Graeme John Cohen
Philips Microelectronics
Hendon, S.A., Australia

Presented at
1984 Australian
Regional Convention
September 25-27
Melbourne

AN AUDIO ENGINEERING SOCIETY PREPRINT

DOUBLE BALANCED MICROPHONE AMPLIFIER

GRAEME J. COHEN

Assoc(Prof)IREE

PHILIPS MICROELECTRONICS, HENDON 5014, SOUTH AUSTRALIA

The combination of a balanced microphone transformer and a low noise amplifier is normally used for studio microphones. A design is shown to replace the transformer and associated amplifier. This incorporates the balanced and floating virtues of a transformer without the signal loss introduced ahead of a low noise amplifier. Gain and input stage current is adjustable. High common mode rejection is achieved with a balanced output and a very good noise figure at all gain settings.

0. INTRODUCTION

The microphone amplifier presented has been designed to meet the criteria of a microphone transformer and includes power gain which normally follows the transformer. Some of the important microphone characteristics are :-

1) Good differential balance with common mode rejection of input signals.

2) Impedance match with low loss.

3) Wide signal bandwidth.

4) Isolation of output from input.

5) Large signal handling without distortion.

The isolation of output from input and good common mode rejection are the obvious difficulties in substituting a transformer. A good transformer has a very good voltage isolation and can also have a floating input and/or output. It can also have a floating differential input and output, that is double balanced.

An attempt has been made to duplicate the above performance. The obvious limitation is the very large common mode signal the transformer can handle. The design presented can handle volts of common mode signal, has an exceptionally good noise figure and a wide dynamic range. A transformer can limit the dynamic range of an amplifier.

1. GENERAL DESCRIPTION

Two noise optimised single ended input stages are used, each using power driven feedback with a single gain setting resistor between them. A single resistor is used to optimise the current for the input stage to suit the source impedance which may vary from tens of ohms to several thousand ohms. The differential gain of the input stage can be varied from unity to more than five hundred whilst preserving dynamic range, common mode rejection and noise figure. The input stage uses two very low noise and well matched LM194 dual transistors in parallel. The design has provision for optimising and matching the input stage operating current.

The signal from each input transistor is amplified in the very low noise operational amplifier NE5532. One amplifier section being used for each input transistor. The good noise figure and balance is maintained along with a power capability for driving the feedback network.

From the two input stages the signal is combined in the classic differential to single ended amplifier configuration. This stage is duplicated with mirror like connections and uses another NE5532 dual amplifier. Power feedback is also used in this stage to preserve the low noise figure at low gain settings. The differential gain of this stage is two with a balanced output, and is capable of driving a balanced load of less than two thousand ohms at full output swing. Good common mode rejection is achieved on both outputs.

2. TYPICAL INPUT STAGES

Two conventional balanced input stages are shown in Figure 1 and Figure 2. The first design has emitter degeneration to define the gain. It has been shown elsewhere, Ref. 1, that collector-base modulation (early-effect) can cause considerable distortion. In Figure 1 it can be seen that the second transistor has a voltage gain. Due to the summing nature of the operational amplifier, the first transistor also has a voltage gain on its collector and hence collector voltage swing. This can cause early-effect distortion in both transistors, particularly with inductive sources. The gain is not well defined and can vary with frequency due to the source impedance particularly at higher gain settings. These gain variations can be small, but may not be tolerated. This microphone-amplifier interaction may be one of the reasons microphone amplifiers sound different or coloured.

A feedback stabilised input stage is shown in Figure 2. This design uses overall series feedback to each input transistor. The voltage swing on the collectors is now very small, either due to the large gain of the operational amplifier or the effect of shunt feedback to the collectors as shown. The gain is now well defined and the circuit is less dependant on source impedance effects.

In Figure 3 a typical instrumentation amplifier configuration is shown. This type of amplifier usually has a limited bandwidth as it is designed for high D.C. gain with low D.C. offset and drift. It can be seen that no coupling capacitors are needed as in the other two circuits. A good recovery from an input overload is thus obtained. The noise figure however, is not usually very good, having low operating currents and high impedances particularly in the feedback network. Also there are four base-emitter junctions in series between the inputs.

3. NEW INPUT STAGE

A new configuration has been developed to overcome many of the shortcomings used in previous designs. A balanced input requires a minimum of two transistors, one for each input. The problem is to have direct coupling of signal and feedback without the use of coupling capacitors and without having two more transistors, to make long-tailed pairs, on each input.

By using the well matched, for offset voltage (Vos = 25μV), pair of transistors LM194 and a dual operational amplifier a circuit has been developed as described below, and with reference to Figure 4.

The current through Q1 is adjusted by "sinking" its emitter current into the output of I.C.1. The collector current of Q1 sets up a voltage across R1 which is compared to a voltage across R2 which has been preset by R4. The two voltages on the inputs of I.C.1 are compared and the output voltage of I.C.1 sets a current through R5 so that the two input voltages of I.C.1 are equal.

The loop has negative feedback. The same conditions apply to Q2 and I.C.2 using R2, R3 and R7.

The important points to note are :-

1. *Resistor R4 sets both Q1 and Q2 current.*
2. *Resistors R1 and R3 are equal as are Resistors R5 and R7. Also I.C.1 and I.C.2 are a dual operational amplifier having low input offset voltage and tends to be equal being a dual amplifier.*
3. *Transistors Q1 and Q2 are a very well matched low noise pair of transistors having an extremely low offset.*

4. *The offset voltage on the output of I.C.1 and I.C.2 is similar and can be cancelled in a differential to single ended amplifier having good common mode rejection.*

5. *Resistor R6 sets the differential gain and is direct coupled due to the excellent balance in the amplifier.*

6. *Transistors Q1 and Q2 currents flow down the feedback resistors R5 and R7 and therefore there is no extra noise generators due to resistors or current sources.*

7. *Resistors R5 and R7 are a low a value as possible that the operational amplifier can drive and therefore R6 is very low. This ensures a very low noise figure.*

The virtues of this design is the excellent balance achieved both D.C. and A.C. The collector currents of Q1 and Q2 are made equal, due to both operational amplifiers acting as voltage comparators across R1 and R3 as to R2. Signal power is applied to the feedback network R5, R6 and R7 thus assuring a very good noise figure. Resistor R6 can become very low, depending on gain, and can be in the order of an ohm. The circuit, however, maintains D.C. stability due to the method of comparing collector currents.

The circuit has a response to D.C. with excellent stability even though low value resistors are used, thus minimising noise figure.

The input transistors are biased to a common voltage with equal resistors.

The outputs can be connected to a differential to single ended amplifier to cancel any offset voltage present and to give common mode rejection of input signals.

The gain is equal to R5 + R6 + R7 divided by R6.

4. NOISE COMPONENTS

The input transistors used have a base spreading resistance rb of approximately 10 ohms, Ref. 2. For a short circuit input the intrinsic emitter resistance is :-

$$r_e \; opt = \frac{Rs + R6 + Rb}{hfe^{\frac{1}{2}}}$$

With R6, gain setting resistor, set at 4 ohms to give 50dB overall gain, then r_e opt is approximately one ohm for each transistor. As the maximum current for the LM194 is 10 milliamps a reasonable current would be 4 milliamps giving r_e of approximately 6 ohms. It was decided two LM194 dual transistors connected in parallel would allow r_e to approach the optimum value for very low source resistances and effectively halving of rb. Several pairs were tried and all gave a 2dB improvement in noise. A total of 8 milliamps can be used for each input side while still allowing low currents to be set for high input source impedances. The noise performance of the LM194 (LM394) has been well documented in Ref. 3.

It has been shown in Ref. 4 and Ref. 5 that a transistor connected in common emitter or common base configuration has very nearly equal input noise components. The feedback network also produces a similar amount of noise when connected to either the base or emitter of a transistor. When connected to the emitter it reduces the stage gain by virtue of emitter degeneration. The feedback network therefore needs to be of low value and in this case is less than 4 ohms for 50dB gain.

Two BFW16A transistors were substituted for the two LM194 dual transistors. A large blocking capacitor was connected in series with the common feedback resistor R6 for the following noise test.

Figure 11 shows the noise performance for two BFW16A and for two parallel connected LM194 transistors. The BFW16A transistors give a 1.5dB improvement and not 2.5dB as shown, for higher frequencies, due to a lower overall gain obtained at this gain setting.

Figure 10 shows the excess flicker noise (1/f) of the high frequency specified BFW16A device. This transistor has a rb of 4 ohms and was included to show how close to the theoretical ideal the design has approached for source resistances of tens of ohms, see Ref. 4.

The following amplifiers, the NE5532 operational amplifier, contribute little to the overall noise for gains of 20dB or more. At unity gain the input transistors have a 10dB gain, compared to the NE5532, set by the feedback resistors as to the collector load resistors. The NE5532 amplifiers have been compensated with feedback capacitors, for the 10dB increase in loop gain at a unity gain setting for the stage. A capacitor of equal value has been added from the other input of the NE5532 to ground to give the same time constant on both inputs. The NE5532 is the dual of the NE5534 and is compensated for unity gain.

It can be seen that no current sources are used that can act as noise generators and that the operational amplifiers act as comparators for setting the collector current of the input transistors. The stage is unconditionally stable with a good phase margin. The square wave response shown in Figure 9, of the output signal shows little overshoot into a capacitive load and has no sign of slew induced distortion during rise and fall times.

The input transistors are cross connected in parallel, section A in parallel with section B of the other, to balance any second order effects of the dual transistors. It can be seen the remainder of the circuit is also balanced.

5. OUTPUT STAGE

The requirement of the output stage is to combine the differential signal, and not common mode signals, and to supply sufficient output capability for driving balanced lines. Also a differential output is useful for phase reversing of the signal.

Each amplifier is connected in the standard differential to single ended configuration with the second amplifier having the opposite phase output. The four resistors associated with each amplifier are equal and of low value for best noise performance at low gain settings. The resistors going to each non-inverting input have small value capacitors across them to compensate for phase shift in the amplifiers. The internal phase shift effectively reduces negative feedback to the inverting input at high frequencies. The overall common mode balance is maintained to 100KHz.

Output resistors of a low value are used to isolate capacitive loads from the feedback network and to present a 50 ohm resistive output impedance to terminate balanced output lines. The output drive capability is 20 volts RMS into 1K5 ohms at 100KHz.

The output at 20 volts peak to peak of 100KHz square wave is shown in figure 8 and of adding a capacitance of 3n3 is shown in Figure 9.

The output stages in the NE5532 operational amplifier are biased in class AB mode and no sign of crossover distortion is evident. The residual distortion of 100KHz sine wave is shown in the centre of Figure 7.

Measurement of the intermodulation distortion is shown in Figure 12. The output from a HP8903A Audio Analyser, used to reject the 60Hz and 750Hz signals, was observed on a HP3582A Spectrum analyser. Top of the screen is 60dB below the original 750Hz signal. The 60Hz signal is not fully rejected and accounts for half the 0.006% meter reading. Intermodulation products appear around the second and third harmonies of the 750Hz fundamental signal. Second harmonic distortion is >100dB below fundamental, <0.002%, and shows the overall balance of the amplifier.

Transient Intermodulation Distortion (TIM) cannot be detected and in view of the slew rate of 18V/µS, as measured, this is understandable, see Ref. 6.

6. CONCLUSION

The design of a balanced input and balanced output low noise microphone amplifier has been described and is a serious attempt to replace the microphone transformer.

Field tests in various studios have shown good results, with no apparent problems normally associated with transformerless input stages. When compared to some of the top recording desks the only obvious difference is the low noise figure of the amplifier. If used with phantom power of greater than eight volts then blocking capacitors need to be used. These should be large, and matched, to preserve the noise figure and common mode rejection.

An alternative output stage is suggested as an area for further development, Figure 6. This addition generates an output centre tap and allows either output terminal to be grounded giving a true floating, balanced output. This version has been tried and needs some high frequency phase correction. A dual operational amplifier version, with separate feedback paths, is being investigated to provide a more attractive alternative. This approach will give a centre tap output instead of only floating the outputs. Reference 10 has details of floating outputs.

A microcircuit hybrid of the amplifier is being developed as an alternative to the microphone transformer. A phase two microcircuit is being studied to give a true floating output with a centre tap.

APPENDIX 1 PERFORMANCE OF AMPLIFIER

Noise Voltage Spectral Density	*0.7nV/√Hz*
Noise Current Spectral Density	*1PA/√Hz*
Equivalent Input Noise in a 20KHz Bandwidth with Shorted Input	*100nV*
Corresponding Voltage Level	*-140dBV*

Total Harmonic Distortion at 20V RMS into 1K5 Balanced Load and 60dB Gain

i) 20 KHz	*0.03%*
ii) 100 KHz	*0.2 %*

Large Signal Bandwidth	*140 KHz*
Slew Rate	*18V/µS*
Overshoot into Capacitive Load	*10%*

Maximum Output into 1K5 Balanced Load at 100KHz	*20V RMS*
Corresponding Voltage Level	*+26 dBV*
Maximum Output into 600 Ohms Balanced Load at 100KHz	*14V RMS*
Common Mode Input Range	*+8V -10V*
Supply Voltage	*± 18V*
Supply Current with no Input	*30mA*

Gain Range Single Output	*0dB to 60dB*
Gain Range Balanced Output	*6dB to 66dB*

$$\text{Gain} = \frac{600}{RG} + 1 \quad \text{for a Single Output}$$

$$\text{Gain} = 2\left(\frac{600}{RG} + 1\right) \quad \text{for a Balanced Output}$$

Note, RG Minimum is 0.6 Ohm

$$\text{Input Stage Collector Current} = \frac{Vs\ \text{Total} - 3.4V}{4K9 + RI}$$

RI = 1K5 (5mA) Used for all Measurements.

APPENDIX 2 *TYPICAL APPLICATIONS*

	MICROPHONE/TAPE HEAD PREAMPLIFIER	*MICROPHONE/MOVING COIL CARTRIDGE PREAMPLIFIER*
Gain Required for Balanced Output	*20 dB*	*40 dB*
Output Stage Gain	*6 dB*	*6 dB*
Input Stage Gain	*14 dB*	*34 dB*
Feedback Resistor, RG	*150 Ω*	*12 Ω*
Feedback Noise Resistance	*120 Ω*	*11.5 Ω*
Allowing r_e of 10 and rb of 10 , then Feedback + r_e + rb	*140 Ω*	*31.5 Ω*
Equivalent Noise Voltage for 3dB Noise Figure in 20KHz Bandwidth and Allowing for Output Stage	*200 nV*	*140 nV*
Referred to Output	*2 μV*	*14 μV*
Maximum Output Level	*20 V*	*20 V*
Input Overload	*2 V*	*200 mV*
Dynamic Range	*140 dB*	*123 dB*
Source Resistance	*140 Ω*	*31.5 Ω*

REFERENCES

1. *TAYLOR, E.F.*
   "Distortion in low noise amplifiers"
   Wireless World - August, 1977.

2. *BASKIND, D. : RUBENS, H. : KELSON, G.*
   "The design and integration of a high performance voltage controlled attenuator".
   AES Preprint 1585 (D.2) November, 1977

3. *JENSEN, D.*
   "JE-990 Discrete operational amplifier".
   Journal of the AES VOL. 28 No. 1/2 Jan/Feb 1980

4. *NORDHOLT, E.H. : VAN VIERZEN, R.M.*
   "Ultra low noise preamplifier for moving coil phono cartridges".
   Journal of the AES VOL. 28 No. 4 April 1980

5. *GRAY, P.R. : MEYER, R.G.*
   "Analysis and design of analog integrated circuits".
   Wiley, New York. 1977

6. *CORDELL, R.R.*
   "Understanding transient intermodulation (TIM) distortion".
   Journal of Electrical and Electronics Engineering, Australia - I.E. Aust. and IREE Aust., Vol. 1 No. 4 December 1981

7. *JUNG, W.G. : STEPHENS, M.L. : TODD, C.C.*
   "Slewing induced distortion and its effect on audio amplifier performance".
   AES Preprint 1252 (G.7) May 1977

8. *HARVEY, B.*
   "An improved preamplifier design".
   AES Preprint 1793 (I-4) May 1981

9. *FOORD, A.*
   "Introduction to low noise amplifier design".
   Wireless World, April 1981

10. PONTIS, G.D.
"Floating a source output"
Hewlett-Packard Journal P.P. 12,13. August, 1980

11. Van Der ZIEL, A.
"Noise : Sources, Characterization, Measurement".
Prentice-Hall Inc., N.J. 1970

12 BURWEN, R.S.
"A low-noise high-output capacitor microphone system".
Journal of the AES Vol. 25 No. 5, May 1977

13. SIGNETICS
Analog data manual 1982.

14. NATIONAL SEMICONDUCTOR
Linear Databook.

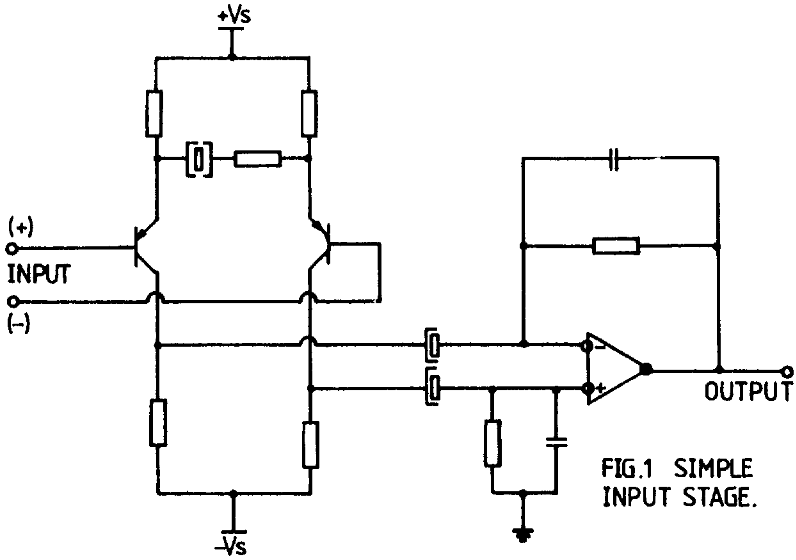

FIG.1 SIMPLE INPUT STAGE.

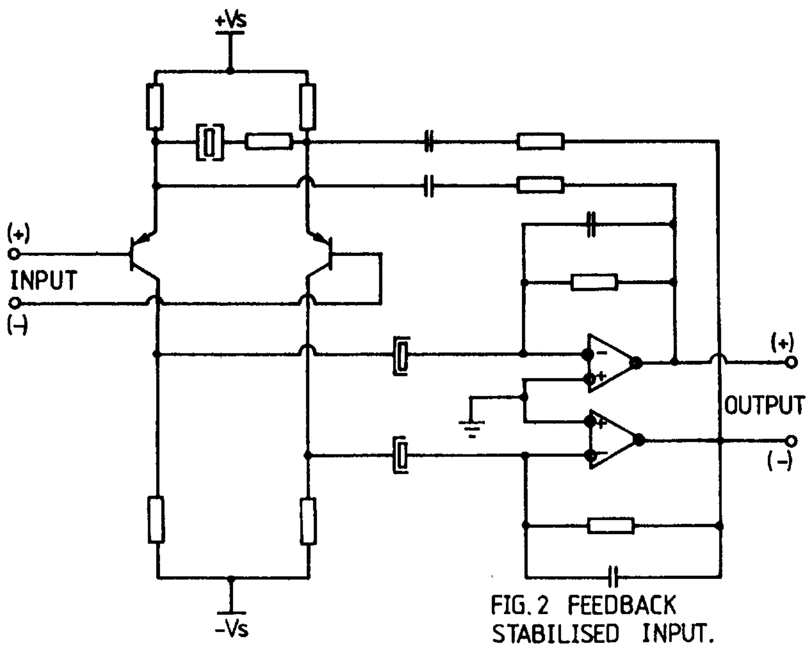

FIG. 2 FEEDBACK STABILISED INPUT.

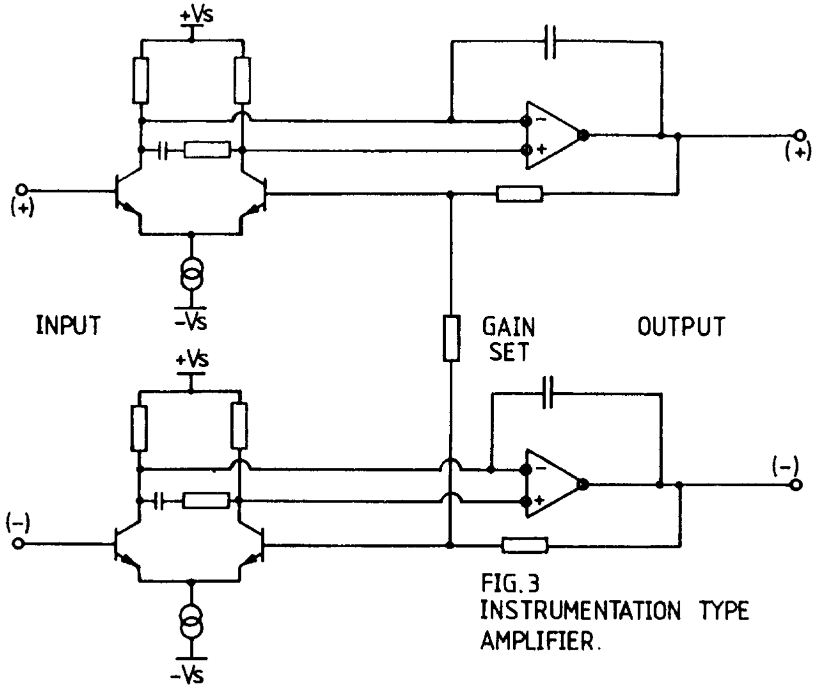

FIG. 3
INSTRUMENTATION TYPE AMPLIFIER.

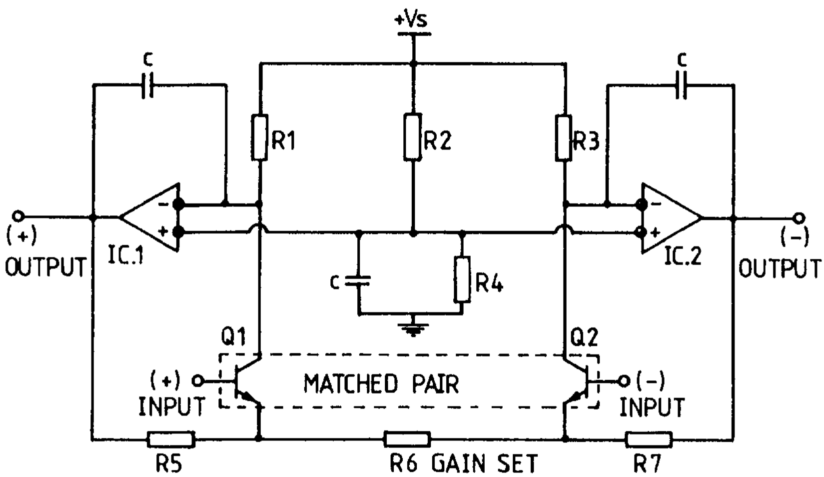

FIG. 4 NEW INPUT STAGE.

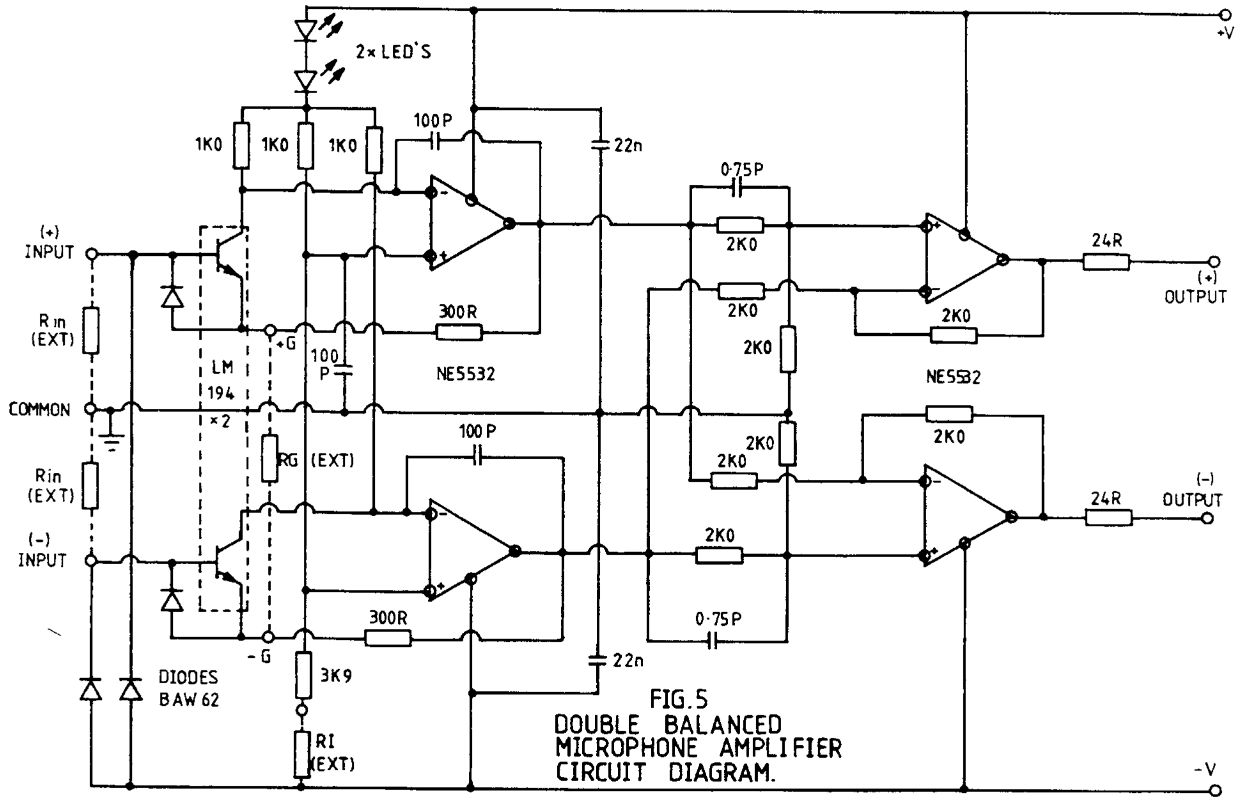

FIG.5
DOUBLE BALANCED
MICROPHONE AMPLIFIER
CIRCUIT DIAGRAM.

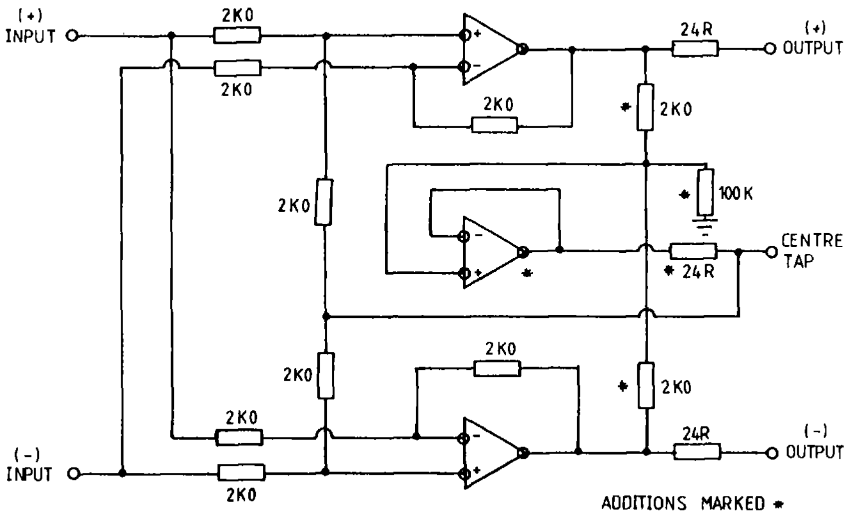

FIG. 6
PROPOSED OUTPUT STAGE
WITH CENTRE TAP.

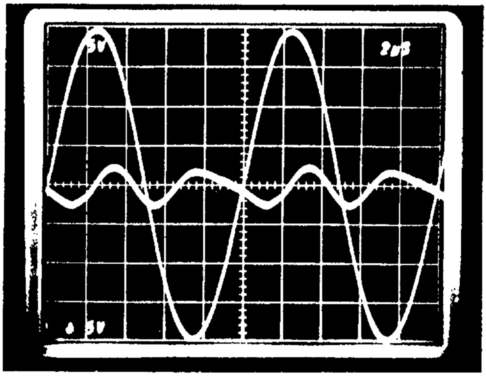

FIGURE 7

100 KHz Sine Wave

60 dB Gain

1K5 Balanced Load

40V P.P. Output

0.15% T.H.D.

Only Low Order Distortion Visible.

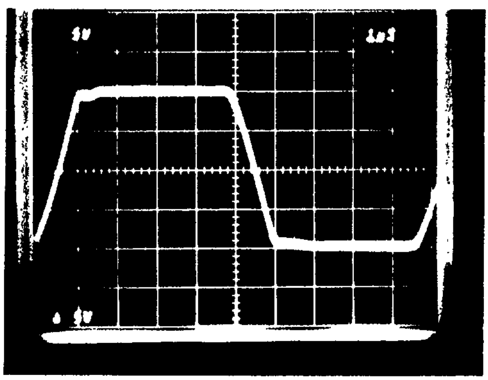

FIGURE 8

100 KHz Square Wave

60 dB Gain

1K5 Balanced Load

< 1µS Rise Time

No Slew Induced Distortion Visible.

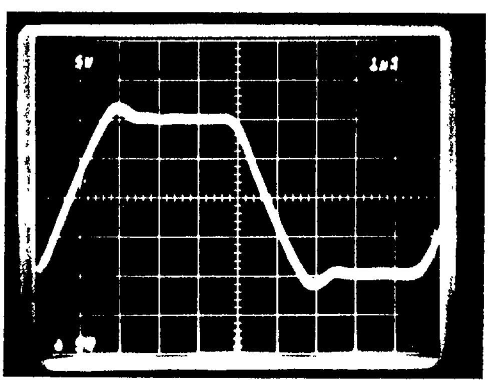

FIGURE 9

100 KHz Square Wave

60 dB Gain

1K5 and 3n3 Balanced Load

< 1.5µS Rise Time

< 10% Overshoot

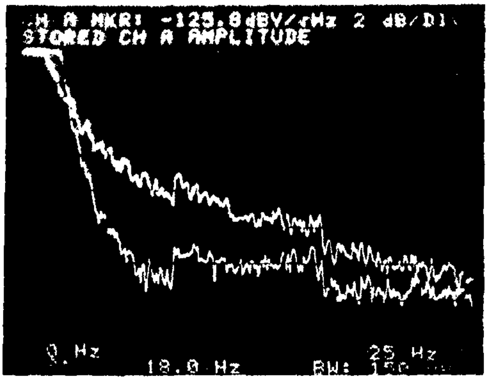

FIGURE 10

DC to 25 Hz Noise

54 dB Gain

Top Trace

2 x BFW16A add +1.5 dB for Lower Gain Devices

Bottom Trace

2 x LM194

4mA Collector Current
Input Short Circuit

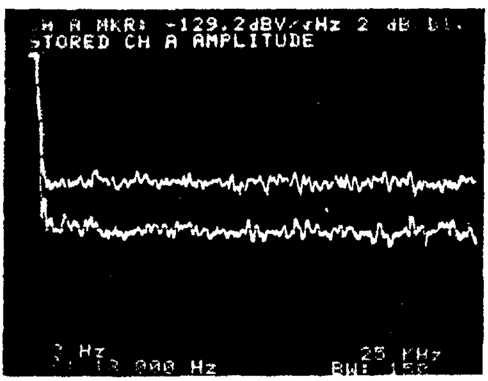

FIGURE 11

DC to 25 KHz Noise

54 dB Gain

Top Trace

2 x LM194

Bottom Trace

2 x BFW16A add +1.5 dB for Lower Gain Devices

4mA Collector Current
Input Short Circuit.

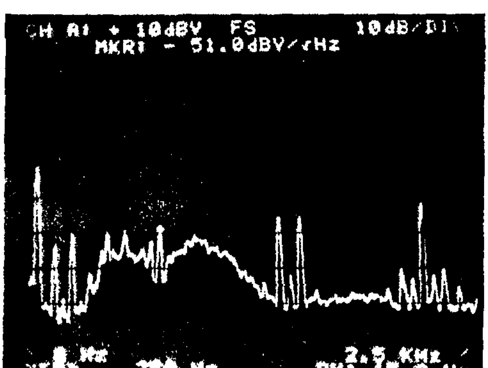

FIGURE 12

Intermodulation Distortion

60 Hz and 750 Hz

1:4 Amplitude

20 Volt Peak to Peak

1K5 Balanced Load

54 dB Gain

0.006% Total
Residual Components