DIGITAL STETHOSCOPE & HEART-RATE AMPLIFIER

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INTRODUCTION

This project aimed at designing a circuit which amplifies and filters heart sounds to signals listenable through headphones, as well as digitalising these noises to correspond to individual heart pulses in order to either drive a microcontroller-based counter, flash an LED or activate a beeper at each pulse. The design process consisted of 3 main stages:

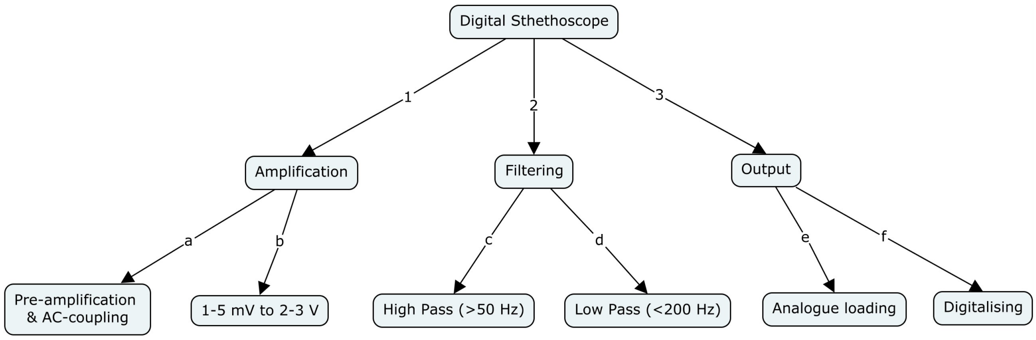

Figure 1: Circuit structure

Initial circuit drafts were first simulated using LTspice to obtain reasonable results, and then implemented physically on a breadboard. This report will first explore the design and simulation of the original model of the circuit, and then compare it with the actual build and any changes made, both in brief. The appendices that follow will attempt to explain the finer theory and reasoning behind each stage.

DESIGN

The following circuit was designed[1] using a bottom-up approach, where individual functional components such as the amplification stages, followed by the filtering stages and so on were designed separately and then put together:

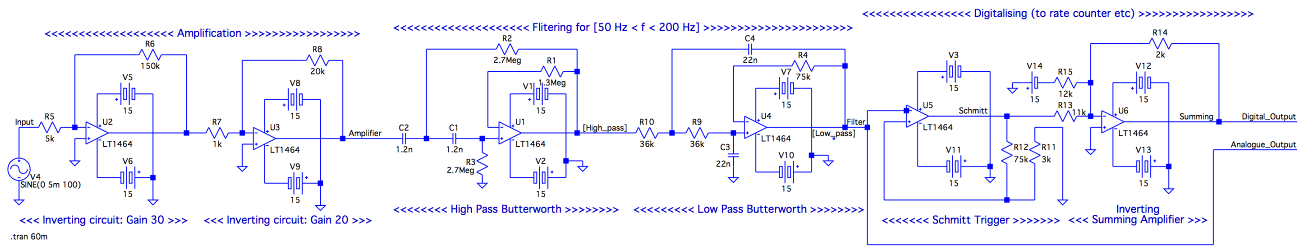

Figure 2: The design model (see *Appendix 11* for full-sized image)

[1] All resistor and capacitor values used in both design and implementation were with reference to the table in *Appendix 1* which lists out commonly available standard decade resistances and capacitances, to (a) make a feasible design model and (b) make real-life building of the circuit time-efficient.

1. AMPLIFICATION:

Amplification involved boosting the 1-5 mV range signals heart sounds usually result in, to a range of 2-3 V within which headphones best operate.

In order to determine the required gain, the following values of V_{IN} and V_{OUT} were chosen (see *Appendix 2* for reasoning):

$$V_{IN} = 5\,mV, \quad V_{OUT} = 3\,V$$

Hence;

$$\therefore Gain = \frac{V_{OUT}}{V_{IN}} = \frac{3}{5\times10^{-3}} = 600$$

This was achieved using two inverting amplifier circuits in series (see *Appendix 3*).

2. FILTERING:

On average, the normal frequency range of heart sounds is 40 – 200 Hz, with a peak (mode) at ≈100 Hz. The biggest source of noise in the circuit is the 'mains hum', i.e. the mains frequency noise at ≈50 Hz. A compromise was made and all sounds < 50 Hz and > 200 Hz was filtered out:

$$\therefore 50\,Hz < f_{required} < 200\,Hz$$

Filtering involved a high-pass Butterworth filter with $f_{cut\text{-}off}$ = 50 Hz and a low-pass Butterworth filter with $f_{cut\text{-}off}$ = 200 Hz (see *Appendix 4*). A high gain means small DC offsets towards the input of the circuit may be amplified to saturate the signal, but this wouldn't be a problem now as a DC signal has a frequency of 0 Hz. This is automatically filtered out by the high-pass filter.

Headphones may be used to listen to the amplified heart sounds signal, at this stage.

3. DIGITALISATION:

The circuit can also be used to trigger a rate counter, an LED or a buzzer which will produce a signal every time a heart sound is detected, the same pair of headphones which would make a "beep beep" at regular intervals, or a microprocessor which could calculate the heart rate using the period between each positive-going pulse. For this, digitalisation of the analogue signal is required to produce a 'spike' only when the input exceeds a certain characteristic threshold above which it is certainly a voltage due to a heart sound.

- A *Schmitt Trigger* is used for the first part of this purpose. As the smallest signal in the range that could be heard is around 1 mV, when amplified by a gain of 600, it gives out a corresponding 0.6 V. Hence, anything heard ≥ 0.6 V is most likely a heart sound: so, the $V_{threshold}$ was considered to be 0.5 V. Whenever this threshold is crossed, the trigger jumps between $V_{saturation}$ = ±13 V. Refer to *Appendix 5* for the theory behind Schmitt triggers, how the right resistor values were determined and a graphical representation of this stage's processing.

  Assuming that V_{OUT} of the Schmitt trigger was originally at +13 V, this means that V_I (potential difference between the two op-amp terminals) was also positive, so V_{IN} was < 0.5 V threshold. So whenever the input signal exceeds the 0.5 V threshold, the output goes to – 13V (negative $V_{saturation}$), and vice versa. The problem here is that the V_{OUT} signal is hence inverted.

- Digital often works between 0 V and 5 V, so the ±13 V saturation signal must be translated up to a zero to positive voltage scale by adding a DC bias, and 'shrunk' down. Hence an *inverting summing amplifier* was used to do both steps in one go. It acts as an inverting amplifier with a gain < 1 to attenuate the signal, followed by adding in a DC bias via another of its inputs (refer to *Appendix 6* for clear explanation and the two-step approach that was taken in designing the same).

The final circuit was assembled as a combination of all the different stages listed above, and tested on an AC sweep analysis. At its peak frequency of 100 Hz within the desired 50 – 200 Hz range,

$$dB = 20log_{10}\left(\frac{V_{OUT}}{V_{IN}}\right) = 20log_{10}(gain)$$

$$\therefore dB = 20log_{10}\left(\frac{3\ V}{5\ mV}\right) = 20log_{10}(600) = 55.6$$

Transient analysis[2] was also performed with a 5 mV, 100 Hz[3] voltage source as the simulation input. This was also repeated with a *.wav* file of a sample heart sound, as seen in *Appendix 7*.

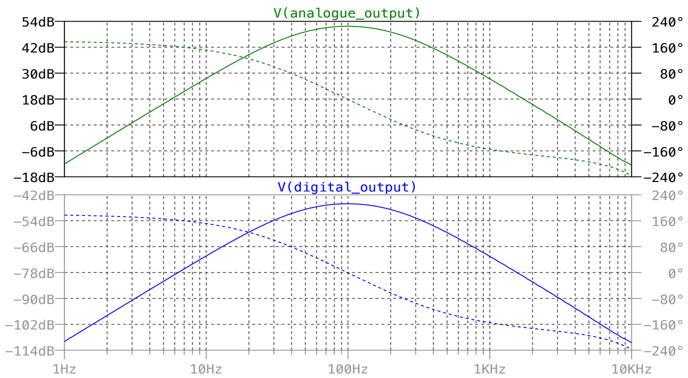

Figure 4: AC analysis and frequency response of the final circuit

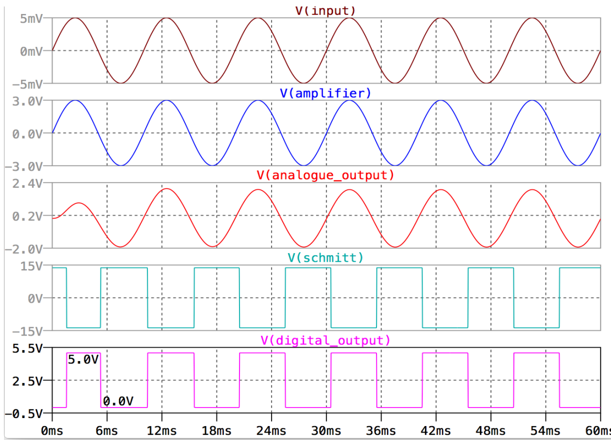

Figure 4: Transient analysis at each segmental output node of the final circuit

IMPLEMENTATION & DISCUSSION

The electrical stethoscope was built in accordance with the design simulated in SPICE. However, some changes had to be made to the original design to account for problems that occurred during the parallel implementation & testing processes (see *Appendix 8* for screenshots of the oscilloscope as the circuit was tested at each stage).

1. AMPLIFICATION:

a. *Microphone pre-amplification and AC-coupling*

The carbon microphone used in the circuit, assembled into a stethoscope bell, is a transducer: it acts as a variable resistor that changes resistance depending on the amount of sound energy it receives. It required a defined, constant low supply voltage supplied via a 10k resistor, forming a potential divider to convert this sound energy into corresponding electrical signals.

The capacitor is used as a method of AC-coupling, i.e. to remove the DC bias from the supply voltage.[4]

[2] Note that in Figure 4, *V(analogue_output)* also corresponds to the cumulative output signal after the filtering stage, and *V(digital_output)* corresponds to that after the inverting summing amplifier segment.
[3] 100 Hz: the most prominent heart sound frequency.

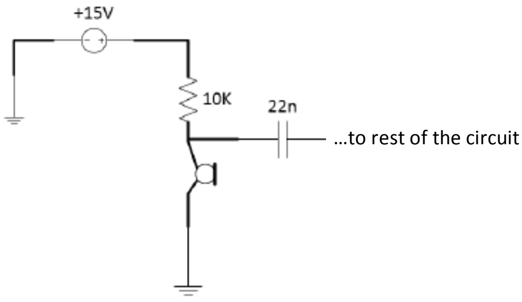

Figure 5: 'Initialising' the stethoscope bell microphone (Dickinson, 2015)

b. Gain stage

Whilst the circuit fundamentals were retained as in the design, the gain was boosted to 1200 by using one inverting amplifier of gain 30x and the other of gain 40x, i.e. by modifying R_1 to 1.2 kΩ and R_2 to 47 kΩ, due to (a) the signal being attenuated too low at the filtering stage, and (b) the analogue output of the heart sound being too faint to be easily heard, when tested practically.

All of the amplification was performed before any other stages when the required signal is still strong and noise is relatively low. The signal picks up more and more electrical noise as it progresses through the circuit, but since amplification was carried out right at the start, the amplitude of the desired signal far outweighs that of the cumulated noise, giving a high signal-to-noise-ratio. The filtering stage removes the mains hum, but afterwards there is still potential for the signal to be distorted, so filtering occurs after amplification.

2. FILTERING:

No changes were made in implementing the two filters as designed on SPICE. However, while the capacitor value used remained 22 nF as in the original design, the 68 kΩ and 33 kΩ resistors were substituted for with a 75 kΩ and 36 kΩ resistor respectively, considering the availability of resistors in the laboratory. This was acceptable as the ratio remained appreciably constant, and since the 'gold' resistors have a tolerance of ±5%.

The LM324 integrated circuit used in the lab was in itself a quad-core operational amplifier, i.e. each component was made up of four separate op-amps that could each be used separately. As a compromise between minimising and compacting the circuit at the trade-off of more exhaustive debugging and analysis of the same whenever required, two of the four available op-amps were used within each LM324 IC component (for instance, both gain stages were fitted in to a single LM324, and so were both filters)[5].

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[4] This works because a capacitor has an impedance of $\frac{1}{j\omega C}$ where $\omega = 2\pi f$ corresponds to the frequency of the signal. Since DC signals have a frequency of 0 Hz (constant supply), its impedance tends to ∞, thereby not letting any DC biases to pass through.

[5] Refer to *Appendix 10* for a LM324 block diagram.

3. OUTPUT:

a. Analogue loading and volume control

The following circuit was added in order to 'drive' the headphones –

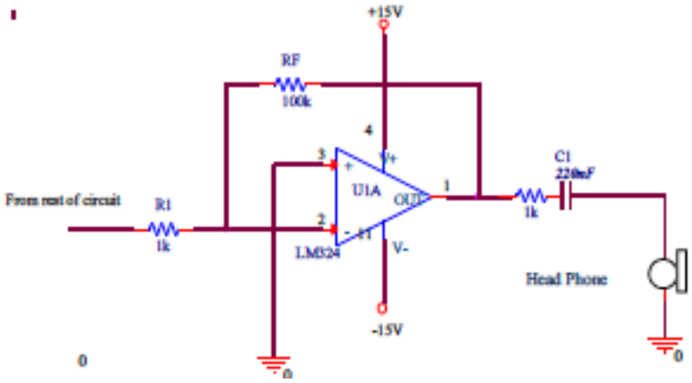


Figure 6: Driving the headphones (Dickinson, 2015)

Replacing R_f with a 100 kΩ variable resistor allowed volume control. The inverting amplifier feature of the circuit was a trade-off between more noise but a loud signal, or a clearer but heavily attenuated signal if resistors of equal value were used; but an op-amp was necessary at this stage (see *Appendix 9*).

b. Digitalising

The Schmitt trigger's $V_{threshold}$ was modified to 1 V by observation, so a potential divider was formed between a 100 kΩ and 680 kΩ ohm resistor instead[6]. The inverting summing amplifier's schematic and purpose remained unchanged albeit the resistor values were altered to suit the new supply voltage and Schmitt $V_{saturation}$: R_f = 8.2 kΩ, $R_{\text{loading the -9V DC bias}}$ = 30.9 kΩ, $R_{schmitt}$ = 22 kΩ.

The final built circuit, as a result of the numerous trial-and-error testing and debugging, yielded acceptable results. An image of the final breadboard and the final waveform observed is seen below.

Having an amplification stage just before driving the headphones, whilst preventing unnecessary *loading*, was at the cost of more noise in the circuit, so a possible improvement to the circuit is to use a unidirectional microphone instead, or to apply a sonic conductive gel between the chest skin layer and the microphone.

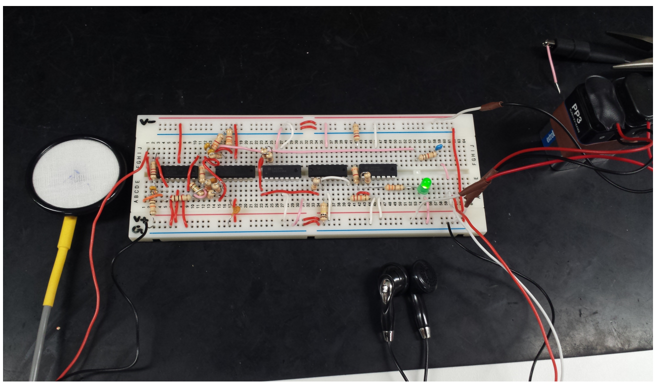


Figure 8: The final breadboard view

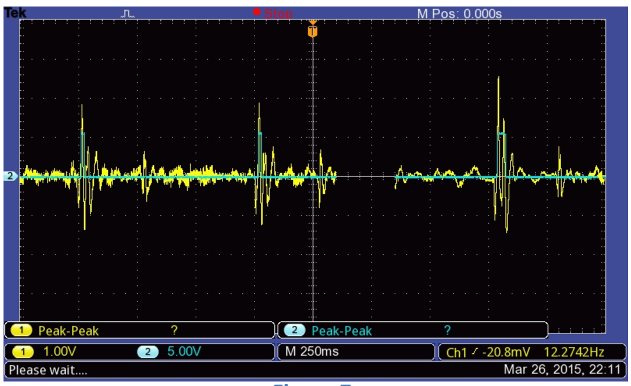


Figure 7:
Analogue output *(yellow)* vs. Digital output from 0-5 V *(blue)*

[6] Note that the $V_{saturation}$ is ±7 V since two 9 V batteries were used as supply voltages instead of a ±15 V power supply as designed, to increase mobility of the device.
This also affects the behavior of the microphone which produces electrical signals corresponding to the sound it receives depending on the defined supply voltage it is connected to: originally designed to have been 15V.

REFERENCES

3M™ Littmann® Stethoscopes. (2015). *Listen to Heart Sounds - Lung Sounds Audio - Breath Sounds - 3M™ Littmann® Stethoscopes - 3M Europe*. Retrieved 03 22, 2015, from 3M™ Littmann® Stethoscopes > Education > Auscultation Lessons > Heart and Lung Sounds: http://solutions.3m.com/wps/portal/3M/en_EU/3M-Littmann-EMEA/stethoscope/littmann-learning-institute/heart-lung-sounds/

Beavis, D. (n.d.). *Beavis Hi-Fi*. Retrieved 03 16, 2015, from http://beavishifi.com/articles/headphonejack/

Department of Electrical, Computer and Energy Engineering, University of Colorado Boulder. (n.d.). *Department of Electrical, Computer and Energy Engineering (ECEE) | University of Colorado at Boulder.* Retrieved 03 21, 2015, from http://ecee.colorado.edu/~mcclurel/resistorsandcaps.pdf

Dickinson, D. R. (2015, 02 13). BE1.HEEL - Electrical Engineering 1st Year Laboratory Handbook. 10.

Public Domain. (2009, 01 26). *Wikipedia.* Retrieved 03 22, 2015, from http://commons.wikimedia.org/wiki/File:Op-Amp_Inverting_Amplifier.svg

Texas Instruments. (1975, 09). *Quadruple Operational Amplifiers - LM324 Data Sheet.* Retrieved 03 22, 2015, from Texas Instruments: http://www.ti.com/lit/ds/symlink/lm324.pdf

The writers would also like to extend their deep appreciation towards Mr Paschal Egan and Mr Niraj Kanabar for their insight and assistance during the building of this project.

Appendix 1

Standard Resistor Values (±5%)						
1.0	10	100	1.0K	10K	100K	1.0M
1.1	11	110	1.1K	11K	110K	1.1M
1.2	12	120	1.2K	12K	120K	1.2M
1.3	13	130	1.3K	13K	130K	1.3M
1.5	15	150	1.5K	15K	150K	1.5M
1.6	16	160	1.6K	16K	160K	1.6M
1.8	18	180	1.8K	18K	180K	1.8M
2.0	20	200	2.0K	20K	200K	2.0M
2.2	22	220	2.2K	22K	220K	2.2M
2.4	24	240	2.4K	24K	240K	2.4M
2.7	27	270	2.7K	27K	270K	2.7M
3.0	30	300	3.0K	30K	300K	3.0M
3.3	33	330	3.3K	33K	330K	3.3M
3.6	36	360	3.6K	36K	360K	3.6M
3.9	39	390	3.9K	39K	390K	3.9M
4.3	43	430	4.3K	43K	430K	4.3M
4.7	47	470	4.7K	47K	470K	4.7M
5.1	51	510	5.1K	51K	510K	5.1M
5.6	56	560	5.6K	56K	560K	5.6M
6.2	62	620	6.2K	62K	620K	6.2M
6.8	68	680	6.8K	68K	680K	6.8M
7.5	75	750	7.5K	75K	750K	7.5M
8.2	82	820	8.2K	82K	820K	8.2M
9.1	91	910	9.1K	91K	910K	9.1M

Standard Capacitor Values (±10%)						
10pF	100pF	1000pF	.010μF	.10μF	1.0μF	10μF
12pF	120pF	1200pF	.012μF	.12μF	1.2μF	
15pF	150pF	1500pF	.015μF	.15μF	1.5μF	
18pF	180pF	1800pF	.018μF	.18μF	1.8μF	
22pF	220pF	2200pF	.022μF	.22μF	2.2μF	22μF
27pF	270pF	2700pF	.027μF	.27μF	2.7μF	
33pF	330pF	3300pF	.033μF	.33μF	3.3μF	33μF
39pF	390pF	3900pF	.039μF	.39μF	3.9μF	
47pF	470pF	4700pF	.047μF	.47μF	4.7μF	47uF
56pF	560pF	5600pF	.056μF	.56μF	5.6μF	
68pF	680pF	6800pF	.068μF	.68μF	6.8μF	
82pF	820pF	8200pF	.082μF	.82μF	8.2μF	

(Department of Electrical, Computer and Energy Engineering, University of Colorado Boulder)

Appendix 2

Secondary research[7] returned an average driving voltage of headphones similar to those used in the lab to be roughly 2.84 V, so a 3 V V_{OUT} value was chosen.

Since the maximum output of 3 V was chosen, V_{IN} was set to its upper bound of 5 mV in calculation, because if 1 mV was picked instead and amplified to 3 V, this would result in a gain of 3000:

$$Gain\ required = \frac{V_{OUT}}{V_{IN}} = \frac{3}{1\times10^{-3}} = 3000$$

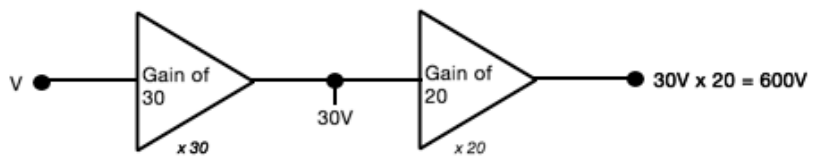
If so, signals close to 5 mV would be overly amplified along with any accompanying noise, creating unwanted distortions and clipping:

$$V_{OUT} = V_{IN} \times Gain = 5\ mV \times 3000 = 15\ V$$

Appendix 3

<u>Using op-amps in series</u>

The LT1464 op-amp used in simulation, as well as the similar LM324 IC used in the lab, has a maximum gain of ≈200, before behaving unstably. Hence two op-amps connected in series, with gains 30 and 20, were used in design:



$$Total\ Gain = 30 \times 20 = 600$$

Op-amps follow a relationship similar to:

$$(200)^n = 600$$
$$log_{200}(600) = n$$
$$\therefore n \approx 1.2$$

where n = number of op-amps required, given that the first op-amp is at maximum gain of 200. This is not just impractical and rather difficult, but also 'pushes the limit' of the op-amp.

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[7] (Beavis); see references.

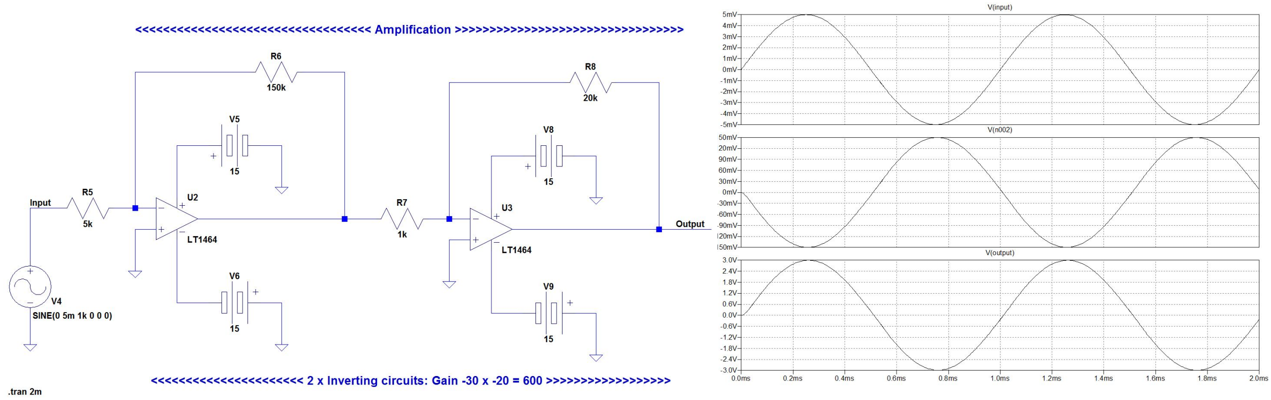


Figure 9: The amplification design stage and the output of each op-amp

NOTE:

This works since the V_{IN} of both stages is still < 1 V and the op-amp continues to operate in its linear region.

The inverting amplifier

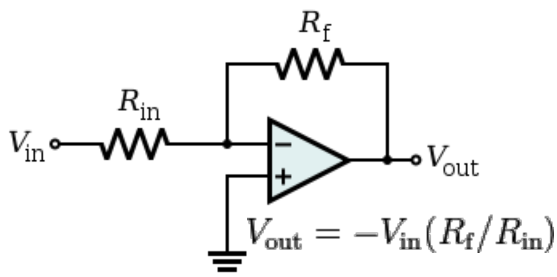
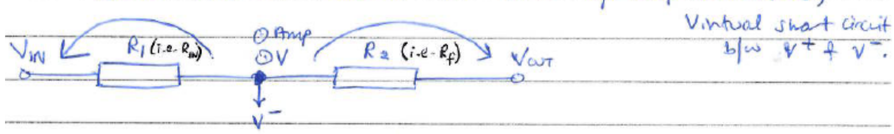


Figure 10: The inverting amplifier (Public Domain, 2009)

Since the op-amp operates in its linear region with infinitesimally small V_I between its positive and negative terminals, a *Virtual Short Circuit* can be assumed between V^+ and V^-. Hence, its characteristic equation can be derived as follows:

Zero current drawn at either op-amp terminals,

Virtual short circuit b/w V^+ & V^-.



$V = IR$; $I = V/R$

$$\frac{V_{IN}}{R_1} = \frac{-V_{OUT}}{R_2}$$

$$\therefore \frac{V_{OUT}}{V_{IN}} = -\frac{R_2}{R_1}$$

For op-amp 1,

$$Gain = -\frac{R_f}{R_{in}}$$
$$-30 = \frac{150\Omega}{5\Omega}$$
$$\therefore R_2 = 150k\Omega$$
$$\therefore R_1 = 5k\Omega$$

Rather than 150 Ω / 5 Ω being used, 150 kΩ / 5 kΩ was chosen for adequate resistive behaviour, as even a normal wire may have resistances in the same order of 10^0 - 10^1 Ω (≈1 Ω).

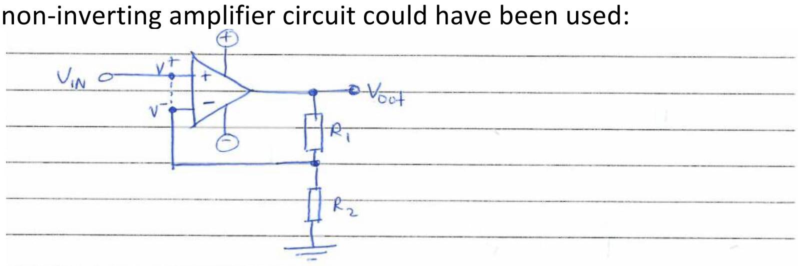
600 kΩ/20 kΩ or any combination giving a quotient of 30 can be used.

Similarly, for op-amp 2,

$$Gain = -\frac{R_f}{R_{in}}$$
$$-20 = \frac{12k\Omega}{1k\Omega}$$
$$\therefore R_2 = 20k\Omega$$
$$\therefore R_1 = 1k\Omega$$

The non-inverting alternative and the reasons against it

Alternatively, a non-inverting amplifier circuit could have been used:



Assuming a virtual short circuit b/w V^+ and V^- and since no current is drawn in at the (−) terminal;

$$V^- = \frac{R_2}{R_1+R_2} \times V_{out} \quad \text{(voltage divider)}$$

And; $V_{IN} = V^+ = V^- = \frac{R_2}{R_1+R_2} V_{out}$

$$\frac{V_{OUT}}{V_{IN}} = \frac{R_1+R_2}{R_2} = \frac{R_1}{R_2} + \frac{R_2}{R_2} = \frac{R_1}{R_2} + 1$$

$$\therefore \frac{V_{OUT}}{V_{IN}} = 1 + \frac{R_1}{R_2} \quad \text{(amplification)}$$

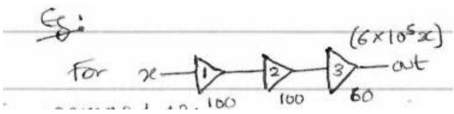
However, its characteristic equation yields the following:

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

$$30 = \frac{R_1}{R_2} + 1$$

$$\therefore \frac{R_1}{R_2} = 29$$

Choosing a pair of available resistors to accurately make this ratio can be difficult. Also, a small change in gain in one op-amp can have large effects on the total gain, as demonstrated below.



For op-amps 1 and 2:
If required gain was 100 and R_1=100Ω, then

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

$$\frac{R_1}{R_2} = gain - 1 = 100 - 1 = 99$$

$$R_2 = \frac{R_1}{99} = \frac{100\Omega}{99}$$

$$R_2 = 1.01\Omega$$

If instead the R_2 was chosen to be 1Ω (a seemingly good approximation),

$$gain = \frac{100\Omega}{1\Omega} + 1 = 101$$

If this approximation was to be made for both amplification stage 1 and 2 the total gain would be given by:

$$gain_{total} = 101 \times 101 \times 60 = 612060$$

This is to be compared to the intended gain:

$$gain_{total} = 100 \times 100 \times 60 = 600000$$

This shows that even a small offset of the resistor values can result in a gain that may significantly differ from the intended gain.

Hence, its use was avoided.

Appendix 4

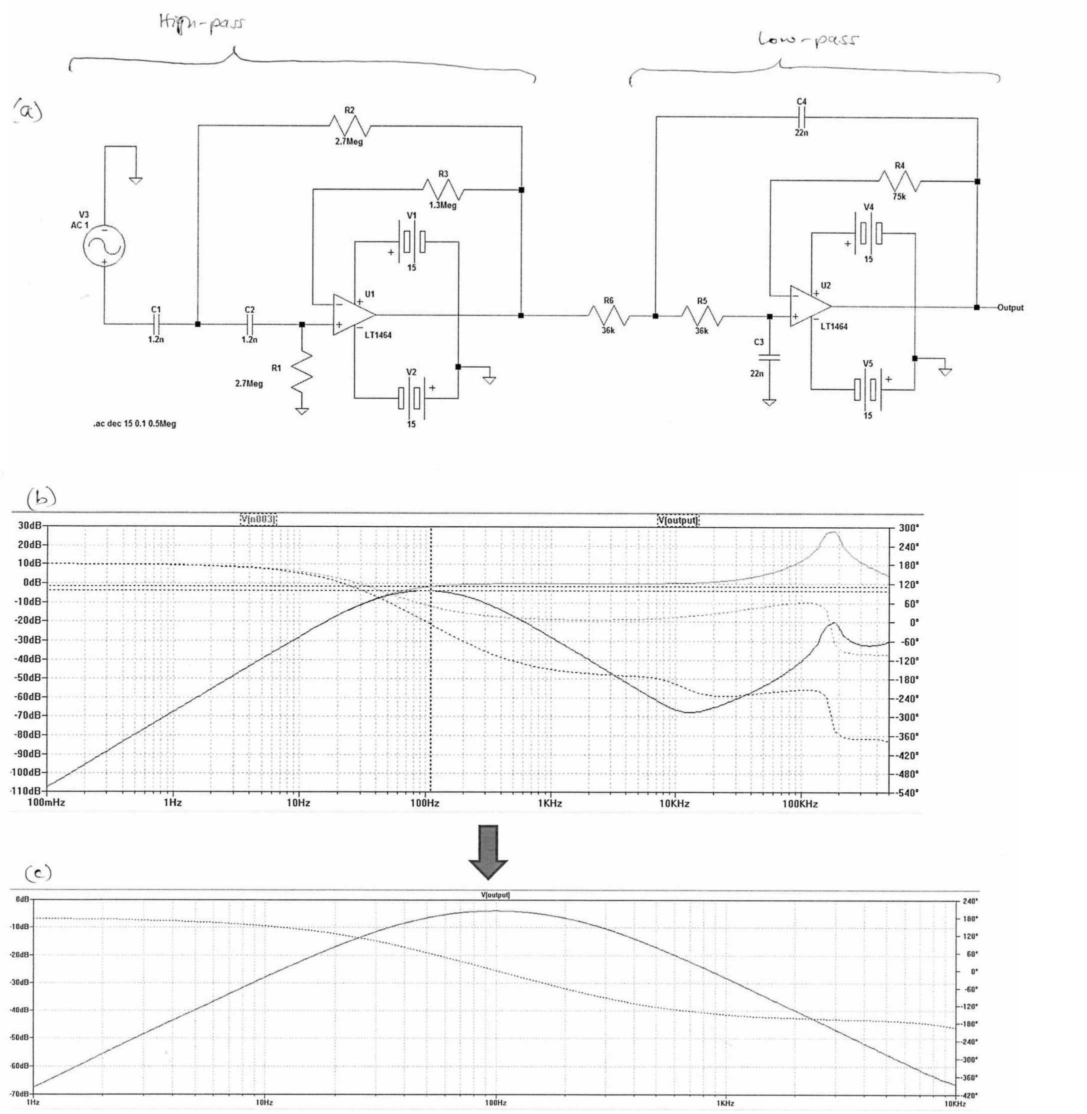


Figure 11: (a) the Filtering stage, (b) distortions at high frequencies, (c) its corrected frequency response curve

A Butterworth filter is a circuit that controls how much of an input signal is passed back out as an output signal depending on its frequency. Those seen in Figure 5(a) are *second-order* Butterworth filters with roll-offs of -40 dB/decade. This means that after exceeding the cut-off frequency of the filter, the output voltage (or amplitude) drops by a factor of 100 (i.e. 20log[100] = 40 dB) for every tenfold increase (or decrease) in frequency: corresponding to the slope of the cut-off region in the filter's frequency response curve.

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If at a certain frequency, a filter results in no amplitudinal change in its output with respect to its input signal (i.e. the signal is not *attenuated*), then it is at 0 dB:

$$dB = 20log_{10}\left(\frac{V_{OUT}}{V_{IN}}\right);\ V_{OUT} = V_{IN}$$
$$dB = 20log_{10}(1)$$
$$= 20 \times 0$$
$$\therefore dB = 0$$

<u>High-pass filter</u>

Butterworth with cut-off frequency, f_C = 50 Hz:

$C_1 = C_2 = \underline{\underline{1.2}}\ nF$ (i.e. 1200pF) arbitrarily.

$R_1 = R_2 = \frac{1}{2\pi f_c C_{1=2}} = \frac{1}{2\pi \times 50 \times 1.2 \times 10^{-9}}$

$= 2.652... \times 10^{6}$

$= \underline{\underline{2.7}}\ M\Omega$

(this value exists on the resistor look-up table)

$R_3 = \frac{1}{2} R_1 = \underline{\underline{1.3}}\ M\Omega$

<u>Low-pass filter</u>

Butterworth with cut-off frequency, f_C = 200 Hz:

$C_1 = C_2 = \underline{\underline{220}}\ nF$ (i.e. $0.22 \mu F$) chosen via trial & error, such that both cap. & res. values calculated exist in the data sheet.

$R_1 = R_2 = \frac{1}{2\pi f_c} = \frac{1}{2\pi \times 200 \times 220 \times 10^{-9}}$

$= 3617.16...$

$= \underline{\underline{3.6}}\ k\Omega$ (exists!)

$R_3 = 2R_1 = \underline{\underline{7.2\ k\Omega}}$ (so, ≈ 6.8k – 7.5k should work since resistors have ±5% tolerance)

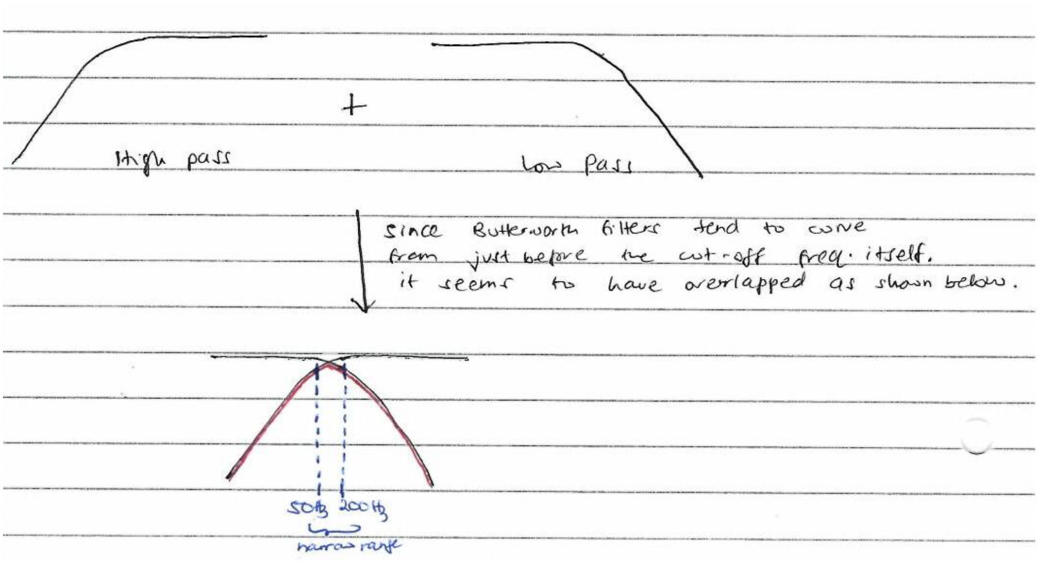
<u>Distortions</u>

As seen in Figure 5(b), unwanted peaks at > 100 kHz were observed when the AC sweep analysis was carried out on LTspice. These distortions are normal as the LT1464 behaves abnormally at very high frequencies; but this wouldn't affect the circuit's functionality, as the heart doesn't produce such high-pitched noises.

However, to minimise this, $C_{1,2}$ was changed to 22 nF (decreased by a factor of 10) and $R_{1,2}$ increased to 36 kΩ & R_3 to 75 kΩ (factor of 10 as well). Hence, the combined high and low-pass filters in series produced the much more refined curve seen in Figure 5(c).

Attenuation at its peak

Since the required frequency range of 50 Hz – 200 Hz is very narrow, and its roll-offs of -40 dB/decade were relatively low for this context, the filter combination never reached 0 dB.

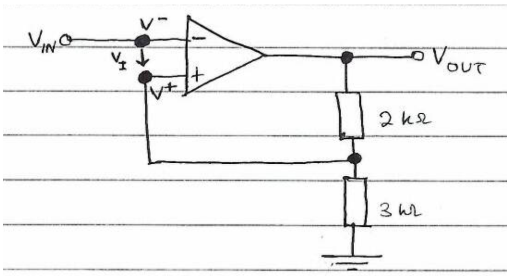


However, the peak at 100 Hz has an attenuation of -4 dB, so the pass band (generally considered to be within -3 dB of the gain at the curve's peak) is -4 dB – 3 = the corresponding frequencies at -7 dB.

At -7 db, f = 46 Hz and 214 Hz, which is well within the required 50 – 200 Hz range.

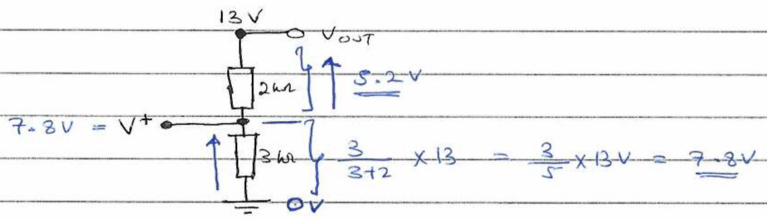
This behaviour is normal of wide band-pass filters.

Appendix 5



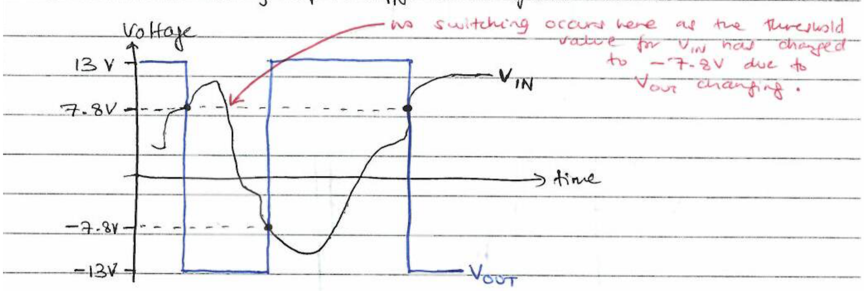
Usually, for this model of op-amp with a steady 15V power supply, its $V_{saturation}$ is $\pm 13V$. So, if V_I is ⊕ve it would go to +13V & if ⊖ve, to −13V.

In this example (and often otherwise too) we assume V_{OUT} is initially at +13V. Because no current flows into either V^+ or V^-, the current $\left(\frac{13\,V}{5\,k\Omega} = 2.6\,mA\right)$ flowing through both resistors is the same. Modelling it as a voltage divider, we find $V^+ = 7.8\,V$.



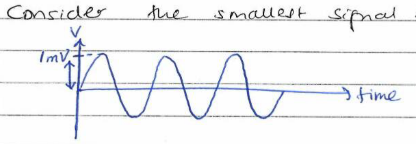
Since we have assumed V_{OUT} to be ⊕ve, from the nature of an op-amp's characteristics, then V_I must also be positive. This means, $V^- = V_{IN}$ must be less than $V^+ = 7.8\,V$.

If V_{IN} exceeds 7.8 V, V_I becomes ⊖ve (as V^+ is still at 7.8V), so V_{OUT} quickly changes to $-V_S$, i.e. −13V. Now, the voltage divider formed by the two resistors places a new threshold voltage of −7.8V at V^+. Similar to before, for V_{OUT} to now remain at −13V, V_I must be ⊖ve this time round, and this is the case only if V_{IN} is greater than −7.8V.

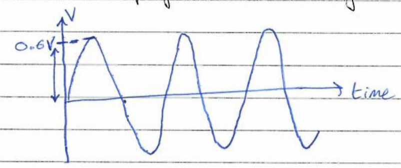


In the context of the heart-rate amplifier circuit,

A) Consider the smallest signal that could be heard:



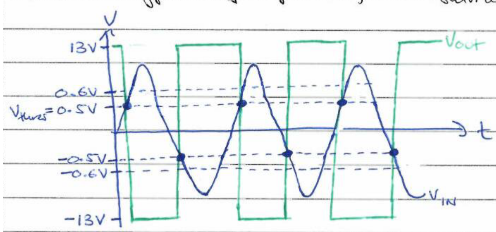
When amplified with a gain of 600:
(i.e. 1mV × 600)



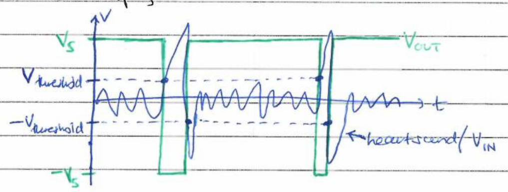
So anything heard ≥ 0.6V is certainly a heart sound.

The $V_{threshold}$ can be considered ≈ 0.5V then.

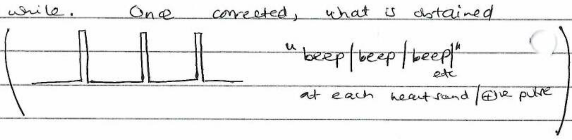
Whenever this threshold (0.5V) is crossed, the trigger jumps b/w $V_{saturation}$ of $\pm V_s = \pm 13V$.



In real life;



The problem here is that the V_{OUT} signal is inverted - it 'dips' at each heart sound, and is at its peak all the other while. Once corrected, what is obtained resembles:



$$V_{threshold} = \left(\frac{R_2}{R_1 + R_2}\right) \cdot V_{saturation}$$

$$\frac{0.5V}{13V} = \frac{R_2}{R_1 + R_2}$$

$$\frac{1}{26} = \frac{R_2}{R_1 + R_2} = \frac{3}{78} = \frac{3k\Omega}{75k\Omega + 3k\Omega}$$

$$\therefore R_2 = 3k\Omega$$

$$\therefore R_1 = 75k\Omega \quad \text{(they exist!)}$$

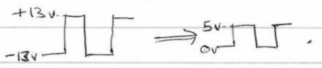
Note that the trigger works since the input signal is now well over a volt (after amplification) and hence the op-amp is *not* operating within its linear region, but acts as a comparator, switching between its two well-defined states of $V_{saturation}$.

Appendix 6

Digital often works between 0v & 5v (GND & VCC).

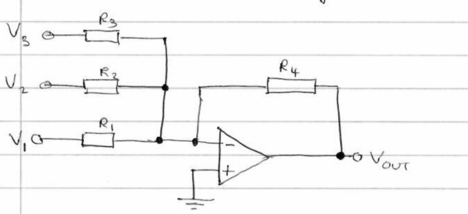
Anything < 0.2v is considered 0 (i.e. 0v) and anything > 3.5v is considered 1 (i.e. 5v).

In order to convert the ± 13 V saturation signal to a 0 − 5v signal, a DC bias must be added, and the signal "shrunk" to be



Another issue is that the Schmitt Trigger's signal is inverted – it peaks (+13 v) when there is no heart signal, and 'dips' into a trough (−13 v) at a pulse.

An inverting summing amplifier is used:



A resistor R_5 between the V^+ terminal of the op-amp, and Ground, is optional, such that $R_5 = R_4$.

$$V_{OUT} = -R_4\left(\frac{V_1}{R_1} + \frac{V_2}{R_2} + \frac{V_3}{R_3} + \ldots + \frac{V_n}{R_n}\right)$$ for any 'n' inputs.

Two-step approach –

Step ① →

Assume only one input: this was identical to the inverting amplifier.

$$V_{OUT} = \frac{-R_4}{R_1} V_1$$

$$-2.5\ v = \frac{-R_4}{R_1} \times 13.86\ v$$

Schmitt trigger saturates b/w ± 13.81v (never exactly ±13 v or ±15v).

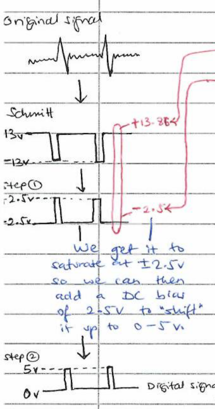
$$\frac{R_1}{R_4} = \frac{13.86}{2.5} = 5.544 \approx 5.5$$

$$\Downarrow$$

$$\frac{11\,k\Omega}{2\,k\Omega} = 5.5$$

$$\therefore R_1 = \underline{\underline{11\,k\Omega}}, \quad R_4 = \underline{\underline{2\,k\Omega}}$$

(see Fig. "Exe 9 – ③ b")



Step ② →

Use the inverting summing amplifier, and a new DC input, to "shift" the wave up to 0 − 5v (using a DC bias).

PTO →

$$V_{OUT} = -R_4\left(\frac{V_1}{R_1} + \frac{V_2}{R_2}\right)$$

$$5V = -2k\Omega\left(\frac{-13.36V}{11k\Omega} + \frac{V_2}{R_2}\right)$$

$$-\frac{5}{2} + \frac{13.36}{11} = \frac{V_2}{R_2}$$

$$-\frac{31}{25} = \frac{V_2}{R_2}$$

$$-1.24 = \frac{V_2}{R_2}$$

Let V_2 also be (−)15 V;

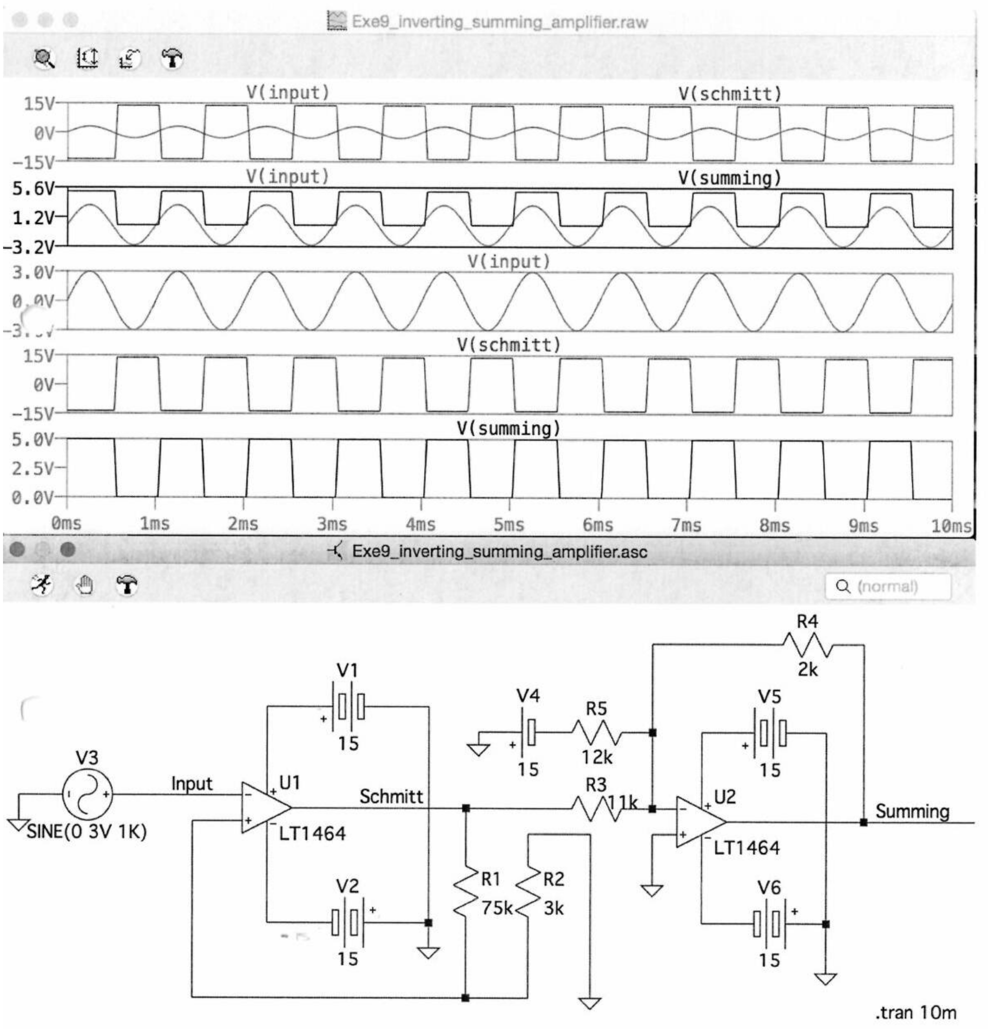
$$R_2 = \frac{-15\,V}{-1.24}$$

$$R_2 = 12.096\ldots$$

$$\therefore R_2 \approx 12\ k\Omega$$

$$\therefore \underline{\underline{V_2 = -15\,V}},\quad R_2 = \underline{\underline{12\ k\Omega}}$$

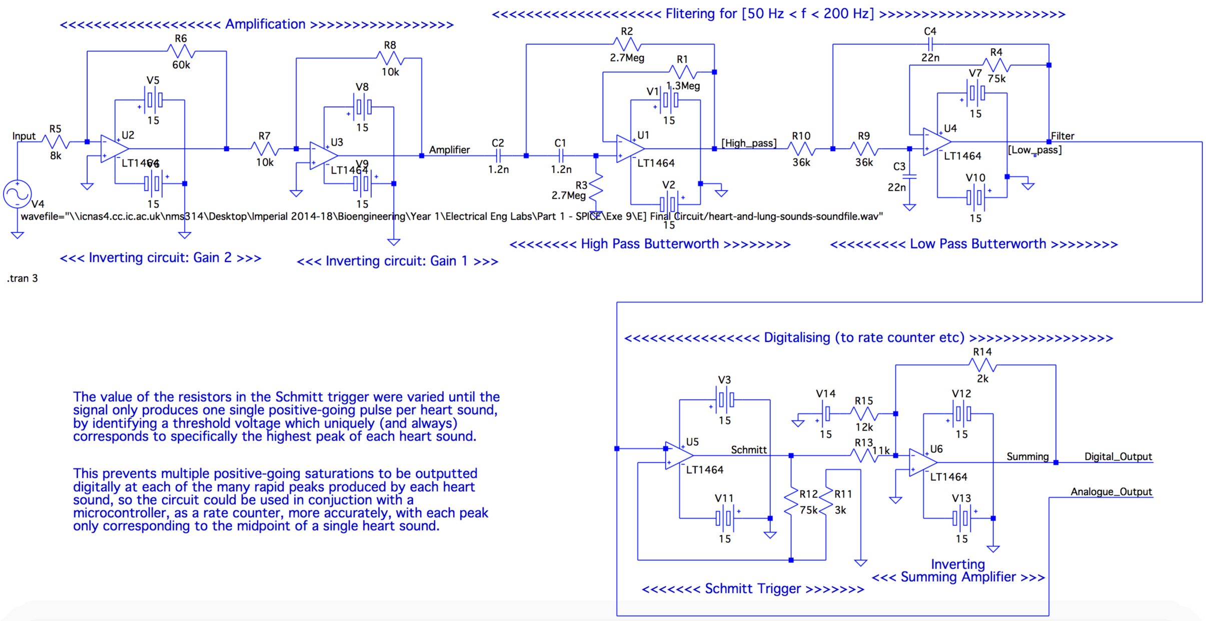
Simulation of the digitalisation segment (Schmitt trigger + inverting summing amplifier) yielded:

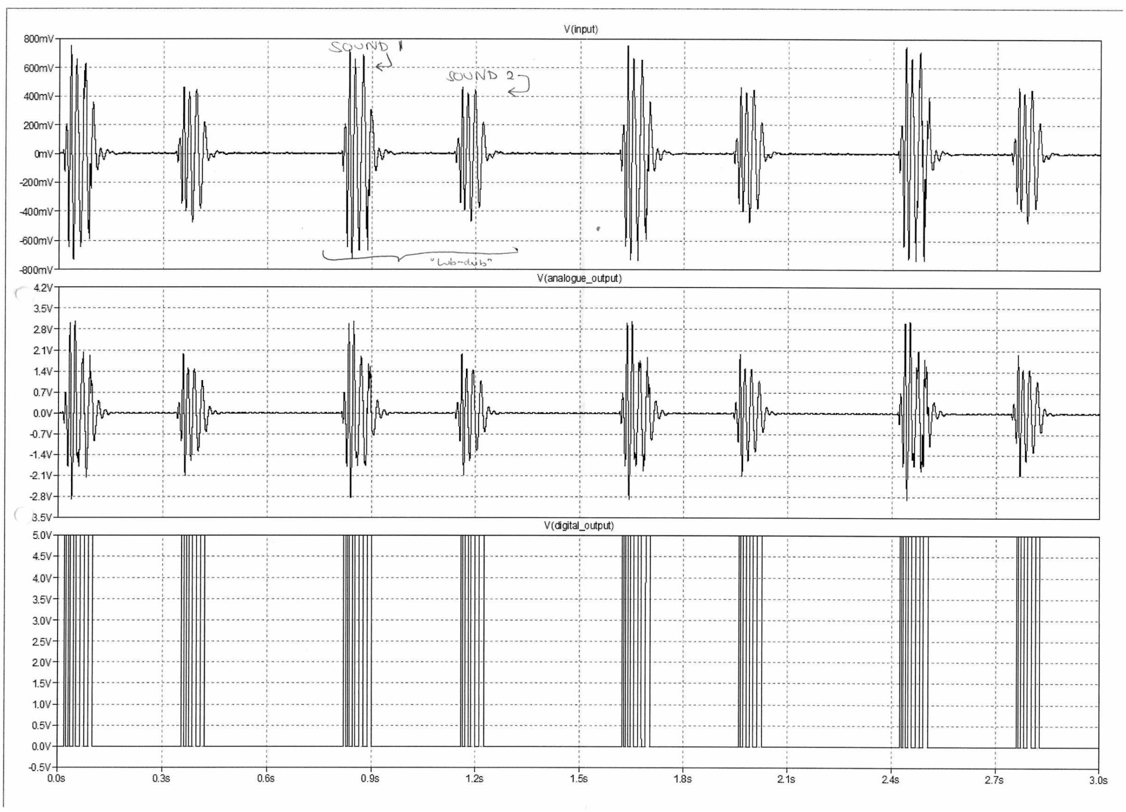


Appendix 7

The *.wav* file used was courtesy of 3M™ Littmann® Stethoscopes.

It was already amplified to ≈800 mV. This triggers the digital output to always tend towards +5V, as 800 mV yields an output of 480 V, which is well above the $V_{threshold}$ of the Schmitt trigger. Hence, the amplification was greatly reduced, as seen in the circuit used for simulation below. Since only one op-amp was adequate, the other was set to simply re-invert the signal back to normal with no change in amplitude (i.e. gain of 1, R_1 = R_2).





Appendix 8

<u>Stage by stage testing of the final circuit</u>

Amplification stage:

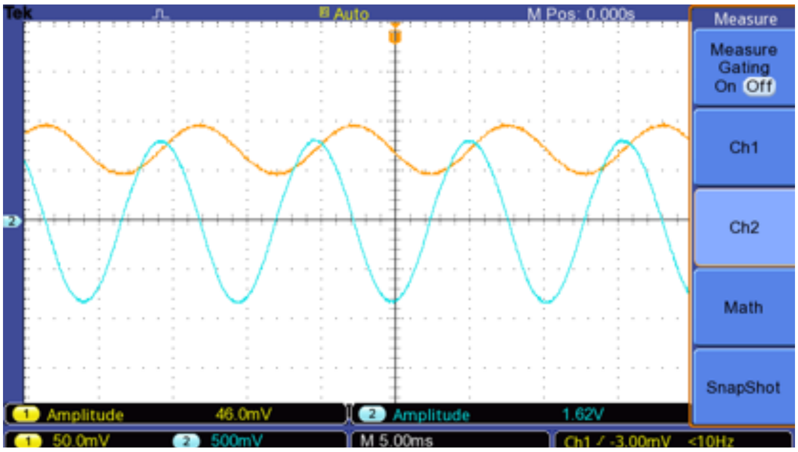


Figure 12: The gain of the final circuit was first increased to 900

The amplification was first increased to 30 x 30 = 900 gain for smaller signals in the range of 1 mV to be picked up better, as it was observed through the SPICE simulation itself that the filtering tends to attenuate the signal too low below 0 dB. Later it was further increased to a gain of 1200.

It is evident that the input signal straight from the microphone (i.e. Channel 1 – orange) has a DC offset seen by its translation upwards in the above diagram. However, the output signal after the amplification signal (i.e. Channel 2 – blue) is centered at zero, proving that the AC-coupling capacitor has indeed removed off any DC biases.

Filtering stage:

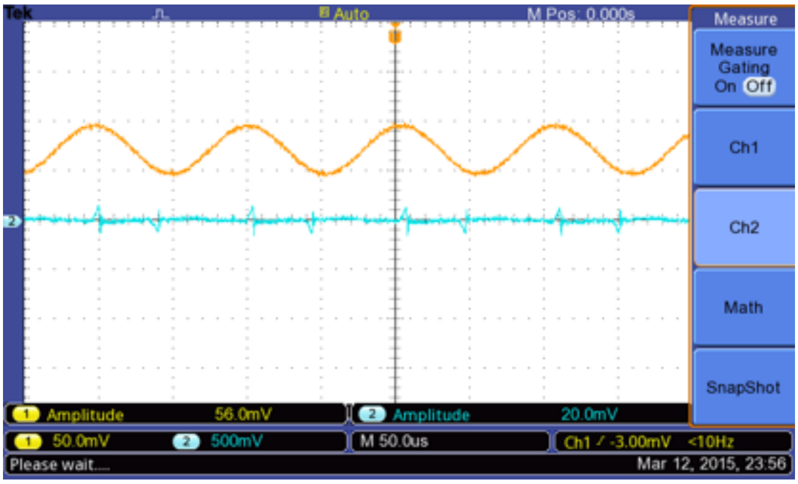


Figure 13: Testing the filtering stage, 50 < f < 200

It is evident that at very low or high frequencies (below or above a peak of 100 Hz), the output signal (blue) is heavily attenuated to almost 0.

The images below describe the behavior of the low-pass filter specifically:

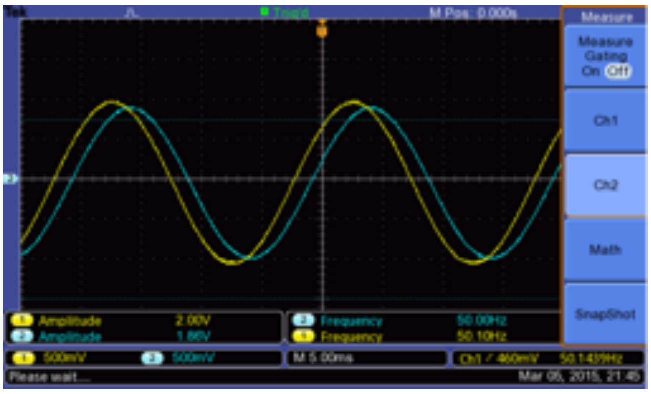


Figure 14: Low-pass, 50 Hz

At a frequency of 50 Hz, well below the 200 Hz cut-off, there is hardly any attenuation and ‘what goes in comes out’ at 0 dB.

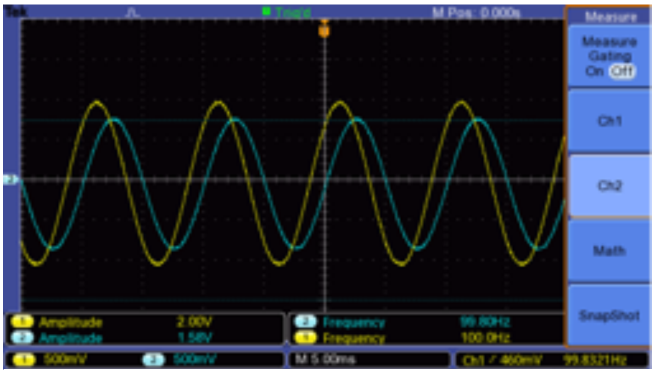


Figure 15: Low-pass, 100 Hz

As the frequency approaches f_C , at 100 Hz, V_{IN} = 2.0 V and V_{OUT} = 1.6 V.

So, dB = 20log(1.6/2.0) = -1.9

This demonstrates the roll-off nature of 2nd order Butterworth filters and how, since its frequency response ‘curves’ about f_C, there is always some attenuation.

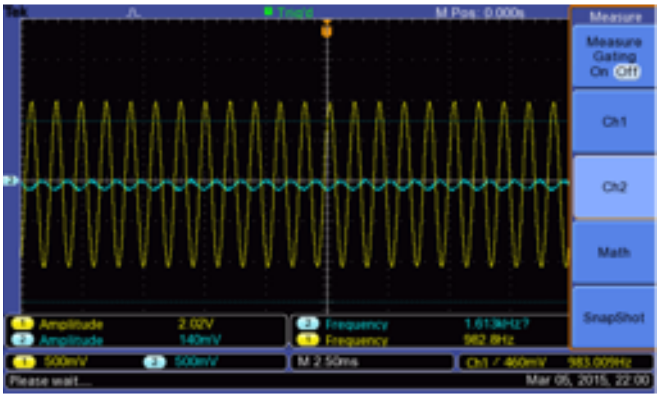
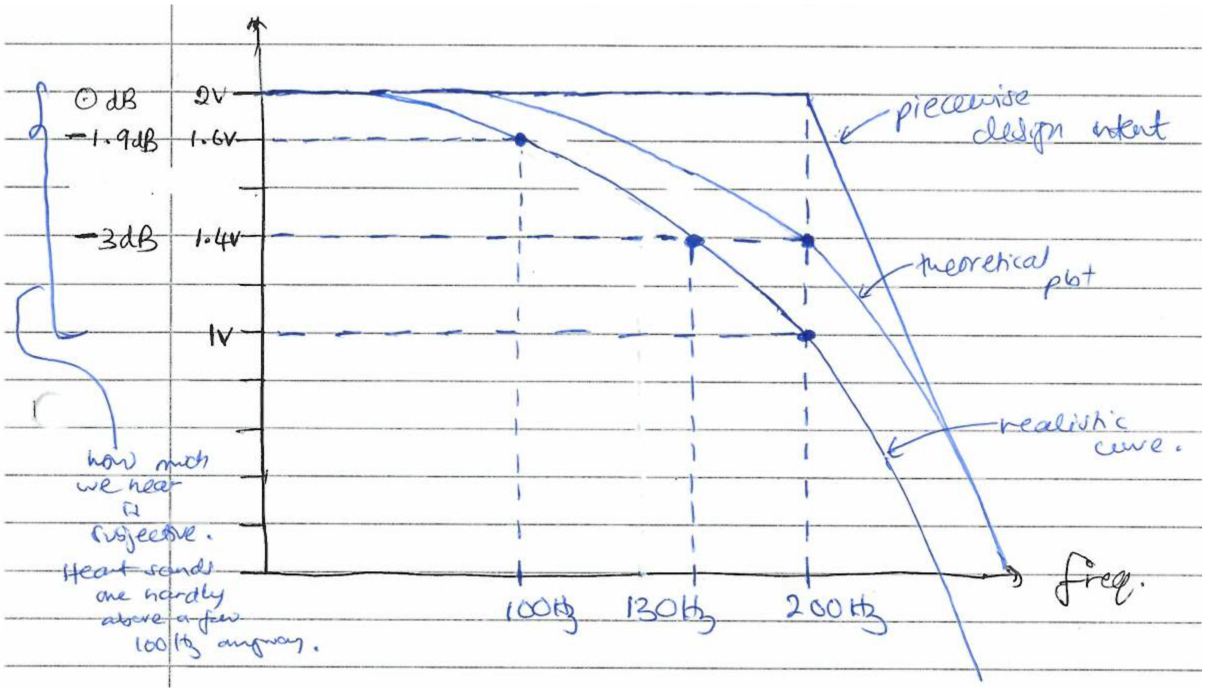


Figure 16: Low-pass, 1 kHz

For frequencies way above the low-pass cut-off of 200 Hz, such as in Figure 12 (≈1000 Hz), the output is heavily attenuated and no more than a few millivolts is obtained out of the original 2 V.

[PLEASE TURN OVER]

Thus, by inspection:



Schmitt trigger:

As a result of the relatively high signal-to-noise ratio, it was possible to use a high and distinct $V_{threshold}$ of 1 V, as shown below.

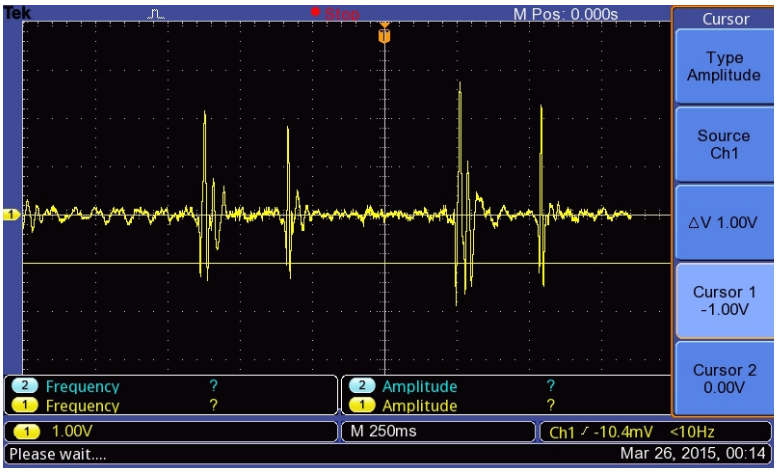


Figure 17: $V_{threshold}$ = 1.00 V

[PLEASE TURN OVER]

The resulting waveform,

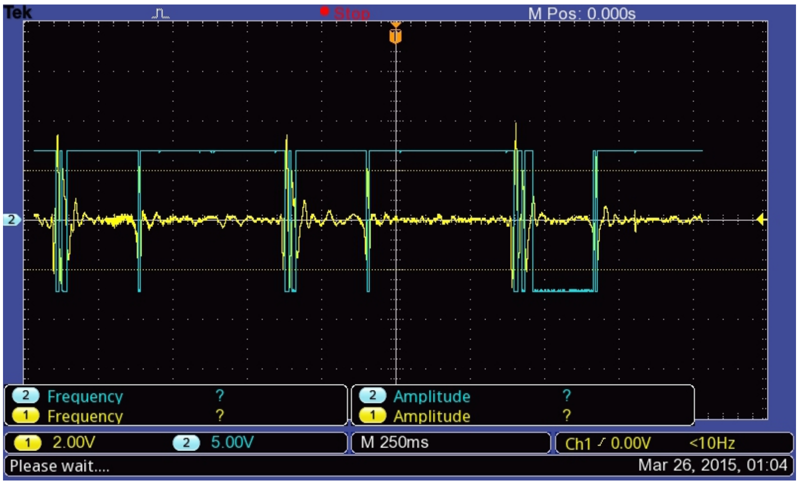


Figure 18: Testing the Schmitt trigger, $V_{threshold}$ = 1 V, $V_{saturation}$ = ±7 V

This was finally connected to the *inverting summing amplifier*, screenshot equivalent to the final output seen in Figure 7.

Appendix 9

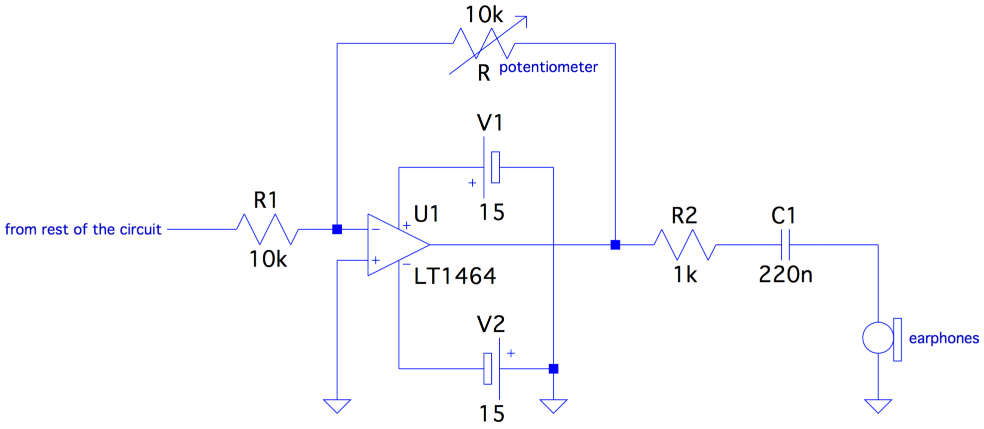


Figure 19: Schematic of the volume control unit

The original design idea for a volume control & headphone driving unit, shown above, was an inverting amplifier circuit which used a potentiometer[8] with maximum resistance 10 kΩ equivalent to the value of R1, to form a volume control unit with a gain that could be varied between 0 and 1. This was to be an important safety feature to avoid over-amplification of large noises, if at all, and to also avoid clipping. Another 1 kΩ resistor and 220 nF capacitor was implemented just before the headphone to provide a loading voltage and to further remove any and all DC biases as this would result in an unnecessary constant 'buzz'.

However, this circuit seemed to attenuate the signal way below a listenable range. It was tested that the 220 nF capacitor further added to this impedance. Attempting to increasing R_f and R up to 100 kΩ also still failed.

It was then decided that this stage was removed and the output from the filters were directly routed through a 10 kΩ variable resistor as a volume control. Since a notable DC noise was observed, this was routed via a capacitor, to the headphones.

$$Capacitor\ impedance = \frac{1}{j\omega C}$$

Since its reactance is inversely proportional to Capacitance, a 2200 µF capacitor was used to ensure its attenuation on the signal was at minimum. The problem here is that, higher the capacitance, relatively much larger its physical size, reducing practicality and portability. This is an issue even in the mobile phone industry etc. hence research into producing semiconductor-based nano-sized capacitors is a budding new field.

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[8] By using only two 'legs' of the potentiometer, i.e. one fixed pin and the 'wiper', it acts as a variable resistor or rheostat.

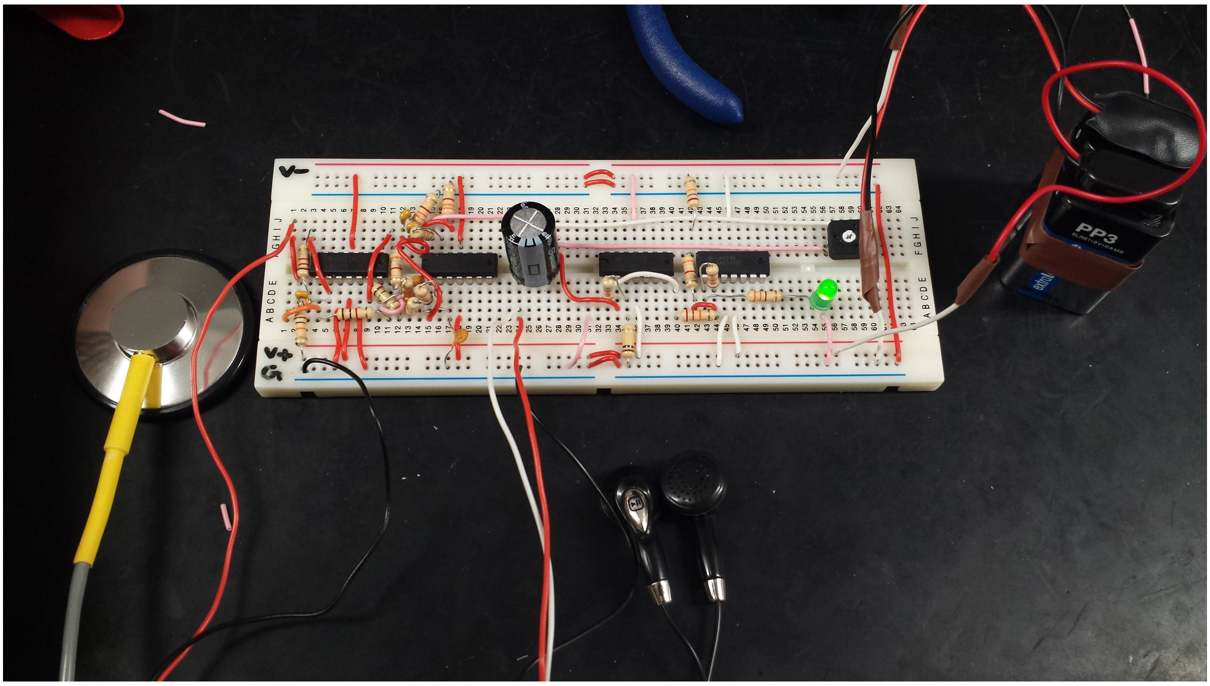

Figure 20: The first complete circuit produced a very clear signal but had one main flaw, headphone *loading*

This produced a very clear signal with a very high signal-to-noise ratio. However, an interesting phenomenon was noticed:

- When the headphones were completely disconnected, the LED, which was meant to flash at each instance of a heart sound (connected to the inverting summing amplifier), worked flawlessly. This is the waveform observed at the output of the filtering stage, at the junction which splits into the analogue output stage and the digitalising stage –

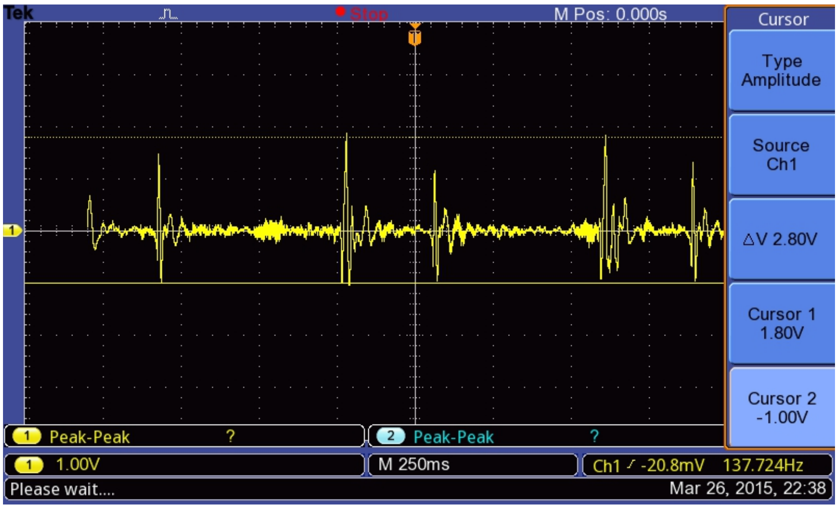

Figure 21: 2.80 V peak-to-peak with no headphones

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- When one channel of the headphones were connected, the signal at the same junction was slightly attenuated, but the LED wasn't badly affected –

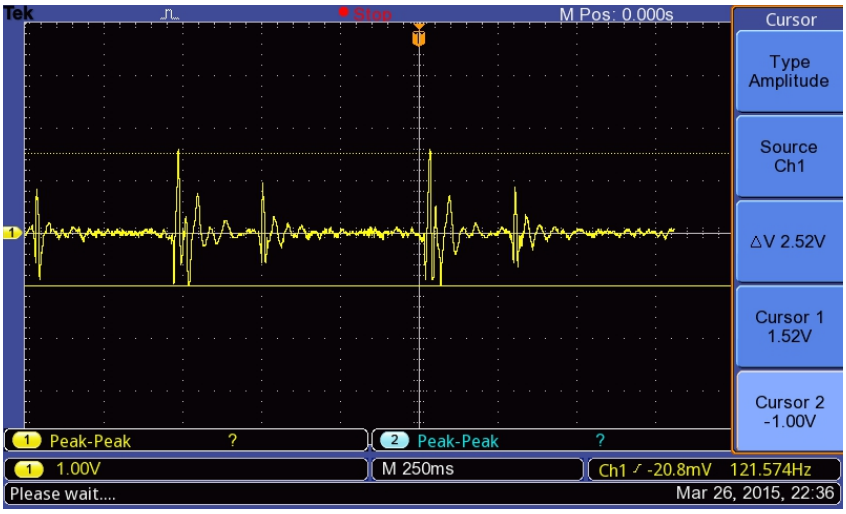

Figure 22: 2.52 V peak-to-peak with one audio channel

- When both audio channels were connected, the attenuation worsened, and the LED simply remained switched off throughout –

Figure 23: 1.88 V peak-to-peak with both audio channels connected

The reason behind this lies in the behaviour of potential dividers. The headphones have an impedance of ≈100 Ω. The op-amp parallel to this stage also has its own high input impedance.

As the impedance across the headphone stage increases with the addition of each audio channel, the voltage drop across the headphones increase so more of the heart sound is diverted to the headphone stage as more work is done by the headphones, producing sound.

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On the other hand, the voltage drop across the digitalizing stage decreases, so the signal the Schmitt trigger receives hardly ever crosses the threshold voltage. Hence, it is always positively saturated, meaning the LED is always off.

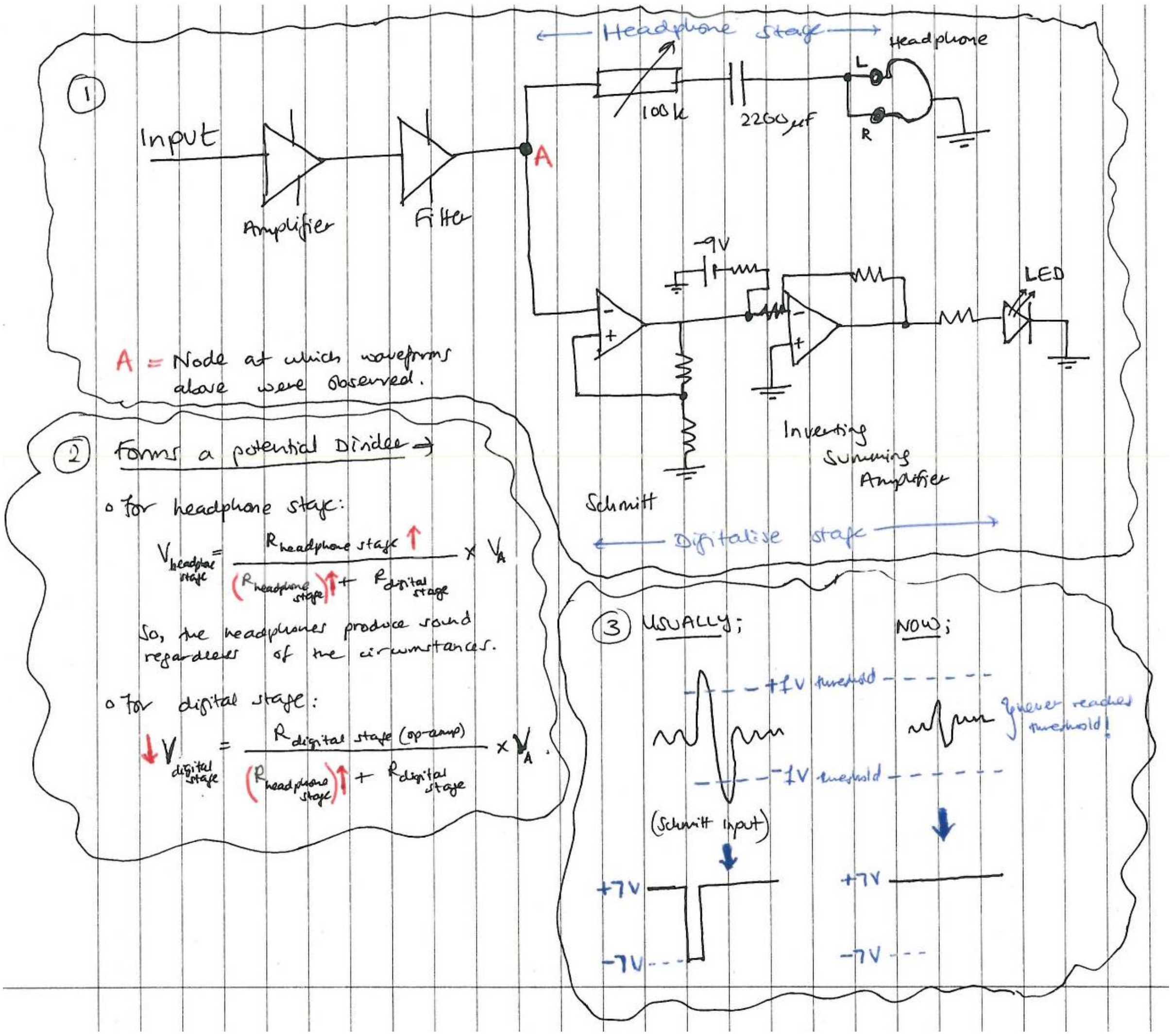


Figure 24: Despite a clear signal, the headphone stage affects the parallel digitalising stage

This demonstrates the need for an op-amp at the start of the headphone driving stage, to ensure that one parallel 'wing' does not affect the other. This is a real-life implication because, if not, the sensitivity of the Schmitt trigger will have to always be recalibrated when different pairs of headphones are used, as they may have varying impedances.

Appendix 10

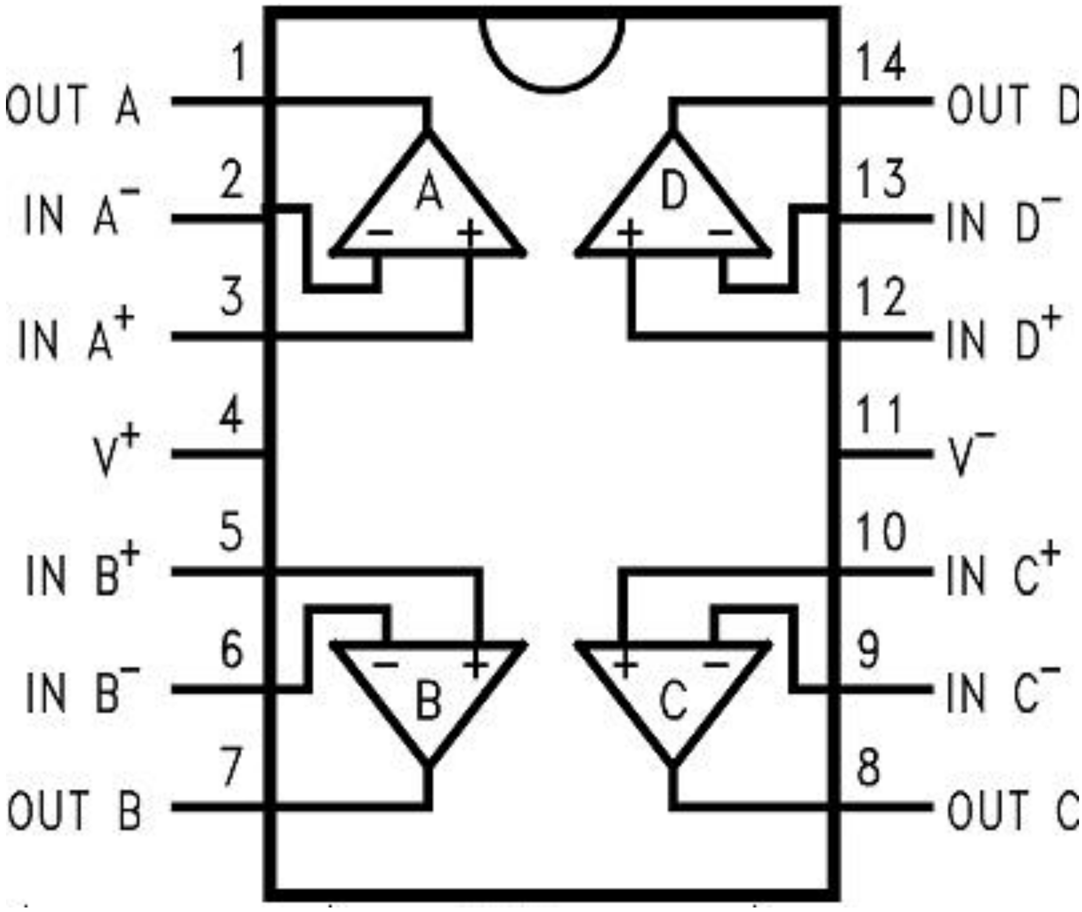
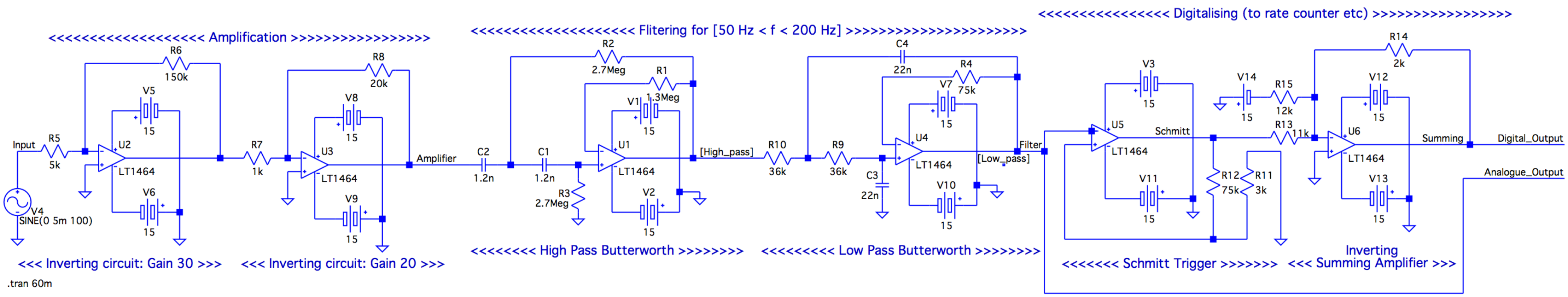


Figure 25: LM324 pin configuration (Texas Instruments, 1975)

Appendix 11



<<<<<<<<<<<<<<<<<<<< Amplification >>>>>>>>>>>>>>>>>>
<<<<<<<<<<<<<<<<<<<<< Flitering for [50 Hz < f < 200 Hz] >>>>>>>>>>>>>>>>>>>>>>>
<<<<<<<<<<<<<<<< Digitalising (to rate counter etc) >>>>>>>>>>>>>>>>>
Input
R5
5k
V4
SINE(0 5m 100)
U2
LT1464
R6
150k
V5
15
V6
15
R7
1k
U3
LT1464
R8
20k
V8
15
V9
15
Amplifier
C2
1.2n
C1
1.2n
R3
2.7Meg
R2
2.7Meg
R1
1.3Meg
V1
15
U1
LT1464
V2
15
[High_pass]
R10
36k
R9
36k
C3
22n
C4
22n
R4
75k
V7
15
U4
LT1464
V10
15
[Low_pass]
Filter
U5
LT1464
V3
15
V11
15
Schmitt
R12
75k
R11
3k
V14
15
R15
12k
R13
11k
U6
LT1464
V12
15
V13
15
R14
2k
Summing
Digital_Output
Analogue_Output
<<< Inverting circuit: Gain 30 >>>
<<< Inverting circuit: Gain 20 >>>
<<<<<<<< High Pass Butterworth >>>>>>>>
<<<<<<<<< Low Pass Butterworth >>>>>>>>
<<<<<<< Schmitt Trigger >>>>>>>
Inverting
<<< Summing Amplifier >>>
.tran 60m