A Harmonic Distortion Measurement System



A Third Year Design Project
For The University of Sheffield

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Abstract

Total harmonic distortion is used as a measure of fidelity in low frequency (audio) amplifiers. This report presents a system which can be used to find the total harmonic distortion of an amplifier using active (operational amplifier based) filters and a spectrum analyser (as an output device). The system is specified, described, tested (physically and by simulation) and evaluated, within the boundaries of the eighteen week project length. The system can present 4, 8 and 16Ω loads to the amplifier being tested. Tests can be carried out at up to 50 watts load dissipation. Any given harmonic can be measured with amplitude as low as -95dBV. The system has the capacity to be used to find its own linear distortion. This is achieved by introducing calculable nonlinear distortion, then comparing the result found from the system with that which is expected. This self test is an integral part of the system.

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Chapter 1 Necessity of Distortion Analysis

The total harmonic distortion (THD) of an amplifier is, in low frequency applications, used as a measurement of its fidelity.

1.1 Quantifying Harmonic Distortion

The harmonic distortion of an amplifier is found by applying a single frequency with constant amplitude to the amplifier under test. The output of the amplifier will contain the amplified signal and multiples of the input frequency. These additional frequencies are the distortion residual. They are a direct result of the shortcomings of the amplifier, assuming the source is perfect. The total harmonic distortion is defined [1]

$$THD = \sqrt{\frac{V_{2f}{}^2 + V_{3f}{}^2 + V_{4f}{}^2 + \ldots + V_{nf}{}^2}{V_f{}^2 + V_{2f}{}^2 + V_{3f}{}^2 + V_{4f}{}^2 + \ldots + V_{nf}{}^2}} \qquad \text{- [1.1]}$$

$n \to \infty$

V_f is the Voltage at the fundamental frequency. V_{2f} is the Voltage at twice the fundamental frequency. To have useful meaning and to ensure repeatability of experiment the THD is quoted as a percentage at an output power e.g. 0.01% at 20 watts. This implies a known load will be used. The input frequency is usually 1 kHz. The fundamental frequency is attenuated after it has passed through the amplifier under test. This is to avoid overloading the input mixer of the analyser. It is necessary to find the distortion that the system will introduce. The distortion residual must be displayed, in this project a spectrum analyser is used.

1.2 System Diagram

The statements in section 1.1 can be represented in a block diagram.

A full system diagram is located in Appendix 1

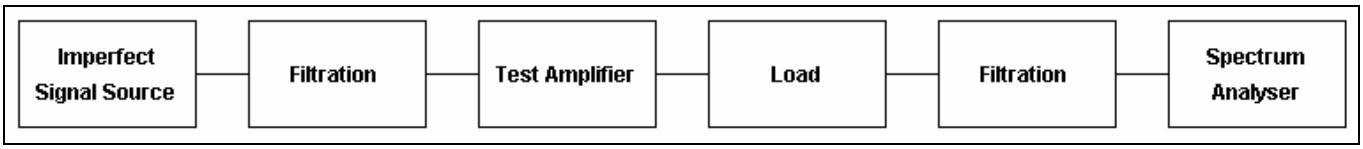


Figure 1: Partial Block Diagram formed from requirements in 1.1

1.3 Specification

Primary Objectives

- Facilitate the measurement of distortion in a test device to a minimum of -95dB for any given harmonic.
- Accept or generate a sine function of frequency 1 kHz having distortion not more than -50dB in the second and all higher harmonics.
- Reduce distortion in the input sine wave by filtration to -100dB or better.
- Apply the filtered sine wave to a DUT.
- Load the DUT resistively to simulate normal operation (16 / 8 / 4) Ω
- Accept and attenuate a signal from the DUT such that its magnitude becomes around a volt.
- Attenuate the fundamental (1 kHz) by filtration.
- Present the output to a spectrum analyser
- System noise should present no barrier to precise measurement to a minimum of -95dB
- Implement the above system entirely in one or more enclosures in the style of current test equipment

Secondary Objectives

- Produce dummy DUTs, which generate calculable distortion at their output. The reading from the spectrum analyser can be used to calculate the distortion analyser's performance
- Incorporate a signal generator of distortion not more than -80dB so that the ability of the equipment is not limited by the quality of an external generator.
- Produce the above system for use at 100Hz, 300Hz, 1kHz and 3KHz
- Incorporate suitable overload protection wherever necessary.

Chapter 2 Signal Sources

Distortion analysis requires a single frequency. A standard laboratory signal generator is not suitable; its own distortion is too high. A suitable signal source must be implemented or available signal sources must be made suitable.

2.1 Analogue vs. Digital

Analogue signal generators that produce sinusoids operate by applying regenerative feedback at the input, from the output, of a second order system, so that oscillation begins. The impedance of the feedback pathway is dynamically controlled such that the oscillator does not saturate [2].

Using this method signals can be generated with THD around -80dB. It is difficult to maintain the linearity that can be achieved with these circuits over a wide range of frequencies. Signal generators in laboratories are often required to have frequency ranges of six decades or more.

A typical figure for distortion of an analogue oscillator that has a wide frequency range is -50dB for any given frequency. (See section 2.3)

Digital Signal Synthesisers use a process of coordinate rotation to manufacture frequencies. [3] The way in which the signal is generated leads to no noticeable pattern in the harmonic content. It is plausible that the third harmonic could be 20dB greater in magnitude than the second.

2.2 The Perfect Source

To generate valid data the source must have known distortion, which can be accounted for by the operator, or have distortion low enough that the result is not significantly affected by it.

To meet the primary specification it is permissible that no source is included in the system, the user must provide one. It is not possible to estimate with the certainty that is required (for calculation of distortion) the shortcomings of an unknown source. Therefore assumptions must be made about the source. A system that makes the nonlinearities in the assumed source negligible must be implemented.

2.3 Signal Generators Currently on the market

A sample of generators supplied by Farnell In-One and Radio Spares was taken to establish the likely distortion present in an average signal source.

Cost	Company	Model	Distortion (%)	Max Frequency
£ 162.00	Thandar	TG200	< 0.5	20K
£ 254.00		TG230	< 0.5	20K
£ 203.00		TG315	< 0.5	30K
£ 350.00		TG500	< 0.5	50K
£1,117.00		TGA1241	< 0.1	100K
£ 695.00		TG1010A	< 0.3	500K
£ 98.00	GW Instek	GFG-8015G	< 1	200K
£ 187.00		GFG-8219A	< 1	200K
£ 165.00	Digimess	FG200	< 2	
£ 549.00	Tecstar	FGA2030	< +/- 2	100K

Table 1: Table of Signal Generators

The ‘Max Frequency’ column shows the highest frequency to which the distortion is guaranteed to be less than the amount specified in the adjacent column (‘Distortion %’). From this small sample two conclusions can be drawn. Firstly price and distortion are loosely linked. Secondly price is the dominant factor in deciding the bandwidth that the distortion figure is guaranteed over.

Half a percent is an adequate figure to work with. This is -46dB. Given the specification requires measurement at -95dB the source must have around 50dB of distortion removed from it to make the error *comparable* with the minimum amplitude to be measured. Methods of achieving this are considered in Chapter 4.

Chapter 3 Operational Amplifiers

Filters must be implemented to remove distortion in the source. And attenuate the fundamental from the output of the amplifier. The obvious way of implementing these filters is using operational amplifiers (op-amps).

3.1 Considerations

3.1.1 Non Linear Distortion

The distortion introduced by the op-amp should be negligible. A suitable op-amp will have distortion + noise around -120dB. This is an arbitrary value that is sufficiently low with respect to the minimum magnitude of a measured signal (-95dB). A THD for the system of -120dB would produce a 5% uncertainty in the voltage measured (-95dB). Therefore a measured signal of -95dB must have a value in the range -94.57dB to -95.44dB. It is appropriate that the error, due to non linear distortion, becomes comparable to the uncertainty in the measurement equipment for the lowest expected input signal magnitude.

3.1.2 Noise

The ‘noise floor’ must be substantially below the lowest signal value of interest. Johnson noise must be considered within the tuned bandwidth of the receiver in the spectrum analyser. Within the bandwidth of the receiver (probably between 10 Hz and 100 Hz) the noise increases the uncertainty of the result. Fortunately noise and bandwidth are related [4]; low bandwidth can help reduce noise.

3.2.1.1 Interference

Interference is the term that covers signals that appear in circuits due to external sources. It is also a term that is used to describe “hum”. Hum is a term associated with audio frequencies; it can be generated when two pieces of equipment are connected such that there is a small resistance in the ground lead. A small voltage may be developed across the ground resistance between the two units. A current would then flow in the “ground loop”. This current can introduce signals in wires that operate on signal voltages. Circuits with low PSRR are susceptible to signals in the power rails appearing in the

signal path of the circuit. Avoiding hum is achieved by following constructional techniques these are addressed in chapter 7

3.2.1.2 Signal to Noise Ratio (SNR)

When considering Johnson noise in amplifiers it is normal to refer the noise to the input. This allows a comparison between signal and noise without considering the gain of the amplifier. The signal to noise ratio is given as [5]

$$SNR = 10\log\left(\frac{Vs^2}{Vn^2}\right) \qquad \text{- [3.1]}$$

An acceptable SNR must be defined. Considering the display on the spectrum analyser is capable of showing differences of 1dB a signal to noise ratio of 25dB (assuming the minimum input signal of -95dB) will show clearly the harmonic peak, and the noise superimposed on that peak will (by the same reasoning as in 3.1.1) be a perturbation of around 1dB or one half minor division of the amplitude axis. Any improvement on -120dB would allow more precise measurements, and measurements of harmonics with smaller amplitudes.

3.1.3 Frequency Response

The gain bandwidth product (GBP) of the device is not critical. The system will operate at one Volt peak therefore with 15 Volt supplies at the point of saturation the maximum undistorted gain is 15. The maximum frequency of interest is 20 kHz consequently the minimum feasible GBP is 300000. This is easily achievable.

The maximum slew rate is a measure of the greatest $\frac{dv}{dt}$ at the inputs that the op-amp can accurately reproduce at its output. At 20 kHz a ¼ period (the rise from 0 to the peak of a sinusoid) takes 25μS, the expected operating voltage is 1 Volt consequently 0.04V/μS minimum slew rate is required. This is not entirely accurate as it assumes that the waveform is triangular, however it provides an estimation. Slew rates of many V/μS are easily achieved by operational amplifiers currently in use.

3.1.4 PSRR

The PSRR (Power Supply Rejection Ratio) is a measurement of the devices ability to reject signal frequencies in its power supply. It is expected that this equipment will be operated from a regulated supply. High PSRR is not vital however it is desirable.

3.2 Devices

During the course of investigation 24 operational amplifiers specifications were consulted. A full list can be found in Appendix 4. The 'best' 10 are presented below

Op-Amp	Distortion (%)	Noise
TL071	0.003	$18nV/Hz^{1/2}$
OPA604	0.0003	$10nV/Hz^{1/2}$
AD797	0.0001	$1.2nV/Hz^{1/2}$
LM833	0.002	$4.5nV/Hz^{1/2}$
NE5532		$8nV/Hz^{1/2}$
NJM2068	0.001	$0.56uV/Hz^{1/2}$
NJM4562D		$0.6uV/Hz^{1/2}$
LF412	0.02	
OP27		$3nV/Hz^{1/2}$
OPA134	0.00008	$8nV/Hz^{1/2}$

Table 2: Table of Best Ten Operational Amplifiers

Where no data is included none was available. Manufacturers rarely specify distortion and noise in the same way. Limitations normally apply such as frequency range or maximum signal level or minimum load resistance. The list was reduced to three selected for distortion and noise, the OPA 604 AD797 and the OPA134. Their costs respectively are 1.41, 6.92 and 0.98. (GBP) (Prices are from Farnell). All of these devices meet the requirements discussed, the OPA134 was chosen.

3.2.1 Total Harmonic Distortion + Noise

The OPA134 Datasheet [6] gives THD+N for a 1 Volt output into a 2kΩ as 0.00008 % which is -121.9 dB. The most likely place for the 2kΩ load requirement to be violated is at the input of the test amplifier. Power amplifiers, are routinely designed with input resistances of tens of kilohms. No steps to avoid this possibility will be taken.

Chapter 4 Filtration Methods

4.1 Band Pass

The system diagram in chapter 1.2 shows filtration after the imperfect source, which was discussed in chapter 2. The distortion is a number of additional harmonics which must be attenuated. It is also advantageous to remove any DC offset that the imperfect source may have. Therefore a band pass filter should be used.

4.1.1 Butterworth Band Pass Design

Analogue signal generators usually do not have digital meters showing frequency. It is quite difficult to obtain accurately a desired frequency. To avoid unnecessary attenuation of the fundamental a band pass filter with a maximally flat pass band 100Hz wide centred on 1 kHz was designed.

This design was rejected due to the likelihood that sensitivity would make it impossible to achieve the flat pass band. It is not a significant improvement to have a maximally flat pass band. The operator can adjust the “level” control on the generator to obtain 0 dBV, should the filter have any loss in its pass band.

4.1.2 Cascaded Single Tuned Band Pass Filtration Systems

If two or more filter sections that have a single centre frequency ω_0 are placed end to end the result is a cascade of single tuned filters. This filter topology is suitable for the reduction in harmonics of the signal generator and is simpler than the Butterworth response in that only one set of values must be calculated. Figure 2 shows the response that is required, the brick wall [7] response is outlined and a likely response is sketched.

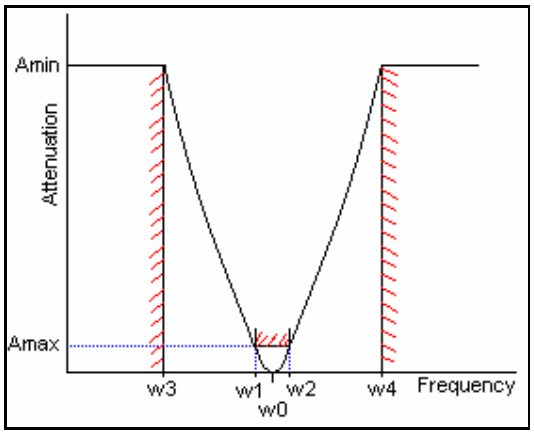


Figure 2: Brick Wall Response of Band Pass Filter

ω0 is the centre frequency, ω1 and ω2 define the edges of the pass band and ω3 and ω4 define the beginning of the stop band. A_{max} is the maximum permissible attenuation in the pass band, A_{min} is the minimum permissible attenuation in the stop band. *In the following discussion upper case* BW *represents the -3db bandwidth and lower case* bw *represents bandwidth at an attenuation given in the surrounding text.* It is necessary to attenuate frequencies above 1900 Hz and pass frequencies between 950Hz and 1050Hz. Assuming the input frequency is as low as is likely (approximately 950Hz). The second harmonic will be 1900Hz. Therefore ω_4 should be 11938.1 $rads^{-1}$. Knowledge of ω_2 and ω_4 allows the other frequencies to be calculated [8]. The exact values do not matter, provided at DC the attenuation is very large and the lower -3dB point is close to 950Hz. The sensitivity of the filter section will cause greater errors, in response, than the error in calculation.

$$\omega_0^{\,2} = \omega_1\omega_2 = \omega_3\omega_4 \qquad \text{- [4.1]}$$

$$\omega_0^{\,2} = 39478417.6$$

$$\omega_1 = \frac{39478417.6}{6579.3} = 6000.4 rads^{-1} = 954.9 Hz$$

$$\omega_3 = \frac{39478417.6}{11938.1} = 3307 rads^{-1} = 526.3 Hz$$

From the specification it is necessary to have the minimum attenuation in the stop band, α_{min}, equal to -50dB. The maximum pass band attenuation, α_{max}, should be insignificant with respect to the minimum graduation on the spectrum analyser. -0.1dB was chosen.

It is necessary to transform the requirements for the band pass filter into the equivalent requirements for a low pass filter, by using a frequency transformation [9].

$$\Omega_p = 1$$

$$\Omega_s = \frac{\omega_4 - \omega_3}{\omega_2 - \omega_1} = \frac{11938.1 - 3307}{6597.3 - 6000.4} = \frac{8631.1}{596.9} = 14.45 \quad \text{- [4.2]}$$

Having found the equivalent low pass specifications the number of stages can be found [10]

$$n = \frac{\log\left[\left(10^{\alpha_{\min}} - 1\right)/\left(10^{\alpha_{\max}} - 1\right)\right]}{2\log\left[\Omega_s/\Omega_p\right]} \quad \text{- [4.3]}$$

$$n = \frac{\log[99999/0.02329]}{2\log[14.45]}$$

$$n = 2.86 \therefore n = 3$$

When filter stages are stacked the properties that represent Q and Bandwidth (BW) in a second order system are affected the bandwidth is reduced (or the Q is increased) according to [11]

$$BW = \frac{\omega_0}{Q}\sqrt{2^{1/n} - 1} \quad \text{- [4.4]}$$

BW is the new bandwidth given a centre frequency of $\omega_{0.}$ n is the number of stages. Q is the quality factor. The term under the root is the bandwidth reduction factor. For a three stage cascade it is approximately 0.5.

If the bandwidth of the system is halved due to the cascading of stages, then the Q of all the stages must be halved to compensate for this. The ratio (Eq^n 4.5) must be unchanged.

$$BW = \frac{\omega_0}{Q} \quad \text{- [4.5]}$$

4.1.3 Band Pass Filtration Methods

The Friend band pass circuit [12] is a second order filter implemented using one op-amp. The Friend circuits' advantage over other possibilities such as a bi quad is it operates by the cancellation of signals through their phase shifts as apposed to magnitudes. This allows circuits of high Q to be used when in other circuit topologies the supply voltages would not allow such operation.

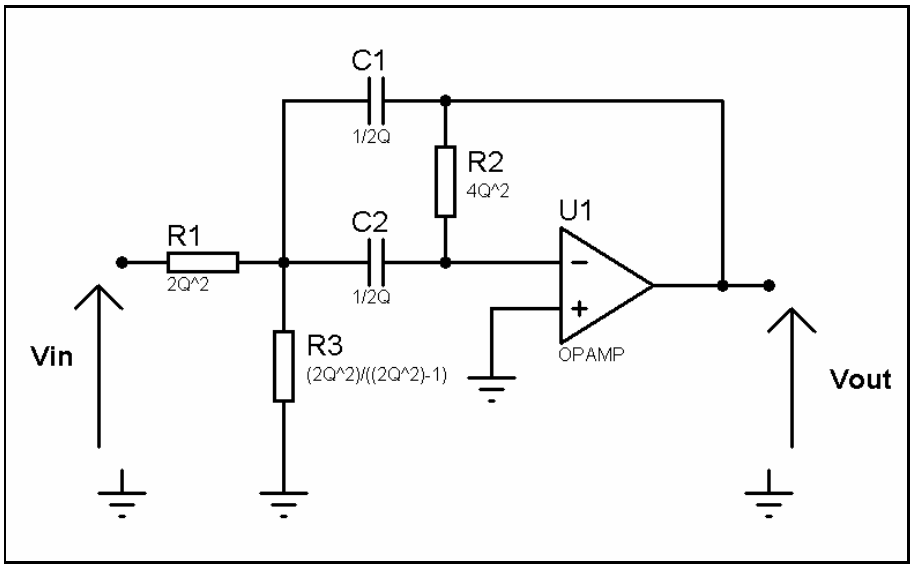


Figure 3: Single Stage Friend Band Pass Circuit

The values of components in Figure 3 are related to a defined value of Q and to a centre frequency of 1 rads^{-1}. Frequency and magnitude scaling are employed [13]. For a centre frequency of 1 rads^{-1} [14]:

$$C_1 = C2 = \frac{1}{2Q} \quad \text{- [4.6]}$$

$$R2 = 4Q^2 \quad \text{- [4.7]}$$

$$R1 = 2Q^2 \quad \text{- [4.8]}$$

$$R3 = \frac{2Q^2}{(2Q^2 - 1)} \quad \text{- [4.9]}$$

It is necessary to choose one component value arbitrarily, and then find the other values. If they are realistic, a solution has been found. Van Valkenburg suggests choosing the capacitance. After several iterations 4.7nF was chosen.

Component	R1	R2	R3	C1,2
Equation ω=1 Q=5	$2(5)^2$	$2(5)^2/(2(5)^2-1)$	$4(5)^2$	½(5)
Value ω=1Q=5	50Ω	1.0204Ω	100Ω	0.1F
Scaled values	16.93kΩ	345.5Ω	33.8kΩ	4.7nF

Frequency and magnitude scaling [15] are then employed:

$$k_f = 2\pi \times 1000 = 6283.1 \quad \text{- [4.10]}$$

$$k_m = \frac{1}{k_f \times C_{new}} \times C_{old} = \frac{1}{6283.1 \times 4.7 \times 10^{-9}} \times 0.1 = 3386.3 \quad \text{- [4.11]}$$

$$R_{(1,2,3)new} = R_{(1,2,3)old} \times k_m \quad \text{- [4.12]}$$

Having chosen C1 & C2 k_f is calculated. Then k_m can be found. It is then possible to determine values of the resistors.

The Schematic of the complete cascade can be found in Appendix 3.1

A Spice frequency response using the model from Burr Brown [16] can be found in Appendix 2.2.1

Data collected on a Network Analyser is given for the pass region in A2.1.1 and graphed in A2.2.5

A plot taken from the network analyser is given, for the region 0 Hz to 20 kHz, in A2.2.13

4.1.3.1 Sensitivity to Component Variation

Spice can be used to find the change in frequency response due to component variation. The component tolerances used are 1% for capacitors and resistors. The slight offset of the middle trace is due to the limitations of real component values. The right trace represents a 1% increase in *all* component values. The left a 1% decrease.

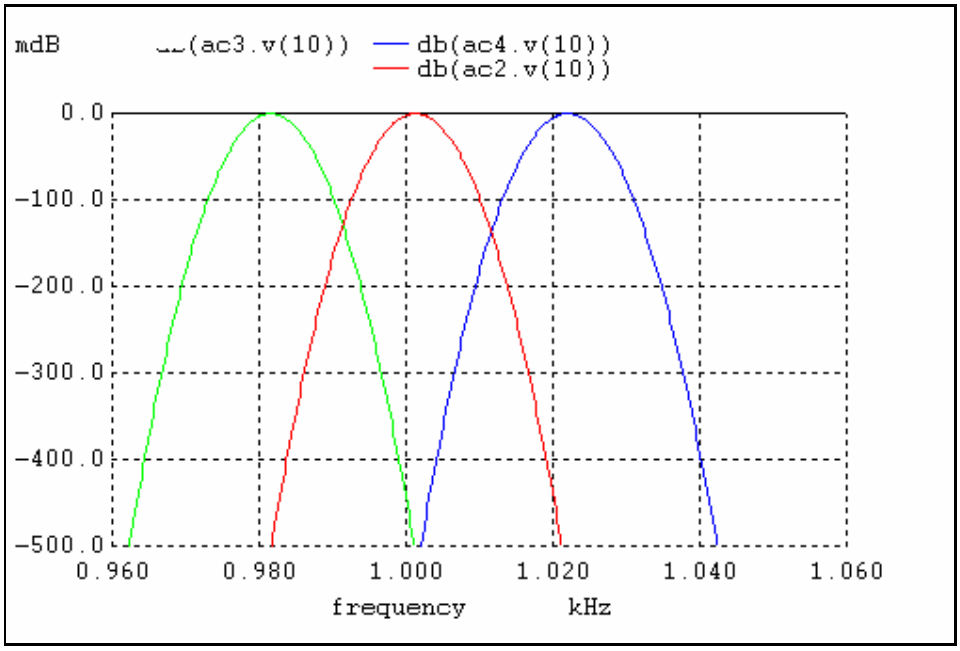


Figure 4: Worst-case Frequency Responses Due to Component Tolerances

4.1.4 Conclusion

The Spice data suggests that the filter will perform adequately, the pass and stop bands are correctly positioned and have the correct attenuations. The maximum attenuation in the pass band due to component variation is around 0.5dB. This can be corrected by adjusting the 'level' (Output Voltage) on the signal generator.

4.2 Band Elimination

It is necessary to attenuate the fundamental (1 kHz) this is due to the way analogue spectrum analysers operate. There is normally a stepped attenuator followed by a mixer [17] & [18]. If the mixer is overloaded harmonics will be generated in quantities that cause significant error in results.

4.2.1 Butterworth Band Elimination Design

Several Butterworth designs were considered none were taken forward because the sensitivity of the friend notch circuit is such that results would be unpredictable – the worst case was unacceptable. Six stages would be required to obtain an attenuation of 25dB over a 100 Hz bandwidth with a centre frequency of 1 kHz. A suitable attenuation is more than 50dB. Hence twelve stages would be required. If such a filter were constructed, it is unlikely to work as required. However the pass and stop bands (assuming perfect components) would be well defined, it is unlikely that such a design would require the addition of corrective data to the measurements as is the case with the cascade that was eventually used.

4.2.2 Cascaded Single Tuned Band Elimination Filtration Systems

Band Elimination design follows a different route, it is not possible to choose Q and the bandwidth at A_{max} (see Figure 7). It is not possible to 'orthogonally tune' the stage.

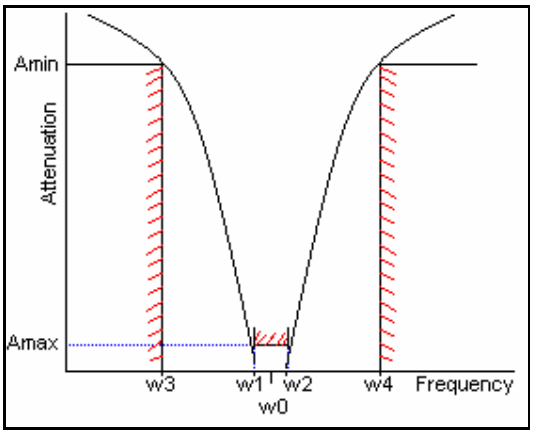


Figure 5: Band Elimination Brick Wall Response

Considering a single stage and arbitrarily selecting a centre frequency of 1 kHz with a bandwidth (bw) at 45db attenuation of 5Hz (5Hz is chosen to allow for drift in analogue generators) there is enough information to find BW (-3dB bandwidth) and the Q then a circuit realisation can be generated.

Van Valkenburg shows [19] that for a second order band elimination response

$$BW = bw\sqrt{10^{\alpha_{min}/10} - 1} \qquad \text{- [4.13]}$$

Where BW is the 3dB bandwidth, and bw is the bandwidth at the required minimum attenuation in the stop band, α_{min}.

$$BW = 5\sqrt{10^{45/10} - 1} \qquad \text{- [4.14]}$$

$$BW = 889.1Hz$$

$$Q = \frac{f_0}{BW} = 1.1247$$

If two such filters are cascaded it is intuitive to assume that the 'depth' of the notch will be doubled. On paper that is the effect. However the band elimination filter operates on the cancellation of phase and for perfect cancellation the values of capacitance in the circuit must be equal. This is unlikely to be the case with real components. As before there is a change in bandwidth due to the cascading of two stages. Using equation 4.4 the cascade of two should lead to a change in 'bandwidth reduction factor' of 0.64. Spice simulation doesn't agree with this calculation, possibly due to the frequency independent attenuation introduced as a result of the models [20] ability to represent the imperfections in the circuit. Therefore Matlab® has been used to model the notch as the product of transfer functions of the individual filter sections. (See Figure. 6). Measurements taken from 'zooming' on parts of this graph give

Number of Sections	-3dB lower	-3dB upper	Bandwidth(3dB)
1	525	1908	1383
2	652	1538	886

Table 3: -3dB Data from a single and a cascade of two band elimination filter sections

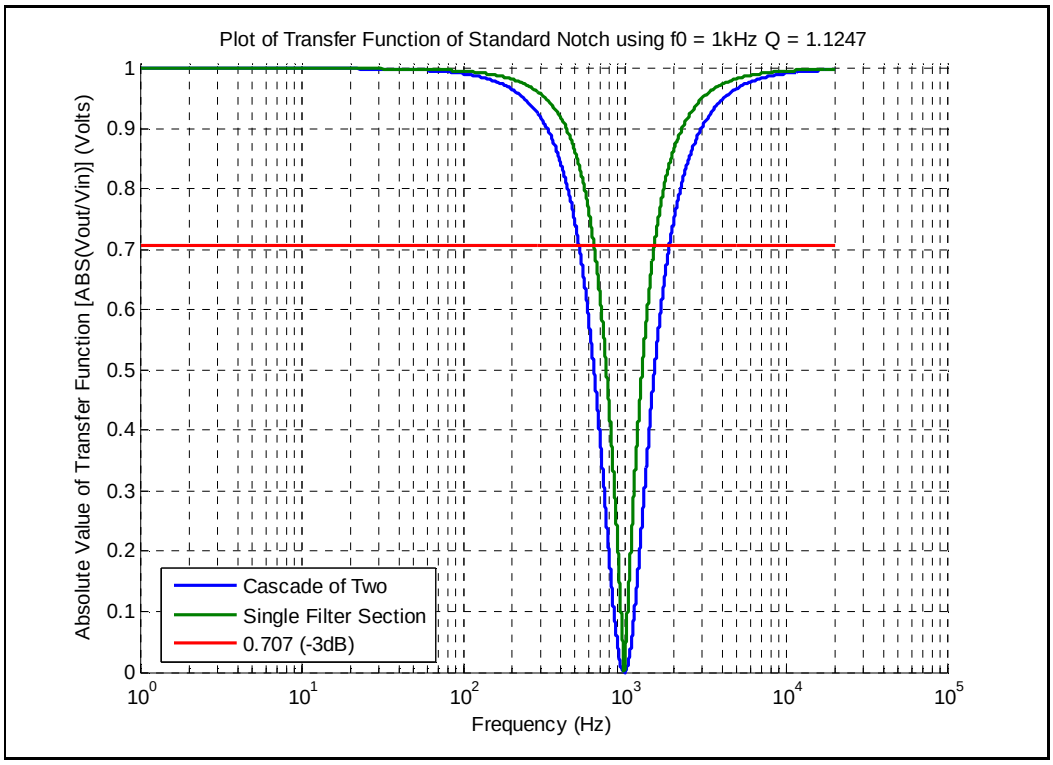


Figure 6: Transfer Function Model of Standard Notch

The data in Table 3 agrees with equation 4.4, suggesting that for the purposes of calculation the -3dB point should be taken at 0.707 of the full output voltage, as opposed to 0.707 of the lowest point in the stop band (which is more logical, given that in a band pass circuit of similar specification Q would be measured at 0.707 of the output voltage in the pass band). Resolution to this has not been sought as it has no effect on the outcome of the implementation of the solution presented.

4.2.3 Band Elimination Filtration Methods

The Friend notch circuit, based on the same bridged T RC network as the band pass, is chosen for the band elimination filter section. This filter section does not benefit from the reduced sensitivity of the band pass version [21]. 0.1% tolerance resistors and 1% tolerance capacitors are used to minimise the variation in response.

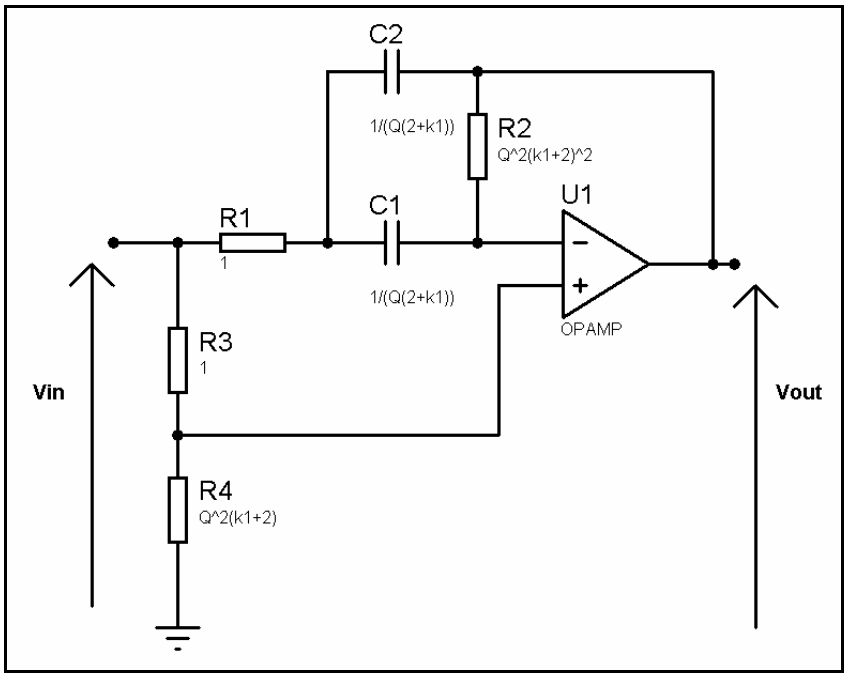


Figure 7: Single Stage Band Elimination Filter Section

The values of components in Figure 7 are related to a defined value of Q and to a centre frequency of 1 rads^{-1}. Frequency and magnitude scaling are employed. For a centre frequency of 1 rads^{-1}[22]:

$$C1 = C2 = \frac{1}{Q(2+k1)} \quad \text{- [4.15]}$$

$$R2 = Q^2(k1+2)^2 \quad \text{- [4.16]}$$

$$R4 = Q^2(k1+2) \quad \text{- [4.17]}$$

$$R1 = R3 = 1 \quad \text{- [4.18]}$$

$$k1 = \frac{\omega_0^{\,2}}{\omega_z^{\,2}} - 1 \quad \text{- [4.19]}$$

k1 is a variable used to define high pass notches. The pass band, that occurs lower in frequency than the stop band, has a higher attenuation than the pass band that occurs higher in frequency than the stop band. Using other circuit topologies it is possible to implement a low pass notch. For a standard notch $\omega_z = \omega_0$ therefore k1 = 0.Design proceeds as in the band pass case in section 4.1.3. Choosing a capacitance of 4.7nF:

Component	R1 & R3	R2	R4	C(1,2)
Equation	1	$Q^2(k1+2)^2$	$Q^2(k1+2)$	1/(Q(2+k1))
Value ω=1 Q=1.1247	1	5.0598	2.5299	0.4446
Scaled Values	15,050Ω	76,170Ω	38,085Ω	4.7nF

Frequency and magnitude scaling are used as in the band pass case.

$$k_f = 2\pi \times 1000 = 6283.1$$

$$k_m = \frac{1}{k_f \times C_{new}} \times C_{old} = \frac{1}{6283.1 \times 4.7 \times 10^{-9}} \times 0.4446 \approx 15050$$

$$R_{(1,2,3,4)new} = R_{(1,2,3,4)old} \times k_m$$

The schematic of the complete cascade can be found in Appendix 3.2

Data for the cascade can be found in Appendix 2.1.2 and a graph of this data is in A2.2.6

A simulated frequency response using the Burr Brown model {16] is given in A2.2.2

A plot taken from the network analyser for the range 0 Hz to 20 KHz is given in A2.2.14

4.2.3.1 Sensitivity to Component Variation

Again Spice is used to find the worst case response given 0.1% resistors and 1% capacitors. The minimum attenuation for both worst cases is where the upper and lower frequency traces intersect with the 1 kHz line. -70dB is, from simulation, the least attenuation achieved.

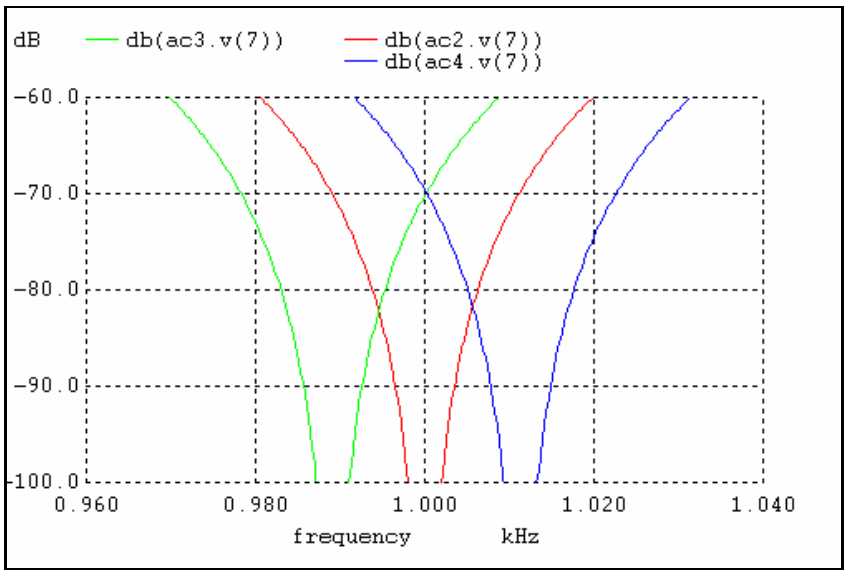


Figure 8: Spice Data for Worst Case Component Variation.

4.2.4 Conclusion

The Spice data suggests that the filter will perform adequately. The pass and stop bands are correctly positioned and have the correct general shape. The Q of the filter cascade forces the post processing of data gathered using the system. This would be unavoidable using a standard notch even if the pass band was flat from 1.9 kHz to 20 kHz due to the frequency independent attenuation. Ian Hickman suggests an interesting solution [23]

Chapter 5 Loading the Device under Test

Harmonic Distortion is quoted as a percentage at a power rating (e.g. 0.001% at 50 Watts). The device under test must be loaded, a resistive load will suffice. The system should provide common loudspeaker impedances. The maximum power that the load should dissipate must be set. The size of a heat sink is proportional to the power it can dissipate for a fixed temperature rise. 50 Watts is acceptable, and will require a moderate heat sink.

5.1 Ensuring fair test conditions

The large current in the load make it necessary to separate the output of the DUT from the signal level circuitry. It is possible that large currents (up to several Amps) flowing in wires positioned close to signal carrying wires would compromise the validity of the data due to coupling. Two cases are used, one to house the signal sub circuits, the other to house the power resistors and heat sink. A variable resistor must also be housed in this case to reduce the output of the DUT to a signal voltage. The attenuation required is dependant on the power that the test is conducted at and the load used. It is preferable to avoid connecting the ground used for output of the DUT to the ground used by the signal stages (a schematic showing interconnections between sub circuits is in Appendix 3.3). Consider the ground as a perfect wire with a parasitic resistance in series. If large currents flow from the load into ground they will develop a voltage across the parasitic resistance. The sinusoidal ripple voltage across the ground wire will be presented to the ground connections used in the filter sections, which is undesirable. Consider a current of 3.15A and a wire resistance of 0.02Ω/meter for 1 meter with an op-amp PSRR of 90dB. This leads to -114dB signal injection, not far below the range of measurements. Interference from sources in the lab (e.g. fluorescent lights) can be reduced by connecting the case to ground.

5.2 Loading, Power Dissipation & Thermal Considerations

Common load impedances currently in use are (4, 8 & 16) Ω. These are implemented using 4.7Ω 6.8Ω and 15Ω 50 Watt maximum dissipation resistors. Referring to the Vishay datasheets a 1.2℃/Watt heat sink was chosen, and accordingly fitted to the resistors and the case.

The calculations associated with this decision are located in Appendix 7

Chapter 6 Measuring Frequency

A frequency counter would be useful in synchronising analogue generators with the notch filter. Two methods were considered. Neither was implemented due to time constraints.

6.1 Counting Methods

It is possible to amplify the signal to saturation, making it square. It can then be used in a counting system based in hard wired logic. A moving average counter, with the restriction is that the length of the average is one second, is shown schematically in Appendix 3.6. Referring to A3.6 the first set of vertically aligned blocks are synchronous decade counters with reset, the second set are latches connected to the output of the counters. The third set are display drivers for the seven segment displays. The RC networks add appropriate delays. The timing of the signals must ensure that data is not cleared before it is latched. This system is unlikely to work without considerable 'tweaking', capacitors are not perfect. The square wave generator in the top left is a one second clock. It is assumed perfect.

The signal of interest is first saturated; it is then fed into the counters which are cascaded. After an effective period of (1-δ)* seconds the output of the count is latched. The latched data is presented to the seven segment drivers and displayed then the counters are reset. Each time new data is latched the display is updated. Experiments are required.

* (where δ is small W.R.T the expected period of the measured signal (around 1ms))

6.2 Sampling Methods

Another possible solution is to use a micro processor to sample the input wave form. The processor can then perform any required calculations to produce a result. The processor could drive an LCD which may be a more suitable output device, given the available space.

Chapter 7 Construction

7.1 Cases

The cases used are cast aluminium. Aluminium is easier to work than steel, and more durable than plastics such as ABS. Some modifications were made to the case containing the signal circuitry to accommodate the number of controls that were required on the front panel. With hindsight a larger single case, with an electrostatic screen inside to reduce interference, may be more ergonomic (it would be easier to carry around). It would also allow the extra room required for subsidiary systems such as an internal signal generator and a frequency counter similar to those previously described.

7.2 Circuit Board Layout

The low frequencies used in this project do not necessitate careful attention to layout of components; the boards were not optimised for area, but instead for simplicity of understanding the layout. This greatly aids fault finding, wire links were favoured over a second copper layer.

7.3 Constructional techniques

In amplifier construction a commonly used grounding arrangement is called ‘star’ (not to be confused with 3Φ systems) where all ground connections have a separate wire linked to a common ground point normally connected directly to the ground side of the smoothing capacitors in the power supply. This was not possible as the power supply is in a separate case; the 4mm terminal (used to connect the ground terminal on the case containing the filters to the power supply) is the closest point that is feasible. Using separate wires for every ground connection is profligate with wire and makes construction more tedious. However this arrangement is likely to have the minimum interference as no current from any subsystem would enter any other subsystems power supply rails as there will always be a path to ground of lower resistance.

The ground shared by the filter and distortion introducing circuits is separate from the ground shared by the load resistors and the amplifier (see Appendix

3.3) the potentiometer that attenuates the output of the amplifier shares ground with the filter stages. This method of grounding ensures that there is no return path into the filter stage ground from the input of the DUT. Therefore it is not possible for large currents to flow in the filter stage ground.

Chapter 8 Characterisation

8.1 Introducing Calculable Distortion

In order to quantify the linear distortion due to the notch filter a known distortion can be used in place of the DUT. The calculated non linear distortion and the measured non linear distortion should differ by the linear distortion of the system, assuming there is no measurable non linear distortion due to the system (see chapter 3.1.1).

Another possible method of finding the linear distortion is to use a network analyser to produce a bode plot for the notch filter.

8.1.1 Introduced Distortion Test

The first test is split into odd harmonics and even harmonics. Two circuits were produced to generate the required harmonics. Matlab® and Spice were used to simulate the likely distortion generation. The circuits were tested in isolation using a spectrum analyser. The distortion from Spice is comparable to the distortion measured in the experiment, the error being 0.37% in the case of even harmonics and 2.09% in the case of odd harmonics. This is sufficient to conclude that the circuits operate correctly. The circuits were used (independently) in place of a test amplifier. The data which is gathered is compared with the data gathered from the circuit in isolation. The difference is the attenuation due to the notch which must be added to any results collected from tested devices.

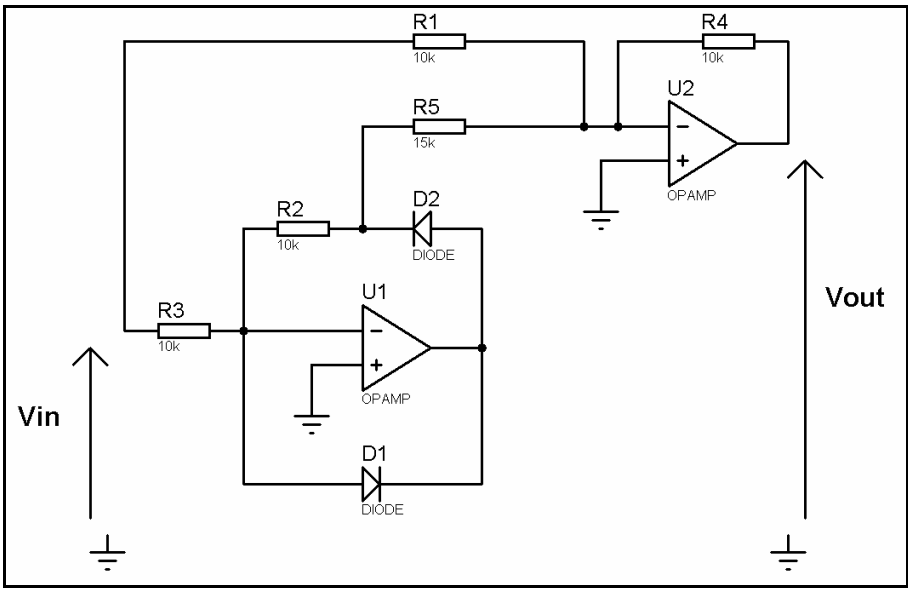


Figure 9: Even Harmonic Distortion Generating Circuit (D1 & D2 are 1N4148) [after R C Tozer, University of Sheffield]

The even harmonic generating circuit (Figure 9) is a limiting circuit followed by a summing amplifier. It generates even harmonics by reducing the amplitude of positive going half cycles compared to negative going half cycles.

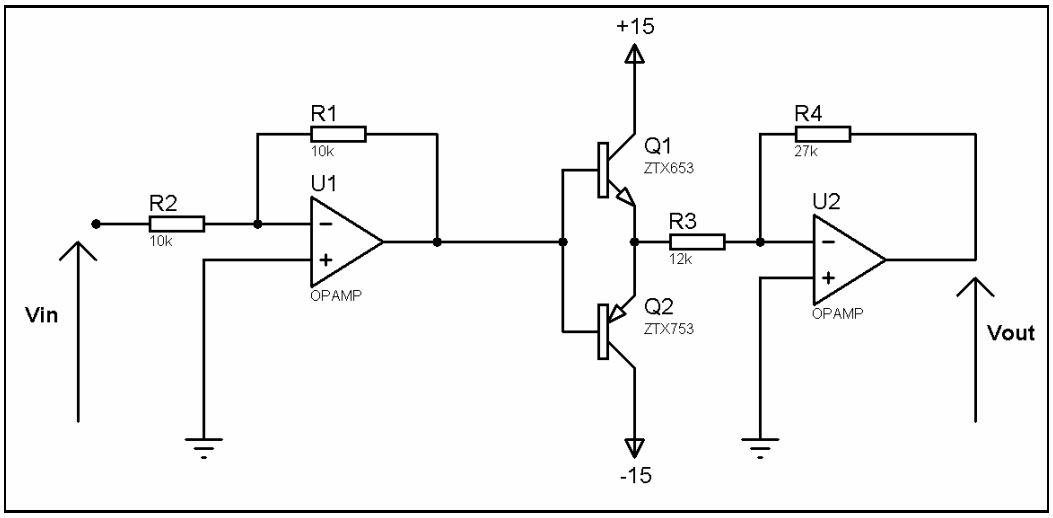


Figure 10: Odd Harmonic Distortion Generating Circuit

Figure 10 is the odd harmonic generating circuit; it generates crossover distortion due to the lack of biasing between the bases of Q1 and Q2. The op-amp that follows the stage is used to restore the peak of the distorted signal to 0 dBV.

Data associated with the distortion generation circuits and characterisation of the system is listed in the contents

Matlab® data is not included as it adds nothing to the Spice data. The Matlab® data was generated by representing each circuit as a function whose output was the piecewise linear equivalent of the expected output from the circuit.

8.1.2 Network Analyser Test

The notch filter is connected to the analyser and the transfer function is plotted against frequency. The addition of the modulus of this data to results gained from an experiment on an amplifier should lead to the correct values of harmonics. These tests (8.1.1 & 8.1.2) should produce the same results, if this is not the case some or all of the data is invalid. Several repeat experiments were conducted before valid data was collected.

8.2 Other Testing

The band pass filter was tested using the network analyser to ensure that the distortion introduced by the source was acceptable. This was conducted using the same method as the notch filter in chapter 8.1.2. The filters centre frequency was measured as 982 Hz, the attenuation at 1 kHz is 0.5dBV. The smallest graduation on the spectrum analyser that was used (HP 141T) is 2dB. Consequently the 0.5dBV error is comparable to the width of the trace on the CRT when the focus is adjusted to its optimal position. If the error becomes significant the output voltage on the signal generator should be adjusted. The –3dB points are offset by the same amount as the centre frequency. The attenuation at 1.9 kHz is -53dB. This indicates that the design functions as intended.

8.3 Result

The result of the experiments performed to characterise the linear distortion due to the system is a graph. This graph should be used to find the addition that should be made to any result given the frequency of the harmonic.

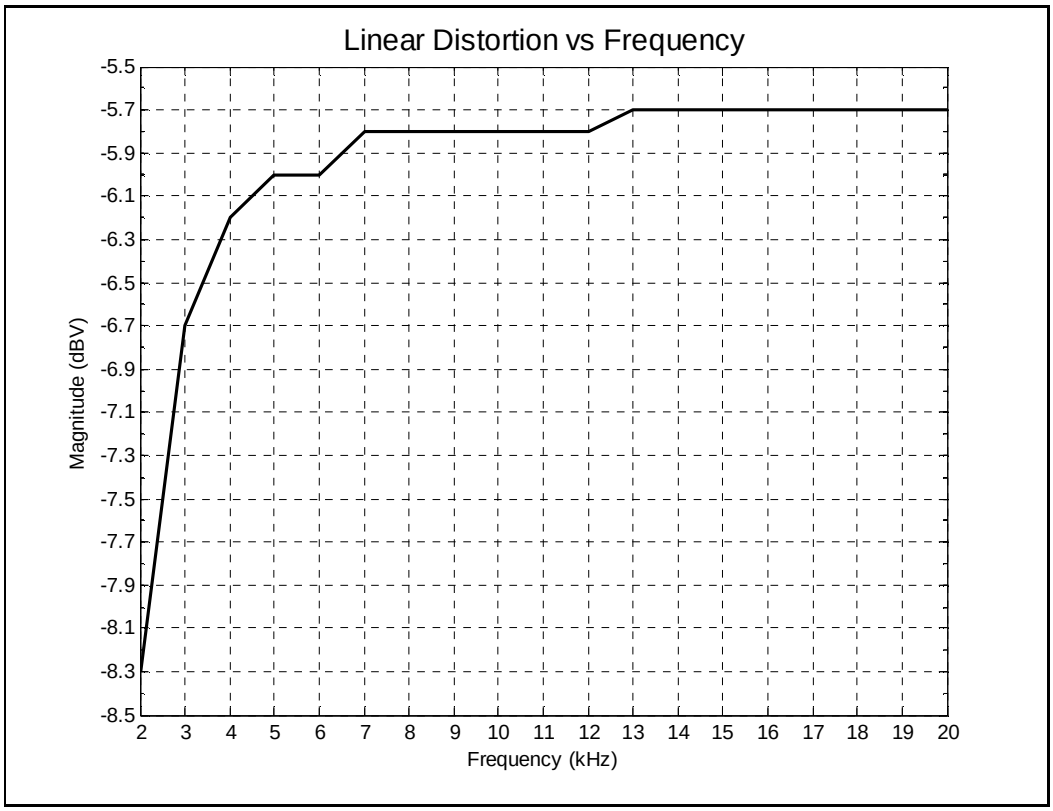


Figure 11: Linear Distortion vs. Frequency (To correct measurements from system)

Chapter 9 Example Test Procedure

This chapter can be treated as a user guide. It describes how data can be gathered, using this project, by example. The amplifier used for this test is a Sansui A-700. The test procedure can be split into five steps.

1. Calculate the voltage required to give the power dissipation in the chosen load.
2. Connect everything appropriately
3. Adjust the amplifier, attenuator (in the load case)
4. Set up the spectrum analyser
5. Conduct the experiment & record results.

9.1 Power Dissipation

The experiment will be carried out at 50 Watts with an 8Ω load.

$$P = \frac{V^2}{R}$$

$$V^2 = P \cdot R = 400$$

$$V = 20V_{rms}$$

$$V_{pk} = 20\sqrt{2} = 28.28V$$

9.2 Connect the Experiment

The experiment should be connected according to the schematic in Appendix 3.3. The essential selections are: load resistance, whether the distortion generating circuits are included and the location of the potentiometer's wiper. The potentiometer should be fully anticlockwise as should the load resistance. Distortion generating circuits should not be connected. On the front of the load box are two sets of 4 mm terminals the left most set connect to the DUT's output. The right most pair monitor the voltage presented to the notch filter's input; they only do this when 100K is selected as the load! At all other times they are disconnected. The 100K load will be selected at this point so a meter set appropriately or an oscilloscope should be connected to these terminals.

The signal generator and test amplifier should be connected as should the power supply for the filters and the spectrum analyser

9.3 Adjust the Amplifier and Attenuator

Ensure that any volume control on the test amplifier is set to minimum. Connect an oscilloscope or meter to the output of the amplifier under test. Increase its volume control until the voltage is required for the correct power dissipation.

Connect the amplifier under test to the left most set of 4mm terminals on the front of the 'load box' connect the oscilloscope or meter to the right most set of terminals. Ensure 100K is the selected load. Slowly advance the pot until the reading on the scope/meter is 1Volt peak.

9.4 Set up the Spectrum Analyser

Assuming that the frequency scale is calibrated (or doesn't require calibration) the following settings should provide useful data:

Name	Value
Bandwidth	30Hz
Scan Width	500Hz
Input level	-60dB
Log Ref. Level	-60dB
Scan Time / Div	5 Seconds

9.5 Results

Results should now be gathered by selecting each harmonic in turn and recording its value. To produce a graph that looks like a spectrum analysers display the user must also estimate the noise. The author feels that the results look satisfactory if the noise is taken to be a line drawn at the average of the difference between the highest noise spike and the lowest noise spike. In Figure 12, 200 points were placed around the 20 data points to produce the

graph. The fundamental was also removed to aid clarity, its value is -58dBV. The experiment was repeated with similar results the data for both experiments can be found in Appendix 2.2.10 & 2.2.11

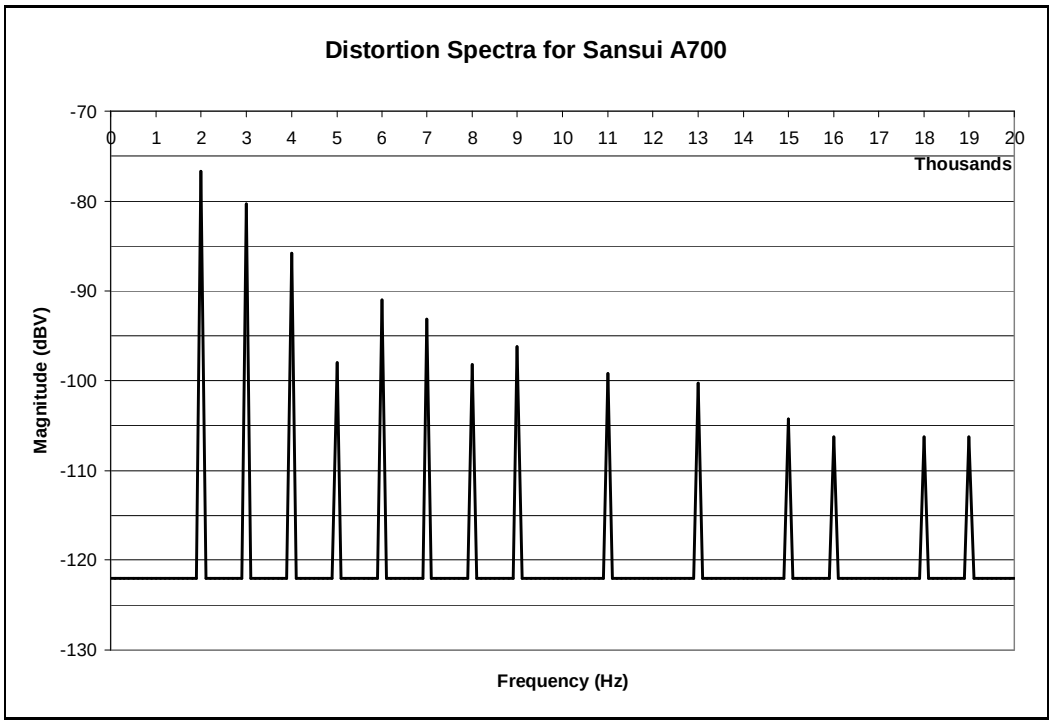


Figure 12: Results of Test 1 on Sansui A700

The total harmonic distortion can be found using equation 1.1 and is 0.0198% which is within the expected range for this amplifier.

Chapter 10 Conclusions

10.1 Conclusion

The primary specification points have been met as have some of the secondary points:

- The solution facilitates the measurement of distortion in a test device to a minimum of -95dB for any given harmonic. Although data has been produced below this value its validity is not certain and should be considered critically. This is not a limitation. In the graph in section 9.5, 98% of the distortion is due to the harmonics at 2, 3, & 4 kHz.
- The solution can accept a sine function of frequency 1 kHz having distortion not more than -50dB in the second and all higher harmonics.
- The band pass filter discussed in section 4.1.2 does reduce distortion in the input sine wave by filtration to -100dB or better.
- Application of the filtered sine wave to a DUT has been achieved.
- Resistive loads to of 4Ω 8Ω and 16Ω have been included.
- The solution can attenuate the signal from the DUT; its magnitude is set by the operator.
- Attenuation of the fundamental (1 kHz) by filtration is achieved by the notch filter discussed in Chapter 4.2.2
- Noise presented no limitation during the measurements documented in Chapter 9.5, where harmonics around -95dBV were successfully measured.
- The system has been implemented in two cases. These cases are adequate for the purpose of ‘proof of design’ use. However a different style of case could be considered in the future.
- Circuits for use in the characterisation of the system have been produced and used. The data gathered from them is consistent with simulation. They were then used to characterise the linear distortion due to the notch filter. The data gathered was validated against data gathered using a network analyser. All the available data was then used to construct a graph of the addition that must be made to results given the frequency such that the results are corrected.

- A signal generator was not implemented due to the constraints of time.
- Only 1 kHz filters were produced, it is the authors belief that a system which has been shown to work under one set of conditions, is likely to be adaptable to a similar set of conditions, given time. A system that may operate under a set of conditions, but is unfinished is worth less as it is unproven.
- Protection was not included due to the constraints of time.

10.2 Ideas for further development

Further development of this project could take two directions, continuing in the current vain, protection, a signal generator, frequency counter and option to operate at multiple frequencies could be developed as well as a more aesthetically pleasing case to house everything. An electrostatic screen would be required to separate the filters from the load resistors.

Alternatively the spectrum analyser could be dispensed with and replaced by a micro processor and an electronically tuned superheterodyne receiver of the type found in spectrum analyser front ends. Using this arrangement the microprocessor could be programmed to perform the entire test with minimal user interaction. The result could then be displayed on a graphical LCD as a spectrum (as in Figure 12) or as a percentage THD. There are several benefits to these modifications. The user cannot make an error that will endanger the equipment as the user doesn't have that level of control. The user cannot miss-interpret the results. The time taken for the test to run would be reduced as the act of writing results down and tuning the spectrum analyser are not necessary. It may even be possible to take the average of several tests by storing the data in memory. It would also be relatively easy (compared to a purely analogue solution) to interface the equipment with output devices such as a plotter. It is not plausible to place a DSP in the signal path to do the test using Fourier transforms. Analogue to digital converters cannot faithfully represent signals of such widely differing magnitudes as are found in this project

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Appendix

A1 System Diagram

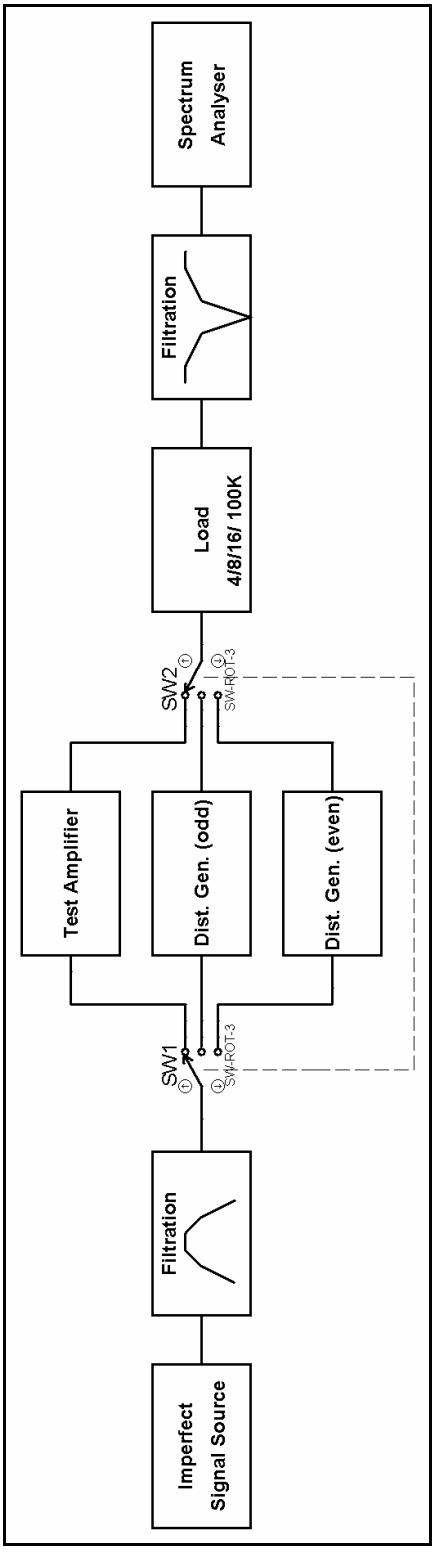


Figure 13: System Diagram

A2 Results

A2.1 Tables of Results

A2.1.1 Measured Data from Band Pass Filter

Frequency (Hz)	Magnitude (dBV)	Phase (°)
950	-1.723	-113.06
955	-1.322	-120.46
960	-0.945	-128.62
965	-0.608	-137.11
970	-0.351	-145.83
975	-0.149	-154.61
980	-0.037	-163.66
985	-0.038	-172.72
990	-0.105	178.22
995	-0.282	169.45
1000	-0.508	161.46
1005	-0.81	153.04
1010	-1.126	145.29
1015	-1.463	137.76
1020	-1.824	130.39
1025	-2.218	123.18
1030	-2.642	116.16
1035	-3.092	109.36
1040	-3.574	102.77
1045	-4.082	96.457
1050	-4.609	90.367

Table 4: Band Pass (pass band only) Data collected on Network Analyser see A2.2.5 for Graph

A2.1.2 Measured Data from Band Elimination Filter

Frequency (kHz)	Magnitude (dBV)	Phase (°)
2	-8.289	60.887
3	-6.668	36.488
4	-6.195	26.154
5	-6.013	20.23
6	-5.911	17.079
7	-5.828	14.315
8	-5.843	12.286
9	-5.798	10.668
10	-5.757	9.454
11	-5.758	8.531
12	-5.756	7.741
13	-5.749	6.962
14	-5.74	6.352
15	-5.739	5.877
16	-5.722	5.298
17	-5.709	4.916
18	-5.733	4.564
19	-5.712	4.185
20	-5.725	3.803

Table 5: Band Elimination (HF pass band only) Data collected on Network Analyser see A2.2.6 for Graph

A2.1.3 Spice Data from Even Harmonic Circuit

Harmonic	Frequency	Magnitude	Phase	Norm. Mag	Norm. Phase
0	0.00E+00	-2.33E-01	0.00E+00	0.00E+00	0.00E+00
1	1.00E+03	6.33E-01	-1.80E+02	1.00E+00	0.00E+00
2	2.00E+03	1.56E-01	8.99E+01	2.46E-01	2.70E+02
3	3.00E+03	4.61E-05	-9.07E+01	7.27E-05	8.93E+01
4	4.00E+03	3.11E-02	8.98E+01	4.92E-02	2.70E+02
5	5.00E+03	4.01E-05	-9.22E+01	6.32E-05	8.78E+01
6	6.00E+03	1.34E-02	8.97E+01	2.11E-02	2.70E+02
7	7.00E+03	3.71E-05	-9.36E+01	5.86E-05	8.64E+01
8	8.00E+03	7.43E-03	8.96E+01	1.17E-02	2.70E+02
9	9.00E+03	3.51E-05	-9.50E+01	5.54E-05	8.50E+01
10	1.00E+04	4.74E-03	8.94E+01	7.48E-03	2.69E+02
11	1.10E+04	3.35E-05	-9.65E+01	5.29E-05	8.35E+01
12	1.20E+04	3.29E-03	8.92E+01	5.19E-03	2.69E+02
13	1.30E+04	3.22E-05	-9.80E+01	5.08E-05	8.20E+01
14	1.40E+04	2.42E-03	8.90E+01	3.82E-03	2.69E+02
15	1.50E+04	3.10E-05	-9.95E+01	4.90E-05	8.05E+01
16	1.60E+04	1.85E-03	8.88E+01	2.93E-03	2.69E+02
17	1.70E+04	3.00E-05	-1.01E+02	4.73E-05	7.89E+01
18	1.80E+04	1.47E-03	8.86E+01	2.32E-03	2.69E+02
19	1.90E+04	2.90E-05	-1.03E+02	4.58E-05	7.73E+01
20	2.00E+04	1.19E-03	8.83E+01	1.88E-03	2.68E+02

THD: 25.2014 %

Table 6: Fourier Analysis of Spice Transient Data in Appendix 2.22

A2.1.4 Spice Data from Odd Harmonic Circuit

Harmonic	Frequency	Magnitude	Phase	Norm. Mag	Norm. Phase
0	0.00E+00	-2.35E-03	0.00E+00	0.00E+00	0.00E+00
1	1.00E+03	9.63E-01	3.16E+00	1.00E+00	0.00E+00
2	2.00E+03	1.67E-03	7.36E+01	1.73E-03	7.04E+01
3	3.00E+03	2.69E-01	1.71E+02	2.80E-01	1.68E+02
4	4.00E+03	9.38E-04	4.63E+01	9.73E-04	4.31E+01
5	5.00E+03	7.12E-02	1.62E+02	7.39E-02	1.59E+02
6	6.00E+03	5.02E-04	-7.16E-01	5.21E-04	-3.88E+00
7	7.00E+03	5.61E-03	1.20E+02	5.82E-03	1.17E+02
8	8.00E+03	3.63E-04	-5.94E+01	3.76E-04	-6.25E+01
9	9.00E+03	1.34E-02	-2.28E+01	1.39E-02	-2.60E+01
10	1.00E+04	2.85E-04	-1.08E+02	2.95E-04	-1.11E+02
11	1.10E+04	1.17E-02	-3.97E+01	1.21E-02	-4.29E+01
12	1.20E+04	1.93E-04	-1.60E+02	2.00E-04	-1.63E+02
13	1.30E+04	5.23E-03	-6.37E+01	5.43E-03	-6.69E+01
14	1.40E+04	1.53E-04	1.36E+02	1.59E-04	1.33E+02
15	1.50E+04	1.63E-03	-1.57E+02	1.69E-03	-1.60E+02
16	1.60E+04	1.44E-04	8.27E+01	1.50E-04	7.95E+01
17	1.70E+04	2.65E-03	1.34E+02	2.75E-03	1.31E+02
18	1.80E+04	1.16E-04	3.59E+01	1.21E-04	3.27E+01
19	1.90E+04	2.27E-03	1.04E+02	2.36E-03	1.01E+02
20	2.00E+04	8.37E-05	-1.98E+01	8.69E-05	-2.30E+01

THD: 29.0022 %

Table 7: Fourier Analysis of Spice Transient Data in Appendix 2.23

A2.1.5 Measured Data from Even Harmonic Circuit (with band pass without band elimination)

Frequency (kHz)	Magnitude (dBV)	Voltage (V)	Voltage^2 (V^2)	THD (%)
1	-4	6.310E-01	3.981E-01	24.93
2	-16	1.585E-01	2.512E-02	
4	-30	3.162E-02	1.000E-03	
6	-38	1.259E-02	1.585E-04	
8	-43	7.079E-03	5.012E-05	
10	-47	4.467E-03	1.995E-05	
12	-50	3.162E-03	1.000E-05	
14	-52	2.512E-03	6.310E-06	
16	-54	1.995E-03	3.981E-06	
18	-56	1.585E-03	2.512E-06	
20	-58	1.259E-03	1.585E-06	

Table 8: Measured Data from Even Harmonic Circuit (with band pass without band elimination)

A2.1.6 Measured Data from Odd Harmonic Circuit (with band pass without band elimination)

Frequency (kHz)	Magnitude (dBV)	Voltage (V)	Voltage^2 (V^2)	THD (%)
1	0	1.000E+00	1.000E+00	30.89
3	-10	3.162E-01	1.000E-01	
5	-23	7.079E-02	5.012E-03	
7	-48	3.981E-03	1.585E-05	
9	-36	1.585E-02	2.512E-04	
11	-38	1.259E-02	1.585E-04	
13	-48	3.981E-03	1.585E-05	
15	-52	2.512E-03	6.310E-06	
17	-50	3.162E-03	1.000E-05	
19	-54	1.995E-03	3.981E-06	

Table 9: Measured Data from Odd Harmonic Circuit (with band pass without band elimination)

A2.1.7 Measured Data from Even Harmonic Circuit (with band pass and *with* band elimination)

Frequency (kHz)	Magnitude (dBV)	Difference (dBV)
1	---	---
2	-25	9
4	-37	7
6	-44	6
8	-49	6
10	-54	7
12	-56	6
14	-58	6
16	-61	7
18	-64	8
20	-64	6

Table 10: Measured Data from even Harmonic Circuit (with band pass and with band elimination)

A2.1.8 Measured Data from Odd Harmonic Circuit (with band pass and *with* band elimination)

Frequency (kHz)	Magnitude (dBV)	Difference (dBV)
1	---	---
3	-17	7
5	-28	5
7	-54	6
9	-43	7
11	-44	6
13	-54	6
15	-59	7
17	-56	6
19	-59	5

Table 11: Measured Data from odd Harmonic Circuit (with band pass and with band elimination)

A2.1.9 Measured results from Sansui A-700 (First Test)

Frequency (kHz)	Magnitude (dBV)	Correction (dBV)	Sum (dBV)	Voltage (V)	Voltaege^2 (V^2)
1				1.000E-01	1.000E-02
2	-85	8.3	-76.7	1.462E-04	2.138E-08
3	-87	6.7	-80.3	9.661E-05	9.333E-09
4	-92	6.2	-85.8	5.129E-05	2.630E-09
5	-104	6.0	-98.0	1.259E-05	1.585E-10
6	-97	6.0	-91.0	2.818E-05	7.943E-10
7	-99	5.8	-93.2	2.188E-05	4.786E-10
8	-104	5.8	-98.2	1.230E-05	1.514E-10
9	-102	5.8	-96.2	1.549E-05	2.399E-10
10		5.8			
11	-105	5.8	-99.2	1.096E-05	1.202E-10
12		5.7			
13	-106	5.7	-100.3	9.661E-06	9.333E-11
14		5.7			
15	-110	5.7	-104.3	6.095E-06	3.715E-11
16	-112	5.7	-106.3	4.842E-06	2.344E-11
17		5.7			
18	-112	5.7	-106.3	4.842E-06	2.344E-11
19	-112	5.7	-106.3	4.842E-06	2.344E-11
20		5.7			

Numerator	3.5486E-08
Denominator	1.0000E-02
THD	0.0001884
THD (%)	0.01884
THD (dB)	-74.50

Table 12: Results of First Test on Sansui A-700

A2.1.10 Measured results from Sansui A-700 (Second Test)

Frequency (kHz)	Magnitude (dBV)	Correction (dBV)	Sum (dBV)	Voltage (V)	Voltaege^2 (V^2)
1				1.000E-01	1.000E-02
2	-84	8.3	-75.7	1.641E-04	2.692E-08
3	-88	6.7	-81.3	8.610E-05	7.413E-09
4	-91	6.2	-84.8	5.754E-05	3.311E-09
5		6.0			
6	-96	6.0	-90.0	3.162E-05	1.000E-09
7	-108	5.8	-102.2	7.762E-06	6.026E-11
8	-102	5.8	-96.2	1.549E-05	2.399E-10
9	-110	5.8	-104.2	6.166E-06	3.802E-11
10	-104	5.8	-98.2	1.230E-05	1.514E-10
11		5.8			
12	-106	5.7	-100.3	9.661E-06	9.333E-11
13		5.7			
14	-110	5.7	-104.3	6.095E-06	3.715E-11
15	-107	5.7	-101.3	8.610E-06	7.413E-11
16	-110	5.7	-104.3	6.095E-06	3.715E-11
17		5.7			
18	-112	5.7	-106.3	4.842E-06	2.344E-11
19		5.7			
20	-110	5.7	-104.3	6.095E-06	3.715E-11

Numerator	3.9432E-08
Denominator	1.0000E-02
THD	0.0001986
THD (%)	0.01986
THD (dB)	-74.04

Table 13: Results of Second Test on Sansui A-700

A2.2 Set of Graphs

A2.2.1 Spice Data from Band Pass Filter

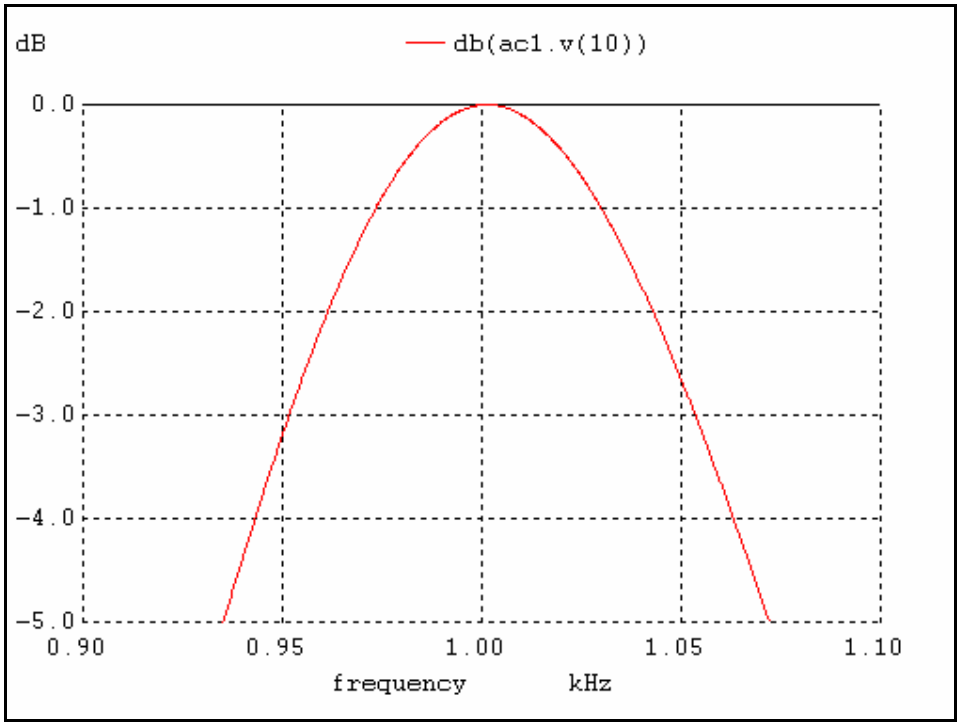


Figure 14: Band Pass Cascade Response 3dB Points (uses obtainable component values)

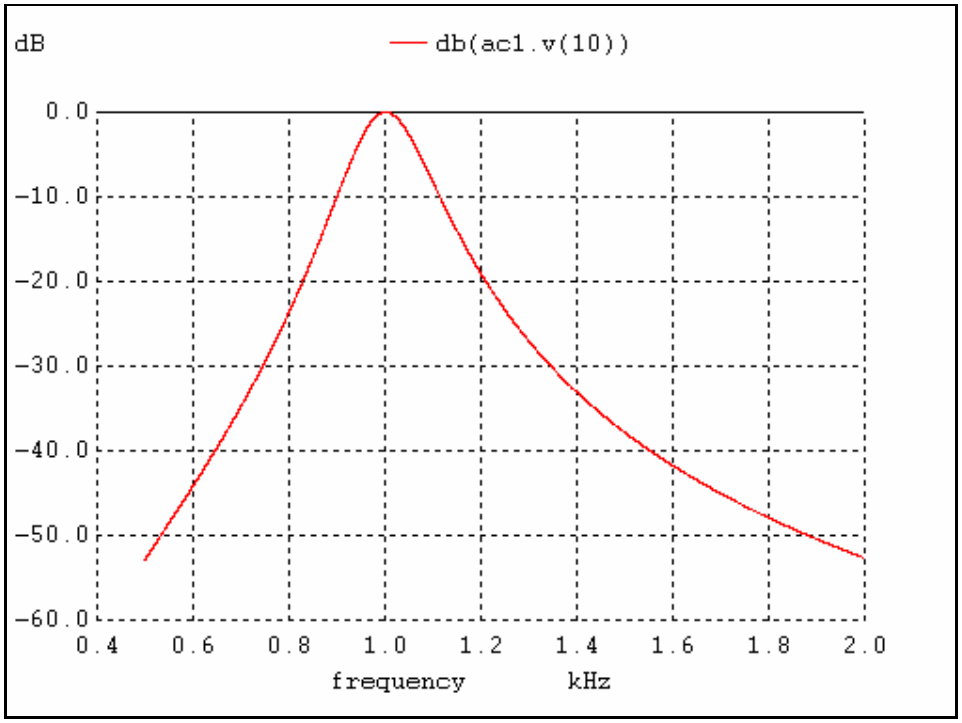


Figure 15: Band Pass Responce Showing -50dB Points

A2.2.2 Spice Data from Band Elimination Filter

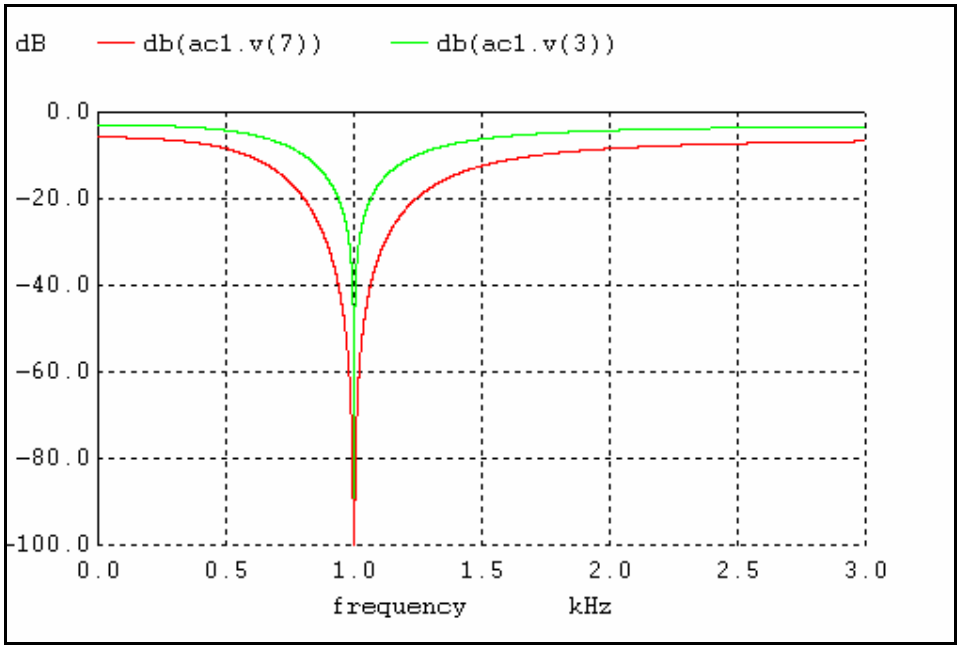


Figure 16: Spice data from the output of the first and second filter section sections

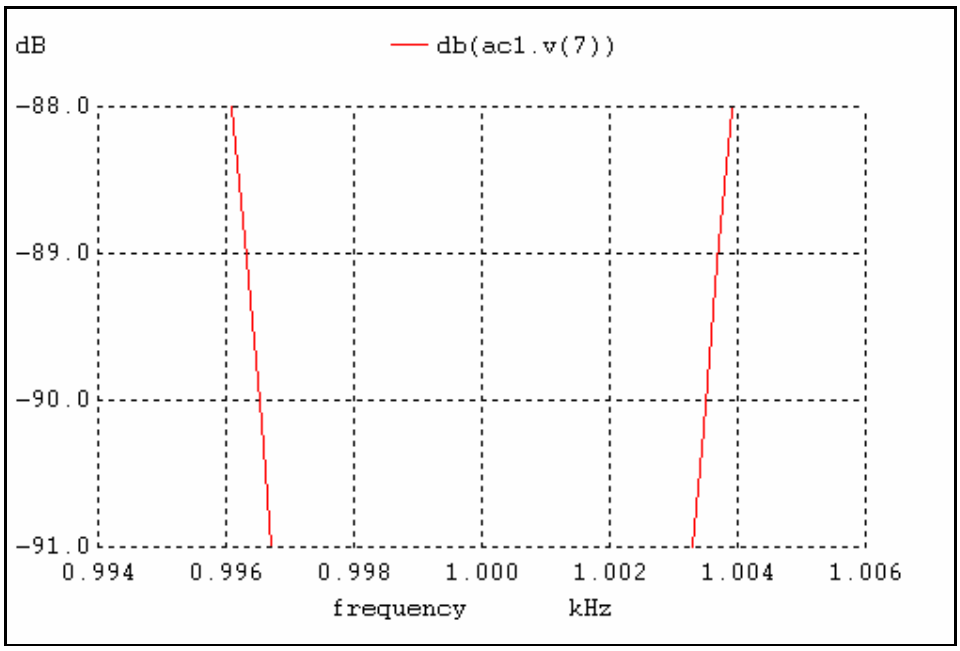


Figure 17: Spice data showing the bandwidth at the anticipated attenuation

A2.2.3 Spice Data from Even Harmonic Distortion Circuit

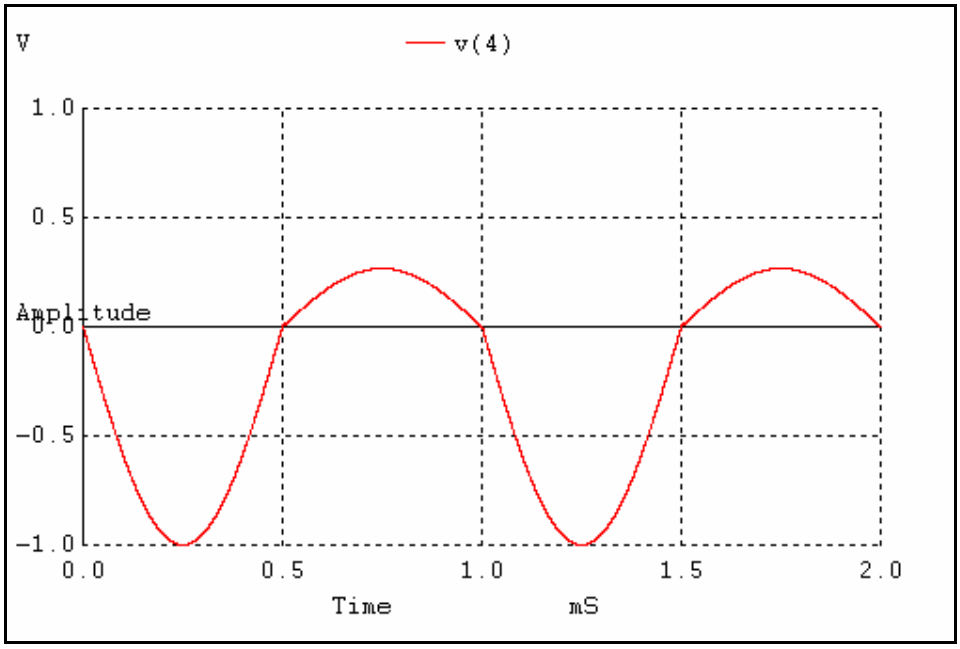


Figure 18: Transient Analysis of Even Harmonic Distortion Circuit

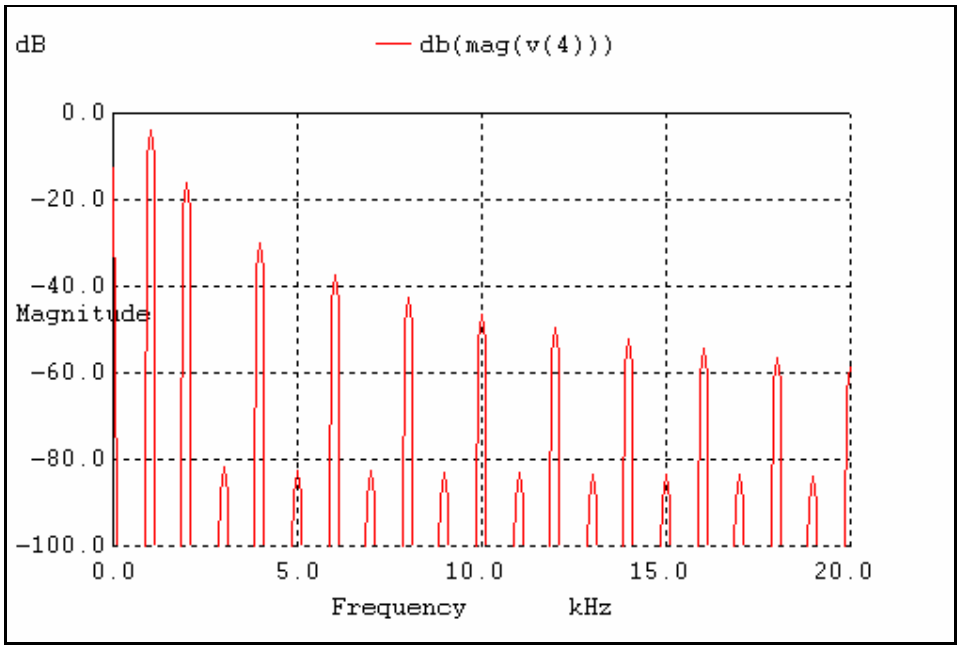


Figure 19: Fourier Transform of above Transient Data

A2.2.4 Spice Data from Odd Harmonic Distortion Circuit

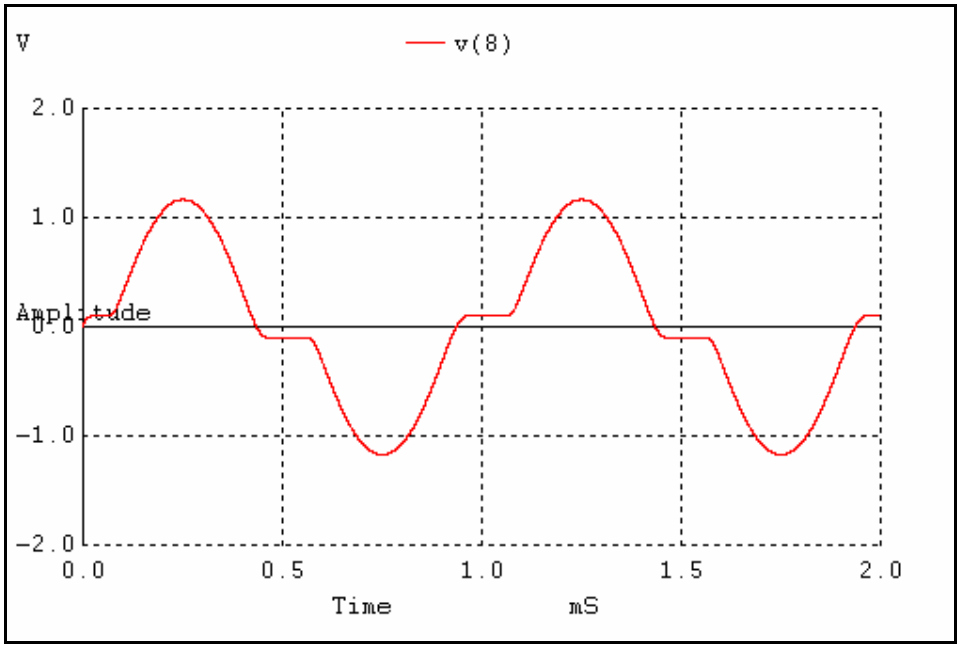


Figure 20: Transient Analysis of Odd Harmonic Distortion Circuit

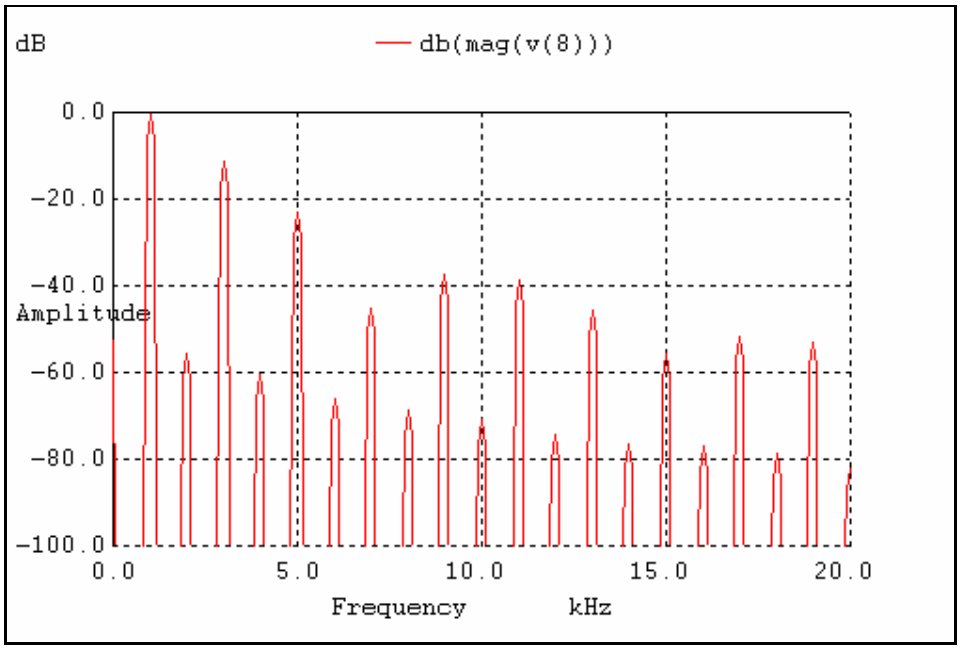


Figure 21: Fourier Transform of above Transient Data

A2.2.5 Measured Results from Band Pass Filter on Network Analyser

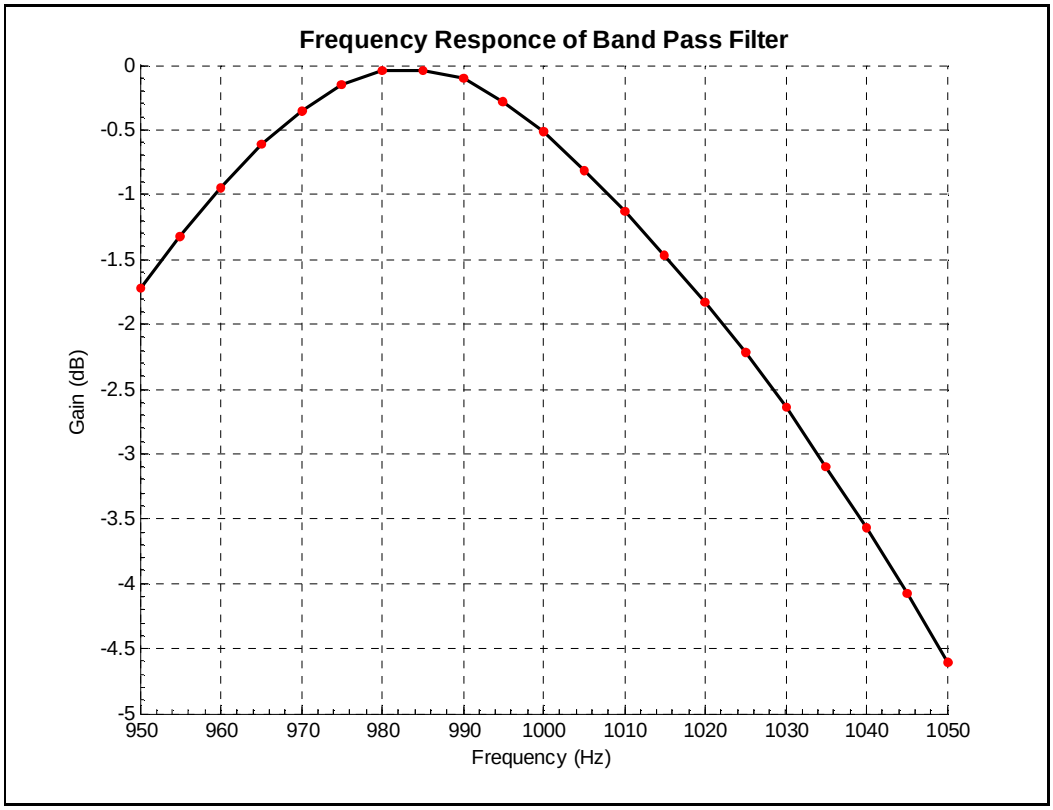


Figure 22: Frequency Response of Band Pass Filter measured on Network Analyser

A2.2.6 Measured Results from Band Elimination filter on Network Analyser

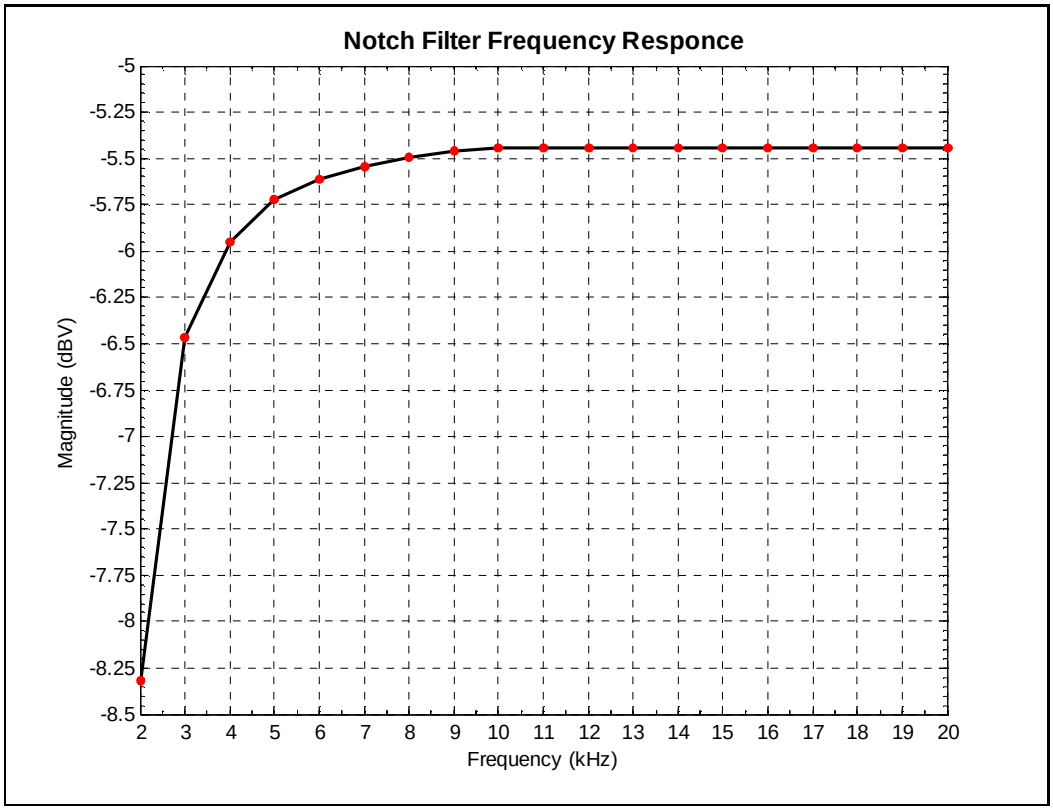


Figure 23: Frequency Response of Band Elimination Filter measured on Network Analyser

A2.2.7 Measured Results from Even Harmonic Distortion Circuit

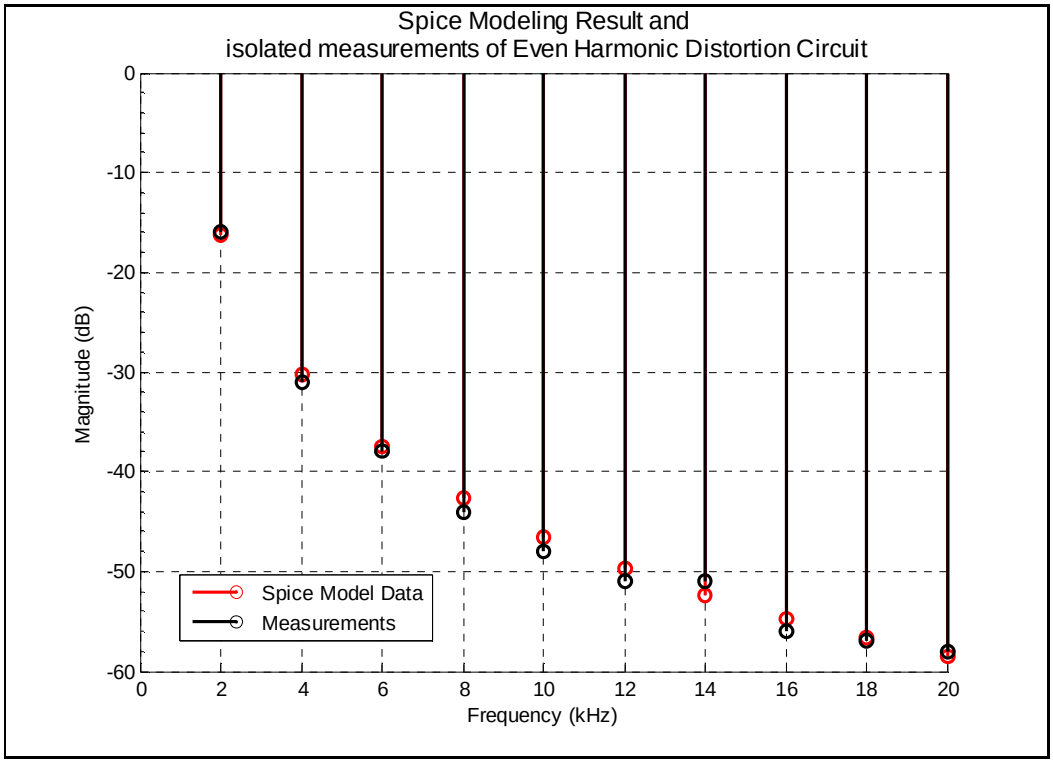


Figure 24: Even Harmonic distortion data (without BE filter) compared with spice model data.

A2.2.8 Measured Results from Odd Harmonic Distortion Circuit.

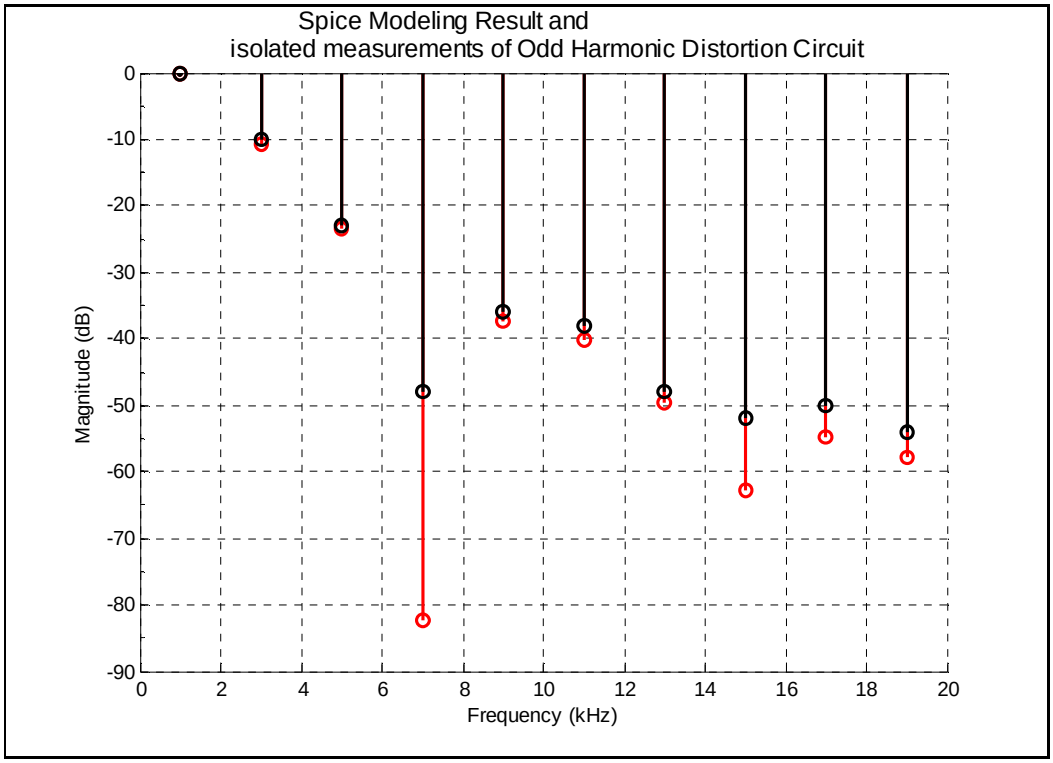


Figure 25: Odd Harmonic distortion data (without BE filter) compared with spice model data.

A2.2.9 Measured Even Harmonic Distortion Circuit without BE Filter Vs. with BE Filter

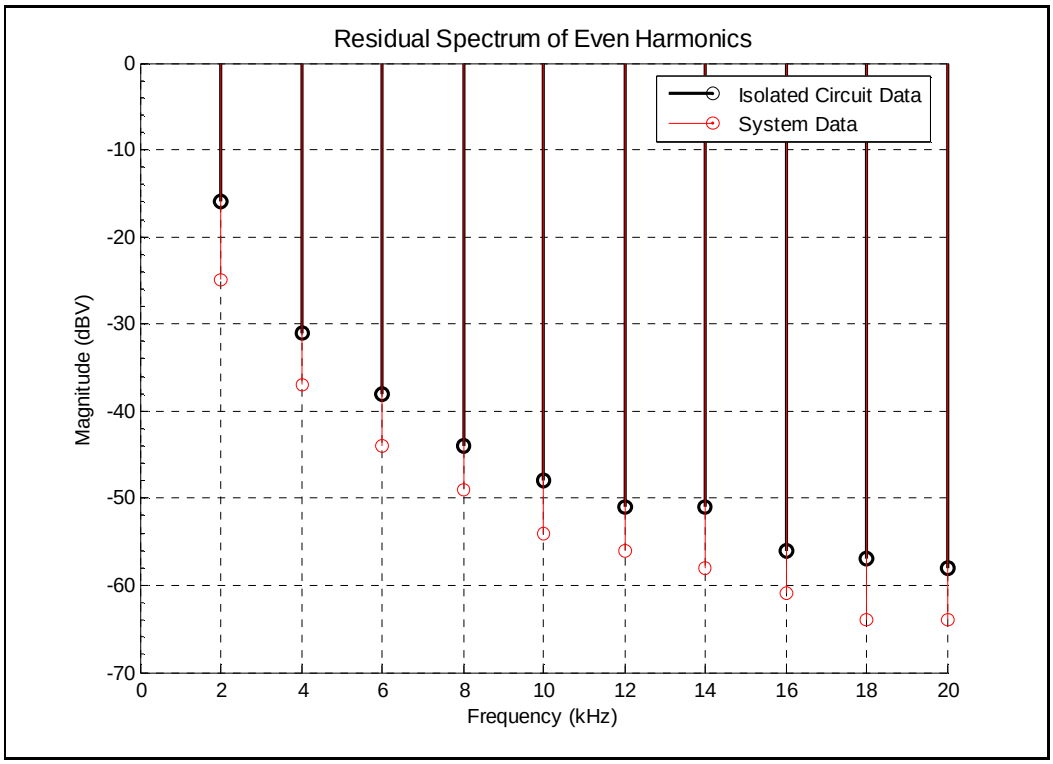


Figure 26: Measured Even Harmonic Distortion Circuit without BE Filter Vs. with BE Filter

A2.2.10 Measured Odd Harmonic Distortion Circuit without BE Filter Vs. with BE Filter

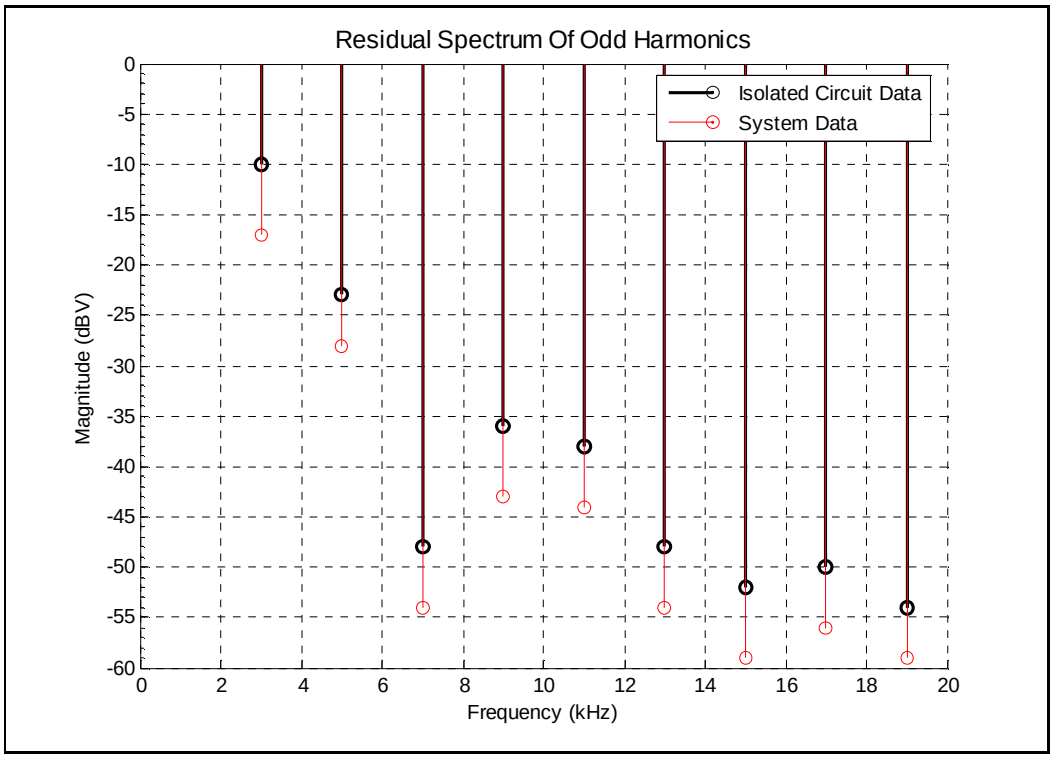


Figure 27: Measured Odd Harmonic Distortion Circuit without BE Filter Vs. with BE Filter

A2.2.11 Measured Results from Sansui A 700 (First Test)

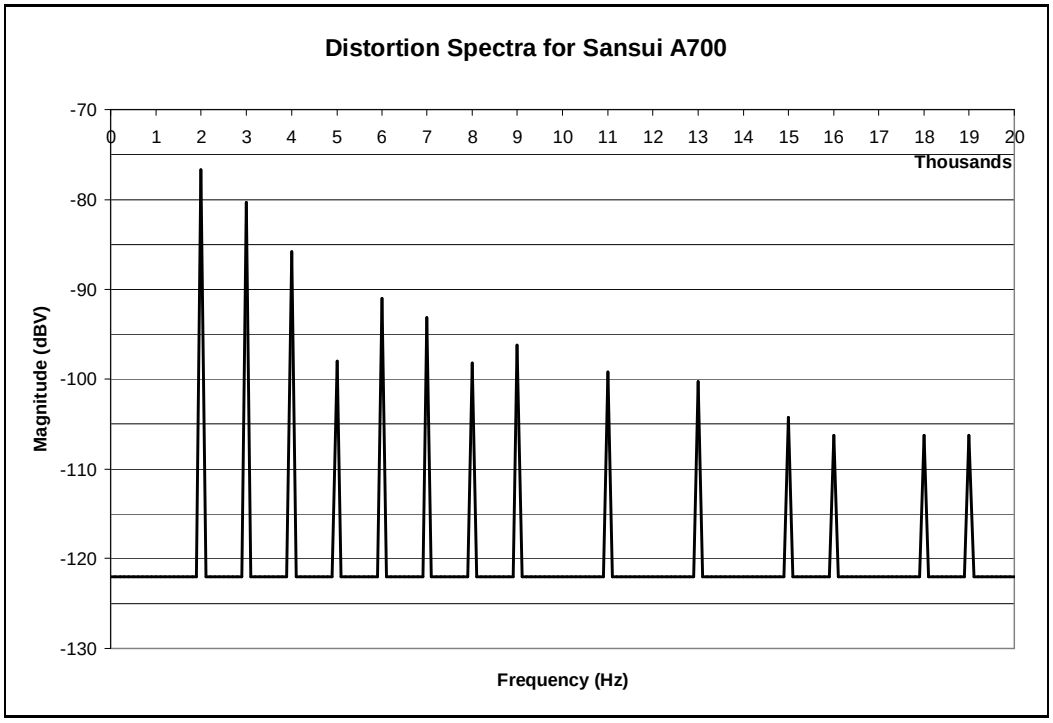


Figure 28: Measured results from Sansui A 700 (First Test)

A2.2.12 Measured Results from Sansui A 700 (Second Test)

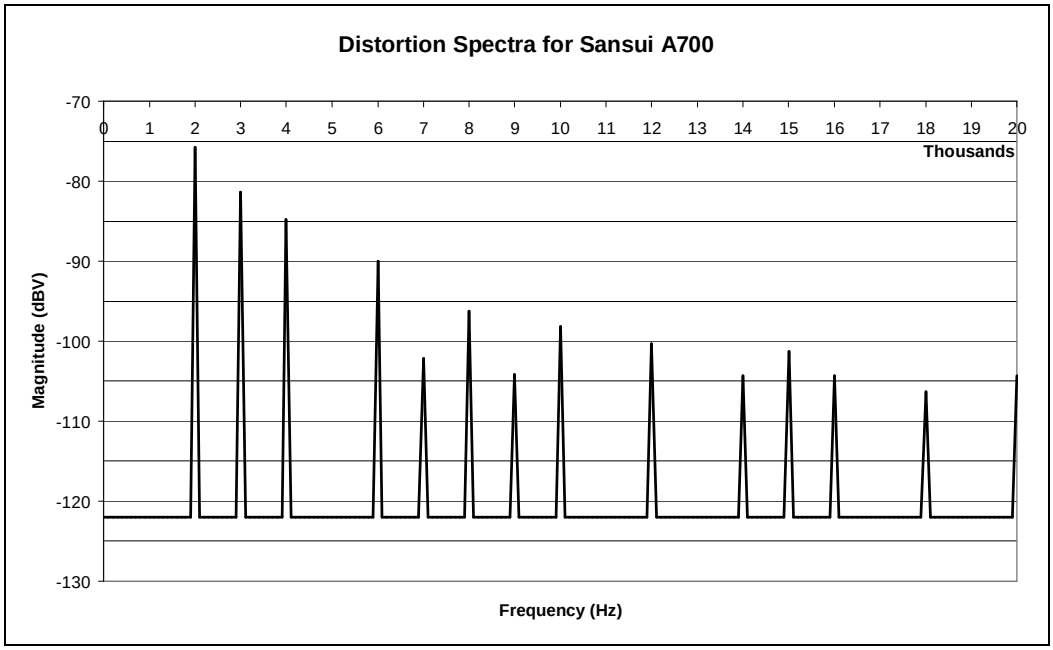


Figure 29: Measured results from Sansui A 700 (Second Test)

A2.2.13 Network Analyser Plots of Band Pass Filter

A2.2.14 Network Analyser Plots of Band Elimination Filter

A3 Complete Circuit Diagrams

A3.1 Band pass Filter Circuit Diagram

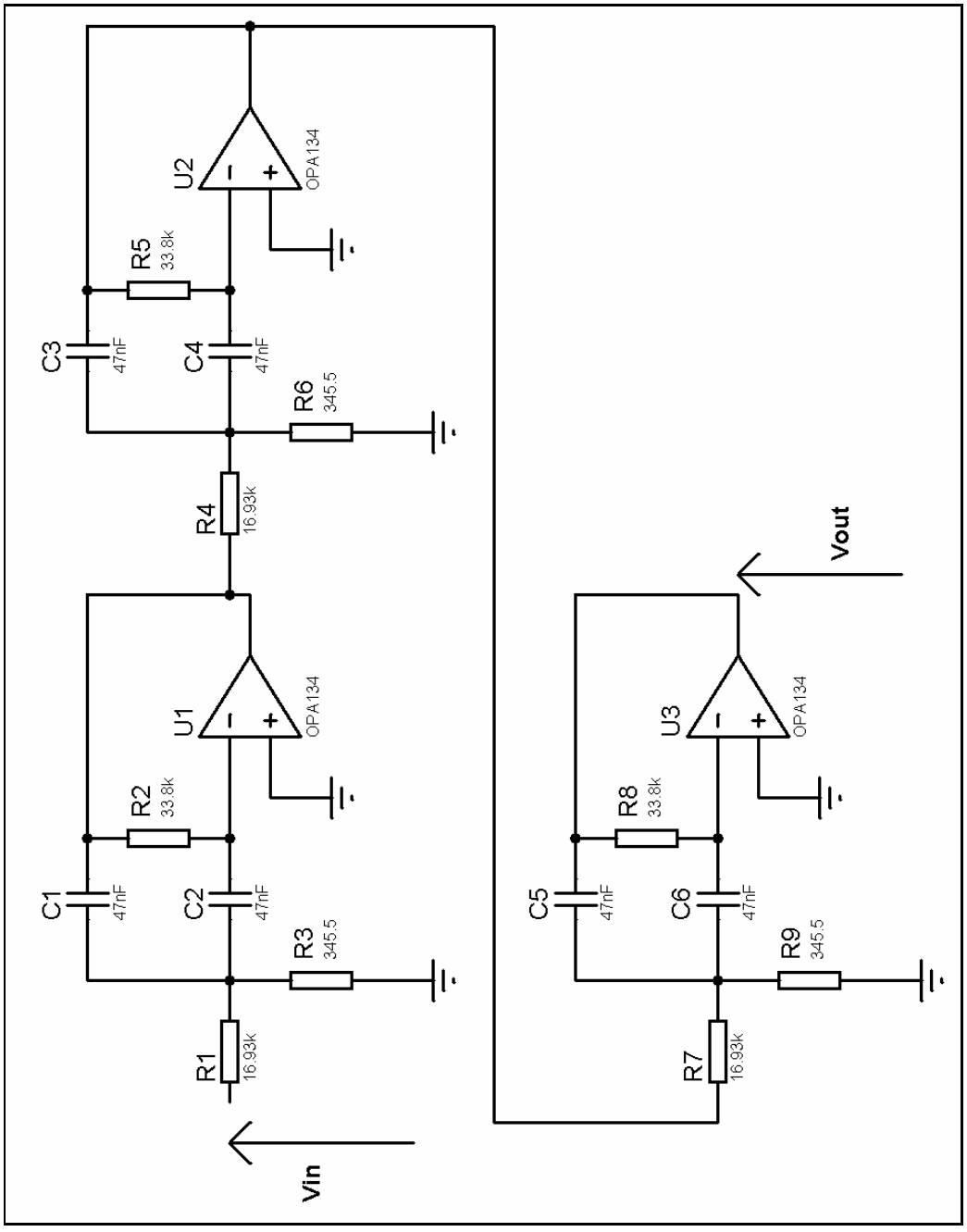


Figure 30: Band Pass Filter Schematic

A3.2 Band Elimination Cascade Schematic

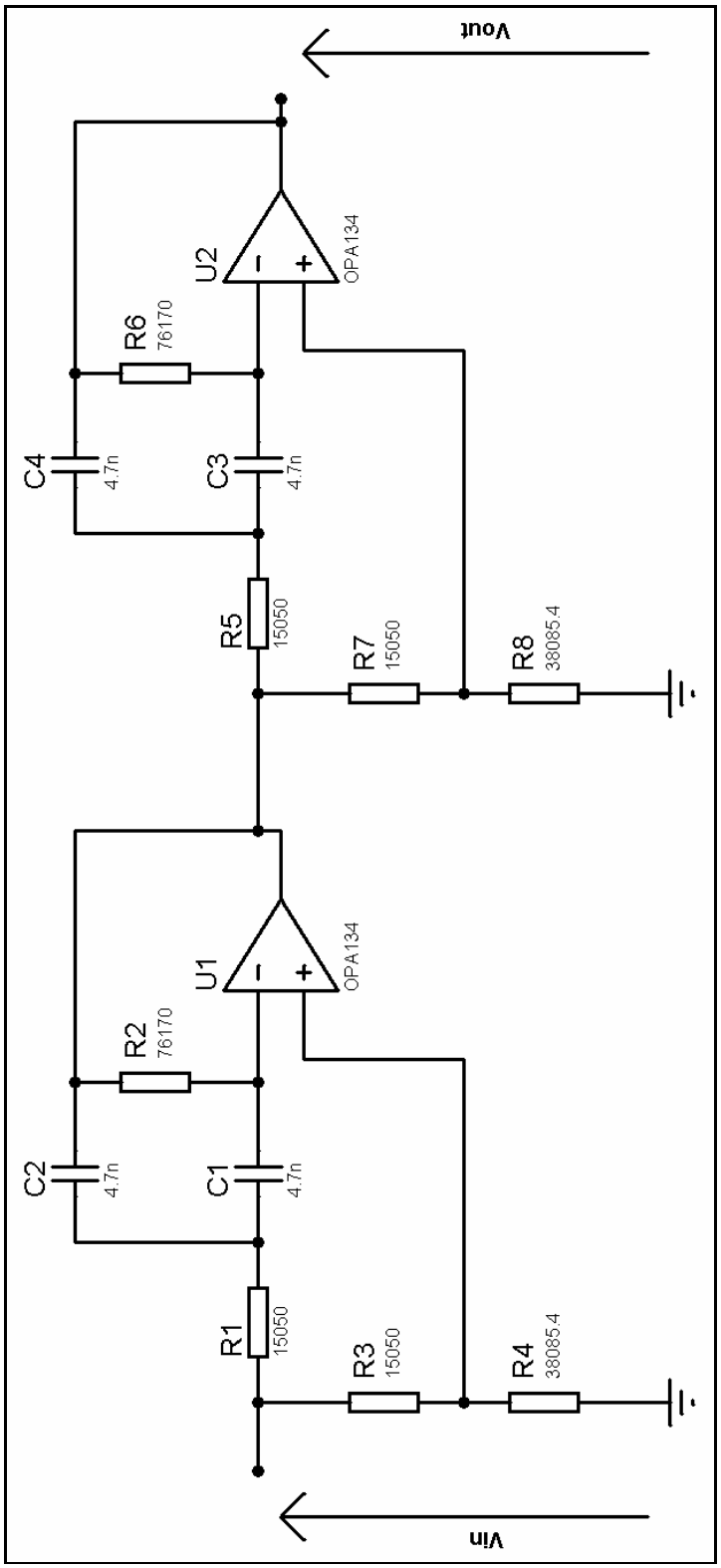


Figure 31: Band Elimination Schematic

A3.3 System Schematic

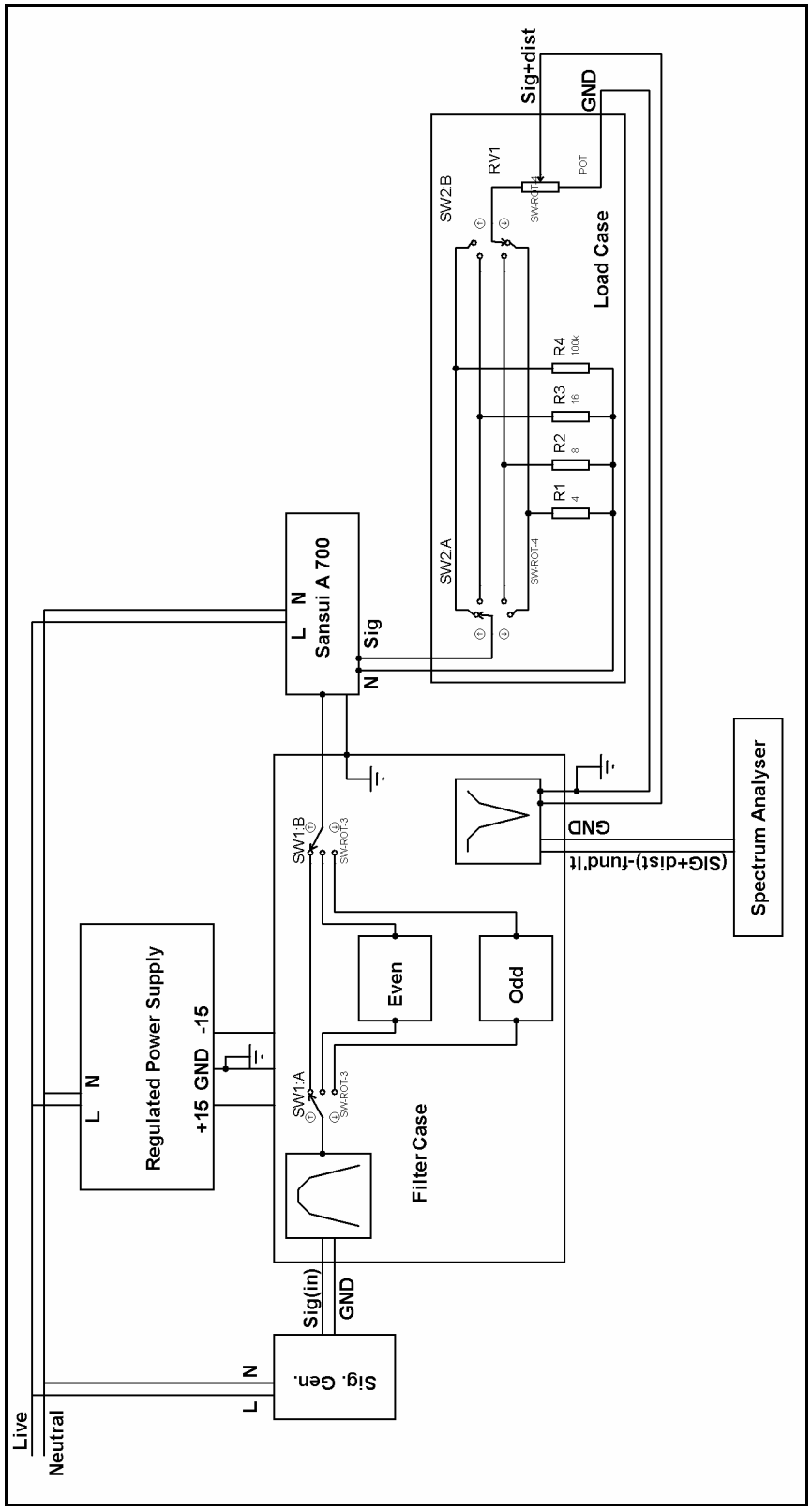


Figure 32: Diagram Showing Connections Between Subsystems, and Grounding

A3.4 Even Harmonic Distortion Circuit

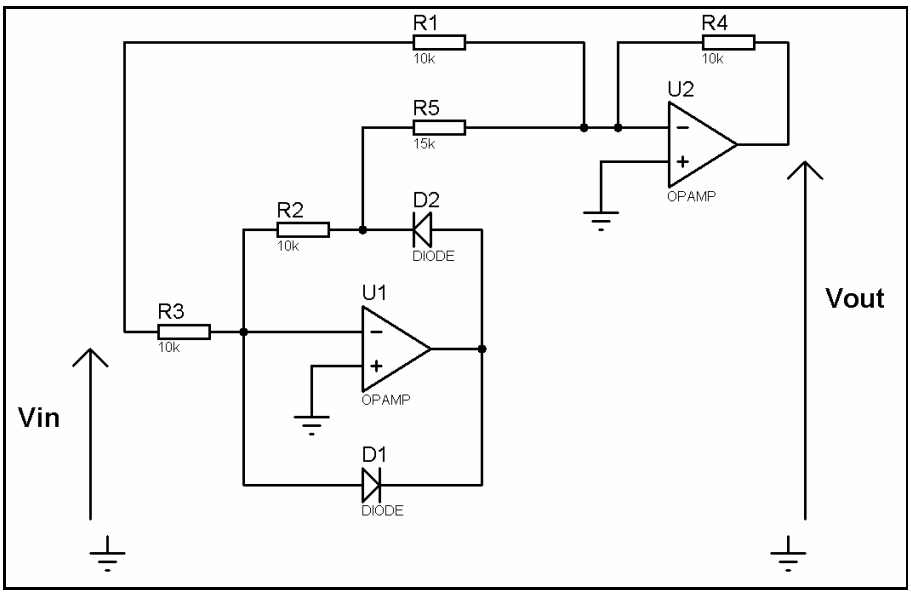


Figure 33: Even Harmonic Distortion Circuit

A3.5 Odd Harmonic Distortion Circuit

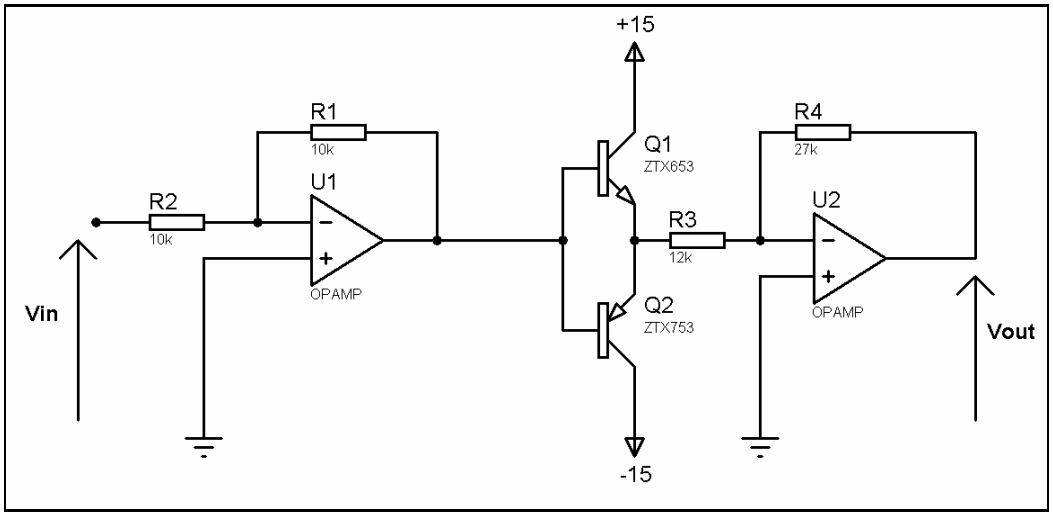


Figure 34: Odd Harmonic Distortion Circuit

A3.6 Hardwired Moving Average Counter (Example)

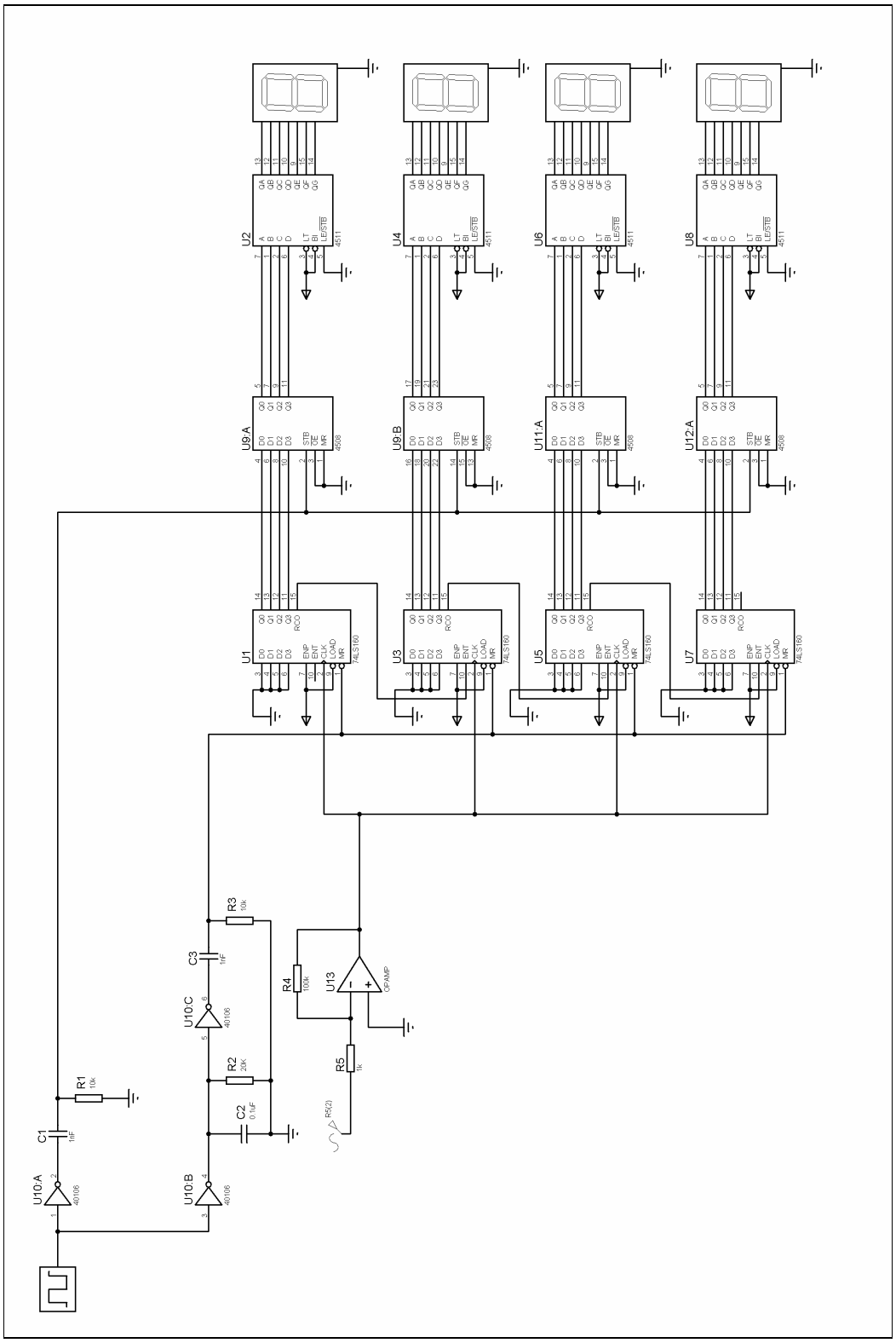


Figure 35: Example Frequency Counter Using Logic

A3.7 Loading Circuitry

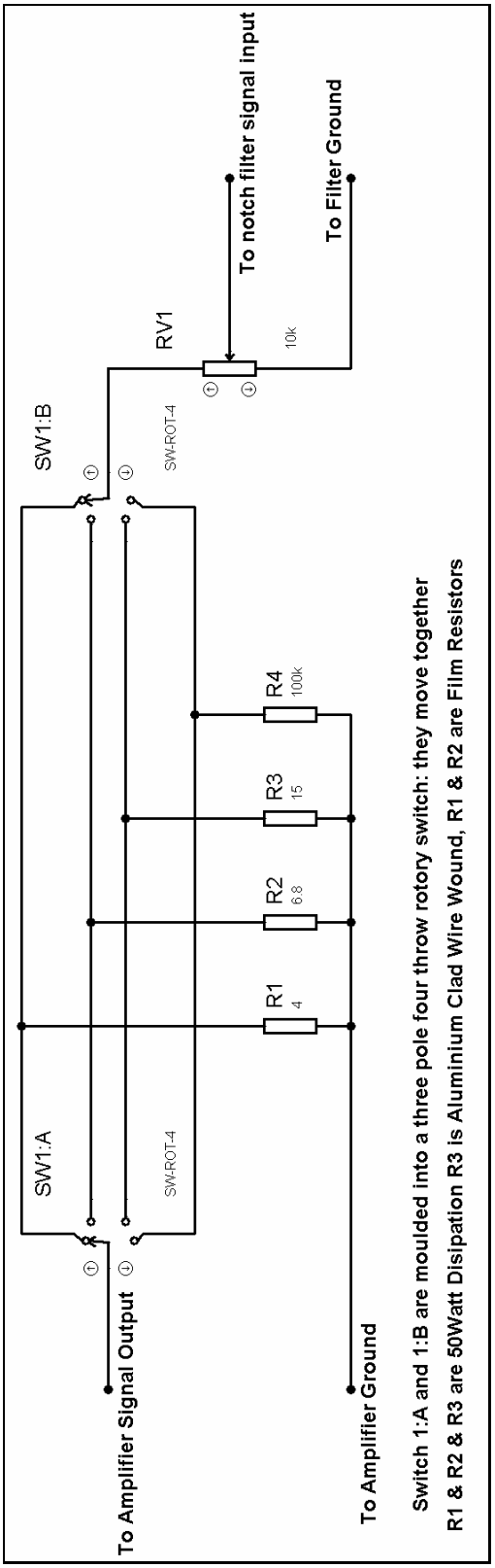


Figure 36: Loading Schematic

A4 List of Operational Amplifiers

5532
AD648JN
AD708JN
AD797
CA3130
LF412
LM307
LM833N
LT1013CN8
NE5532
NJM2068D
NJM2122D
NJM4562D
OP27
OP29
OP490GP
OPA134
OPA604
TL071
TL081
LM741
TLC272
TLE2072CD

Table 14: List of Considered Operational Amplifiers

A5 Thermal Calculations

Having consulted datasheets for the Vishay and Alcol resistor(s) the following thermal diagrams were drawn.

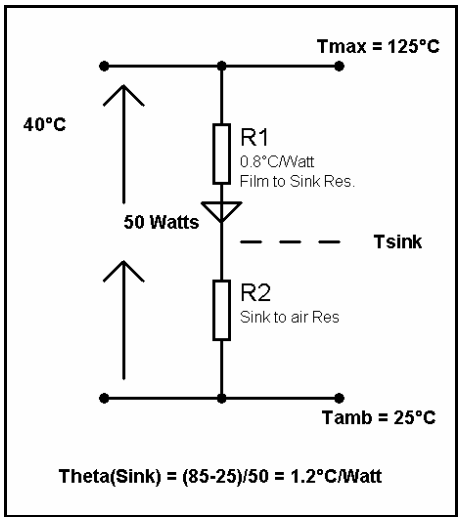


Figure 37: Vishay Film Resistors

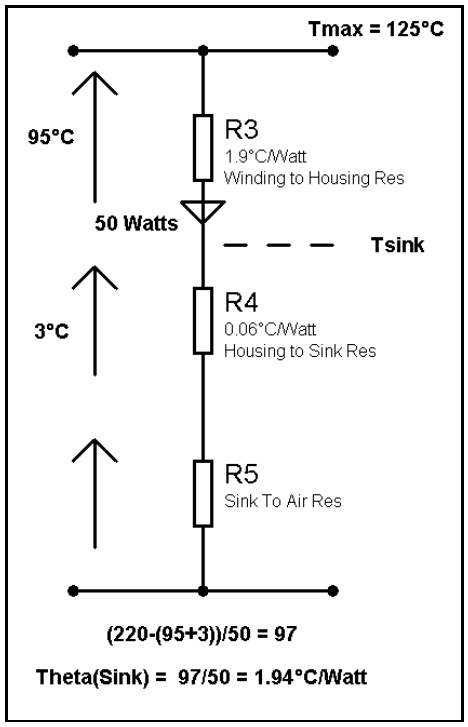


Figure 38: Alcol Wire Wound Aluminium Clad Resistor

The above diagrams are drawn on the principle that the current in the load is DC. This is not the case. However the heatsink rating is given for

"Vertical, in free air."

The case used does not permit this. The sink is horizontal and fills the aluminium case to which it is attached. These conditions make accurate estimation of the heatsink size given the power dissipated by a sinusoid (using calculus) invalid. The DC condition is a more suitable 'worst case' estimation.

A6 Interim Report

Third Year Project

Interim Report

(Week 12)

James Green

030119646

Contributing to Module EEE360

Project Supervisor DR. R C Tozer

Second Marker Prof. Z Zhu

Project Description

The aim of the project is to produce a piece of equipment whose purpose is to allow harmonic distortion measurements to be made using a spectrum analyser and a signal generator. The equipment should be designed for use over the range of audible frequencies (20 Hz to 20 kHz).

'Traditional' Distortion Measurement

Traditional distortion measurements were made using 'audio test sets' such as the HP 339A and 334A [1]. Test sets such as these normally include some form of oscillator which is passed to the test amplifier. Having been distorted in the amplifier, the returning signal is presented to a tuned circuit of the same form that is found in RF receivers. The user can change the resonant frequency of this tuned circuit to pick out the harmonic of interest. The result of the filtration is displayed on a moving coil meter normally in RMS Volts.

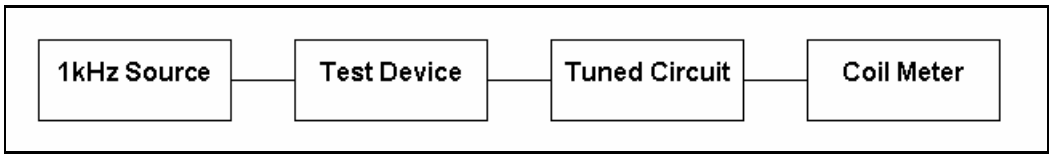


Figure 39: Simplified System diagram for 'audio test set'

Such distortion measurement sets persist to the present. The moving coil meter is sensitive to all voltage amplitudes in the pass band of the tuned filter. If the signal that is received has a noise voltage of 30mV, and a distortion residual of 10mV at the resonant frequency of the tuned circuit, the RMS volt meter will not differentiate the noise from the distortion and the reading is unhelpful. [2] Morgan Jones suggests that a spectrum analyser could be used to present the distortion residual (that is, the output from the test device with the input signal removed). This would allow the distortion components to be detected against the noise in the system. It also removes the necessity of a user operated tuned circuit. It could be replaced by a filter which removes the fundamental and leaves the residual.

Most of these audio test sets had a signal source built in. The distortion of such sources could be made remarkably low perhaps as low as -80dB even when employing valves as the active devices. The most likely choice of circuit is the Wein Bridge oscillator.

Progress

As part of the project specification presented in 'Report 1' it is necessary to provide the user with a means to connect an external signal source, such as a function generator. This signal would be passed to the test device. Modern function generators do not work on the same principle as the generators in 'test sets'.

The Wein bridge is good at generating a single frequency with minimal distortion [3] Horowitz and Hill quote 0.003% THD. Its principle of operation is to cause regenerative feedback in an amplifier at the required output frequency. It then measures the output voltage, converts it to a DC value and uses that value to change the magnitude of negative feedback on the amplifier, hence keeping the amplifier from saturating, while generating a sine wave with negligible distortion.

A modern signal generator, which is required to have a usable frequency range of many decades, is usually a digital system that constructs its sine wave using a process known as coordinate rotation. Other methods include approximation by series and lookup tables. Continuous time solutions mainly use RC circuits that are modified by some filtration.

Signal generators currently being sold produce between 0.1% and 2% THD. These figures are the extreme examples from a set of ten generators currently available from RS, Farnell, and Rapid Electronics. Considering a well designed power amplifier could have distortion as low as 0.005% at 1 kHz at full output power, it is clear some signal processing must be applied to the signal generators output. To this end a band pass filter was designed

Filter Responses

The Butterworth response's notable point is it has 'maximally flat' pass band(s). The band pass is the logical choice as it allows the attenuation of low frequencies (including DC) at no extra cost. The Butterworth response is also a logical choice as it can be designed to have a pass band of, for example, 1000Hz +/- 50Hz; thus allowing for the inaccuracy and drift associated with signal generators (especially analogue).

However, the Butterworth response was not carried forward, as a simpler solution having very minor drawbacks is available. A Butterworth response is produced by implementing a number of second order filter circuits, each of them having a different frequency of resonance and quality factor. Real components have a tolerance and the response of all electronic systems is dependant on the tolerance of the components used in their construction. The sensitivity of a circuit is defined as the percentage change in the output frequency upon the change in the component value. This is considered for all the components in the system. This allows graphs to be drawn which show the maximum deviation from the expected value. The Butterworth response will deviate from the maximally flat condition and hence it is more logical to

use a filter whose stages are all tuned to the same frequency. Using this arrangement means there will always be some loss in the pass band, but the loss can be compensated for by increasing the output level on the generator. Having analysed both systems the single resonant frequency design was carried forward, as it is easier to implement and has no complicating drawbacks when considered against the stagger tuned Butterworth response. The single resonant frequency tuned filter has been recently constructed and performs agreeably when compared to the results of spice modelling. The resonant frequency is only around 5Hz from its anticipated value and the maximum expected attenuation in the pass band has not been exceeded (less than 0.5 dB. The attenuation at the beginning of the stop band is 40dB while the filter was designed for 50dB. This discrepancy is currently an open avenue of investigation. It is unlikely to require any redesign. A filter of 40db attenuation in conjunction with a second harmonic distortion from a generator of -40db would lead to a final second harmonic distortion of -80db which is acceptable.

Notch Filter

The user will not be able to select the harmonic of interest as the spectrum analyser will be used to show all the harmonics of interest simultaneously. Hence a filter response is required that can be used to attenuate significantly the fundamental and leave only the residual. The band elimination, or notch, is the only obvious solution. The notch is different from other filter responses in that the attenuation is linked to the bandwidth. There is no way around this. A Butterworth band elimination filter was considered the optimal solution for several weeks; however the ideal response and the likely response (having discussed the problem with Dr. Tozer) were considered to be so different that the solution was impractical. As a result, a fourth order notch with a single resonant frequency was designed and constructed. This notch has a disadvantage in that it still has significant attenuation at almost all harmonics of interest. It will be necessary to measure the reduction in these frequencies with considerable accuracy and precision. A method of accomplishing this is yet to be devised. The printed circuit board (hereafter PCB) was constructed, but didn't function as expected. A short inspection revealed an irreparable error in the board, rendering it useless. A corrected board has been manufactured, and measurements have been made using a network analyser. The notch depth is 63 dB at the frequency of interest. Most test sets attenuate the fundamental by around 30dB.

System Distortion

The circuitry that is currently under construction will distort the signal. It is necessary to find the result of this distortion, such that it can be factored into the results of any test. This distortion is inherent in all circuitry with nonlinear devices. In traditional test sets it would be necessary to calibrate the equipment using a circuit with some known (calculable) distortion. The system currently being designed can't be calibrated in the same way as a coil meter. The idea of using a circuit of known distortion to find the 'characteristic' distortion of the system is valid. It would be impossible to measure accurately the characteristic distortion directly. Using a circuit of

known distortion, in place of the test device, makes it possible to find the 'error' which will be the difference between the calculated value and the value read on the spectrum analyser.

Two circuits have been devised to implement the distortion required and Matlab has been used to find their likely distortion components. The values have been crosschecked with spice modelling using Texas instruments macro model of the op-amp in question (OPA134).

Loading

Generally, distortion measurements are quoted in the form 'x.xx' % at 'f' kHZ at 'n' Watts. This makes it necessary to load the test device. A resistive load is all that is required to produce a result; there is little point in designing a loudspeaker model for only one speaker when undoubtedly amplifiers will use different loudspeakers and crossover networks. The most common loudspeaker impedances currently in use are 4Ω, 8Ω and 16Ω. These impedances have been provided with a maximum power dissipation of 50 watts. An open circuit connection has also been provided. This allows the measurement of the signal level at the output of the attenuator. It is critical not to overload the notch filter. Should the voltage difference between the non inverting and inverting inputs of the first stage of the notch be greater than 30 volts, it will be damaged. Furthermore, excessive signals will lead to a larger portion of the output characteristic of the op-amp being traversed. This will make the large signal nonlinearity of the op-amp more significant with respect to the signal that is leaving the amplifier. Clearly, this is a situation that should be avoided; therefore the signal voltage at the input of all filters should be close to one volt.

Mechanical Considerations

To avoid interference due to the high current that is likely to flow in the load, the loading and attenuation circuitry will be housed in a different enclosure from the filters. Two cast aluminium boxes have been used to house the subsystems that have already been constructed.

Planning Changes

In 'Report 1' a time plan was outlined in the form of a Gant chart. It was anticipated that building would start in week seven, and by this time all of the design phase would be complete. On reflection this was a short-sighted idea. In any project there is an element of testing in the design phase and then some redesign and more testing, as simulators can't tell us everything. The design phase was 'complete' midway through week eight, and construction commenced in earnest at the start of week nine. Since then building and testing have been the chief concerns of the author. It is likely that the building phase will not take as long as anticipated. Given that there is an entire week devoted to building and testing in week one of semester two, it is likely that the system will be in its finished state by week three. This leaves three weeks in case of overshoot, as building must be complete by week six. Assuming

this expectation is realised there will be an extra three weeks to draft the final project report and consider the presentation.

References

[1] HP Datasheet for 339A Distortion test set found at "http://www.testequipmentdepot.com"

[2] Jones, M "Valve Amplifiers" Third Edition (Newnes, London, 2003) Pages 143 - 151

[3] P Horowitz & W Hill "The Art of Electronics" Second Edition (Cambridge University Press, Cambridge, 1989) Pages 296 – 297

[4] Van Valkenburg M "Analog Filter Design" (Holt Rienhart and Winston, 1982) Page 262