UNIVERSITY OF NAIROBI

COLLEGE OF ARCHITECTURE AND ENGINEERING

SCHOOL OF ENGINEERING

DEPARTMENT OF ELECTRICAL AND INFORMATION ENGINEERING

PROJECT INDEX: 096

PROJECT TITLE: DESIGN OF A 50W POWER AMPLIFIER DRIVEN BY AN ACTIVE 4-WAY CROSS-OVER NETWORK AND A PRE-AMPLIFIER

Compiler:

OPIYO STEPHEN OMONDI

F17/29477/2009

SUPERVISOR: MR. S. L. OGABA
EXAMINER: DR. G.S.O.

Project report submitted in partial fulfilment of the requirement for the award of the degree of **Bachelor of Science inELECTRICAL AND ELECTRONIC ENGINEERING** of the University of Nairobi 2015

Submitted on: 24THApril, 2015

DECLARATION OF ORIGINALITY

NAME OF STUDENT: OPIYO STEPHEN OMONDI
REGISTRATION NUMBER: F17/29477/2009
COLLEGE: Architecture and Engineering
FACULTY/SCHOOL/INSTITUTE: Engineering
DEPARTMENT: Electrical and Information Engineering
COURSE NAME: Bachelor of Science in Electrical and Electronic Engineering
PROJECT TITLE: DESIGN OF A 50W POWER AMPLIFIER DRIVEN BY AN ACTIVE 4-WAY CROSS-OVER NETWORK AND A PRE-AMPLIFIER

1) I understand what plagiarism is and I am aware of the university policy in this regard.
2) I declare that this final year project report is my original work and has not been submitted elsewhere for examination, award of a degree or publication. Where other people's work or my own work has been used, this has properly been acknowledged and referenced in accordance with the University of Nairobi's requirements.
3) I have not sought or used the services of any professional agencies to produce this work.
4) I have not allowed, and shall not allow anyone to copy my work with the intention of passing it off as his/her own work.
5) I understand that any false claim in respect of this work shall result in disciplinary action, in accordance with University anti-plagiarism policy.

Signature………………………………………………………………

Date…………………………………………………………………

CERTIFICATION

It is certified that Mr. OPIYO STEPHEN OMONDI, Registration No. F17/29477/2009, student at The University of Nairobi, Department of Electrical and Information engineering, has submitted the project entitled "DESIGN OF A 50W POWER AMPLIFIER DRIVEN BY AN ACTIVE 4-WAY CROSS-OVER NETWORK AND A PRE-AMPLIFIER'' under my guidance towards partial fulfilment of the requirements for the award of the Undergraduate degree of B.Sc. in Electrical and Electronic Engineering.

This is a record of project work carried out by him under my guidance and supervision. His work is found to be outstanding and has not been done earlier.
I wish him success in all his endeavors.

Mr. S. L. OGABA
Department of Electrical and Information Engineering
The University of Nairobi

Signature: …………………………………………………………

Date: ……………………………………………………………….

DEDICATION

I would like to dedicate this work to my family and friends and thank them for the financial and moral support they gave me during my entire quest in this degree course.

ACKNOWLEDGEMENTS

First and foremost, all thanks goes to the Almighty God for His guidance and being by my side throughout my academic life.

My special gratitude goes to the Dean-Faculty of Engineering, Chairman-Department of Electrical and Information Engineering and all my lecturers as well as all non-teaching staff at the Department for their contributions during my time in the facilities.

Sincere thanks to my supervisor Mr. S. L. Ogaba who guided and provided me with invaluable advice and tips while undertaking this project.

I am also thankful to a selection of my classmates for their help in understanding the project.

Finally I would like to thank my friends and family for their help and support throughout my academic life.

TABLE OF CONTENTS

LIST OF TABLES

LIST OF FIGURES

LIST OF ABBREVIATIONS/ACRONYMS

LPF	Low-pass filter
BPF	Band-pass filter
HPF	High-pass filter
Hz	hertz
KHz	Kilo-hertz
MHz	Mega-hertz
BW	bandwidth
Ω	SI unit for resistance, ohm
KΩ	Kilo-ohm
nF	Nano-Farads
μF	Micro-Farads
mF	Milli-Farads
C	capacitance
R	resistance
L	inductance
dB	decibel
s	Complex frequency
A(s)	Transfer function
f_c	Corner frequency
f_c	Center frequency
f_m	Mid frequency
Q	Quality factor
MFB	Multiple Feedback topology

ABSTRACT

Most speaker boxes have an electronic circuit inside them called a crossover. Its job is to split the sound into frequency ranges; it sends the low frequency sounds to the woofer, and the high frequency sounds to the tweeter. This is important because each of the loudspeakers is built to produce certain frequency ranges, and they don't sound very good outside of their proper range. Most speaker boxes come with passive crossovers built in. The crossovers are made up of some inductors and capacitors that filter the low frequencies from the highs. But passive crossovers have a number of very significant drawbacks. The biggest problem with passive crossovers is that they significantly degrade the quality of the signals that pass through them by introducing distortion.

Music systems of today have separate speakers (woofer, midrange and tweeter) connected to the output of amplifiers. This project entails the design of a 50W power amplifier driven by an active 4-way cross-over network. An active crossover, is built from the highest quality operational amplifiers (op-amps) and the filter functions are applied to the millivolt level, high impedance, line-level preamp input instead of the low impedance, high voltage level amplifier output. The active filters are to separate the sounds into 4-band i.e. the Bass-band, midrange-1, midrange-2 and tweeter-band. The biggest advantage of active crossovers is the sound quality. A vanishingly small amount of noise and distortion is introduced in the op-amp filter circuits, on the order of 0.00005% total harmonic distortion.

CHAPTER 1

INTRODUCTION

1.1Filters

A filter is an analog circuit which performs signal processing function of removing unwanted frequency components from a signal, enhances wanted ones or both.
Electronic filters can be:

- Active or passive
- High-pass, low-pass, band-pass, band-stop or all-pass
- Analog or digital
- Linear or non-linear
- Infinite impulse response or finite impulse response
- Discrete-time or continuous-time

In this literature we are going to dwell mostly on the first two points and especially on the Active filter.

1.2 Passive Filters

Passive filters are circuits made up of passive components that include resistors, capacitors and inductors. The main disadvantage of Passive Filters is that the amplitude of the output signal is less than that of the input signal, i.e. the gain is never greater than unity and that the load impedance affects the filters characteristics. With passive filter circuits containing multiple stages, this loss in signal amplitude called “Attenuation” can become quiet severe. One way of restoring or controlling this loss of signal is by using amplification through the use of Active Filters.

1.3 Active Filters

As their name implies, **Active Filters** contain active components such as operational amplifiers, transistors or FET’s within their circuit design. They draw their power from an external power source and use it to boost or amplify the output signal. These active devices are used in combination with some resistors and capacitors in their feedback loops, to synthesize the desired filter characteristics and provide an LRC-like filter performance at low frequencies.Filter amplification can also be used to either shape or alter the frequency response of the filter circuit by producing a more selective output response, making the output bandwidth of the filter narrower or even wider. Then the main difference between a “passive filter” and an “active filter” is amplification.

Active filters can have high input impedance, low output impedance, and virtually any arbitrary gain. They are also usually easier to design than passive filters. Possibly their most important attribute is that they lack inductors, thereby reducing the problems associated with those components. Still, the problems of accuracy and value spacing also affect capacitors, although to a lesser degree.

Unlike a passive high pass filter which has in theory an infinite high frequency response, the performance of an active filter at high frequencies is limited by the gain-bandwidth product (or open loop gain)of the amplifying elements, but within the amplifier's operating frequency range, the op-amp-based active filter can achieve very good accuracy, provided that low-tolerance resistors and capacitors (i.e. 1% or less) are used. Moreover, active filters generate noise due to the amplifying circuitry, but this can be minimized by use of low-noise amplifiers and careful circuit design. Still, active filters are generally much easier to design than passive filters, they produce good performance characteristics, very good accuracy with a steep roll-off and low noise when used with a good circuit design.

1.4 Practical Applications

A common need for filter circuits is in high-performance stereo systems, where certain ranges of audio frequencies need to be amplified or suppressed for best sound quality and power efficiency. Equalizers allow the amplitudes of several frequency ranges to be adjusted to suit the listener's taste and acoustic properties of the listening area. Crossover networks block certain ranges of frequencies from reaching speakers. A tweeter (high-frequency speaker) is inefficient at reproducing low-frequency signals such as drum beats, so a crossover circuit is connected between the tweeter and the stereo's output terminals to block low-frequency signals, only passing high-frequency signals to the speaker's connection terminals. This gives better audio system efficiency and thus better performance. Both equalizers and crossover networks are examples of filters, designed to accomplish filtering of certain frequencies.

Another practical application of filter circuits is in the “conditioning” of non-sinusoidal voltage waveforms in power circuits. Some electronic devices are sensitive to the presence of harmonics in the power supply voltage, and so require power conditioning for proper operation. If a distorted sine-wave voltage behaves like a series of harmonic waveforms added to the fundamental frequency, then it should be possible to construct a filter circuit that only allows the fundamental waveform frequency to pass through, blocking all (higher-frequency) harmonics.

1.5 Statement of Problem

Most modern music systems have separate speakers (woofer, mid-range & tweeter) connected to output of amplifiers. The modern consumer is currently deviating away from passive crossover filters and towards active filters due to their numerous advantages over the former. This project employs design of four active filters of different cut-off frequencies.

1.6 Objectives

The aim of this project is to design an amplifier which operates from both AC and DC voltages (12v) which drives the four channels.

The four filters are as follows:
One Active Low Pass Filter
One Active High Pass Filter
Two Band-pass Filters.

CHAPTER 2

LITERATURE REVIEW

2.1Fundamentals of Filters

The simplest low-pass filter is the passive RC low-pass network shown in Figure 1.

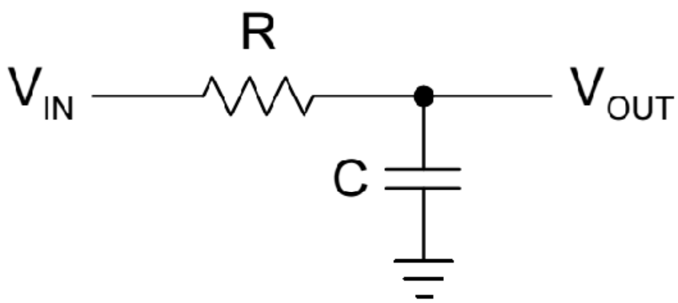

Figure 1: First-Order Passive RC Low-Pass Filter

Its transfer function is:

$$A(s) = \frac{\frac{1}{RC}}{s + \frac{1}{RC}} = \frac{1}{1 + sRC}$$

where the complex frequency variable, $s = j\omega + \sigma$, allows for any time variable signals. For pure sine waves, the damping constant, σ, becomes zero and $s = j\omega$. For a normalized presentation of the transfer function, s is referred to the filter's corner frequency, or –3 dB frequency ω_C, and has these relationships:

$$s = \frac{s}{\omega_C} = \frac{j\omega}{\omega_C} = j\frac{f}{f_C} = j\Omega$$

With the corner frequency of the low-pass in Figure 1 being $f_C = 1/2\pi RC$, s becomes $s = sRC$ and the transfer function A(s) results in:

$$A(s) = \frac{1}{1 + s}$$

The magnitude of the gain response is:

$$|A| = \frac{1}{\sqrt{1 + \Omega^2}}$$

For frequencies, $\Omega >> 1$, the roll-off is 20 dB/decade. For a steeper roll-off, n filter stages can be connected in series as shown in Figure 2. To avoid loading effects, op amps, operating as impedance converters, separate the individual filter stages.

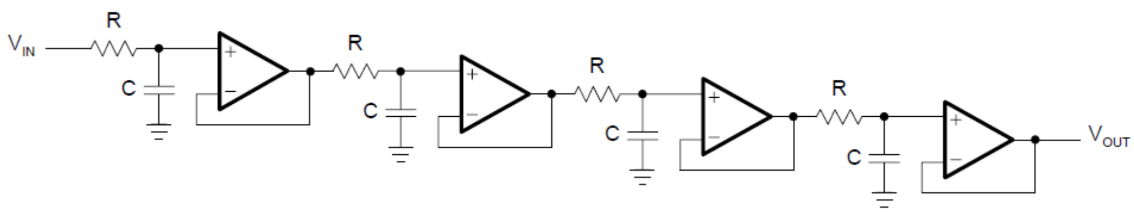

Figure 2: Fourth-Order Passive RC Low-Pass with Decoupling Amplifiers

The resulting transfer function is:

$$A(s) = \frac{1}{(1 + \alpha_1 s)(1 + \alpha_2 s)\ldots(1 + \alpha_n s)}$$

In the case that all filters have the same cut-off frequency f_C, the coefficients become $\alpha_1 = \alpha_2 = \cdots \alpha_n = \alpha = \sqrt{\sqrt[n]{2} - 1}$ and f_C of each partial filter is $1/\alpha$ times higher than f_C of the overall filter.

Figure 3 shows the response of a fourth-order RC low-pass filter. The roll-off of each partial filter (Curve 1) is –20 dB/decade, increasing the roll-off of the overall filter (Curve 2) to 80 dB/decade.

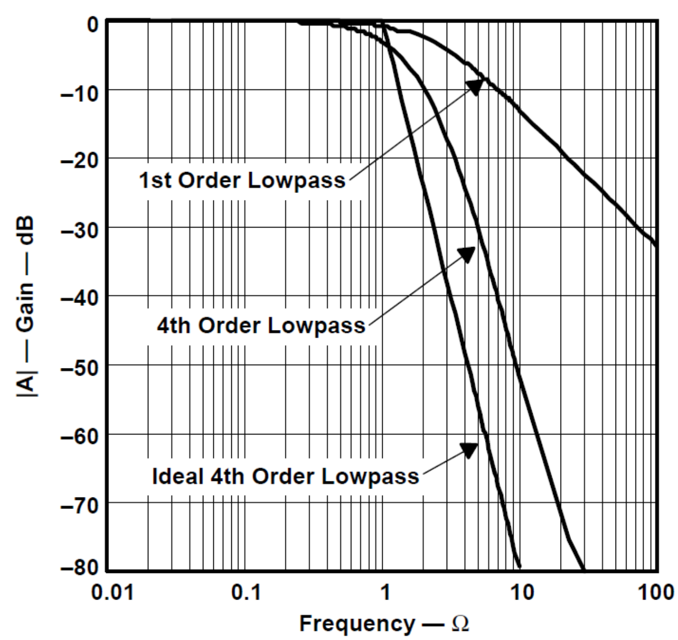

Figure 3: Frequency response of a Fourth-Order Passive RC LPF

The corner frequency of the overall filter is reduced by a factor of $\alpha \approx 2.3$ times versus the –3 dB frequency of partial filter stages.
In addition, Figure 3 shows the transfer function of an ideal fourth-order low-pass function (Curve 3).

In comparison to the ideal low-pass, the RC low-pass lacks in the following characteristics:

- The pass-band gain varies long before the corner frequency, f_C, thus amplifying the upper pass-band frequencies less than the lower pass-band.
- The transition from the pass-band into the stop-band is not sharp, but happens gradually, moving the actual 80-dB roll off by 1.5 octaves above f_C.
- The phase response is not linear, thus increasing the amount of signal distortion significantly.*(Kugelstadt)*

The gain and phase response of a low-pass filter can be optimized to satisfy one of the following three criteria:

1) A maximum pass-band flatness,
2) An immediate pass-band-to-stop-band transition,
3) A linear phase response.

For that purpose, the transfer function must allow for complex poles and needs to be of the following type:

$$A(s) = \frac{A_0}{(1 + a_1 s + b_1 s^2)(1 + a_2 s + b_2 s^2)\dots(1 + a_n s + b_n s^2)} = \frac{A_0}{\prod_i (1 + a_i s + b_i s^2)}$$

where A_0 is the pass-band gain at dc, and ai and bi are the filter coefficients.
Since the denominator is a product of quadratic terms, the transfer function represents a series of cascaded second-order low-pass stages, with ai and bi being positive real coefficients. These coefficients define the complex pole locations for each second-order filter stage, thus determining the behavior of its transfer function.
The following three types of predetermined filter coefficients are usually used and are available listed in table format.

- The **Butterworth coefficients**, optimizing the pass-band for maximum flatness
- The **Tschebyscheff coefficients**, sharpening the transition from pass-band into the stop-band.
- The **Bessel coefficients**, linearizing the phase response up to f_C

The transfer function of a passive RC filter does not allow further optimization, due to the lack of complex poles. The only possibility to produce conjugate complex poles using passive components is the application of LRC filters. However, these filters are mainly used at high frequencies. In the lower frequency range (< 1 MHz) the inductor values become very large and the filter becomes uneconomical to manufacture. In these cases active filters are used.

The following paragraphs introduce the most commonly used filter optimizations.

2.1.1 Butterworth Low-Pass Filters

Butterworth filters are termed maximally-flat-magnitude-response filters, optimized for gain flatness in the pass-band. The attenuation is –3 dB at the cutoff frequency. Above the cutoff frequency the attenuation is –20 dB/decade/order. The transient response of a Butterworth filter to a pulse input shows moderate overshoot and ringing.Therefore, a Butterworth low-pass is often used as anti-aliasing filter in data converter applications where precise signal levels are required across the entire pass-band.*(Karki, September 2002)*
Figure 4 plots the gain response of different orders of Butterworth low-pass filters versus the normalized frequency axis, ▯ (▯ = ▯ / ▯▯); the higher the filter order, the longer the pass-band flatness.

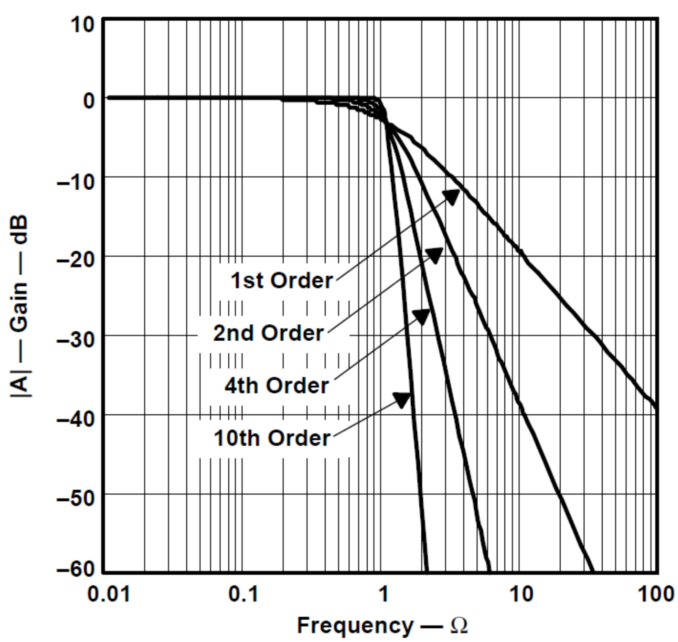

Figure 4: Amplitude Response of Butterworth Low-Pass Filters

2.1.2 Tschebyscheff Low-Pass Filters

The Tschebyscheff low-pass filters provide an even higher gain roll-off above fC. Tschebyscheff or **Chebyshev** filters as some books would call them, are designed to have ripple in the pass-band, but steeper roll-off after thecutoff frequency. Cutoff frequency is defined as the frequency at which the response falls belowthe ripple band. For a given filter order, a steeper cutoff can be achieved by allowing more pass-band ripple. The transient response of a Chebyshev filter to a pulse input shows more overshoot and ringing than a Butterworth filter. However, as Figure 5 shows, the pass-band gain is not monotone, but contains ripples of constant magnitude instead. For a given filter order, the higher the pass-band ripples, the higher the filter's roll-off.*(Karki, September 2002)*

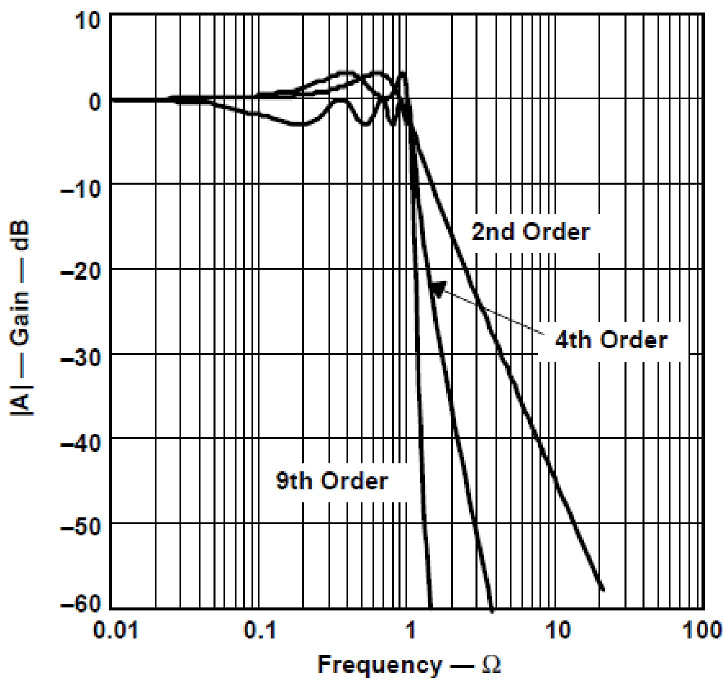

Figure 5: Amplitude Response of Tschebyscheff Low-Pass Filters

(Kugelstadt)

With increasing filter order, the influence of the ripple magnitude on the filter roll-off diminishes.

Each ripple accounts for one second-order filter stage. Filters with even order numbers generate ripples above the 0-dB line, while filters with odd order numbers create ripples below 0 dB.

Tschebyscheff filters are often used in filter banks, where the frequency content of a signal is of more importance than a constant amplification.

2.1.3 Bessel Low-Pass Filters

Bessel filters are optimized for maximally-flat time delay (or constant-group delay). This means that they have linear phase response and excellent transient response to a pulse input. This comes at the expense of flatness in the pass-band and rate of rolloff. The cutoff frequency isdefined as the –3-dB point.
(Karki, September 2002)

The Bessel low-pass filters have a linear phase response (Figure 6) over a wide frequency range, which results in a constant group delay (Figure 7) in that frequency range. Bessel low-pass filters, therefore, provide an optimum square-wave transmission behavior. However, the pass-band gain of a Bessel low-pass filter is not as flat as that of the Butterworth low-pass, and the transition from pass-band to stop-band is by far not as sharp as that of a Tschebyscheff low-pass filter (Figure 8).

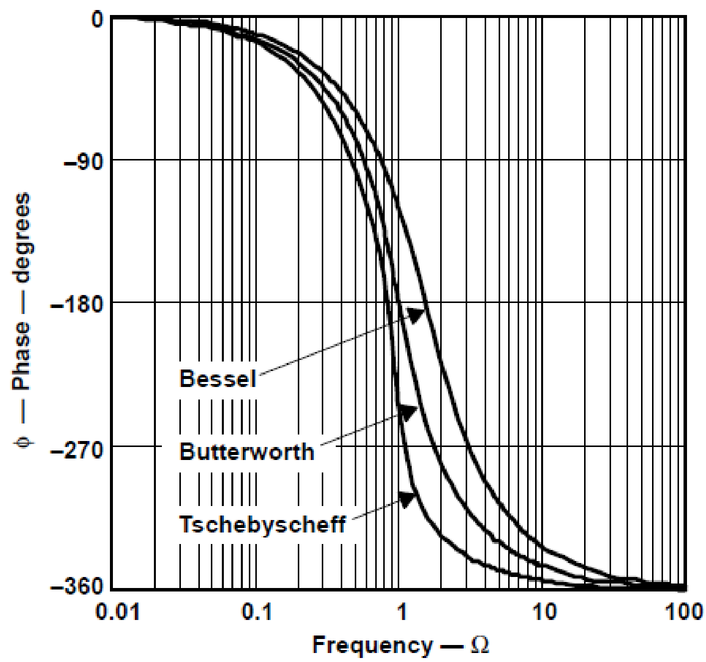

Figure 6: Comparison of Phase Responses of Fourth-Order Low-Pass Filters

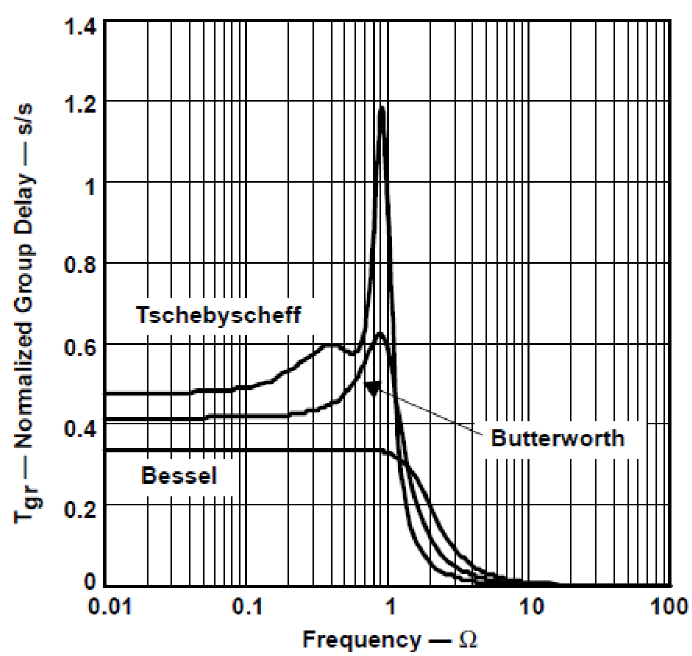

Figure 7: Comparison of Normalized Group Delay (Tgr) of Fourth-Order Low-Pass Filters

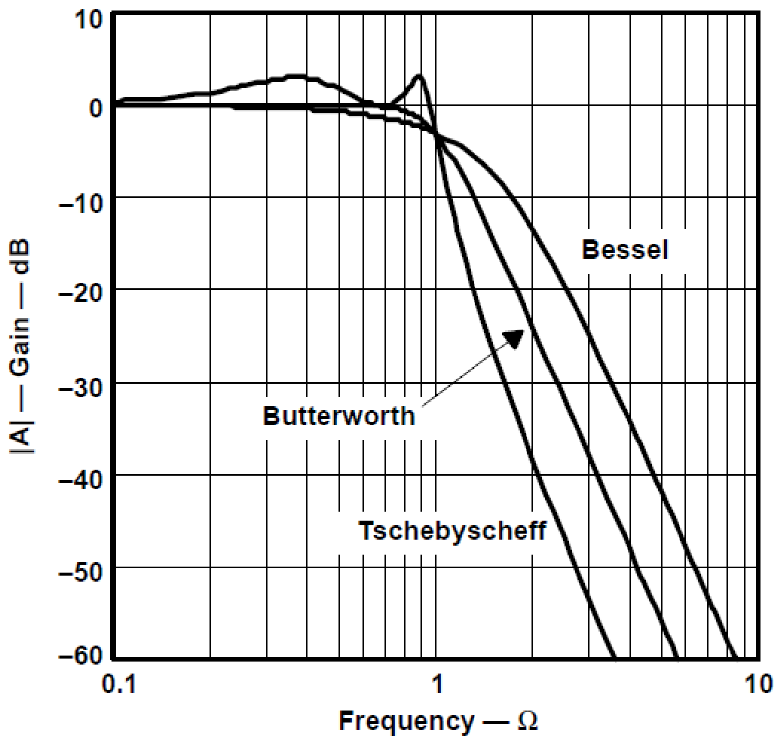

Figure 8: Comparison of Gain Responses of Fourth-Order Low-Pass Filters

(Kugelstadt)

2.1.4 Quality Factor Q

The quality factor Q is an equivalent design parameter to the filter order n. Instead of designing an nth order Tschebyscheff low-pass, the problem can be expressed as designing a Tschebyscheff low-pass filter with a certain Q.

For band-pass filters, Q is defined as the ratio of the mid frequency, fm, to the bandwidth at the two –3 dB points:

$$Q = \frac{f_m}{(f_2 + f_1)}$$

For low-pass and high-pass filters, Q represents the pole quality and is defined as:

$$Q = \frac{\sqrt{b_i}}{a_i}$$

High Qs can be graphically presented as the distance between the 0-dB line and the peak point of the filter's gain response. An example is given in Figure 9, which shows a tenth-order Tschebyscheff low-pass filter and its five partial filters with their individual Qs.

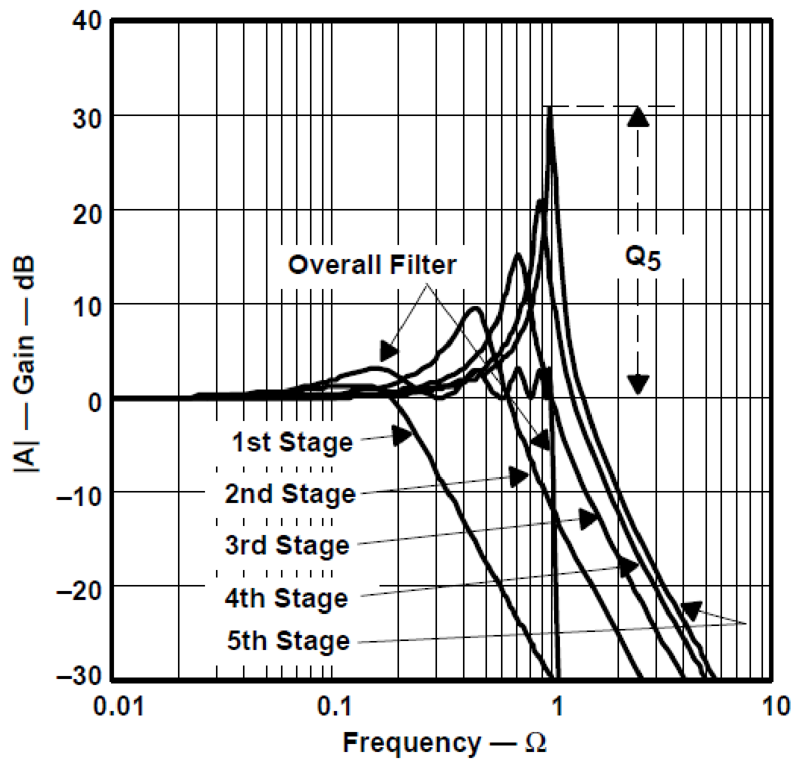

Figure 9: Graphical Presentation of Quality Factor Q on a Tenth-Order Tschebyscheff Low-Pass Filter with 3-dB Pass-band Ripple

The gain response of the fifth filter stage peaks at 31 dB, which is the logarithmic value of Q5:

$$Q_5|_{dB} = 20\log Q_5$$

Solving for the numerical value of Q5 yields:

$$Q_5 = 10^{\frac{31}{20}} = 35.48$$

The graphical approximation is good for Q > 3. For lower Qs, the graphical values differ from the theoretical value significantly. However, only higher Qs are of concern, since the higher the Q is, the more a filter inclines to instability.

2.2 Low-Pass Filter Design

Low-pass filters are the most widely applied filter type. They are designed to readily pass all frequencies extending from dc to a set cutoff frequency, fc. This region where the frequencies readily pass through the filter is called the pass-band, and fc is defined as the filter bandwidth.

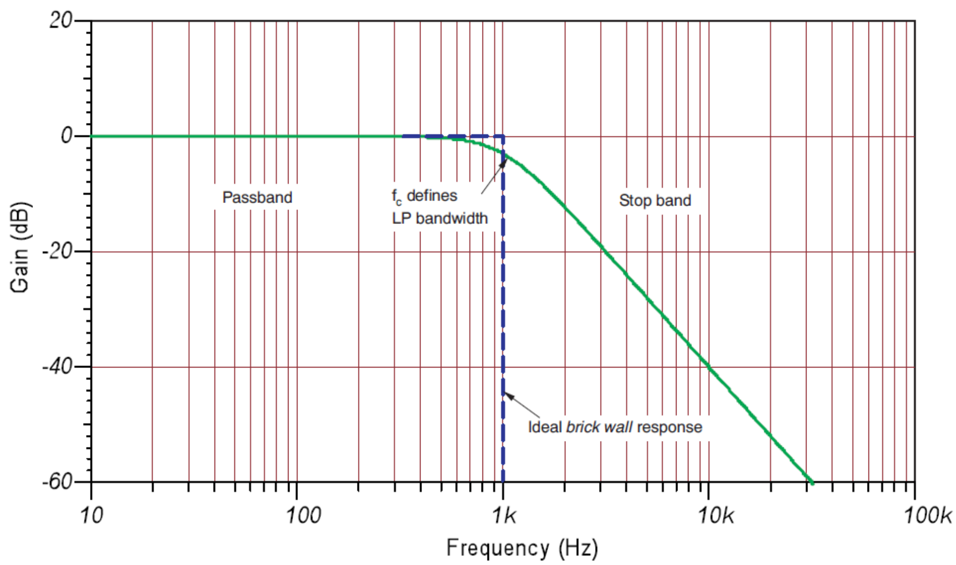

Figure 10: An example of a Low-Pass Filter Response

Once the cutoff frequency is reached, the filter begins to attenuate any frequency higher than fc. The region above fc is called the stop-band. Most often the attenuation increases, or rolls off, and achieves very high levels which are limited only by the non-ideal electronics used to construct the filter. These include the stray and lead inductance of the capacitors, non-ideal operational amplifier limitations, and circuit parasitics associated with the physical circuit layout.

The equation below represents a cascade of second-order low-pass filters. The transfer function of a single stage is:

$$A_i(s) = \frac{A_0}{(1 + a_i s + b_i s^2)}$$

For a first-order filter, the coefficient b is always zero (b1=0), thus yielding:

$$A(s) = \frac{A_0}{(1 + a_1 s)}$$

The first-order and second-order filter stages are the building blocks for higher-order filters. Often the filters operate at unity gain (A0=1) to lessen the stringent demands on the op amp's open-loop gain.
Figure 11 shows the cascading of filter stages up to the sixth order. A filter with an even order number consists of second-order stages only, while filters with an odd order number include an additional first-order stage at the beginning.

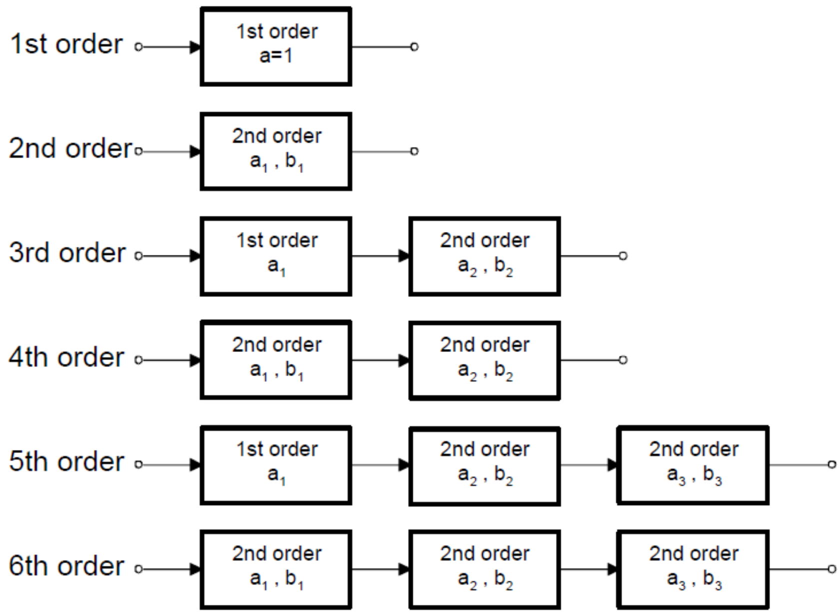

Figure 11: Cascading Filter Stages for Higher-Order Filters

2.2.1 First-Order Low-Pass

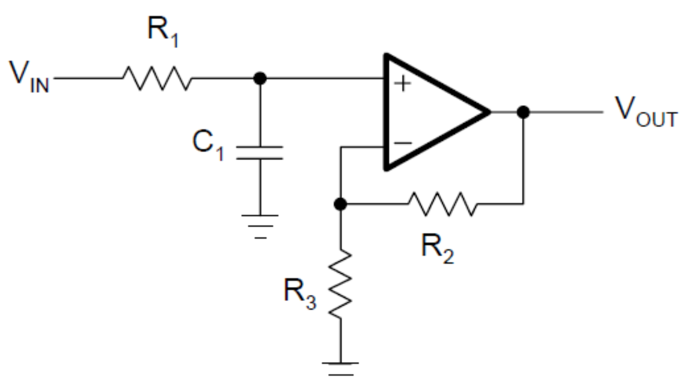

Figure 12:First-Order Non-inverting Low-Pass Filter

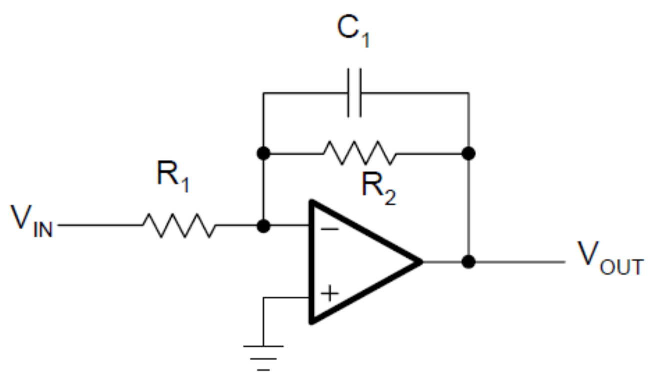

Figure 13:First-Order Inverting Low-Pass Filter

The figures above show a first-order low-pass filter in the non-inverting (Figure 12) and in the inverting (Figure 13) configuration.

The transfer functions of the circuits are as follows:
For the non-inverting:

$$A(s) = \frac{1 + \frac{R_2}{R_3}}{1 + \omega_c R_1 C_1 s}$$

For the inverting:

$$A(s) = \frac{-\frac{R_2}{R_1}}{1 + \omega_c R_2 C_1 s}$$

The negative sign indicates that the inverting amplifier generates a 180° phase shift from the filter input to the output.

In this paper, however, the non-inverting configuration is used in design to avoid this phase shift.
The coefficient comparison between the non-inverting transfer function and

$$A(s) = \frac{A_0}{1 + a_1 s}$$

Yields the following:

$$A_0 = 1 + \frac{R_2}{R_3}$$

$$a_1 = \omega_c R_1 C_1$$

To dimension the circuit, the corner frequency (fC), the dc gain (A0), and capacitor C1 need to be specified and then solve for resistors R1 and R2:

$$R_1 = \frac{a_1}{2\pi f_C C_1}$$

$$R_2 = R_1(A_0 - 1)$$

The coefficient a1 is taken from the Butterworth coefficients table in the appendix. Note, that all filter types are identical in their first order and a1 = 1. For higher filter orders, however, a1 ≠ 1 because the corner frequency of the first-order stage is different from the corner frequency of the overall filter.

2.2.2 Second-Order Low-Pass Filter

There are two topologies for a second-order low-pass filter, the Sallen-Key and the Multiple Feedback (MFB) topology. The topology chosen for the design a second-order low-pass filter in this paper is the Sallen-Key topology.

The general Sallen-Key topology in Figure below allows for separate gain setting via A0 = 1+R4/R3. However, the unity-gain topology in Figure is usually applied in filter designs with high gain accuracy, unity gain, and low Qs (Q < 3).

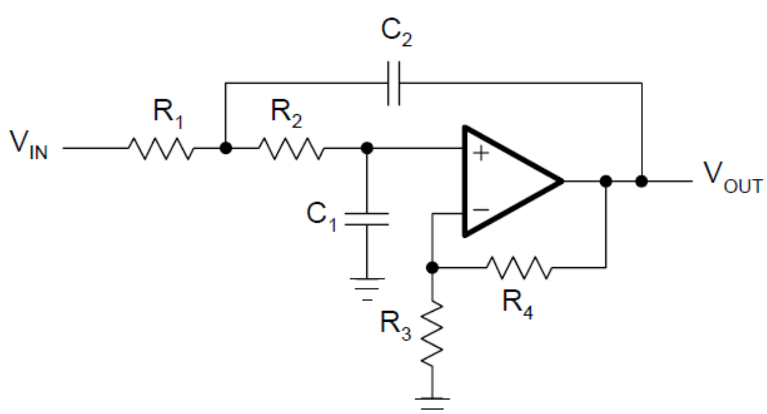

Figure 14: General Sallen-Key Low-Pass Filter

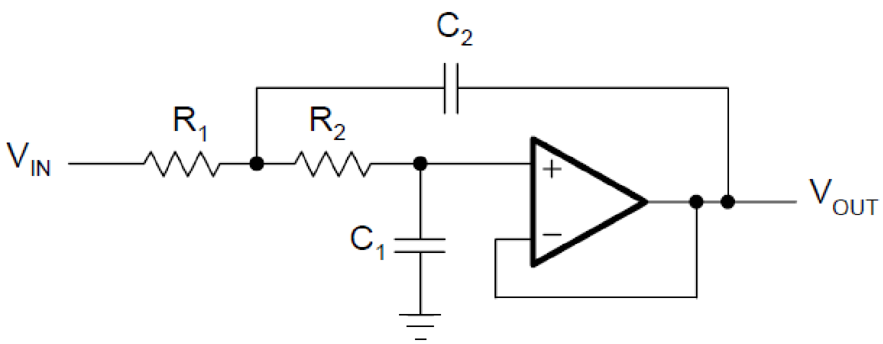

Figure 15: Unity-Gain Sallen-Key Low-Pass Filter

The transfer function of the general circuit is:

$$A(s) = \frac{A_0}{1 + \omega_c \left[C_1 (R_1 + R_2) + (1 - A_0) R_1 C_2 \right] s + \omega_c^2 R_1 R_2 C_1 C_2 s^2}$$

For the unity-gain circuit, the transfer function simplifies to:

$$A(s) = \frac{1}{1 + \omega_c C_1 (R_1 + R_2) s + \omega_c^2 R_1 R_2 C_1 C_2 s^2}$$

The coefficient comparison between this transfer function and Equation 16–2 yields:

$$A_0 = 1$$

$$a_1 = \omega_c C_1 (R_1 + R_2)$$

$$b_1 = \omega_c^2 R_1 R_2 C_1 C_2$$

Given C1 and C2, the resistor values for R1 and R2 are calculated through:

$$R_{1,2} = \frac{a_1 C_2 \mp \sqrt{a_1^2 C_2^2 - 4 b_1 C_1 C_2}}{4 \pi f_c C_1 C_2}$$

In order to obtain real values under the square root, C2 must satisfy the following condition:

$$C_2 \geq C_1 \frac{4 b_1}{a_1^2}$$

A special case of the general Sallen key topology is the application of equal resistor values and equal capacitor values: R1 = R2 = R and C1 = C2 = C

Thus the transfer function results to;

$$A(s) = \frac{A_0}{1 + \omega_c RC(3 - A_0)s + (\omega_c RC)^2 s^2}$$

With;

$$A_0 = 1 + \frac{R_2}{R_1}$$

The coefficient comparison with equation;

$$A_i(s) = \frac{A_0}{(1 + a_i s + b_i s^2)}$$

Yields the following;

$$a_1 = \omega_c RC(3 - A_0)$$

$$b_1 = (\omega_c RC)^2$$

Given C and solving for R and A0 results in:

$$R = \frac{\sqrt{b_1}}{2\pi f_c C}$$

$$A_0 = 3 - \frac{a_1}{\sqrt{b_1}} = 3 - \frac{1}{Q}$$

Thus, A0 depends solely on the pole quality Q and vice versa; Q, and with it the filter type, is determined by the gain setting of A0:

$$Q = \frac{1}{3 - A_0}$$

2.2.3 Higher-Order Low-Pass Filters

Higher-order low-pass filters are required to sharpen a desired filter characteristic. For that purpose, first-order and second-order filter stages are connected in series, so that the product of the individual frequency responses results in the optimized frequency response of the overall filter.

In order to simplify the design of the partial filters, the coefficients ai and bi for each filter type are listed in the Butterworth coefficient table in the Appendix section which provides coefficients for the first 10 filter orders.

2.3 High-Pass Filter Design

The high-pass filter, as shown below has a pass-band where all frequencies above the cutoff frequency (fc) pass with little to no attenuation. Below fc, within the filter's stop-band, the signals are attenuated at ever greater levels as the frequency moves lower

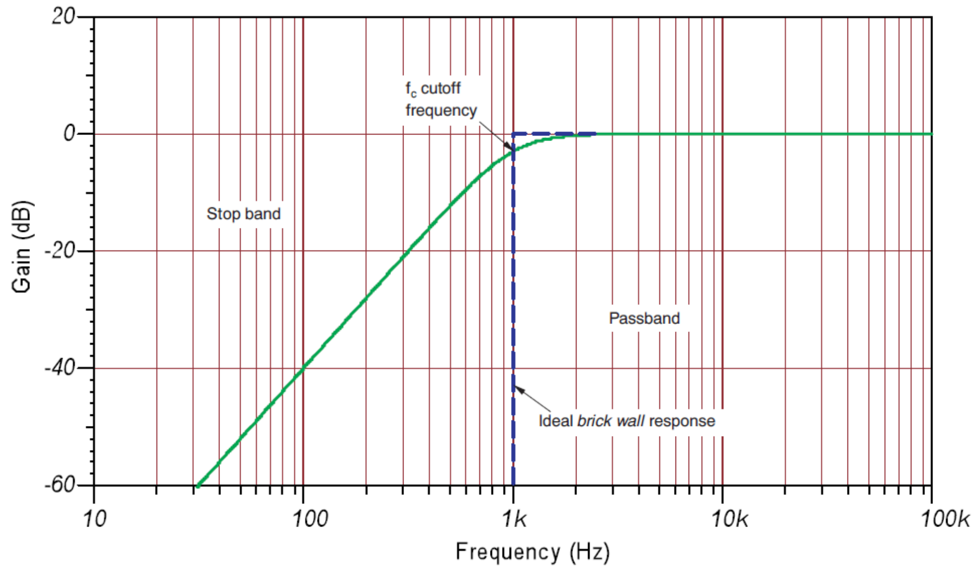

Figure 16: An Example of a High-Pass Filter Response

By replacing the resistors of a low-pass filter with capacitors, and its capacitors with resistors, a high-pass filter is created.

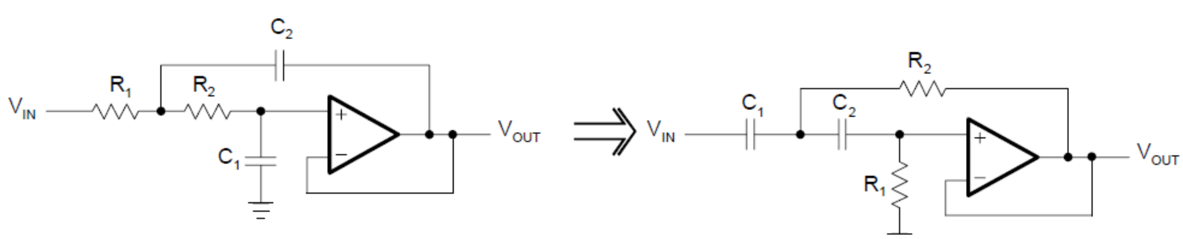

Figure 17: Low-Pass to High-Pass Transition through Component Exchange

To plot the gain response of a high-pass filter, mirror the gain response of a low-pass filter at the corner frequency as shown in Figure 18. This only applies in the case of a two-way crossover network. For a 4-way crossover network, the corner frequency of the Low-pass filter will have to be different from that of the High-Pass filter.

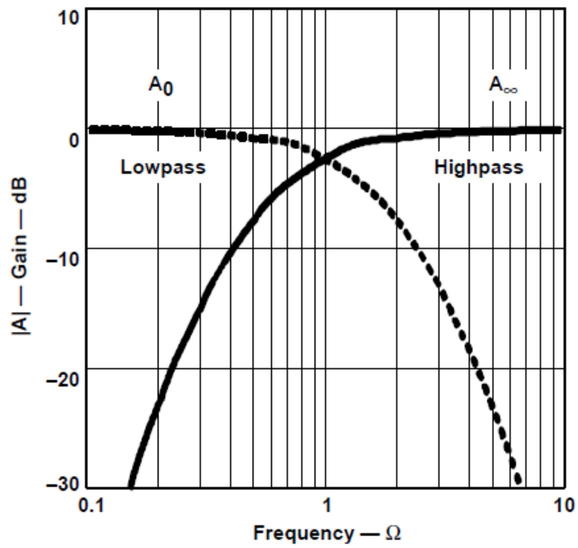

Figure 18: Developing the Gain Response of a High-Pass Filter

The general transfer function of a high-pass filter is then:

$$A(s) = \frac{A_\infty}{\prod_i \left(1 + \frac{a_i}{s} + \frac{b_i}{s^2}\right)}$$

A_∞ is the pass band gain.

2.3.1 First-Order High-Pass Filter

Figure 19 and 20 show a first-order high-pass filter in the non-inverting and the inverting configurations respectively.

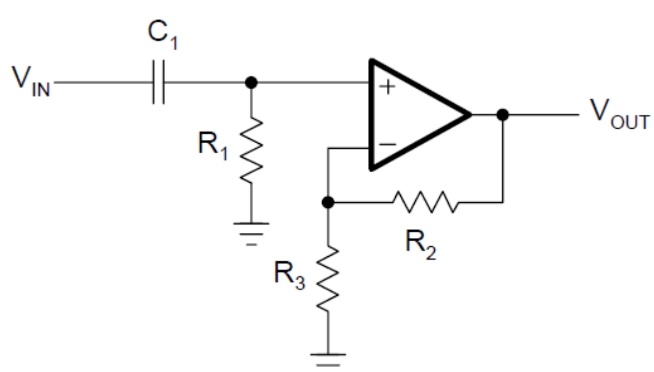

Figure 19: First-Order Non-inverting High-Pass Filter

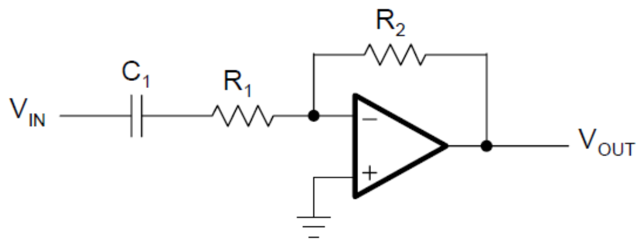

Figure 20: First-Order Inverting High-Pass Filter

With b=0 for all first-order filters, the transfer function of a first-order filter simplifies to:

$$A(s) = \frac{A_\infty}{1 + \frac{a_1}{s}}$$

The transfer functions of the circuits are as follows:
For the non-inverting:

$$A(s) = \frac{1 + \frac{R_2}{R_1}}{1 + \frac{1}{\omega_c R_1 C_1} \cdot \frac{1}{s}}$$

For the inverting:

$$A(s) = \frac{-\frac{R_2}{R_1}}{1 + \frac{1}{\omega_c R_1 C_1} \cdot \frac{1}{s}}$$

The negative sign indicates that the inverting amplifier generates a 180° phase shift from the filter input to the output.

In this paper, however, the non-inverting configuration is used in design to avoid this phase shift.
The coefficient comparison between the non-inverting transfer function and the general transfer function of a first-order filter provides the following pass-band gain factor:

$$A_\infty = 1 + \frac{R_2}{R_1}$$

while the term for the coefficient a_1 is given by:

$$[illegible]$$

To dimension the circuit, specify the corner frequency (fC), the dc gain (A∞), and capacitor (C1), and then solve for R1 and R2:

$$[illegible]$$

$$[illegible]$$

2.3.2 Second-Order High-Pass Filter

High-pass filters use the same Sallen-Key topology as the low-pass filters. The only difference is that the positions of the resistors and the capacitors have swapped positions.

The general Sallen-Key topology in Figure 21 below allows for separate gain setting via A0 = 1 + R4/R3.

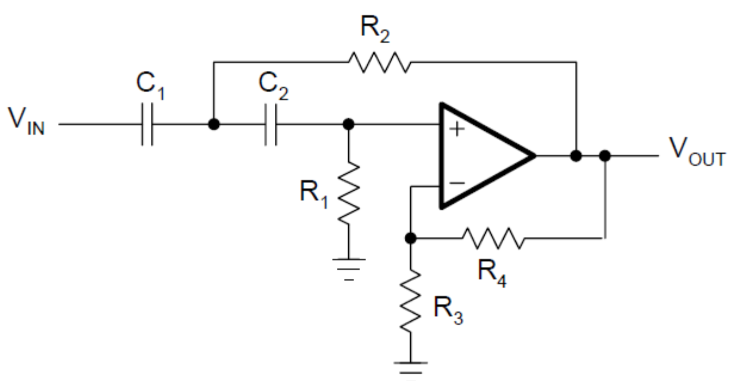

Figure 21: General Sallen-Key High-Pass Filter

The transfer function of the circuit is:

$$[illegible]$$

Where $[illegible]$

The unity-gain topology in the circuit below (Figure 22) is usually applied in low-Q filters with high gain accuracy.

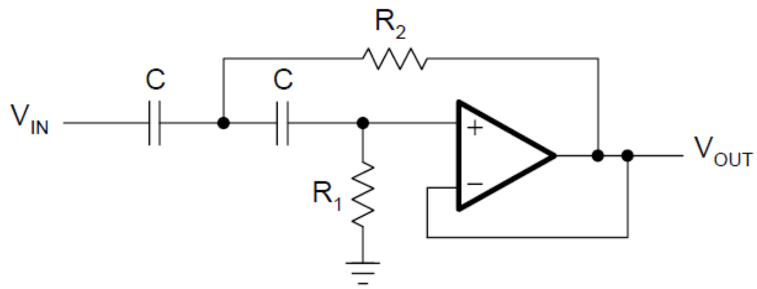

Figure 22: Unity-Gain Sallen-Key High-Pass Filter

To simplify the circuit design, it is common to choose unity-gain (α = 1), and C1 = C2 = C. The transfer function of the above circuit then simplifies to:

$$A(s) = \frac{1}{1 + \frac{2}{\omega_c R_2 C s} + \frac{1}{\omega_c^2 R_1 R_2 C^2 s^2}}$$

The coefficient comparison between this transfer function and the general transfer function of a single stage of a high-pass filter yields:

$$A(s) = \frac{A_\infty}{1 + \frac{a_i}{s} + \frac{b_i}{s^2}}$$

This is the general transfer function, with;

$$A_\infty = 1$$

$$a_i = \frac{2}{\omega_c R_2 C}$$

$$b_i = \frac{1}{\omega_c^2 R_1 R_2 C^2}$$

Given C, the resistor values for R1 and R2 are calculated through:

$$R_2 = \frac{1}{\pi f_c a_i C}$$

$$R_1 = \frac{a_i}{4\pi f_c b_i C}$$

2.3.3 Higher-Order High-Pass Filter

Likewise, as with the low-pass filters, higher-order high-pass filters are designed by cascading first-order and second-order filter stages. The filter coefficients are the same ones used for the low-pass filter design, and are listed in the coefficient table in the Appendix section.

2.4 Band-Pass Filter Design

The band-pass filter has a pass-band that allows a select band of frequencies that fall within the pass-band to pass with little, or no, attenuation. An upper (f_H) and lower (f_L) cutoff frequency define the bandwidth of the filter (f_H to f_L). Frequencies beyond the pass-band lie in the two stop-bands and receive greater attenuation as the frequency moves further away from the pass-band in either direction.

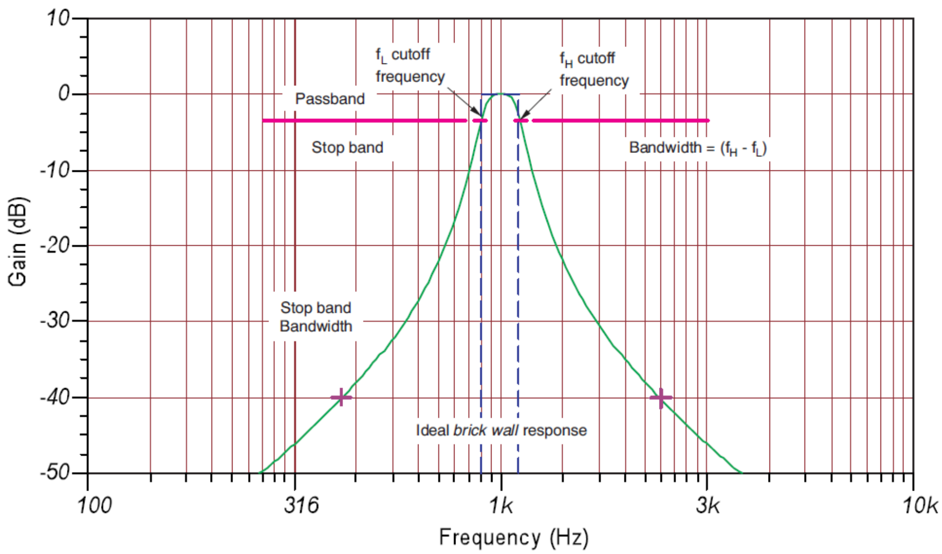

Figure 23: An example of a Band-pass Filter Response

The center frequency fc the frequency at which the maximum gain of the filter occurs. Theoretically, the center frequency is the geometric mean of the two half-power frequencies, fL and fH.
The lower cut-off frequency, fL, is the lower frequency at which the gain is 3dB less than the gain at the center frequency. Similarly, the upper cutoff frequency, fH, is the upper frequency at which the gain is 3dB less than the gain at the center frequency.

For the case of a Band-Pass Filter, the pass-band characteristic of a low-pass filter is transformed into the upper pass-band half of a band-pass filter. The upper pass-band is then mirrored at the center frequency, into the lower pass-band half. This is shown in Figure 24.

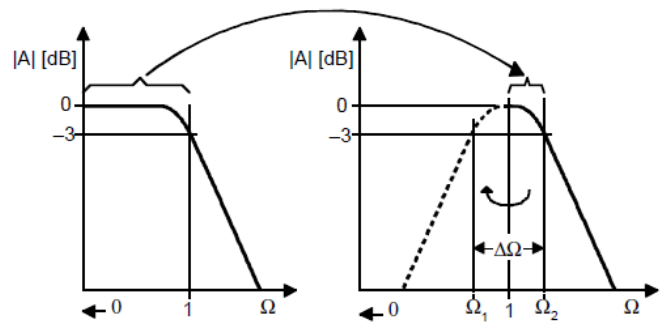

Figure 24: Low-Pass to Band-Pass Transition

The corner frequency of the low-pass filter transforms to the lower and upper –3 dB frequencies of the band-pass, Ω1 (fL) and Ω2 (fH). The difference between both frequencies is defined as the normalized bandwidth ΔΩ.

$$\Delta\Omega = \square_{\square} - \square_{\square}$$

The normalized center frequency, where Q = 1, is:

$$\Omega_{\square} = 1 = \square_{\square} . \square_{\square}$$

In analogy to the resonant circuits, the quality factor Q is defined as the ratio of the center frequency (fc) to the bandwidth (ΔΩ):

$$\square = \frac{\square_{\square}}{\Delta\Omega} = \frac{\square_{\square}}{\square_{\square} - \square_{\square}}$$

An 'ideal' band-pass filter has constant gain within the pass-band, zero gain outside the pass-band, and an abrupt boundary between the two. This response characteristic is impossible to realize in practice, but it can be approximated to varying degrees of accuracy by real filters.

The rate of change of attenuation between the pass-band and the stop-band also differs from one filter to the next. The slope of the curve in this region depends strongly on the order of the filter, with higher-order filters having steeper cut-off slopes.

Band-pass filters are used in electronic systems to separate a signal at one frequency or within a band of frequencies from signals at other frequencies.
In comparison to wideband filters, narrow-band filters of higher order consist of cascaded second order band-pass-filters that use the Sallen key or multiple feedback (MFB) topology. At high Q and Gains the Sallen type may not be very stable.

The simplest design of a band-pass filter is the connection of a high-pass filter and a low-pass filter in series, which is commonly done in wide-band filter applications. Thus, a first order high-pass and a first-order low-pass provide a second-order band-pass, while a second-order high-pass and a second-order low-pass result in a fourth-order band-pass response.

2.4.1 Second-Order Band-Pass Filter

To develop the frequency response of a second-order band-pass filter, apply the transformation equation;

$$\frac{1}{\Delta\Omega}\left(s + \frac{1}{s}\right)$$

to a first-order low-pass transfer function:

$$A(s) = \frac{A_0}{1 + s}$$

This yields the general transfer function for a second-order band-pass filter:

$$A(s) = \frac{A_0 . \Delta\Omega . s}{1 + \Delta\Omega . s + s^2}$$

When designing band-pass filters, the parameters of interest are the gain at the mid frequency (A_m) and the quality factor (Q), which represents the selectivity of a band-pass filter. Therefore, replacing A_0 with A_m and ΔΩwith 1/Q, we obtain:

$$A(s) = \frac{\frac{A_m}{Q} . s}{1 + \frac{1}{Q} . s + s^2}$$

The figure 25 below shows the normalized gain response of a second-order band-pass filter for different Qs.

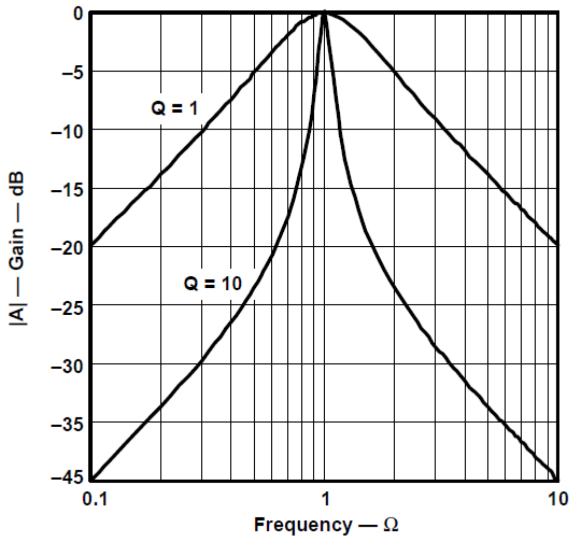

Figure 25: Gain-Response of a Second-Order Band-Pass Filter

The graph shows that the frequency response of second-order band-pass filters gets steeper with rising Q, thus making the filter more selective

The analysis of a second order Sallen-Key Band-Pass circuit is greatly simplified if all frequency-determining capacitors are set equal to each other, as well as all frequency-determining resistors. Under these conditions, the filter is sometimes referred to as Equal Component Sallen-Key Filter. Thus, taking advantage of this trick by letting

$$R_1 = R_2 = \frac{R_f}{2} = R \text{ and } C_1 = C_2 = C$$

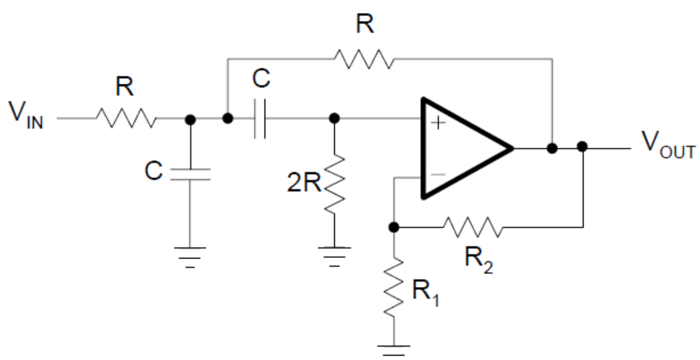

Figure 26: Sallen-Key Band-Pass

The Sallen-Key band-pass circuit in Figure 26 has the following transfer function:

$$A(s) = \frac{\frac{G}{RC} \cdot s}{s^2 + \left(\frac{3-G}{RC}\right)s + \left(\frac{1}{RC}\right)^2}$$

Where

Center frequency: $f_m = \frac{1}{2\pi RC}$

Inner Gain: $G = 1 + \frac{R_2}{R_1}$

Gain at f_c: $A_m = \frac{G}{3-G}$

Quality factor: $Q = \frac{1}{3-G}$

The Sallen-Key circuit has the advantage that the quality factor (Q) can be varied via the inner gain (G) without modifying the mid frequency (fm). A drawback is, however, that Q and Am cannot be adjusted independently.

Care must be taken when G approaches the value of 3, because then A_m becomes infinite and causes the circuit to oscillate.

To set the mid frequency of the band-pass, specify fm and C and then solve for R:

$$R = \frac{1}{2\pi f_m C}$$

Because of the dependency between Q and Am, there are two options to solve for R2: either to set the gain at mid frequency:

$$R_2 = \frac{2A_m - 1}{1 + A_m}$$

or to design for a specified Q:

$$R_2 = \frac{2Q - 1}{Q}$$

2.4.2 Fourth-Order Band-Pass Filter (Staggered Tuning)

It was previously shown that the frequency response of second-order band-pass filters gets steeper with rising Q. However, there are band-pass applications that require a flat gain response close to the mid frequency as well as a sharp pass-band-to-stop-band transition. These tasks can be accomplished by higher-order band-pass filters.

Of particular interest is the application of the low-pass to band-pass transformation onto a second-order low-pass filter, since it leads to a fourth-order band-pass filter.
Replacing the S term in Equation 16–2 with Equation 16–7 gives the general transfer function of a fourth-order band-pass:

$$A(s) = \frac{\frac{s^2 A_m (\Delta\Omega)^2}{b_1}}{1 + \frac{a_1}{b_1}\Delta\Omega . s + \left[2 + \frac{\Delta\Omega^2}{b_1}\right] . s^2 + \frac{a_1}{b_1}\Delta\Omega . s^3 + s^4}$$

Similar to the low-pass filters, the fourth-order transfer function is split into two second-order band-pass terms. Further mathematical modifications yields:

$$A(s) = \frac{\frac{A_{mi}}{Q_i} . \alpha s}{\left[1 + \frac{\alpha s}{Q_i} + (\alpha s)^2\right]} . \frac{\frac{A_{mi}}{Q_i} . \frac{s}{\alpha}}{\left[1 + \frac{1}{Q_i}\left(\frac{s}{\alpha}\right) + \left(\frac{s}{\alpha}\right)^2\right]}$$

This equation represents the connection of two second-order band-pass filters in series, where

- A_{mi} is the gain at the mid frequency, f_{mi}, of each partial filter
- Qi is the pole quality of each filter
- α and 1/αare the factors by which the mid frequencies of the individual filters, fm1 and fm2, derive from the mid frequency, fm, of the overall band-pass.

In a fourth-order band-pass filter with high Q, the mid frequencies of the two partial filters differ only slightly from the overall mid frequency. This method is called staggered tuning.

Factor α needs to be determined through successive approximation, using equation

$$\alpha^2 + \left[\frac{\alpha . \Delta\Omega . a_1}{b_1(1 + \alpha^2)}\right]^2 + \frac{1}{\alpha^2} - 2 - \frac{(\Delta\Omega)^2}{b_1} = 0$$

With a1 and b1 being the second-order low-pass filter coefficients of the desired filter type, and ΔΩ = 1⁄QBP the overall quality factor of the filter.

To simplify the filter design, Table 1 lists those coefficients, and provides the α values for three different quality factors, Q = 1, Q = 10, and Q = 100.

Bessel				Butterworth				Tschebyscheff			
a_1	1.3617			a_1	1.4142			a_1	1.0650		
b_1	0.6180			b_1	1.0000			b_1	1.9305		
Q	100	10	1	**Q**	100	10	1	**Q**	100	10	1
ΔΩ	**0.01**	**0.1**	**1**	ΔΩ	**0.01**	**0.1**	**1**	ΔΩ	**0.01**	**0.1**	**1**
α	1.0032	1.0324	1.438	α	1.0035	1.036	1.4426	α	1.0033	1.0338	1.39

Table 1: Values of α for Different Filter types and Different Qs

After α has been determined, all quantities of the partial filters can be calculated using the following equations:

The mid-frequency of filter 1 is:

$$f_{m1} = \frac{f_m}{\alpha}$$

The mid-frequency of filter 2 is:

$$f_{m2} = f_m \cdot \alpha$$

With fm being the mid-frequency of the overall fourth-order band-pass filter.
The individual pole quality, Qi is the same for both filters:

$$Q_i = Q \frac{(1 + \alpha^2) b_1}{\alpha . a_1}$$

With Q being the quality factor of the overall filter.
The individual gain (Ami) at the partial mid frequencies, fm1 and fm2, is the same for both filters:

$$A_{mi} = \frac{Q_i}{Q} . \sqrt{\frac{A_m}{b_1}}$$

With Ambeing the gain at the mid frequency fm of the overall filter.

2.5 Practical Design Recommendations

2.5.1 Op Amp Selection

The most important op amp parameter for proper filter functionality is the unity-gain bandwidth. In general, the open-loop gain (AOL) should be 100 times (40 dB above) the peak gain (Q) of a filter section to allow a maximum gain error of 1%.

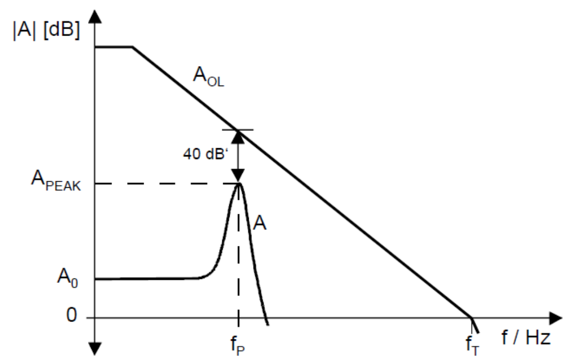

Figure 27:Open-Loop Gain (AOL) and Filter Response (A)

The following equations are good rules of thumb to determine the necessary unity-gain bandwidth of an op amp for an individual filter section.

First-order filter:

$$f_T = 100 \cdot Gain \cdot f_C$$

Second-order filter (Q < 1):

$$f_T = 100 \cdot Gain \cdot f_C \cdot k_i \quad \text{with} \quad k_i = \frac{f_{Ci}}{f_C}$$

Second-order filter (Q > 1):

$$f_T = 100 \cdot Gain \cdot \frac{f_C}{a_i} \sqrt{\frac{Q_i^2 - 0.5}{Q_i^2 - 0.25}}$$

Texas Instruments offers a wide range of op amps for high-performance filters in single supply applications.

2.5.2 Capacitor and Resistor Selection

The tolerance of the selected capacitors and resistors depends on the filter sensitivity and on the filter performance.

Sensitivity is the measure of the vulnerability of a filter's performance to changes in component values. The important filter parameters to consider are the corner frequency, fC, and Q.

For example, when Q changes by $\pm$ 2% due to a $\pm$ 5% change in the capacitance value, then the sensitivity of Q to capacity changes is expressed as:

$$S^{Q}_{C} = \frac{2\%}{5\%} = 0.4\frac{\%}{\%}$$

The following sensitivity approximations apply to second-order Sallen-Key and MFB filters:

$$S^{Q}_{R} \approx S^{Q}_{C} \approx S^{f_C}_{R} \approx S^{f_C}_{C} \approx \pm 0.5\frac{\%}{\%}$$

Although 0.5 %/% is a small difference from the ideal parameter, in the case of higher-order filters, the combination of small Q and fC differences in each partial filter can significantly modify the overall filter response from its intended characteristic.

To minimize the variations of fC and Q, NPO (COG) ceramic capacitors are recommended for high-performance filters. These capacitors hold their nominal value over a wide temperature and voltage range. The various temperature characteristics of ceramic capacitors are identified by a three-symbol code such as: COG, X7R, Z5U, and Y5V.

COG-type ceramic capacitors are the most precise. Their nominal values range from 0.5 pF to approximately 47 nF with initial tolerances from $\pm$ 0.25 pF for smaller values and up to $\pm$ 1% for higher values. Their capacitance drift over temperature is typically 30ppm/°C.

X7R-type ceramic capacitors range from 100 pF to 2.2 µF with an initial tolerance of +1% and a capacitance drift over temperature of $\pm$15%.
For higher values, tantalum electrolytic capacitors should be used.

Other precision capacitors are silver mica, metallized polycarbonate, and for high temperatures, polypropylene or polystyrene.

Since capacitor values are not as finely subdivided as resistor values, the capacitor values should be defined prior to selecting resistors. If precision capacitors are not available to provide an accurate filter response, then it is necessary to measure the individual capacitor values, and to calculate the resistors accordingly.

For high performance filters, 0.1% resistors are recommended.

2.5.3 Component Values

Resistor values should stay within the range of 1 kΩ to 100 kΩ. The lower limit avoids excessive current draw from the op amp output, which is particularly important for single-supply op amps in power-sensitive applications. Those amplifiers have typical output currents of between 1 mA and 5 mA. At a supply voltage of 5 V, this current translates to a minimum of 1 kΩ.

The upper limit of 100 kΩ is to avoid excessive resistor noise.

Capacitor values can range from 1 nF to several µF. The lower limit avoids coming too close to parasitic capacitances. If the common-mode input capacitance of the op amp, used in a Sallen-Key filter section, is close to 0.25% of C1, (C1 / 400), it must be considered for accurate filter response. The MFB topology, in comparison, does not require input-capacitance compensation.

CHAPTER 3

DESIGN

3.1 Design Specifications

A four way active filter network is desired with the following characteristics.

a) A 3rd Order Low Pass Filter with a cut-off frequency of 1kHz.
b) A 2nd Order Band Pass Filter with a Mid-frequency of 1.875 kHz.
c) A 2nd Order Band Pass Filter with a Mid-frequency of 4.5 kHz.
d) A 3rd Order High Pass Filter with a cut-off frequency of 6 kHz.

All of the above networks are Butterworth types of filters with Sallen-Key configuration. Both the Low Pass and the High Pass filters are of unity-gain Sallen-Key topology while the two Band Pass filters each have their own gains according to design

The power supply to be designed is required to operate from both AC and DC voltages (12V) and the amplifier to be connected to the output of each network is a 14W monolithic integrated circuit in penta-watt package known as the TDA2030.

3.2 Power Supply Design

A 3-terminal output transformer is the most suitable in obtaining both positive and negative voltages required to power up different parts of the circuitry. A 15V 0 15V, 2A transformer was used in this particular design. The full bridge rectifier was constructed using four 1N5401 power diodes due to their ability to pass more than 1A of current.

Two smoothing capacitors are used to smoothen the resulting DC voltages from the rectifier and large values are recommended. (C1 and C2 are both 6800uF). Input supply capacitor is recommended for filtering noise on the input and the output supply decoupling capacitor for stabilizing the output.

Two voltage regulators (LM78xx series and LM79xx series) were used as shown below, one for the positive voltage rail and another for the negative voltage rail. Each of these regulators can deliver up to 1.5 A of output current. The internal current-limiting and thermal-shutdown features of these regulators essentially make them immune to overload.Heat sinks were used on the 7815 and 7915 ICs, each one with its own isolated from the other.

In many cases, a regulator powers a load that is not connected to ground but, instead, is connected to a voltage source of opposite polarity (e.g., operational amplifiers, level-shifting circuits, etc.). In these cases, a clamp diode should be connected to the regulator output as shown with D7 and D8 in Figure 28. This protects the regulator from output polarity reversals during startup and short-circuit operation.

Occasionally, the input voltage to the regulator can collapse faster than the output voltage. This can occur, for example, when the input supply is crowbarred during an output overvoltage condition. If the output voltage is greater than approximately 7 V, the emitter-base junction of the series-pass element (internal or external) could break down and be damaged. To prevent this, a diode shunt can be used as shown in the Figure 28 by diodes D5 and D6.

Different orderable part numbers will be able to tolerate different levels of voltage. It is also recommended to have a decoupling capacitor on the output of both of the devices' power supply to limit noise on the device input i.e. the C5 and C6 capacitors.

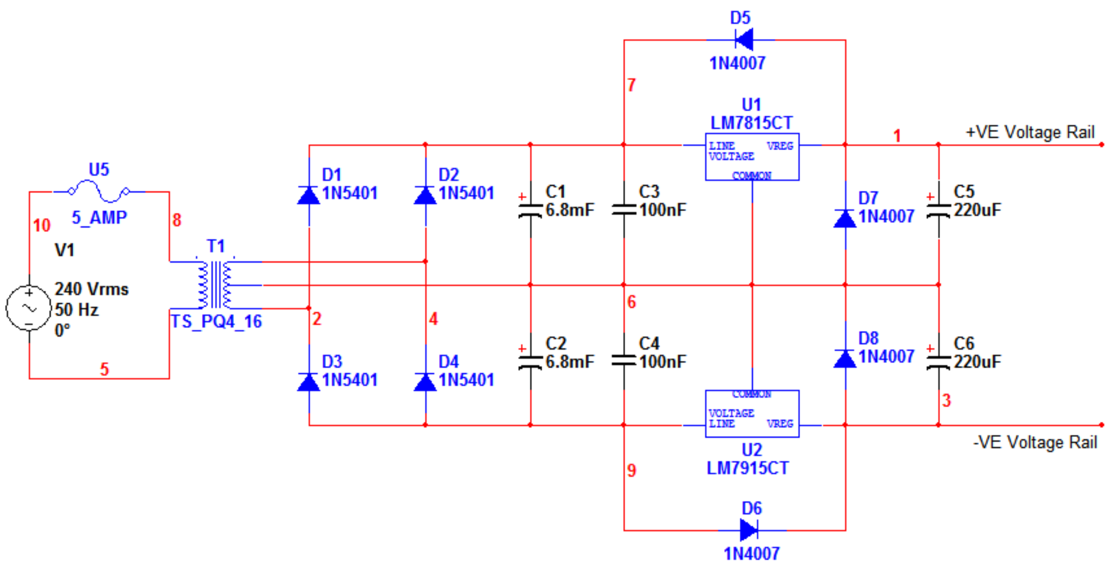

Figure 28: Power Supply Schematic as shown in Multisim

At the output of each of the regulators, an external dc source can be connected as necessary to cater for the requirement of both AC and DC supply of power. This is shown by the terminal blocks J5 and J8 in Figure 29 for the positive and negative voltage rails respectively. For the positive voltage regulator output, the positive of the dc source is to be connected to the positive voltage rail and the negative, to the ground. For the negative voltage regulator output, the positive of the dc source is to be connected to ground and the negative is to be connected to the negative voltage rail.

The UF5400 series power diodes are connected in each of the external dc source terminals as shown in Figure 29 in order to prevent current flowing into the battery in case both the AC and DC sources are connected to the system simultaneously. D7 and D8 represent this.

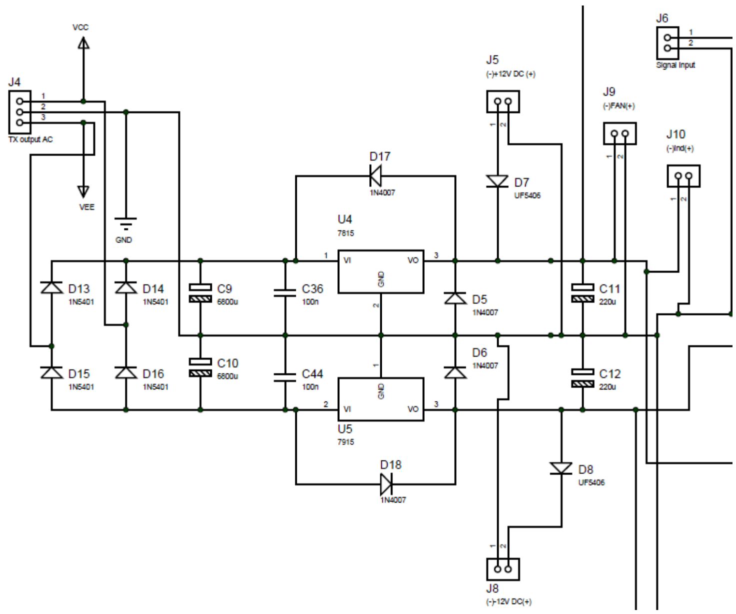

Figure 29: Power Supply as shown in Proteus

3.3 Filter Designs

The LM358, Fig. 31 consisting of two independent, high gain, internally frequency compensated operational amplifiers was used to implement the Band-Pass 1 and Band-Pass 2 filters.

The LM324, Fig. 30,a quadruple high-gain, frequency-compensated operational amplifier, implements the Low-Pass Filter and the High-Pass Filters.

The selection of these two operational amplifiers was influenced by their ability to operate on a split power supply as well as compact design consideration.

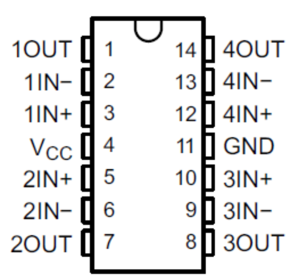

Figure 30: LM324 Top View

(Texas Instruments - Quadruple Operational Amplifiers, SEPTEMBER 1975–REVISED JANUARY 2014)

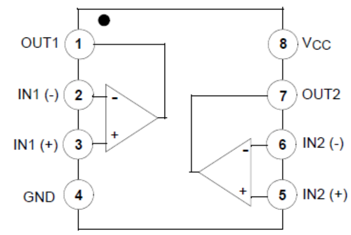

Figure 31: Internal Block Diagram of LM358

(Fairchild Semiconductors, 2010 Fairchild Semiconductor Corporation)

3.3.1 Low Pass Filter Design

This was designed using a third order low pass filter of Butterworth type. It constitutes of two filters; a first and second-order filters cascaded.
The cut-off frequency was set at 1 kHz to achieve the maximally flat pass-band response and a 60db/decade roll-off from pass-band to stop-band.

For the first order filter: $a_1 = 1$, $b_1 = 0$
Take C_1= 22nF standard value, then;

$$R_1 = \frac{a_1}{2\pi f_c C_1} = \frac{1}{2\pi \times 1000 \times 22 \times 10^{-9}} = 7234 \approx 7.2k$$

7.5k is the equivalent standard value.

For the second order filter, the Sallen-Key, unity-gain Topology was chosen for high gain accuracy and low Q (Q<3).

From the Butterworth coefficient table in the appendix: $a_2 = 1$, $b_2 = 1$
Take C_2 = 22nF standard value, then;

$$R_{2,3} = \frac{a_2 C_3 \mp \sqrt{a_2^2 C_3^2 - 4 b_2 C_2 C_3}}{4\pi f_c C_2 C_3} \qquad \text{and} \qquad C_3 \geq C_2 \frac{4 b_2}{a_2^2}$$

$$C_3 \geq 22nF \times \frac{4}{1} = 88nF \cong 100nF$$

The $R_{2,3}$ equation above means that R_2 takes the negative sign and R_3 takes the positive sign.
The closest standard capacitor value that could be obtained for C_3 was 100nF. Therefore;

[illegible equation]

[illegible equation]

The closest standard resistor values that could be obtained for R_2 and R_3were 2.4k and 4.5k respectively.

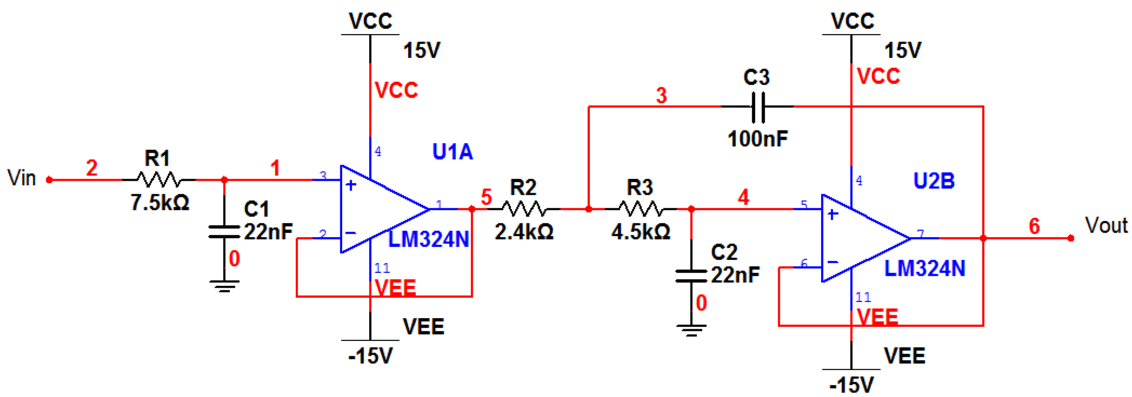

Figure 32:3rd Order Active Low-Pass Filter

3.3.2 High Pass Filter Design

Like the Low-Pass filter, the High-Pass Filter design was of the third order Butterworth type and constituted of two filters; a first and second-order filters cascaded.
The cut-off frequency was set at 6 kHz.

For the first order filter: $a_1 = 1$, $b_1 = 0$
Take C_4= 22nF standard capacitor value, then;

[illegible equation]

R_4 standard value is given by 1.2k

Similar to the Low Pass Filter design, the Sallen-Key, unity-gain Topology was chosen for the second order filter. From the Butterworth coefficient table in the appendix: $a_2 = 1$, $b_2 = 1$

Take $C_5 = C_6 = C = 22nF$ standard capacitor values, then;

[illegible]

[illegible]

Standard values for R5 and R6 are 620Ω and 2.4k respectively.

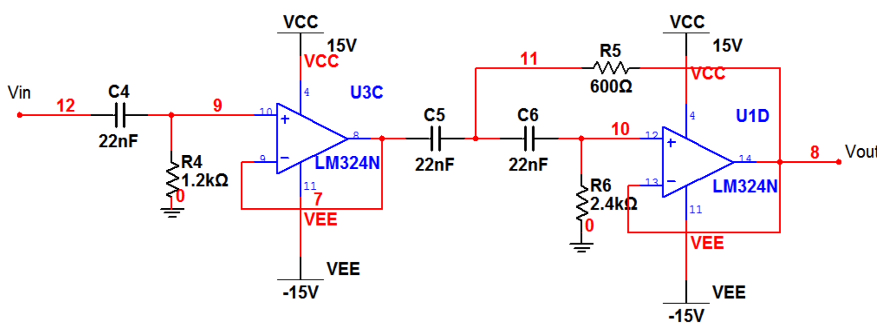

Figure 33: Active High-Pass Filter

3.3.3 Bandpass-1 Filter Design (Mid-1)

This is a 2nd order Sallen-Key Topology Band-pass filter with a specified mid frequency of 1.875 kHz. Bandwidth is 2250Hz and C is 100nF standard value.

[illegible]

[illegible]

[illegible]

[illegible]

[illegible]

[illegible]

[illegible]

[illegible]

[illegible]

For convenience, R_6 and R_7 were chosen as 1k to give a gain of at least 2.

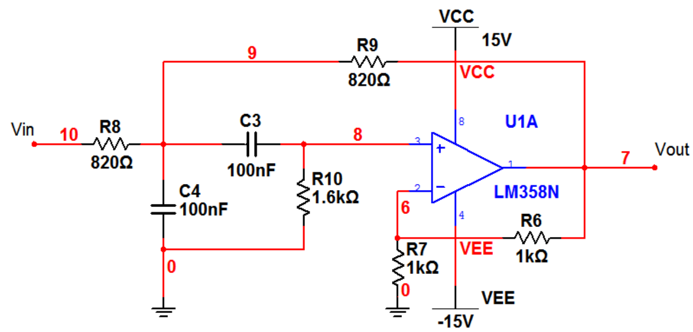

Figure 34: Active Band pass Filter 1

3.3.4 Bandpass-2 Filter Design (Mid-2)

Likewise, this is also a 2nd order Sallen-Key Topology Band-pass filter. Specified mid frequency is 4.5 kHz. Bandwidth is 3 kHz and C is chosen as 100nF for convenience of purchasing components.

[illegible]

[illegible]

[illegible]

[illegible]

[illegible]

$$□ = \frac{□}{□ - □} \rightarrow □ = \frac{□□ - □}{□} = \frac{□□□.□□ - □}{□.□} = □.□□$$

$$□□□□\ □_{□} = □□, □□□□$$

$$□ = □ + \frac{□_{□}}{□_{□}} \rightarrow □.□□ = □ + \frac{□_{□}}{□□} \rightarrow □_{□} = □.□□□$$

$$□ = \frac{□}{□□ □_{□} □} = \frac{□}{□□ \times □□□□ \times □□□ \times □□^{□□}} = □□□.□□ \cong □□□□$$

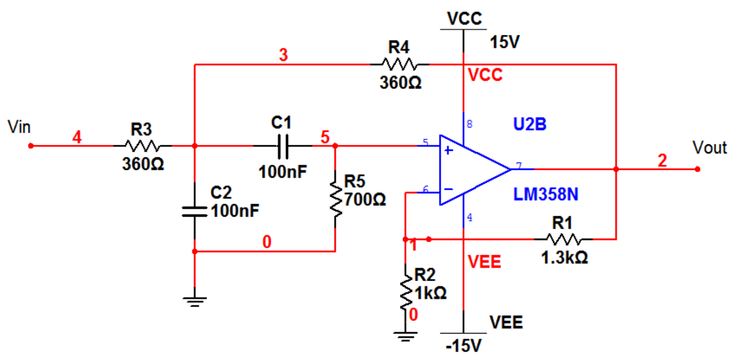

Figure 35: Active Band pass Filter 2

R was approximated at 360 Ω standard value. R_2 was taken as 1k for convenience of purchasing components.

3.4 TDA2030Power Amplifier Design

The TDA2030 is a monolithic integrated circuit in Pentawatt package, intended for use as a low frequency class AB amplifier. It has the following features

* Very low external component required.
* High current output and high operating voltage.
* Low harmonic and crossover distortion.
* Built-in Over temperature protection.
* Short circuit protection between all pins.
* Safety Operating Area for output transistors.

It is essential to take into account all the working conditions, in particular mains fluctuations and supply voltage variations with and without load. The TDA2030(V_{smax}=44V) is particularly suitable for substitution of the standard IC power amplifiers (with V_{smax}=36V) for more reliable applications. Typically it provides14W output power (d = 0.5%) at 14V/4W; at ± 14V or 28V, the guaranteed output power is 12W on a 4Ω load and 8W on an 8Ω load.

A regulated supply is not usually used for the power output stages because of its dimensioning must be done taking into account the power to supply in signal peaks. They are not only a small percentage of the total music signal, with consequently large over-dimensioning of the circuit.
Even if with a regulated supply higher output power can be obtained(V_s is constant in all working conditions),the additional cost and power dissipation do not usually justify its use. Using non-regulated supplies, there are fewer designee restriction. In fact, when signal peaks are present, the capacitor filter acts as a flywheel supplying therequired energy.
In average conditions, the continuous power supplied is lower. The music power/continuous power ratio is greater in case than for the case of regulated supplied, with space saving and cost reduction.

The TDA2030has an original circuit which limits the current of the output transistors. This function can be considered as being peak power limiting rather than simple current limiting. It reduces the possibility that the device gets damaged during an accidental short circuit from AC output to Ground. The device incorporates an original (and patented) short circuit protection system comprising an arrangement for automatically limiting the dissipated power so as to keep the working point of the output transistors within their safe operating area. A conventional thermal shut-down system is also included. The presence of a thermal limiting circuit offers the following advantages:

1) An overload on the output (even if it is permanent),or an above limit ambient temperature can be easily supported since the Tj cannot be higher than 150°C
2) The heat sink can have a smaller factor of safety compared with that of a congenital circuit, there is no possibility of device damage due to high junction temperature

increase up to 150°C, the thermal shut-down simply reduces the power dissipation and the current consumption.

(SGS-Thomson Microelectronics GROUP OF COMPANIES- TDA2030, 1994)

This device is used in between each crossover network and its respective load (Speaker).

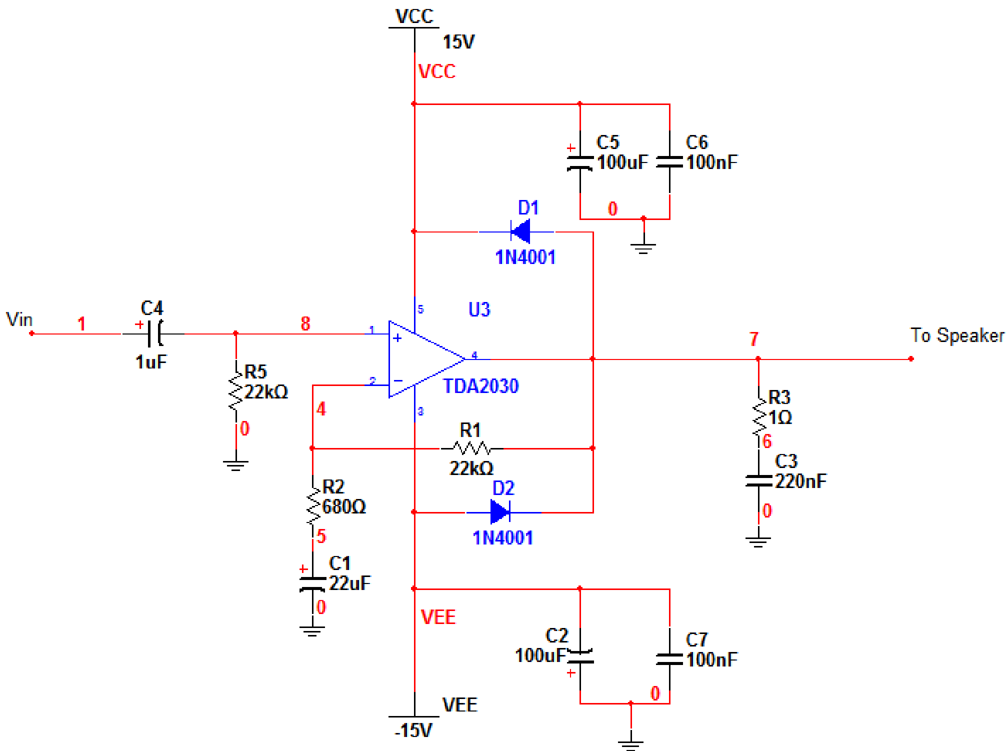

Figure 36: TDA2030 Circuitry

The C4 value in the circuit is replaced with a 1 nF NPO capacitor for the High Pass Filter Stage.

The following table shows the recommended values for the design of a dual supply TDA2030 power amplifier.

Table 2: TDA2030 dual supply component functions

(SGS-Thomson Microelectronics GROUP OF COMPANIES- TDA2030, 1994)

Component	Recomm. value	Purpose	Larger than recommended value	Smaller than recommended value
R1	22 kΩ	Closed loop gain setting	Increase of gain	Decrease of gain (*)
R2	680 Ω	Closed loop gain setting	Decrease of gain (*)	Increase of gain
R3	22 kΩ	Non inverting input biasing	Increase of input impedance	Decrease of input impedance
R4	1 Ω	Frequency stability	Danger of osccilat. at high frequencies with induct. loads	
R5	≅ 3 R2	Upper frequency cutoff	Poor high frequencies attenuation	Danger of oscillation
C1	1 µF	Input DC decoupling		Increase of low frequencies cutoff
C2	22 µF	Inverting DC decoupling		Increase of low frequencies cutoff
C3, C4	0.1 µF	Supply voltage bypass		Danger of oscillation
C5, C6	100 µF	Supply voltage bypass		Danger of oscillation
C7	0.22 µF	Frequency stability		Danger of oscillation
C8	$\cong \frac{1}{2\pi \text{ B R1}}$	Upper frequency cutoff	Smaller bandwidth	Larger bandwidth
D1, D2	1N4001	To protect the device against output voltage spikes		

(*) Closed loop gain must be higher than 24dB

3.5 Printed Circuit Board Design

The whole circuit schematic was drawn and the PCB designed using Proteus 8 Proffessional. All the components were arranged on the board as shown in Figure 37. Allowance of heat sinks as well as terminal block connectors were taken into consideration.

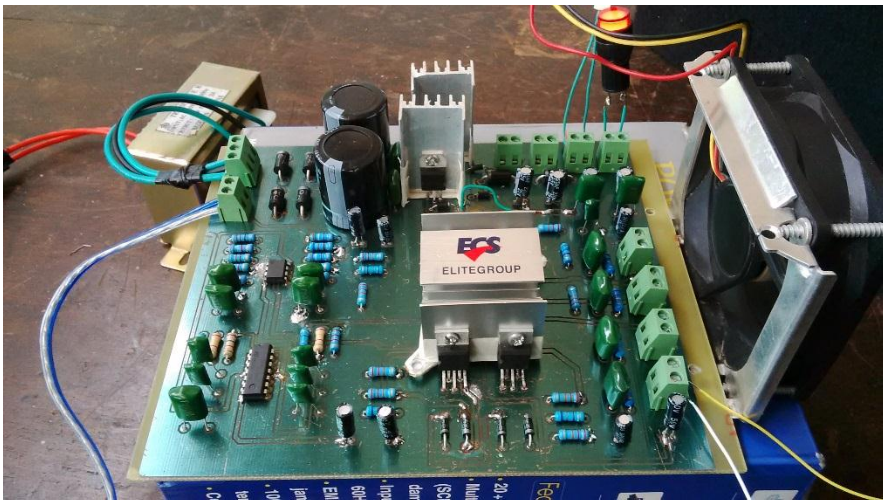

Figure 37: Completely built circuit with a cooling fan

All the components were soldered on the board according to the schematic on Appendix 2. Two different heat sinks were connected to each of the voltage regulators separately to avoid shorting.

One heat sink was used on the four TDA2030 ICs for the purpose of a compact design and to save on space. An additional cooling fan was installed in order to cater for the heat dissipated by each of these four ICs.

Appropriate troubleshooting was done in regards to looking for tracks shorted together as well as making sure components are soldered properly onto their respective tracks.

CHAPTER 4

TESTING AND EXPERIMENTATION

4.1 Computer Simulation Results

NI Multisim version 10 was used to draw and simulate the four cross-over networks. Bode plots were used to check the gain response performance of each of the circuits as shown below.

Low Pass Filter (Bass Band)

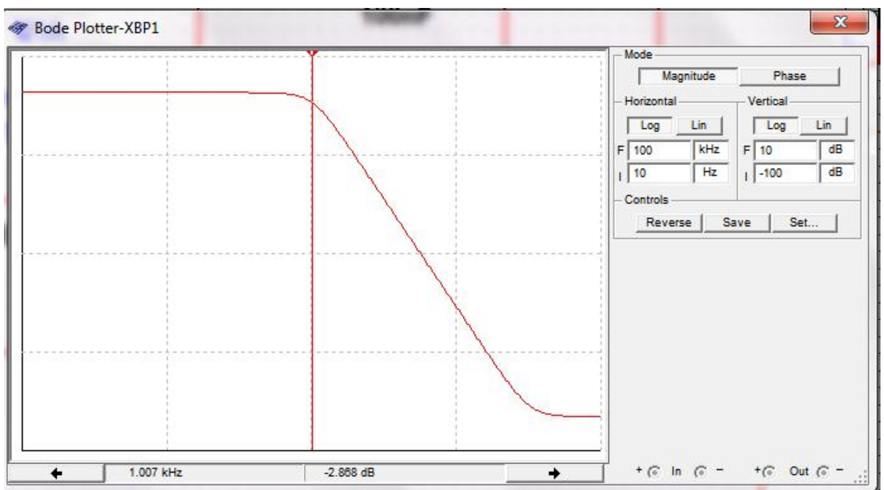

Figure 38: Gain Response of a 3rd Order LPF (Fc = 1 kHz)

Mid-Range Band 1 (Band Pass)

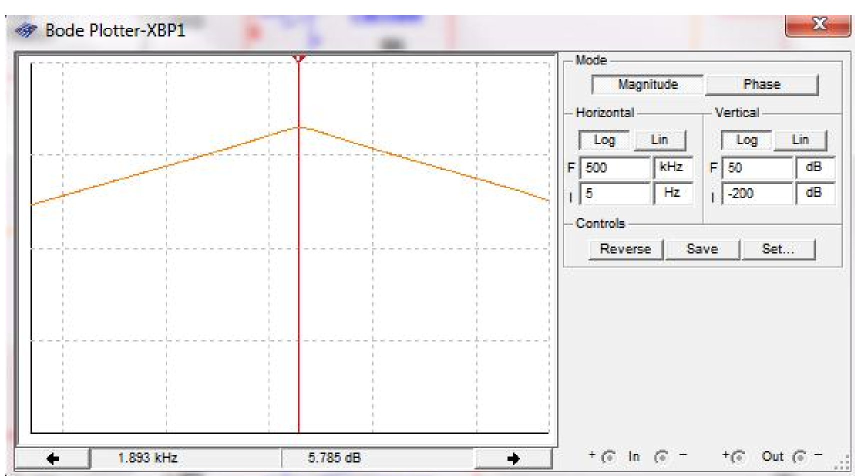

Figure 39: Gain Response of a 2nd Order BPF (Fm = 1.875 kHz)

Mid-Range Band 2 (Band Pass)

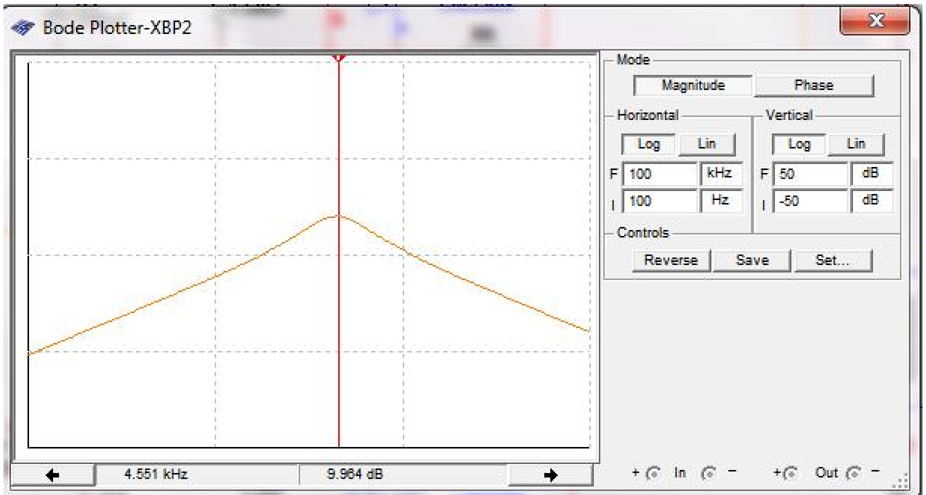

Figure 40: Gain Response of a 2nd Order BPF (Fm = 4.5 kHz)

High Pass Filter (Tweeter Band)

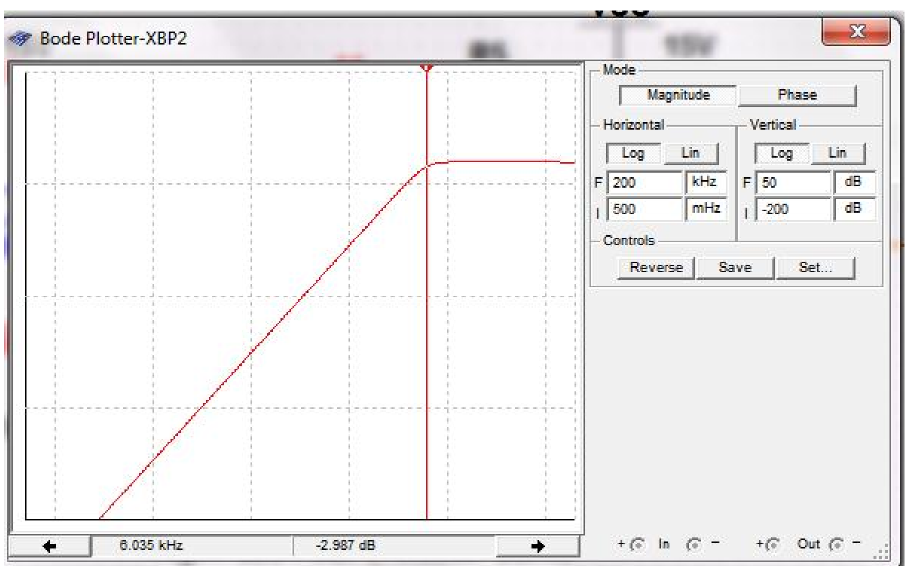

Figure 41: Gain Response of a 3rd Order HPF (Fc = 6 kHz)

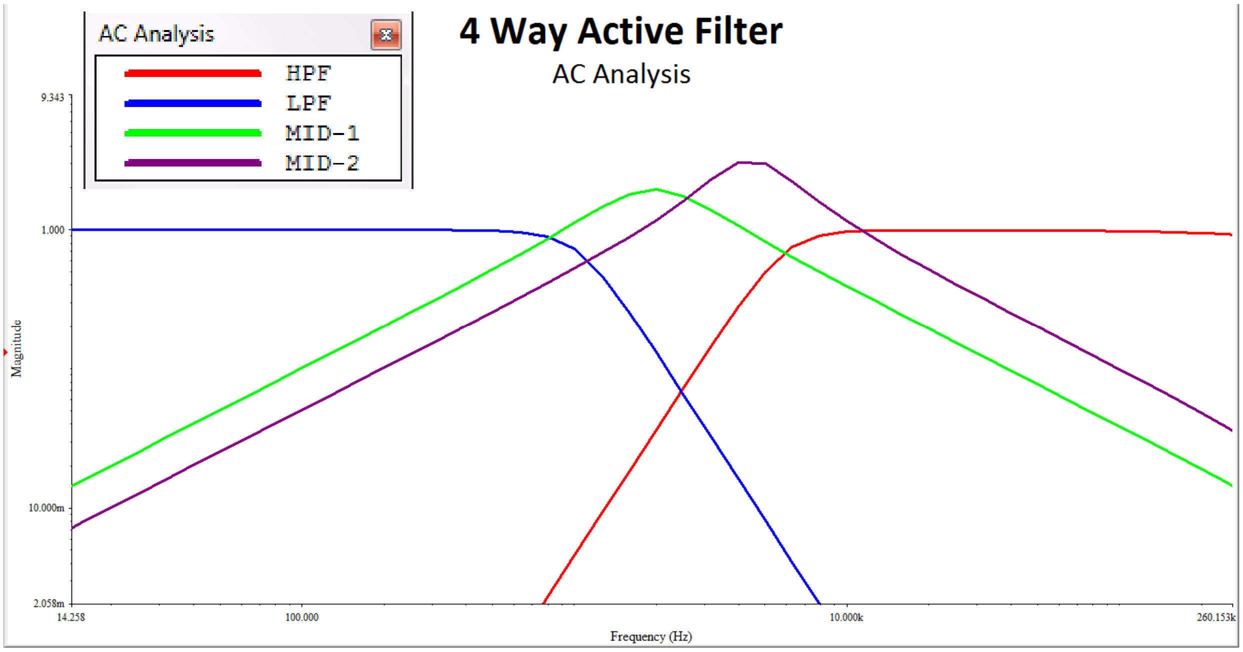

Figure 42: Gain Response comparisons of the four networks as analyzed by Multisim.

4.2 Experimental Results

A signal generator was connected to the input of the circuit and a CRO at the output as shown below. The signal generator was adjusted such that it gave out a voltage of 6 divisions peak-to-peak at 0.2v/div. This translated to an input voltage of 1.2 Volts peak-to-peak.

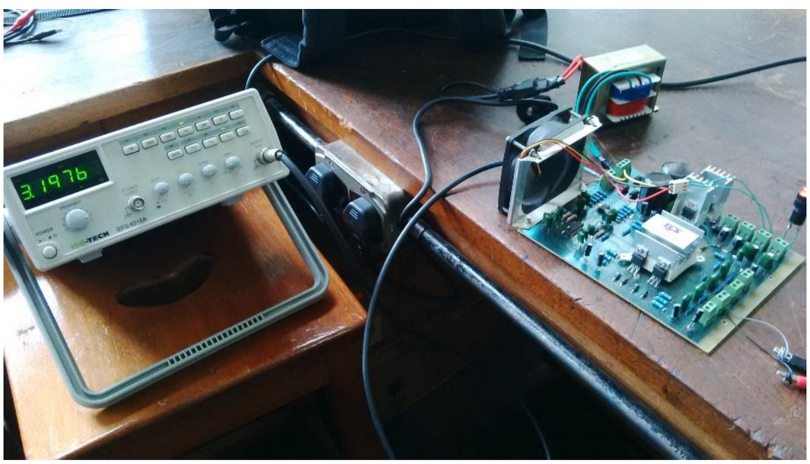

Figure 43: Signal Input connected to a Signal Generator

The frequency of the signal generator was varied and for each frequency step, the output at each of the four networks was measured. The results were tabulated as shown in the tables below.

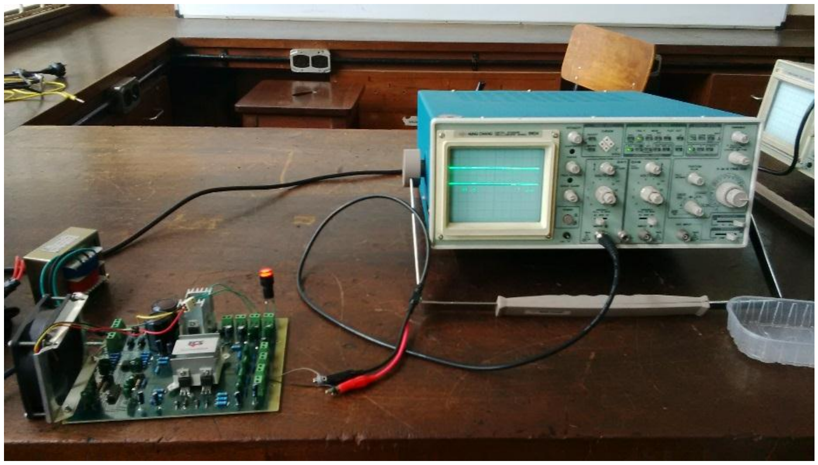

Figure 44: LPF Output connected to a CRO

Graphs of the results of each of the four networks was also plotted.
The following tables and graphs show how the magnitudes of each of the crossover networks vary with different frequencies.

Low Pass Filter

Table 3: Low Pass Filter Magnitudes

Frequency (Hz)	Output Voltage (Volts) (Peak-to-peak)
50	10
100	10
200	10
400	10
600	9.6
800	9.6
1000	9.4
1200	9.6
1400	8.4
1600	6.5
1800	5.0
2000	3.5

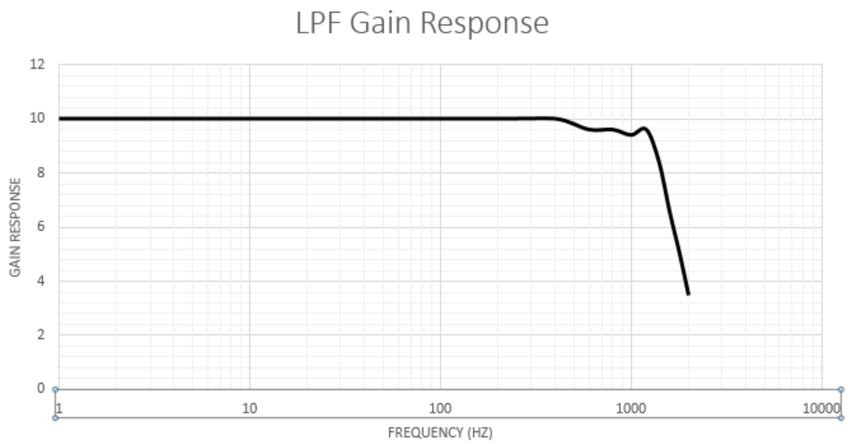

Figure 45: LPF Gain Response

Mid-Range Band 1 (Band Pass)

Table 4: Mid-Range 1 Magnitudes

Frequency (Hz)	Output Voltage (Volts) (Peak-to-peak)
200	9.5
400	9.7
600	9.9
800	10.2
1000	10.3
1200	10.5
1400	10.5
1600	10.5
1800	10.5
2000	10.5
2200	10.5
2400	10.5
2600	10.3
2800	10.2
3000	10.2
3200	10.2
3400	10
3600	9.9
3800	10
4000	9.7

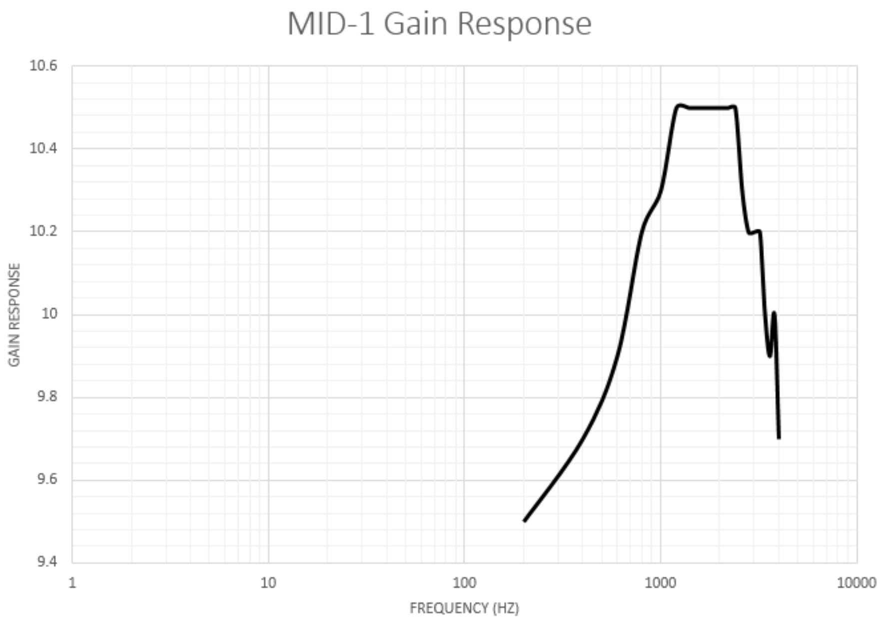

Figure 46: Mid-1 Gain Response

Mid-Range Band 2 (Band Pass)

Table 5: Mid-Range 2 Magnitudes

Frequency (Hz)	Output Voltage (Volts) (Peak-to-peak)
3200	10.1
3400	10.7
3600	11
3800	11.2
4000	11.4
4200	11.4
4400	11.4
4600	11.4
4800	11.4
5000	11.4
5200	11.3
5400	11.2
5600	10.8
5800	10.8
6000	10.3

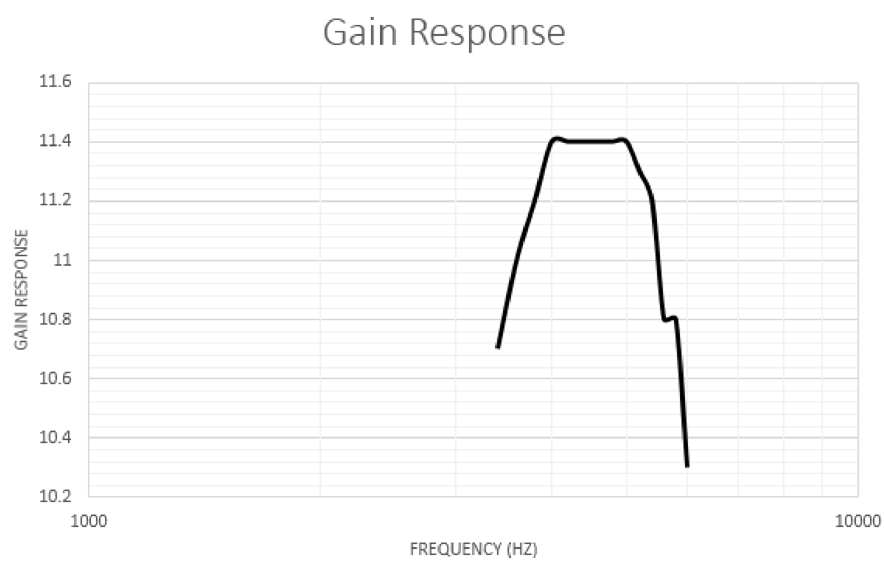

Figure 47: Mid-2 Gain Response

High Pass Filter

Table 6: High Pass Filter Magnitudes

Frequency (Hz)	Output Voltage (Volts) (Peak-to-peak)
4200	8.5
4400	8.5
4600	8.5
4800	9
5000	9
5200	9
5400	9.1
5600	9.1
5800	9.5
6000	9.8
6200	10
6400	10
6600	10
6800	10
7000	10

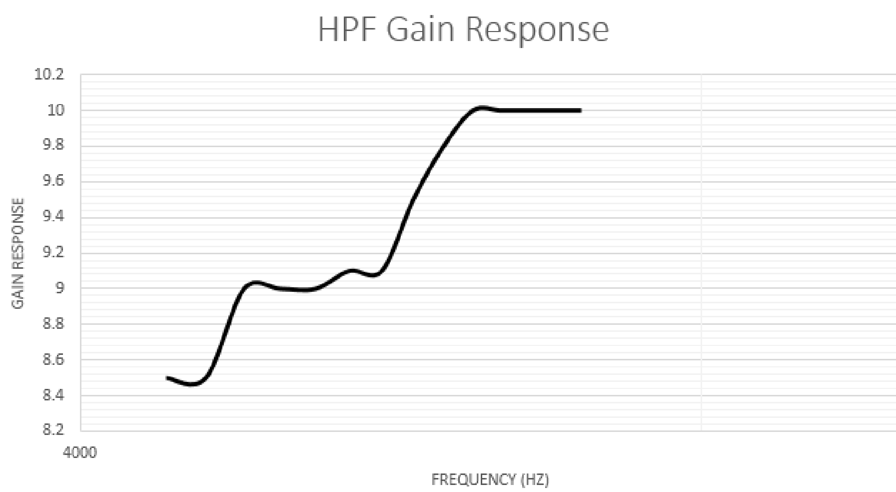

Figure 48: HPF Gain Response

The graphical results above show fairly similar performances of each of the crossover networks. The low pass and high pass filters show a unity gain over the pass-band up until the cutoff frequency.

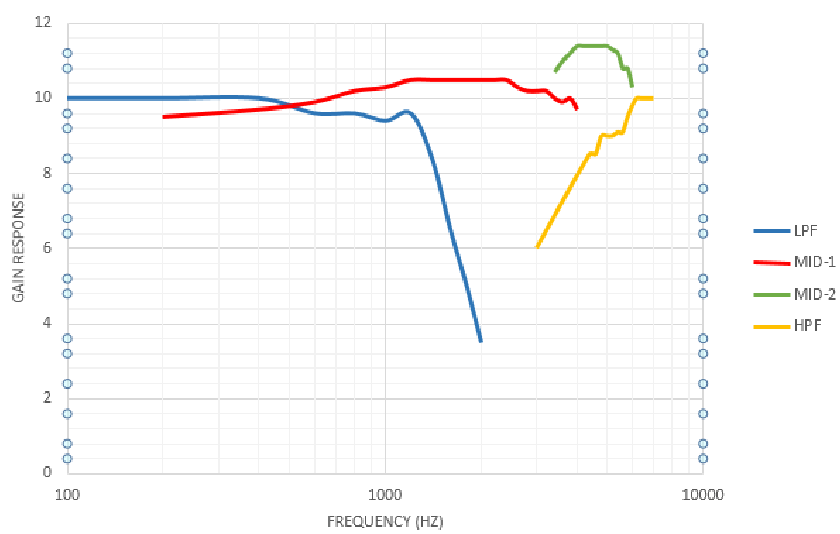

Figure 49: Comparison of the four Gain Responses

The experimental results differ a bit from the simulated results when the two gain responses of each network are compared. This is because some of the passive components used in the circuit are not the exact values of the simulation components and moreover, some of them were rounded off to the nearest standard value. Component tolerance could also be a cause of the difference as they are about 5% and above.

CHAPTER 5

CONCLUSION AND RECOMMENDATION

The filters were constructed from the given design specifications; Sallen-key topology with Butterworth response was used to achieve maximally flat pass-band with steep roll-off from pass- band to stop-band. The Bass (Low pass filter) and tweeter (High pass filter) designed are third order unity-gain filters while mid-ranges one and two designed are second order band-pass filters whose gain is dependent on the bandwidth of a given filter. Thus, midranges 1 and 2 have different gains. The values for the passive components of the respective filters were calculated. A LM324 and LM358 op-amps were used to implement the design in place of six LM741s to minimize congestion on the board. The LM358 was used in the design of the Band-pass 1 and 2 while the LM324 represented the active element in both the Low Pass and High pass filters.

The circuits designed were simulated with using Multisim and the computer simulation results practically confirmed the design specifications. Proteus 8 Professional was used to generate the PCB schematic from the complete circuit. Fabrication and implementation was done onto the board so that the device was ready for presentation.

It is recommended that both the L7815 and L7915 voltage regulators be connected to separate heat sinks. All the four TDA2030s should be connected to a common heat sink as they share a common pin. An addition of a fan is also advised in order to cool down the quickly heating TDAs and their heat sink.
Further enhancement can be made to the circuit like addition of potentiometers for control of each of the four crossover networks. This would ensure individual volume control of each of the outputs to the satisfaction of the user. Addition of a temperature controlled cooling system would also be suitable. A fan that would come on once the main volume surpasses the 50% mark would go some way in conserving power.

REFERENCES

1. Integrated Electronics Analog and Digital Circuits and Systems – Jacob Millman, Christos C Halkias.
2. First Courses on Power Electronics and drives – Ned Mohan
3. Analog Electronics 2nd Edition – Malestrom
4. Power Electronics Handbook - Rashid
5. R. M. -. C. Editor, Op-amps for Everyone, Dallas, Texas: Texas Instruments, 2002.
6. Texas Instruments, "Electrical Engineering Junior Lab," *Active Filter Design (Sallen KeyFilter),* vol. 2, p. 8, 2002.
7. H. R. M.E. Van Valkenburg, Analog Filter Design, 1982.
8. H. Zumbahlen, "Analog Devices," *Using the Analog Devices Active Filter Design Tool,* p.28, 2003.
9. K. Lacanette, "A Basic Introduction to Filters - Active, Passive and switched capacitors," vol. Application Note 779, p. 22, April 1991.
10. T. Kungelstadt, Active Filter Design Techniques, Texas: Texas Instruments, 2012.
11. H. Zumbahlen, Analog Filters.
12. M. E. Meserve, "Band-Pass Active Filter Designer," *Band-Pass Active Filter Designer,* p. 5.
13. JJC, "Active Filter Design (Sallen-Key Filter)," *Electrical Engineering Junior Lab,* p. 8, 9/7/02.
14. A.B. Williams, Electronic Filter Design Handbook, McGraw-Hill, 1981.
15. H. Zumbahlen, Passive and Active Filtering, Analog Devices.
16. S. P. Laboratory, Analog Electronics, Active Filters, CALIFORNIA: CALIFORNIA INSTITUTE OF TECHNOLOGY, Revision December 2012.
17. STMicroelectronics, "TDA 2030," *TDA 2030,* p. 13, 1998.
18. S.Franko, Design with Operational Amplifiers and Analog Integrated Circuits, McGraw–Hill, 1988.
19. J. W. A.I. Zverev, Handbook of Filter Synthesis, 1967.
20. D. a. J.Hilburn, Rapid Practical Designs of Active Filters, University of Michigan: Wiley, 1975.
21. B. Carter, Filter Design in Thirty Seconds, Texas instruments, December 2001.
22. D.Johnson and J.Hilburn, Rapid Practical Designs of Active Filters, John Wiley & Sons, 1975.
23. U.Tietze and Ch.Schenk, Halbleiterschaltungstechnik, Springer–Verlag, 1980.
24. H.Berlin, Design of Active Filters with Experiments, Howard W.Sams & Co., 1979.
25. M.Van Falkenburg, Analog Filter Design, Oxford University Press, 1982.

APPENDIX 1

FILTER COEFFICIENT TABLE

The table headers consist of the following quantities:

- **n** is the filter order
- **i** is the number of the partial filter.
- **a_i, b_i** are the filter coefficients.
- **ki** is the ratio of the corner frequency of a partial filter, fCi , to the cornerfrequency of the overall filter, fC. This ratio is used to determine theunity-gain bandwidth of the op amp, as well as to simplify the test ofa filter design by measuring fCi and comparing it to fC.
- **Q_i** is the quality factor of the partial filter.
- **fi / fC** this ratio is used for test purposes of the all-pass filters, where fi is thefrequency, at which the phase is 180° for a second-order filter,respectively 90° for a first-order all-pass.

Table 7: Butterworth Coefficients

n	i	a_i	b_i	$k_i = f_{Ci} / f_C$	Q_i
1	1	1.0000	0.0000	1.000	—
2	1	1.4142	1.0000	1.000	0.71
3	1	1.0000	0.0000	1.000	—
	2	1.0000	1.0000	1.272	1.00
4	1	1.8478	1.0000	0.719	0.54
	2	0.7654	1.0000	1.390	1.31
5	1	1.0000	0.0000	1.000	—
	2	1.6180	1.0000	0.859	0.62
	3	0.6180	1.0000	1.448	1.62
6	1	1.9319	1.0000	0.676	0.52
	2	1.4142	1.0000	1.000	0.71
	3	0.5176	1.0000	1.479	1.93
7	1	1.0000	0.0000	1.000	—
	2	1.8019	1.0000	0.745	0.55
	3	1.2470	1.0000	1.117	0.80
	4	0.4450	1.0000	1.499	2.25
8	1	1.9616	1.0000	0.661	0.51
	2	1.6629	1.0000	0.829	0.60
	3	1.1111	1.0000	1.206	0.90
	4	0.3902	1.0000	1.512	2.56
9	1	1.0000	0.0000	1.000	—
	2	1.8794	1.0000	0.703	0.53
	3	1.5321	1.0000	0.917	0.65
	4	1.0000	1.0000	1.272	1.00
	5	0.3473	1.0000	1.521	2.88
10	1	1.9754	1.0000	0.655	0.51
	2	1.7820	1.0000	0.756	0.56
	3	1.4142	1.0000	1.000	0.71
	4	0.9080	1.0000	1.322	1.10
	5	0.3129	1.0000	1.527	3.20

APPENDIX 2

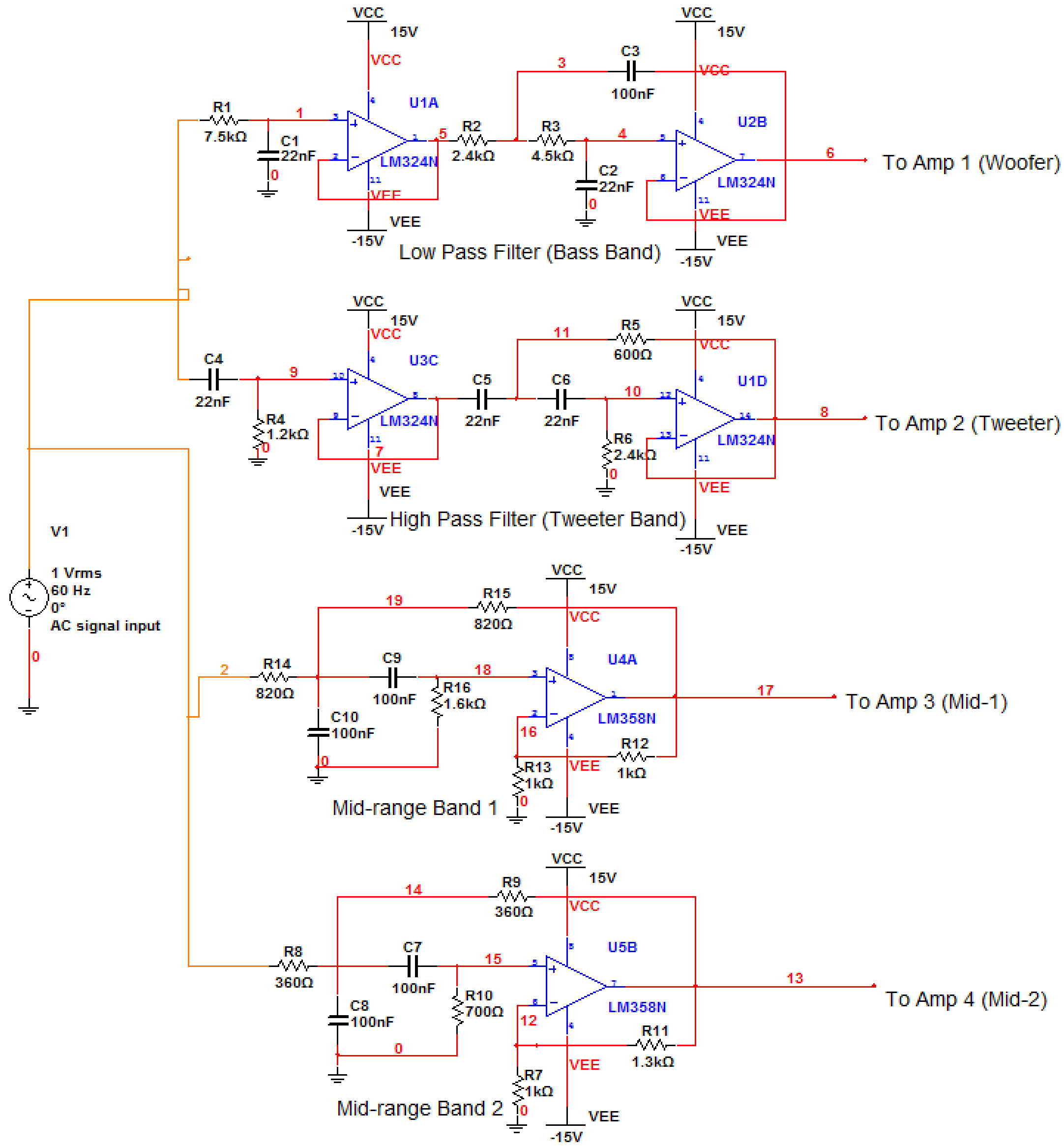

Figure 50: 4 Way Active Filter Pre-amplifier stage

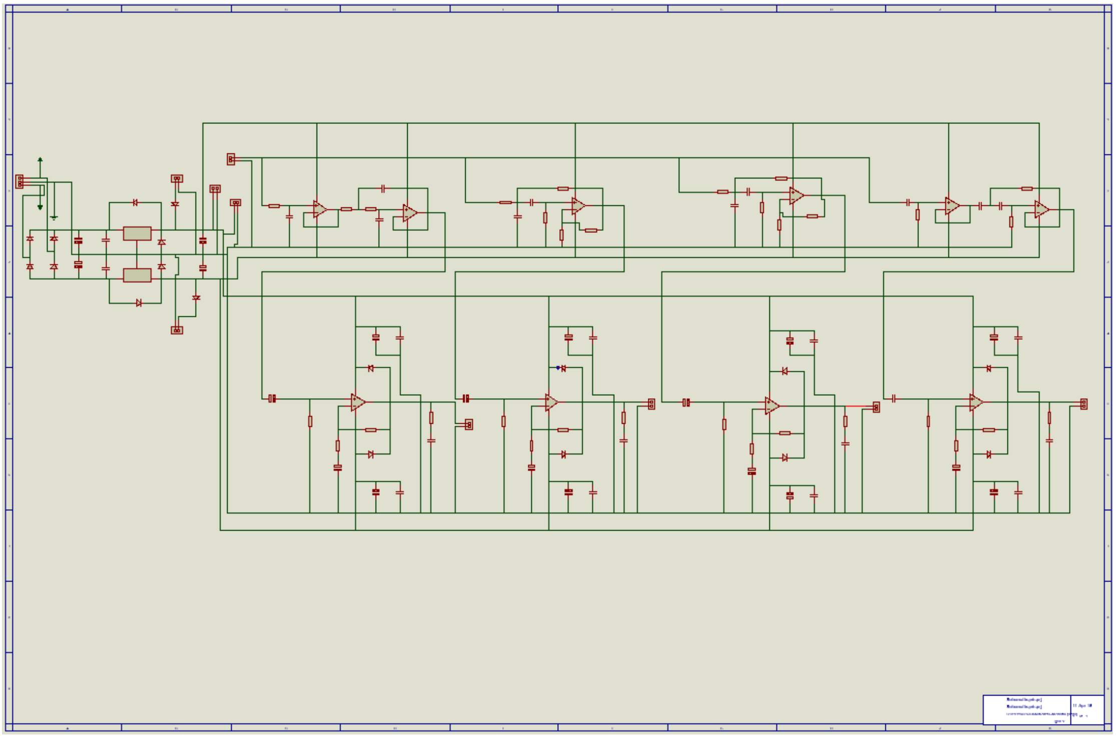

Figure 51: A schematic of the complete circuitry as shown in Proteus