Distortion Measurement

Bridge-Null Method

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Summary

The requirements for the measurement of non-linear distortion in audio amplifier are examined. The limitations of conventional measurement methods are discussed and several improved test methods described in which complex test signals can be used. A detailed examination is made of one method in which the input and output signals of an amplifier are combined in such a way that the undistorted component of the output signal is cancelled by the input signal and the distortion component isolated. The existing literature concerning this method is surveyed. The sources of error then using this technique are examined. These include phase end gain errors at high and low frequencies, earth connections arrangements and the effects of complex loads. Methods of reducing the errors are explained and a practical measuring instrument circuit designed.

The instrument has a differential input so that inverting, non-inverting or differential amplifiers can be tested and uses a simple adjustable second order high frequency phase and gain compensation network. The distortion and noise of the instrument are analysed. The practical performance of the instrument is evaluated and its distortion contribution shown to be of an extremely low value. The rejection of test signal distortion is calculated for a particular amplifier test and shown to be more than adequate even when measuring extremely low harmonic distortion. The effectiveness of the load effect compensation arrangement derived is illustrated in tests on a typical class-B power amplifier to detect crossover distortion, transient intermodulation distortion and phase modulation.

Chapter 1. - Distortion Measurement Techniques

1.1)- Requirements for the measurement of distortion in audio frequency power amplifier

In an ideal audio frequency power amplifier the output signal voltage would be identical to the input signal voltage multiplied by a constant. The input impedance would be infinite so that there would be no loading effect on the signal source, and the output impedance would be zero so that the output voltage would be independent of the load used. In the design of a practical amplifier it is necessary to decide to that extent the requirements can be reduced and to be able to measure the deviation from the ideal when using input signals having similar characteristics to those for which the amplifier is designed. Some types of distortion are more audible than others and a test method is required in which the resulting distortion specification gives a good indication of how the amplifier will sound in practical use.

The input and output impedance requirements depend on the signal source impedance. Z_S, and load impedance, Z_L, respectively. For a given degree of non-linearity in the input impedance the distortion introduced will be minimized by the use of a low value of Z_S so distortion should be measured using the largest value of Z_S to be used in the intended application of the amplifier to give a worst-case figure. Comparison of the results with those obtained using a much lower value of Z_S will indicate the relative significance of this source of distortion.

For a given degree of non-linearity in the output impedance the resulting distortion will be dependent on the load used. Although many loudspeakers are specified as having an impedance of 8Ω there is usually a large variations in a given loudspeakers from 5 to 40Ω [1,2] and significant reactive components. The distortion produced by an amplifier with a loudspeaker load will therefore be different from that with an 8Ω resistor and the total harmonic distortion figure is typically 10dB higher over much of the frequency range [3] and in some cases considerably worse [4]. If possible distortion measurements should therefore be carried out using various typical loudspeakers loads to give an indication of the performance under normal operating conditions. Comparison with measurements made with no load connected will give an indication of the relative significance of distortion due to non-linearity in the output impedance. It should be noted however that in class-B amplifiers crossover distortion can occur at low output currents and may sometimes be more significant when using high load impedances if operation is then confined to the non-linear crossover region.

A distortion component would still be present even with a zero source impedance and no load. There are therefore three separate components of distortion to be considered and for a general-purpose amplifier in which different source and load impedances may be used it would be an advantage to obtain measurements of the three components separately or at least to present the total distortion figure as a function of source and load impedances.

1.2)- Distortion specification and measurement methods

It is convenient to specify the distortion of an amplifier in a form which enables comparison with other amplifiers or with some known standard requirements, e.g. the DIN 45500 [5] standard for audio power amplifiers specifies a maximum RMS total harmonic distortion of 1% for sine wave signals from 40Hz to 12.5kHz for any power output between 100mW and 10W outputs and intermodulation distortion no more than 3% at 10W output for inputs of 250Hz and 8kHz with an amplitude ratio of 4:1. Other minimum standards for high quality sound reproduction have been proposed [6,7,8] which demand much lower levels of distortion, although there is some disagreement about the audibility of various quantities and types of distortion in amplifier when used for the reproduction of music [8 to 13].

Total harmonic distortion (THD) measurements can be made using a distortion factor meter, which is basically a variable frequency notch filter. A low distortion sine wave is applied to the input of the amplifier being tested and the output fed to the input of the notch filter, which is adjusted to eliminate the frequency of the test signal. The remaining signal contains distortion and noise from the amplifier and can be measured on a RMS meter to give a percentage THD. This signal will contain components over a wide frequency range and as the RMS value will depend on the bandwidth of the measuring system a band pass filter is generally incorporated in the instrument. The bandwidth used must be stated as part of the distortion specification.

The limitations of this technique are:

1)- The inability of the instrument to distinguish between the distortion added by the amplifier and that already present in the test signal or added by the input stages of the instrument itself, A signal generator with very low distortion must be used.

2)- The inclusion of amplifier noise over the wide bandwidth which may be needed to include all significant distortion components. In the extreme cases the THD measured may be predominantly noise. The standard method of THD specification makes no allowance for this possibility (14) and can therefore be misleading.

3)- For frequencies above 10kHz the harmonics will be outside the audible frequency range and therefore knowledge of their amplitudes can only give an indirect indication of the importance of high frequency non-linearity. When reproducing complex audio signals any audible distortion due to high frequency signals will consist of intermodulation products, which are only produced when two or more frequencies are present in the input signal.

4)- As a single frequency input signal must be used the method can detect static distortion but not the dynamic distortion, which is caused by variations in the nature of the input signal (15).

The advantages of the technique are:

1)- The output from the distortion factor meter can be displayed on an oscilloscope. This gives additional information concerning the nature of the distortion, particularly if a dual trace oscilloscope is used to display both amplifier output and the distortion waveform since the phase relationships of the distortion components are changed very little and therefore the waveform indicates the error voltage present at any position on the output signal.

2)- A single THD measurement can be carried out quickly as the only critical adjustment is that of the notch filter to give maximum attenuation at the input frequency.

As an example of the standard of performance possible with this type of instrument, the Radford Series 3 Distortion Measuring Set (16) has a measurement frequency range of 5Hz to 50kHz and is capable of measuring distortion products as low as 0.001%.

An alternative method, which to some extent avoids the disadvantages of the distortion factor meter is the use of a wave analyser. This instrument is basically a band pass filter with a very narrow bandwidth, typically 5HZ, and high rejection of frequencies outside this band. The output of the filter is measured on a meter. The centre frequency of the band can be varied so that the amplitudes of the individual frequency components of a signal can be measured provided they are not to close in frequency for the filter to separate then. The wave analyser is most useful for the measurement of intermodulation distortion (IMD) in which input frequencies f_1 and f_2 generate distortion at frequencies $nf_1 + mf_2$ where n and m are positive or negative integers.

The advantages of such measurements are:

1)- Provided the two signal sources used are connected in such a way as to avoid any significant interaction the signal applied to the amplifier will consist only of frequencies f_1 and f_2, their individual harmonics produced by the generators and noise. Provided one test frequency is not an integer multiple of the other the IMD products will not be at the same frequencies as the harmonics and therefore the products detected at frequencies $nf_1 + mf_2$ will be entirely due to the non-linearity of the amplifier. Extremely low harmonic distortion signal generators are therefore not essential.

2)- The contribution of noise to the measurement will be low due to the narrow bandwidth of the wave analyser. For a white noise interfering signal the RMS noise voltage is proportional to the square root of the bandwidth. i.e. For total RMS noise voltage v_n in a 20kHz bandwidth an instrument with a 5Hz bandwidth will detect only

$$\left(\frac{5}{2\times10^4}\right)^{\frac{1}{2}} v_n = 0.016\ v_n$$

The noise introduced by the measuring instrument in the stages after the filter will not be reduced and this, together with the finite attenuation of frequencies outside the filter bandwidth and distortion introduced by the input stages of the instrument will limit the lowest level of distortion, which can be detected.

3)- High frequency test signal IMD products can be measured within the audio frequency range and are therefore directly related to the audible effects.

Disadvantages are:

1)- The distortion waveform is not obtained; e.g. Crossover distortion in class-B amplifier generally consists of short spikes on the distortion waveform where the amplifier output current passes through zero. These can be clearly seen an a distortion factor meter output displayed on an oscilloscope but will only be observed as high order harmonic or intermodulation products in wave analyser tests with no direct indication of their source.

2)- Even with test signals consisting of only two frequencies a large number of IMD products may be produced which must be measured individually. This problem can however be reduced by the use of a spectrum analyser in which the centre frequency of the filter is automatically swept through a given range. The various frequency components can be displayed on a CRT as a graph of amplitude against frequency.

Typical instruments are: [(17)]

The Marconi Wave Analyser type TF455D/1, which has a bandwidth of 4Hz with a response 40dB down at 30Hz off tune. Distortion introduced by the instrument is at least 70dB below the signal level.

The Marconi TF2370 spectrum analyser has an amplitude display range of 100dB and a minimum filter bandwidth of 5Hz with 70dB attenuation at 100Hz off tune.

There are two widely used standard IMD test methods. These are the SMPTE and CCIF methods [(18)]. The SMPTE method uses a low and a high frequency test signal e.g. 200Hz and 8kHz and the CCIF method uses two closely spaced high frequency signals e.g. 14kHz and 15kHz.

It is possible to plot swept IMD curves, e.g. If two test frequencies f_1 and f_2 are separated by a constant frequency f_0 and swept through a given frequency range then the IMD product $f_1 - f_2$ is a constant frequency f_0. The amplitude of this product can be measured and displayed as a function of one of the test frequencies. Such swept IMD plots can reveal distortion problems, which occur in narrow frequency ranges and may not show up on conventional IMD or THD tests at a limited selection of test frequencies [(4)].

1.3)- Improved Distortion Specification Methods

The distortion specifications obtained for several operational amplifiers by a variety of test methods have been compared by Leinonen, Otala and Curl [(18)]. The THD at 1kHz and the CCIF and SMPTE IMD specifications were compared with the “noise-transfer” and “dynamic IMD” tests.

The noise-transfer method uses an input signal consisting of filtered noise having a frequency range of 11 kHz to 20 kHz. The intermodulation products of this noise in the amplifier output in the 0 to 10 kHz range are measured and the ratio of the RMS value to that of the input signal calculated.

The dynamic IMD test uses a low-pass filtered square wave and a high frequency sine wave with peak amplitude ratio of 4:1. A sine wave of 15 kHz was used and a 3.18kHz square wave chosen for good separation between the sine wave, its harmonics and the intermodulation products. The total RMS IMD voltage is expressed as a percentage of the RMS amplitude of the 15 kHz sine wave. The figure includes the dynamic IMD components caused by the rise-time portion of the square wave.

The comparison of the methods showed that the CCIF, noise-transfer and dynamic IMD figures gave good agreement in the order in which the tested amplifiers were places. i.e. the amplifiers with the worst figures in the one test method were also the worst in the others. The THD and SMPTE IMD tests gave very low distortion levels in all the amplifiers tested. The same peak amplitude of output signal was used for each test method. e.g. The μA709 operational amplifier with unit gain compensation and 20dB gain in the non-inverting mode gave less than 0.02% THD at 1 KHz and 0.11% SMPTE IMD while the dynamic IMD using a 30 KHz low-pass filtered square wave gave a figure of 62%, the CCIF figure was 26% and the noise-transfer was 50%. Clearly the 1 KHz THD and the SMPTE methods do not give a good indication of the levels of distortion under other signal conditions.

All of the test methods so far mentioned apart from the noise-transfer method have the disadvantage that the ratios of peak to RMS amplitudes of the test signals used are small and therefore the majority of high quality loudspeakers cannot be used as loads during high peak amplitude tests as they could be damaged by the heat generated. The maximum power available from an amplifier is in practice only used during short transient peaks of audio signals, which loudspeakers can be designed to handle safely. During typical orchestral music signals the amplitude is 20dB above the long term RMS amplitude for only 0.01% of the duration and 10dB above for 1% of the time [(6)]. If the +20dB level is generated by 100 W from the amplifier then it can be seen that the power level to be handled by the loudspeaker will be less than 10 W for 99% of the time and the long-term average power level will be 1 W.

It is possible to construct networks of passive components with some of the impedance characteristics of typical loudspeakers [(3,4)] but these will only be approximated equivalents due to the extreme complexity of even the simplest loudspeaker. A moving coil loudspeaker has an equivalent circuit, which consists not only of the resistance and inductance of the coil but also of components due to the motion of the coil in the magnetic field. This is influenced by external factors such as resonance, reverberation and reflections of acoustic energy [(19)]. A distortion measurement method in which a signal with a high peak to RMS amplitude ratio can be used will therefore have some advantage in evaluating amplifier performance under normal operating conditions with a loudspeaker load.

Several systems have been developed by the BBC Research Department. In one system two or more frequency bands of white noise are used and the IMD product generated in other regions of the audio band are measured [(20)]. The experiments carried out indicated that these measurements correlated much better with subjective assessment of distortion than did standard THD figures. As with the noise-transfer system described earlier the signal does not cover the whole band simultaneously and an alternative method was developed [(21)] which used a pseudo-random binary sequence as a test signal. This gives components at equal frequency intervals. As the IMD components produced by this signal would occur at the same frequencies as components of the test signal the whole signal is first shifted by a constant frequency before application to the amplifier and then shifted back after passage through the amplifier. The test signal components are then eliminated by a comb filter to leave only distortion components. The amplitudes of test signal and distortion were measured on a standard BBC peak-programme meter and the ratio in dB described as the “noise-separation” figure, N. A test signal with components at 150Hz intervals shifted by –33Hz was used for good agreement between N ands= subjective assessment. The value of N is related to sine wave harmonic distortion for a system with an n^{th} order non-linearity by the equation:

$$N\ (dB) = |D| - 3(n-1) - 10\log n!$$

Where D is the level of the n^{th} harmonic in dB relative to the fundamental. D and N are measured at the same test power level. The equation shows that such tests give preference to higher order non-linearities compared to THD tests at a given power level. An increase in the importance of higher order non-linearities has also been found in subjective evaluations (8). The subjective agreement was said to improve with increased ratio of peak to RMS amplitudes of the test signal. This is to be expected considering the high ratio found in typical music signals mentioned earlier.

A further method is the direct comparison of the input and output signals. If the output signal of an amplifier is attenuated to the level of the input signal and then in a test instrument added to the input signal in the case of an inverting amplifier or subtracted for a non-inverting amplifier then the remaining signal will consist of the distortion and noise added by the amplifier. The phase and gain variations in the amplifier at the high and low frequency extremes can be compensated for by similar characteristics being applied to the signal used for comparison. Any type of test signal can be used provided the phase and gain characteristics of the amplifier can be duplicated with accuracy over the frequency range covered by the signal. The direct comparison method (sometimes described as the null or bridge method) will now be investigated and some of the practical uses demonstrated.

Chapter 2. - Literature Survey

A number of articles have been written during the past 25 years concerning the use of the direct comparison test method, and the following is a summary of the significant contributions, which have been found.

REF. 22.
E. R. Wigan, “Diagnosis of Distortion” Wireless World, 59, pages 261-266. June 1953.

A system is described in which distortion is extracted by direct comparison method using a single sine wave test signal. A simple addition of input and attenuated output signals is obtained using a transformer to produce the phase reversal needed to test non-inverting amplifiers. For testing systems, which generate a dominant harmonic a system is illustrated which can cancel this component of distortion by the use of a variable phase oscillator set to the frequency of the harmonic and locked in phase with the test signal generator. The oscillator output is adjusted in amplitude and phase so that on addition to the distortion signal the dominant harmonic is cancelled. Other distortion components can then be observed more clearly. The display method used gives an indication of the harmonic structure of the distortion. The test signal is applied to the X-amplifier of an oscilloscope via a phase shifter and the distortion signal applied to the Y-amplifier giving a trace, which is essentially a Lissajous figure. Many types of distortion can be diagnosed by an interpretation of the trace and several examples are given.

REF. 23.
D. C. Pressey, “Measuring non-linearity.” Wireless World, 60, pages 60-62, February 1954.
(+Correction: Wireless World, March 1954, page128)

The method is formulated mathematically:
If an input signal V_i produces an output V_o such that:

$$Vo = aV_i + bV_i^2 + cV_i^3 + dV_i^4 + \ldots\ldots.$$

Then adding $-aV_i$ gives the error voltage

$$V_e - bV_i^2 + cV_i^3 \ldots\ldots..$$

The distortion can be expressed as a percentage like the equation below, or as a percentage of the output voltage.

$$N = \frac{V_e}{aV_i} \times 100\%$$

A simple circuit arrangement is shown with only an attenuator and a summing amplifier, no phase compensation being used. Only inverting amplifier can be tested with this simple arrangement. A display similar to that used in REF 22 is employed.

REF. 24.
M.G. Scroggie, Radio and Electronic Laboratory Handbook.
Iliffe Books Ltd, 1961. (7th Edition) Chapter 11.

The direct comparison method used in REF’s 22 & 23 is briefly described. If the same display method is used, but with suitable phase correction to give a single line trace, and then it is possible to calculate the amplitudes of the individual harmonics for a sine wave test signal. Formulae are given for the first seven harmonics in terms of measurements on the display, but it is stated that the results are very inaccurate. The use of a wave analyser to analyse the distortion waveform is far better.

REF. 25.

F. Jones, "Dynamic testing of audio amplifiers."
Hi-Fi News & Record Review, November 1970, pages 1655 - 1657.

The use of the direct comparison method of distortion extraction to compare the performance of different amplifier is explained. The input signal to the amplifier being tested and the signal to be compared with the amplifier output are obtained from two separated secondary windings on the output transformer of a power amplifier, which is used to amplify the test signal. By first listening to the output of the tested amplifier at normal listening level and then listening to the distortion signal alone without change in its level the seriousness of the distortion produced under normal operating conditions is assessed. For a typical high quality amplifier (the Quad 303) the distortion signal by itself was said to be inaudible.

The test method had been used for the previous 25 years by the Acoustical Manufacturing Company Ltd, of Huntingdon, England, the manufactures of the Quad amplifiers. The main limitation was found to be the difficulty in achieving cancellation when using complex reactive loads. Cancellation of the undistorted signal of about 40dB was obtained with an electrostatic loudspeaker used as load, although this has a relatively simple impedance characteristic being almost a pure capacitance at high frequencies. For testing the QUAD II valve amplifier only a first order phase correction was used, with a circuit arrangement as shown in figure 2.1.

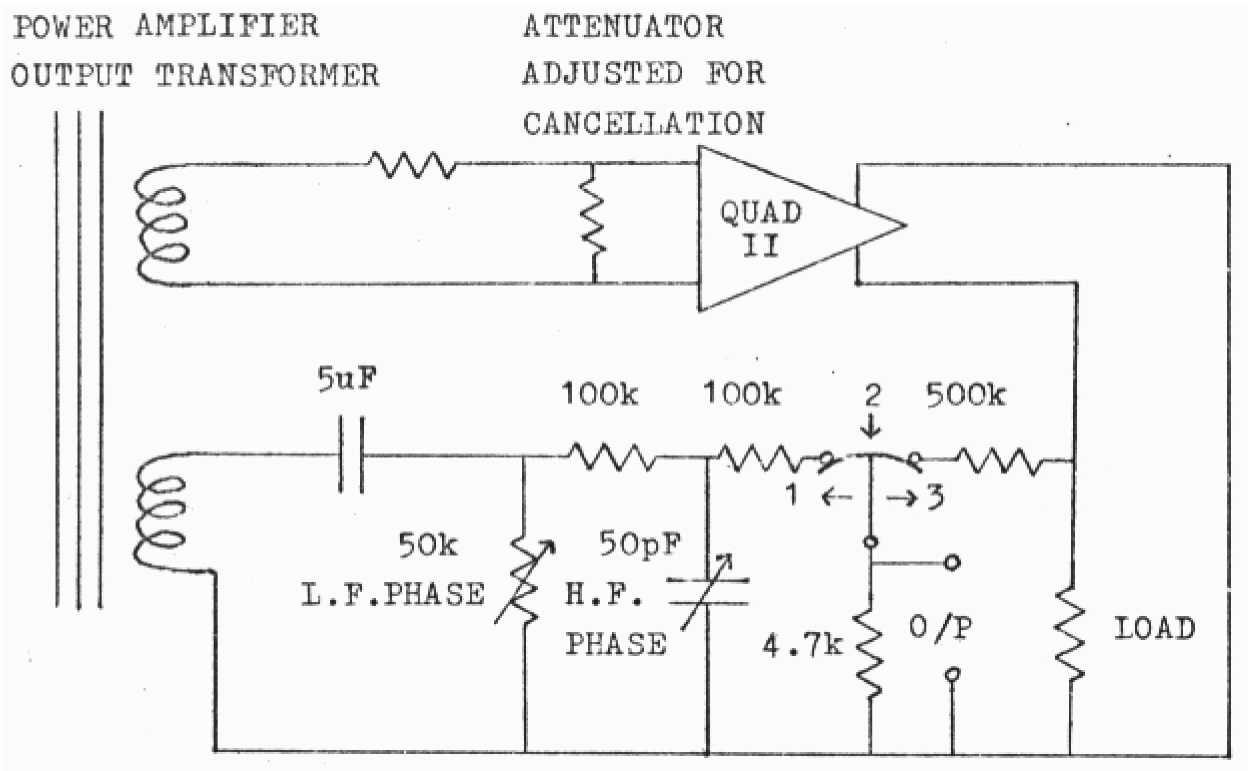


Fig. 2.1 – Bridge Circuit For Quad II Amplifier test

With the switch in position 1 the test signal is obtained at the output. In the position 2 the distortion signal alone is obtained. In position 3 the amplifier output is obtained. By passing the output through a further power amplifier to a loudspeaker the three signals can be therefore each be listened to. By suitable choice of the relative phases of the two transformer winding outputs either inverting or non-inverting amplifiers can be tested.

REF. 26.

P. Blomley, "New approach to class-B amplifier design", Wireless World, March 1971.
Reprinted in "High Fidelity Designs", (I.P.C. Electrical-Electronic Press Ltd.)

A distortion waveform obtained using the null method at 3 KHz at an amplifier output power level of 10W is shown and indicates that measurements of distortion levels less than 0.003% are possible. Problems mentioned are the phasing of the signals and the presence of earth loops and “spurious pick-up difficulties”.

REF. 27.

A. R. Collins, “Testing amplifiers with a bridge”, Audio, March 1972, pages 28 –32.

Some of the limitations of THD and IMD tests are given. The inability to detect dynamic distortions is mentioned and one of these types of distortion explained. This is known as “dynamic crossover distortion” and is due to large amplitude signals causing power dissipation in the output transistors of a class-B amplifier and a consequent rise in the temperature and increase in collector current for a given bias voltage. The quiescent current therefore increases and assuming that it had been adjusted to its optimum values with no signal applied the crossover distortion will get worse. Crossover distortion tends to be more significant for small amplitude signals as the output stage is then operating mainly in its non-linear crossover region. Therefore a high amplitude signal followed by a low amplitude signal will cause a rise in crossover distortion in the low amplitude signals until the output stage returns to thermal equilibrium. The effect occurs only for changes in the signal characteristics, hence the name “dynamic” crossover distortion.

The same test method as that used in REF. 25 is described and a block diagram of a general purpose testing arrangement given as shown in figure 2.2.

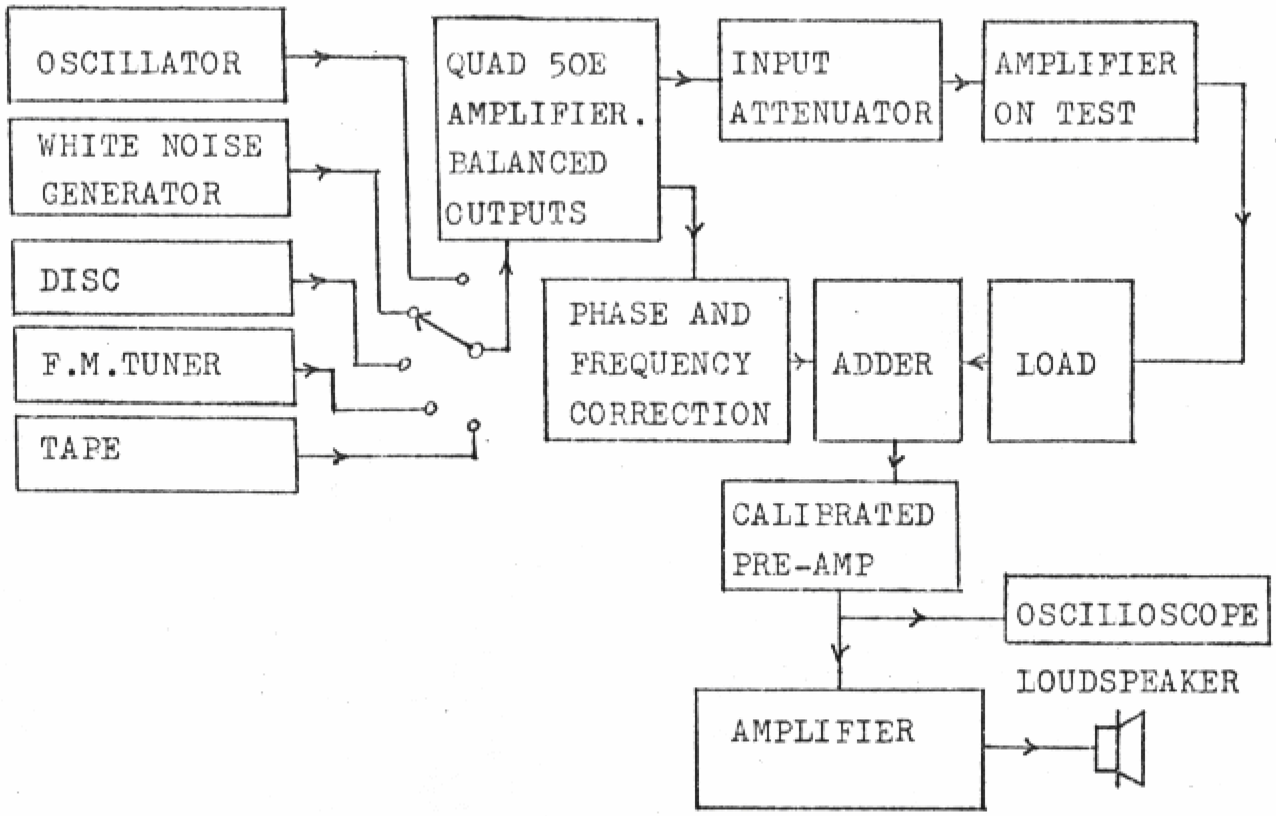


Fig. 2.2 – Complete test equipment

A variety of signal sources can be used and the distortion signal extracted can be displayed on an oscilloscope or amplified and listened to. The Quad 50E amplifier shown has an output transformer with two secondary windings as in REF. 25.

REF. 28.
Jan Lohstroh and Matti Otala, "An audio power amplifier for ultimate quality requirements." IEEE transactions on Audio and Electroacoustics, Vol.AU-21, No. 6, December 1973, pages 545 - 551.

A practical circuit is given for inverting amplifier test as shown in figure 2.3. The component values are chosen for use with the amplifier design described. High frequency amplitude and phase compensation is provided by a combination of a 150ns delay a first order RC filter. Distortion less than 0.01% was observed. The majority of the output signal at lower distortion level was due to incomplete phase compensation. Test signals used: Sinusoidal, Noise and Music signal.

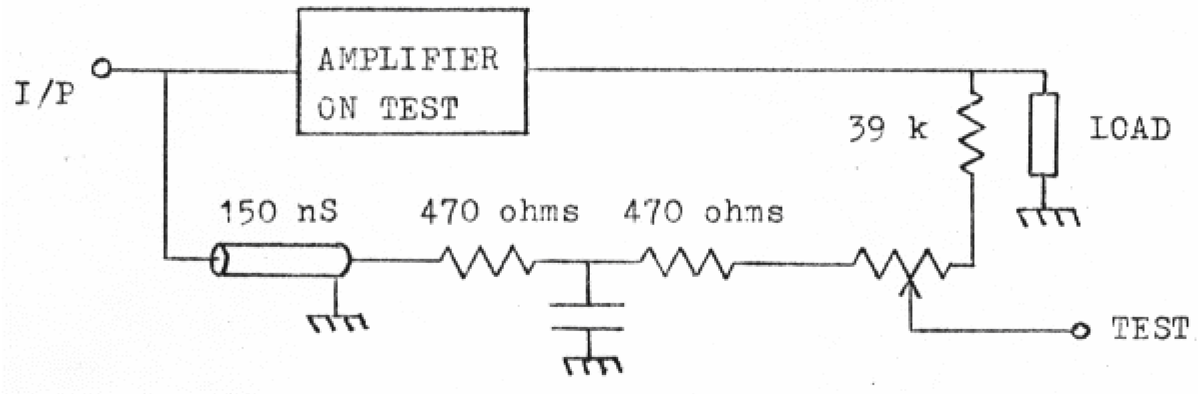


Fig 2.3 – Test circuit for an inverting amplifier

REF. 29
P.J. Baxandall, "Audible amplifier distortion is not a mystery"
Wireless World, November 1977, pages 63 – 66

A practical circuit for testing inverting amplifiers is given as shown in figure 2.4. Amplitude and phase compensation are provided at both high and low frequencies by second order RC filters.

Several uses of this system are described in connection with an attempt to assess the subjective audibility of the levels of distortion produced by modern high quality amplifiers.

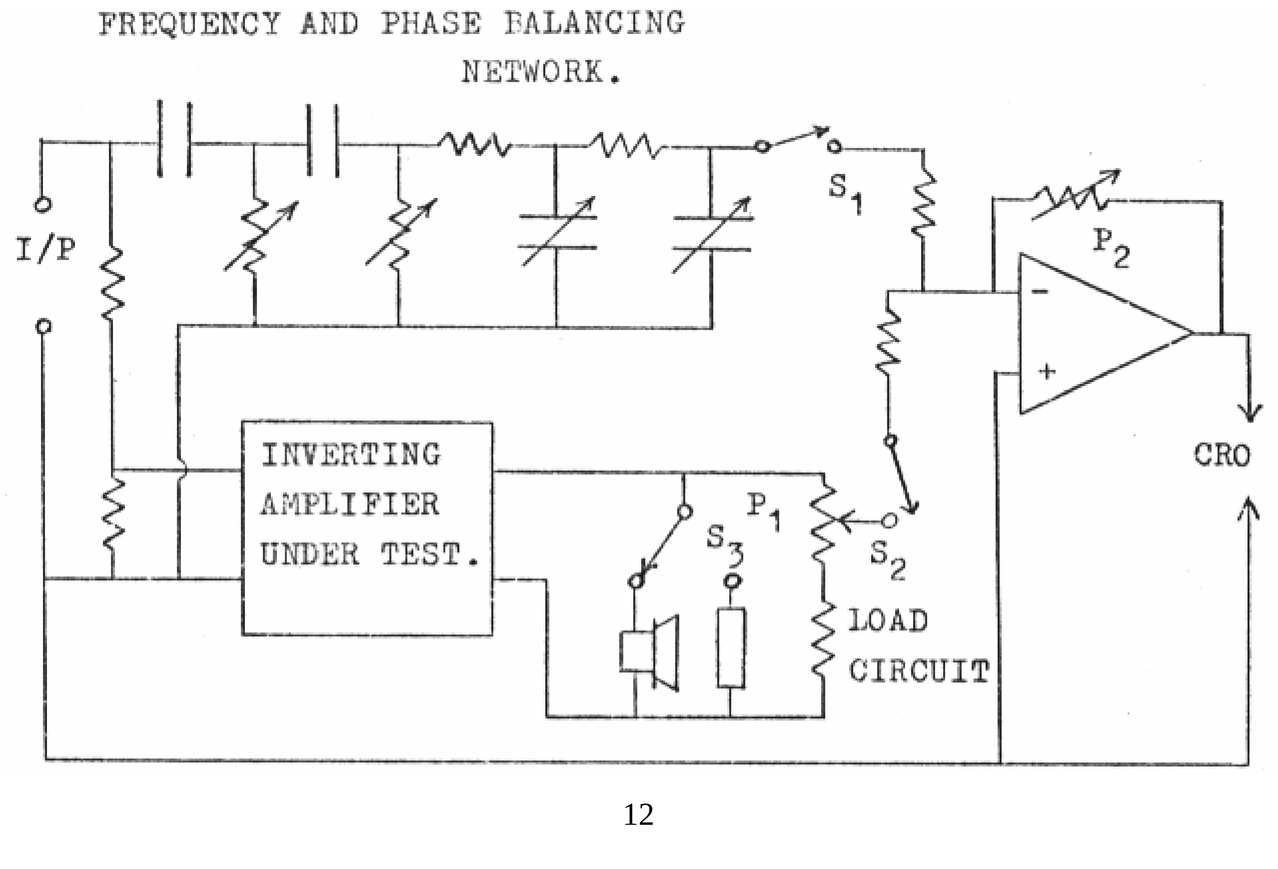


Fig 2.4 – A test circuit for an inverting amplifier

With S_1 and S_2 both closed the undistorted signal is cancelled by adjusting of P_1 and the phase compensation components, and the distortion obtained at the output of the test circuit. With S_2 closed and S_1 open the output of the amplifier under test is obtained. With S_1 closed and S_2 open the test signal is obtained with frequency and phase characteristics similar to those of the amplifier under test applied. P_2 determines the gain of the instrument and can be adjusted to give suitable gain for the display on an oscilloscope of whichever of the signals is selected.

Chapter 3. - Sources of error and their reduction.

3.1)- Phase compensation

The low frequency amplitude and phase response of an amplifier can be calculated from knowledge of the component values used in the amplifier, and a suitable network constructed to give similar characteristics. Alternatively it may be possible to modify the amplifier or choose the points in the circuit from which input and output waveforms are taken to minimise low frequency effects. e.g. consider the arrangement in figure 3.1.

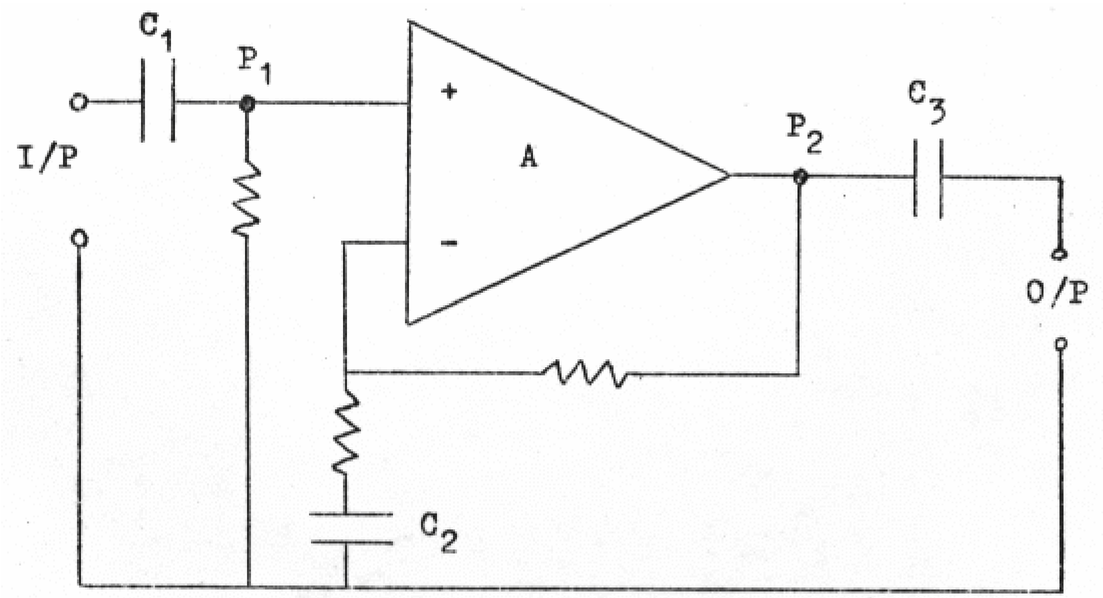


Fig. 3.1 – Low frequency response determining components of a typical amplifier

If the internal circuit of amplifier A is direct-coupled so that it has no significant low frequency gain or phase variations, then by taking the input and output waveforms from P_1 and P_2 instead of from the I/P and O/P terminals the remaining source of low frequency phase shift will be C_2. In many amplifiers C_2 is not essential and is only used to increase the DC negative feedback and thereby reduce the output offset voltage. During tests it may be possible to short out C2 without interfering seriously with the operation of the amplifier. There will be a small effect due to C3 at low frequencies as it reduces the output current and therefore reduces the voltage drop across the output impedance of the amplifier. This effect can be compensated for as described later in section 3.4 (page 38). Such methods are more suitable to the design stage than to the testing of complete amplifiers.

At high frequencies compensation is applied to an amplifier to maintain stability of the negative feedback loop. The most common method of maintaining stability is to use a 6dB/octave fall in loop response above a certain frequency giving a phase lag reaching 90 degrees. Provided the gain round the feedback loop falls to unity before other phase lags introduce a further 90 degree the amplifier will be stable. The open loop response is then predominantly first order. For measurements of the highest accuracy the higher order effects must be taken into account. A closer approximation may therefore be possible using a second order network.

There will also be a time delay between the input and output signals, i.e. an extra phase lag with no associated fall in gain. The relationship between attenuation and phase has been examined by Bode [30] who states that a unique relation exists between any given attenuation characteristic and the minimum phase shift which must be associated with it. Under certain conditions an excess phase lag can exist. One such condition is when the active devices, network elements and wiring cannot be considered to obey a lumped constant analysis and the distributed reactances must be taken into account. The excess phase lag may be regarded as a time delay at a given frequency, but it is not necessarily the same value of time delay at all frequencies.

At a given frequency the response of a second order approximation can be shown to have an amplitude and phase response equal to that of a first order response plus a time delay. Consider the network in the figure 3.2, which is equivalent to that used in REF.29:

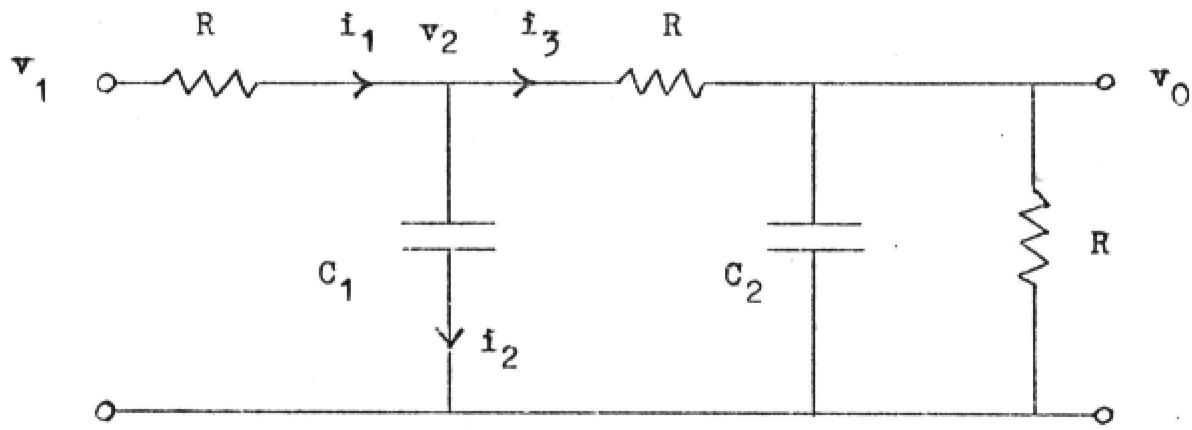


Fig. 3.2 – Second order network

$i_1 = (v_1 - v_2) / R$
$i_2 = v_2\, s\, C_1$
$i_3 = (v_2 - v_0) / R$
$i_1 = i_2 + i_3$

It means:

$(v_1 - v_2) / R = v_2\, s\, C_1 + (v_2 - v_0) / R$

$2{*}v_2 /R + v_2\, s\, C_1 = v_1/R + v_0/R$

$v_2 = (v_1 + v_0) / (2 + s\, C_1\, R)$

$v_0 = v_2 / (2 + s\, C_2\, R)$

$$\therefore v_0 = \frac{v_1 + v_0}{(2 + s\, C_1\, R)(2 + s\, C_2\, R)}$$

$$\therefore v_0 = \frac{v_1}{3 + 2\, s\, R\, (C_1 + C_2) + s^2\, R^2\, C_1\, C_2}$$

Replace s by ***jw***:

$$v_0 = \frac{v_1}{3 - w^2\, R^2\, C_1\, C_2 + j\, w\, 2\, R\, (C_1 + C_2)}$$

It means that the second transfer function:

$$G_2(w) = \frac{v_0}{v_1} = \frac{1}{3 - w^2\, R^2\, C_1\, C_2 + j\, w\, 2\, R\, (C_1 + C_2)}$$

At low frequencies the value of the term $w^2R^2C_1C_2$ in the denominator will be small and the response is approximately that of a first order system:

$$G_1(w) = \frac{1}{3 + j\, w\, 2\, R\, (C_1 + C_2)}$$

(Which is produced by the network of figure 3.2 if C_1 is replaced in parallel with C_2).

As ***w*** increases the real part of the denominator of $G_2\,(w)$ decreases and therefore the phase lag given by:

$$\theta_2 = \tan^{-1}\left[\frac{2\, R\, w\, (C_1 + C_2)}{3 - w^2\, R^2\, C_1\, C_2}\right]$$

Increases relative to that of $G_1\,(w)$ given by:

$$\theta_1 = \tan^{-1}\left[\frac{2\, R\, w\, (C_1 + C_2)}{3}\right]$$

While $|G_2\,(w)|$ becomes greater than $|G_1\,(w)|$.

Therefore for a given attenuation the second order filter gives a greater phase lag and if at a given frequency the attenuations are set equal by adjustment of the capacitors then the second order response is equivalent to a combination of the first order response is equivalent to a combination of the first order response and a time delay at that frequency.

Consider the above equation for θ2 is a maximum when the value of $C_1 + C_2 = K$ then value of θ_2 is a maximum when the value of $C_1 * C_2$ is a maximum.

$C_1 * C_2 = C_1\,(K - C_1)$ and is a maximum when:

$$\frac{d}{dC_1}\,(C_1\, K - C_1^{\,2}) = 0$$

Therefore $K - 2C_1 = 0$.

Therefore $K = 2C_1$ and $C_1 = C_2$.

Therefore, the greatest additional time delay occurs when $C_1 = C_2$ while none occurs for $C_1 = 0$.

The compensation methods used in REF 28 and 29 could therefore give identical results at one frequency but would not match exactly throughout an extended frequency range. In practice the response of an amplifier being tested will not be given exactly by either of the two alternatives and therefore the choice between them can be based on other considerations.

The relative simplicity of providing a second order compensation network makes this choice more attractive than the variable time delay solution. Adjustment of C_1 and C2 may be difficult however due to the fact that each affects both amplitude and phase. In the time delay alternative the time delay adjustment changes only the phase relationship and there is therefore less interaction between this adjustment and that of the first order network.

3.2)- Earth connections

The circuits arrangements shown in REF 28 and 29 are based on the assumption that what are to be compared are the input and output voltages relative to the same earth. In practice an amplifier will have separate input and output earth terminals and it cannot be assumed that both will be at the same potential or that any difference will be an undistorted product of the input signal. One of the possible sources of distortion at the output earth terminal is illustrated in figure 3.3.

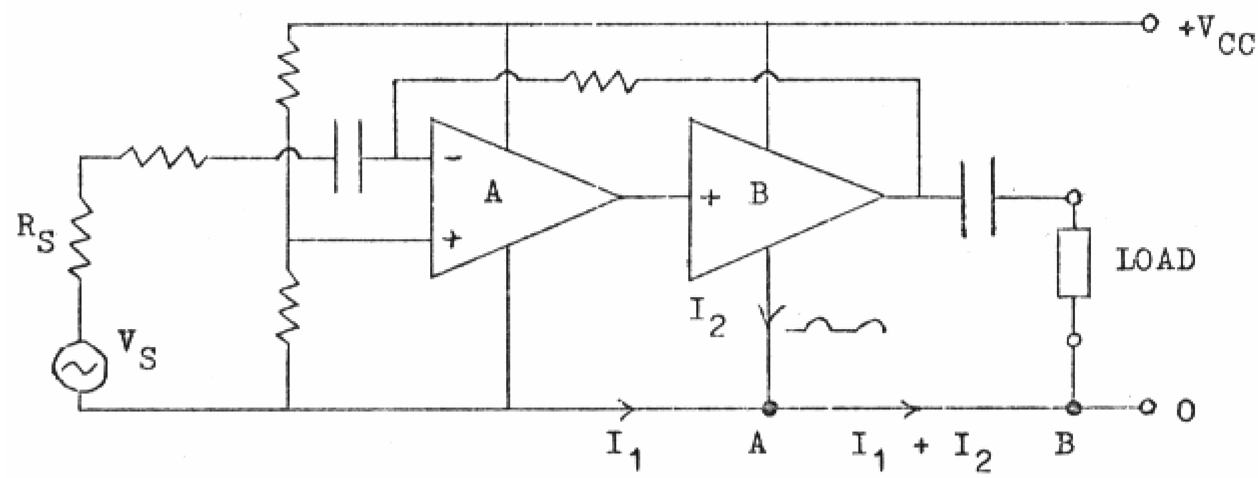


Fig. 3.3 – A distortion generating arrangement

The diagram shows a badly chosen circuit arrangement in which the extremely distorted waveform I_2 in the class-B output stage passes through AB. I_2 has a peak amplitude about equal to the peak amplitude of the current through the load. e.g. For 30W output into 8Ω load the peak current I_P for a sine wave is given by:

$$\text{power} = \frac{I_P^2 R}{2}$$

Therefore, in this case I_P is equal to 2.7A.

The connection from A to B may have significant resistance. For a resistance of 0.1Ω the peak voltage drop due to 2.7A would be 0.27V while the peak voltage across the 8Ω load is 21.6V. The voltage across AB is therefore a significant percentage of the output signal, about 1.2%, although this is entirely distortion. A test circuit in which the input and output voltages relatives to the input earth were compared would not reveal the seriousness of this effect. A circuit arrangement is required in which the potential difference across the output terminals is compared with the potential difference across the input terminals.

The arrangement used in REF 25 (see chapter 2) has the required properties. In this case the input and output signals of the amplifier bring tested are not directly compared. The output signal is instead compared with a signal obtained from the same transformer as the input signal. Whether or not the two signals are sufficiently similar for high accuracy measurement will depend on the properties of the transformer used. It was stated in the reference that there was difficulty in extracting distortion of about 0.1% when using this type of circuit. A simple alternative circuit arrangement was designed and is shown in figure 3.4.

In the circuit shown all voltages are measured relative to the output earth terminal voltage v_6. The input differential signal then becomes $(v_1 - v_6) - (v_2 - v_6) = v_1 - v_2$ and this is compared to $(v_5 - v_6)/A$. Comparison of the two difference signals is therefore achieved as required. There is an additional advantage that similar interference signals picked up by the two connections to the amplifier input will cancel.

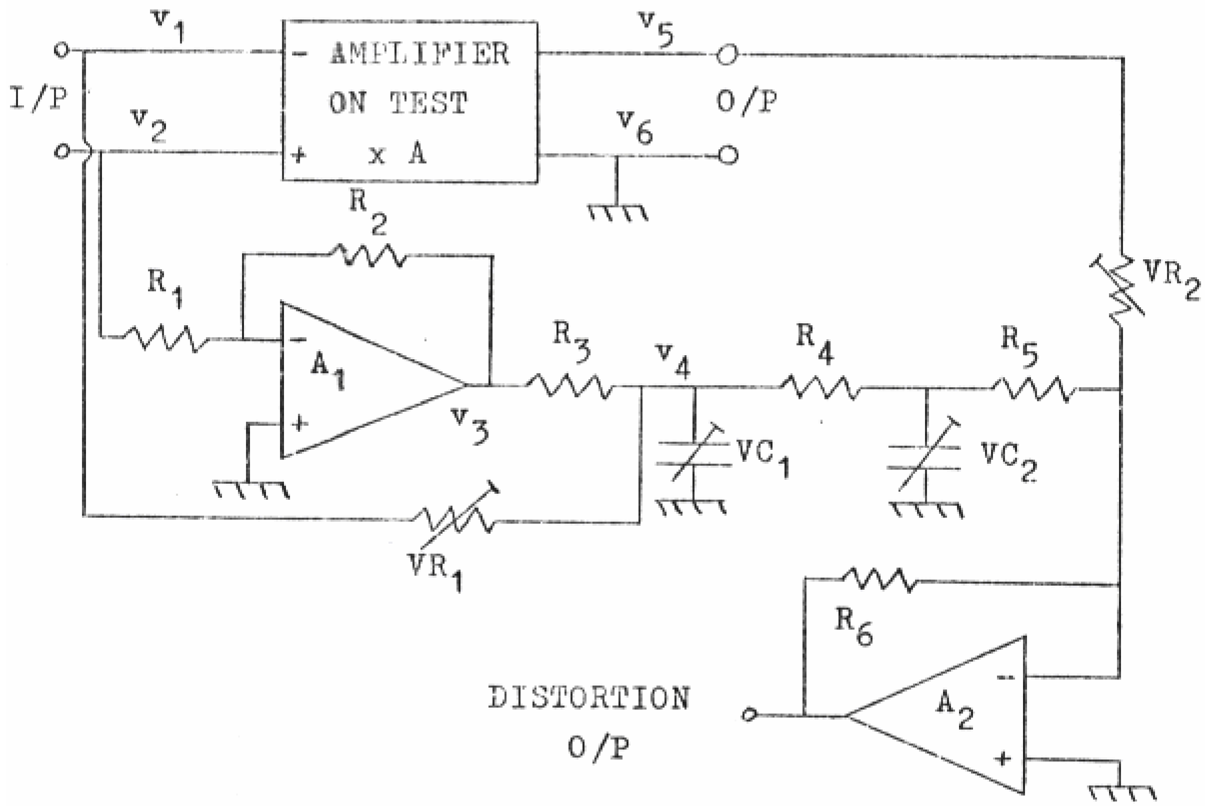


Fig. 3.4 – Distortion measuring circuit with differential input

Differential input or output amplifiers can also be tested with this circuit, but with differential output types amplifier A_1 must handle the difference between an input and an output terminal and may therefore limit the maximum output signal which can be used. The distortion output is also obtained relative to an output terminal voltage v_6 and a measuring instrument with a differential input must be used to measure the distortion output. The circuit of REF. 25 is more suitable for testing differential output amplifiers as there is then no restriction on the output voltage provided the breakdown voltage of the transformer used is not exceeded.

It was decided to use the circuit of figure 3.4 for the practical evaluation of the test method.

A second order high frequency compensation network is shown using VC_1 and VC_2 for adjustment. No low frequency compensation is shown, as this may be unnecessary, as explained earlier. If required such compensation can be included most easily at the input before R_1 and VR_1.

A1 is connected as a unit gain inverting amplifier with $R_1 = R_2$.

Potentiometer VR_1 is used to set v_4 to zero when $v_1 = v_2$.

If A_1 gives a gain of exactly –1 then $v_3 = -v_2$, and $v_4 = 0$ for $v_1 = v_2$ if $VR_1 = R_3$.

The combination of R_1, R_2, R_3, VR_1 and A_1 acts as a differential amplifier. There will be very little distortion added by this circuit when testing inverting amplifiers since then A_1 only amplifies the small difference in potential between the input and output earth terminals while the full input signal is only applied to VR_1. The use of a standard differential amplifier in this position would therefore give an inferior performance unless it was capable of generating as little distortion as a resistor.

3.3)- Resistor Characteristics

A description of the characteristics of the most common type of resistor is given in REF. 31. There are several of the characteristics, which are relevant to the accuracy of distortion measurements using the type of instrument to be described. These are:

1- Voltage coefficient

The resistance of some types of resistor can change significantly as a result of an applied voltage. The voltage coefficients of carbon composition and carbon film resistors are given as typically 3000 and 100 parts per million per volt respectively. It means, for a 1-volt amplitude signal applied the incremental resistance will change by 0.3% and 0.01% respectively. These changes would have a significant effect on the signal cancellation if they occurred in R_3, VR_1, R_4, R_5, or VR_2 in figure 3.4. R_1 and R_2 are of equal value and have equal voltages applied. They give a gain of R_2/R_1 for amplifier A_1 and therefore provided R_1 and R_2 have similar properties they will not introduce large errors.

There are other types of resistor, which have negligible voltage coefficients. These include metal oxide, cermet and metal film types. Wirewound resistors also have very small voltage coefficients but may have significant reactive components depending on the winding technique used in their construction.

2- Thermal effects.

The temperature coefficients of metal oxide, cermet and metal film resistors are given as 50 to 250, 100 and 15 to 100ppm/^{o}C respectively. When a signal is applied across a resistor its temperature changes due to the power dissipated. REF. 31 gives typical graphs of temperature change as a function of power dissipation. The relationship is linear over the temperature range shown with a ½ watt resistor increasing in temperature by 50^{o}C at ½ Watt dissipation. The change is therefore 100^{o}C per Watt. For a temperature coefficient of 100ppm/^{o}C the change is therefore 10^4ppm/W. E.g. A 2KΩ resistor with a 1 V applied dissipates 1/2000W. The incremental resistance will therefore change by 10^4/2000ppm = 0.0005%. For measurements using constant amplitude test signals such changes can be compensated for by adjustment of the potentiometers in figure 3.4. When varying amplitude or very low frequency test signals are used the thermal effects may become significant, so the resistors and potentiometers should be low temperature coefficient types. Metal film and metal oxide fixed resistors and cermet potentiometers are readily available and are therefore to be recommended in this application. The use of resistors with high specified maximum power dissipation would also reduce the thermal effects. The most critical resistance in figure 3.4 is VR_2, which has the full output of the amplifier, applied across it. For testing high power amplifiers therefore particular attention must be paid to the thermal properties of VR_2.

3- Resistor noise

A resistor with current running through produces thermal noise and current noise. The thermal noise voltage is a function of temperature, resistance and bandwidth and is independent of the applied signal except for the effect of the resulting temperature change. Thermal noise will be considered later (Section 4.4). Current noise is a function of the applied signal voltage. The typical total current noises for metal oxide, cermet and metal film resistors are given as 0.03, 0.4 to 1.0 and 0.015µV/V respectively. The use of metal film or metal oxide resistors will therefore give negligible current noise. Even cermet types will give noise less than 0,0001% of the applied voltage.

Measurement of third harmonic distortion generated by solid carbon, carbon film and metal film resistors has been made by Takahisa, Yanagisawa and Shiomi [(32)]. They suggest that in general passive elements have non-linear V-I characteristics due to the presence of electrode contacts and potential barriers in the current path.

The third harmonic voltage was found to be proportional to $J_1^n L/D^m$ where J_1 is the current density of the fundamental (10kHz was used), L is the length and D the thickness of the film. The n is between 2.2 and 2.8 and m = 3.0 for metal film resistor.

For 250KΩ resistor with a 250V signal applied the third harmonic voltage were:

Metal film (½ W)	0.03 to 0.15mV	=	0.12 to 0.6ppm.
Carbon Film (½ W)	1.5 to 4.0mV	=	6.0 to 16ppm.
Solid Carbon (¼ W)	400 to 800mV	=	0.16 to 0.32%.

The fourth, fifth and sixth harmonics are shown for a high distortion carbon film resistor sample as –60dB, -26dB and –74dB respectively relative to the third harmonic.
The inferior performance of carbon resistor is confirmed by these results.

3.4)- Effects of load impedance

When using a load such as a loudspeaker in which the impedance is a complex function of frequency the voltage drop across the amplifier output impedance will also be a complex function of frequency. By adding an impedance Z' as shown in figure 3.5 a voltage can be obtained which is a function of the voltage drop across the amplifier output impedance, Z_o.

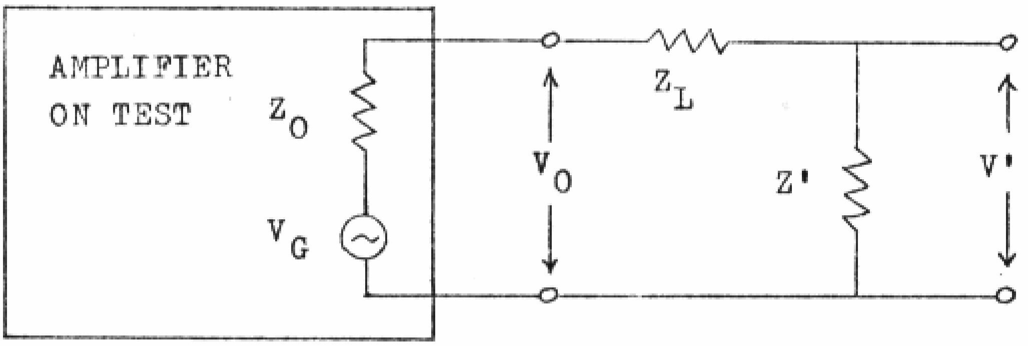


Fig. 3.5 – Extraction of signal for load effect compensation.

$V' = (v_G - v_o) * Z' / Z_o$
Therefore, $v_o + v' * Z_o / Z' = v_G$

By adding $v' * Z_o / Z'$ to v_o the amplifier output voltage can be obtained without its load dependence. i.e. the open circuit output voltage v_G is obtained.

In general Z_o will not be linear. The non-linear component Z_D generates distortion, which it is required to measure and therefore there is no need to compensate for this. If $Z_o = Z_{LIN}$ is the linear component of Z_o then it is Z_{LIN} which must be compensated for, and $v' * Z_{LIN} / Z'$ must be added to v_o to achieve this.

The value of Z_{LIN} is not, however a constant. Suppose the voltage drop across Z_o is given by:
$v = Z_1I + Z_2I^2 + Z_3I^3 + \ldots\ldots$ Equ. 3.1

For a current A sin *wt* the first three terms give:
$v = AZ_1\sin wt + A^2Z_2 \sin^2 wt + A^3Z_3 \sin^3 wt$
$= AZ_1 \sin wt + A^2Z_2(1 - \cos 2wt) + A^3Z_3(3 \sin wt - \sin 3wt)$
i.e. The undistorted component is:

$$(A\, Z_1 + 3\, A^3 Z_3 / 4) \sin wt$$

If Z_{LIN} is defined as the ratio of undistorted voltage to current then $Z_{LIN} = Z_1 + 3\, A^2 Z_3 / 4$. i.e. Z_{LIN} is a function of signal amplitude A.

The compensation can therefore only be carried out at a single signal amplitude using linear components for Z'. At other amplitudes an undistorted signal component will remain. The effectiveness of this method is therefore dependent on the degree and type of non-linearity of Z_O. The even order terms in the power series (equation 3.1) do not contribute an undistorted component and it is the odd order coefficients Z_3, Z_5, Z_7, etc. which are relevant.

A suitable addition to the measurement circuit of figure 3.4 to include the compensation for varying load impedance is shown in figure 3.6. The rest of the circuit is as figure 3.4.

(The above analysis of the effect of non-linearity on the undistorted signal component applies equally to the amplifier non-linearity being measured and when the test signal amplitude is changed will lead to the requirement for readjustment of the test circuit for optimum signal cancellation. What is then being done is in the effect to select the best straight line through the transfer characteristic rather than regard all non-linear power series terms as distortion [23].)

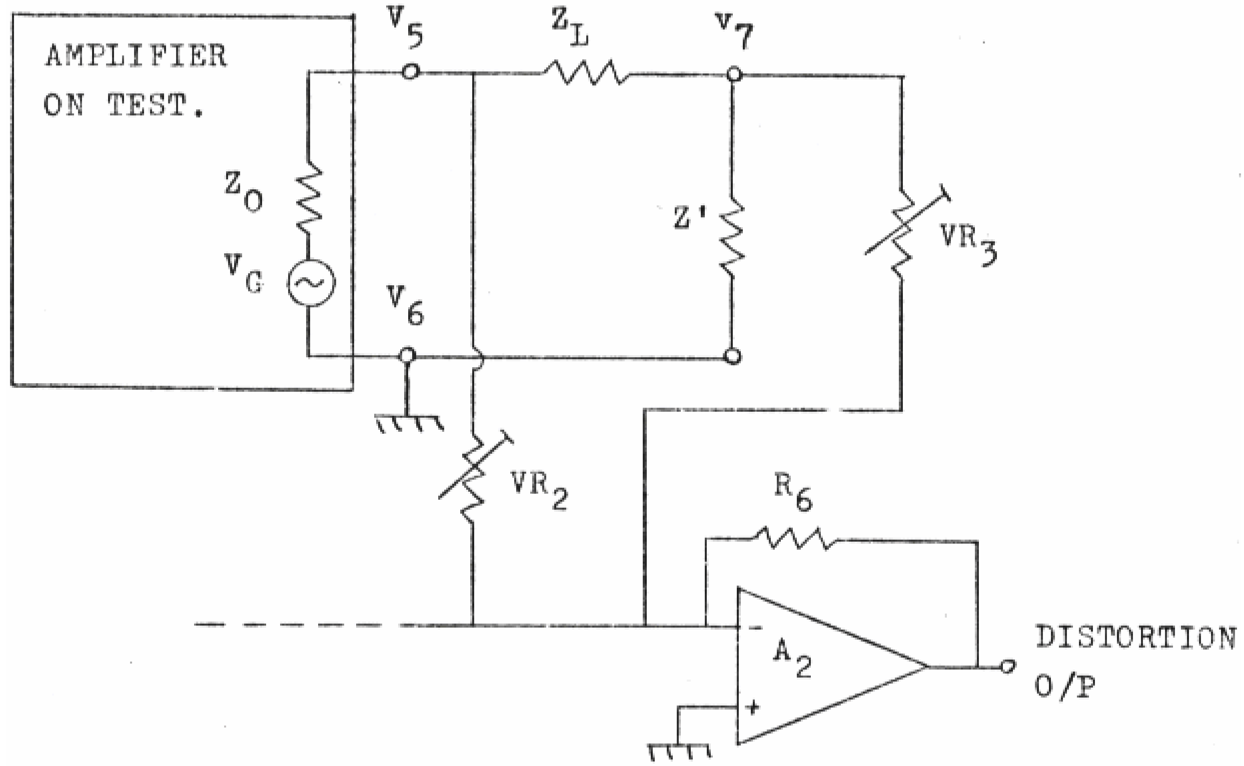


Fig.3.6 – Load effect compensation addition

The output of A2 resulting from v5 and v7 is:

$$R_6\, v_5 / VR_2 + R_6\, v_7 / VR_3$$

It is required to add v7 ZLIN /Z' to v5 and therefore it is required that the output becomes:

$$R_6\, (v_5 + Z_{LIN} * v_7 / Z') / VR_2$$

Therefore, it is required that:

$$\frac{R_6\, Z_{LIN}}{VR_2\, Z'} = \frac{R_6}{VR_3}$$

i.e.

$$\frac{Z'}{VR_3} = \frac{Z_{LIN}}{VR_2}$$

If Z_{LIN} has reactive components and it is required that Z' is small then it will be more convenient to use a small value resistance (say 0.1Ω) for Z' and add the reactive compensation components to VR_3. As VR_3 will be relatively large only small capacitances will generally be needed to give the necessary results.

In general the output impedance of an amplifier will increase at high frequencies due the fall in overall negative feedback. As the output impedance is reduced by a factor (1 – AB) where A is the open loop gain and B the feedback network gain it is possible for the output impedance to have a negative real part if the real part of AB becomes more than +1. This can only occur for phase lags in the gain round the feedback loop of more than 90^o and is therefore unlikely to happen within the audio frequency range when using the usual first order high frequency compensation. If it did occur then it would be necessary to use a different arrangement to that of figure 3.6. e.g. A proportion of v_7 could be taken via an inverting amplifier to the input of A_2 to give a subtraction from v_5 instead of an addition.

Sometimes an amplifier is intentionally designed to have a negative output resistance at low frequencies to give better damping of a loudspeaker resonance. This effect is produced by the use of positive current feedback.

The easiest way to avoid having to compensate for a negative output resistance is to add a small resistor in series with the amplifier output equal to or greater than the largest negative value of the real part of Z_O within the frequency range of interest. This resistor can then be treated as part of Z_O and v_5 obtained from the end of the resistor connected to the load. The effective output resistance is then never negative within the frequency range used.

To set up the circuit balance VR_2 should be adjusted for cancellation of the undistorted signal with the load disconnected (then $v_7 = 0$). Connection of the load will then introduce an additional undistorted signal component due to the voltage drop across Z_O. This can be eliminated at a given signal amplitude (for a sine wave signal) by adjustment of the values of VR_3 and Z'.

In the above analysis it has been assumed that $VR_2 >> Z_L$ and $VR_3 >> Z'$ so that VR_2 and VR_3 do not significantly alter the voltages being measured. This condition will usually be met when measuring power amplifiers.

To use the above compensation method it is convenient to measure Z_O and provided this has only a small non-linear component this can be done by first balancing the circuit of figure 3.4 with no load connected and using a sine wave signal. Addition of a resistive load, R_L, of known value will give a voltage drop across Z_O, which will be amplified by A_2. Provided the voltage drop is significantly greater than the amplifier distortion it can be compared in amplitude and phase with the signal across R_L by displaying both signals on a dual trace oscilloscope. The gain and phase shift of A_2 must be taken into account. Knowing the voltage drop across Z_O for a given voltage across R_L the value of Z_O can be calculated at the frequency used since the current through Z_O is the same as that through R_L. Plotting Z_O against frequency will make it possible to work out the value of Z_O as a function of frequency and derive suitable component value for Z' and VR_3.

Many power amplifiers include components in their output circuit to assist in the high frequency stabilization when using capacitive loads or to protect the output stage against the effects of an inductive load. An output coupling capacitor is also used in some circuits. The effects of all these components can be eliminated by including them as part of the load as shown in figure 3.7.

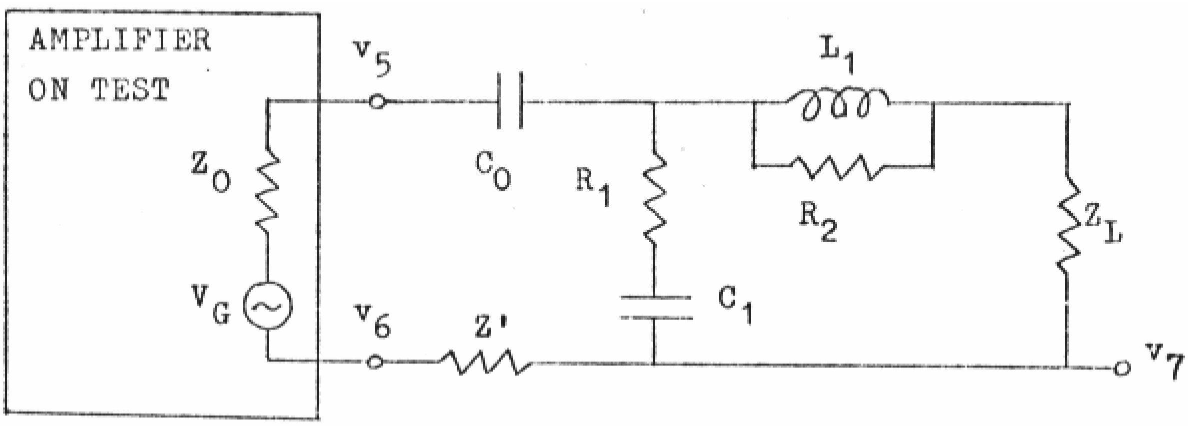


Fig. 3.7 – Output component effect elimination

C_O is the output coupling capacitor.

L_1 and R_2 compensate for capacitive load effects.

C_1 and R_1 compensate for inductive load effects.

The compensation components are generally referred to as Zobel networks.

For changing levels of power output from the amplifier being tested the temperature of the load will change and for a load with a non-zero temperature coefficient its impedance will change. These changes will also be compensated for by the method given.

Output stage protection circuits within the amplifier may cause problems when attempting to drive reactive loads at high power due to the resulting high voltage-current product across the output transistors (1). Although the effect is a function of the load used the compensation method will not eliminate the results and the distortion generated will be observed.

3.5)- Requirement for accurate balance adjustment

Adjustment of VR_1, VR_2 and VR_3 to give cancellation of the undistorted signal will be very critical when measuring low levels of distortion and even multi-turn potentiometers may give insufficiently fine adjustment. One solution is to use a small value fine adjustment potentiometer in series with the main potentiometer. Fixed value resistors generally have superior stability characteristics and it is possible to carry out the balancing using these as follow:

1)- To cancel the undistorted signal component first use a resistance box, RB, as the potentiometer and adjust for balance using a low frequency test signal. (A resistance box may have significant reactive components, which would affect the high frequency balance.)

2)- Take a fixed resistor, R_1, of a slightly larger value than the BR setting and connect in parallel with RB (see figure 3.8A).

3)- Readjust RB for balance to give the value of resistance required in parallel with R_1 and connect a slightly lower value, R_2, in series with RB (figure 3.8B).

4)- Readjust RB for balance to give the value of resistance required in series with R_2. Place a slightly larger value in parallel with RB.

Then continue balancing and adding alternating series and parallel resistors to build up a network of fixed resistors giving a closer and closer approximation to the exact value needed.

To compensate for drift in component values during tests (e.g. due to temperature changes) the network can be terminated with a potentiometer to make fine adjustment possible. (see figure 3.8C)

Figure 3.8A

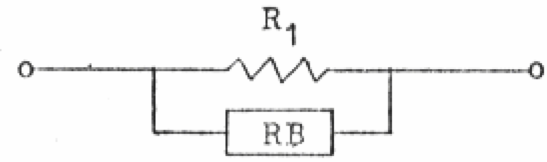


Figure 3.8B

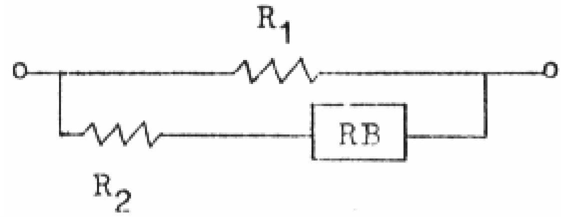


Figure 3.8C

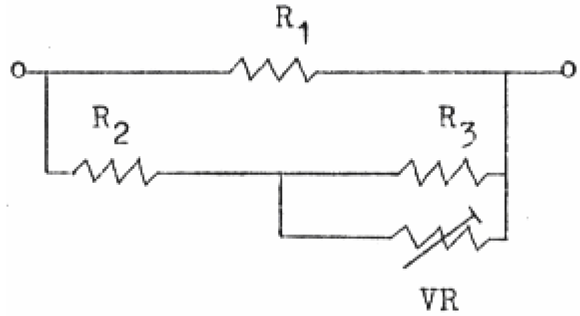


Fig. 3.8 – Successive approximation resistor adjustment

While metal film resistors are recommended for use as R_1 and R_2 in figure 3.8 the other resistors in the network have little effect on the total value and the inferior carbon film types could be used for these.

3.6)- High frequency phase and gain of amplifier A_1

At high frequencies there will be a phase shift and fall in gain associated with A1 (figure 3.4). This will introduce errors in the accuracy of the input difference signal extraction and also in the final signal being used for comparison with the output signal v_5. There are several methods of reducing these errors:

1)- A frequency and phase characteristic can be applied to v_1 and v_5 similar to that applied to v_2 by A_1. This requires the addition and adjustment of two further sets of compensation components.

2)- Feedforward error correction can be applied by an additional inverting amplifier [33]. Distortion, phase shift and gain variations can all be reduced using this method.

3)- Phase compensation of A_1.

The phase and gain errors of A_1 can be reduced in the frequency range of interest by the addition of a capacitor as in figure 3.9.

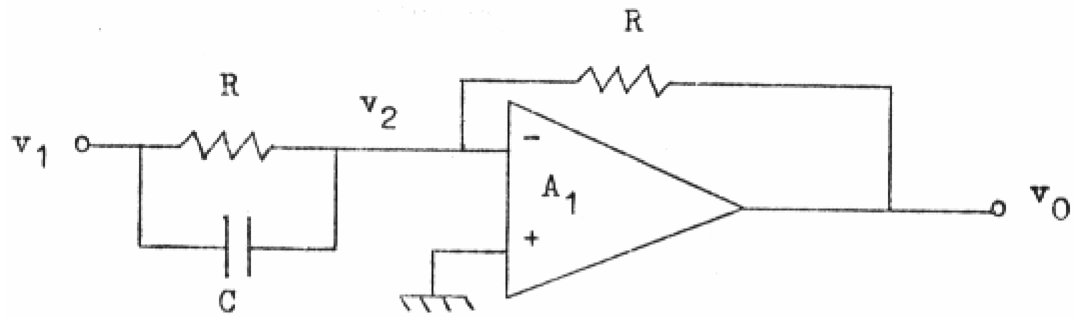


Fig. 3.9 – Compensation of A_1

Let the open loop gain of A_1 be $-A/s$

Summing the currents at the amplifier input:

$$\frac{v_1 - v_2}{R} + (v_1 - v_2)\, sC = \frac{v_2 - v_O}{R} \quad \ldots\ldots\ldots \quad \text{Equ.3.2}$$

$v_O / v_2 = - A / s$
Therefore, $v_2 = - s\, v_O / A$
Let CR = T
Substitution for v2 and CR in equation 3.2 gives:
$v_1 (1 + sT) = - v_O (s (2 + sT) / A + 1)$
Therefore:

$$\frac{-v_O}{v_1} = \frac{1 + sT}{1 + s(2 + sT)/A} = \frac{1 + jwT}{1 - w^2T/A + jw2/A}$$

Phase error $= \tan^{-1} wT - \tan^{-1} (2w /(A - w^2T))$
$= 0$ when $wT = 2w / (A - w^2T)$
i.e. when $w^2T^2 - AT + 2 = 0$

e.g. Choosing T for phase error = 0 at $w = 10^5$ (f = 16kHz)

Let $A = 10^8$

Therefore, $10^{10}T^2 - 10^8T + 2 = 0$

This gives $T = 2.000005 \times 10^{-8}$ or 9.99998×10^{-3}

Using $T = 2 \times 10^{-8}$ (e.g. R = 2KΩ, C = 10pF)

$$\frac{-v_O}{v_1} = \frac{1 + jw \times 2 \times 10^{-8}}{1 + jw \times 2 \times 10^{-8} - w^2 \times 2 \times 10^{-16}}$$

The deviation from unity gain and zero phase shifts is therefore introduced by the term $w^2 \times 2 \times 10^{-16}$ and will increase with increasing frequency. e.g. At $w = 10^5$ radians / sec.:

Phase angle $\theta = -2.3 \times 10^{-7}$ degrees.

Gain = 1.000002

Therefore, Gain error = 0.0002%

Using the more precise value of T would give zero phase error at this frequency but increase the gain error.

While the above calculation indicates that this method is capable of high accuracy it should be noted that several approximations have been made:

a) The input impedance of A_2 has been neglected.
b) The open loop gain of a practical amplifier will be more complicated than the expression – A/s used.
c) The resistors used may have significant reactive components at high frequencies [31].
d) Components used will have a wide tolerance, generally ± 1% or worse.

As the characteristics of a practical circuit will not be accurately predictable it is appropriate to include variable adjustments to compensate for the unknown factors. A variable capacitor can be used for C and adjusted to give good common mode rejection for high frequencies in the differential input stage formed by A_1, R_1, R_2, R_3 and VR_1 (see figure 3.4). If the two inputs are connected together, and a low frequency signal applied to them, then VR_1 can be adjusted to give cancellation of the signal at A_2 output. With a high frequency signal (about 20kHz) applied the variable capacitor connected across R_1 can be adjusted to give the best cancellation of this signal.

The inclusion of C changes the phase versus gain characteristics of the A_1 feedback loop and must be taken into account when calculating the high frequency compensation necessary for stability.

Due to its simplicity it was decided to use this method of compensation in the practical design to be produced.

Chapter 4. - Measuring Instrument Circuit Design

4.1)- Unit Gain Inverting Amplifier, A_1

The requirements for A_1 (Figure 3.4) are that it has low distortion, low noise and give a constant gain throughout the audio frequency range. The maximum signal amplitude to be handled depends on the input signal required by the amplifier being tested to give its maximum output. A value of 1V peak amplitude will be used for the analysis of the distortion of the design to be produced.

There are many small signal amplifiers available in integrated circuit form, designed with emphasis on a variety of parameters such as low frequency gain, noise, bandwidth, distortion, common mode rejection, etc. The performance with regard to distortion of several integrated circuit amplifiers has been compared [34,35]. The 741, LM301 and µA739 were tested with closed loop gains of –3. The µA739 gave the lowest distortion level of 0.013% at an output of 1V RMS at 20kHz, reducing at lower frequencies. Figures given for a simple three transistor discrete component amplifier in REF. 34 showed that under similar conditions a distortion level only a third that of the µA739 was produced. Clearly the use of this integrated circuit would severely limit the usefulness of the instrument for testing low distortion amplifiers. It would be possible to use two amplifiers in a feedforward error correction circuit as in REF. 33 but it was decided to use a discrete component amplifier designed to optimise the parameters of interest in this application.

The design finally produced is shown in figure 4.1.

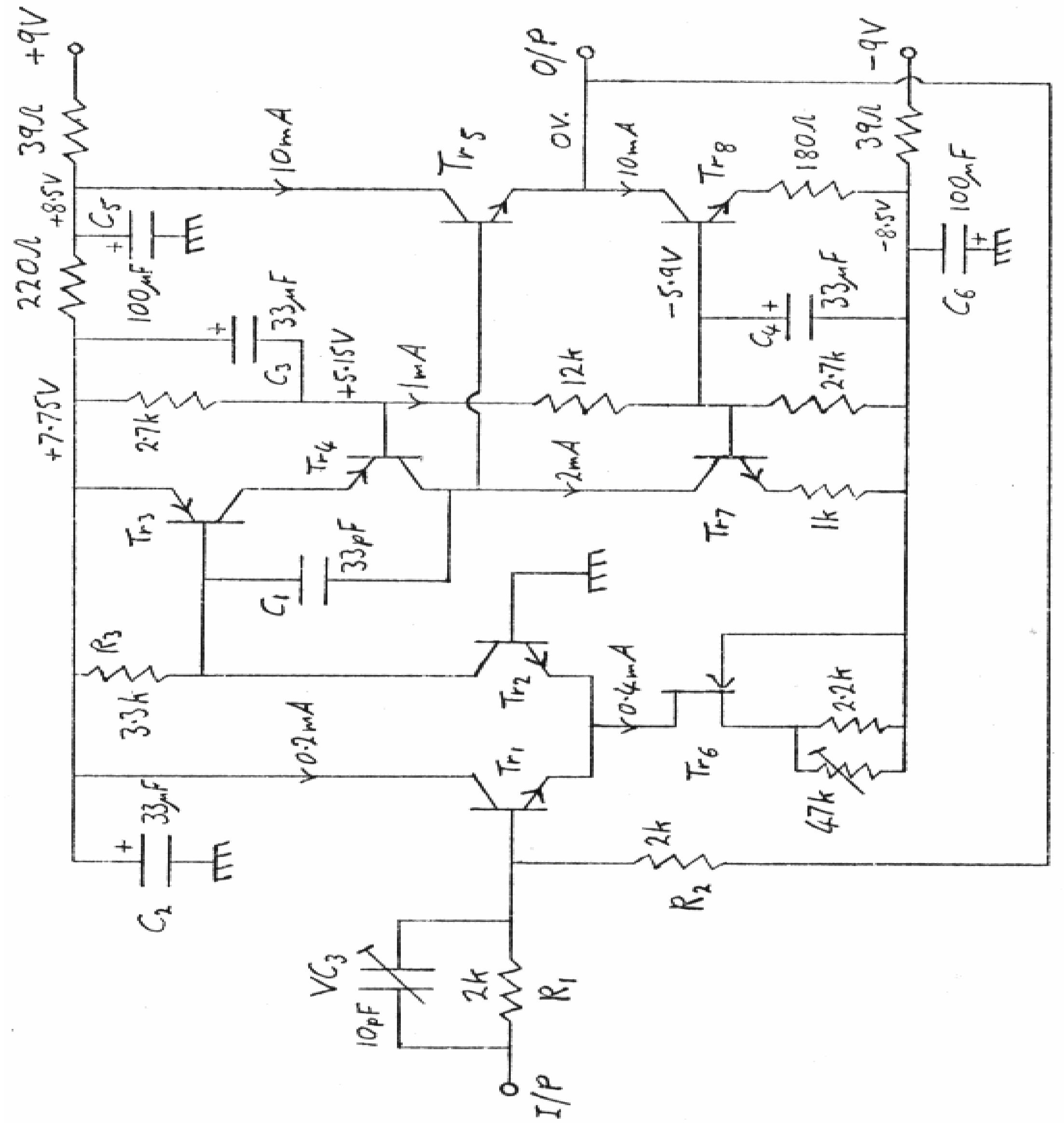


Fig. 4.1 Circuit Diagram of Unity Gain Inverting Amplifier, A1

Tr6, Tr7 and Tr8 act as current sources.

The transistors used should be low noise types and the one chosen were:

NPN: BC169C (h_{FE} = 450 to 900 at I_C = +2mA, V_{CE} = +5V)
PNP: BC259B (h_{FE} = 240 to 500 at I_C = -2mA, V_{CE} = -5V)

VC_3 provides gain and phase corrections as described in Section 3.6, while C_1 gives high frequency negative feedback loop stability. As the open loop gain falls at 6dB/octave at high frequencies the amount of overall negative feedback reduces and distortion is consequently reduced less. Calculation of distortion will therefore be made at the top end of the audio frequency range (20kHz) to give a worst-case figure. Each stage will now be considered separately. Only an approximate analysis will be given, as all the factors affecting performance are not known to a high degree of accuracy.

4.1.1)- Differential Input Stage Analysis

a)- Distortion: The distortion of this type of stage is analysed in REF. 36 where it is shown that minimum distortion occurs for equal collector currents in the two transistors. The distortion is then predominantly third harmonic and is less than 0.005% for a peak sine wave input amplitude of 1mV if the collector currents are matched to within 0.6%. For an output of 1V peak amplitude the gain from Tr_1 base to output of the amplifier must be 1000 at 20kHz if the input signal is to be 1mV at this frequency. This value of gain was chosen for the design. The expression used for the gain in section 3.6 was $-\boldsymbol{A / s}$. For modulus of gain = 1000 at 20kHz this gives:

$$A = 1000 * 2 * \pi * 2 * 10^4 = 1.2 \times 10^8$$

The value of 10^8 used for the calculation of VC_3 was therefore sufficiently accurate as this is a variable component.

b)- Noise: Graphs of noise figure against I_C are given for the BC169C transistor in REF. 37. At 10kHz a noise figure of 0.5dB is obtained at I_C = 0.2mA and a source resistance of 1KΩ. Using $R_1 = R_2$ = 2KΩ to give this source resistance Tr_1 will only increase the effective input noise by 0.5dB while Tr_2 will contribute even less as its equivalent input noise current generator is effectively shorted and only its noise voltage contributes [(38)]. The contribution of the transistors to the noise voltage of the complete circuit will be neglected.

c)- Frequency Response: At I_C = 0.2mA f_T is given [(37)] as 60MHz at V_{CE} = 10V. Tr_2 operates as a common base stage and therefore will have a current gain only a lithe less than unity up to 60MHz. Tr_1 is a common collector stage and has an input impedance which falls at high frequencies due to the presence of input capacitances C_{CB} and C_{BE}. C_{CB} is given as 2.7pF at V_{CB} – 10V and C_{BE} can be calculated from the formula:

$$f_T = \frac{1}{2\pi C_{BE} R_e} \quad \text{where } R_e = \frac{25}{I_E} \text{ ohms. } (I_E \text{ in mA})$$

Taking f_T = 60 MHz, I_E = 0.2mA gives C_{BE} =21pF. The signal voltage v_E at emitter of Tr_1 is half the input voltage at the base, v_B. As a result of this only half the input voltage appear across C_{BE} and it takes a current equal to that which would be taken by ½ C_{BE} connected from base to earth. The total effective input capacitance is therefore:

$C_{IN} = C_{CB} + 0.5C_{BE}$ = 13pF

The effect of this on the overall negative feedback of A_1 can be seen from figure 4.2 where R_1 and R_2 have been replaced by their Thevenin equivalent and the signal source impedance taken as zero.

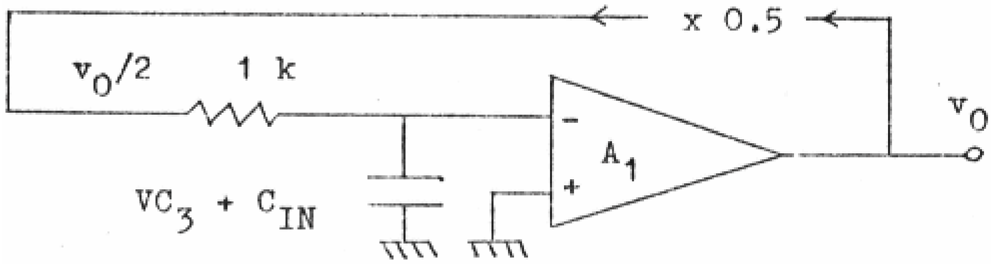


Fig. 4.2 – Effect of Input Capacitance on the Feedback

C_{IN} =13pF and VC_3 10pF (see section 3.6) The feedback network response is therefore that of a first order low pass filter with gain $0.5/(1 + jw / w_O)$, where $w_O = 1 / 10^3(C_1 + C_{IN})$.

Therefore, $f_O = w_O / 2\pi$ = 7MHz.

There is therefore a 45° phase lag, and a gain of – 9dB at 7MHZ due to the feedback network. The open loop gain of – 1.2×10^8 /s chosen in Section 4.1.1 gives a gain of 8.7dB and a phase lag of 90° at 7MHz. The total gain round the feedback loop at this frequency is about unity while the phase lag is 135°. As the gain has fallen to unity before the phase lag has reached 180° the amplifier will be stable.

4.1.2)- Cascode Stage Tr_3 and Tr_4

a)- Introduction: In this part of the circuit Tr_4 operates in common base mode and has a low input impedance giving Tr_3 a low impedance collector load. The voltage gain of Tr_3 is –1 since Tr_3 and Tr_4 have approximately equal emitter currents and therefore approximately equal base to emitter voltages. The low voltage gain of Tr_3 reduces distortion due to the dependence of the output admittance, h_{oe} on V_{CE}. The base to collector capacitance of Tr_3 is a function of V_{CE} and therefore will introduce distortion. A high voltage gain for Tr_3 would increase the effect of this non-linear capacitance due to the Miller Effect.

The output admittance of the common base stage Tr_4 is given by $h_{ob} = h_{oe} / h_{fe}$. The input impedance of Tr_5 is given by $h_{fe} * R_L$. For BC259B transistor [37] h_{fe} = 240 to 500 and $h_{oe} < 70\mu S$ (both at 1kHz). Therefore, $1/h_{ob} > 3.4M\Omega$.

For a total output load for Tr_5 of 1KΩ given by the 2KΩ feedback resistor in parallel with the 2KΩ load resistance to be used, the input resistance of Tr_5 is greater than 450KΩ.

b)- Open Loop Gain: C_1 is chosen to give the required open loop gain of 1000 at 20kHz. The signal current through C_1 is approximately equal to the collector signal current of Tr_2 at 20kHz.

Therefore the voltage gain is given by $gm_1 * 1/jwC_1$, where gm_1 is the mutual conductance of the input stage.

$$gm_1 = 1 / 2R_e = 4mS$$

Therefore, at 20KHz for a voltage gain of modulus 1000:

$$(4 \times 10^{-3}) / (2\pi \times 10^4 \times C_1) = 1000$$

Therefore, C1 ≈ 33pF.

c)- Distortion: The value of R_3 is given by the collector current of Tr_2 (0.2mA) and v_{BE} of Tr_3 (0.64V) as 3.3KΩ (neglecting the base current of Tr_3). Input impedance of Tr_3 at I_C = 2mA is given by $h_{fe} * R_e$.

$R_e = 25 / I_E = 12.5\Omega$

Using the minimum value of h_{fe} for the BC259B of 240 gives a total input impedance including R_3 of about 1.6KΩ. The minimum impedance at Tr_4 collector is 3.4MΩ in parallel with 450KΩ (see section 4.1.2) and the impedance of the current source Tr_7. For BC169C transistor used as Tr_7 $h_{oe} < 110\mu S$ and $h_{fe} > 450$. Therefore, $1 / h_{ob} > 4M\Omega$.

The total impedance at Tr_4 collector is therefore a minimum of 360KΩ. The minimum open loop voltage gain of the stage with C_1 disconnected is $gm*R_L = I_E * 360 \times 10^3 /25$. With I_E = 2mA this becomes 28800.

The feedback loop therefore can be represented by the equivalent circuit shown in figure 4.3.

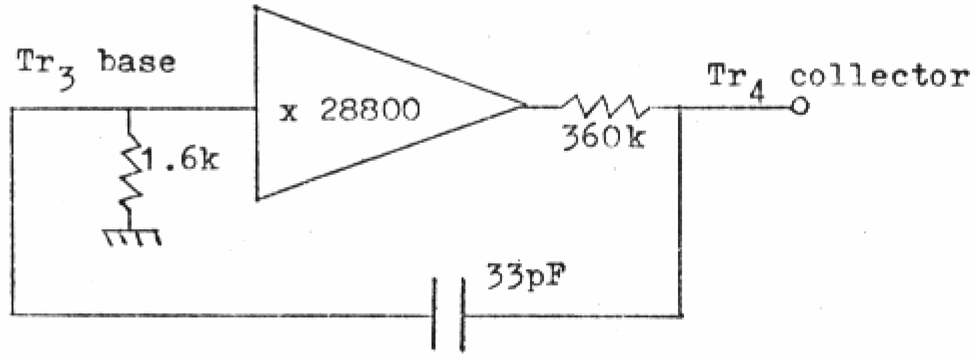


Fig. 4.3 – Equivalent Circuit of Feedback Loop

Impedance of 33pF at 20kHz = -j240 KΩ.

Minimum gain round the feedback loop is:

$$\frac{28800 \times 1.6}{1.6 + 360 - j240}$$

This gives the modulus of the gain as 106.

Including C_1 the load at Tr_4 collector is 360KΩ in parallel with approximately –j240KΩ. Therefore the modulus of impedance is 200KΩ. Therefore the modulus of open loop gain with the effect of C_1 included is given by $R_L * gm = 200 \times 10^3 * 0.08 = 1.6 \times 10^4$.

For an output of 1V the input is $1 /1.6 \times 10^4$V = 0.06mV.

Sine wave distortion in a common emitter stage due to the exponential relationship between v_{BE} and I_C is mostly second harmonic, which expressed as a percentage of the fundamental, is equal in magnitude to the peak amplitude of the fundamental in mV [36,39]. i.e. 0.06mV peak input signal gives 0.06% second harmonic distortion. With large negative feedback loop gain, AB, the distortion is reduced by approximately AB. The value of AB to be used in the calculation of second harmonic distortion should, however, be the value at the second harmonic frequency of 40kHz, i.e. 212. The loading effect of C_1 on the output of the stage is the open loop condition also gives a reduction by a factor of about 2. The closed loop second harmonic distortion is therefore about 0.06 / (212 * 2) = 0.00014%.

4.1.3)- Output Stage

This stage adds distortion due to the exponential relationship between v_{BE} and I_C. At collector current I_{C5}: gm = (I_{C5} / 25) S. (I_{C5} in mA). For R_L = 1KΩ and a 1V signal applied I_{C5} changes by a value I_{C5} = 1mA.

Changes in $v_{BE} = 10^{-3}$ / gm since gm = $\delta I_C / \delta v_{BE}$. Therefore, $v_{BE} = (10^{-3} \times 25/I_{C5})$ V = 25 / I_{C5} mV.

Percentage second harmonic distortion = v_{BE} in mV = 25 / I_{C5} %.
This is, however a percentage of v_{BE}. Therefore, as a percentage of the 1V output signal:
Second harmonic distortion = $25 \times 25 \times 10^{-3} / I_{C5}^2 = 0.625/ I_{C5}^2$ %. (I_{C5} in mA).
For I_{C5} = 10mA second harmonic distortion = 0.006%.

4.1.4)- Total Distortion

There are several sources of distortion [42], which have not been considered. These include the variations of barrier and diffusion capacitances in the transistor resulting from changes in V_{CB} and I_C respectively, and changes in h_{fe} resulting from changes in V_{CB} and I_C. At low frequencies thermal modulation effects may also become significant [43]. Rough calculations suggest that the most important of these is the change in h_{fe} of Tr_5 due to its changing V_{CB}. This gives second harmonic distortion of the order of 0.001%.

The total RMS distortion depends on the relative phases of the separate components at each harmonic. The worst-case figure is the sum of the individual components. The second harmonic distortion derives for Tr_3 (0.00014%) and Tr_5 (0.006% and 0.001%) for a 20kHz signal give a maximum total of about 0.007%. The overall negative feedback loop gain is about 250 at 40kHz and therefore the total closed loop second harmonic distortion will be a maximum of about (0.007 / 250)% = 0.00003%.

The third harmonic distortion of the input stage (0.005%) is reduced by the –6dB/octave response of the second stage and the feedback loop gain (167 at 60kHz) to about 0.00001%. Total RMS distortion D = SQR $(d_2^2 + d_3^2)$ where d_2 and d_3 are second and third harmonic percentages respectively. The total RMS harmonic distortion at an input signal frequency of 20kHz and an output peak amplitude o 1V is therefore a maximum of about 0.000032%.

In general in class-A amplifier second harmonic distortion is proportional to the signal amplitude while the third harmonic distortion is proportional to the square of the signal amplitude [39].

4.1.5)- Circuit Details

The source of Tr_5 is shown with a variable resistance connected. As the distortion produced by the input stage is critically dependent on the matching of the collector currents of Tr_1 and Tr_2 the optimum current through Tr_6 is most easily set by adjusting it to give minimum total amplifier distortion. By applying sine wave common mode signal to the differential input stage of the measuring instrument (i.e. applied to R_1 and VR_1 in figure 3.4) the input signal can be cancelled to leave a signal, which include the distortion of A_1. Adjustment of the current through Tr_6 alter the second harmonic distortion and therefore by observing the amplitude of the total distortion at the output of A_2 the optimum current can be set. A field effect transistor current source was used rather than a further bipolar transistor similar to Tr_7 and Tr_8. This was done because a battery supply was used to give low hum and noise and consequently the supply voltage reduced slowly with the time and the resistive bias arrangement used for Tr_7 and Tr_8 bases gave slowly changing collector currents. Also it was found that when using three bipolar transistors as current sources with their bases biased by the same resistive voltage divider the amplifier had two stable states and on being connected to the power supply it went into one of these states in which the output voltage becomes about –8V. Shorting the bases to 0V line for a moment triggered the circuit into the required operating condition with a 0V output for no input signal. The effect was due to Tr_8 taking a large base current until its V_{CE} increased sufficiently for the current gain to become significant. The resulting low base voltage on the input stage current source transistor prevented it from conducting sufficiently to make Tr_2, Tr_3, Tr_4 and Tr_5 conduct and increase V_{CE} of Tr_8. The negative output state was consequently stable. The use of a field effect transistor for Tr_6 completely solved the problem.

The approximate source resistance required by Tr_6 was calculated by first measuring I_{DSS} and V_P for the device used (an E202 type). A Tektronix Type 576 Curve Tracer was used and values of I_{DSS} = 1.71mA and Vp = 1.4V obtained.

Using the formula below we found V_{GS} = 0.72V at I_D = 0.4mA.

$$I_D = I_{DSS} (1 - V_{GS} / V_P)^2$$

Therefore the source resistance required = (0.72 / 0.4) KΩ = 1.8 KΩ. A value of 2.2 KΩ in parallel with a 47 KΩ preset potentiometer was used.

The voltage and current levels relevant to the choice of resistor value are shown in figure 4.1 together with the nearest standard value of resistance corresponding to these levels. As the transistor used are all very high current gain types the base currents are very small and have been ignored.

C_3 and C_4 ensure that Tr_4, Tr_7 and Tr_8 operates in common base mode by effectively earthing the bases at frequencies within the audio range. The capacitor values are not critical and were chosen as 33µF. As electrolytic capacitors have significant impedance at very high frequencies a small value ceramic disc capacitor (0.02µF) was connected in parallel with each electrolytic to maintain a low impedance. The supply decoupling capacitors C_2, C_5 and C_6 were included to prevent interaction via the supply connections with other sections of the test instrument. The test circuit including amplifier A_1 and A_2 (figure 3.4) was build on Veroboard and mounted in a diecast aluminium box connected to the 0V line to reduce interference pickup.

4.2)- Output amplifier, A_2

The output amplifier (A_2 in figure 3.4) must have low noise as the distortion signal to be amplified may be at a very low level. Low noise transistors are therefore required. Extremely low distortion is however not essential and only other requirement is for sufficiently wide bandwidth to pass all distortion of interest. In some applications it may be useful to limit the bandwidth of the amplifier to reduce the effects of noise. It was decided to use a separate filter, which could be disconnected when not required, and not add any high frequency gain reducing components to A_2. The choice of suitable filters is described in section 4.3.

The gain required for A_2 depends on the maximum level of distortion to be measured, and also on the sensitivity of the measuring equipment to be connected to its output. A gain of 100 was chosen so that a distortion level of 1% produced by a 1V input test signal will give a distortion output of 1V. A published design was chosen (REF. 40) which is shown in the necessary form for this application in figure 4.4. VR_2, R_5 and R_6 are as shown earlier in figure 3.4. Tr_{10} is a low noise field effect transistor used as a 0.1mA constant current source. Measurement of I_{DSS} and V_P as 1.92mA and 1.4V respectively lead to the calculation of the source resistor value as 11KΩ.

The operation of the circuit is described in detail in REF. 40. A 4.7µF input coupling capacitor, C_7, is used to eliminate the effects of any offset voltages in the amplifier being tested. With $R_3 = VR_1$ = 2KΩ and $R_4 = R_5$ = 1KΩ (figure 3.4) the resistance in series with C_7 is about 3KΩ at low frequencies giving a high pass first order characteristic with gain reduced by 3dB at about 11Hz.

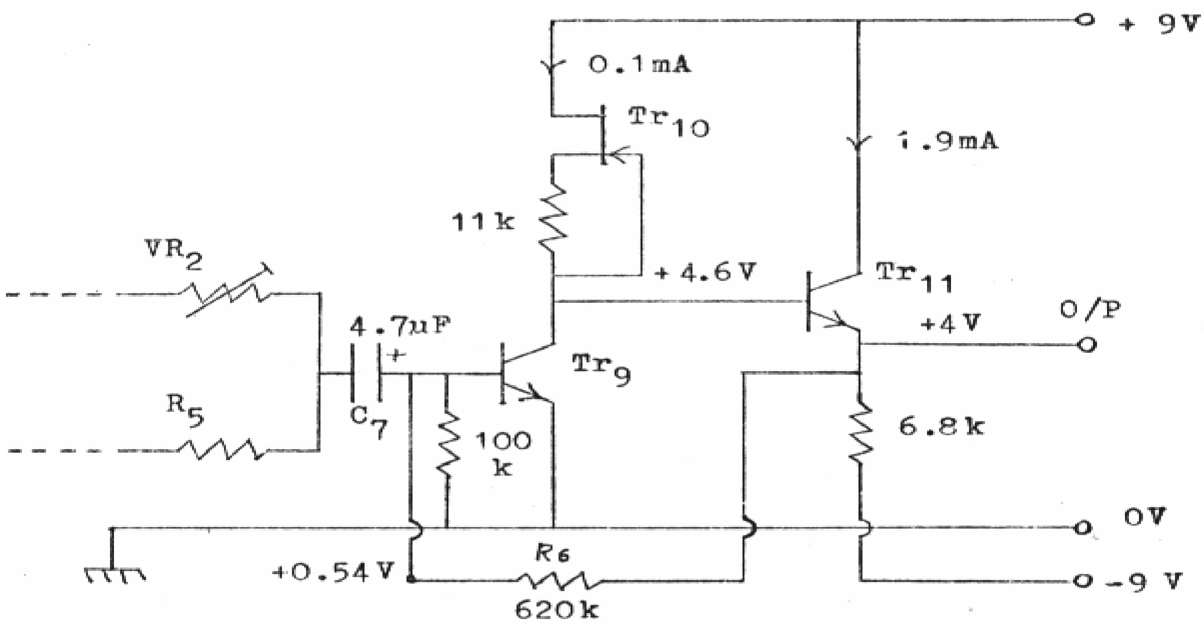


Fig. 4.4 – Circuit Diagram of Amplifier A_2

With a source resistance of 3KΩ a collector current of 0.1mA for a BC169C transistor [37] gives a noise figure of about 0.5dB.

For $v_2 = 0$ and $v_1 = v_i$ in figure 3.4, VR_1 and R_3 from a signal source given by their Thevenin equivalent as 0.5 v_i in series with 1KΩ (for $VR_1 = R_3 = 2KΩ$). Including R_4 and R_5 (1KΩ each) gives a total source resistance of 3KΩ and an output from A_2 of (R6 x 0.5 v_i) / 3KΩ assuming A_2 to have a very high open loop gain.

Therefore, for a gain of 100, R6 = 600KΩ.

When testing an amplifier with gain $-A$ the output of the amplifier for input v_i will be $-Av_i$ + distortion, D.

The output voltage of A_2 is then:

$$100\, v_i + (-Av_i + D)\, R_6 / VR_2.$$

For the undistorted signal to cancel it is required that $A * R_6 / VR_2 = 100$ therefore, using R_6 = 600KΩ gives VR_2 = 6A KΩ

The minimum recommended value of V_{CE} for Tr_9 and V_{DG} for Tr_{10} is given in REF. 40 as 2V. Below this value the impedance at Tr_9 collector falls significantly and consequently the open loop gain also falls. The values of V_{CE} and V_{DG} are about 4.6V and 4.4V respectively giving a maximum recommended peak output voltage of about ±2.4V.

The base to collector capacitance of Tr_9 is effectively in parallel with the 620KΩ feedback resistor. The value of C_{CB} is about 5pF at V_{CE} = 4.6V for the BC169 transistor used for Tr_9 [37]. This gives a closed loop –3dB frequency of approximately 50kHz.

4.3)- Optional Filter Stage

The design of simple high and low pass filters is described in REF. 41. A variety of response shapes are possible but the most useful in this application are the Bessel (maximally flat time delay) and the Butterworth (maximally flat attenuation). The Butterworth type should be used when the relative amplitudes of distortion components in the pass band are to be preserved, e.g. when the total output of the instrument is to be measured using a RMS reading meter. The Bessel response is more suitable when the distortion waveform is to be displayed on an oscilloscope as the alteration to the shape of the waveform is minimised.

Suitable filters are the active filter modules made by Barr & Stroud Ltd. Their Series EF10/20 filters offer a choice of Bessel, Butterworth or Chebyschev (equal ripple in the pass band gain) and second, third or fourth order responses determined by externally added resistors and capacitors. The range of high and low pass –3dB frequencies is from 1Hz to 30kHz with a maximum input signal of 5V peak.

4.4)- Total Circuit Noise

A resistance of R ohms at temperature T in Kelvin generates thermal noise of RMS value SQR (4KTRB) volts, where K is the Boltzmann's constant (1,38 x 10^{-23} J/^{o}C) and B is the noise bandwidth in Hz.

At T = 300K and B = 20kHz the RMS noise voltage, vn = SQR (4 x 1,38 x10^{-23} x 300x2 x10^{4} x R) = 1.8 x 10-8 x SQR (R) Volts.

For R in KΩ v_n = 0,58 SQR (R) µV. Therefore, 2KΩ gives v_n = 0.82µV.

The thermal noise voltages in the complete circuit of the test instrument are shown in figure 4.5.

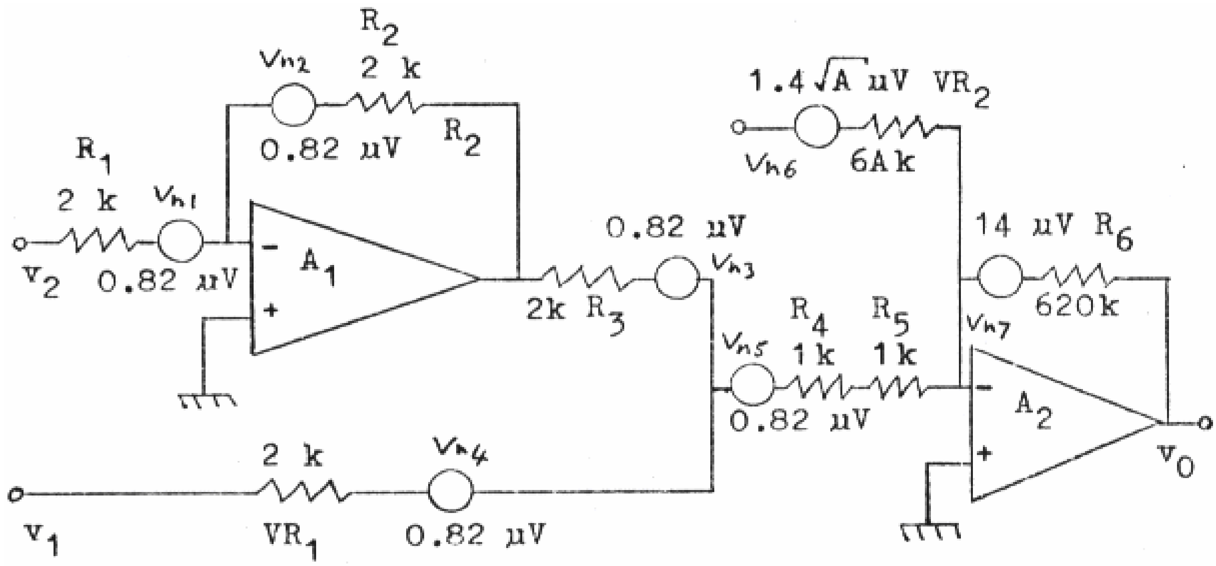


Fig. 4.5 – Thermal Noise Voltages in the Test Circuit

(A is the voltage gains of the amplifier being tested). The noise voltages, v_{n1} and v_{n2} are added by A_1 and appear in series with R_3 and therefore add to v_{n3}. The addition of RMS voltages vn1, vn2 and vn3 gives a total of SQR ($v_{n1}^2 + v_{n2}^2 + v_{n3}^2$) = 1.4µV.

The Thevenin equivalent of R_3 and VR_1 gives the sum of noise voltages (1.4 / 2) µV and (0.82 / 2) µV in series with 1KΩ. The addition of vn5 gives a total of 1.1µV in series with 3KΩ.

The noise voltages added by A_2 gives output noise voltages of about: (1.1 x 600) / 3µV from the input stage, (1.4 SQR x 600) / 6AµV from VR_2, and (14 x 600)/600µV from R_6.

Generally A >> 1. The greatest contribution from v_{n6} is for the minimum value of A. Therefore for a worst-case output noise voltage put A = 1. The noise voltage to be added is then 220µV, 140µV and 14µV. The total output noise voltage is then 260 µV.

For a test signal of 1V a distortion signal of 2.6μV representing 0.00026% distortion will be multiplied by 100 to give 260μV output from A_2. Therefore the distortion signal will fall below the level of the thermal noise voltage at distortion levels of less than about 0.00026% with a 1V test signal. This does not however represent a lower limit of THD or IMD measurement possible as the individual components of a distortion signal can be extracted from the wide band noise by the use of a wave analyser as explained in section 1.2.

The noise generated by the amplifier being tested may be more significant than that of the test circuit. If not then an increase in distortion to noise ratio can be gained by placing an attenuator at the input of the amplifier being tested and using the larger amplitude signal at the input of the attenuator for comparison in the test instrument. For an attenuation 1/K at the amplitude input the test signal must be increased by a factor K and therefore the maximum amplitude signal which the unity gain inverting amplifier, A_1, can handle without introducing significant distortion.

Chapter 5. - Practical Evaluation of the Measuring Circuit

A series of measurements were made to investigate the practical abilities and limitations of the test circuit developed in Chapter 4.

Apparatus used

Signal generator: Farnel type ESG1
Oscilloscope: Telequipment D67 dual trace.
Oscilloscope camera: Telford type A, with Polaroid 107C Land Film.
Stabilised Power Supply: Weir Minoreg type 325 (for 741 op-amp supply)
Batteries type PP6 9 volts (for instrument supply)
Dymar A.F. Wave Analyser Type 1771
Resistors used in test instruments (R_1, R_2, R_3, R_4, R_5, VR_1 and VR_2 in figure 3.4): 0.5W metal oxide from RS Components Ltd.
Operational Amplifier type ML741CS.
Resistance box

Procedure:

The measuring instrument was built using the circuit described in Chapter 4. For most of the measurements to be described an additional amplification stage was used to increase the output to a sufficient level for display on the oscilloscope, which has a maximum sensitivity of 10mV/cm. This amplifier stage is shown in the figure 5.1. The gain provided is –10 giving a total instrument gain of 60dB. The –3dB frequency range of the stage is about 15Hz to 100kHz.

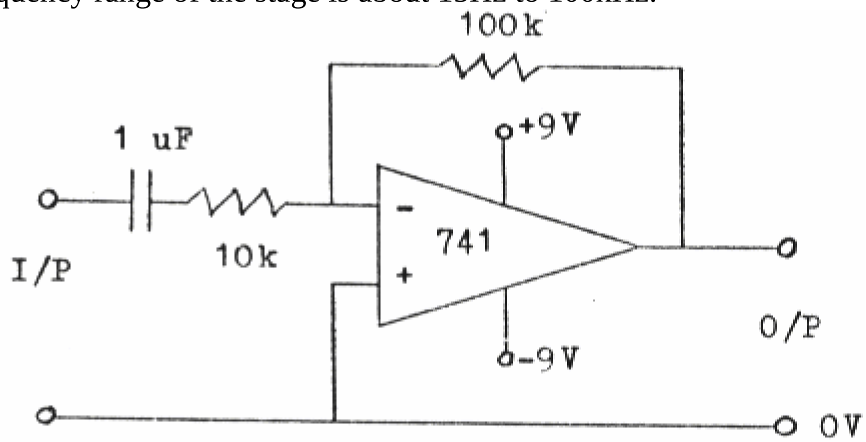


Fig. 5.1 – Additional Gain Stage

The following measurements were made:

5.1)- Frequency Response of the Instrument

The frequency response of the test instrument was measured by applying a sine input signal of 10mV peak amplitude across the differential input (with VC_1 and $VC_2 = 0$) and observing the output of amplifier A2 on the oscilloscope. This gain at 1kHz was found to be close to 100 when the feedback resistor, R6, was chosen as 680KΩ. The gain relative to that at 1kHz is shown for a range of frequencies in Table 5.1.

Table 5.1 – Frequency Response of Instrument	
Frequency (kHz)	Gain relative to 1kHz (dB)
20	-0.45
30	-0.92
40	-1.4
50	-1.9
60	-2.5
70	-3.1
80	-4.1
90	-4.8
100	-5.2

5.2)- Distortion of the Instrument

The instrument distortion can be measured using a common mode input signal as described in Section 4.1.5. The distortion of the signal generator was first measured using the Dymar wave analyser. At an output of 1V RMS at 2kHz the distortion at the individual harmonic frequencies is shown in Table 5.2.

Table 5.2 Distortion of the signal generator at 2kHz	
Frequency of Harmonic (kHz)	Amplitude Relative to Fundamental
4	0.48% (-46dB)
6	0.13% (-58dB)
8	0.037% (-69dB)
10	0.028% (-71dB)

The higher harmonics could not be measured as they were below the noise level of about 0.01% (-80dB). (The manufacturer specification states that the wave analyser can make measurements down to –74dB.) At an output of 1 V RMS at 20kHZ the distortion at the second harmonic frequency of 40kHz was 0.47% (-47dB). The wave analyser has a maximum frequency of 50KHz so the higher order harmonics could not be measured.

Applying a 2kHz common mode signal of peak amplitude 1 V to the differential input of the instrument the fundamental output was reduced by adjustment of VC_3 (figure 4.1) and VR_1 (figure 3.4). For VR_1 a network of fixed resistors and a potentiometer were used as described in section 3.5. The network used in this case is shown in figure 5.2.

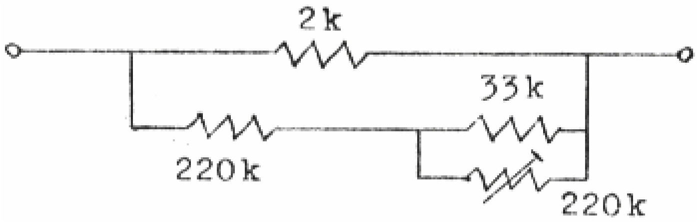


Fig. 5.2 – Network used for VR_1

The capacitor used for VC_3 was of the Philips "beehive" air –spaced trimmer type, of maximum value 8pF in parallel with a fixed capacitor of value 5.6pF.

The output of the instrument (overall gain 60dB) at 2kHz was reduced to about –51dB relative to the input signal (measured with the wave analyser) giving a common mode rejection ratio (CMRR) of 111dB at this frequency. This is only a typical value as the ratio changed slowly with time. With no further adjustments to VR_1 and VC_3 the CMRR was measured at higher frequencies giving results as in Table 5.3.

Table 5.3 – CMRR with value optimised at 2kHz.	
Frequency (kHz)	CMRR (dB)
2	111
4	106
6	104
8	103
10	102
20	99

VC_3 and VR_1 were adjusted with an input of 20kHz to optimise the value of CMRR at this frequency. The CMRR then became 110dB. With the same settings of VC_3 and VR_1 the CMRR at 40KHz was found to be 92dB.

From the values of CMRR the effect of the generator distortion can be calculated when measuring the distortion of the instrument at 2Khz or 20KHz. For a common mode input signal of 2kHz with distortion d_n at the n^{th} harmonic and CMRR c_n at the harmonic frequency (c_n and d_n in dB) the effect of the generator distortion will be identical to that of $-d_n - c_n$ (in dB) distortion generated by the amplifier A_1 at the n^{th} harmonic relative to the common mode input signal.

For CMRR optimised at 2kHz the distortion level in A_1, which would give the same instrument output as the common mode generator distortion, is given in Table 5.4.

Table 5.4 – Effect of generator distortion	
Frequency (kHz)	Equivalent A_1 distortion
4	-46 –106 = -152dB
6	-58 – 104 = -162dB
8	-69 – 103 = -172dB
10	-71 –102 = -173dB

Similarly for CMRR optimised at 20kHz the second harmonic generator distortion will give the same instrument output amplitude at 40kHz as second harmonic distortion in A_1 of (-47 –92) dB = -139dB. These figures give the limit of measurements of distortion of A_1 when using the signal generator tested. The distortion of the later stages of the instrument is not likely to be significant as only the distortion and the common mode breakthrough will be distorted and these are all at very low levels.

The harmonic distortion was measured using a 2kHz common mode signal with CMRR optimised at that frequency. The distortion levels measured at the harmonics are shown in Table 5.5.

Table 5.5 - A_1 Distortion. - 2kHz Input.	
4kHz	- 140 to –146dB
6kHz	- 141 to – 144dB
8kHz	Not measurable

Similarly at 20KHz the second harmonic distortion (40kHz) was measured as –118dB (including a correction for the fall in gain of the instrument at 40kHz from Table 5.1).

The distortion figures are all well within the limits set by the generator distortion breakthrough. The figures for 2kHz signal distortion are, however, very approximate as the distortion is of the same order of magnitude as the noise detected by the wave analyser and the readings fluctuated considerably. Only the second and third harmonics were of sufficient amplitude to be distinguishable from noise when displaying the wave analyser output on an oscilloscope (with its timebase triggered by the signal generator output).

At lower test signal amplitudes the distortion reduced as expected and became unmeasurably low at all harmonics. At higher amplitudes the distortion remained very low up to about 4.5V common mode signal beyond which it rose rapidly.

Adjustment of the input stage current source FET (see figure 4.1) gave no significant change in the second harmonic distortion indicating that the input stage contribution to this must be small.

5.3)- Measurements on a 741 Operational Amplifier

A 741 operational amplifier (type ML741CS) was used in the circuit arrangement shown in figure 5.3 to demonstrate both the effectiveness of the load compensation arrangement (Section 3.4) and the ability of the test instrument to extract low-level distortion waveforms.

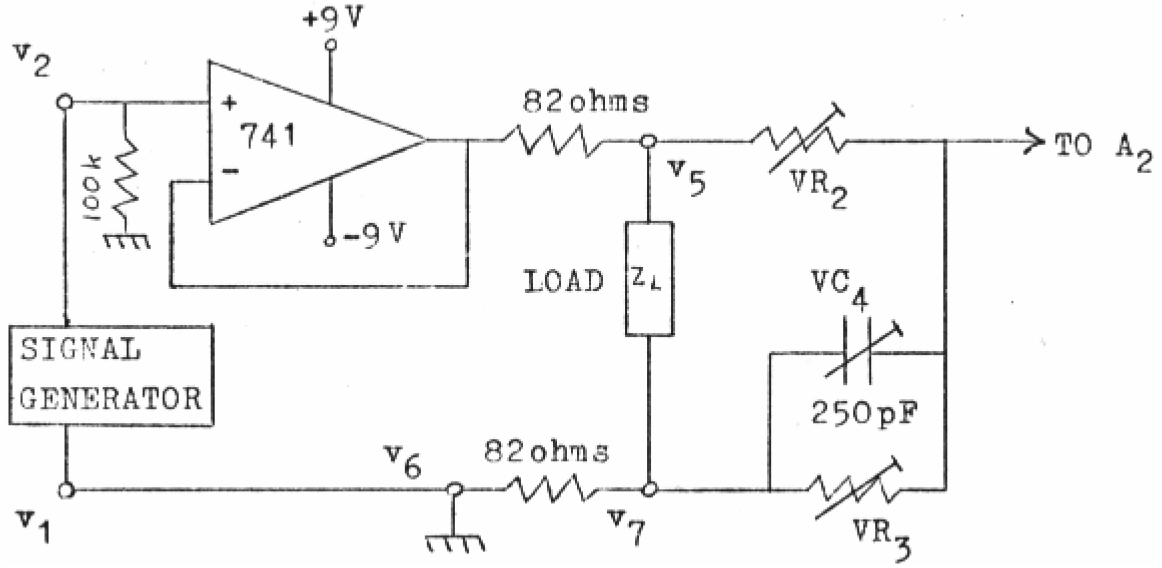


Fig. 5.3 - Circuit of 741 Amplifier Used in Tests

The voltages v_1, v_2, v_5, v_6 and v_7 are as shown in figure 3.4 and figure 3.6 which also shows the connections to the test instrument. The 82Ω output resistor is included to increase the effect of the load on the output signal and thereby make the test more demanding.

The phase compensation components VC1 and VC2 (see figure 3.4) were in this case each a 150pF fixed capacitor in parallel with a 56pF trimmer capacitor.

With no load connected VR_2, VC_1 and VC_2 were adjusted in an attempt to cancel out the fundamental for test frequencies of 2kHz and 20kHz but it was found that when adjusted at 2kHz for optimum cancellation an extra phase lag was always required in the phase compensation network for optimum cancellation at 20kHz. The best results were obtained with VC_1 and VC_2 approximately equal. This gives the maximum effective time delay as shown in Section 3.1. With optimum adjustment at 2kHz the fundamental breakthrough at 20kHz was observed on the oscilloscope and seen to be of the same order of magnitude as the distortion produced by that frequency.

The networks used for VR_2 and VR_3 in this case are as shown in figure 5.4. The 250pF variable capacitor, VC4, was a compression type trimmer used to compensate for high frequency phase errors.

VR_2=

VR_3=

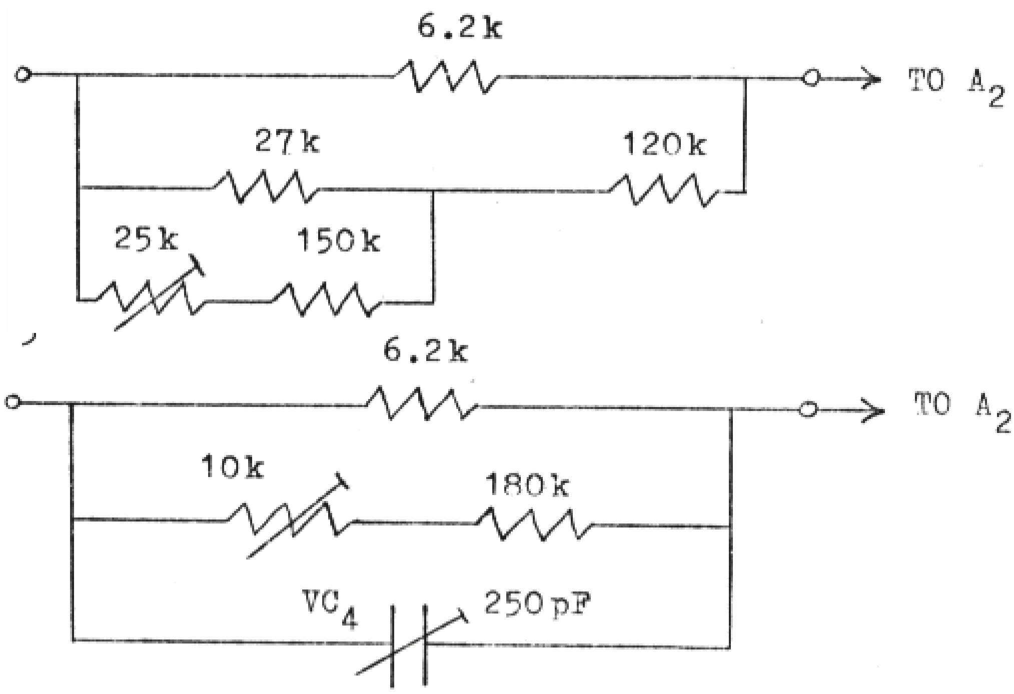


Fig. 5.4 – Networks Used for VR_2 and VR_3

Measuring the output of the test instrument with the wave analyser the rejection of a 2kHz test signal was optimised using VR_2, VC_1 and VC_2. The values of VC_1 and VC_2 were estimated as 170pF each after adjustment. A 1V peak amplitude test signal was used for all the tests on the 741 operational amplifier. The instrument output at 2kHz was reduced to –53dB relative to the test signal input. Taking into account the 60dB gain of the instrument the rejection of the 2kHz signal is –113dB. With the same component settings the signal rejection was measured at higher frequencies and the results are shown in Table 5.6.

Table 5.6 – Best Obtainable Rejection of Signals with Rejection Optimised at 2kHz.

Frequency	Rejection
2 kHz	- 113 dB
4 kHz	- 104 dB
6 kHz	- 94 dB
8 kHz	- 87 dB
10 kHz	- 83 dB
20 kHz	-71 dB

Increasing one of the phase compensation capacitors VC_1 or VC_2 improved the rejection at 20KHz to –90dB. With this setting the rejection at an input frequency of 40kHz was found to be –58dB.

These results can be combined with the measured values of generator distortion to calculate to the limit to distortion measurements set by generator distortion breakthrough. The limits calculated are shown in Table 5.7 for a 2kHz test frequency. For a 20kHz test frequency the limit to second harmonic distortion measurement is (-47 –58) dB = -105dB.

The distortion waveform obtained at the test instrument output at a test frequency of 2kHz was photographed and is shown in figure 5.5. No load was connected but it should be noted that the test instrument has an input impedance of about 6KΩ, which may have a significant effect on the 741-amplitude distortion.

Table 5.7 – Limit of Distortion Measurement with 2kHz signal	
Frequency	Limit to measurement due to generator distortion
4 kHz	(-46 –104) dB = -150dB
6 kHz	(-58 –94) dB = -152dB
8 kHz	(-69 –87) dB = -156dB
10 kHz	(-71 –83) dB = -154dB

A 1KΩ load was then connected in the position shown in figure 5.3 and the load effect compensation components VR_3 and VC_4 adjusted to cancel the 2kHz fundamental. The distortion waveform then obtained is shown in figure 5.6. The peak-to-peak amplitudes of test signal and distortion waveform were compared to give a total peak distortion figure. The distortion waveforms contain significant noise components and the peak-to-peak distortion values used are therefore only estimates attempting to neglect the noise. With the 2kHz test signal the total peak distortion is found to be:

No load, 0.0014% (-97dB).
1kΩ load, 0.0024% (-92dB).

The measurement procedure was repeated at a test signal frequency of 20kHz. The result with no load is shown in figure 5.7 and with a 1KΩ load in figure 5.8. The peak distortion levels are:

No load, 0.03% (-70dB).
1KΩ load, 0.04% (-68dB).

Both can be seen to be predominantly second harmonic. The value for both 2KHz and 20KHz are well within the limits set by signal generator distortion breakthrough.

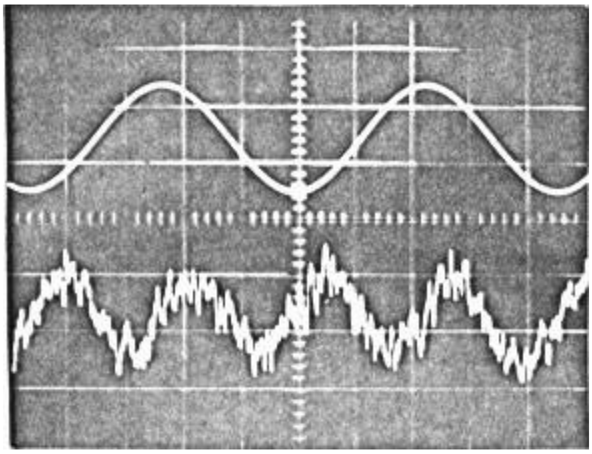


Fig 5.5 Upper: Test signal, 2V p-p, 2kHz, (1V/div).
Lower: Distortion waveform, no load, 28µV p-p, (20µV/div).
Peak distortion: 0.0014% (-97dB)

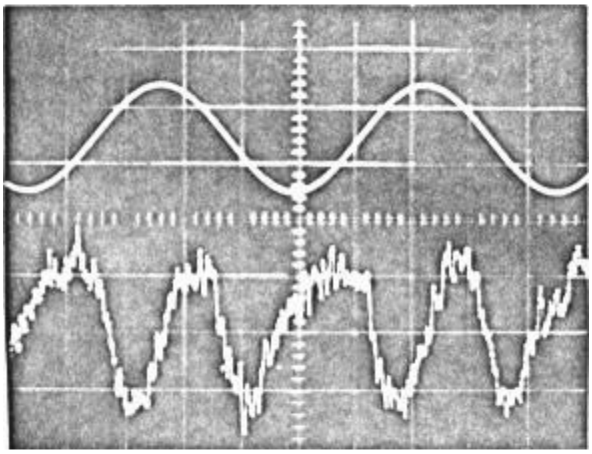


Fig. 5.6 – **Upper: Test signal, 2V p-p, 2kHz, (1V/div).**
Lower: Distortion waveform, 1KΩ load, 48µV p-p, (20us/div).
Peak distortion: 0.0024% (-92dB)

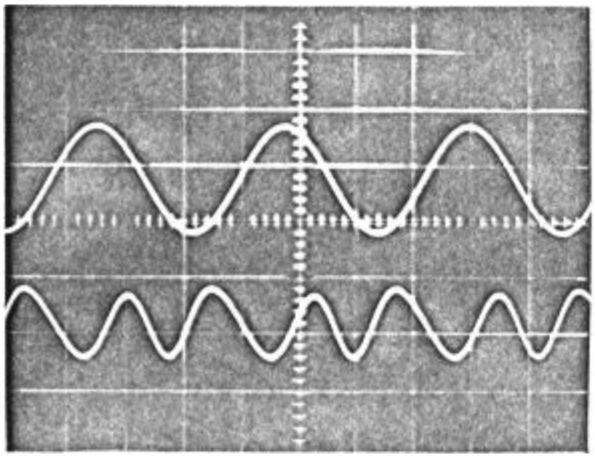


Fig. 5.7 - **Upper: Test signal, 2V p-p, 20kHz, (1V/div).**
Lower: Distortion waveform, no load, 0.6mV p-p, (0.5mV/div).
Peak distortion: 0.03% (-70dB).

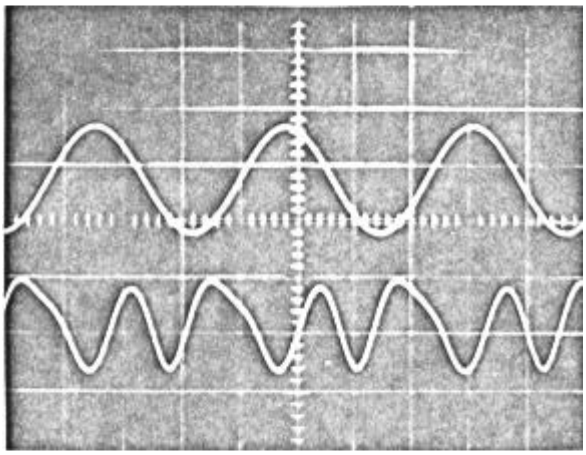


Fig. 5.8 - **Upper: Test signal, 2V p-p, 20kHz, (1V/div).**
Lower: Distortion waveform, 1KΩ load, 0.8mV p-p, (0.5 mV/div).
Peak distortion: 0.04% (-68dB).

To demonstrate the ability of the load compensation circuit arrangement to eliminate the effects of loads with reactive and non-linear components the load shown in figure 5.9 was used.

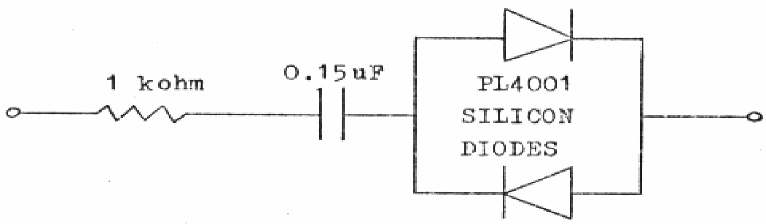


Fig. 5.9 – Load with resistive, reactive and non-linear components.

The circuit was first adjusted with a test signal of 2kHz giving an output with no load connected similar to that shown previously in figure 5.5. On connecting the load (figure 5.9) with the load compensation components VR_3 and VC_4 disconnected the test instrument output waveform became as shown in figure 5.10. i.e. The waveform peak-to-peak amplitude increased by a factor of about 1000. (The extra gain stage (figure 5.1) was not used for figure 5.10) Connecting VR_3 and VC_4 and adjusting them for minimum load peak distortion gave the result shown in figure 5.11.

The distortion waveform of figures 5.5 to 5.8 and 5.11 were analysed using the wave analyser and the results are presented in Table 5.8.

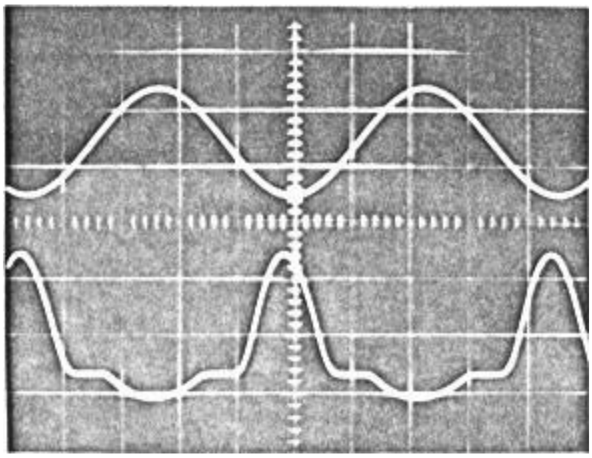


Fig. 5.10 - Upper: Test signal, 2V p-p, 2kHz, (1V/div.).
Lower: Distortion, Figure 5.9 load, 26mV p-p, (10mV/div.).
Peak distortion: 1.3% (-38dB)

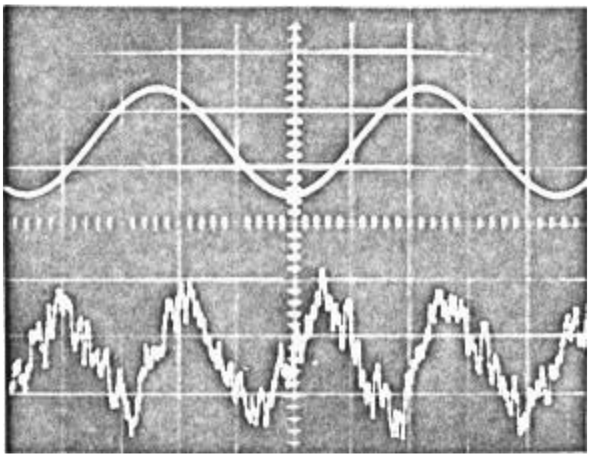


Fig. 5.11 - Upper: Test signal, 2V p-p, 2kHz, (1V/div.).
Lower: Distortion, Figure 5.9 load compensated, 40uV p-p, (20µV/div)
Peak distortion: 0.0020% (-94dB).

Table 5.8 – Wave analyser measurements of the distortion					
Fundamental:	2kHz	2kHz	20kHz	20kHz	2kHz
Load:	None	1KΩ	None	1KΩ	Fig. 5.9
Harmonic N^{o}	Amplitudes in dB relative to fundamental				
2	-96	-91	-71	-69	-94
3		-100			-111
4		-109			-111
5		-115			-115
6		-119			-119
7		-129			

The higher order distortion components not shown in the table for the 2kHz test signal were all below the noise level.

The Dymer A.F. Wave Analyser Type 1771 used to measure the amplitudes of the harmonics has the following specifications:

Frequency range:	20Hz to 50kHz.
Selectivity:	-3dB ± 5Hz.
	-40dB ± 50Hz
	-60dB ± 100Hz
	-70dB ± 200Hz
Residual noise:	< -80dB.
IMD:	<-70dB.

Chapter 6. - Measurements of Power Amplifier Distortion

6.1)- Power Amplifier Design

To illustrate some of the uses of the test instrument in the testing of power amplifiers an amplifier was designed and built. The circuit is shown in figure 6.1 and is the usual arrangement of differential input stage, driver and class-B output stage. An additional common collector stage, Tr_3, is included. This gives the driver stage, Tr_4, a low source impedance and consequently reduces distortion caused by the Early effect. It also reduces the effect of feedback from collector to base via C_{CB} giving a wider bandwidth. The use of this technique is a suitable alternative to the cascode arrangement used earlier (Amplifier A_1, figure 4.1) having the advantage that the available output voltage swing is not reduced. Transistors Tr_5 and Tr_6 form a temperature dependent current source. The collector current of Tr_6 is proportional to the base to emitter voltage of Tr_5, which has the same temperature coefficient as the base to emitter voltages of the output stage transistors. By placing Tr_5 in thermal contact with the output transistor heat sink the effect on the quiescent current of temperature changes caused by output stage power dissipation can be partly compensated for, although there will be a delay in the compensation due to the non-zero thermal time constant involved. The output stage quiescent current produced can be set to the required value by VR_4. Transistors Tr7 and Tr8 are shown as single transistors but are actually Darlington pairs. The current gain is specified as a minimum of 750 at I_C – 3A.

As current limited power supplies were used with a maximum current of about 0.5A a load resistance of 22Ω was chosen with a supply voltage of ± 15V so that an output voltage of 20V peak to peak could be used for the tests. For a sine wave signal this represent an output power of 2.3W. While this is not representative of the usual operating condition with a loudspeaker it is sufficient for the intended test. The high frequency stabilisation was arrived at largely by trial and error. The use of a feedback capacitor to the base of Tr1 reduces the possibility of the occurrence of transient intermodulation distortion [(6)] (TID) or slew rate limiting as described in REF. 12.

The connections to two signal generators are shown for the demonstration of intermodulation distortion. A low pass first order filter is included at the input. This is required when using square wave test signals so that their harmonics beyond the audio frequency range can be attenuated. Using the 0.068µF capacitor shown the –3dB frequency of this filter is 17.5kHz. This was used in all the tests to be performed. The need for such a filter when carrying out TID tests is described in REF. 18, although this suggests a higher –3dB frequency (30kHz). The choice depends on the frequency range of the signal, which the amplifier is intended to handle.

The amplifier has a low frequency gain of –10 and is direct coupled to avoid the need for low frequency gain and phase compensation in the test instrument. The amplifier was build on Veroboard and the output transistor mounted on a heat sink consisting of a 1/8” thick sheet of aluminium about 3” x 4”. The input and feedback resistors were 0.5W metal oxide types to give stable closed loop gain and low distortion.

6.2)- Tests Results

The tests carried out were to detect the presence of crossover distortion, TID and phase modulation. As the intention is only to illustrate some of the uses of the test instrument no attempt has been made to analyse the distortion generating processes within the amplifier or to obtain accurate numerical distortion specifications. The results are presented as photographs of the oscilloscope traces obtained together with brief comments on the interpretation.

The instrument was adjusted to compensate for the gain and phase response of the amplifier. No load effect compensation was required as a resistive load was used. VR_2 was a network of fixed resistors and a potentiometer similar to those used previously but with a total value of about 60KΩ. For optimum high frequency phase and gain compensation VC_1 was zero while VC_2 was a 56pF variable capacitor in parallel with a total of 537pF. When adjusted the total value of VC_2 was estimated as 580pF.

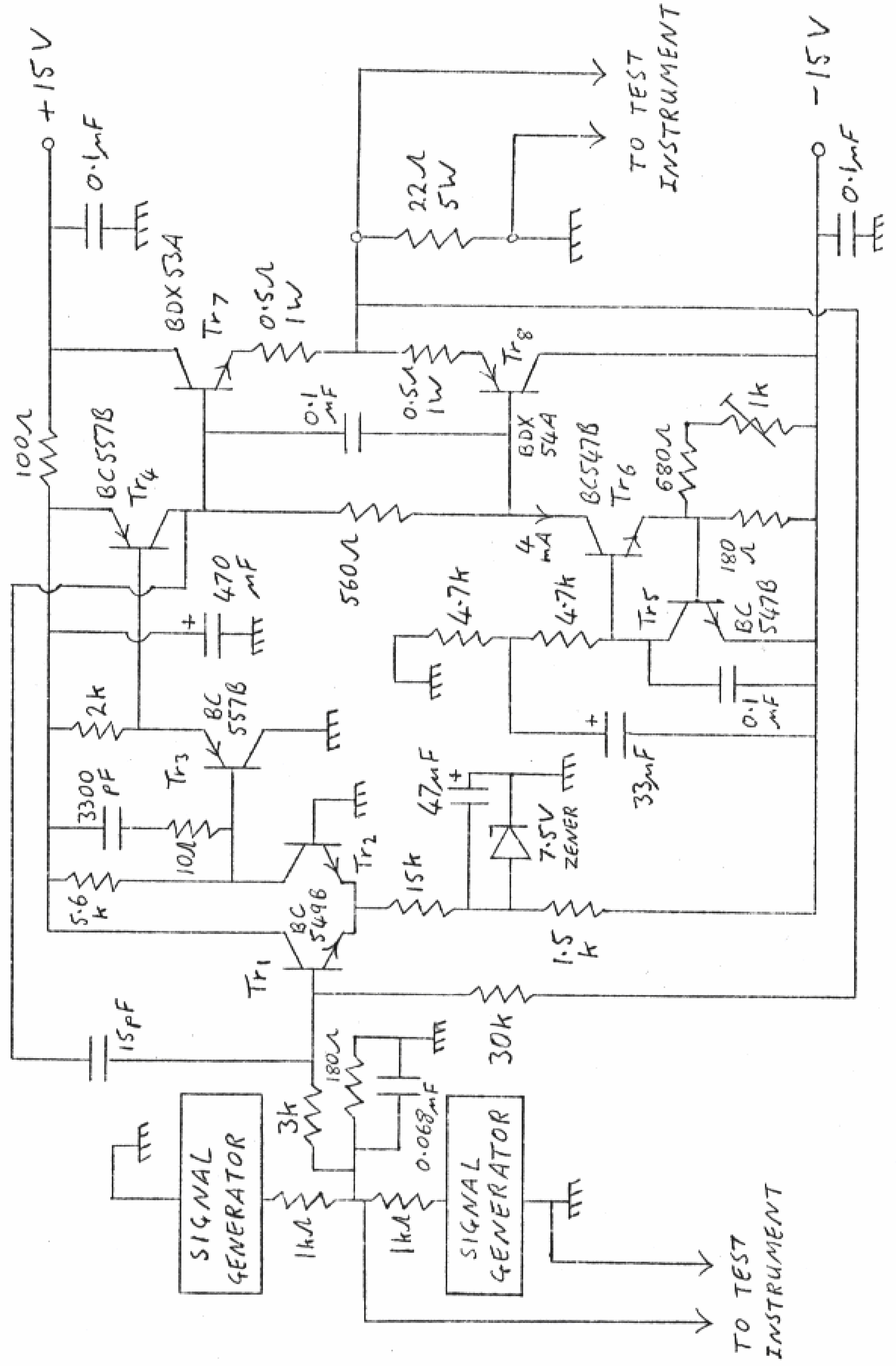


Fig. 6.1 – Circuit of Power Amplifier

The total peak-to-peak instrument output signal was measured at a range of frequencies with the compensation adjusted for good rejection of fundamental frequencies throughout the whole audio frequency range. The values obtained include distortion but give an indication of the accuracy of the compensation obtained as at all frequencies the fundamental breakthrough predominated. The rejection can be seen to be better than 90dB up to 20kHz. The results are shown in Table 6.1.

As the distortion was below the test frequency breakthrough with this optimum wide band compensation it was found to be necessary to adjust the compensation to best indication of the distortion in each of the tests performed.

Table 6.1 – Rejection of Fundamental With Compensation Optimised. For Wide Band Signal Use. Amplifier Output = 20V p-p

Frequency	Rejection (dB)
100 Hz	98
500 Hz	98
1.0 kHz	98
2.0 kHz	98
4.0 kHz	95
6.0 kHz	92
8.0 kHz	92
10.0 kHz	91
12.5 kHz	94
15.0 kHz	95
17.5 kHz	92
20.0 kHz	90
25.0 kHz	84

Figures 6.2 and 6.3 show the sine wave distortion obtained at frequencies of 2kHz and 20kHz respectively. The output stage quiescent current, I_Q, was adjusted for minimum peak-to-peak distortion in each case and measured as 36mA for 2kHz and 40mA for 20kHz. Reducing I_Q to 25mA gave the result shown in figure 6.4 (2kHz test signal) while increasing it to 100mA gave figure 6.5.

Note: All photographs with the exception of figures 6.3, 6.4, 6.5, 6.6 and 6.7 (including those in Chapter 5) were obtained with the oscilloscope in the alternate trace mode using the single shot facility. This gives good indication of waveforms with a high noise content but does not always give accurate relative phase of two traces as there is a time delay between photographing them and the oscilloscope was found to not always trigger on exactly the same part of the waveform with single shot operation. The single shot trigger control must be pressed twice with the alternate trace mode to give the two traces. The camera shutter is held open (exposure time control setting "B") while the traces are triggered. Figures 6.3, 6.6 and 6.7 used a continuous trace with camera exposure time setting n^{o} 8. Figures 6.4 and 6.5 used single shot with chop mode dual trace to give accurate relative phase indication together with minimised noise effect.

By feeding the amplifier output into the external timebase input of the oscilloscope and the test instrument output into the usual vertical input, traces similar to those described in Chapter 2 can be obtained. The test and distortion signals of figures 6.2 and 6.4 were used to show the effect of reducing I_Q below its optimum value at 2kHz. The results are shown in figures 6.6 and 6.7 for optimum and reduced I_Q respectively. Attempts to produce single line traces (i.e. traces without loops) by adjustment of phase shifts were not successful.

In an attempt to detect TID a square wave and sine wave were added at the amplifier input. Adjustment of the compensation components to cancel the 15kHz sine wave used gave figure 6.8. To find whether any intermodulation was taking place the instrument output with the square wave alone applied to the amplifier was observed and is shown in figure 6.9. The two distortion traces appear to be only the sum of the individual distortion components, suggesting that TID is not significant in this amplifier. It has been shown [6] that TID is caused by the overshoot in the input stage closed loop input signal resulting from the square wave component of the test signal. This overshoot occurs unless the open loop voltage gain –3dB frequency is greater then or equal to the –3dB frequency of a low pass first order filter through which the square wave has first been passed. The relevant open loop voltage gain figure is that measured from the input transistor base to the amplifier output in the case of a shunt feedback design. In the circuit of figure 6.1 the open loop gain is reduced by the 3300pF high frequency compensation capacitor in parallel with 5.6KΩ giving a –3dB frequency of 8.6kHz which is less than the 17.5kHz low pass filter –3dB frequency used. TID should therefore occur, but is not necessarily significant provided the input stage can handle the overshoot without becoming excessively non-linear. The overshoot can be observed by amplifying and displaying the waveform at the input transistor base with the square wave test signal input applied to the amplifier in the closed loop condition. This was done using the test instrument as a high gain low noise amplifier and the result is shown in figure 6.10 in which the overshoot is clearly visible. The result may, however, have been affected significantly by the loading due to the test instrument and a high input impedance buffer stage should ideally be used for such tests.

Finally the traces shown in figure 6.11 and 6.12 were obtained to illustrate the detection of phase modulation. 2kHz and 20kHz signals were used and the phase modulation caused by the 2kHz signal is indicated by the rise in the 20kHz component of the distortion waveform at certain parts of the 2kHz wave. Adjustment of the phase compensation capacitor VC_2 (figure 3.4) moved the position of the maximum 20kHz breakthrough from the position shown in figure 6.11 to that in figure 6.12. Adjustment of the gain by VR_2 could not move the position to this extent and this suggests that the observation is primarily of phase rather than amplitude modulation.

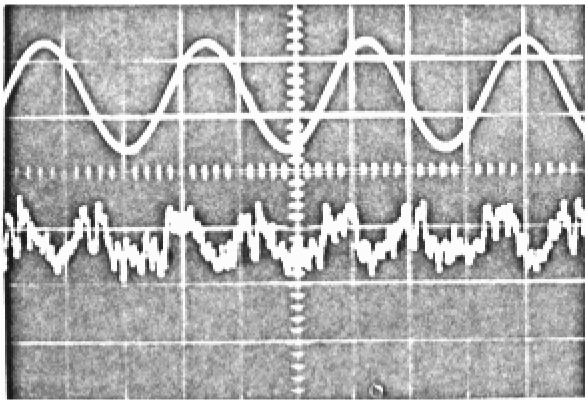


Fig. 6.2 - **Upper: Amplifier output, 20V p-p, 2kHz.**
Lower: Distortion, 120µV p-p (At amplifier O/P).
Peak distortion: 0.0006% (-104dB), IQ = 36mA.

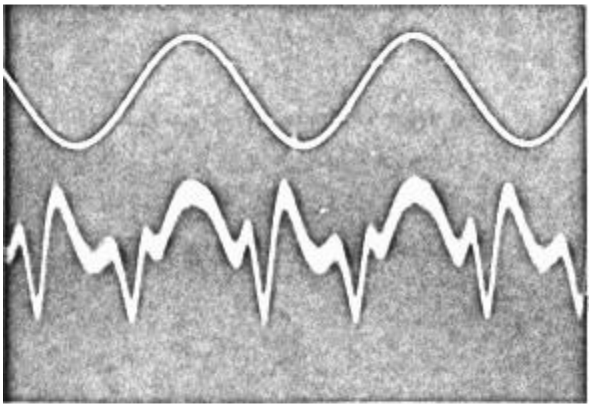


Fig. 6.3 - **Upper: Amplifier output, 20V p-p, 20kHz.**
Lower: Distortion, 250µV p-p.
Peak Distortion: 0.0013% (-98dB). I_Q = 40mA.

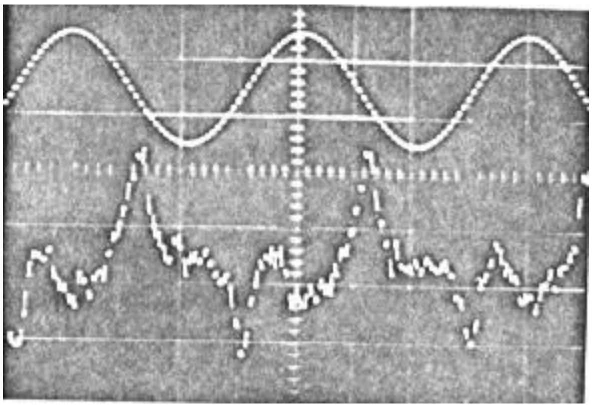


Fig. 6.4 - **Upper: Amplifier output, 20V p-p, 2kHz.**
Lower: Distortion, 240µV p-p.
Peak distortion: 0.0017% (-95dB). I_Q = 25mA

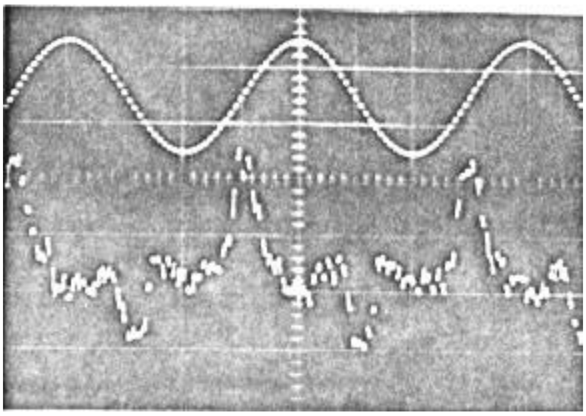


Fig. 6.5 - **Upper: Amplifier output, 20V p-p, 2kHz.**
Lower: Distortion, 340µV p-p.
Peak distortion: 0.0017% (-95dB). I_Q = 100mA

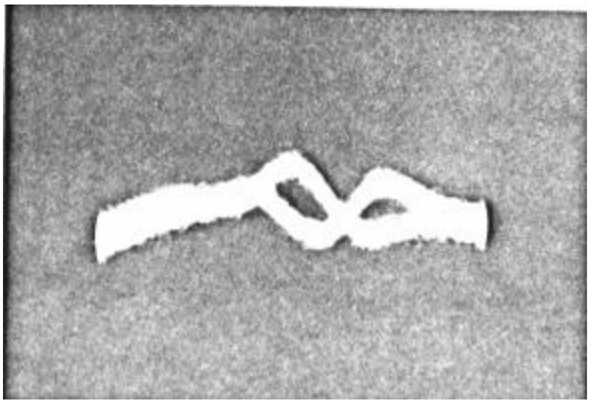


Fig. 6.6 - **Horizontal: Amplifier output, 20V p-p, 2KHz.**
Vertical: Distortion, 120µV p-p.
(Signals used are as in Figure 6.2). I_Q = 36mA

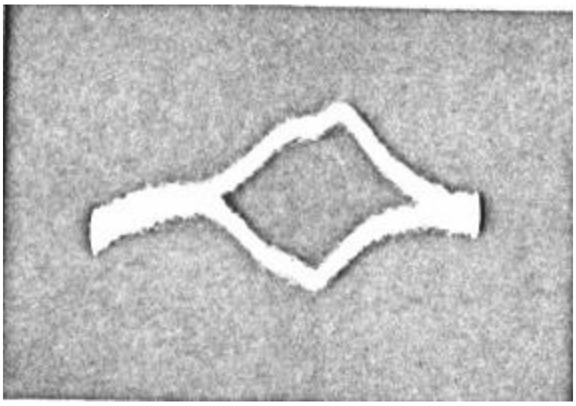


Fig. 6.7 - **Horizontal: Amplifier output, 20V p-p 2KHz.**
Vertical: Distortion, 340µV p-p.
(Signals used are as in figure 6.4) IQ = 25mA.

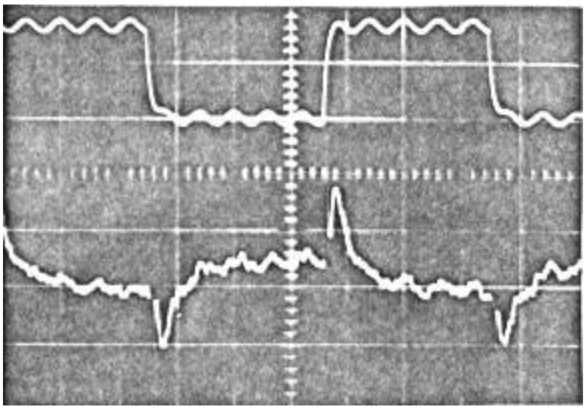


Fig. 6.8 - **Upper: Square wave, 16V p-p, 1.5KHz, plus sine wave, 2V p-p, 15KHz. (amp output)**
Lower: Test instrument output, equivalent to amplifier output distortion of 1mV/div.

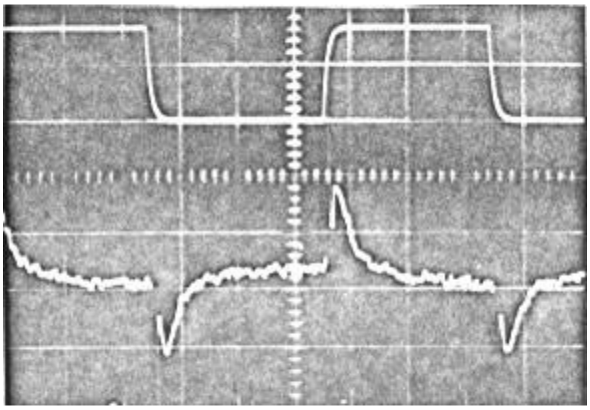


Fig. 6.9 - **Upper: Square wave, 16V p-p, 1.5kHz.**
Lower: Test instrument output as for figure 6.8.

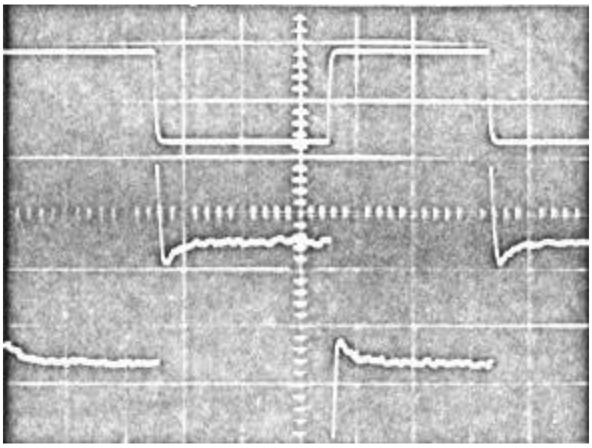


Fig. 6.10 - Upper: Square wave amplifier output, 16V p-p, 1.5 kHz.
Lower: Tr1 (figure 6.1) base voltage, 200µV/div.
Overshoot: 460µV amplitude.
Steady state: 210µV amplitude.
Percentage overshoot: 120%

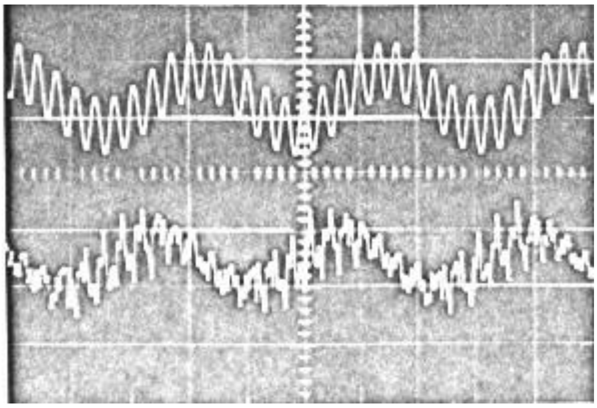


Fig. 6.11- Upper: Amplifier output, sum of 2kHz and 20kHz sine waves, each 10V p-p
Lower: Test instrument output, equivalent to amplifier output distortion of 1mV/div.

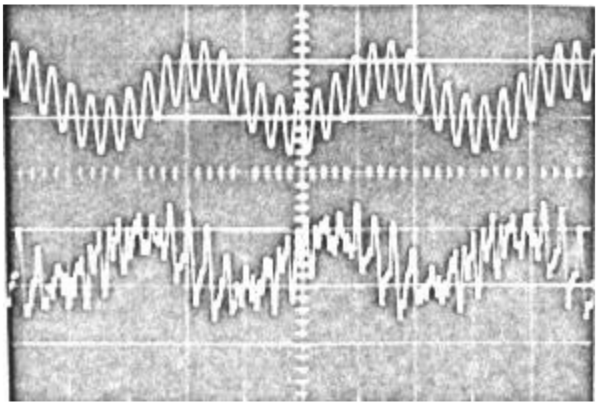


Fig. 6.12 – As for figure 6.11 but with phase compensation capacitor (VC_2) readjusted.

Conclusions

The measurements performed confirm that the direct comparison method is capable of a high level of performance. The harmonic distortion generated by the instrument developed is of the order of –140dB at 2kHz and –118dB at 20kHz (second harmonic) at an input signal peak amplitude of 1V when testing amplifiers the unity gain amplifier in the instrument (A_1) does not contribute significantly to the instrument distortion and the ultimate limit of measurement is set by the non-linearity of the passive components used, particularly the resistors. Even using a relatively high distortion signal generator (typically 0.5% THD) it has been shown that the limit to measurements of harmonic distortion due to breakthrough of harmonic distortion of the generator is not a significant factor in typical low distortion measurements.

The second harmonic instrument distortion measured using a common mode input signal of 1V peak amplitude at 20kHz is about 12dB higher than the figure calculated in Chapter 4. This suggests that either the analysis given was not very accurate or that the sources of distortion analysed are not the most important. The effects of the various barrier and diffusion capacitances in the circuit were not investigated in details, and the non-linearity of the metal oxide resistors used was assumed to be negligible. There are also other effects not mentioned. e.g. The currents in the circuit produce magnetic fields which can affect the resistivity of a conductor in various ways. The relevant phenomena are described in REF. 44 and include the magneto resistance effect. The significance of such effects is, however, difficult to estimate. Despite the disagreement between calculated and measured distortion values the performance of the circuit is more than adequate for testing of audio amplifiers.

Using a wave analyser to analyse the test instrument output it was shown in Chapter 5 that distortion components down to about –130dB below the fundamental could be measured (see Table 5.8), limited primarily by the noise from the amplifier being tested in the case of the tests on the 741 operational amplifier. The use of phase detection techniques could extend the range even further as very small effective noise bandwidths can then be used. The range of the wave analyser used was only about 80dB and it can be seen therefore that the test instrument can be used in conjunction with other test apparatus of only moderate specifications to obtain very high performance.

The effectiveness of the load effect cancellation arrangement developed in Section 3.4 is demonstrated by the waveforms in figures 5.10 and 5.11 in which the cancellation of a high level of distortion generated by the use of a load with non-linear and reactive components is shown. The distortion generated by the amplifier itself is the revealed. Only a sine wave test signal was used for this demonstration. The effectiveness when using wide band signals depends on the accuracy with which the output impedance of the particular amplifier to be tested can be compensated for. Without such compensation the use of a loudspeaker load during the testing of a power amplifier would considerably reduce the signal cancellation possible as described in REF. 25.

The tests carried out in the Chapter 6 were to illustrate some of the many uses of the instrument. The power amplifier designed for these tests had a very low level of distortion, which presented a severe test of the distortion waveform extraction ability of the instrument. The distortion waveform shown in figure 6.2 is at a level of –105dB relative to the fundamental (i.e. almost 60dB below the distortion level of the signal generator used) and yet features of the waveform are clearly visible, including the second harmonic and crossover effects. The variations of quiescent current in the amplifier shown that a reduction from the optimum value of about 30% gave a similar distortion amplitude to an increase of about 180%. This demonstrates that too much quiescent current is less serious than too little.

The traces in figures 6.6 and 6.7 show the error voltage as a function of amplifier output voltage as described in REF. 22. While this in theory gives a direct indication of the non-linearity of the transfer characteristic of the amplifier the interpretation of the practical results may not be easy. The separations of the trace into two lines is generally due phase shifts. The changes in separation in figure 6.6 suggest that the phase shift may be a function of signal amplitude. The barrier and diffusion capacitances in the amplifier can produce such an effect. The increased separation in figure 6.7 with reduced quiescent current may be due to the effects of charge storage in the bases of the output transistors (secondary crossover distortion). Further investigation of the results of such tests would, however, be required to learn the correct means of interpretation. The presence of variable phase shift is confirmed by the results given in figures 6.11 and 6.12 in which phase modulation is demonstrated.

The attempt to detect TID was not successful as shown in figures 6.8 and 6.9 where no intermodulation effects can be seen, although such effects may be masked by the high instrument output signal produced by the square wave. The square wave overshoot to occur is revealed by direct extraction of the input waveform as in figure 6.9. The overshoot amplitude is less than 0.5mV and is therefore not of sufficiently large value to cause significant input stage non-linearity.

The main limitation of the instrument in practical use was found to be the difficulty in providing accurate phase and gain compensation to match the characteristics of the amplifier being tested. The use of switched ranges and multi=turn controls would make the adjustments easier. The methods used during the experimental evaluation were very time consuming and are not to be recommended for more general use.

The 741 operational amplifier tested needed more effective time delay in the compensation circuit than was available from the simple network used. The use of more complex all-pass active filter circuits for the provision of time delay could solve this problem if required although these would use further amplifiers which would add distortion and may reduce the minimum levels which can be measured.

Distortion Measurement

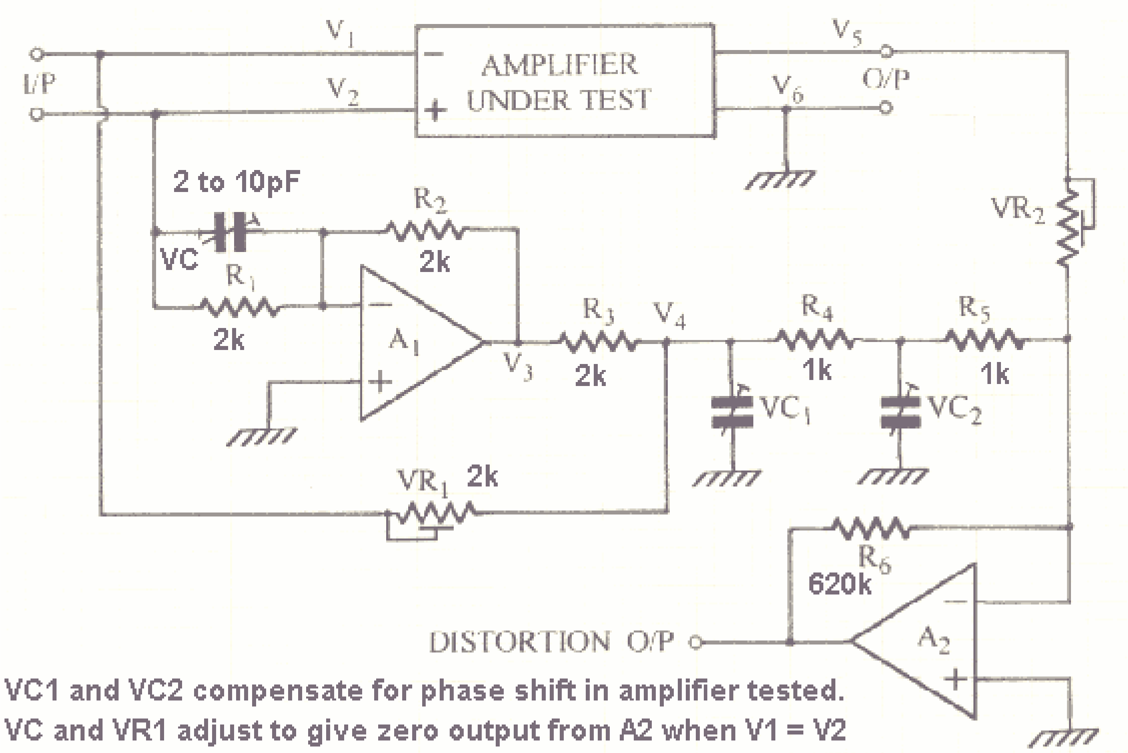
There are several widely used methods of measuring distortion in audio amplifiers, but probably the most useful is the direct comparison of the input and output signals, sometimes referred to as the bridge, or null method. This involves attenuating the output of the amplifier under test to the level of the input, and then adding or subtracting the input signal, depending on whether the amplifier is inverting or non-inverting respectively, so that the undistorted component of the output signal is cancelled leaving only distortion and noise.

There are several advantages to this method:

An ultra low distortion signal generator is not needed. The signal generator distortion and noise are cancelled along with the undistorted component. On one occasion I measured distortion components around 0.00001% using a signal generator with over 0.5% THD. It was the noise level, which eventually limited the accuracy, not the generator distortion.

Another advantage is that we are not restricted to using sine wave test signals, and with adequate amplitude and phase adjustment in the nulling circuit we can use a wide range of signals, including music, which will give a good indication of the performance in the intended application.
It is not intended to give a detailed description of the nulling method. This was the basis of my M.Sc. dissertation. I will merely present the relevant circuit diagrams, to show how the distortion traces published for my design were extracted.

The first diagram shows the basic circuit, which can be used for both inverting and non-inverting amplifiers, and also has the advantage of comparing the voltage difference between the output terminals with the voltage difference between the input terminals. In this way any earth line problems will not be missed, which is possible if we assume the output earth is identical to the input earth.

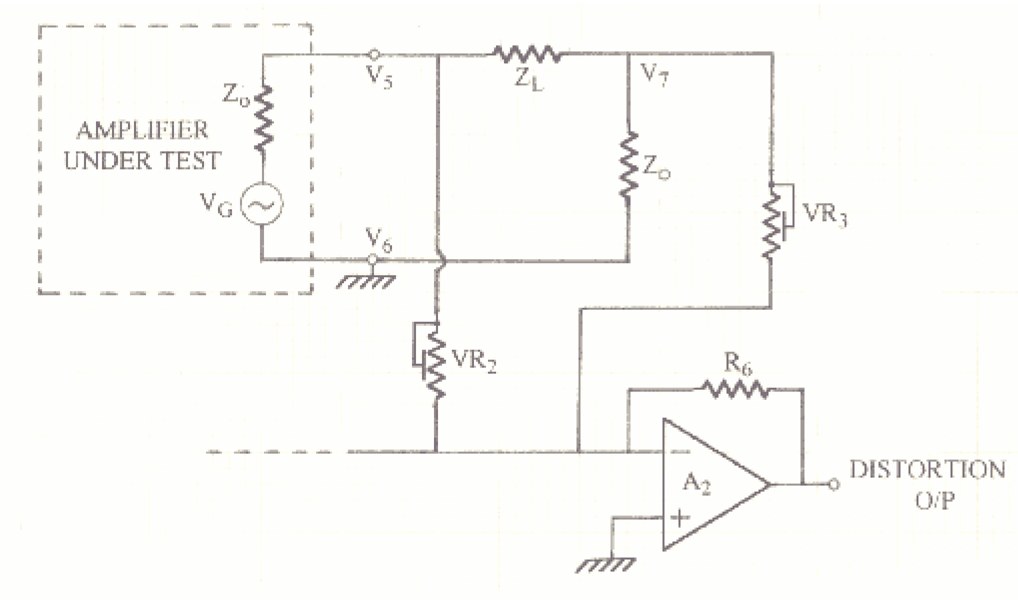


VC1 and VC2 compensate for phase shift in amplifier tested.
VC and VR1 adjust to give zero output from A2 when V1 = V2
If gain of amplifier under test = A, then VR2 = 6A kohm

Adjustment of the variable components is difficult and takes a lot of time and patience. Simple preset pots are not good enough, and for ultimate results a network of fixed resistors needs to be built up to get as close as possible to the final value needed, and just a very small trimmer used for final fine adjustment. VR_1 and VC are adjusted to give accurate common-mode rejection, and are set by applying a sine wave signal to both inputs V1 and V2 (with no test amplifier connected). With a 1kHz signal VR_1 is adjusted to give zero o/p from A_2, then using 20kHz VC is adjusted to give zero o/p. VR_1 may now need to be reset slightly, so that there is zero o/p at both 1kHz and 20kHz. When testing an amplifier, VR_2 is determined by the gain of the amplifier, and is about 6k multiplied by the amplifier gain. Again accurate adjustment is needed.

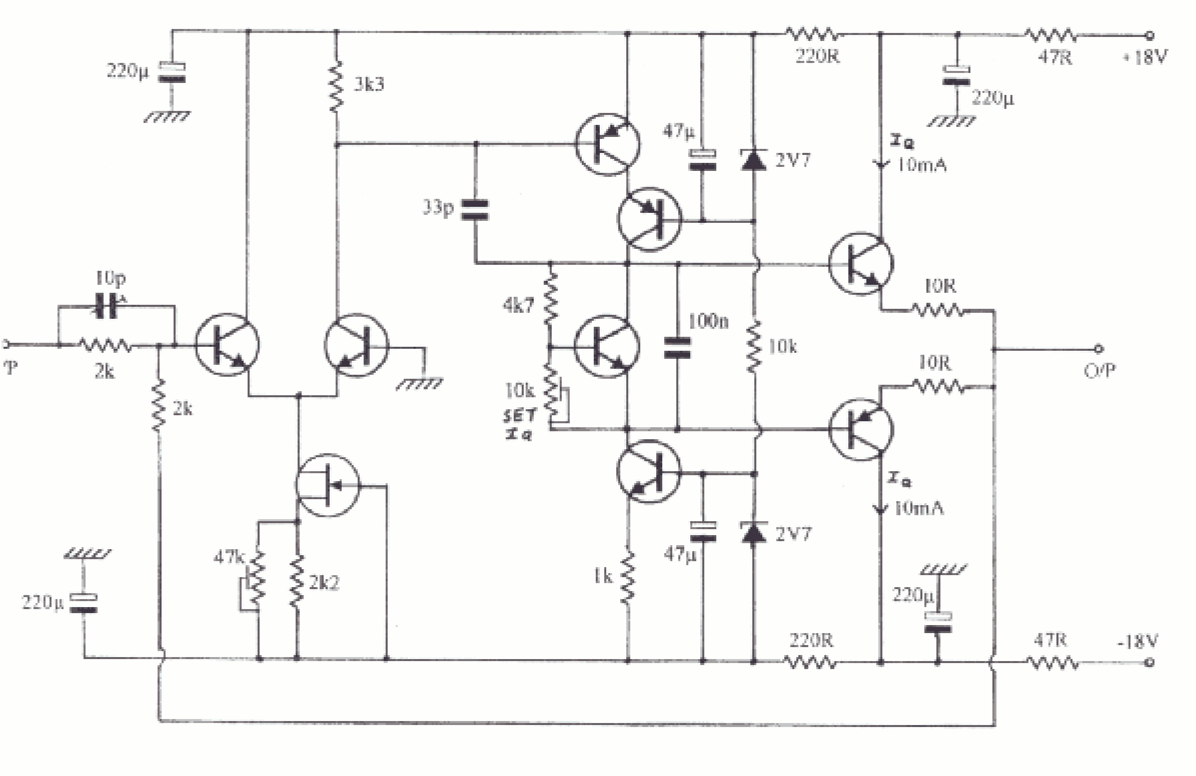
VC_1 and VC_2 adjust to compensate for phase shift in the amplifier, and are widely different for different designs, so little guidance can be given. Two variable controls are shown, but this is only really needed for accurate compensation over a wide bandwidth, e.g. if testing with a music signal. For single sine wave tests VC_2 can be omitted. For wide bandwidth testing the two capacitors may need to be similar or very different in value, depending on the nature of the phase shift in the amplifier, which may be close to a simple first order low pass filter, or may have something similar to a time delay element. (I found that an old 741 op-amp actually had more excess phase shift than a medium power class-B amplifier.) The real time delay in audio amplifiers is insignificant, the observed effect being just additional phase shift for a given attenuation compared to a first order low pass response at low attenuation. The second order filter has greater phase shift for a given attenuation and can adequately match the amplitude and phase response of most amplifiers over a wide bandwidth, though it is usually easier to do this if any low pass filter at the input of the amplifier is disabled so that we are not also having to compensate for this. The output amplifier A_2 is only amplifying the distortion voltage, so almost any low noise wide bandwidth op-amp will be good enough. R_6 determines the gain of the whole instrument, and 620k is typical, giving gain about 100 relative to the input signals V_1 and V_2

A problem with the simple phase adjustment used is that although it can reasonably well cope with the response of an amplifier driving a resistive load, if we use a real loudspeaker load the complex impedance will give variations in output level and phase sufficiently large to swamp the small distortion signals we are interested in. To compensate for the speaker impedance would be very difficult, but it is still possible to reduce the effects to a low level if we instead concentrate on the output impedance of the amplifier, and compensate for the voltage drop across this impedance, which is really the only effect of the load impedance. The output impedance of the amplifier will generally be much simpler than that of the speaker, and a simple modification to the previous test circuit can be used in which Z_0 is an impedance identical to the amplifier output impedance, and Z_L is the load impedance. The rest of the circuit is as in the previous diagram.

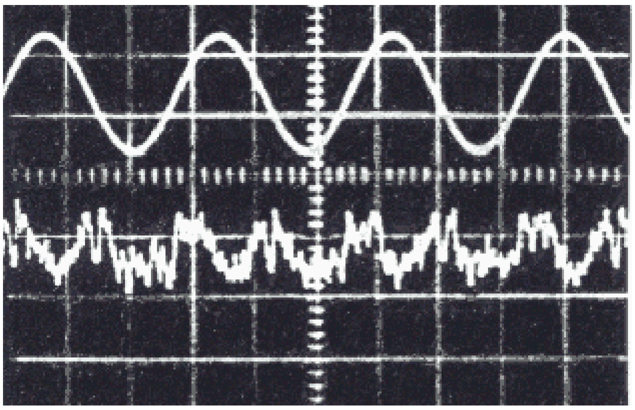


Adjustment of the controls can then give the necessary nulling. In practice some variation on this basic circuit may be more convenient. In tests a distortion trace was clearly revealed for an amplifier loaded with a highly non-linear and frequency dependant load, the effects of which had previously totally hidden the distortion.

The inverting amplifier used in the test circuit must of course be designed for the minimum possible distortion, and although op-amps with very low distortion are available, I found that a discrete component design could easily give better results by a factor of 10 or more at 20kHz. A circuit I have used is shown next, and although I do not have accurate distortion figures for it, the 20kHz distortion is certainly well below 0.0001% and at 1kHz is better than 0.00001%. The 47k control adjusts the total current in the input differential stage to set minimum distortion. For the input stage alone minimum distortion is with equal currents through the two transistors, but of course this is not necessarily the minimum for the whole amplifier, and in principle a slight offset in the input stage may generate a small non-linearity sufficient to cancel opposite non-linearity in later stages. This is why I left this as an adjustment rather than just set the transistor currents equal using the usual current mirror techniques. In practice I found adjusting this control gave little change in distortion, but I never investigated further.



Finally, to give an example of the distortion traces obtainable by this method, here is a trace using an earlier, inferior design, which even so shows that noise level is the eventual limiting factor in obtaining useful information. The distortion level here is around 0.0006% at 2kHz test frequency using a fairly standard class-B amplifier. I used the single shot facility on the oscilloscope to improve clarity in this case.



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