**UNIVERSITY OF NAIROBI**

**SCHOOL OF ENGINEERING**

**DEPARTMENT OF ELECTRICAL AND INFORMATION ENGINEERING**

# DESIGN AND IMPLEMENTATION OF 100 W CLASS AB POWER AMPLIFIER

BY

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**F17/1439/2011**

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**Project report submitted in partial fulfillment of the requirement for the award for the degree of Bachelor of Science in Electrical and Electronics Engineering of the University of Nairobi 2016**

**Submitted on: 13th May 2016**

# Declaration of Originality

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**DEPARTMENT:** Electrical and Information Engineering
**COURSE NAME:** Bachelor of Science in Electrical & Electronics Engineering
**NAME OF STUDENT:** Okwado Festus Walufu
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## Dedication

To my family especially my dad, Mr. Matthew Ademba and my mum, Mrs. Magdalene Wangatia for their relentless support in my University education.

# Acknowledgement

I would like to express my heartfelt gratitude to Mr. Ogaba, who was my supervisor, for his constant guidance in the implementation of this project.

I would like to express great gratitude to all the lecturers in the Electrical Engineering department, University of Nairobi for equipping me with all the knowledge and skills that provided enough competence to complete the project.

I would also like to acknowledge my examiner, Mr. Ombura for taking his time to examine this project.

Thirdly, I would like to thank members of staff in the Electrical department including Mr. Rotich for providing support in acquisition of components and Mr. Wangai of mechanical workshop for his assistance in making of the PCB

Finally, I would like to thank my classmates & colleagues especially David Kibet, Gideon Mwendwa, Khaemba John and Tonny Silvance for their much valued advice and great contribution towards the success of the project.

Above all, I would like to thank the almighty God for enabling me achieve this success and showing me light at the end of the tunnel.

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# Chapter one: Introduction

## 1.1 History of power amplifiers

Vacuum tubes (valves) amplifiers were by far the dominant until the 1960s, when semiconductors (transistors) started taking over for performance and economic reasons, including heat and weight reduction and improved reliability. At first all silicon power transistors were NPN, and for a time most transistor amplifiers relied on input and output transformers for push-pull operation of the power output stage. These transformers were as always heavy, bulky, expensive, and non-linear, and added insult to injury as their LF and HF phase-shifts severely limited the amount of negative feedback that could be safely applied. Proper complementary power devices appeared in the late 1960s, and full complementary output stages soon proved to give less distortion than their quasi-complementary predecessors as mentioned in [1].

### 1.1.1 Vacuum tubes versus solid state designs

The following are disadvantages of vacuum tubes over solid state audio amplifiers as mentioned in [2]

1. Vacuum tube amplifiers cost from 3 to 10 time as much as solid-state designs.
2. Vacuum tubes are heavier and bulkier than solid state designs
3. Vacuum tubes are more fragile compared to solid state
4. Vacuum amplifier do not provide any protection against continuous short-circuit conditions at the output
5. Vacuum tubes amplifier waste more energy than solid state designs, due to the necessity of heating the tube filaments.

## 1.2 Objectives

### 1.2.1 Main project objective

The main objective of this project was to design a 100W power amplifier with feedback-pair complementary symmetry output power transistors.

### 1.2.2 Specific project objectives

- To design three class AB power amplifiers with feedback-pair complimentary symmetry output driven by an active 3-way cross-over network and a pre-amplifier.
- The total power of the three power amplifiers should some up to a total of 100W each driving a resistive load of 8 ohms.
- The filters making up the active network should be second order with Butterworth response comprising of Sallen-Key Topology.

## 1.3 Problem statement

Many power amplifiers are affected by noise and distortion. A simple class AB power amplifiers with feedback-pair complimentary was designed to overcome the mentioned problems.

## 1.4 Project scope

Class AB power amplifiers with feedback-pair complimentary was designed. The designed circuit was the simulated using Protues software. The results were then compared and found to be similar.

The circuit was then implemented on PCB after which results were taken and analysis done.

# Chapter two: literature review

## 2.1 Audio Power Amplifier

The audio power amplifier is a kind of electronic amplifiers that amplify low-power audio signals (the frequencies of the low-power signals are always between 20 Hz to 20 KHz) to a level that can be suitable for driving the loudspeakers. Nowadays all types of electronics that could make sounds are widely using the audio power amplifier, such as mobile phones, MP4 players, laptops, television, audio equipment, etc.

For better performance, an audio amplifier should have the following characteristics:

1. It should have very low harmonic distortion and intermodulation distortion.
2. It should have uniform frequency response over the whole audio range from 20 Hz to 20 kHz, +/-5 Db.
3. It should have maximum power output over the whole audio range without noticeable distortion.
4. It should be stable under practical load condition.
5. It should add very little or no noise to the input signal. Meaning that the amplifier should have higher signal-to-noise ratio.

## 2.2 Basic performance specifications of audio amplifier

The performance specifications listed by the manufacturer of an audio power amplifier range from a very sparse set to a fairly detailed list. The primary specifications include maximum power, frequency response, signal to noise ratio, and distortion.

### 2.2.1 Amplifier distortion

There are three types of distortion namely; phase distortion, frequency distortion and amplitude/ harmonic distortion [3].

Frequency distortion occurs when circuit elements and devices respond to the input signal differently at different frequencies.

Phase distortion is characterized by a change in the phase angle between the fundamental and one or more harmonic components of a complex waveform.

Harmonic or amplitude distortion occurs when the amplitude of the output signal is not directly related to the input signal by a constant factor over the complete 360° of an input signal. It is produced by the non-linearity of the characteristic of amplifier. This is the most significant distortion in power amplifiers. If the fundamental frequency has a amplitude $A_1$ and the *n*th frequency component has an amplitude An, a harmonic distortion can be defined as

$$\% \; n\text{th harmonic distortion} = \% \; D_n = \frac{A1}{An} \times 100\%$$

The fundamental component is typically larger than any harmonic component as mentioned in [4]

### 2.2.2 Frequency response

Frequency response is the term used to describe the range of tones that a stereo system can reproduce.

There are two requirements for frequency response:

One requirement is that the range of frequency response should be wide enough. The lower frequency should be as low as possible, and the upper frequency as high as possible. Typically, the specified frequency range for audio components is 20Hz to 20 KHz, which is the approximate range of human hearing.

The other requirement is that the frequency response should be flat. It means being linear. A well-designed amplifier is linear across the whole operating range, and its frequency response just varies a little between 20Hz to 20 KHz.

### 2.2.3 Signal to noise ratio

Signal-to-noise ratio is the ratio of signal power to noise power. There is no doubt that the higher signal-to-noise ratio, the better performance of the amplifier.

Noise is a term used to describe any unwanted random signal variations. The most common noise in amplifier is thermal noise and shot noise. Thermal noise is caused by any passive components due to movement of charged carriers in a circuit. Shot noise is caused by random arrival of charge carriers crossing a potential barrier, as in diodes and transistors.

Signal-to-noise ratio can be expressed as shown in equation below.

SNR= $^{Psignal}/_{Pnoise}$

### 2.2.4 Output power

Strictly speaking, the output power for amplifiers is usually regarded as maximum RMS-power output per channel, at a specified distortion level at a particular load, which is considered as the most meaningful measure of power.

In general, a power amplifier for loudspeakers will typically be measured at 4 and 8 ohms.

### 2.2.5 Slew Rate

Slew rate is a measure of how fast the output voltage of the amplifier can change under large-signal conditions. It is specified in volts per microsecond. Slew rate is an indicator of how well an amplifier can respond to high-level transient program content. A less capable amplifier might have a slew rate of 5 V/µs, whereas a really high-performance amplifier might have a slew rate on the order of 50 to 300 V/µs. For a given type of program material, a higher-power amplifier needs to have a higher slew rate to do as well as a lower-power amplifier, since its voltage swings will be larger. A 100-W amplifier driving a loudspeaker whose efficiency is 85 dB will need to have 3.16 times the amount of slew rate capability as a 10-W amplifier driving a 95-dB speaker to the same sound pressure level.

As a point of reference, the maximum voltage rate of change of a 20-kHz sine wave is 0.125 V/µs per volt peak. This means that a 100-W amplifier that produces a level of 40-V peak at 20 kHz must have a slew rate of at least 5 V/µs. In practice a much larger value is desirable for low-distortion performance on high-frequency program content.

## 2.3 Power amplifier stages.

An audio power amplifier is divided into three main stages

1. Input stage
2. Voltage amplification stage
3. Output stage

Before designing every stage of power amplifier, it is very important to consider if BJTs or FETs are the best devices for the job. Below is a summarized comparison of the two [1]

<u>Advantages of the FET input stage</u>

There is no base current with FETs, so this is eliminated as a source of DC offset errors. However, the FET gate leakage currents increase very rapidly with temperature, and under some circumstances may need to be allowed for.

<u>Disadvantages of FET input stage</u>

1. The undegenerated transconductance is low compared with BJTs. There is much less scope for linearizing the input stage by adding degeneration in the form of source resistors, and so an FET input stage will be very nonlinear compared with a BJT version degenerated to give the same low transconductance.
2. The Vgs offset spreads will be high. Having examined many different amplifier designs, it seems that in practice it is essential to use dual FETs,which are relatively very expensive and not always easy to obtain. Even then, the Vgs mismatch will probably be greater than Vbe mismatch in a pair of cheap discrete BJTs; for example the 2N5912 N-channel dual FET has a specified maximum Vgs mismatch of 15mV. In contrast the Vbe mismatches of BJTs, especially those taken from the same batch (which is the norm in production) will be much lower, at about 2–3mV, and usually negligible compared with DC offset caused by unbalanced base currents.

3. The noise performance will be inferior if the amplifier is being driven from a low-impedance source, say 5K or less. This is almost always the case.

The predictable Vbe/Ic relationship and much higher transconductance of the bipolar transistor make it, in my opinion, the best choice for all three stages of a generic power amplifier.

### 2.3.1 Input Stage

The input stage of an amplifier performs the critical duty of subtracting the feedback signal from the input, to generate the error signal that drives the output. There are two main possibilities for an input stage for a power amplifier.

a) Long tail pair
b) Single transistor

#### 2.3.1.1 Long Tail Pair (LTP)

Use of the long-tailed (or differential) pair in an amplifier means that the amplifier will operate with what is generally called 'voltage feedback' (VFB). The feedback is introduced as a voltage, since the input impedance of both inputs is high (and approximately equal), and input current is (relatively speaking) negligible.

The feedback resistor and capacitor are selected to allow the circuit to operate at full open loop gain for the applied AC, but unity gain for DC to allow the circuit to stabilize correctly with a collector voltage at (or near) 0V. The circuit diagram for LTP is shown below.

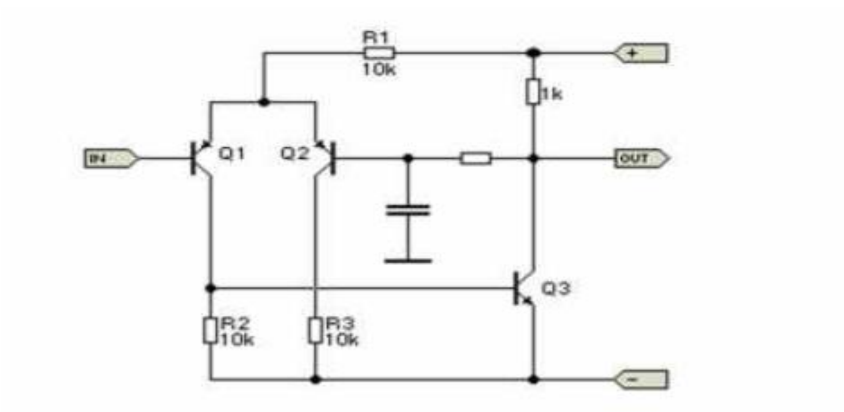


*Figure 1: Long Tail Pair Circuit*

The gain of an LTP is typically highest in its balanced state and decreases as the signal goes positive or negative away from the balance point. This symmetrical behavior is in contrast to the asymmetrical behavior of the common-emitter stage, where the gain increases with signal swing in one direction and decreases with signal swing in the other direction. To first order, the symmetrical distortion here is third harmonic distortion [5]

#### 2.3.1.2 Single Transistor

This is a single transistor, with the feedback applied to the emitter.

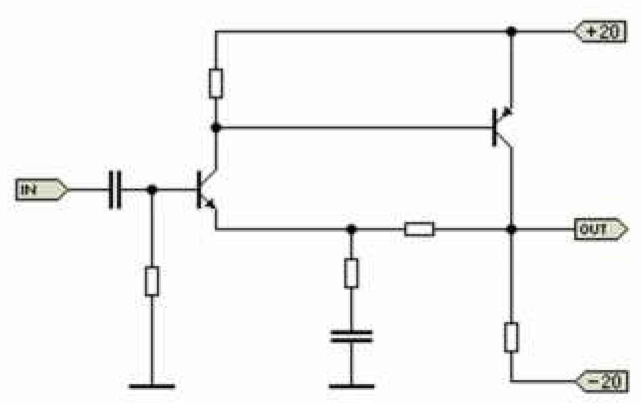


*Figure 2: Single Transistor Circuit*

An amplifier using this input stage requires little or no additional stabilization (the Miller Capacitor) which is mandatory with amplifiers having LTP input stages. An amplifier using this input stage is referred to as a 'current feedback' (CFB) circuit, since the feedback 'node' (the emitter of the input transistor) is a very low impedance.

The base circuit is the non-inverting input, and has a relatively high input impedance - but not generally as high as the differential pair. The +ve and -ve inputs are therefore asymmetrical. CFB amplifiers are used extensively in extremely fast linear ICs, and are capable of bandwidths in excess of 300MHz.

This input stage cannot be DC coupled (at least not without using a level shifting circuit), because of the voltage drop in the emitter circuit and between the emitter-base junction of the transistor. Since these cannot be balanced out as they are with an LTP input stage, the input must be capacitively coupled.

### 2.3.2 Voltage Amplification Stage

The Voltage-Amplifier Stage (or VAS) has often been regarded as the most critical part of a power-amplifier, since it not only provides all the voltage gain but also must give the full output voltage swing. A well-designed VAS stage will contribute relatively little to the overall distortion total of an amplifier [1].

This stage normally consist of class A driver, a currents source and a bias servo.

#### 2.3.2.1 Class A driver

The driver is normally a common- emitter transistor as shown in figure below

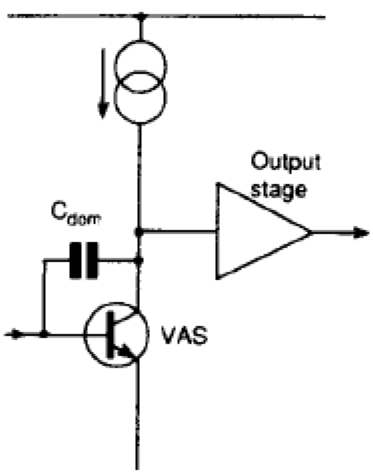


*Figure 3: Class A driver Circuit*

Capacitor Cdom is the so-called Miller compensation capacitor CM. It plays a critical role in stabilizing the global negative feedback loop around the amplifier. It does this by rolling off the high-frequency gain of the amplifier so that the gain around the feedback loop falls below unity before enough phase lag builds up to cause instability [1].

#### 2.3.2.2 Current sources

The distinguishing feature of a current source is that it is an element through which a current flows wherein that current is independent of the voltage across that element. Types of current sources are as follows:

##### 2.3.2.2.1 Simple Current Source Using voltage divider

A simple current source is shown in Figure below.

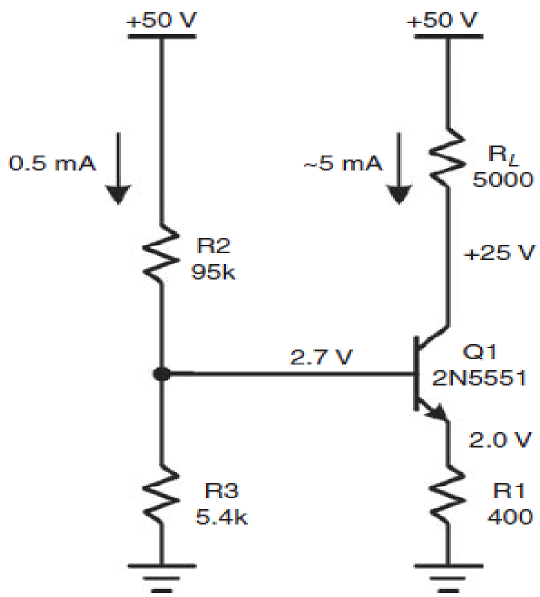


*Figure 4: Simple Current Source Using voltage divider Circuit*

The voltage divider composed of R2 and R3 places 2.7 V at the base of Q1. After a Vbe drop of 0.7 V, about 2 V is impressed across emitter resistor R1. If R1 is a 400-Ω resistor, 5 mA will flow in R1 and very nearly 5 mA will flow in the collector of Q1. The collector current of Q1 will be largely independent of the voltage at the collector of Q1, so the circuit behaves as a decent current source.

##### 2.3.2.2.2 Current source using LED

The figure for current source using LED is shown below

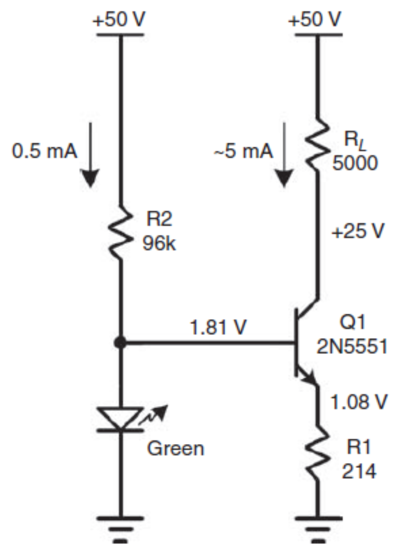


*Figure 5: Current source using LED Circuit*

The green LED is used as a voltage reference of about 1.8 V, putting about 1.1 volts across R1. Therefore the value determines the constant current Ic that will flow through Q1. In the figure above, R1 was set to 214 ohms hence giving Ic of about 5 mA. The green LED is biased by a current of 0.5 mA.

### 2.3.2.2.3 Bootstrap Current Source

Figure below shows the circuit of a bootstrap constant current source.

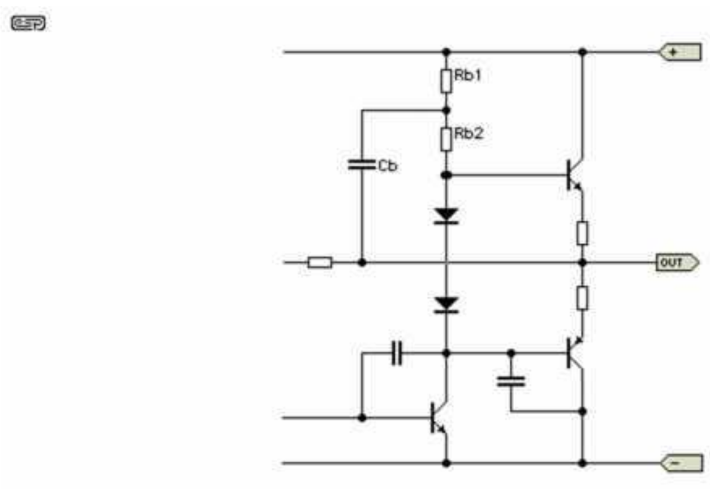


*Figure 6: Bootstrap Current Source circuit*

Under quiescent conditions, the output is at zero volts, and the positive supply is divided by Rb1 and Rb2. The base of the upper transistor will be at about +0.7V - just sufficient to bias the transistor. As the output swings positive or negative, the voltage swing is coupled via Cb, so the voltage across Rb2 remains constant. The current through Rb2 is therefore constant, since it maintains an essentially constant voltage across it.

#### 2.3.2.3 Bias Servo Circuit

The figure below is a bias servo circuit.

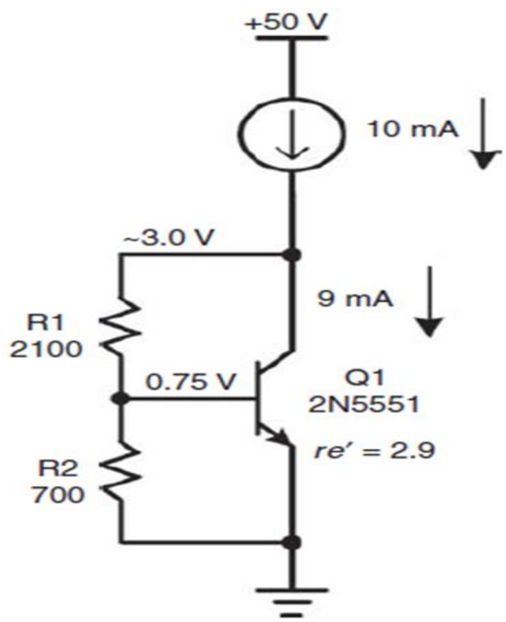


*Figure 7: Bias servo circuit*

In the circuit shown, the Vbe of Q1 is multiplied by a factor of approximately 4. The voltage divider formed by R1 and R2 places about one-fourth of the collector voltage at the base of Q1. Thus, in equilibrium, when the voltage at the collector is at four Vbe, one Vbe will be at the base, just enough to turn on Q1 by the amount necessary to carry the current supplied [5].

The design of many audio amplifier require that the bias servo is made adjustable by using a variable R2 in order to account for differing characteristics of output power transistors.

### 2.3.3 Output stage

The output stage essentially consists of a power amplifier and its purpose is to transfer maximum power to the output device. If a single transistor is used in the output stage, it can only be employed as class A amplifier for faithful amplification. Unfortunately, the power efficiency of a class A amplifier is very low ( 25% or less).

Class B push-pull amplifier has a better efficiency of 78.5% (maximum efficiency) which is way higher than class A amplifier. However class B push- pull amplifier suffers from crossover distortion. To eliminate crossover distortion, both transistor of class B push- pull amplifier are biased slightly above cut-off when there is no signal hence forming class AB push pull amplifier.

Class AB amplifier is therefore the most suitable among the three since it does not suffer from crossover distortion while at the same time maintaining a high efficiency of 78.5% (maximum efficiency).

There are three main types of output class AB stages:

1. Complementary Feedback Pair (CPF)
2. Darlington Complementary
3. Quasi-complementary

#### 2.3.3.1 Complementary Feedback Pair (CPF)

Complementary Feedback Pair (hereinafter CFP) sometimes called the Sziklai-Pair, seen in Figure below.

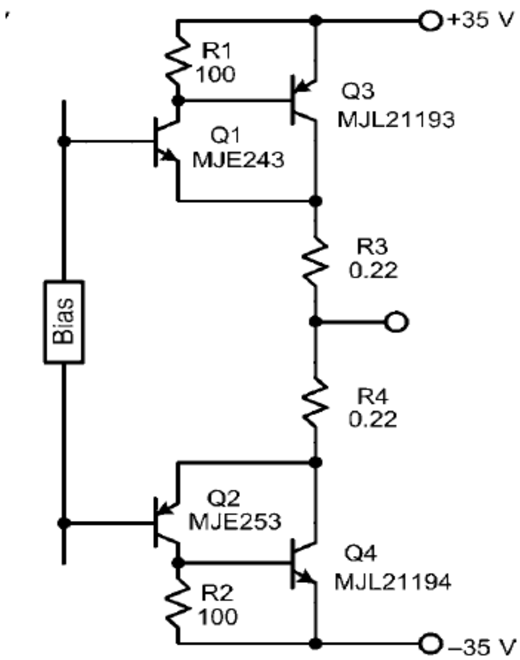


*Figure 8: Complementary Feedback Pair (CPF)  Circuit*

Here both output devices are operated in the common-emitter (CE) mode. The CFP output stage has a high degree of local feedback that linearizes each half of the output stage [5]. This results in very low output impedance for each half of the output stage and thus, presumably, reduced crossover distortion.

The compound feedback pair output stage has only one controlling Vbe, and is thus far easier to stabilize. Since the single Vbe is that of the driver the requirements for the Vbe multiplier are less stringent and thermal stability is generally very good to excellent.

#### 2.3.3.1.1 Miller Effect in the CFP Output Stage

The collector-base capacitance in the CE-operated output devices creates a Miller effect in the CFP. The small-signal effect is to reduce CFP bandwidth and to partially compensate its feedback loop. The large-signal effect is to cause high-frequency distortion due to the nonlinearity of Ccb of the output transistor. Moreover, there is also the large-signal action of the Miller effect that opposes turn-off. Consider the case where total Ccb is 500 pF and the output voltage slew rate is 50 V/µs. This current will be 25 mA, more than what is often run through the base-emitter resistor.

#### 2.3.3.2 Darlington complementary symmetry output

The circuit of Darlington complementary symmetry output is shown in figure below.

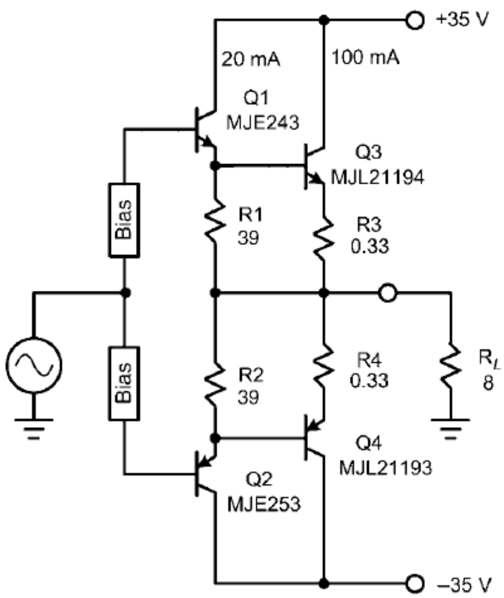


*Figure 9: Darlington complementary symmetry output Circuit*

This configuration is well known for its high current gain. However it suffers from thermal instability because the Darlington configuration has two emitter-base junctions for each output device. Since each has its own thermal characteristic (a fall of about

2mV per degree C), the combination is difficult to stabilize.

#### 2.3.3.3 Quasi-Complementary symmetry output

The output uses one pair of Darlington and another pair of compound in its implementation as shown in figure below.

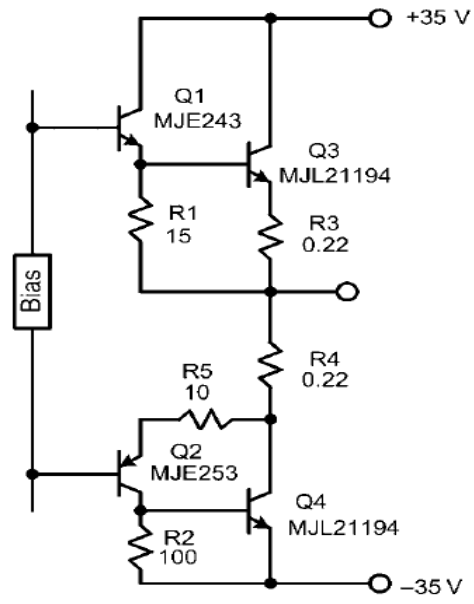


*Figure 10: Quasi-Complementary symmetry output Circuit*

This configuration suffers from distortion because of its fundamental asymmetry hence it is rarely used in modern days.

## 2.4 Protecting the Amplifier and Loudspeaker

One of the tougher practical design considerations for a power amplifier is protection. There are two major aspects here. The first, and most important, is that the amplifier should not fail in such a way that it will destroy the expensive loudspeakers to which it is connected. The second is that the amplifier should not self-destruct when driving a difficult load at high power levels or when it is subjected to a short circuit at the output.

### 2.4.1 Loudspeaker Protection

It is very important to protect loudspeakers from high DC voltages at the output of the amplifier. An output stage that fails will often do so by shorting the output of the amplifier to one of the rails. Speaker fuses are often used, but they can be unreliable and introduce low-frequency distortion. Active circuits that sense a DC level at the output of the amplifier can be used to open a loudspeaker relay or disable the power supply to the output stage. An alternative is to crowbar the output of the amplifier to ground with a TRIAC that is fired when DC is sensed.

### 2.4.2 Short Circuit Protection

This is the most fundamental form of protection for the amplifier itself. With a rugged output stage, a loudspeaker fuse or rail fuses or circuit breakers may be sufficient. How long it takes them to act is the key here. Current limiting and active amplifier shut down circuits can also play an important role in protecting the amplifier against short circuits.

## 2.5 FILTERS

A filter is defined as any circuit that produces a prescribed frequency response characteristic, of which the most common objective is to pass certain frequencies while rejecting others [6]. Filters can be classified as:

- 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

### 2.5.1 Passive Filters versus Active Filters

Passive filters consists of combinations of resistance, capacitance and inductance. Passive RLC structures are capable of achieving relatively good filter characteristics in application ranging from the audio frequency range to the upper limit of the lumped parameter range. However at the lower end of audio frequency range, inductance values increase as the required frequency decreases causing increase in internal losses since inductors are imperfect devices. This losses cause the filter response to have large deviations from their desired forms.

Active filters consist of combination of resistance, capacitance and one or more active device (such as opamp) employing feedback. While active filters are capable of operating at low frequency, they have a few disadvantages of their own. This include requiring power to operate and also employing feedback which can cause instability in the circuit.

### 2.5.2 Band Classifications of active filters

#### 2.5.2.1 High pass filter

A high-pass filter is an electronic filter that passes signals with a frequency higher than a certain cutoff frequency and attenuates signals with frequencies lower than the cutoff frequency. A second order high pass filter is shown below with unity-gain sallen key configuration.

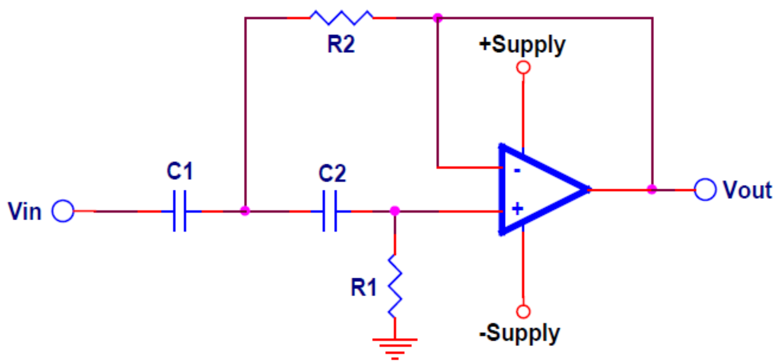


*Figure 11: A second order high pass filter Circuit*

#### 2.5.2.2 low-pass filter

This is a filtere that passes signals with a frequency lower than a certain cutoff frequency and attenuates signals with frequencies higher than the cutoff frequency. Below is a second order low pass filter with sallen key configuration.

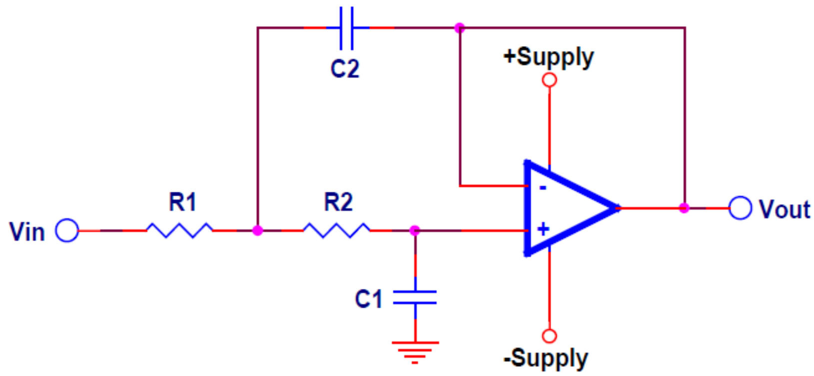


*Figure 12: A second order low pass filter Circuit*

#### 2.5.2.3 Band pass filter

This is a filter that passes frequencies within a certain range and attenuatres frequencies outside that range. It can be implemented by cascading a high pass and a low pass filter as shown below.

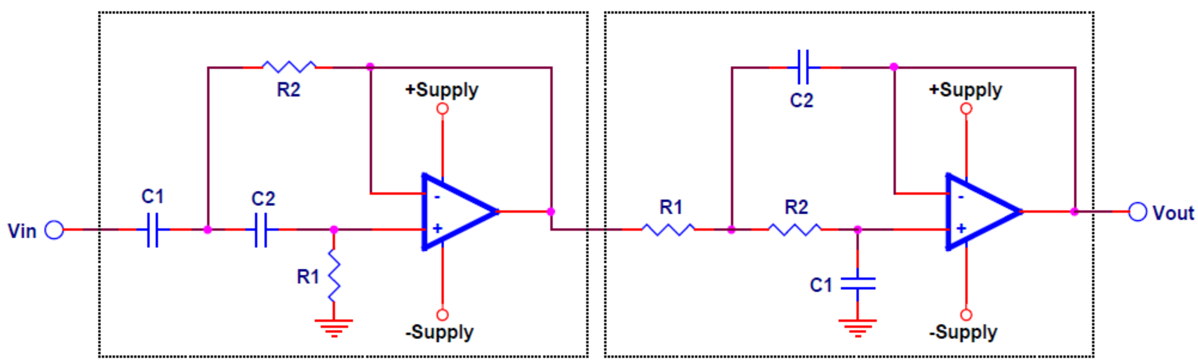


*Figure 13: A second order band pass filter Circuit*

### 2.5.3 Filter optimization

#### 2.5.3.1 Butterworth response

This response is also referred to as maximally flat amplitude response. It provides maximum band pass flatness hence suitable for designing active crossover network for audio signal.

Below is a gain response of different orders of Butterworth low-pass filters versus the normalized frequency axis, Ω (Ω = f / $f_C$); the higher the filter order, the longer the passband flatness.

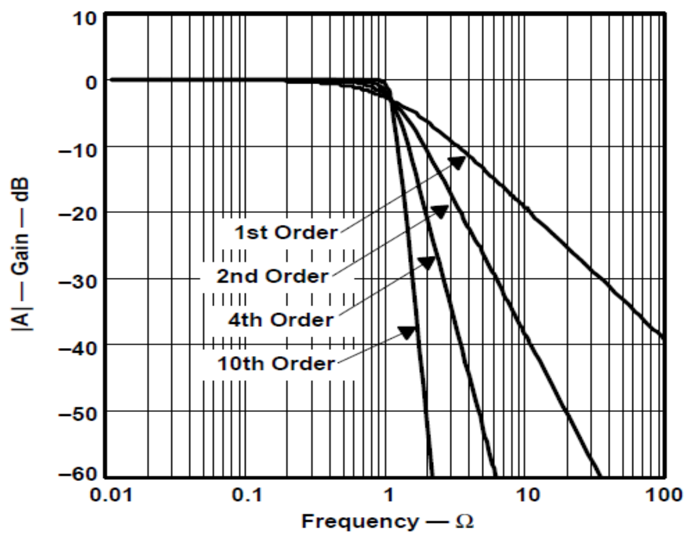


*Figure 14: Butterworth response graph*

#### 2.5.3.2 Chebyshev Response

Its response is referred to as an equiripple response because the pass band is characterized by a series of ripples that have equal maximum levels and equal minimum levels. The number of ripples is a function of the reactive elements in the design. This filters have a sharper slope than Butterworth and are thus are capable of achieving more attenuation in the stop band for a given number of reactive elements. However their time delay and phase characteristics are less ideal than those of Butterworth filter, and they tend to exhibit a ringing effect with transient signals [6].

For a given filter order, the higher the passband ripples, the higher the filter's roll-off as shown in figure below. With increasing filter order, the influence of the ripple magnitude on the filter roll-off diminishes. This kind of filters are used in filter banks, where the frequency content of a signal is of more importance than a constant amplification. Below is gain versus frequency response of different orders of Chebyshev low pass filters.

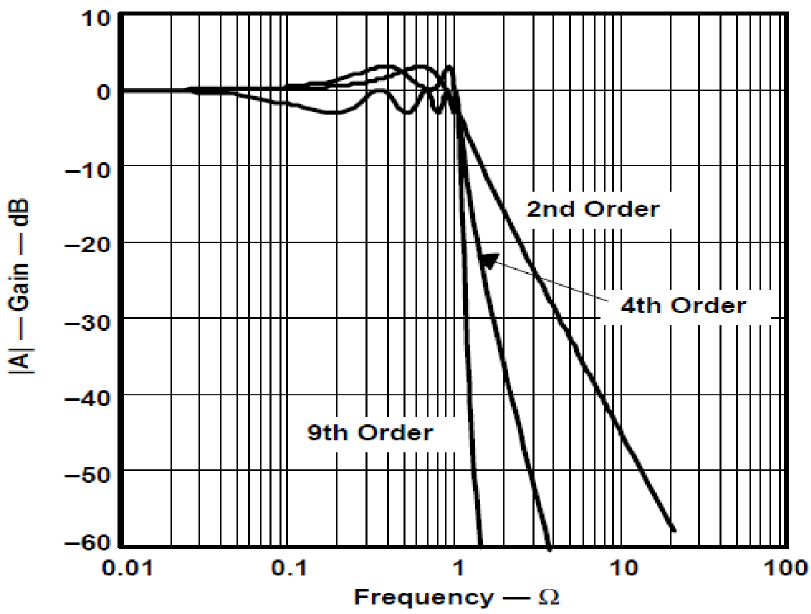


*Figure 15: Chebshev response curve*

#### 2.5.3.3 Bessel filter response

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

### 2.5.4 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.

# Chapter three: Methodology and circuit design

## 3.1 Circuit design

The circuit was divided into four stages:

1. Preconditioning stage
2. Input stage
3. Voltage amplification stage
4. Output stage

The circuit was designed stage by stage, based on the specification requirements. The requirements were as follows.

1. The amplifier should produce 100-watts.
2. The power amplifier should employ complementary feedback-pair output power transistors.

3. The amplifier should be able to operate within audio frequency range which is between 20 Hz to 20 KHz.
4. The amplifier should have low distortion with low noise.

The circuit was designed from the output stage all the way to the precondition circuit.

### 3.1.1 Output Stage

The output stage provides a gain of slightly less than unity. Its main job is to provide buffering in the form of current gain between the output of the VAS and the loudspeaker load. The output stage of audio amplifier has to be designed in such a way that it suffers from very little or no distortion. The class of amplifier chosen should also have relatively high efficiency. Class AB was chosen because it meets the above conditions. It does not suffer from cross over distortion and it has a relatively high efficiency of 78% (maximum efficiency).

There are various types of output class AB stages. However, compound feedback pair output stage was chosen since it is easier to stabilize compared to quasi-complimentary and Darlington complimentary output stages. This is because CFP has only one controlling Vbe, and is thus far easier to stabilize.

BJTs were used because of their predictable Vbe/Ic relationship and much higher trans-conductance compared to FET.

Design of output stage

The output stage was split into 3 sections

1. 60 watts power amplifier to drive the woofer.
2. 35 watts power amplifier to driver the mid-range
3. 5 watts power amplifier to drive the tweeter

1. Design of 65 watts woofer

2. Design of 35 watts power amplifier to drive mid-range

The design circuit is shown in the figure below

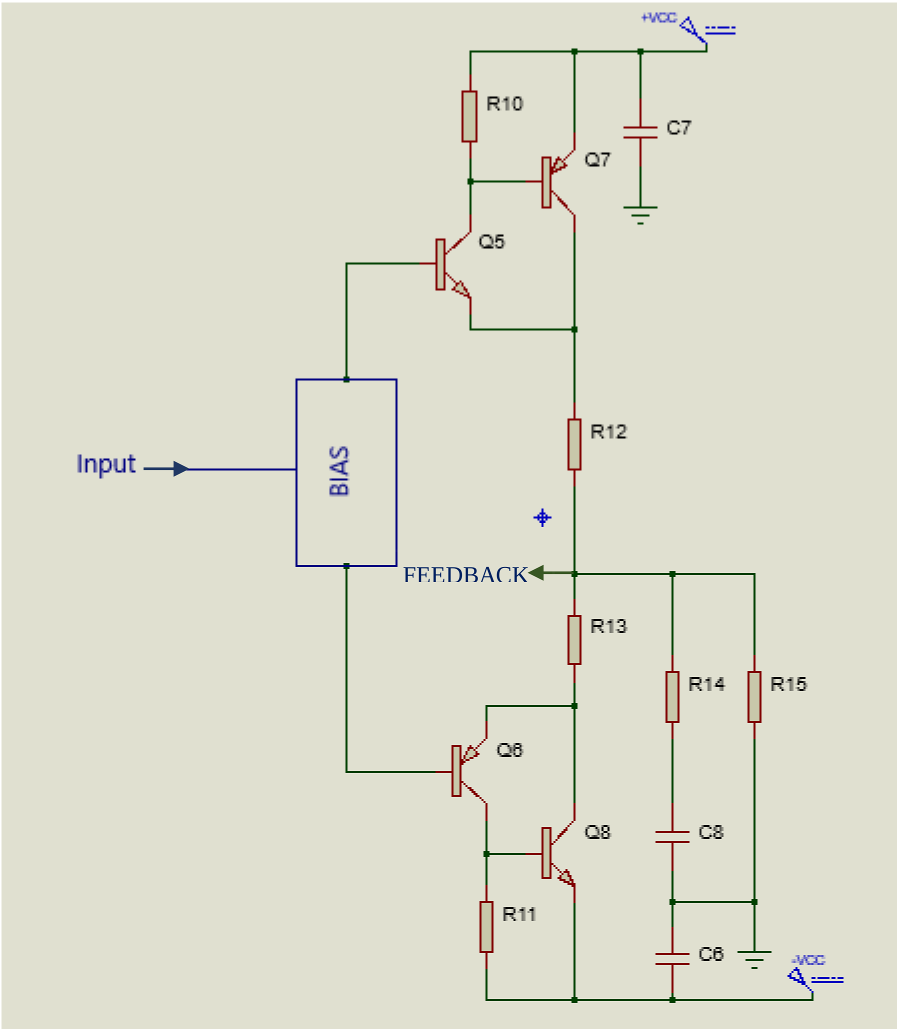


*Figure 16: Class AB power amplifier with CFP Circuit*

Amplifier specifications

1. 30 watts output power
2. 8 ohms load (R15)

Vcc calculation:

$$\text{Power} = \frac{(\text{Vrms})^2}{2\text{Rl}}$$

$= \frac{(Vout)^2}{2Rl}$ where Vout is the peak output voltage on the speaker.

Vout = $\sqrt{2Rl \times power}$

Vout = $\sqrt{2 \times 30 \times 8}\ V$

$= 21.9\ V$

The peak value for output current is given by

Iout = $\frac{\text{Vout}}{\text{Rl}}$

$= \frac{21.9}{8} A$

= 2.74 A

To calculate the required positive rail supply voltage (+Vcc) we have to assess the additional voltage drop at Q5, Q7, R10 and R12. We can assume that the total drop is around 3V. Meaning that:

+Vcc = 21.9 V + 3V

≈ 25 V

Since the circuit is symmetrical it means that the negative rail supply voltage (-Vcc) is:

-Vcc = - 25V

Taking into account the calculated values above, we can choose an appropriate transistor. Knowing that the peak collector current is 2.74 A, to be on a safe side we will use transistors that have peak collector current at least two times higher, and there will suit those which have that parameter in the range of 7...10A. Maximum voltage between collector and emitter of output transistor we have at the moment when the complementary transistor is inactive, is then given by:

Vce = Vcc + Vout

= 25 V + 21.9 V

= 46.9 V

To be on a safe side we should add at least another 30%, and therefore Vce should be chosen to be at least 60V.

In my case a chose Q7 as BD712 and Q8 as BD711 power transistors because their ratings meet the minimum requirements, i.e. power rating is 75Watts, Vce is 90 V and Ic is 10A.

To determine the parameters of driver stage transistors Q5 and Q6 we must proceed from assumptions that the current gain of output transistors is 40 which is quite realistic and even a modest value for today's output transistors. Assuming this value, we can calculate the maximum collector current of the drivers as:

$$Icpdr = \frac{Iout}{Bp}$$ where Bp is the current gain of power transistor

$$= \frac{2.74}{40} \times 1000\ mA$$

$$= 68.5 \text{ mA}$$

To be safe, we should use transistors with at least two times higher collector current, let's say 150mA, and even better to choose even higher value which will not be a problem for modern driver transistors. For maximum collector-emitter voltage of the drivers we should choose the same as for the output power transistors, i.e. about 60V.

The best driver transistor that meets all the above requirements is BD139 and BD140 power transistors.

Assuming that the current gain for BD139 and BD140 is 50, we can get its base current as:

$$Ibpdr = \frac{Icpdr}{Bpdr}$$ where Bpdr is current gain of power drive transistors

$$Ibpdr = \frac{68.5}{50} \text{ mA}$$

$$= 1.37 \text{ mA}$$

Resistor R10 and R11 are added to prevent output transistor collector to base leakage current from allowing the device to turn itself on, and also speeds up the turn-off time. The value must be selected with reasonable care, if it is too low, the output transistor will not turn on under quiescent (no signal) conditions, the driver transistors will be subject to excessive dissipation, and crossover distortion will result. If too high, turn-off performance of output devices will be impaired and thermal stability will not be as good.

Values of between 100 Ohms up to a maximum of 1k should be fine for most amplifiers, with lower values used as power increases. In my case I chose R10 and R11 to be 220 ohms. The value is well within the range.

The output stage emitter resistors R12 and R13 provide thermal bias stability and also play a role in controlling crossover distortion. Larger values of emitter resistor RE promote better current sharing. However, it is often possible to achieve lower crossover distortion with smaller emitter resistors. There is thus a trade-off between bias stability and crossover distortion. The use of small-value emitter resistors can lead to current hogging by one of the output transistors (0.15 ohms or less is considered to be small). The effect of thermal runway is more pronounced with output stages where very low-value emitter resistors are used.

I choose 0.33 ohm resistor for my design for both R12 and R13.

## 3.2 Voltage amplification stage

VAS is a high-gain common-emitter (CE) stage that provides most of the voltage gain of the amplifier. It is loaded with a current source rather than a resistor so as to provide the highest possible gain (voltage gain of 100 to 10,000). This means that the difference signal needed to drive the input stage does not need to be very large to drive the output to its required level. If the difference signal is close to zero, and 1/20 of the output is compared to the input, it follows that the output would be almost exactly 20 times the input. Below is a circuit diagram of VAS. It has a class A driver, spread bias and a constant current source (bootstrap).

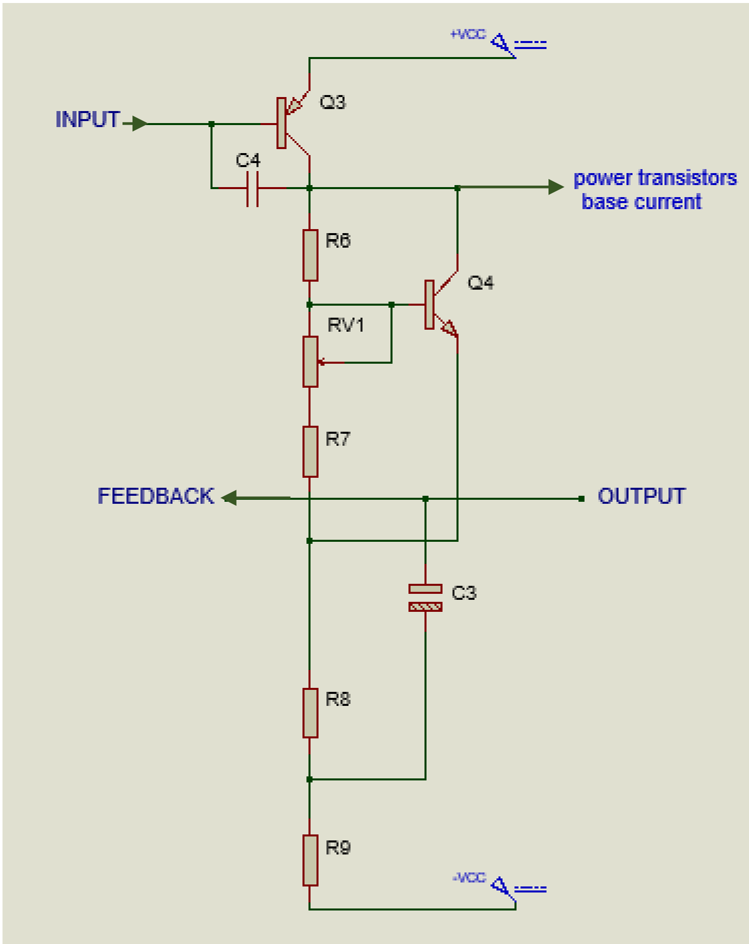


*Figure 17: VAS stage circuit design*

### 3.2.1 Design of bootstrap

The bootstrap circuit is formed by C3 capacitor and resistors R8 and R9.

From previous section the current to power transistors driver was found to be 1.37mA.

We have to design the current to provide at least 4 times the power transistors drive current.

Ibtsrp = 4× 1.37 mA ; where Ibstrp is the bootstrap current

= 5.48 mA

Icdr = 5×1.37

= 6.85 mA

Voltage drop across R9 + R8 is given by:

$V_{cc} - Q_{6\ Vbe}$ = 25V – 0.65V

= 24.35 V

Therefore the values of R8 and R9 can be calculated as:

$$R9 + R8 = \frac{24.35}{5.48}\ K$$

$$= 4.44\ K$$

For simplicity, we design such that R9 = R8. Since the two resistor are in series it implies that:

$$R9 = R8 = \frac{4.43}{2}\ K = 2.217\ K$$

$$\approx 2.2\ K \text{ (the nearest resistor standard value)}$$

Since R8 and R9 were designed to be equal the voltage rating of C3 of should be a minimum of ½ the positive supply voltage, but preferably greater. The circuit was designed with C3 rated at 100 nF (35V). The C3 capacitor is used from the output to maintain a relatively constant voltage across a resistor.

### 3.2.2 Design of Bias Servo

The bias servo is made up transistor Q4 and resistors; R6, R7 and Rv1. If we want to be able to completely stop any current flow through the output transistors by turning RV1 preset, we have to bring voltage higher than some 0,6-0,7 V to the base of Q4. To be safe, let it be 0.75V. For that voltage to appear on the base of Q4, we have to turn Rv1 slider completely on one end that is connected to the collector of the Q3.

Voltage drop across RV1 + R7 is equal to the Vbe of Q4 which is approximately equal to 0.65 V. Through a serial link Rv1, R6 and R7 should not pass more than about 1/10 of that total current of Q4 which is approximately 0.55 mA and the total resistance Rv1 + R7 can be calculated as:

$$RV1 + R7 = \frac{0.65}{0.55}\ K$$

$$= 1.2\ K$$

For practical purpose we choose RV1 to be 1K and R7 to be 390 ohms (slightly higher than 200 ohms for protection purpose, plus the slight increase in resistance will have un noticeable effect on the signal).

If we ever move Rv1 slider all the way to the end connected to base of Q4, we will actually make a short circuit between the base and emitter of Q4, which will stop its conduction and its collector emitter junction will behave as a very high resistance. Because of that, voltage between the bases of power transistor drivers (Q5 and Q6) will be much higher than necessary 1.3V, which will in turn produce that the output transistors immediately begin to draw very high current, and probably blow of immediately! To prevent that, resistor R7 is inserted in series with RV1.

The value of R6 was chosen to be 1k since the voltage drop across it is 0.65 V and current through it is approximately 0.6 mA.

The type of transistor to be used for Q4 should be a low power transistor since the transistor is not handling much power (approximately 5.5mA × 1.4V = 7.7 mA). In this case BC546BP was chosen.

The driver class A resistor chosen should be able to pass more than 30mA collector current and a Vce of greater than 50V. The power transistor used in this case was BD140.

## 3.3 Input stage

The input stage of an amplifier performs the critical duty of subtracting the feedback signal from the input, to generate the error signal that drives the output. It is almost invariably a differential transconductance stage; a voltage-difference input results in a current output that is essentially insensitive to the voltage at the output port. The circuit for input stage of audio amplifier is shown below.

Transistors Q1 and Q2 form the input differential pair. This arrangement is often called a long-tailed pair (LTP) because it is supplied with a so-called tail current from a very high-impedance circuit like the current source shown. We will often take the liberty of referring to the amplifier’s input stage as the IPS. The input differential amplifier usually has a fairly low voltage gain, typically ranging from 1 to 15.

The IPS compares the applied input signal to a fraction of the output of the amplifier and provides the amount of signal necessary for the remainder of the amplifier to create the required output. This operation forms the essence of the negative feedback loop. The fraction of the output to which the input is compared is determined by the voltage divider consisting of R5 and R2. If the fraction is 1/20 and the forward gain of the amplifier is large, then very little difference need exist between the input and the fed back signal applied to the IPS in order to produce the required output voltage. The gain of the amplifier will then be very nearly 20. This is referred to as the closed-loop gain of the amplifier.

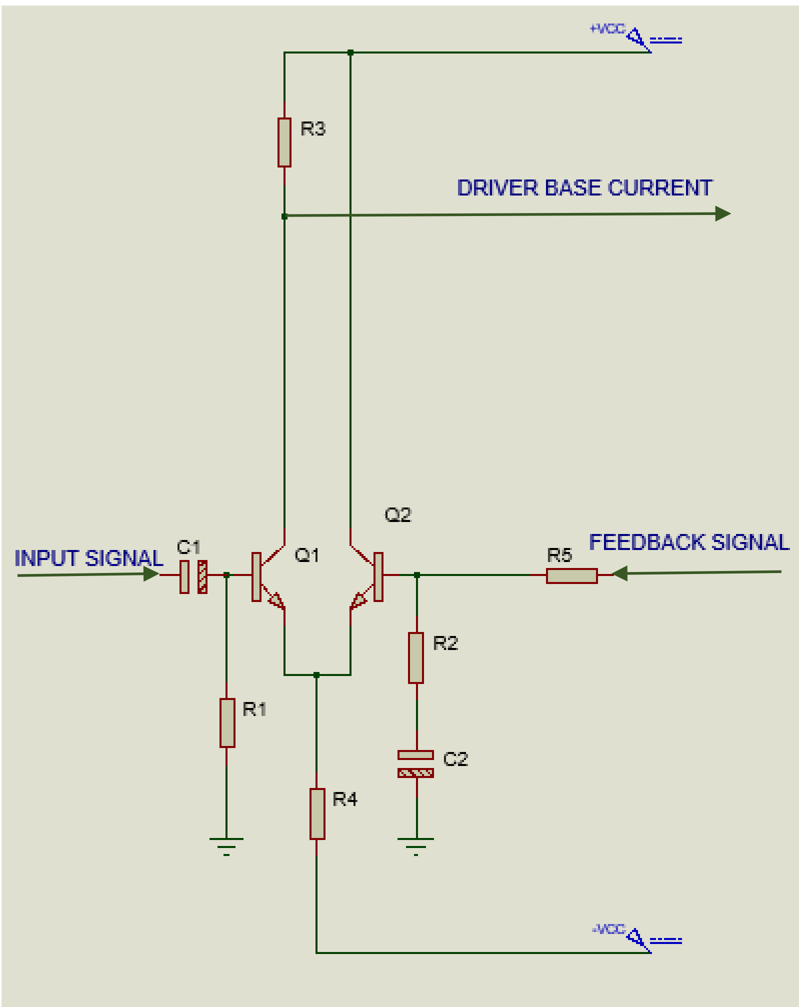


*Figure 18: Input stage design circuit*

Calculation of component values:

From previous we know that the driver collector current is 6.85mA (Icdr) and taking the current gain (Bdr) of driver transistor as 50, then

$$\text{Ibdr} = \frac{Icdr}{Bdr}$$

$$= \frac{6.85mA}{50}$$

$$= 0.137 \text{ mA}$$

We make collector current of Q1 to be 5 times Ibdr.

$$\text{Icq1} = 5\times 0.137 \text{ mA}$$

$$= 0.685 \text{ mA}$$

We choose R3 such that the current through it is five times greater than 0.685 mA. We can choose the value to be 1 mA which is now the collector current for Q1.

The voltage across R3 ($V_{R3}$) is equal to Vbe of Q3 which can be approximated as 0.65 V.

Therefore the value of R3 can be found as follows:

$$\text{R3} = \frac{VR3}{IR3}$$

$$= \frac{0.65}{1} \times 1000 \text{ Ohms}$$

$$= 650 \text{ Ohms}$$

R3 was chosen as 680 ohms because it was the closest standard value to 650 Ohms

The values of R4 was calculated as follows:

$$\text{R4} = \frac{Vcc - Vbeq1}{2\times Icq1}$$

$$R4 = \frac{25-0.65}{2\times1}\ K$$

R4 = 12.175 K

≈ 12 K (the approximate resistance standard value

The value of R5 was chosen to be 22 K so that to balance the inverting terminal input impedance. The value of R2 was chosen as 1.3K so that to give a voltage gain (Av) of 18.

$$Av = \frac{R5+R2}{R2}$$

$$= \frac{22+1.3}{1.3}$$

= 17.923 ≈ 18

R1 and C1 form a high pass filter to filter out frequencies less than 20Hz, so does R2 and C2. This means that the product of R1 and C1 should be approximately equal to the product or R2 and C2.

C1 was chosen to be 4.7 uF (35V) while C2 was chosen to be 100 uF (35V). The voltage rating were chosen to be slightly higher than Vcc so that the capacitors can charge fully without blowing up.

Q1 and Q2 were chosen as BC546 because they are low frequency and low power transistor since Q1 and Q2 are handling low power.

## 3.4 Design of three way active crossover network

The following table was used to select crossover frequencies for the active filters.

| Crossover frequency | % Power to Bass | % Power to Mid + High |
|---|---|---|
| 250 Hz | 40 | 60 |
| 350 Hz | 50 | 50 |
| 500 Hz | 60 | 40 |
| 1,200 Hz | 65 | 35 |
| 3,000 Hz | 85 | 15 |
| 5,000 Hz | 90 | 10 |

The above table came from a loudspeaker manual "LOUDSPEAKER ENCLOSURE DESIGN AND CONSTRUCTION" published by FANE.

The crossover frequency for woofer was chosen to be 1.2 KHz because from our power amplifier design, the woofer amplifier was designed to supply 65% of total power. The crossover frequency from mid-band to tweeter (f2) was chosen to be 5 KHz.

The filters were designed using Butterwoth response with Sallen Key configuration. For simplicity of the circuit, all the filters were second order

### 3.4.1 Design of Low Pass filter

Below is a second order low pass filter circuit diagram with cutoff frequency (f1) 1.2 KHz.

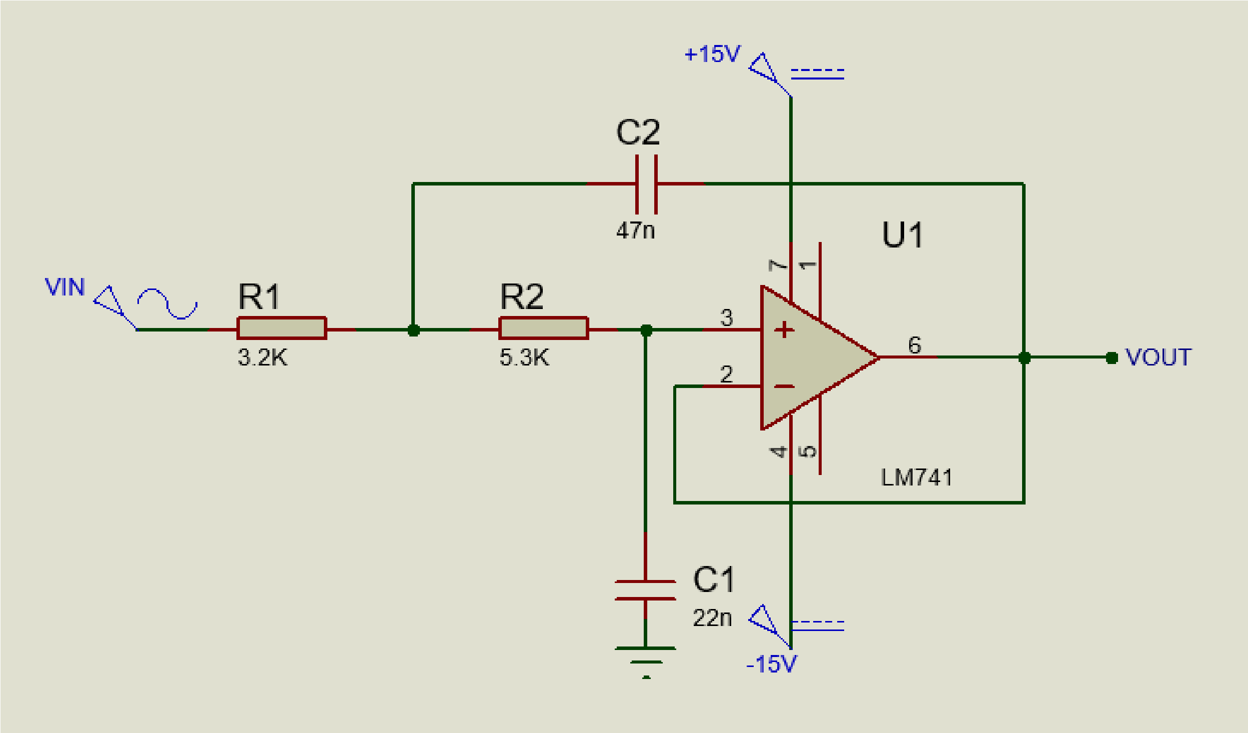


*Figure 19: second order low pass filter circuit diagram with cutoff frequency (f1) 1.2 KHz.*

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

$$R_{1,2} = \frac{a1c2 \pm \sqrt{(a1)^2\,(C2)^2 - 4b1C1C2}}{4\pi f1C1C2}$$

C1 was chosen to be 22 nF

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

$$C2 \geq = \frac{C1 \times 4b1}{(a1)^2}$$

For second order Butterworth response, $a_1$=1.4142 and $b_1$= 1

$$C2 \geq \frac{22 \times 4 \times 1}{1.4142^2}\ \text{nF}$$

$C2 \geq 44$ nF

C2 = 47 nF (nearest capacitor standard value)

$$R_{1,2} = \frac{1.4142\times47\times10^{-9} \pm \sqrt{(1.4142)^2\,(47\times10^{-9})^2 - 4\times47\times22\times10^{-18}}}{4\pi\times1.2\times47\times22\times10^{-15}}$$

$R_1 \approx 3.2$ K

$R_2 \approx 5.3$ K

### 3.4.2 Design of high pass filter

Below is a second order high pass filter circuit diagram with cutoff frequency (f2) 5 KHz.

It has Butterworth response with Sallen Key topology.

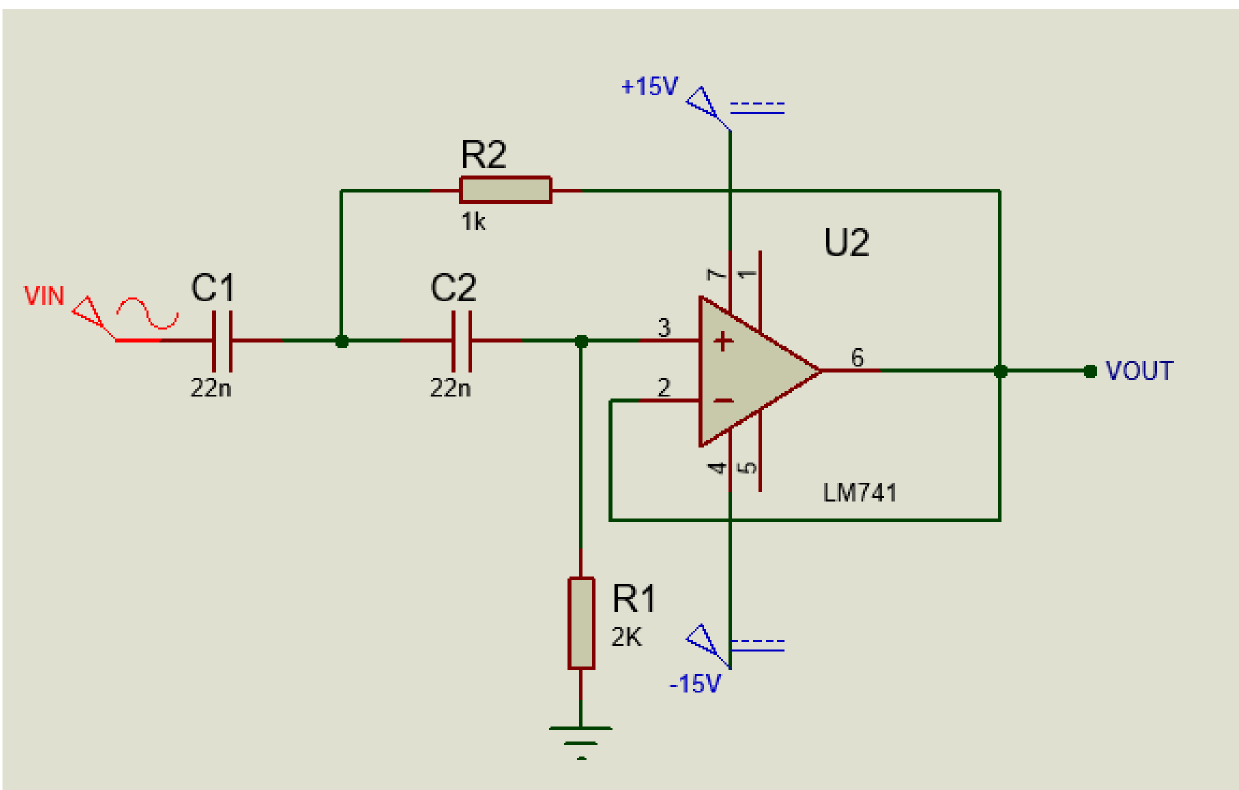


*Figure 20: second order high pass filter circuit diagram with cutoff frequency (f2) 5 KHz.*

We chose C = C1 = C2 = 22nF

For second order Butterworth, a1=1.4142 and b1=1

$$R1 = \frac{1}{\pi f2 \times C \times a1 \times 1000} \text{ K}$$ ; where $f_2$ is cutoff frequency for high pass filter.

$$R1 = \frac{1}{\pi \times 5000 \times 22 \times 1.4142 \times 10^{-9} \times 1000} \text{ K}$$

R1 ≈ 2 K

$$R2 = \frac{a1}{4\pi f2 \times C \times b1 \times 1000} \text{ K}$$

$$R2 = \frac{1.4142}{4\pi \times 5000 \times 22 \times 10^{-9} \times 1 \times 1000} \text{ K}$$

R2 ≈ 1 K

### 3.4.3 Design of band pass 2nd active order filter

The bassband is a cascade of high pass a followed by a low pass filter with cutoff frequency of 1.2 KHz and 5 KHz respectively

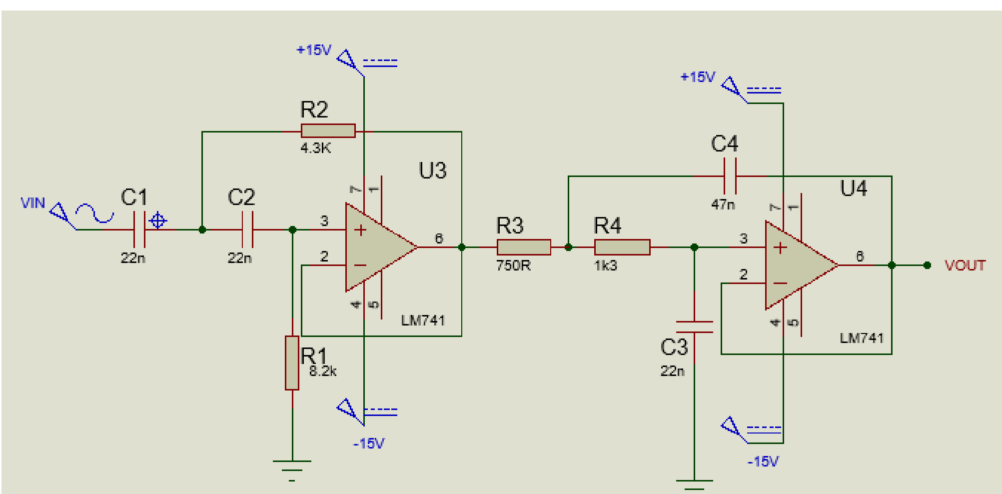


*Figure 21: band pass filter with cutoff frequency of 1.2 KHz and 5 KHz respectively*

After choosing the high pass filter capacitor values as follows

C = C1 = C2 = 22 nF, the resistor values were calculated as follows.

$$R1 = \frac{1}{\pi f1 \times C \times a1 \times 1000} \text{ K; where } f_1 \text{ is cutoff frequency for high pass filter.}$$

$$R1 = \frac{1}{\pi \times 1200 \times 22 \times 1.4142 \times 10^{-9} \times 1000} \text{ K}$$

R1 = 8.53 K

R1 ≈ 8.2 K (nearest standard value)

$$R2 = \frac{a1}{4\pi f2 \times C \times b1 \times 1000} \text{ K}$$

$$R2 = \frac{1.4142}{4\pi \times 1200 \times 22 \times 10^{-9} \times 1 \times 1000} \text{ K}$$

R2 ≈ 4.3 K

For the low pass filter, C3 was chosen as 22 nF

$$R_{2,3} = \frac{a1c4 \pm \sqrt{(a1)^2\,(C4)^2 - 4b1C3C4}}{4\pi f2C3C4}$$

In order to obtain real values under the square root, C3 must satisfy the following condition:

$$C4 \geq = \frac{C3 \times 4b1}{(a1)^2}$$

For second order Butterworth response, $a_1$=1.4142 and $b_1$= 1

$$C4 \geq \frac{22 \times 4 \times 1}{1.4142^2} \text{ nF}$$

C4 ≥ 44 nF

C4 = 47 nF (nearest capacitor standard value)

$$R_{3,4} = \frac{1.4142 \times 47 \times 10^{-9} \pm \sqrt{(1.4142)^2 (47 \times 10^{-9})^2 - 4 \times 47 \times 22 \times 10^{-18}}}{4\pi \times 5 \times 47 \times 22 \times 10^{-15}}$$

$R_3 \approx 0.75$ K

$R_4 \approx 1.3$ K

### 3.4.4 Design of preconditioning circuit

The preconditioning circuit is designed to give a gain or attenuation to the input signal in order to control the volume of audio signal.

Below is a simple implementation of the circuit using an opamp and resistor to give a maximum voltage gain of 5.

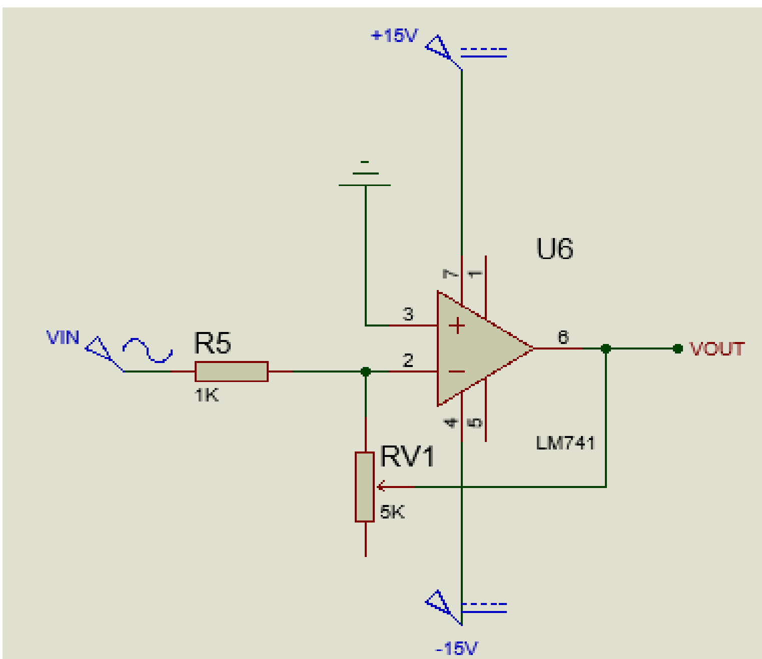


*Figure 22: Preconditioning circuit*

Maximum gain = $\frac{Rv1}{R5}$

$= \frac{5K}{1K}$

$= 5$

## Discussion and Conclusion

The simulated, calculated and hardware results were in harmony. The power transistors started heating a lot when the power amplifier was powered on. This was because the heat sink was small. A new heat sink was designed which was able to handle the power dissipated by transistors.

## Recommendation

All though a class AB power amplifier was build and functioned well, more improvements can be made to the circuit to make it better. They include:

1) Use of active elements to make a bootstrap circuit so that the amplifier can be more stable.
2) Use heat sinks with wider surface area to increase the power dissipation.
3) Power amplifiers with higher output power could be built by connecting more output power transistors in parallel.

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