TECHNICAL NOTE 124

SGS ATES

TDA 1054M - CASSETTE RECORDER PREAMPLIFIER WITH ALC

by G. Mercatelli, B. Sandoli

This note describes the characteristics and principal applications of the TDA 1054M multifunction integrated circuit.
The first part deals with a description of the circuit and of the problems associated with the choice of possible external connections. Some typical applications for recorders are then given and suggestions made for applications in which the TDA 1054M can be used to particular advantage.
The appendix deals with general problems of noise in preamplifiers and summarizes the typical electrical performance of the circuit.

INTRODUCTION

The TDA 1054M is a monolithic integrated circuit presented in a 16-pin dual in-line plastic package. The device is aimed principally at the tape recorder/player market as a preamplifier with automatic control of the recording level. Other applications in which the TDA 1054M can be used are dictaphones, dynamic compressors in telephone equipment, hi-fi preamplifiers, wired transmission receivers etc.
The device incorporates the following functions:

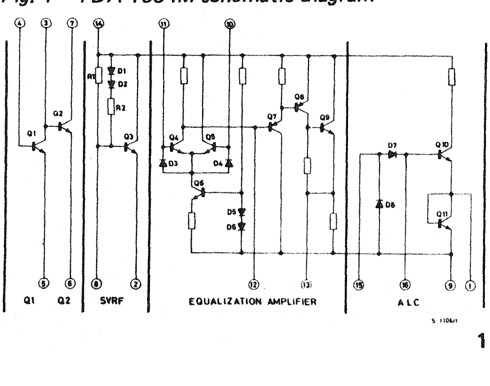
- low noise preamplifier
- operational amplifier with high open-loop gain (~60 dB)
- automatic record level control circuit (ALC)
- supply ripple rejection circuit (SVRF).

Before going into a detailed description of the circuit let us consider how these functions are used in a magnetic recorder. The low noise preamplifier (comparable to the best discrete component amplifiers) amplifies the signal from the head (or from the microphone when recording). This signal is then amplified again and equalized by the high gain operational amplifier (the high gain is necessary to obtain correct equalization and low distortion in closed-loop operation). When recording, the ALC circuit providing feedback round the operational amplifier acts to maintain constant the amplitude of the signal to be recorded (at the terminals of the record head) even when the input signal shows wide variations of level.
Finally the SVRF circuit allows high ripple rejection on the supply and rapid switch-on.
This section is characterized by low voltage drop ($\simeq$ 1V) between supply and filter output, which is of considerable importance since the whole system can thus be used on reduced supply voltages.

CIRCUIT DESCRIPTION

The internal equivalent circuit is shown in fig. 1.

Fig. 1 - TDA 1054M schematic diagram



Low-noise preamplifier

The input stage is formed by the transistors Q1 and Q2. Since these can be connected in different ways they allow a high performance/cost ratio for each application.
Usually in the design of a preamplifier it is necessary to meet a series of requirements which sometimes call for a compromise solution.
In particular the following features are required:

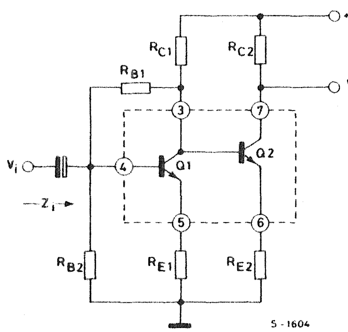
- high open-loop gain, which, when a feedback configuration is used, allows high stability and precision of the closed-loop gain, as well as low distortion
- the best possible signal to noise ratio
- high dynamic range
- excellent rejection of supply-ripple interference
- relatively high input impedances ($Z_i = 47\ k\Omega$)
- lowest possible output impedances.

Some connections of the input transistors in relation to the performance/cost ratio will now be analyzed and the characteristics of transistors Q1 and Q2, with regard to noise and current gain presented.

First preamplifier connection

The most economic method of achieving a high gain preamplifier is by connecting transistors Q1 and Q2 as shown in fig. 2.

Fig. 2 - Economical connection of transistors Q1 and Q2 in a high gain preamplifier



Although this circuit gives satisfactory performance it presents certain limitations regarding the following parameters:

- voltage gain: the voltage gain depends on the spread of current gain of Q2 ($h_{fe\,2}$)
- input impedance: the input impedance can vary as a function of the current gain of Q1 ($h_{fe\,1}$) or of Q2 ($h_{fe\,2}$).

In fact:

$$Z_i \simeq R_{B2} \,//\, \frac{R_{B1}}{G_{v1}} \,//\, h_{fe\,1}\,(r_{E1} + R_{E1})$$

where:

$$G_{v1} = \frac{R_{C1} \,//\, h_{fe\,2}\,(R_{E2} + r_{E2})}{R_{E1} + r_{E1}}$$

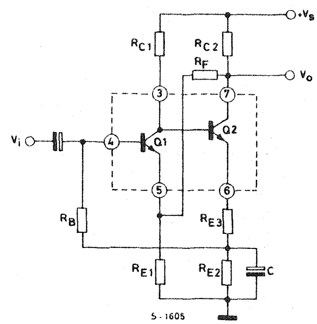
$$r_{E1} = \frac{26}{I_{C1}\ (mA)}\ (\Omega) \qquad r_{E2} = \frac{26}{I_{C2}\ (mA)}\ (\Omega)$$

- output impedance: the output impedance is relatively high ($\simeq R_{C2}$) because a compromise must be reached between noise problems (choice of I_C) and the obtainable dynamic range of the output signal
- stability of the working point of the transistors: by suitable choice of external component values the working point of the first transistor (Q1) is stabilized as regards both temperature variations and current gain spread ($h_{fe\,1}$). The working point of the second transistor (Q2) is stabilized only as regards temperature variation and not as regards spread in the current gain ($h_{fe\,2}$).

Second preamplifier connection

A second method of connecting the transistors Q1 and Q2 is shown in fig. 3.

Fig. 3 - A second method of connecting Q1 and Q2, with better performance than the circuit of fig. 2



This arrangement, which is less economic than the previous one, allows better results to be obtained for the allowing parameters:

- closed-loop voltage gain: the closed-loop voltage gain does not depend on the characteristics of the transistors but on the ratio R_F/R_{E1}
- input impedance: the input impedance is independent of the current gain of the transistors ($h_{fe\,1}, h_{fe\,2}$). By choosing suitable values for the

external components the following is obtained:

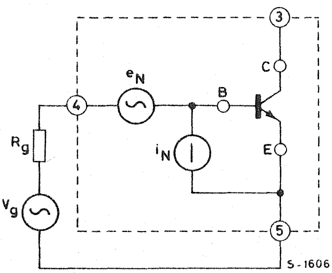
$$Z_i \simeq R_B \quad \text{(fig. 3)}$$

- output impedance: the output impedance is much lower than R_{C2}. In fact by using shunt feedback the open-loop output impedance ($\triangleq R_{C2} \, // \, R_F$) is divided by the loop gain ($A \cdot \beta$) where A is the open-loop gain and $\beta = R_{E1}/R_F$
- working point stability: the working point is stabilized with regard to both temperature variations and the spread of current gains $h_{fe\,1}$ and $h_{fe\,2}$
- ripple rejection: any interference (noise, ripple, etc.) appearing on the collector of Q1 shows up $A \cdot \beta$ times smaller at the output than it would with the previous circuit (fig. 2).

Current noise and gain of transistors Q1 and Q2

As is well known all transistors generate noise. A "noisy" transistor can be represented by an equivalent circuit in which there appears a noise-free transistor with voltage and current noise generators connected at its input as shown in fig. 4.

Fig. 4 - Equivalent circuit of noise voltage and noise current generators at transistor input



This equivalent circuit is particularly useful in the design phase since, knowing i_N and e_N (usually data from the catalogue) it is possible to choose the bias current, the voltage and the source resistance (these last two parameters are usually interdependent) so that the maximum signal to noise ratio at the input of the transistor can be obtained. Figs. 5 and 6 show the equivalent spot-noise voltage and the equivalent spot-noise current as a function of the frequency with the collector current of Q1 as a parameter (the same characteristics are also valid for Q2).
The term "spot" is used to indicate that the measurement is made for each frequency in a band 1 Hz wide.

Fig. 5 - Equivalent input noise voltage vs. frequency (input transistor Q1)

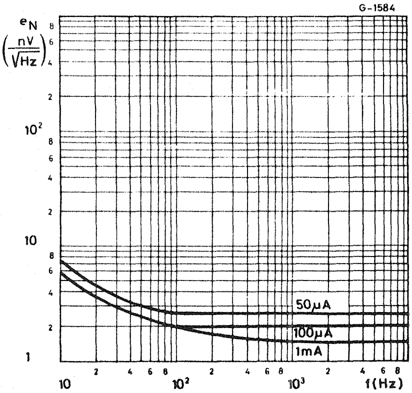
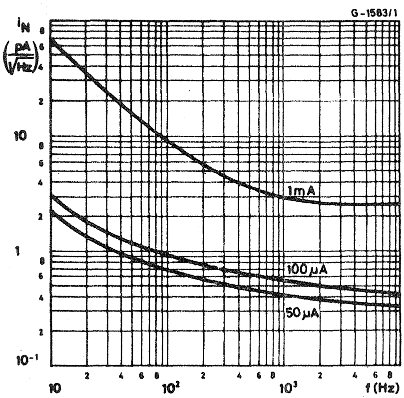


Fig. 6 - Equivalent input noise current vs. frequency (input transistor Q1)



In other words these graphs give the effective value of the voltage and current noise which the transistor generates at a given frequency. Fig. 7 shows the equivalent spot noise voltage e_N and the spot noise current i_N, as functions of the collector current, with frequency as a parameter. Fig. 8 shows the constant noise figure contours as a function of the source resistance and the collector current of Q1. It should be noted that in this case the wide-band noise figure was measured. Fig. 8 shows the excellent noise performance of transistors Q1 and Q2; in fact with $I_C = 100\ \mu A$ and $Rg = 4.7\ k\Omega$ the typical noise figure is 0.5 dB, and reaches an overall maximum of 4 dB. The values of e_N and i_N are comparable to those of the best discrete devices. In fig. 9 both the value of the optimum source resistance and the corresponding minimum spot-noise figure are shown as functions of the collector current of Q1.

Fig. 7 - Equivalent input spot noise voltage and current vs. bias current (input transistor Q1)

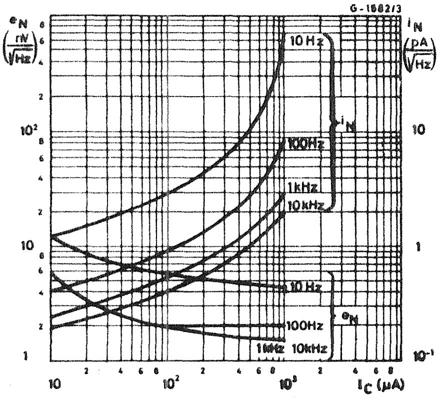


Fig. 8 - Noise figure vs. bias current (input transistor Q1)

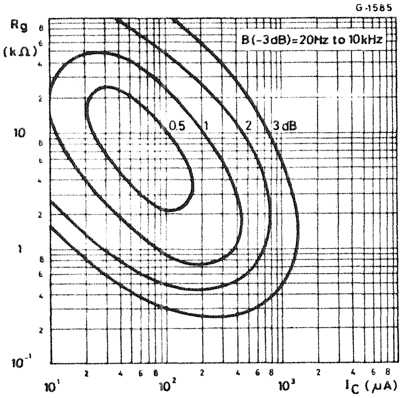
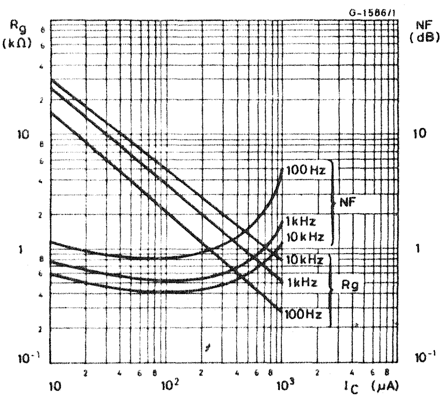
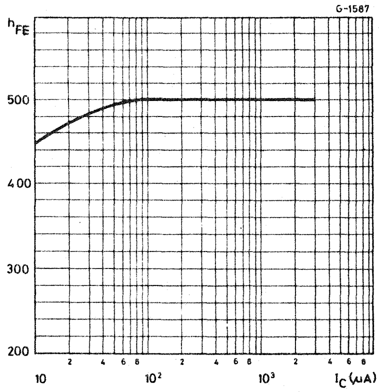


Fig. 9 - Optimum source resistance and minimum NF vs. bias current (input transistor Q1)



Finally fig. 10 shows the current gain h_{FE} as a function of the quiescent collector current. As can be seen, even with low current levels h_{FE} has a high value and the typical value (500) is exceptional for an integrated circuit.

Fig. 10 - Current gain vs. collector current (input transistor Q1)



Operational amplifier

The operational amplifier (fig. 1) consists essentially of a differential stage with the emitters driven by a current generator. The collector circuit is decoupled from the output transistors by means of an emitter follower stage.

This ensures that the gain of the differential stage depends entirely on the value of the collector resistance (pin 12).

The stage formed by the two final transistors operates like an equivalent PNP transistor with voltage gain. The output is taken from the emitter.

Figs. 11 and 12 show the gain and the phase response of the open-loop amplifier as a function of the frequency.

Fig. 11 also shows compensated open-loop gains which can be obtained by inserting a compensation capacitor between pin 12 and pin 13.

It should be noted that if closed-loop gains greater than 40 dB are used (using a purely resistive feedback network) no compensation is required.

Still referring to fig. 11, the upper cutoff frequency for any closed-loop gain and for a given compensation capacitor can be found (the load resistor between pin 13 and ground is 560 Ω).

SVRF stage

This stage (fig. 13) allows high rejection of supply interference to be obtained without the usual problem associated with conventional filters. In particular, for the same attenuation, the SVRF system allows very short charging times for the filter capacitor and low voltage drop ($\simeq$ 1V) between the supply and the filter output.

Fig. 11 - Open loop gain vs. frequency (equalization amplifier)

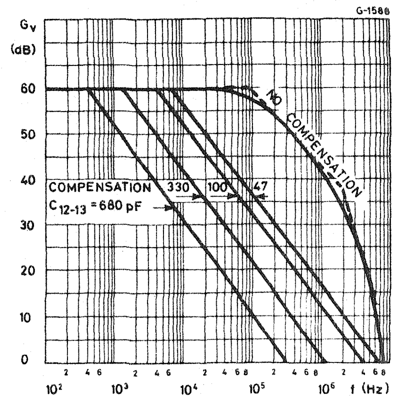


Fig. 12 - Open loop phase response vs. frequency (equalization amplifier)

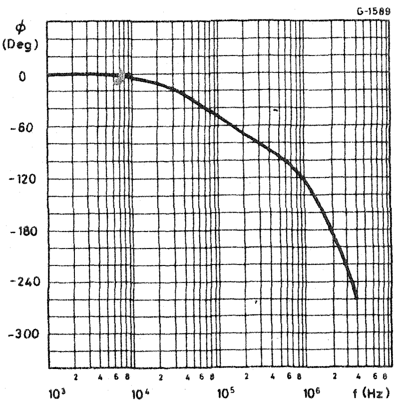
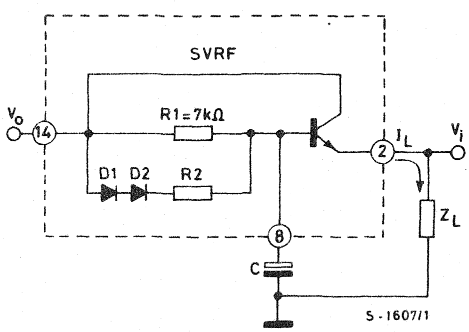


Fig. 13 - Equivalent circuit of the SVRF stage



At the moment of switch-on the capacitor C is charged with a very short time constant by means of the diodes and R2 (R2 $\simeq$ 100 Ω).
When the voltage across C is equal to (V_s - 2 V_{BE}) diodes D1 and D2 are cutoff and a filter section is formed by R1 and C. At the output of this filter, the interference attenuation is approximately equal to:

$$\frac{X_c\ (\Omega)}{7} \cdot 10^{-3} \qquad (X_c \ll 7 \cdot 10^{-3})$$

At the terminals of the load (Z_L) there is now an accurately filtered voltage which is 1V less than the value of V_s.
Note that the current I_L supplied to the load must not be greater than 5 to 10 mA, so that the base current of the transistor is prevented from generating across R1 a voltage sufficient to switch-on diodes D1 and D2. In fact, in this case the filter action is cutoff. The same happens if the peak amplitude of the disturbance present at pin 14 is greater than 1V, in which case the supply voltage has to be prefiltered.

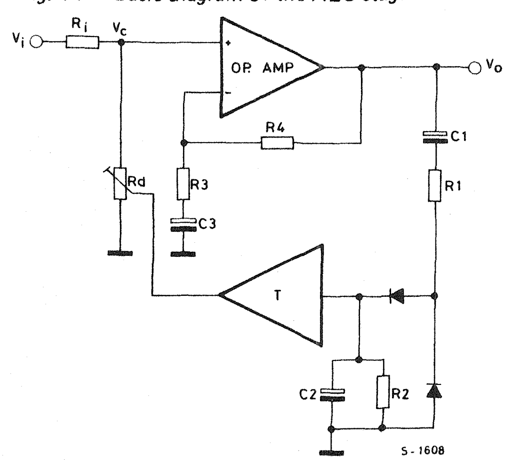
Automatic level control system (ALC)

This system maintains the level of the signal to be recorded at a value which prevents saturation of the tape and which optimizes the signal to noise ratio even when there are notable variations in the input signal.

Principle of automatic level control

Before describing the ALC circuit of the TDA 1054M it is worth describing the operation of the automatic level control as a system and the problems connected with it. A diagram showing the basis of operation is given in fig. 14.

Fig. 14 - Basic diagram of the ALC stage



This consists of an amplifier (OP. AMP.) having constant gain ($G_v = 1 + \frac{R4}{R3}$), which in feedback transforms output signal level information (usually by means of a peak-to-peak detector) into a continuous voltage which drives the networks indicated by T and Rd.

The element T transforms the continuous voltage level into a signal capable of modifying the circuit conditions symbolized by variable resistor Rd. The value assumed by the resistor Rd is a function of the output signal level V_o and is such that the voltage V_c at the input of the OP. AMP. is constant, even when variations of V_i are present. Obviously if V_o is less than a certain value the system is not controlled.

In this case:

$$V_i \simeq V_c = \frac{V_o}{G_v}$$

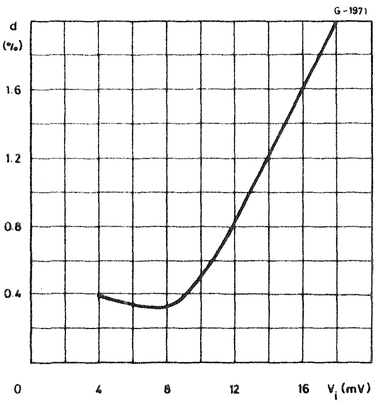
(G_v is the gain of the OP. AMP.).

For the TDA 1054M the value of V_o below which the system is not controlled is around 1 Vrms.

This figure determines the gain which the OP. AMP. must have in order that the system is controlled at a certain input voltage V_i.

With this in mind, it is advisable to fix a threshold voltage $V_i = 10$ mV. In fact, as can be seen from fig. 15 distortion occurs for values less than 10 mV.

Fig. 15 - Distortion of the ALC stage vs. input voltage (pin 11)



Finally the OP. AMP. should have a gain of 40 dB.

Let us now consider the speed of response of the system (when controlled) to positive and negative changes of the input signal i.e. the limiting time, the time for return to nominal level (1 Vrms) and the recovery time.

– Limiting time, and time for return to nominal level.

Let us suppose that at a certain moment t_0, the input signal increases by $+\Delta V_i$ as shown in fig. 16. Usually such an increase drives the OP. AMP. into saturation and the time for which it remains in this condition is called the limiting time (t_l).

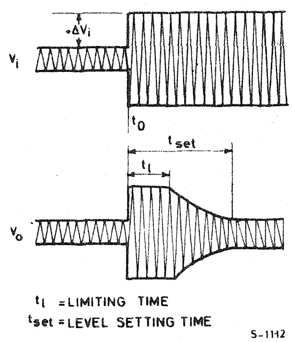
t_l depends on the relationship between the external capacitances, the time constant $\tau = R1 \cdot C1$, the supply voltage and the signal variation.

The criteria for choosing the length of t_l are the result of several compromises. In particular, if t_l is too long, there will be audible distortion during playback (during t_l the output is a square wave), and if it is too short, the sensation of increased level will be lost while dynamic compression phenomena and instability may occur.

The time for return to nominal level is defined as the total time between the instant t_0 and the instant in which the output reassumes the nominal value. This time (t_s) is roughly equal to 5 t_l.

On the basis of tests carried out it has been found that if a musical signal with high dynamic range ($\Delta V_i = +40$ dB) is to be recorded, the best value of t_s is between 200 and 300 ms. In the case of dictaphones or other systems where the signals have lower dynamic range, lower times, for example 20 to 50 ms, are sufficient.

Fig. 16 - Limiting and level setting time

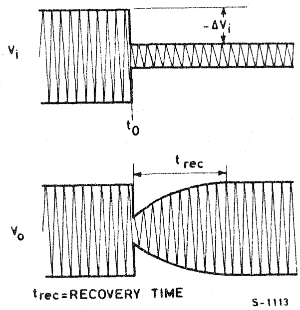


– Recovery time.

Let us now suppose that at the instant t_0 the input signal decreases by $-\Delta V_i$ (fig. 17). The recovery time (t_{rec}) is defined as the time between the instant t_0 and the instant in which the output signal returns to the nominal level. This time depends essentially on the discharge time constant of R2 · C2 (see fig. 14) and on the size of the step $-\Delta V_i$. In this case too, if this time is too long the signal to noise ratio on the tape deteriorates.

If it is too short the sensation of the low signal level is lost during playback. Usually the time is about 2 to 3 minutes ($-\Delta V_i = 40$ dB).

Fig. 17 - Recovery time



The ALC system of the TDA 1054M

Let us now analyze the circuit of the TDA 1054M which, providing feedback around the operational amplifier, allows automatic level control to be obtained.

Fig. 18 shows the particular arrangement employed where the part within the broken line is the circuit concerned and which, for convenience, will be called the ALC. The peak-to-peak detector of fig. 14 is now formed by D7 and D8.

The system which allows a dynamic resistance, varying with the DC voltage level at pin 16 (i.e. inversely proportional to the OP. AMP. output signal), to be seen at pin 1, is formed by transistors Q10 and Q11.

In order to gain a better understanding of the dynamic attenuator, the circuit formed by D7 and D8 seen at pin 1 can be presented as in fig. 19.

Fig. 18 - Diagram of ALC stage with IC circuit shown

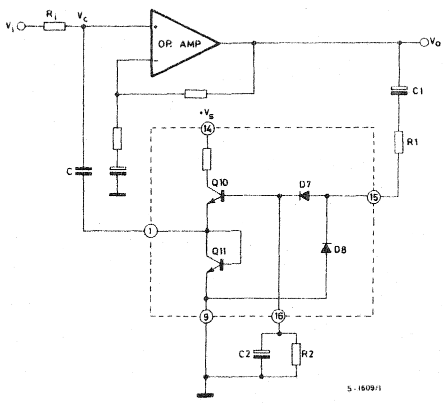
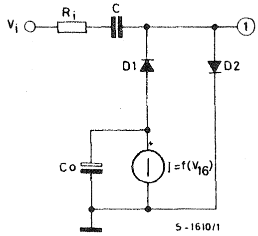


Fig. 19 - Basic dynamic attenuator circuit



As can be seen there is a pair of diodes in anti-parallel supplied by a current generator whose output is proportional to the DC voltage present at pin 16 (Co is a short circuit for the changing components of the signal). Diodes D1, D2 act as a dynamic resistance to the negative and positive half-waves respectively.

Thanks to the excellent matching of transistors Q10 and Q11 both electrically and from the layout point of view, it is possible to obtain very symmetrical behaviour for both half-waves minimising second harmonic distortion.

In comparison with discrete systems this system allows lower distortion, wide control range, long recovery times and good signal to noise ratio (at pin 1) to be obtained.

Let us now make some final comments on the ALC circuit. It should be noted that the generator resistance R_i (see fig. 18) has no influence on the controlled voltage value V_C, although its value should be between 4.7 kΩ and 47 kΩ.

The lower limit is determined by the minimum dynamic resistance at pin 1 of 4.7 Ω and therefore to have a control range of 60 dB for the input signal V_i, R_i must be greater than 4.7 kΩ.

The upper limit results from the necessity to limit the attenuation of the signal by the input impedance of the OP. AMP.

It has already been said that the output voltage at which the ALC goes into operation is approximately 1 Vrms. It is, however, possible to control the output voltage of the operational amplifier at a higher level by drawing the ALC control voltage (pin 15) not directly from pin 13, but from a divider between pin 13 and ground. The output voltage V_o will therefore be given by:

$$V_o\,\text{rms} \simeq 1 + \frac{R'}{R''}$$

where R" is the resistance between pin 15 and ground, and R' is the resistance between pin 15 and pin 13. Obviously the maximum value of V_o depends on the supply voltage V_s ($V_o = 2V$ with $V_s = 9V$).

In order to maintain the limiting time and the time for return to nominal value equal to those obtainable with the circuit in fig. 18, R'/ /R"

should equal R1.
Finally, as can be seen in fig. 18, pin 1 must be DC decoupled from the rest of the circuit.

APPLICATIONS

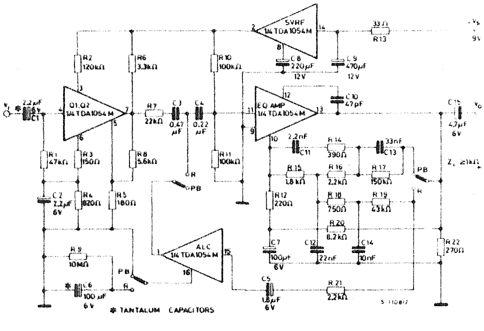
The TDA 1054M is a monolithic integrated circuit designed for use mainly in magnetic recorders with automatic control of the recording level. Because of its versatility however, the device can be used to advantage in all applications where high gain, low noise and low distortion are required.
In this technical note we suggest the following applications.

- Cassette recorder
- Low-cost cassette recorder
- Complete portable recorder circuit
- Hi-Fi preamplifier with feedback tone control
- Hi-Fi preamplifier with passive tone control
- Stereo preamplifier (using only one TDA 1054M) for ceramic pick-ups
- Stereo application (two TDA 1054M)
- Wired transmission receiver
- Dynamic range compressor.

Cassette recorder

The circuit is shown in fig. 20. This circuit allows high performance to be obtained both in recording and playback. In particular, it meets DIN standards 45513, 45511, 45500 for signal to noise ratio, distortion, equalization and input impedance. The general electrical characteristics are given in the appendix.

Fig. 20 - Application circuit for battery/mains tape and cassette recorder/player



Preamplifier

As can be seen in the diagram the recording and playback sections have a common preamplifier. This preamplifier, which has a flat frequency response, is formed by transistors Q1 and Q2 connected in a shunt feedback circuit. This allows the following characteristics to be obtained: transistor working point stability, high input impedance (more or less defined by R1) low output impedance, low distortion, good supply ripple rejection and closed-loop gain depending only on the relationship R8/R5. In this case the gain is fixed at 30 dB.
The nominal input sensitivity is 0.3 mV, although the preamplifier can accept input signals up to 40 mV before saturation occurs. This is necessary because during recording the input dynamic range can be as much as 40 dB.
Particular attention has been paid to the noise performance of the preamplifier, both in choosing the working point of the transistors and by optimizing the values of the external components.

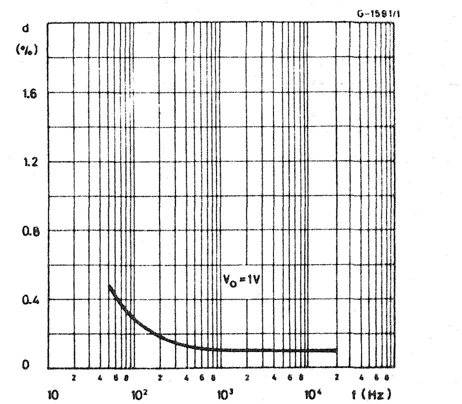
SVRF

The operation of this stage has already been described. It is worth mentioning, however, that there is an attenuation of the ripple (f = 100 Hz) between the supply and pin 2, of 80 dB with a loss of only 0.8 V with respect to supply voltage, (including the filter formed by the 33 Ω resistor and the 470 μF capacitor).

Performance during playback

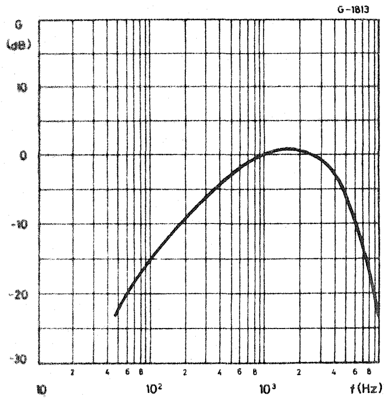
Distortion as a function of frequency is shown in fig. 21. The operational amplifier is used to achieve the required voltage gain between input and output and to obtain correct equalization of the signal as a function of the frequency.

Fig. 21 - Distortion vs. frequency in the circuit of fig. 20 (playback)



The voltage induced at the head terminals by a tape recorded with constant current and running at a speed of 4.75 cm/s, has a frequency dependence as shown in fig. 22.

Fig. 22 - Relative frequency response of playback head (tape recorded with constant current and at a speed of 4.75 cm/s)

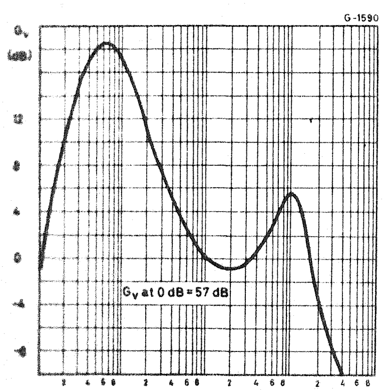


The signal at the terminals increases linearly as a function of the frequency with a slope of 20 dB per decade until about 1.5 kHz, after which it decreases rapidly because of losses due to the finite section of the head gap, the selfdemagnetization of the tape, the thickness of the magnetic coating on the tape, the finite distance between tape and gap and finally due to electrical losses in the copper and iron of the head.

It is therefore clear that in order to have a flat frequency response the amplifier must be equalized correctly.

In the typical application (fig. 20) the relative gain as a function of frequency has a curve as shown in fig. 23. This is obtained with the network formed by C13, R17, C11, R14, R15, R16, C12. The equalization for frequencies lower than 1 kHz is determined essentially by the ratio of the impedance formed by C13/ / R17 to R12.

Fig. 23 - Relative frequency response in the circuit of fig. 20 (playback)

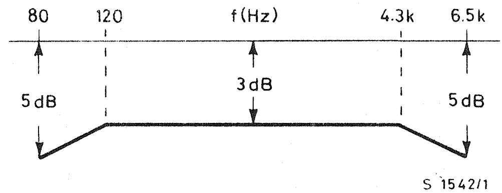


For frequencies greater than 1 kHz the circuit response depends mainly on the transfer impedance of the T circuit (formed by C12 + R15, C11, R16) relative to R12.

The function of the resistance R14 is to prevent high frequency oscillations.

The equalization adopted allows the requirements of DIN standard 45511 to be met with respect to frequency response of magnetic tape recorders (speed 4.75 cm/s). In particular frequency response during playback must fall within the tolerances specified in fig. 24.

Fig. 24 - Limit specifications of playback frequency response



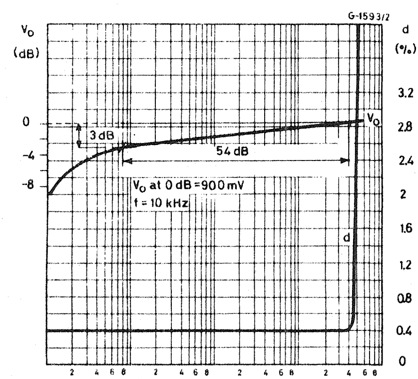
Finally, the use of the T network allows a rapid fall in gain to be obtained at frequencies greater than 10 kHz, minimizing noise contributions at these frequencies.

Performance during recording

As has already been said the TDA 1054M features the possibility of automatic control of the recording level, which allows excellent recordings to be obtained while preventing both tape saturation and degradation of the signal to noise ratio.

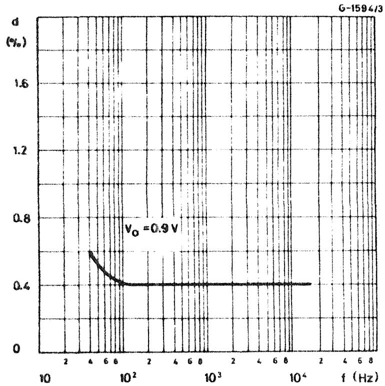
As can be seen in fig. 25 the output signal varies by 3 dB for an input signal variation of 54 dB.

Fig. 25 - Output voltage variation and distortion with ALC vs. input voltage in the circuit of fig. 20 (recording)



Furthermore over the whole control range the harmonic distortion remains very low (d = 0.4%). Fig. 26 shows distortion as a function of frequency when recording with automatic level control.

Fig. 26 - Distortion vs. frequency with ALC in the circuit of fig. 20 (recording)

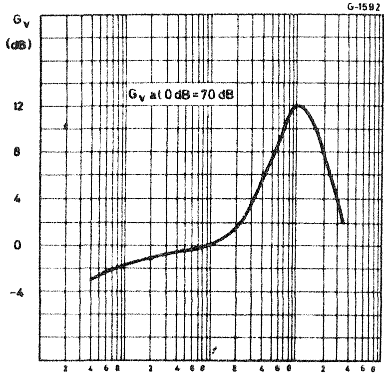


During recording, as in playback, equalization is necessary to compensate for the losses at high frequencies. Fig. 27 shows the relative gain as a function of frequency. This curve is obtained by means of the network formed by R12, R19, R18, R15, C12, C14 (fig. 20). For frequencies lower than 1 kHz the gain of the operational amplifier is determined essentially by the ratio:

$$G_v = \frac{R15 + R18 + R19}{R12}$$

The equalization peak is determined by the network R19, C14, R18, C12.

Fig. 27 - Relative frequency response in the circuit of fig. 20 (recording)



Resistance R20 has the function of rapidly charging capacitor C7, reducing switch-on time as much as possible. It should be mentioned that thanks to low output impedance (typically 12 Ω) the residue of the ultrasonic erase signal is greatly attenuated (pin 13). This is extremely important in order to avoid the ALC controlling the erase signal instead of the signal to be recorded, and also to avoid intermodulation distortion on the signal to be recorded. Fig. 28 shows the curves of limiting time and time for return to nominal level as a function of input level. These times depend on the values chosen for C5, R21, C6, R9. Finally in fig. 29 an example of a printed circuit and component layout for the circuit in fig. 20 is given.

Low cost cassette recorder

Fig. 30 shows the circuit of a low cost cassette recorder which nevertheless gives good performance.

Input circuit

Transistors Q1 and Q2 are directly coupled in cascade as an amplifier having constant gain over the whole frequency range. The input impedance is about 27 kΩ.
As can be seen, bias for the operational amplifier input is obtained by using the quiescent emitter-collector voltage of Q2 (pin 7). In both recording and playback the operational amplifier operates with frequency-dependent gain compensating for losses at low and high frequencies (see previous section). The SVRF and the ALC sections have the same functions as before.
Fig. 31 shows an example of a printed circuit for this circuit.
The general characteristics of the circuit are shown in table 1.

Fig. 28 - Limiting and level setting time vs. input signal variation

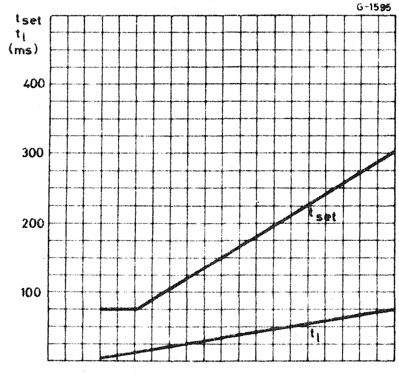


Fig. 29 - P.C. board and component layout for the circuit of fig. 20 (1:1 scale)

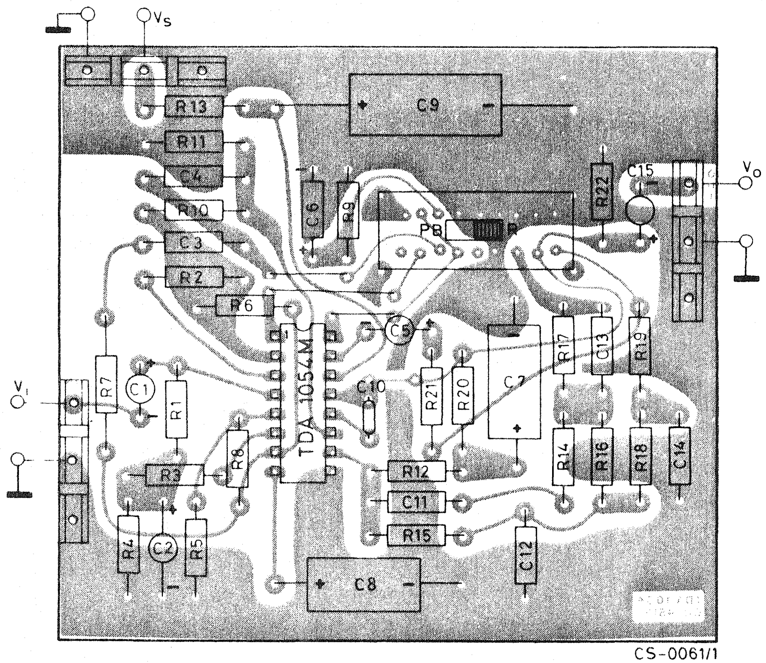


Fig. 30 - Application circuit for low-cost cassette player and recorder

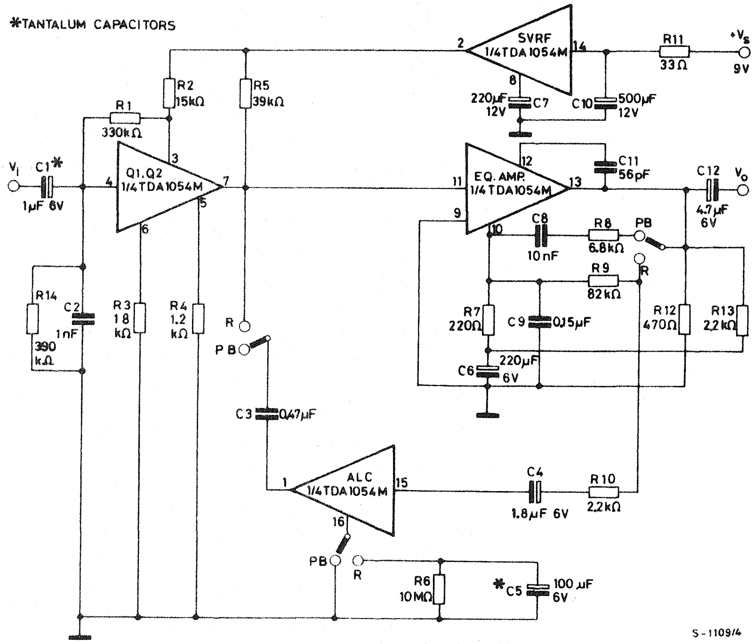


Fig. 31 – P.C. board and component layout for the circuit of fig. 30 (1:1 scale)

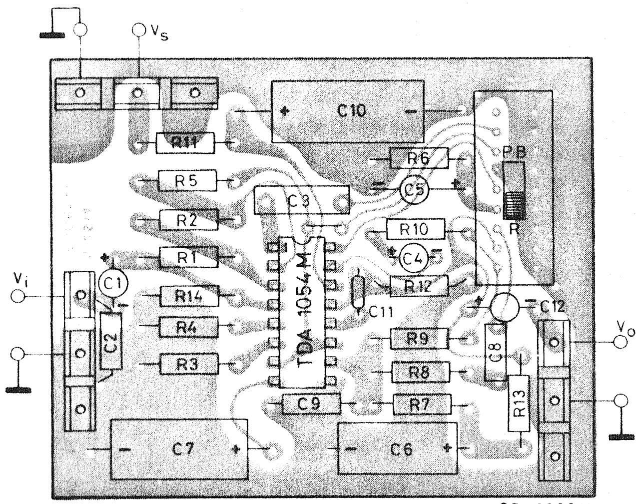


Table 1 — MAIN PERFORMANCE DETAILS FOR CIRCUIT IN FIG. 30 (V_s = 9V)

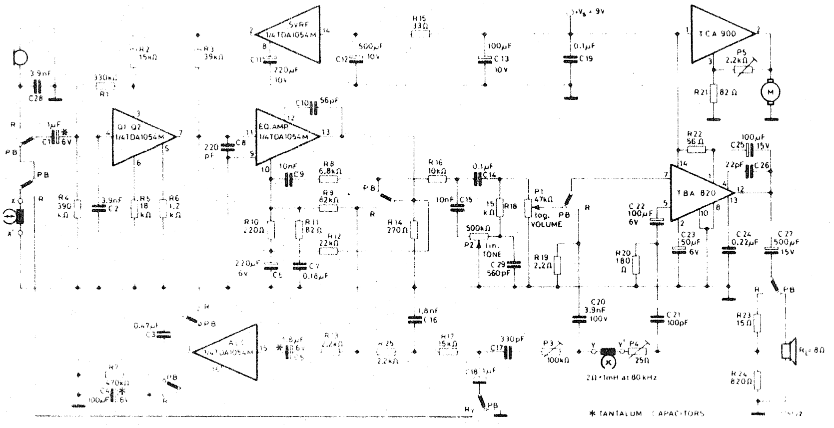
PLAYBACK				
Supply voltage range	=	5 to 12V		
Quiescent drain current	=	18 mA		
Voltage gain (closed loop)	=	54 dB	@	f = 1 kHz
Relative frequency response	=	12 dB	@	f = 100 Hz
		0 dB	@	f = 1 kHz
		5 dB	@	f = 6 kHz
		15 dB	@	f = 10 kHz
		10 dB	@	f = 60 kHz
Distortion	=	0.6%	@	V_o = 1V, f = 1 kHz
Weighted background noise at output	=	1.3 mV	@	Z_g = 300Ω + 120 mH (DIN 45405)
RECORDING				
Voltage gain (closed loop)	=	70 dB	@	f = 1 kHz
Relative frequency response	=	-3 dB	@	f = 140 Hz
		0 dB	@	f = 1 kHz
		4 dB	@	f = 10 kHz
Distortion	=	0.7%	@	V_o = 0.9V, f = 10 kHz
ALC range for 3 dB of output voltage variation	=	54 dB	@	$V_i \leqslant$ 40 mV, f = 10 kHz

Complete portable recorder circuit

Fig. 32 shows a complete portable cassette recorder circuit involving 3 integrated circuits:

- TDA 1054M; record/playback preamplifier
- TBA 820 ; power amplifier in playback and erase oscillator in recording
- TCA 900 ; motor speed controller.

Fig. 32 – Complete cassette player and recorder



Operation during playback

The signal from the head is amplified and equalized by the TDA 1054M (electrical performance of this part of the circuit is similar to the circuit in fig. 30). The signal is sent via the tone and volume controls to the TBA 820 which provides power amplification.

With an input signal of 0.3 mV and with the volume potentiometer set at 2/3 of maximum value an output power of 1.2 W (d = 10%) can be obtained.

Operation during recording

The signal from the microphone is amplified and equalized by the TDA 1054M which also provides automatic control of the recording level (see characteristics of circuit in fig. 30). The signal present at pin 13 is sent (by means of a filter network formed by the 2.2 kΩ and 15 kΩ resistors and the 1.8 nF capacitor) to the recording head. This cell also supplies the recording head with the correct audio current (with the head used the audio current is 30 μA).

The ultrasonic premagnetization signal is also sent to the recording head. As is known AC premagnetization allows better signal to noise ratio, lower distortion, greater residual induction on the tape and finally, a wider frequency response than would be obtained with DC premagnetization.

The premagnetization current can be controlled by the 100 kΩ trimmer (fig. 32) (with the recording head used, $I_{bias} = 250\ \mu A$).

Ultrasonic oscillator

Apart from supplying the premagnetization current to the recording head, the ultrasonic oscillator also supplies the erase head. This circuit is realized by maintaining the TBA 820 in spontaneous oscillation by means of a network (which includes the erase head) providing the TBA 820 with the correct positive feedback.

The output of the TBA 820 (pin 12) is a square wave of very stable amplitude (since it depends only on supply voltage variation). The series resonant circuit formed by the erase head and the 3.9 nF capacitor "extracts" the fundamental from the square wave thus producing an erase (and premagnetization) current with very low harmonic distortion (d = 0.3%). The maximum erase current which can be obtained is about 60 mA rms.

In any case the current in the erase head can be controlled by the 25 Ω trimmer.

Finally it should be noted that the oscillation frequency is in fact the resonant frequency of the circuit formed by the erase head inductance and the series capacitor. (In the case of fig. 32 C = 3.9 nF).

Motor speed controller

The speed of the motor is controlled by the TCA 900 monolithic integrated circuit. More information can be found in technical note No. 113.

Stereo Recorder

For good performance, the fundamental requirements of stereo recorders with ALC are:

- close matching of frequency response and closed loop gain of each channel
- preservation of output dynamic ratios between channels in correspondence with input dynamic ratios between channels.

The TDA 1054M respects these conditions, matching of devices being better than 3 dB (3 dB maximum deviation).

Figs. 33 and 34 show two examples of stereo recorder/player, the first having optimized performance and the second optimised economy.

Fig. 33 - Stereo application circuit for battery-mains tape-cassette player and recorder

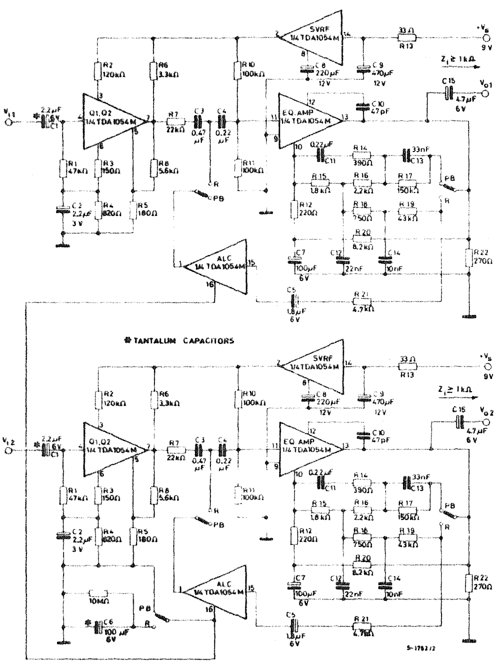
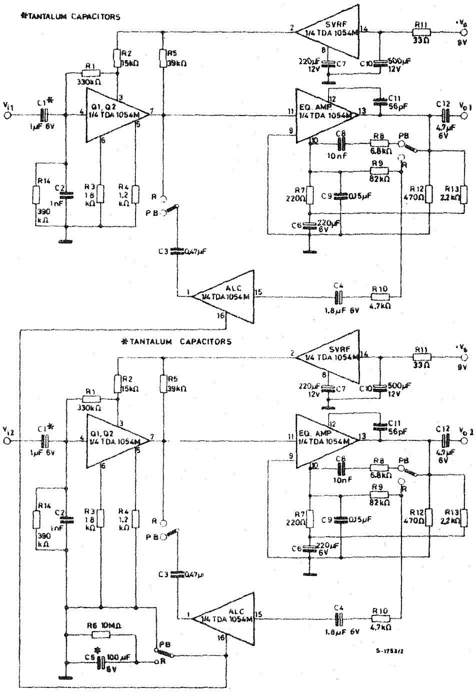


Fig. 34 - Low cost stereo application circuit



Hi-Fi preamplifier with feedback tone control

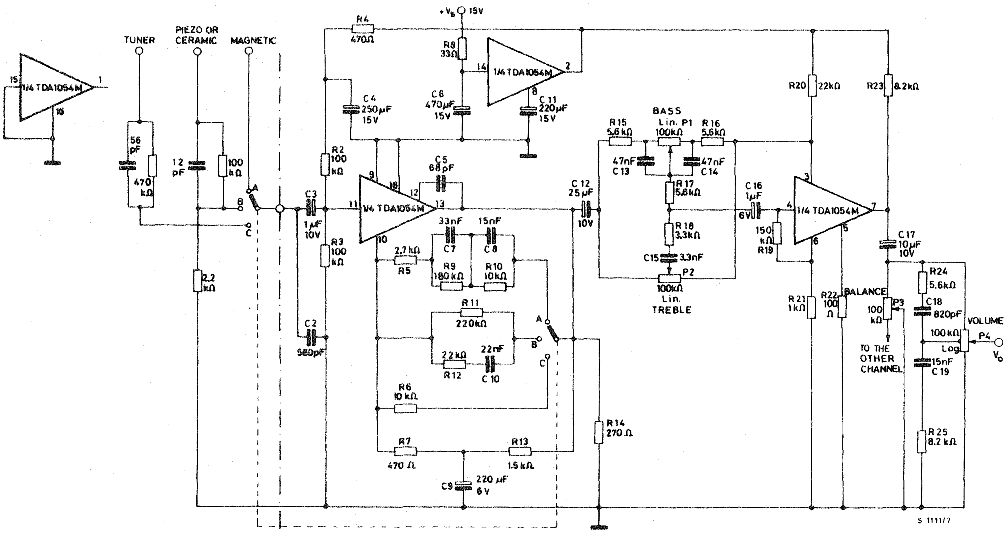
Fig. 35 shows the layout of a hi-fi preamplifier with feedback tone controls (Baxandall type). Inputs are provided for magnetic pick-up (RIAA equalization), tuner (linear response), ceramic and piezoelectric pick-ups.

For the last two a special type of equalization must be used. In fact if the equivalent capacitance of the ceramic or piezoelectric head is between 800 pF and 2000 pF, the equalization adopted allows a 3 dB frequency response from 40 Hz to 20 kHz. The principal characteristics of the circuit are shown in table 2.

Table 2 — MAIN PERFORMANCE DETAILS FOR CIRCUIT IN FIG. 35

Supply voltage range	=	10 to 18V		
Input sensitivity for:				
magnetic pick-ups	=	2.5 mV	@	V_o = 300 mV, f = 1 kHz
ceramic pick-ups	=	100 mV		
Output voltage before clipping	=	2.5 V	@	f = 1 kHz
RIAA equalization for magnetic pick-ups	=	± 1 dB	@	B = 40 to 18,000 Hz
Signal to noise ratio for magnetic pick-ups	=	66 dB	@	$\lvert Z_g \rvert$ = 4.7 kΩ
				B (-3 dB) = 20 to 20,000 Hz
Input impedance for:				
magnetic pick-ups	=	47 kΩ	@	f = 1 kHz
ceramic pick-ups	=	470 kΩ		

Fig. 35 - Hi-Fi preamplifier for magnetic and ceramic pick-ups



Figs. 36 and 37 show respectively distortion as a function of the frequency and frequency response curves obtainable with different tone control positions.

Fig. 36 - Distortion vs. frequency in the circuit of fig. 35

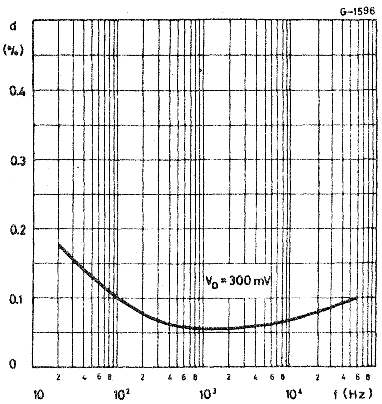
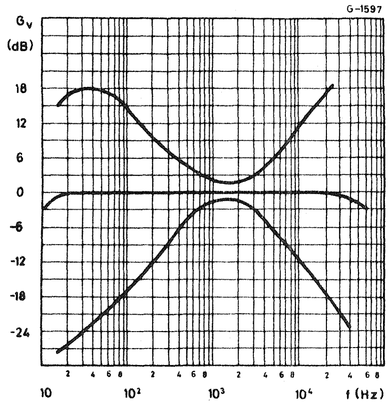


Fig. 37 - Frequency response in the circuit of fig. 35



Hi-Fi preamplifier with passive tone control

Fig. 38 shows the circuit of a hi-fi preamplifier with passive tone controls, allowing the use of logarithmic potentiometers. This circuit is better than the previous one due to in particular lower output impedance and overload protection of the second stage (the volume potentiometer is placed between the two stages).

As in the previous circuit, inputs are provided for:

- magnetic pick-ups ($Z_i = 47\ k\Omega$; RIAA equalization)
- auxiliary input ($Z_i = 470\ k\Omega$; linear response)
- ceramic or piezoelectric pick-ups ($Z_i = 47\ k\Omega$; RIAA equalization).

In this case too, by using ceramic or piezoelectric pick-ups with an equivalent capacitance of between 800 and 2000 pF it is possible to obtain a flat frequency response from 30 Hz to 40 kHz (-3 dB). Table 3 shows the main characteristics of the circuit.

Figs. 39 and 40 show distortion as a function of frequency and output voltage respectively.

Fig. 38 - Hi-Fi preamplifier with passive tone control

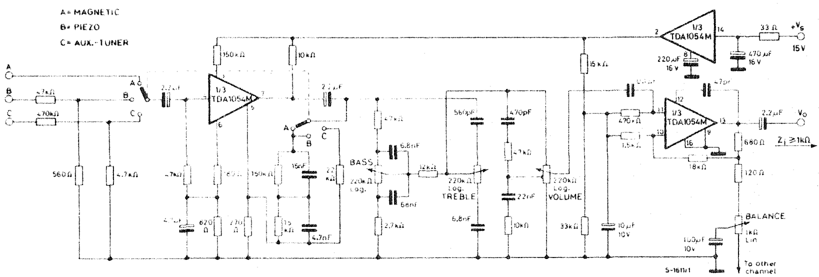


Table 3 — MAIN PERFORMANCE DETAILS FOR CIRCUIT IN FIG. 38

Input sensitivity for:			
magnetic pick-ups	= 4 mV	@	V_o = 500 mV, f = 1 kHz
piezo pick-ups	= 100 mV	@	V_o = 500 mV, f = 1 kHz
auxiliary	= 100 mV	@	V_o = 500 mV, f = 1 kHz
Maximum signal at magnetic pick-ups input	= 35 mV	@	V_o = 3.4V, f = 1 kHz
Signal to noise ratio	= 68 dB	@	V_o = 500 mV, $\|Z_g\|$ = 3.9 kΩ (flat response)
Supply voltage rejection ratio	= 54 dB	@	$f_{(ripple)}$ = 100 Hz (flat response)
Frequency response (-3 dB)	= 30 to 40,000 Hz (tone controls in mid. positions)		

Fig. 39 - Distortion vs. frequency in the circuit of fig. 38

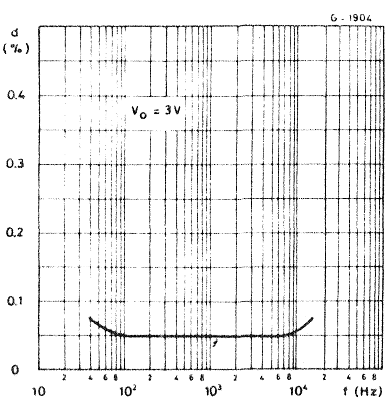


Fig. 40 - Distortion vs. output voltage in the circuit of fig. 38

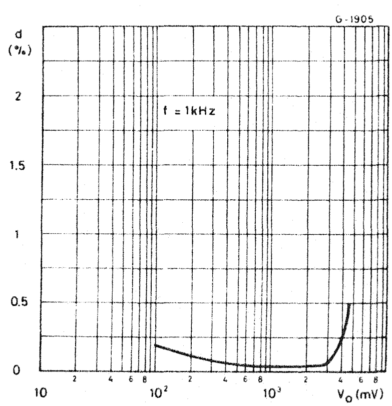


Fig. 41 shows the frequency response for different tone control positions and fig. 42 gives the RIAA equalization curve.

Fig. 41 - Frequency response in the circuit of fig. 38 for various tone control positions

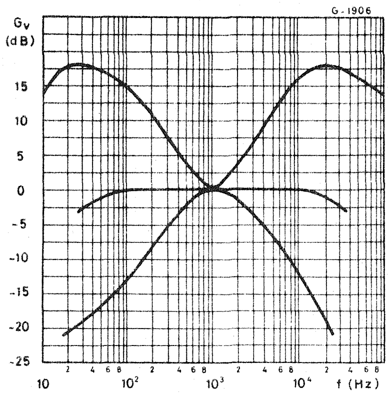
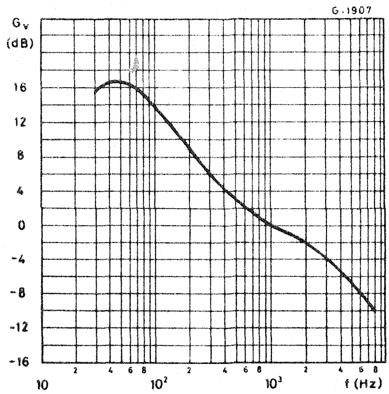


Fig. 42 - Frequency response of the preamplifier equalizer (RIAA) in the circuit of fig. 38



Wired transmission receiver

In fig. 43 the TDA 1054M is used in a receiving circuit for wired transmissions (excluding the input filters). In this case the operational amplifier is used as an RF amplifier. The ALC system maintains the carrier level at pin 11 constant (and therefore at pin 13). This is extremely useful since the input signal can vary from 10 mV to 150 mV depending on the loads on the telephone line. This circuit solution therefore makes it possible to have a detected signal at the output, whose amplitude does not depend on amplitude variations of the carrier signal. This is detected by the diode-connected transistor Q1, (see equivalent circuit in fig. 1). The low frequency signal is then drawn from a common collector stage formed by Q2.

Table 4 gives the principal characteristics of the circuit.

Fig. 43 - Wired transmission receiver circuit (excluding the input filters)

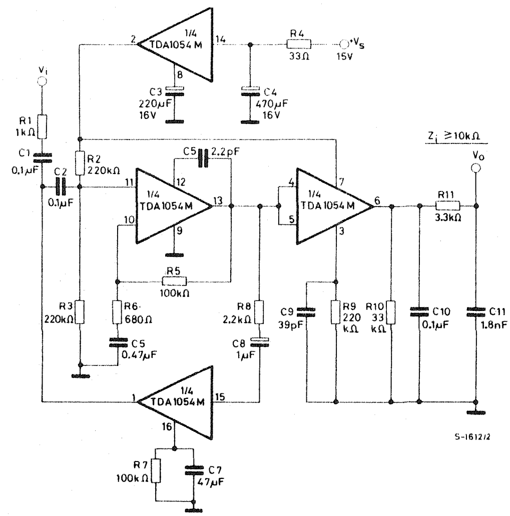


Table 4 — MAIN PERFORMANCE DETAILS FOR CIRCUIT IN FIG. 43

Input sensitivity	=	10 mV	@	V_o = 0.3V, m = 0.5
Distortion	=	0.6%	@	178 kHz ≤ f ≤ 343 kHz
Frequency response (-3 dB)	=	3 to 15,000 Hz		
Signal to noise ratio	=	56 dB	@	V_o = 0.3V, $\|Z_g\|$ = 1 kΩ
ALC range for 3 dB of output voltage variation	=	40 dB	@	V_o = 1V 178 kHz ≤ f ≤ 343 kHz

Stereo preamplifier for ceramic or piezoelectric heads

An interesting application of the TDA 1054M is shown in fig. 44. A stereo preamplifier for ceramic or piezoelectric pick-ups is realized with a single TDA 1054M. By careful choice of component values the same electrical characteristics have been obtained for each channel. In particular it has been arranged so that the open-loop gains are the same (52 dB) and thus any differences between the closed-loop gains and the frequency responses depend entirely on the precision of the external components.

Fig. 45 shows the frequency response curves for different tone control positions, and fig. 46 shows distortion as a function of the frequency.

Fig. 44 - Stereo preamplifier for ceramic or piezoelectric pick-ups

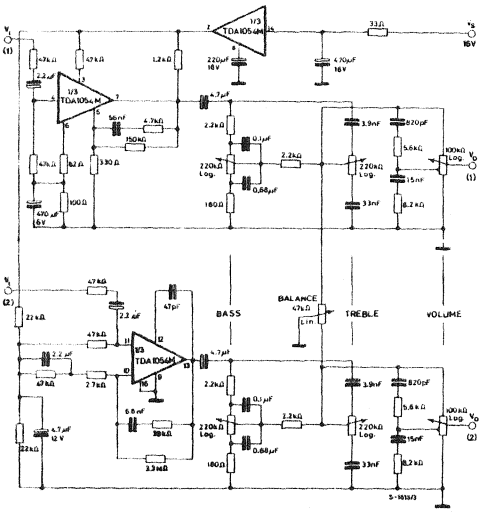


Fig. 45 - Frequency response in the circuit of fig. 44 for various tone control positions

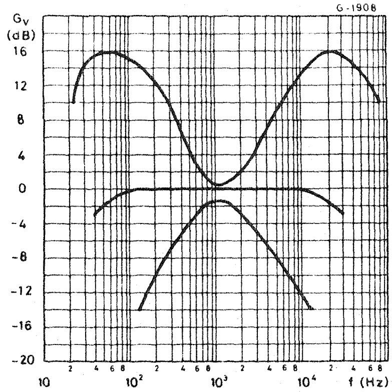
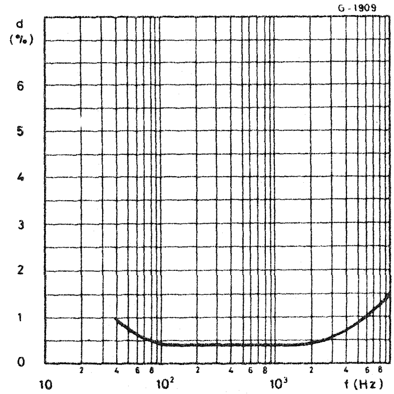


Fig. 46 - Distortion vs. frequency in the circuit of fig. 44



Circuit characteristics are shown in the table 5.

Table 5 — MAIN PERFORMANCE DETAILS FOR CIRCUIT IN FIG. 44

Input sensitivity	=	200 mV	@	V_o = 100 mV, f = 1 kHz
Distortion	=	0.3%	@	V_o = 100 mV 40 Hz ⩽ f ⩽ 15 kHz
Frequency response (-3 dB)	=	40 to 120,000 Hz	@	$C_{(pick-up)}$ = 800 pF
Signal to noise ratio	=	60 dB	@	V_o = 100 mV, $\|Z_g\|$ = 47 kΩ
Supply voltage rejection ratio	=	60 dB	@	$f_{(ripple)}$ = 100 Hz
Tone control range	=	± 14 dB	@	f = 100 Hz, f = 10 kHz

Dynamic range compressor

In fig. 47 the circuit of a dynamic range compressor is shown. This circuit can be used in telephone equipment, modulators and generally in all systems where it is necessary to maintain constant a low frequency signal level. The circuit has a flat response while the input signal level at which control begins is fixed at 50 μV. The circuit performance can however be modified by varying the gain of the first stage:

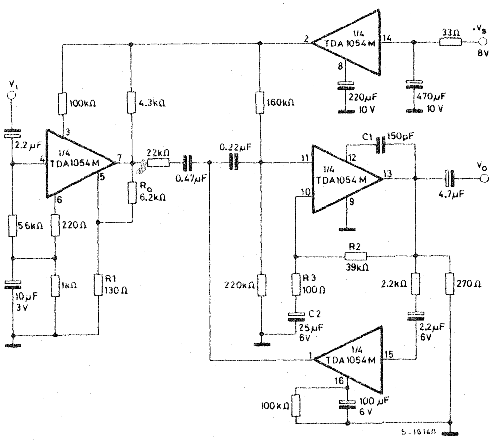
$$G_{v1} = 1 + \frac{Ro}{R1}$$

of the second stage:

$$G_{v2} = 1 + \frac{R2}{R3}$$

and the frequency response (obtaining the value of C1 from fig. 11).

Fig. 47 - Dynamic range compressor circuit



CONCLUSION

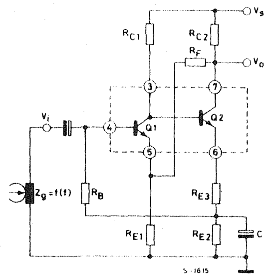
The TDA 1054M not only provides the best solution to the medium and high-quality recorder markets, but thanks to its versatility and performance characteristics can also be used in many other applications, as demonstrated in this technical note.

APPENDIX

Preamplifier noise

We shall discuss here criteria useful to the designer when planning a low noise preamplifier. Fig. 48 shows the circuit of a preamplifier which can be used both in recorders and in hi-fi preamplifiers for magnetic pick-ups.

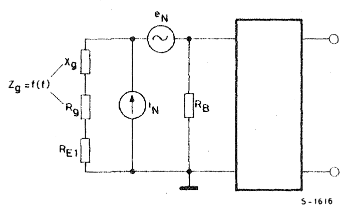
Fig. 48 - Low noise preamplifier circuit for tape recorders and hi-fi preamplifiers with magnetic pick-ups



Signal to noise ratio

The first consideration (provided that a choice of pick-up is possible) is optimizing of the signal to noise ratio at the input of the preamplifier. Fig. 49 shows the preamplifier with the noise generators referred to the input. In this way the preamplifier can be considered as noise-free.

Fig. 49 - Equivalent circuit of noise voltage and noise current generators at preamplifier input (open loop)



For the time being we shall disregard the noise due to the second transistor and for convenience the signal to noise ratio will be calculated considering the open-loop circuit shown in fig. 49. In fact as is known the signal to noise ratio does not depend (for the same band) on the presence of negative feedback.
The noise contributions are given by:

- e_N Equivalent noise voltage of the input transistor (from data sheet)
- i_N Equivalent noise current of the input transistor
- Thermal noise generated by the real part R_g of the source impedance
- Thermal noise generated by the emitter bias resistance R_{E1}.
  It should be noted that resistance R_{E1}, placed between the emitter of the first transistor and earth, generates the same noise as a resistance

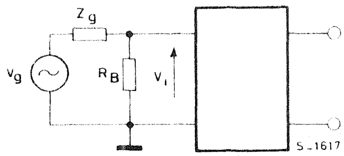
of the same value in series with the input. In fact it has been observed that in feedback systems any interference (including noise) generated at the point where the input signal and the feedback signal are compared, is treated by the amplifier as if it were applied at the input.

– Thermal noise generated by the input resistance R_B.

The total noise voltage V_N present at the input is given by formula 1 * (noise sources not correlated). This noise is obviously a function of the frequency and of the bandwidth Δf. Let us now consider the useful signal present at the input of the preamplifier (fig. 50):

$$V_i = V_g \cdot \frac{R_B}{|R_B + Z_g|}$$

Fig. 50 – Useful signal present at preamplifier input



Hence the signal to noise ratio is given by formula 2**.

On the basis of this result it can be stated that, independent, of the values of e_N and i_N, the highest possible signal to noise ratio can be achieved if the following conditions are met:

– $R_B \gg |Z_g + R_{E1}|$ over the whole range of audio frequencies.
This means choosing a head the modulus of whose impedance at 15 to 20 kHz is much less than the input impedance R_B (usually 47 kΩ).

– R_{E1} should be kept as low as possible. The lower limit for R_{E1} is determined by the need to keep the preamplifier input impedance much greater than R_B (even in open-loop conditions) so that the impedance seen by the source depends almost exclusively on R_B. In practice, with the available current gains ($h_{fe\,min} = 300$) and at the working values of emitter current (80 to 300 μA) R_{E1} should be between 100 Ω and 470 Ω.

– The terms involving $|(Z_g + R_{E1})|^2$ are very small compared with the others. Assuming f = 1 kHz:

$$L \ll \frac{1}{2\pi \cdot 10^3} \cdot$$

$$\cdot \sqrt{\frac{e_N^2 + 4KT\,(R_{E1} + R_g)}{i_N^2 + \dfrac{4KT}{R_B}} - (R_{E1} + R_g)^2} \qquad (3)$$

As can be seen the value of L depends also on the terms e_N^2 and i_N^2 which in turn depend on the frequency and bias current of the transistors (see figs. 5 and 6).
In any case, since manufacturers of heads quote the inductance at a frequency of 1 kHz, the values of e_N and i_N are taken from the figures mentioned above, corresponding to this frequency.
In order to make a first approximation of the value of L which satisfies the previous inequality, average values relative to bias currents of between 50 μA and 300 μA are chosen for e_N and i_N (f = 1 kHz).
A more accurate choice of the bias current is made on the basis of noise figure.

One final comment should be made regarding the effect of the inductance value of the head on the signal to noise ratio. The minimum value of the inductance is linked to the signal voltage obtainable. Bearing in mind the voltage level necessary at the output of the preamplifier and the voltage gain, it would be a mistake to use a head which gives a signal greater than required at the input of the preamplifier. In fact while the usable signal increases in proportion to the square root of inductance, the noise increases in direct proportion to it. This means that by doubling the inductance the signal increases by 3 dB while the noise increases by 6 dB, meaning that the signal to noise ratio deteriorates by 3 dB.

Noise figure

As has been seen the noise present at the input of

* Formula 1: $$V_{N}rms = \frac{R_B}{|R_B + R_{E1} + Z_g|} \cdot \sqrt{\Delta f \cdot \left[e_N^2 \cdot \left|1 + \frac{R_{E1} + Z_g}{R_B}\right|^2 + i_N^2 \cdot |Z_g + R_{E1}|^2 + 4KT\left(R_{E1} + R_g + \frac{|R_{E1} + Z_g|^2}{R_B}\right)\right]}$$

** Formula 2: $$S/N = \frac{V_i}{V_N} \doteq \left|\frac{R_B + R_{E1} + Z_g}{R_B + Z_g}\right| \cdot \frac{V_g}{\sqrt{\Delta f \cdot \left[e_N^2 \cdot \left|1 + \frac{R_{E1} + Z_g}{R_B}\right|^2 + i_N^2 \cdot |Z_g + R_{E1}|^2 + 4KT\left(R_{E1} + R_g + \frac{|R_{E1} + Z_g|^2}{R_B}\right)\right]}}$$

the preamplifier (open-loop) is given by:

$$V_N^2 = e_N^2 + i_N^2 \cdot |Z_g + R_{E1}|^2 + 4KT\,(R_{E1} + R_g) + \frac{4\,KT}{R_B} \cdot |R_{E1} + Z_g|^2$$

This is spot noise i.e. it is calculated for a bandwidth of 1 Hz. The (spot) noise figure is therefore given by:

$$NF = 10\,\lg\left[1 + \frac{i_N^2 \cdot |Z_g + R_{E1}|^2 + e_N^2}{4\,KT\,(R_{E1} + R_g)} + \frac{|R_{E1} + Z_g|^2}{R_B\,(R_{E1} + R_g)}\right]$$

This figure too is a function of the frequency and of the bias current of the transistors. We shall now study how this bias current should be chosen. A technique widely adopted is to evaluate the noise figure by considering the input impedance as purely resistive and of value equal to the geometric mean of the impedance values at the extremes of the audio band. The inequality (3) allows the approximate value of L to be obtained (the resistive component R_g usually lies between 50Ω and 300Ω). The equivalent source resistance can thus be calculated from the following formula:

$$R_{g\,eq} = \sqrt{|Z_g|_{f\,=\,80\,Hz} \cdot |Z_g|_{f\,=\,15\,kHz}}$$

Knowing this value (fig. 8) the bias current, which for a given $R_{g\,eq}$ gives the minimum noise figure, can be chosen. It is now possible to determine the exact values of e_N and i_N from figs. 5 and 6 and hence calculate both the signal to noise ratio (equation 2) and the exact inductance of the head. Therefore in order to achieve the best noise relationship for a preamplifier the procedure to be followed may be summarized thus:

- the input impedance must be made much greater than the impedance of the head at the upper limit of the audio band
- the resistance of emitter resistor R_{E1} must be kept as low as possible while compatible with the input impedance
- the inductance value (f = 1 kHz) must satisfy the inequality (3)
- the value of $R_{g\,eq}$ is then obtained
- on the basis of $R_{g\,eq}$ the bias current which produces the lowest noise figure is calculated
- the inductance value of the head must be as low as possible while compatible with the signal level required at the input of the preamplifier.

Let us now consider the noise at the base of the second transistor. This noise can be referred to the input of the preamplifier if divided by the term:

$$G_{v1} \cdot \sqrt{B \cdot \frac{\pi}{2}}$$

where G_{v1} is the gain of the first transistor and B is the -3 dB bandwidth of the first transistor. The total voltage noise at the input is therefore the mean square value given by this noise plus the generator e_N present at the input of the preamplifier. In order to minimize this noise the second transistor must be biased with the lowest possible current compatible with the dynamic range and the distortion.

Finally it must be remembered that we have discussed noise at the input of the preamplifier and that the noise at the output depends greatly on the frequency response of the preamplifier. Different results will be obtained depending on whether the amplifier operates with a flat response (over the audio range), with RIAA equalization or with the equalization normally used in tape recorders. By optimizing the signal to noise ratio at the input, however, the best signal to noise ratio is automatically given at the output.

Table 6 – TYPICAL PERFORM. DETAILS FOR CIRCUIT IN FIG. 20 (T_{amb} = 25°C, V_s = 9V)

Parameter		Test conditions	Min.	Typ.	Max.	Unit
PLAYBACK						
G_v	Voltage gain (open loop)	f = 20 to 20,000 Hz		110		dB
G_v	Voltage gain (closed loop)	f = 1 kHz		57		dB
$\|Z_i\|$	Input impedance	f = 100 Hz f = 1 kHz f = 10 kHz		10 41 43		kΩ kΩ kΩ

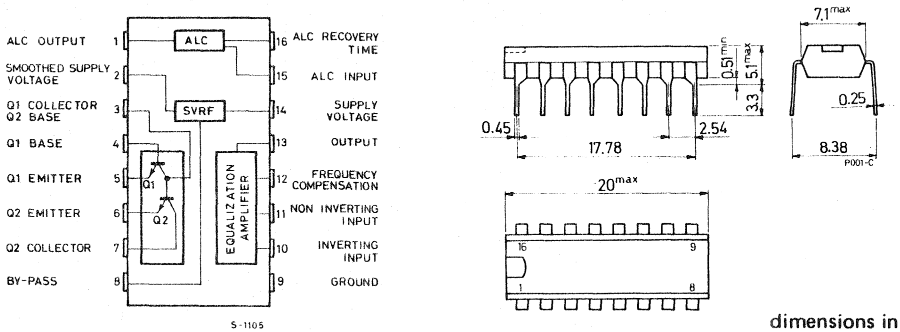
Table 6 – TYPICAL PERFORMANCE OF CIRCUIT IN FIG. 20 (continued)

	Parameter	Test conditions		Min.	Typ.	Max.	Unit
$\lvert Z_o \rvert$	Output impedance	f = 1 kHz			12	35	Ω
B	Frequency response				see fig. 23		
d	Distortion	$V_o = 1V$	f = 1 kHz		0.1		%
	Background noise at the output.	Z_g = 300 Ω + 120 mH (DIN 45405)			1.3		mV
***	Weighted background noise at the output				1.3		mV
$\frac{S+N}{N}$	Signal to noise ratio	$V_o = 1.3V$ Z_g = 300 Ω + 120 mH			60		dB
SVR	Sypply voltage ripple rejection at the output	$f_{(ripple)}$ = 100 Hz			30		dB
t_{on}**	Switch-on time	$V_o = 1V$			500		ms
RECORDING							
G_v	Voltage gain (open loop)	f = 20 to 20,000 Hz			110		dB
G_v	Voltage gain (closed loop)	f = 1 kHz			70		dB
B	Frequency response				see fig. 27		
d*	Distortion without ALC	$V_o = 0.9V$	f = 1 kHz		0.3		%
d	Distortion with ALC	$V_o = 0.9V$	f = 10 kHz		0.4		%
ALC	Automatic level control range (for 3 dB of output voltage variation)	$V_i \leqslant 40$ mV	f = 10 kHz		54		dB
V_o	Output voltage before clipping without ALC	f = 1 kHz			2.3		V
V_o	Output voltage with ALC	V_i = 30 mV	f = 10 kHz		0.9		V
t_l**	Limiting time (see fig. 16)	ΔV_i = +40 dB	f = 1 kHz		75		ms
t_{set}**	Level setting time (see fig. 16)				300		ms
t_{rec}**	Recovery time (see fig. 17)	ΔV_i = -40 dB	f = 1 kHz		150		s
t_{on}**	Switch-on time	$V_o = 0.9V$			500		ms
$\frac{S+N}{N}$****	Signal to noise ratio with ALC	$V_o = 0.9V$	R_g = 470Ω		64		dB

* Measured with selective voltmeter
** This value depends on external network
*** When the DIN 45511 norm for the frequency response is not mandatory the equalization peak at 10 kHz can be avoided-so halving the output noise.
**** Noise measured on weighted (DIN 45405)

Mechanical and electrical characteristics of the TDA 1054M

Fig. 51 – Connection diagram and mechanical data



dimensions in mm

Fig. 52 – Test circuit

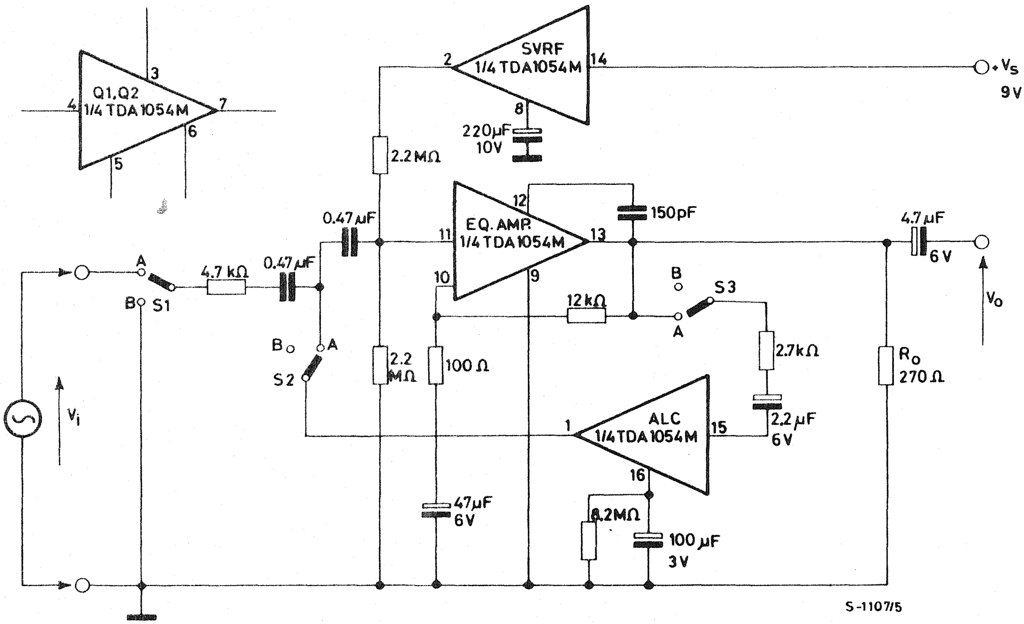


Table 7 – THERMAL DATA				
$R_{th\ j\text{-}amb}$	Thermal resistance junction–ambient	max	200	°C/W

Table 8 – ELECTRICAL CHARACTERISTICS (Refer to the test circuit, $T_{amb} = 25°C$)

	Parameter	Test conditions	Min.	Typ.	Max.	Unit
V_s	Supply voltage		4		20	V
I_d	Quiescent drain current	$V_s = 9V$ $R_o = \infty$ $S1 = S2 = S3 = B$		6		mA
h_{FE}	DC current gain (Q1 and Q2)	$I_C = 0.1$ mA $V_{CE} = 5V$	300	500		–
e_N	Input noise voltage (Q1)	$I_C = 0.1$ mA $V_{CE} = 5V$ $f = 1$ kHz		2		$\frac{nV}{\sqrt{Hz}}$
i_N	Input noise current (Q1)	$I_C = 0.1$ mA $V_{CE} = 5V$ $f = 1$ kHz		0.5		$\frac{pA}{\sqrt{Hz}}$
NF	Noise figure (Q1)	$I_C = 0.1$ mA $V_{CE} = 5V$ $R_g = 4.7$ kΩ B (–3 dB) = 20 to 10,000 Hz		0.5	4	dB
G_v	Open loop voltage gain (equalization amplifier)	$V_s = 9V$ $f = 1$ kHz		60		dB
V_o	Output voltage with ALC	$V_s = 9V$ $V_i = 100$ mV $f = 1$ kHz S1=S2=S3=A		0.9		V
R1	(for SVRF system)			7.5		kΩ
R2	(for SVRF system)			120		Ω
e_N	Equivalent input noise voltage (for equalization amplifier pin 11)	$V_s = 9V$ $R_g = 4.7$ kΩ $G_{v(closed)} = 100$ S1 = B B (–3 dB) = 20 to 20,000 Hz		1.3		μV
	Drop-out (between pins 14 and 2)	$I_d = 6$ mA $V_s = 9V$		0.8		V

SGS – ATES GROUP OF COMPANIES

Italy – France – Germany – Singapore – Sweden – United Kingdom – U.S.A.

PRINTED IN ITALY

124116/1