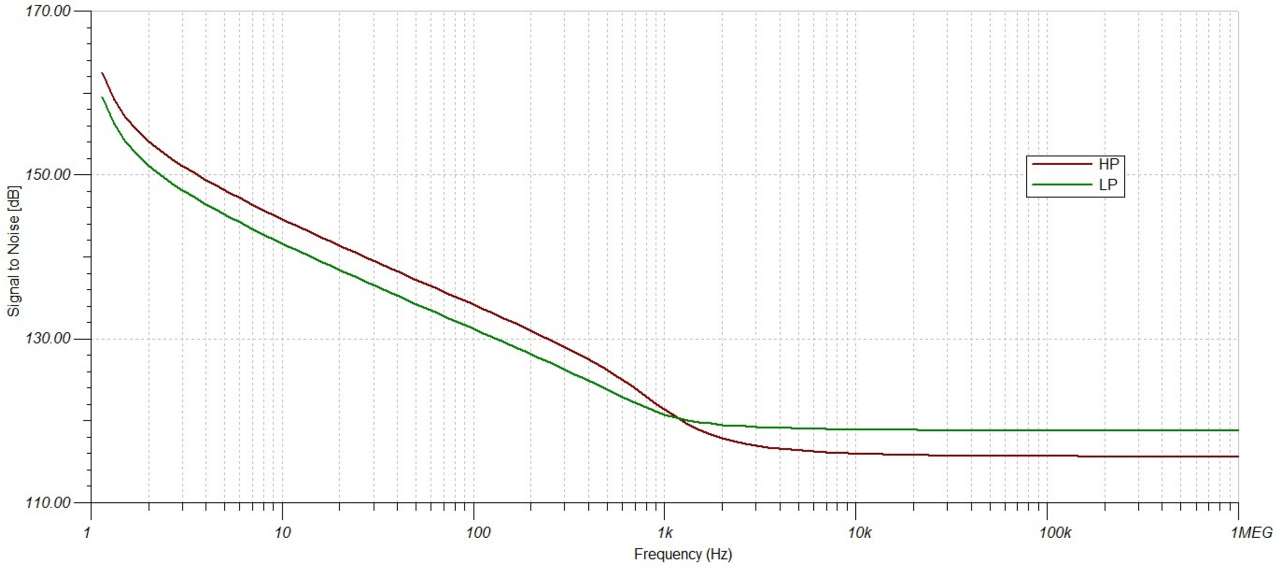


Figure 3.24: Noise-Bandwidth Performance of the Final Linkwitz-Riley cross-over Circuit

3.4 Subtracting Amplifier

As the audio system designed in this report is developed for the motional feedback application, the system needs a method of subtracting the feedback input from the audio input signal. This is accomplished using a subtracting amplifier. The topology of which is shown in figure 3.25.

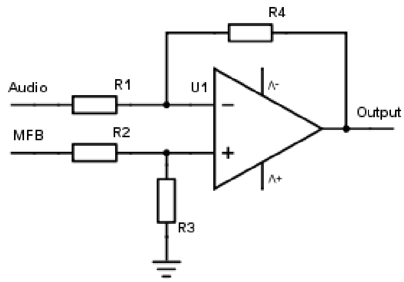


Figure 3.25: The topology of a subtracting amplifier.

This subtracting amplifier can also be used to damp the input signal to the maximal input value of the amplifier. According to specification 2.11 and 2.12 both maximal input voltages are 4 V peak-to-peak. Since in the worst case they are 180° out of phase and maximal magnitude, the maximal peak-to-peak output voltage of the subtracting amplifier with unity gain is 8 V. The audio amplifier has a maximal voltage amplitude at the output of 28 V. Because it amplifies 28 dB (*specification 2.21*), the maximal input voltage amplitude must be:

$$V_{in,A} = \frac{28}{25} = 1.12\,V \tag{3.56}$$

Which is equivalent to 2.24 V peak-to-peak. The gain of the subtracting amplifier needs to be:

$$g = \frac{2.24}{8} = 0.28 \tag{3.57}$$

Choosing $R_1 = R_2$ and $R_3 = R_4$ the gain is described as:

$$g = \frac{R_3}{R_1} = 0.28 \tag{3.58}$$

Setting R_1 to 10 $k\Omega$ yields a value of 2.8 $k\Omega$ for R_3. The output signal of the subtracting amplifier is described as:

$$V_o = 0.28 \cdot (V_{audio} - V_{mfb}) \tag{3.59}$$

Which implements the subtraction required for the implementation of motional feedback.

3.5 Displacement Sensor

To implement motional feedback, a signal that represents the displacement of the cone is required. In order to select the right sensor type a physical analysis [13, pp. 8,9] of the speaker is required.

3.5.1 Speaker Physics

The acceleration of the speaker piston can be described by:

$$a(t) = -\omega\hat{v}\sin\omega t \tag{3.60}$$

Where $\hat{v}$ is the velocity of the piston. Converting to RMS values, this equation can be rewritten to:

$$v_{rms} = \frac{a_{rms}}{\omega} \tag{3.61}$$

The sound pressure received on a distance d, emitted by the speaker is given by:

$$p_{rms} = \frac{\rho A_\phi v_{rms} f}{2d} \tag{3.62}$$

In this equation ρ represents the density of air and A_ϕ is the area of the speaker. Substituting equation 3.61 in equation 3.62 yields:

$$\frac{\rho A_\phi a_{rms}}{4\pi d} \tag{3.63}$$

This equation shows that the air pressure emitted is dual with the acceleration of the speaker. Relating both the current and the acceleration of the speaker to the force applied to the speaker, it can be shown that these quantities are also dual. The magnitude of the force on the speaker is given by:

$$F = ma \qquad F = l \cdot I \cdot B \tag{3.64}$$

3.5.2 Sensor Selection

The result of the equations derived in the previous section is that the voltage on the voice coil (which according to Ohms law is dual with the current) is dual with acceleration of the speaker and the sound pressure emitted. The three sensors that measure these quantities directly are the accelerometer, the microphone and the voltage sensor. Furthermore the displacement or the velocity of the speaker could be measured, but would need a static reference point, since these quantities are not directly dual to the quantity of interest. All possible measurement methods are shown below:

- Voltage Sensor
- Microphone
- Accelerometer
- Speedometer
- Displacement Sensor

These measurement methods are also discussed in the original paper of Philips [1], where the following remarks are made. The voltage sensor measures the voltage on the voice coil itself and subtracts from it the voltage supplied to receive the induced voltage. The problem with this approach lies in the dependency on temperature and the inhomogeneity of the magnetic field, both of which negatively impact the accuracy of the received data. The microphone has the major drawback that the signal received depends on the channel through which the output propagates. Because of this the received signal will not accurately represent the displacement of the speaker.

The latter two options share the same problem, which is that a physical static reference point is required to give meaning to the output. Due to the vibrations in the mechanical placing of the sensor a static reference point is hard to obtain. The accelerometer does not suffer from these problems and is therefore the preferred method of measurement.

3.5.3 Accelerometer Selection

The two main types of accelerometer are:

- The MEMS accelerometer
- The piëzo-electric accelerometer

The first device is cheaper and mechanically more compact than the second device. The main drawback of this sensor type is its frequency behaviour. Most MEMS devices have a low pass characteristic with a cutoff frequency around 400 Hz. This bandwidth is insufficient for the application of this project, since the specified frequency range is 10 $Hz - 1\ kHz$ (*specification 1.3.3.*). The ACH-01[5] is an exception to this behaviour but is expensive.

The piëzo-electric accelerometer is generally more expensive due to the buffer that must be installed to compensate for its high ohmic behaviour. However its frequency behaviour fits the purpose of this project better than the MEMS devices. The mass-spring system formed by the sensor does have a resonance peak located around 1.5 kHz, but because it is not placed inside the frequency range of the applied motional feedback, does not create a significant problem. Because of its fitting characteristics, the piëzo-electric accelerometer is selected over the MEMS accelerometer as accelerometer type. The sensor used in the motional feedback system will thus be a piëzo-electric accelerometer.

3.5.4 DC Filter

The output of a piëzo-electric accelerometer does contain a DC offset of 6 V. This DC offset will need to be filtered without significantly altering the output of the sensor. There are two methods of filtering DC, passively or actively.

Passive filter
A simple first order passive filter is an RC-filter. This RC-filter is a high-pass filter. Passive filters do not need a power supply. The capacitor causes a 90° phase shift. The RC-filter is shown in figure 3.26 (upper circuit).

Active filter
A simple first order active filter is shown in figure 3.26 (lower circuit). This RC-filter is also a high-pass filter. This active filter needs a power supply. The capacitor causes a 90° phase shift. The op amp causes a 180° phase shift. The AC analysis of both circuits is shown in figure 3.27. The amplitude is these filters are the same.

[5] http://www.audiomatica.com/wp/wp-content/uploads/ach_01.pdf

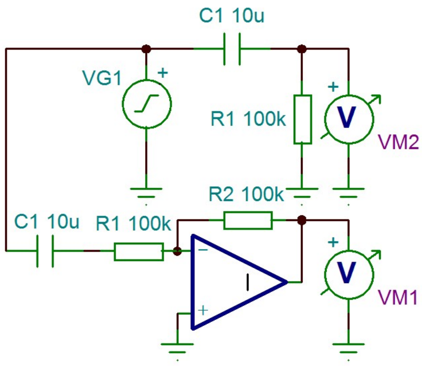


Figure 3.26: Two types of Filters

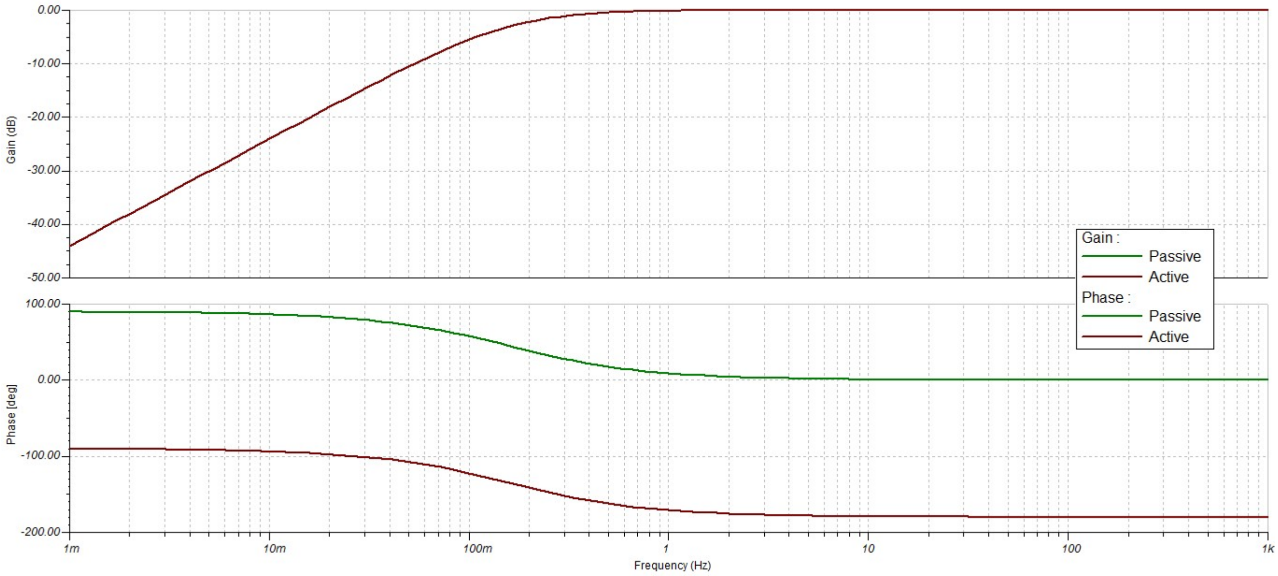


Figure 3.27: AC Analysis of the filters

Decision
A passive DC-filter is simple and effective. According to the figure 3.27, the phase of the passive filter is 0° from 0 Hz. The phase shift of the active filter is $-180°$ from 0 Hz, due to the op amp. This phase shift negatively impacts the closed loop stability of the motional feedback system. The passive DC-filter is therefore preferred over the active DC-filter.

Chapter 4

Implementation and Results

4.1 Nested Feedback Amplifier

As the project is still ongoing test results of the amplifier are not presentable at this stage. The principle has been tested but was not recorded and as such can not be presented in this report.

4.2 4th Order Crossover filter

Before implementing the filter circuit on a PCB, a prototype is made on a prototyping board. The prototype has to meet the given specifications, so the tests are based on:

Transfer function The transfer function must be flat and the bandwidth is 10 Hz − 1 kHz, specification 2.32.

Amplification / Attenuation Attenuation of the low-pass filter at 1.5 kHz at least 15 dB, specification 2.34.

DC Output The DC output is at most 50 mV, specification 2.36.

4.2.1 Transfer Function

Arta[1] (Audio Measurement and Analysis Software) is used to measure the transfer function of the cross-over filters. The audio output of the computer is connected to the input of the crossover filter. The microphone input of the computer is connected to one of the output of the filters. Arta produces white noise and measures the transfer. The result is shown in figure 4.1. The low-pass filter is the red line, the high-pass filter the yellow line and the total transfer is the blue line. However, the results are not as expected. The sound card of the computer has influence on the result. It has a highpass-filter with a cutoff frequency of 100 Hz and a low-pas filter with a cutoff frequency at 20 kHz. The crossover frequency of the crossover filter is around the 1 kHz, which is correct. Shortly, it does not meet specification 2.32, because it is not flat. This is due to the sound card of the computer. The results of section 4.2.2 show amplifications at some frequencies. The amplification are the same at the bandwidth. So finally, it meets the specification.

4.2.2 Amplification / Attenuation

The amplification can be measured with a oscilloscope. The function generator produces a 2 V peak-to-peak sinus with variable frequencies. One of the results is shown in figure 4.2. The generator (blue line) produces a 1.5 kHz

[1] http://www.artalabs.hr/

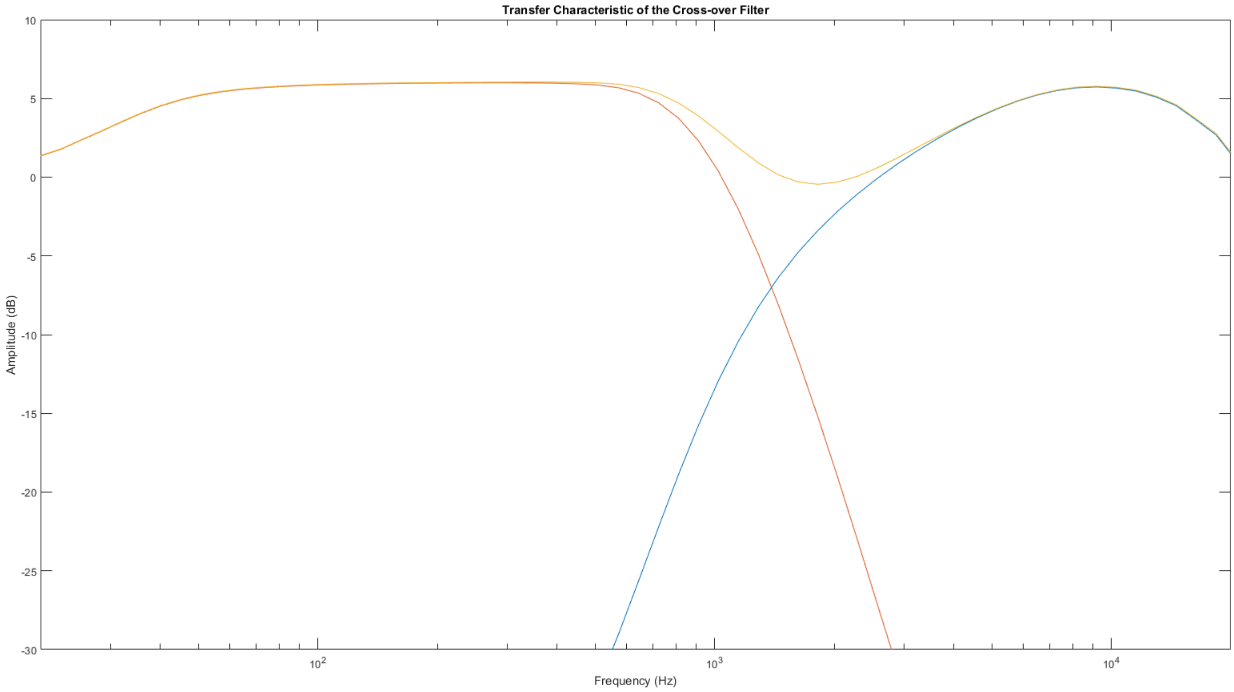


Figure 4.1: Transfer function of the Cross-over

sinus without offset. The output (yellow line) is a 1.5 kHz sinus with a peak-to-peak voltage of 320 mV. The attenuation at 1.5 kHz is:

$$D = 20log(\frac{V_{in}}{V_{out}}) = \frac{2.08}{0.320} = 16.3dB \tag{4.1}$$

This meets specification 2.32, because the attenuation must be at least 15 dB. The other results are in appendix C.

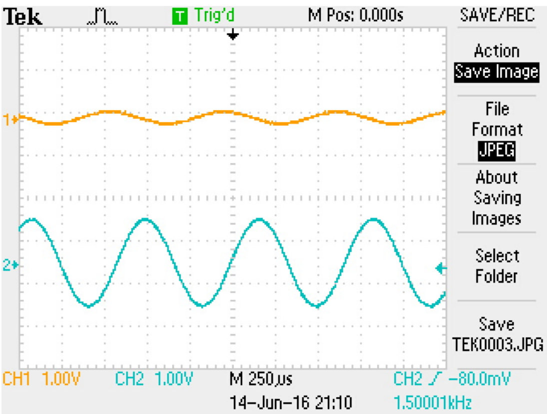


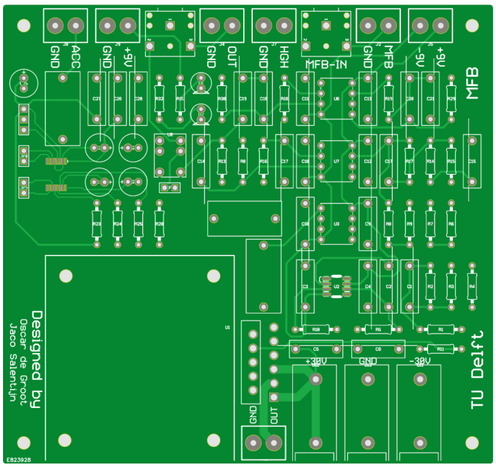
Figure 4.2: Low-pass filter at 1500 Hz

4.2.3 DC Output

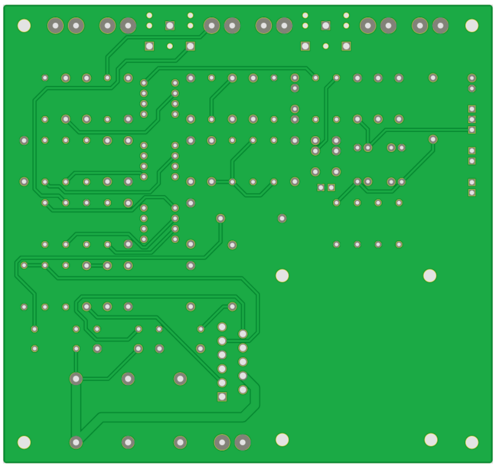
The DC output is measured with a multimeter. The input terminals of the filter are grounded. The multimeter displayed of maximum DC voltage of 40.8 mV. This DC voltage wil be filtered by a DC filter in front of the amplifier. So this circuit is not a limitation to specification 2.36.

4.3 PCB

After testing the crossover prototype, a PCB is designed. A PCB is more professional, easier in use and can be used in the resulting end product. Most of the components are placed in rows to save space. The dimensions are 119.38 *mm* by 111.76 *mm*, which meet specification 2.43.



(a) PCB Design (top) of the amplifier, ADC and crossover



(b) PCB Design (bottom) of the amplifier, ADC and crossover

Chapter 5

Conclusion

5.1 Amplifier

The nested feedback topology suppresses distortion in the output of audio amplifiers at the cost of an extra op amp and cheap passive components. The main issue of the nested feedback topology is its stability. In this report the circuit is stabilised by three measures. Firstly the op amps are selected to have gain bandwidth products that are spaced far apart to prevent synchronous accumulation of phase shift. Secondly lead compensation is applied in the overall feedback of the amplifier, which improves the phase behaviour of the loop gain. Lastly dominant pole compensation is implemented to decrease the loop gain until the circuit reaches a phase margin of $45°$. The large capacitor value necessary for adequate compensation is achieved using the miller effect.

Simulations of the designed circuit show that the closed loop and loop gain behaviour fits the specifications. The circuit is closed loop stable and amplifies 28 dB in the pass band.

5.2 Cross-over

The golden rule for a cross-over: *The cross-over must be inaudible*. The perfect cross-over does not exist, so this crossover is audible. But with some specification, it is possible to make a cross-over with a great quality. With the applied knowledge, the designed cross-over meets all the specifications. The output has a bandwidth of $10\ Hz - 1\ kHz$, and it's phase is linear at this bandwidth. The attenuation at $1.5\ kHz$ is more than $15\ dB$ and the DC output does not exceed $50\ mV$.

A circuit with these properties is hard to make with only passive components. Also small adjustments can simply be made, when needed. It was the right choice to make this filter active, because the need of a power supply isn't a problem. The highest frequency is $20\ kHz$, which doesn't cause problems due to the bandwidth limitation of the op amps. Active filters are easier to design and have comparable results as passive filters.

5.3 Displacement Sensor

An accelerometer is the most practical sensor to measure the motion of the driver. It is directly dual with the measurable quantity and provides an accurate representation of the displacement. A piëzo-electric accelerometer is prefered over a MEMS accelerometer, because its frequency response is more suitable for the motional feedback application.

Chapter 6

Discussion

As mentioned in the introduction, the nested feedback amplifier can be extended to have more than one additional loop. The bottleneck in this design is the phase shift accumulation of the op amps. Using adequate compensation, multiple local feedback loops can lead to a further improvement in distortion reduction.

The parts discussed in this report are designed for a motional feedback application. The parts described here need to be integrated with the other motional feedback system blocks. Furthermore to create a profitable end product the sensor and its wiring can be integrated into a single device that is clicked onto a driver. Such a device omits the need to drill a hole into the driver and adds to the universal character of the motional feedback system.

Bibliography

[1] Klaassen J. A., Koning S. H., (1968). *Motional feedback with louspeakers* [Online], Available:
`http://mfb.piratelogic.nl/data/docs/PTR.1968.MFB.motionalfeedbackspeakers.en.pdf`

[2] Rosema Y., Overwater R. *Digital Implementation of a Motional Feedback Audio System*, Department of Electrical Engineering, Mathematics and Computer Science, Delft University of Technology, Delft, Netherlands, 2016

[3] Bons W., Hol T. *Digital Control in a Motional Feedback Audio System*, Department of Electrical Engineering, Mathematics and Computer Science, Delft University of Technology, Delft, Netherlands, 2016

[4] Linkwitz S.H., (2014). *Crossovers* [Online] Available:
`http://www.linkwitzlab.com/crossovers.htm`

[5] Mancini, R., (2003). *Analyzing Feedback Loops Containing Secondary Amplifiers* [Online], Available:
`http://www.ti.com/lit/an/slyt103/slyt103.pdf`

[6] Cherry E. M., (1977, May 7). *A new result in negative-feedback theory, and its application to audio power amplifiers* [Online], Available:
`http://onlinelibrary.wiley.com/doi/10.1002/cta.4490060305/epdf`

[7] Mancini R., "Stability Analysis of Voltage-Feedback Op Amps," Texas Instruments Inc., Dellas, Tex. Rep.SLOA020A, March. 2001. Available:
`http://www.ti.com/lit/an/sloa020a/sloa020a.pdf`

[8] Lundberg K., (2004). *Internal and External Op-Amp Compensation: A Control-Centric Tutorial* [Online], Available:
`http://web.mit.edu/klund/www/papers/ACC04_opcomp.pdf`

[9] Basso C., (2008, November). *Transient response counts when choosing phase margin* [Online], Available:
`http://powerelectronics.com/site-files/powerelectronics.com/files/archive/powerelectronics.com/power_systems/simulation_modeling/Transient-response-phase-margin-PET.pdf`

[10] Linkwitz S.H., (1976). *Active Crossover Networks for Noncoincident Drivers* [Online], Available:
`http://www.nextdigital.com.br/Linkwitz_%20AES%201976.pdf`

[11] Ting H.W., Wang H.Y., (2010). *Improvement of Stop-Band Attenuation for Sallan-Key Low-Pass Filter* [Online], Available:
`http://ieeexplore.ieee.org/stamp/stamp.jsp?tp=&arnumber=5669175&tag=1`

[12] Lacanette K., (1991). *A Basic Introduction to Filters - Active, Passive and switched-Capacitor* [Online], Available:
`http://www.swarthmore.edu/NatSci/echeeve1/Ref/DataSheet/IntroToFilters.pdf`

[13] Janssen G. J. M., (2015). *EPO-1 Luidspreker Project 2015* [Online], Available: https://blackboard.tudelft.nl/bbcswebdav/pid-2545827-dt-content-rid-8955430_2/courses/37526-151602/LS_EPO-1_2015_NB.pdf

Appendix A

Distortion Mask

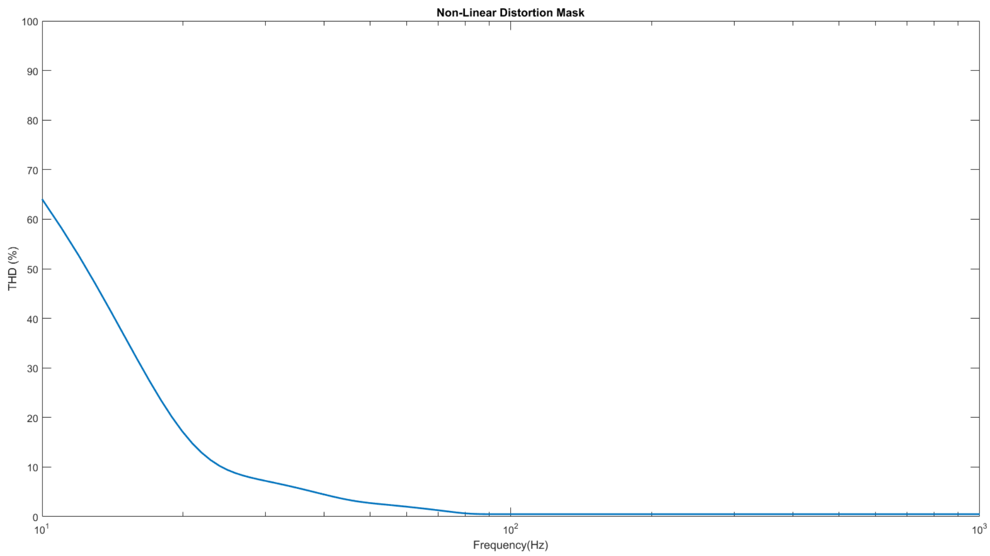


Figure A.1: The distortion mask created to specify the output distortion as a function of frequency.

Appendix B

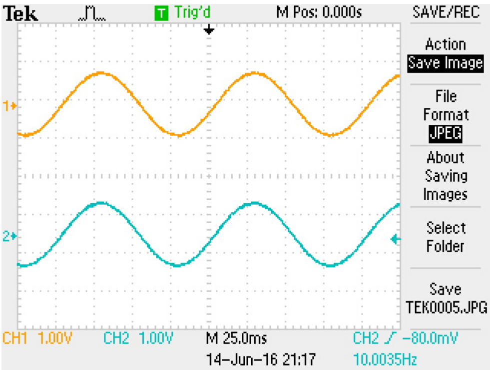
Operational Amplifiers Specifications

Table B.1: Relevant specification for the two selected op amps. The values displayed are the typical values.

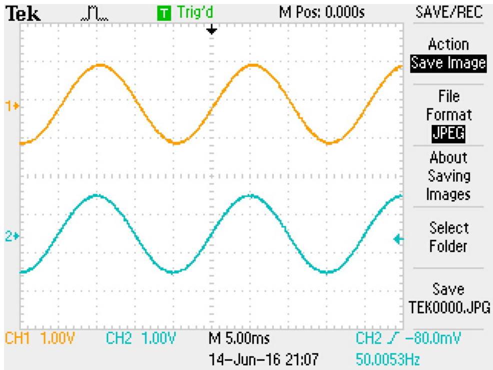
Specification	LM3886	OPA1611	LM833
Gain-Bandwidth Product	8 MHz	40 MHz	15 MHz
Open Loop Voltage Gain	115 dB	130 dB	110 dB
THD + N	0.03%	0.000015%	0.002%
Offset Voltage	1 mV	100 μV	0.3 mV
Bias Current	0.2 μA	60 nA	300 nA
Input Offset Current	10 nA	25 nA	10 nA

Appendix C

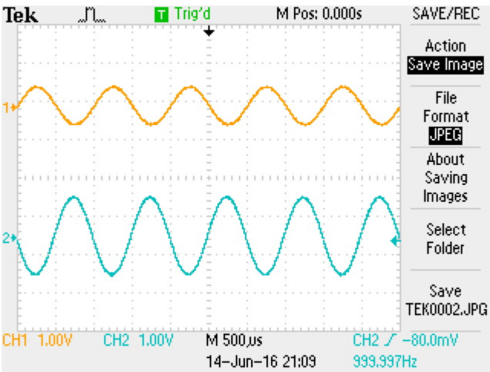
Results Cross-over



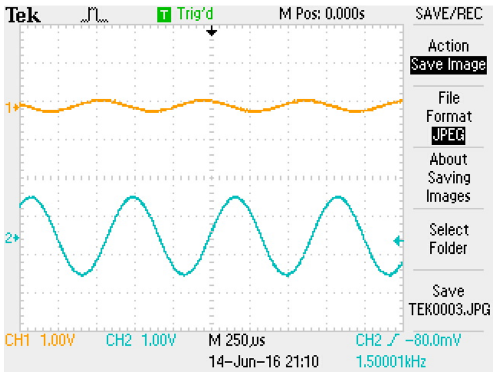
(a) Low-pass filter at 10 Hz



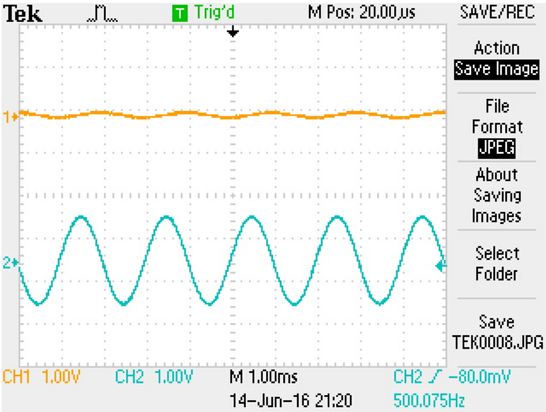
(b) Low-pass filter at 50 Hz



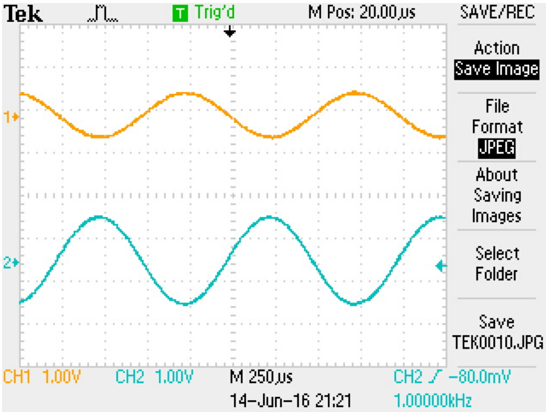
(a) Low-pass filter at 1000 Hz



(b) Low-pass filter at 1500 Hz



(a) High-pass filter at 500 Hz



(b) High-pass filter at 1000 Hz

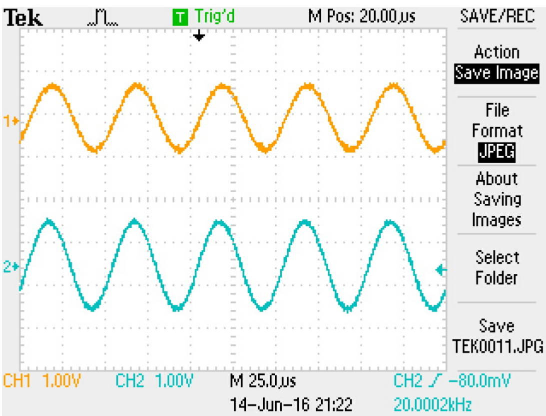


Figure C.4: High-pass filter at 20000 Hz