Wireless World Circard

Series 5: Audio circuits

The vast majority of these low-level audio circuits use the 741 operational amplifiers, which need dual-polarity power supplies. The tape head preamplifier is one exception, where the low-noise National or RCA op-amps are recommended. One card, entitled economy i.c. audio circuits, gives a selection of eight circuits using the LM3900 in a way that allows use of a single-ended power supply. Five cards deal exclusively with preamplifiers of various kinds, one a multi-input type catering for all the usual signal sources. Card 1, detailing an equalizer for magnetic pickups, shows the technique of reducing cable capacitance by using double-screened cable.

Most of the remaining cards deal with higher-level circuits, for mixing, filtering and tone controls. Cards 3 and 7 show simple 12dB/octave scratch and rumble filters, with useful modifications that can give a variable attenuation slope and cut-off frequency with a single potentiometer, the rate varying between 6 and about 18dB/octave.

The two tone control cards also give useful circuits, card 6 particularly with its simple multi-section circuit (the "graphic" or "room" equalizer) using only RC (Wien) networks, rather than the more common RLC circuits. The circuits of card 2 give two variants of the now-famous Baxandall tone control circuit, one using an op-amp and one using a field effect/bipolar transistor combination to give better linearity.

Audio Circuits

Cinemascope or the Magic Lantern? The breadth of choice available to the user of equipment reproducing audio signals is just as great. We are here seeking the happy medium and sidestepping such difficulties as to whether the medium should be disc or tape — or for Menuhin fans, the message.

The starting point is the assumption that the signals though complex can be represented as a mixture of sinusoidal waves of different amplitudes with frequencies lying between certain limits, say 20Hz to 20kHz. Generally the aim of good audio equipment is to produce at the ear of the listener a pattern of sound most closely resembling that which he would have heard as a direct listener to the original sound source. The system has to take account of the characteristics of the transducers at both ends of the chain as well as any intervening media used for storing or transmitting this signal.

If the input transducer had a linear amplitude response and gave the same output voltage for a given sound intensity regardless of frequency, then the following amplifiers could themselves have a linear response. The design of such amplifiers, with the aid of modern technology in the form of operational amplifiers is by now routine. There are three distinct departures from this idealized existence.

- The output voltage for constant signal strength may be frequency dependent in some controlled manner e.g.: tape-head e.m.f. proportional to frequency for constant amplitude recorded signal.
- The signal may have been recorded and/or processed by some preceding stage with some characteristic defined according to some standard (R.I.A.A., B.S., C.C.I.R. etc).
- Imperfections in some other part of the system may have resulted in anomalies in the desired response e.g.: resonance effects in transducers.

Any one of these would call for correcting action in the amplifying chain, though in some cases as in the design of loudspeakers, resonance effects in the speaker itself can be dealt with by careful design of the enclosure. As each transducer is a very complicated mechanism involving the interaction of several electrical and mechanical properties it is common to operate them with amplifiers whose impedance characteristics are closely controlled, thus eliminating one possible source of variation in performance. This article considers only the input transducers, such as microphones, tape-heads, and assumes that any succeeding power amplifier/loud-speaker combination can have its imperfections accounted for by tone controls.

The matching problem at the input reduces the design of amplifiers whose input impedance is either equal to, much less than, or much greater than that of the source. Equal source and input impedances are used in line amplifiers where, for example, input, output and attenuator resistances might be 600Ω. This allows for easy calculation of power levels at all points in a system, and for the interconnection of multiple elements in a system. On the other hand, even within such a system the power output amplifier might be designed to have an output resistance $<<$ 600 ohms so that several such loads might be paralleled without diminishing the power fed to each.

A second important feature of the matched impedance condition is that it maximizes power transfer from a source of given e.m.f., and internal resistance. In most modern circuits using heavy negative feedback, the natural impedances tend to be either very high or very low and there is then no advantage from a power transfer standpoint of artificially modifying their terminal impedances to some arbitrary value. To do so simply throws away power in the passive network added for this purpose.

In the case of very small signals where noise is a severe problem, matching of impedances plays an important part. A moving-coil microphone having a low internal impedance (e.g. 200Ω) generates a low e.m.f. of, say, 100μV r.m.s. Fed directly to a semiconductor amplifier, the input noise voltage would be relatively large, while it would be possible to have a high input impedance using series negative feedback. A step-up transformer of large turns-ratio would greatly increase the signal e.m.f. at the amplifier input and would dominate the noise voltage. However the effective source impedance seen by the amplifier would also be raised by the transformer action and with it the contribution to noise due to the amplifier's input noise current. The optimum condition is when the contributions due to noise voltage and current generation are comparable. Other parameters such as amplitude response are also affected but the condition chosen is often close to the matched condition.

For microphones, the mechanical properties are normally designed so that they are self-equalized, i.e. that they give an output e.m.f. that depends only on the sound intensity and not on frequency. The most common microphones are magnetic in some form, variants including moving-coil, moving-iron and ribbon microphones. Reduction of the moving mass to extend response tends to reduce both sensitivity and impedance with the problems described above. Crystal microphones are used for low-cost applications such as simple cassette recorders and require a high input impedance pre-amplifier to avoid attenuation at low frequencies where the capacitive reactance of the microphone increases. The pre-amplifier design is similar to that for crystal/ceramic pickups, i.e. a flat response and generally an impedance in excess of 1MΩ, possibly up to 10MΩ or more for low capacitance units with extended low-frequency amplitude response. Alternatively, the feedback may contain a capacitance whose change in reactance compensates for that of the transducer.

These ceramic elements could in principle be designed to give a frequency-independent output when used for record reproduction, but another factor enters the argument. During recording, signals are first passed through frequency-dependent amplifiers. These have strictly controlled characteristics, usually referred to as R.I.A.A. and further defined in BS1928. If all signals were recorded with a so-called constant-velocity characteristic it would be found in practice that the amplitude at low frequencies would result in breakthrough between neighbouring sections of the groove. This is because constant velocity fixes the velocity at the zero-crossing point of the signal.

At low frequencies the longer period would allow proportionately larger excursions. Hence low-frequency signals are recorded with amplitude proportional to signal e.m.f. (whereas a velocity-proportional recording would otherwise have merits since a magnetic playback element would re-convert that velocity back into e.m.f. proportionately). This constant amplitude characteristic merges into a constant-velocity region at around 1kHz, but at still higher frequencies the recording again changes to constant amplitude. The reason is different. The majority of the noise in any system is concentrated in the higher octaves as in most cases noise is proportional to bandwidth. By emphasizing high frequency signals during the recording process and reversing the procedure on playback, the overall amplitude response remains linear, but any noise due to the record surface and playback pre-amplifier is diminished as it is relative to a much larger signal. Noise accompanying the original signal emerges from the system at an unchanged ratio.

This recording characteristic of BS1928 accommodates the larger low-frequency amplitudes common in music, and does not lead to distortion at high frequencies as the signal amplitudes are relatively small. The playback transfer function is

$$T_0 = k \frac{(1+j\,\omega\,T_2)}{(1+j\,\omega\,T_1)\,(1+j\,\omega\,T_3)}$$

Where $T_1 = 75$, $T_2 = 318$ and $T_3 = 3180\mu s$. To achieve this with a magnetic cartridge, the preamplifier input resistance should be higher than that of the cartridge at all frequencies of interest, or should have a fixed value that can be allowed for in tailoring the cartridge response in terms of its electro-mechanical properties. A typical value is $50k\Omega$. The voltage gain must fall between 50Hz and 500Hz at 6dB/octave, passing through a point of inflection at 1kHz, and falling again at 6dB/octave beyond 2.2kHz. These three time-constants may be defined by three separate *CR* circuits; in some cases two of the time constants are achieved using a single capacitor in a suitable network of resistors.

With ceramic cartridges, equalization is not a result of circuit design but of the transducer itself. The various parameters such as compliance as well as resonances are carefully combined to provide a good approximation to the desired equalization subject to correct loading as outlined above. Where an amplifier has in-built equalization (i.e. for magnetic cartridge) then a separate network may be inserted between the ceramic cartridge and that input to remove the effect of that equalization — a cumbersome process that might be called re-de-equalization.

Tape-recorded signals followed a C.C.I.R. characteristic recently re-defined and extended in BS1568 part 1. There is a low-frequency time constant identical to that in the R.I.A.A. curves, i.e. a time constant of $3180\mu s$, with one further time-constant depending on tape speed, but giving a response that is constant above a particular frequency. These characteristics are quite independent of any imperfections in particular combinations of heads and tapes though intended to optimize their operating conditions. Feedback networks are then similar to those for magnetic cartridge pre-amps equalized as above though requiring only two *CR* time constants in the ideal case. In practice the imperfections of the system may force for example the addition of some treble boost on playback, operating in the 5 to 20kHz region. This is not covered by any standard, but may readily be incorporated by a further decrease in feedback factor in these regions.

Once these preamplifiers have converted the transducer outputs into voltages bearing a nominally linear relationship to the original sound intensity, it might be thought possible simply to amplify the signals further and apply them directly to an output transducer. Such trusting simplicity exposes one to ridicule for ignorance of that recent discovery. Finagle's axiom on reproducing circuitry and equipment — FARCE for short — viz "All signals are equalized but some are more equalized than others". Most audio systems use one or more circuits to modify the amplitude response of the signals passing through them to correct for this effect.

Where unwanted material occurs at the extremes of the spectrum then low-pass or high-pass filters are used for sharp attenuation of these unwanted signals with minimum attenuation of the desired signals. These filters when used in audio equipment are generally called scratch and rumble filters respectively but the basic principles underlying them are the same (see Circards series 1). Second- or third-order filters are used, and as the ultimate judgement of these audio systems is rightly the subjective one of a listening test, the choice of filter characteristic is often empirical.

During such a listening test, the parameters of the room housing the loudspeaker plays a large part, while sound sources including commercial recordings are not above suspicion in respect of the linearity of amplitude response. Even if all such sources reached the impeccable standards which the engineers concerned strive so successfully to meet, there would remain the personal preferences of the user. It takes a brave man to refrain from just-a-touch on the tone controls when demonstrating the superiority of his latest equipment to a fellow enthusiast (competitor?). Of all the tone controls proposed, the most generally accepted is due to P. J. Baxandall, basing itself on a feedback rather than a passive network. This allows for true boost or cut to either low or high frequencies relative to an unchanging centre frequency, generally 1kHz. Two potentiometers are used, adjusting the feedback in the two frequency regions separately around a virtual earth amplifier such that the gain in these regions varies typically from 0.1 up to 10 i.e. 20dB. The higher the quality of the sources and other links in the chain the smaller the range covered by these tone controls need be.

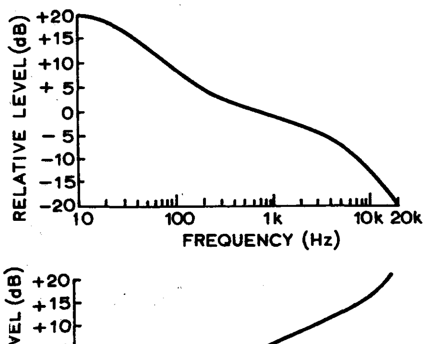
More complex tone controls may be used to sub-divide the frequency spectrum still further; though purists will reject this approach as it smacks of gimmickry, there can be a case for it for various forms of electronic musical instruments and in sound effects. One possibility is the use of parallel channels each consisting of a low-*Q* band-pass filter using sufficient channels that the mixed signal has very little ripple in its overall amplifier response characteristic when all controls are level. It is convenient for producing relatively small amounts of boost and cut at selected regions in the spectrum and may be augmented by active filters with higher *Q* if stronger effects are needed.

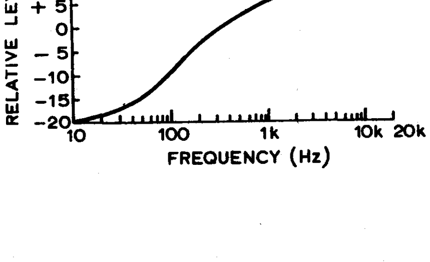
The mixer circuits used in such a system, as when mixing inputs from tape, disc, radio, are now frequently based on the see-saw amplifier feeding to the virtual earth through appropriately scaled resistors. If it is desired to obtain significant voltage gain from the mixer as well as having multiple inputs, the bandwidth restrictions are more severe, being in effect determined by the total gain used, i.e. the sum of the gains with respect to various inputs. Phase shift in the operational amplifier at all frequencies above 10Hz is such as to make the virtual earth point have a largely inductive impedance, i.e. one that rises proportional to frequency.

As a final comment on the possibility of multi-band operation of audio systems it can be argued that limiting the number of loudspeaker drive units to two (a woofer and a tweeter) with the occasional addition of a mid-range squawker is too restrictive. Accordingly we suggest a new classification scheme of dividing the spectrum from 10Hz to 30kHz into seven bands each to be handled by a separate loudspeaker, see table. Combined with quadraphonic operation surely an export boom must be the result?

Table 1: proposed classification for loudspeakers.

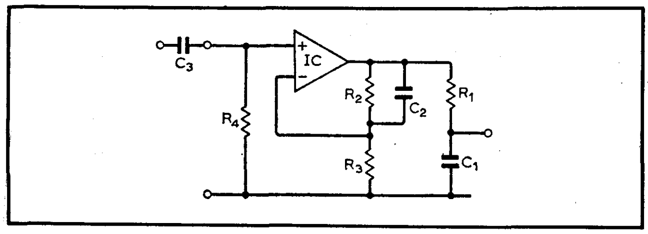
Frequency range (Hz)	Title
10- 30	grunter
30-100	boomer
100-300	roarer
300- 1k	crooner
1k- 3k	howler
3k- 10k	screamer
10k- 30k	screecher





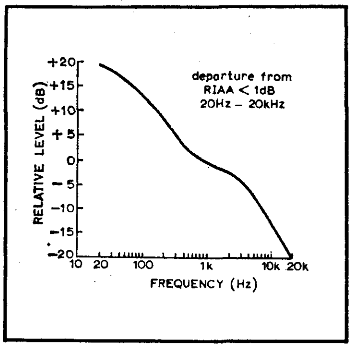
Wireless World Circard

Magnetic cartridges/R.I.A.A. equalization



Typical performance

IC: 741
R_2: 680kΩ
R_3: 76kΩ (82kΩ//1MΩ)
R_1: 15kΩ
R_4: 68kΩ
C_3: 0.33μF.
C_2: 4.7nF
C_1: 4.7nF
Supplies: ±10V
Distortion: 0.05% at 350mV r.m.s. input 100Hz, 1kHz. 0.1% at 10kHz



Circuit description

The required transfer function to meet the playback characteristic of BS 1928: 1965 and RIAA is of the form

$$T = K\frac{(1 + j\omega T_2)}{(1 + j\omega T_1)(1 + j\omega T_3)}$$

with $T_1 = 75$, $T_2 = 318$ and $T_3 = 3180\mu s$. The circuit shown achieves this by using an external passive network R_1, C_1 which gives the attenuation corresponding to $1/(1 + j\omega T_1)$ and a feedback network R_2, R_3, C_2 which controls the gain of the amplifier to give the remaining frequency-dependent terms. In this way three time constants are provided using only two capacitors. At very low frequencies the gain is constant at $(R_2 + R_3)/R_3$ provided the input time constant is low enough. As frequency rises X_{C2} becomes comparable with R_2 and the gain falls, until X_{C2} becomes comparable to R_3. At higher frequencies amplifier gain becomes unity, with the passive network following contributing an increasing attenuation at still higher frequencies. In some cases the limited bandwidth of the amplifier might be relied on to approximate to this final high-frequency performance, eliminating R_1 and C_1. Input impedance is determined by R_4 except at very high frequencies where amplifier input capacitance becomes significant. Output impedance is determined by R_1, C_1. To obtain high-accuracy capacitors at low cost, the time constants may be obtained using resistors for R_2, R_3 that result in a d.c. offset due to input current that conveniently cancels that due to R_4. Lower values of resistor minimize hum pick-up problems in difficult environments.

Component changes

IC: Any general purpose op-amp whose gain-bandwidth product 20kHz.
R_2: 6.8 to 680kΩ
R_3: 680Ω to 100kΩ
R_1: 470Ω to 47kΩ
C_3: 0·1 to 10μF (may be dispensed with if head allowed to carry small input current).
C_2: 1nF to 1μF
C_1: 1nF to 1μF
Constraints: $R_1C_1 = 75\mu s$, $R_2C_2 = 3180\mu s$ and $R_2R_3C_2/(R_2 + R_3) = 318\mu s$.
Within these contraints the input resistance should be around 50kΩ for matching standard magnetic cartridges; the resistor values have to be compromises between drift and hum/noise problems (high R), as well as amplifier loading (low R).

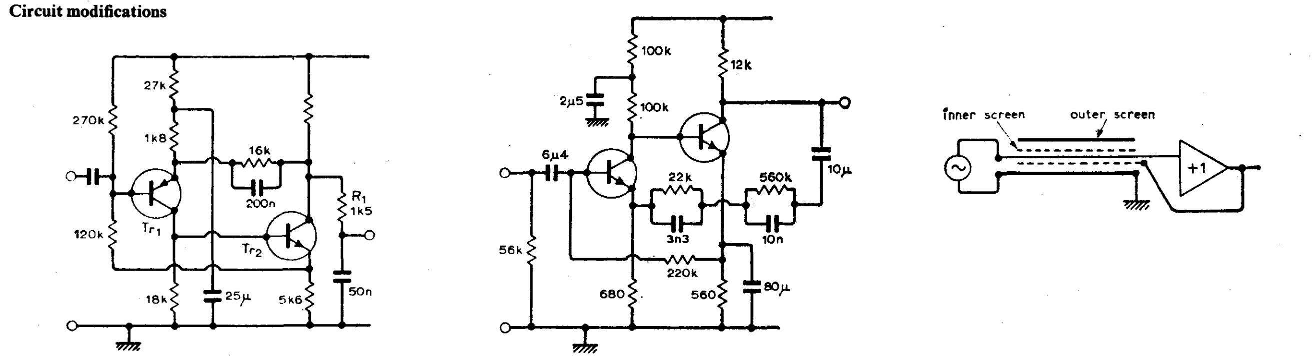
Circuit modifications

- Any amplifier capable of accepting shunt-derived series-applied negative feedback may be used. Feedback is considerable and the amplifier does not need high open-loop gain to provide accurate equalization using this network. A single-ended supply may be used and the circuit is tolerant of wide variations in this supply and of transistor characteristics. The load resistance must be $\gg R_1$ if the 75-μs time constant is not to be modified though in practice loads may be accommodated by raising R_1 and accepting some attenuation at all frequencies.
- Where the amplifier is to accommodate varying types of input shunt-applied feedback leading to a see-saw type amplifier is the usual solution; this allows control over input impedance and transfer function for a wide range of requirements. Again the gain requirements are low and one or two transistors may suffice in the amplifier.
- To minimize cable capacitance effects, twin-screened cable may be used with the outer screen grounded and the inner screen connected to the feedback path. By this means inner-screen capacitance is bootstrapped. Applicable to all series applied feedback circuits.

Cross references

Series 5, cards 4, 8 & 12.

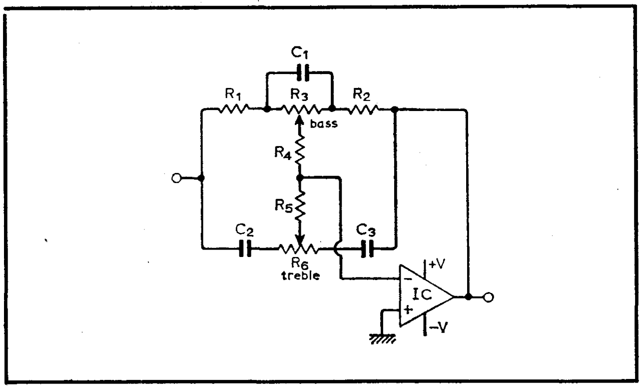
Circuit modifications



Wireless World Circard

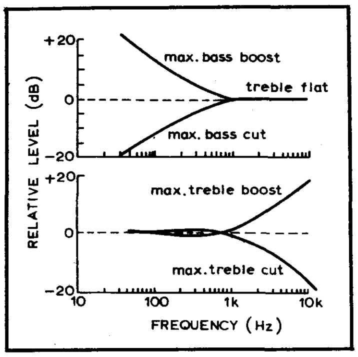
Series 5: Audio circuits 2

Tone control circuits/Baxandall



Typical performance

IC: 741
Supplies: ± 15V
R_1, R_2: 4.7kΩ $\pm 5\%$
R_3, R_6: 100kΩ
R_4: 39kΩ
R_5: 5.6kΩ
C_1: 47nF $\pm 5\%$
C_2, C_3: 2.2nF $\pm 5\%$
Input signal: 2V pk-pk
Source impedance: 60Ω



Circuit description

A tone control is a variable filter in which one or more of the elements is variable and which allows amplitude-frequency response of an amplifier to be adjusted. The above active circuit operates with a frequency-dependent feedback network and is based on the original Baxandall design. It has its greatest effect on the extreme bass and treble parts of the audio spectrum, and allows for separate bass and treble controls between which there is low interaction. The circuit features low distortion with maximum boost, and with the controls in mid-position the overall response is flat to within 1dB over the audio range. To ensure minimum restriction on the range of control available the source impedance of the driving circuit should be low. The component values used give the characteristics shown in the graphs with approximately 20dB of bass and treble boost and cut at 30Hz and 15kHz with respect to 1kHz, where the gain is unity. Excluding hum, the total harmonic distortion and noise is better than 0.5% over the range 100Hz to 10kHz, and better than 0.01% for input signals less than 100mV.

Component changes

To make small changes in overall response of the tone control circuitry, resistors R_4 and R_5 may be increased to double their present value, or reduced to zero, e.g. for R_4 zero bass boost and cut increased by approximately 3dB for frequencies below 1kHz; for R_5 zero, treble boost and cut increased by 2dB for frequencies greater than 1kHz.

Circuit modifications

- To obtain gain with low input signals, connect feedback network to potential divider tapping across the output as shown centre i.e. gain = $(R_7 + R_8)/R_8$. Useful values for a gain of 10: R_8: 470Ω, R_7: 4.7kΩ. To ensure that the input and feedback parts of the network give correct balance with the controls in mid-position, the source resistance should be equal to R_7 and R_8 in parallel.
- For a discrete-component output stage, use a field-effect transistor in preference to a bipolar transistor because of better linearity and lower noise level (see last reference), and the high input impedance minimizes loading on the tone control circuit. Emitter-follower output provides a low output impedance and thus the arrangement can replace the op amp.

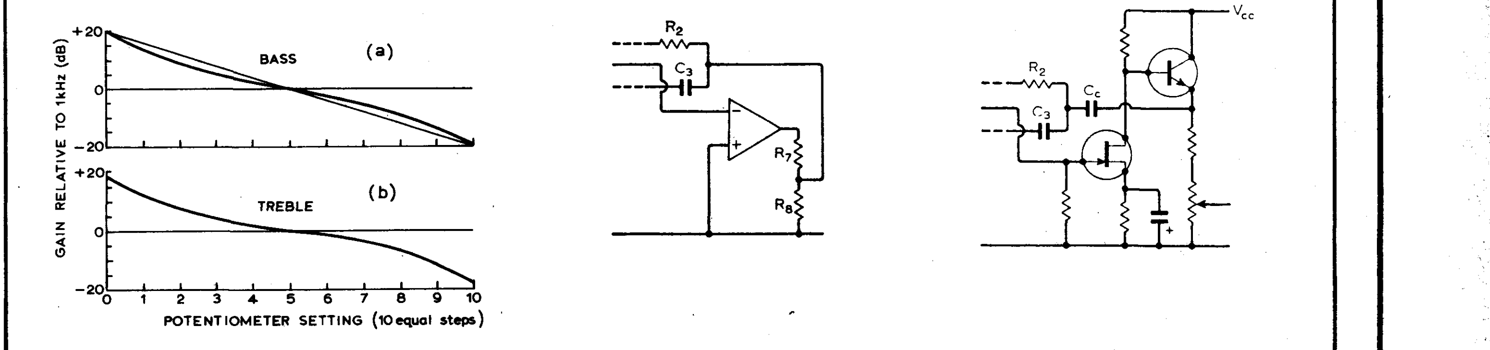
Further reading

Transistor Audio and Radio Circuits, Mullard, 1969, pp.131-9.
Baxandall, P. J., Negative-feedback tone control, *Wireless World*, vol. 58, 1952, pp.
Bailey, A. R., High performance transistor amplifier, *Wireless World*, vol. 71, 1966, p.601.
Quilter, P. M., Low distortion tone control circuit, *Wireless World*, vol. 77, 1971, pp.199-200.
Ambler, R., Tone balance control, *Wireless World*, vol. 76, 1970, pp.125/6.
Meyer, D., Audio pre-amplifier using operational amplifier techniques, *Wireless World*, July 1972, pp.309-12 & p.544 Nov.
Linsley Hood, J. L., Modular pre-amplifier design, *Wireless World*, vol. 75, 1969, pp.306-10.

Cross references

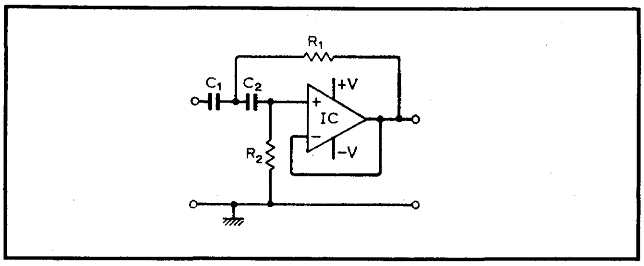
Series 5, card 6.

Circuit modifications



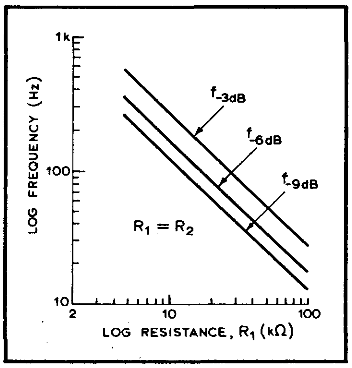
Wireless World Circard

Rumble filters



Typical performance

IC: 741
Supplies: ± 15V
C_1, C_2: 100nF $\pm 10\%$
R_1, R_2: see graphs
Input: 100mV r.m.s.
Source resistance: 60Ω



Introduction

An audio amplifier, having an amplitude-frequency response extending into the sub-audible range, can suffer from the reproduction of sub-audio-frequency signals especially when it is operated in its high output level region. Although these very low frequencies are themselves inaudible they can overload various parts of the system and/or produce intermodulation distortion and even audible tones by mixing with wanted signal components. These rumble signals may be produced from many sources such as turntable units, tape decks and microphones, in the latter case being largely due to sudden changes in breadth or due to wind when outdoors. In disc reproduction, rumble signals are generated due to the transmission of mechanical vibrations from the turntable unit to the stylus or even along the pick-up arm to the cartridge. Sometimes these rumble signals are of large amplitude due to resonances in the turntable mounting plate or plinth or due to transmission of the sub-audible slip frequency of an induction-type drive motor. To reduce the effect of these unwanted signals many amplifiers include a high-pass, or rumble, filter which can be of active or passive form. Often the filter is a fixed-response, passive design permanently connected in the pre-amplifier and consists of high-cost inductors having high-permeability screens and capacitors. A cheaper active filter can normally be designed using transistors or integrated circuits and located between the preamplifier equalization network and tone controls. Such filters can be switched out of circuit if desired and may have a variable response. The circuit shows a simple, second-order active filter having a cut-off frequency that is easily adjusted by means of R_1 and R_2 which could be in the form of a ganged potentiometer.

Component changes

Useful range of supplies: ± 3 to ± 18V. For a required cut-off frequency C_1 and C_2 could be made 1% components and R_1 and R_2 matched to give the same time constants. Ratio C_1/C_2 can be changed to alter response. Operational amplifier 741 can be changed to a 748 or 301 with a 30-pF compensation capacitor. Resistors R_1 and R_2 may be made continuously variable with a ganged log-law potentiometer.

Circuit modifications

It is possible to cascade a number of high-pass sections, of the type shown below, to obtain filters having a higher rate of cut-off, as the output impedance of each section is very low.

- The network shown left is a modification of the basic section requiring the addition of C_3, R_3, R_4 and A_2. Components C_3 and R_3 form an additional first-order section and R_4 is connected between the output of this section and the original second-order section. Amplifier A_2 is connected as a follower having a high input impedance so that the filter response is not significantly altered by the loading of the following stage. When $R_x = R_4$, V_{out} is in the form of a first-order high-pass filter response. When $R_x = 0$, V_{out} follows the signal at the output of A_1 which is in the form of a third-order response as the passive first-order section C_3, R_3 is then in cascade with the active second-order section; R_4 is then, ideally simply a high resistance shunt path. As R_x changes from R_4 to zero, the cut-off frequency and the rate of cut-off increase, so that it could be useful to choose the first-order and second-order section time constants to define the end-points of the response. Graph shows the variation obtained with a 741 as A_1, A_2; supplies of ± 15V; C_1, C_2: 100nF; R_1, R_2: 33kΩ; C_3: 10μF; R_3: 330Ω and R_4: 100kΩ.
- Circuit right shows an alternative high-pass section having a Bessel response and a cut-off frequency of $0.7/2\pi C_4 R_5$ when $C_4 = C_5 = C_6$ and $R_6 = 3R_5$.

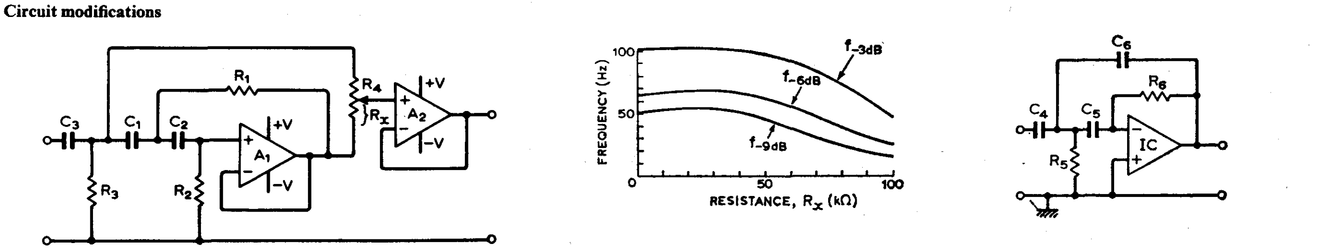
Further reading

Evans, J. H. and Williams, P., Modular integrated-circuit audio mixer, *Wireless World*, Dec. 1972, pp.564-70.
Gayford, M. L., Hi-Fi for the Enthusiast, Pitman 1971.
Earl, J., How to Choose Pickups and Loudspeakers, Fountain Press 1971.
SL403A & SL402A I.C. Audio Amplifier Applications Report, publication no. 4324, Plessey Microelectronics, 1970.

Cross references

Series 5, card 6.
Series 1, cards 3, 5 & 6.

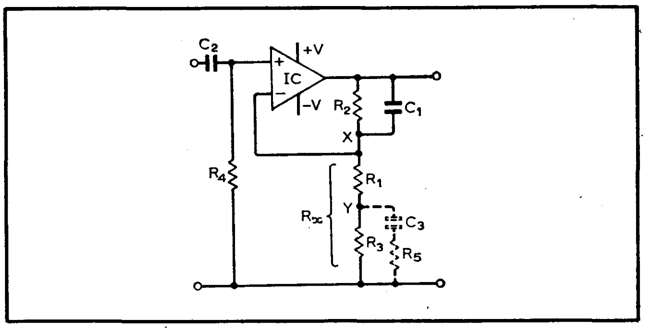
Circuit modifications



Wireless World Circard

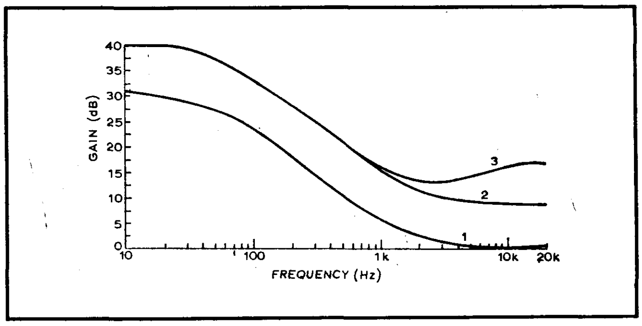
Series 5: Audio circuits 4

Tape-head preamplifier



Typical performance

IC: 741*
Supplies: ± 5 to ± 15V
R_1: 220Ω; R_2: 10kΩ
(9.5kΩ preferred)
R_3: 47Ω
C_1: 0.33μF ±1%
values suitable
for 9.5cm/s
R_4: 100kΩ; R_5: 22Ω
C_2: 47nF; C_3: 1μF
*not optimum for low
noise performance



Circuit description
The circuit is closely related to that of the magnetic pickup preamplifier having a transfer function of the form

$$k(1 + j\omega T_1)/(1 + j\omega T_2).$$

Time constants are determined in accordance with appropriate standards. The circuit shown achieves two time-constants using a single capacitor with $T_1 = C_1(R_X$ in parallel with $R_2)$, $T_2 = C_1R_2$. For widely-spaced time constants $R_2 \gg R_X$ and $T_1 \approx C_1R_X$ i.e. independent control of the time-constants is possible in practice by varying R_X and R_2 separately. At low frequencies the gain rises to a maximum of $(R_2/R_X) + 1$ falling at high frequencies to unity. This fails to make use of the high open-loop gain of the amplifier, providing equalization without further amplification. The circuit also lacks the facility for introducing high-frequency lift to help overcome the practical imperfections of head and recording medium (see modifications over). Input resistance of the circuit should be high and this could be accomplished by direct connection of the tape-head to the non-inverting input. This would however allow the amplifier input current to flow in the head, partially magnetizing it and increasing the resulting tape-noise. This effect may be avoided by using C_2 (low-leakage) and by regularly demagnetizing tape-heads.

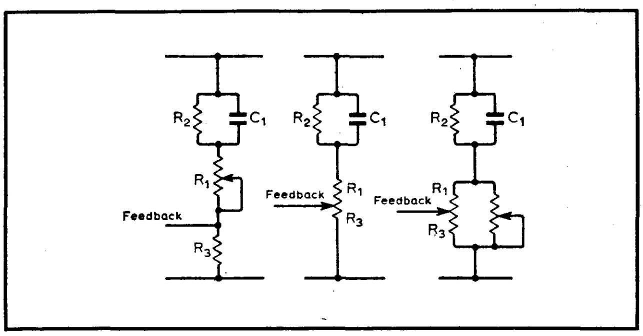
The head characteristic is such that a tape recorded with constant magnetization, irrespective of frequency, would produce an e.m.f. from the playback head proportional to frequency, as e.m.f. is proportional to the rate-of-change of magnetic field. This explains the need for the rising gain at low frequencies. A complex pattern depending on record bias levels, high-frequency fall-off in remanence, play-back gap dimensions, record and playback gap alignments, magnetic domain size in the tape determine the number and extent of corrective actions during the record-replay process. Limiting the gain to a constant value at high frequencies is covered by the equalization standard, other actions, particularly treble boost on record and playback being considered for each specific head, tape and bias combination.

Component changes

- Improved performance obtained by using specially designed low-noise amplifiers such as the National Semiconductor LM380/381 or the RCA quad-amplifier packages CA3048 or CA3052.
- Increase gain by connecting feedback from junction of R_1 and R_3.
- Approximately 6dB of treble boost achieved when C_3/R_5 network inserted, the upper limitation being determined by R_5.

Circuit modifications
These are concerned with the higher frequency range where $X_{C1} \ll R_2$.

- Network shown left is useful for variable tape speeds. If tape speed is reduced, time constant must be increased. With a lower tape speed, the output is reduced, but this will be matched by the increased gain obtained by increasing R_1 i.e. gain given by $(R_1 + R_3)/R_3$.
- Variable gain possible without alteration of time constant is available from circuit shown centre.
- Circuit on right shows an arrangement in which both a variable gain and time constant are independently achieved.

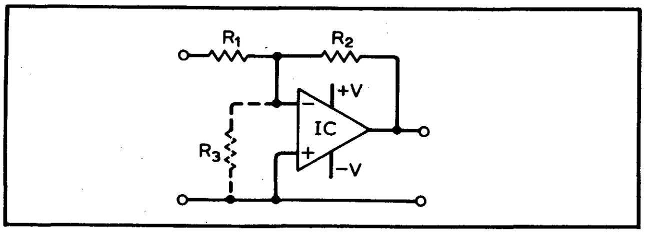


Further reading

Ridler, P. F., Transistor tape preamplifier, *Wireless World*, vol. 64, 1958, pp. 572-3.
Ridler, P. F., High quality tape preamplifier, *Wireless World*, vol. 68, 1962, pp.52-3.
Murray, J. S., Tape recorder circuits, *Wireless World*, vol. 68, 1962, pp.531-4 & 601-4.
Stuart, J. R., Tape noise reduction, *Wireless World*, March 1972, pp.104-110.
Stuart, J. R., High-quality tape recorder, *Wireless World*, vol. 76, 1970, pp.524-9 & 587-91.
BS 1568: Part 1:1970. Magnetic tape recording equipment.
National Semiconductor application note AN-64.
Linear Integrated Circuits and MOS Devices, RCA Databook series SSD-201.

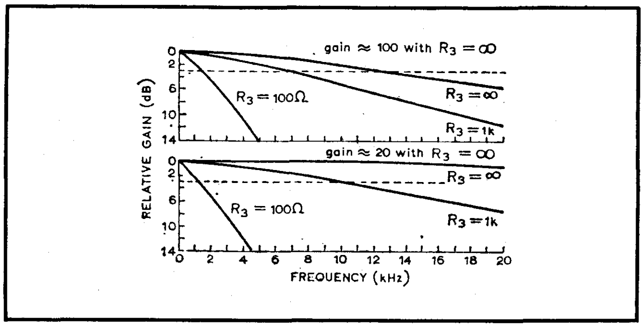
Cross references
Series 5, card 1.

Audio mixer/Operational amplifier



Typical performance

IC: 741
Supplies: ±15V
Input: 200mV pk-pk
Upper graph: R_1: 1kΩ; R_2: 100kΩ ±5%
R_3: ∞; 1kΩ, 100Ω ±5%
Lower graph R_1: 4.75kΩ; R_2: 100kΩ ± 5%
R_3: ∞, 1kΩ, 100Ω ±5%



Circuit description
This is a conventional summing amplifier using a 741 op-amp, with the gain defined by the ratio of the feedback resistor R_2, to the resistance R_1 connected to the inverting input of the amplifier, when a single source is connected. Open-loop gain-bandwidth product of the 741 is 1MHz, and if the aim is an amplifier with a gain of 100, bandwidth is then 10kHz. With multiple inputs the presence of the source resistance of these inputs connected to the nominal virtual earth point means that at high frequencies the signal is further attenuated and the bandwidth may be significantly less than 10kHz. The source resistances are effectively connected in parallel to ground and, for example, for ten 1-kΩ inputs, the effective resistance to ground is 100Ω. Hence with a feedback resistor of 100kΩ, the feedback factor is 1/1000, and from a feedback standpoint, the system corresponds to one for which the gain is 1000, and gain-bandwidth product is reduced to 1kHz. If a 3dB cut-off of 50kHz is desired to allow a safe margin, then this could be achieved for a gain of 20. If it is fed from, say, ten inputs simultaneously, then the allowed gain for each input would be 1/10 of 20, on the basis of the previous argument that it is the sum of the parallel admittances which determines the gain-frequency characteristic of the amplifier, and either a gain of 20 for a single amplifier, or ten separate gains of 2 can be achieved from the mixer. Thus an overall unity-gain mixer is a not-unreasonable circuit. Although it appears to waste bandwidth at first sight, it allows in practice, with ten inputs, a bandwidth approaching 100kHz.

Analysis
This shows why a large bandwidth cannot be achieved in a mixer which is simultaneously fed from a large number of inputs. The mixer may be in the form below left, where there are N inputs and the value of the feedback resistor is kR. As far as feedback is concerned, the circuit reduces to that shown centre and the magnitude of the effective gain is $|G| = kR/(R/N) = Nk$.

For the 741, $Nk \times$ bandwidth = 1MHz. If the bandwidth is 50kHz, Nk must be 20, and if N is 10, k is 2.

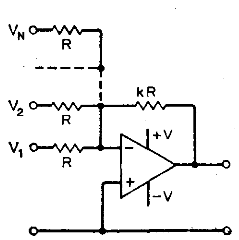
Component changes
Effects of parallel source resistance on magnitude and phase at 1kHz and 10kHz for nominal gain magnitudes of 100 and unit ($R_3 = \infty$) are tabled below R_2 = 100kΩ, R_1 = 1k or 100kΩ.

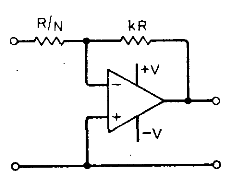
gain		frequency	R_3	phase lag
nominal	measured	(kHz)	(kΩ)	(deg)
100	95	1	∞	−180
100	70	10	∞	−240
100	90	1	1	−200
100	45	10	1	−260
1	unity	1	∞	−180
1	within	10	∞	−180
1	resistor	1	1	−180
1	tolerances	10	1	−215

In the last two observations for unit gain, the 1kΩ shunt resistance causes a d.c. shift of −250mV.

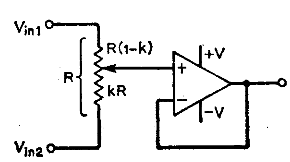
Circuit modification
To mix two signals, use voltage follower as shown below
Output voltage is $V_{out} = kV_{in1} + (1-k)V_{in2}$
An advantage is that the degree of mixing is continuously variable, with equal mixing when $k = 0.5$. However, $R \gg$ output impedance of the signal sources, and suitable drivers would be op-amps with $R >$ 10kΩ.

Further reading
Evans, J. H. & Williams, P., Modular integrated circuit audio mixer, *Wireless World*, vol. 78, Dec. 1972, pp.564-70.
Audio mixing preamplifier circuit, *Mullard Outlook*, vol. 15, no. 1. June 1965.

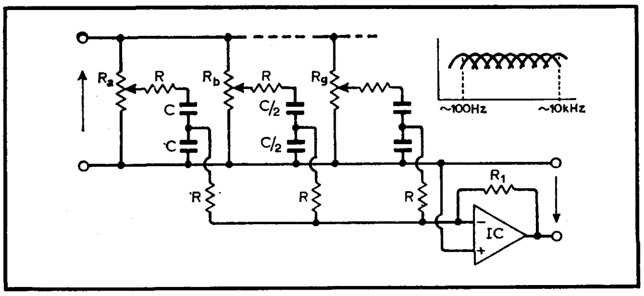




Circuit modification



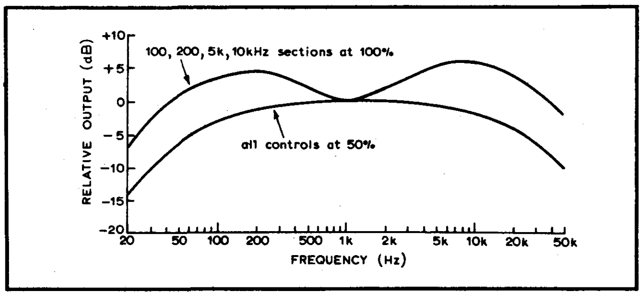
Multi-section tone control system



Typical performance

IC: 741
Supplies: ±15V
R_a to R_n: 10k
R: 12k
C: 100nF, 50nF, 20nF
10nF, 5nF, 2nF, 1nF
R_1: 1kΩ

n: 7 stages
For R_a $-R_g$ set to max.
output +0dB/−3dB
from 120Hz to 15kHz.
centre frequency: $1/2\pi CR$.



Circuit description

To modify the amplitude-frequency characteristic of an amplifier over limited parts of the spectrum, filters may be used centred on the appropriate frequencies. Variations on this method are unlimited, and the circuit shown represents one possibility, in which the output is proportional to the sum of the outputs of a number of passive filters. Centre frequencies of the passive filters are scaled in 1, 2, 5, 10 steps from 100Hz to 10kHz, and the *Q* is low enough that for equal inputs the ripple in the response curve is <1dB with −3dB points below 100Hz and above 10kHz. The summing amplifier operates on the currents flowing in the resistors forming part of what would normally be the parallel arms of the Wien networks. Raising or lowering the contribution of each network singly or in groups by means of the input potentiometers allows for moderate amounts of cut or boost in particular parts of the spectrum. If larger boosts are required then the passive networks can be replaced by active filters (see Circard series 1). The number of sections used must depend on particular requirements; even a single section feeding the summing junction together with direct input via a resistor allows for selective boost or cut at a given frequency. The larger the number of sections the higher the *Q* that may be used for a given ripple in the frequency response. The C, R values chosen would result in unloaded centre frequencies greater than the nominal 100, 200Hz etc. Using relatively high values for R_a to R_g lowered the frequencies closer to nominal for potentiometer settings close to centre positions.

Component changes

Potentiometer resistances should be low to avoid changing characteristics of Wien networks (output resistance of pot. included in series arm of network). A compromise is necessary as multiple pots. present heavy load to the audio source – range 1 to 100kΩ within above constraints.
C: 100 pF to 10μF chosen to give centre frequencies of $1/2\pi CR$.
R: 1k to 1MΩ
R_1: 1 to 100kΩ

Any number of sections may be used in principle with the restriction that the frequency limitations of the summing amplifier dictate the high-frequency response.

Circuit modifications

- Any band-pass amplifier may be added to the system, or replace an existing passive section, to increase the boost attainable at a given frequency. Such filters as those of Circard series 1 nos 1, 3, 6 & 8 may be used. As shown, the output is an inverted version of the input at the centre frequency and adjustment of the *Q* by varying the tapping point results in complex variations in the response, ranging from band-pass to a virtual notch. These methods must be applied with discretion but as the accompanying graph shows even a single active filter can produce dramatic variations in response (applications could include sound-effects, tone forming in electronic musical instruments).
- Combinations of low and high-pass circuits allow the raising and lowering of ranges of frequencies without the peaky response of band-pass circuits as above. Notch filters inserted in series can be used to tune out particular frequencies.

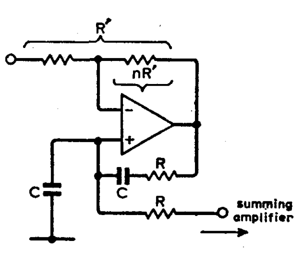
Further reading

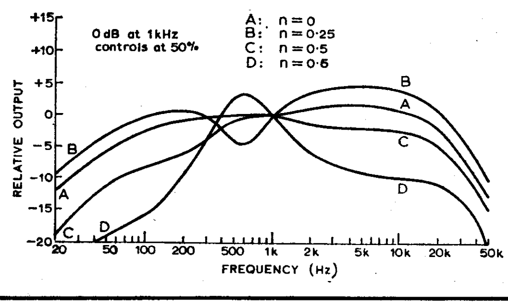
Ambler, R., Tone-balance control, *Wireless World*, vol. 76, 1970, pp.124-6.
Hutchinson, P. B., Tone control circuit, *Wireless World*, vol. 76, 1970, pp.538-40.

Cross references

Series 5, cards 2 & 10.
Series 1, cards 1, 3, 8 & 11.

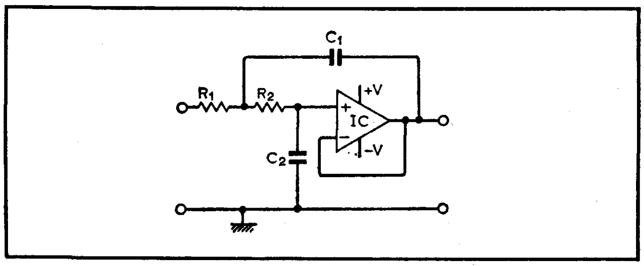
Circuit modifications





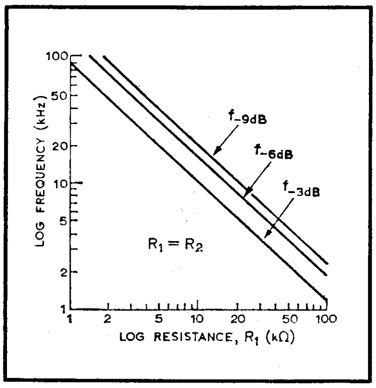
Wireless World Circard

Scratch filters



Typical performance

IC: 741
Supplies: ±15V
C_1, C_2: 1nF ±1%
R_1, R_2: see graphs opposite
Input: 100mV r.m.s.
Source resistance: 60Ω



Circuit description

Many audio amplifiers provide low-pass or top-cut filters to define the upper limit of the frequency response. These filters, normally incorporated between the source equalization pre-amplifier and the tone control circuit, are used to eliminate unwanted signals such as tape hiss, surface noise and scratches from old discs, radio interference, bias oscillator pick-up. Although some undesired h.f. signals are outside the audible range they can nevertheless overload the amplifier or introduce inter-modulation distortion components that may render the output unsatisfactory unless they are attenuated by a low-pass filter. Such filters may be passive or active networks or a combination of both types. Whatever its form, the low-pass filter will normally be switched in or out of circuit as desired. Cut-off frequency of the filter may either be infinitely-variable over a wide range or selected by a multi-position switch, typical values being 4, 8, 10, 12 and 15kHz. Selection of the cut-off frequency will normally be made subjectively depending on the amount and nature of the high frequency noise. While the rate of cut-off could also be chosen subjectively it would not normally exceed about 18dB/octave due to the increasing likelihood of severe transient distortion or ringing in the region of the cut-off frequency as a result of the amplifier becoming conditionally stable.

The circuit shown is an example of a simply-designed, second-order active low-pass filter using RC networks that can provide a wide range of cut-off frequencies with a cut-off rate that is well within the above-stated maximum. By suitable choice of components its input impedance can be adjusted so that it does not significantly load the preceding pre-amplifier and its output impedance is low due to the operational amplifier.

Component changes

Useful range of supplies: ±3 to ±18V.
For defined cut-off frequency C_1 and C_2 may be changed with corresponding change in R_1 and R_2 to give same time constants.
Ratio C_1/C_2 may be varied to change response.
741 operational amplifier may be replaced by a 748 or 301 using a 30-pF compensation capacitor.
R_1 and R_2 may either be switched with a two-pole unit or made infinitely variable with a ganged log-law potentiometer.

Circuit modifications

- Low-pass filters capable of providing a wide range of cut-off frequencies and a variable rate of cut-off tend to become complex networks. Circuit on left shows a modification of the basic circuit which provides a compromise between network complexity, variable cut-off frequency and rate of cut-off. Components A_1, R_1, R_2, C_1 and C_2 form the basic second-order active filter and R_3 and C_3 form an additional first-order passive section. Potentiometer R_4 is connected between the outputs of the first-order and second-order networks. When $R_x = R_4$, V_{out} provides a first-order response cascaded first-order and second-order networks. R_4 must be reasonably large to prevent significant changes in the time constants at the extreme values of R_x and A_2 serves as a high-input-impedance buffer to avoid excessive loading of the network. Between the extreme settings of R_x the filter provides a variable cut-off frequency and a variable rate of cut which increases as the cut-off frequency falls. The graph shows this effect with A_1, A_2:741; supplies of ±15V; C_1, C_2:1nF; R_1, R_2: 10kΩ; C_3: 10nF; R_3: 1kΩ and R_4: 100kΩ.

- Of the many alternative forms of active low-pass filters one example is shown right which provides a Bessel response – a good compromise between sharpness of cut off and transient overshoot – when $R_5 = R_6 = R_7$ and $C_4 = 3C_5$ with a cut-off frequency of $1.4/2\pi C_4 R_5$ Hz.

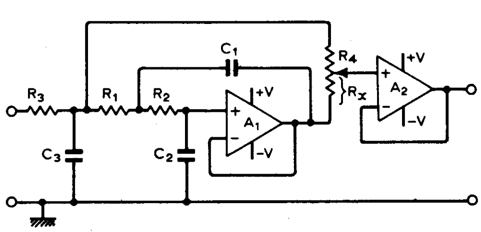
Further reading

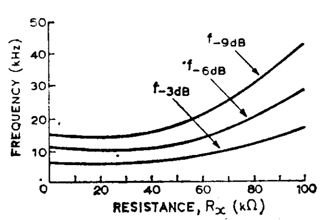
Evans, J. H. and Williams, P., Modular integrated circuit audio mixer, *Wireless World*, Dec. 1972, pp.564-70.
Gayford, M. L., Hi-Fi for the Enthusiast, Pitman, 1971.
SL403A & SL402A I.C. Audio Amplifier Applications Report Publication No. 4324, Plessey Microelectronics, 1970.

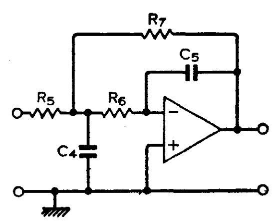
Cross References

Series 5, card 6.
Series 1, cards 3, 4, 6 & 11.

Circuit modifications

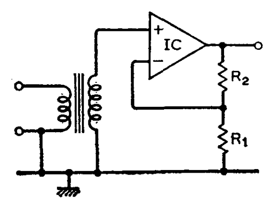






Wireless World Circard

Microphone preamplifiers



Typical performance

IC: 741
Transformer: see note
R_2: 1kΩ
R_1: 100Ω
Source resistance: 200Ω
Response: 45Hz to 20kHz, +0dB/−3dB
(Application of small bass-boost as in card 2 gave 20Hz to 20kHz ±1.5dB
Equivalent input noise: 0.5μV r.m.s. 20Hz to 20kHz
Noise reduced by 2dB by reducing bandwidth to 10kHz.

Note:
Performance of the microphone amplifier is strongly affected by the transformer. Type used was U204 by AKG, having a 15:1 turns ratio, a secondary inductance high enough to avoid l.f. limitation, and sufficiently low self-capacitance to prevent h.f. losses. Microphones may have specific transformers designed for them and the characteristics of the two should be considered together.

Circuit description

High-quality moving-coil microphones have a low output e.m.f. and a low internal resistance (< 1mV and 100-500Ω). Coupled directly to any semiconductor amplifier whether discrete or integrated, the resulting signal/noise ratio would be unacceptable as "hi-fi" but might be adequate for some applications (s/n ratios of about 45-50dB being the limit attainable with low-cost i.cs and low-sensitivity microphones). As the microphone has a relatively low resistance the dominant noise effect is due to the equivalent noise voltage at the amplifier input—the noise current fails to develop any significant p.d. across the microphone resistance. By transformer-coupling the signal to the amplifier input, and using a step-up ratio of 10:1 or greater, the signal e.m.f. can be greatly increased; no change in noise voltage generated at the amplifier input occurs but the noise current produces a proportionately larger contribution because the effective source impedance seen by the amplifier input is n^2R_s where n is the step-up turns ratio and R_s the microphone resistance. The optimum turns ratio which results in equal contributions from voltage and current noise generators may not give the optimum from a standpoint of amplitude response; with too high a turns ratio the resulting high impedance may be shunted heavily by capacitance at high frequencies, while the secondary inductance limits the low-frequency response.

Component changes

The amplifier gain bandwidth product is 10^6Hz and to leave a safe margin for amplifier limitations of upper cut-off, gains of < 50 should be accepted. The transformer has a turns ratio (e.g. 5 to 50) and the amplifier output may be up to 500 × the microphone e.m.f. though 100 to 200 is more likely. Keep R_1, R_2 low from low-noise standpoint e.g.

R_1: 10 to 500Ω.
R_2: 100 to 5kΩ.
R_2/R_1: 5 to 50.

Coupling capacitor may be used at output though d.c. content is small.
Output stage of op amp is class B. Where crossover distortion is a worry, even though negative feedback is large, stage may be biased into class A by taking resistor from output to one supply line, drawing current greater than expected peak load current.

Circuit modifications

- Any non-inverting amplifier may use series-applied feedback in the same way. If conventional transistor amplifiers are used, one approach is to define the d.c. conditions separately and capacitively couple the transformer output to the amplifier. The method shown left uses the low d.c. resistance of the transformer to act as a shunt-applied d.c. negative feedback from second emitter to first base. If R_4/R_5 is relatively large (high closed-loop gain) then p.d. across R_3 is V_{be} of Tr_1, ≈0.6V for silicon. Convenient resistors might have $R_2 = 5R_3$, $R_1 = 10R_2$, $R_2 = 10R_4$, R_4/R_5 = required gain, $0.6V/R_3$ = standing current in Tr_2; all for V_s = 5 to 15V. The values may be chosen to give a standing current in Tr_3 convenient for the required load swing, or if that is not a limiting factor, the currents may be lowered to raise the effective input impedance. The ratios are neither critical nor optimal, but are working guides.
- Using the virtual earth idea, an extremely simple self-biased stage gives no voltage gain but a lower output impedance at negligible cost. It may be similarly applied to the d.c. feedback-pair circuit or as shown to an alternative operational amplifier arrangement.

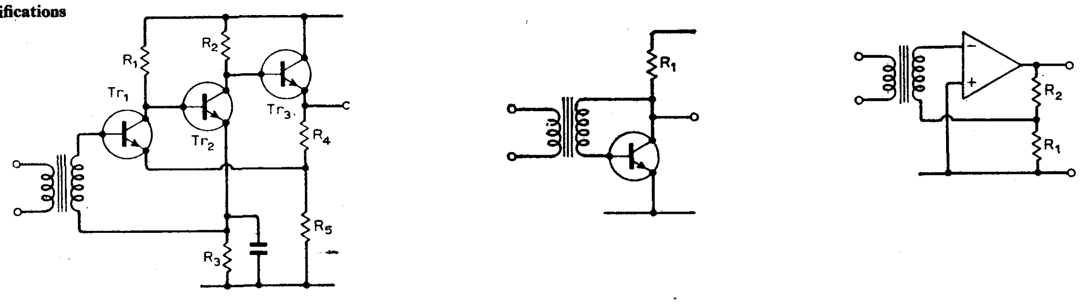
Cross References

Series 5, cards 2 & 12.

Further reading

Walker, H. P., Stereo Mixer - 1, *Wireless World*, vol. 77, 1971, pp.221-5.
Walker, H. P., Low-noise Audio Amplifiers, *Wireless World*, vol. 78, May 1972, pp.233-7.

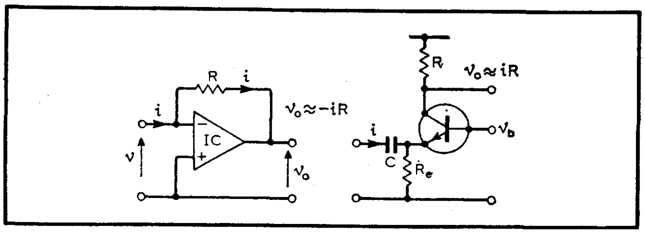
Circuit modifications



Wireless World Circard

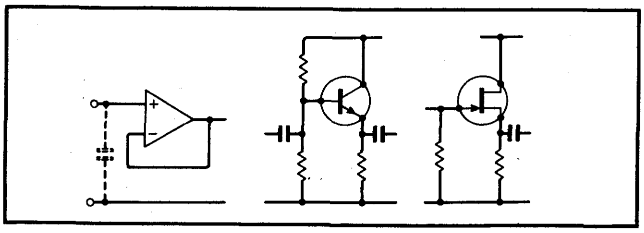
Series 5: Audio circuits 9

Impedance matching and transforming



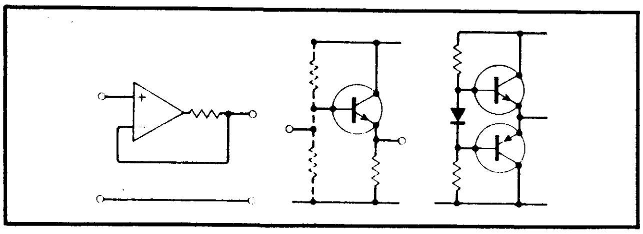
Circuit description

Shunt-applied negative feedback reduces the input resistance of an amplifier. To a first-order, for A large the input impedance is R/A, commonly called Miller effect, properly due to Blumlein. As A is frequently dependent, and for amplifiers such as 741 has 90° phase lag for 10Hz < f < 1MHz, the feedback current lags on the applied voltage i.e. input impedance is low, predominantly inductive and is hence proportional to frequency. The assumption that the input is a virtual earth is justifiable in most cases, but may not be so at upper end of audio range ($Z_{in} \sim j1000\Omega$ at $f = 10$kHz for $R = 100$kΩ). If d.c. conditions not important, common-base amplifier has low input resistance ($\sim 25\Omega$ for $I_c = 1$mA) and excellent frequency response. Lower cut-off frequency determined by C and source resistance. Output may be taken at collector either as current into external load (note quiescent d.c. current) or as voltage developed across R. Values of V_b, R_e determine quiescent conditions. Input resistance non-linear and input current should be restricted to small fraction of standing current. Input resistance $\propto 1/I_c$.



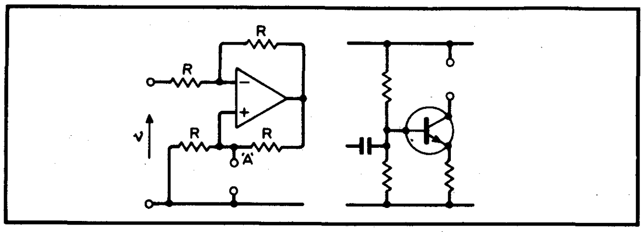
Circuit description

The basic follower circuits (voltage, emitter and source follower) have a.c. voltage gains close to unity, supply appreciable load currents and ideally draw negligible current from the source i.e. giving high input impedance. At high frequencies the finite open-loop gain of the operational amplifier together with its phase lag contribute additional shunt capacitance effects at the input depending on the amplifier input resistance. In most cases the total input capacitance is likely to be dominated by the physical system capacitances; minimum device capacitances below 1 to 3pF are rare, total capacitance may be around 10pF. For high source resistance (>100kΩ) this may bring cut-off frequency below 100kHz but not serious in audio band. Most likely limitation remains falling gain of amplifier if set for high voltage gain initially. See card 5 on mixer circuits. Input impedance of discrete circuits may be high particularly f.e.t. (>1MΩ) but a.c. coupling required, at least at output.



Circuit description

Amplifiers such as the 741, 301 etc. have open-loop output resistances of 50 to 150Ω, relatively frequency independent up to the useful limit of operating frequency. Falling gain of the amplifier at high frequencies does not allow the feedback to reduce the effective output impedance to almost zero as it does at low frequencies. Output impedance is largely inductive and corresponds to a few tens of microhenries e.g. at 90kHz series resonance effects can be observed with a load capacitance of 100nF. Effects are minimal at audio frequencies except where high gain is being attempted i.e. feedback much less than unity. This property is distinct from the current limiting of the amplifier into too low a load resistance, where additional internal transistors clamp the output current to $\sim$25mA. Again emitter followers, Darlington pairs may be used. For high output currents complementary emitter followers in class B are used but the subject is then properly treated as a power amplifier.



Circuit description

There is no simple way of obtaining high output resistance to a grounded load using operational amplifiers that have one output grounded. One possibility shown above uses positive and negative feedback combined to raise the impedance at point A close to infinity. This may also be viewed as a negative impedance converter circuit and has the possibility of an output resistance which while large may have either positive or negative sign. Load resistance should not be too large (say $< 10R$ if possible) so that the negative resistance effect cannot dominate. Output current is then determined by V and R. At high frequencies the amplifier gain/phase shift gives an equivalent output impedance that becomes capacitive. For a 741 using $R = 10$kΩ, a shunt capacitance of about 60pF was observed. Upper cut-off frequencies in excess of 100kHz are possible. The transistor circuit and its complementary version allow high impedance a.c. drive but superposed on a direct standing current.

Further reading

Evans, J. H. and Williams, P., Modular integrated circuit audio mixer, *Wireless World*, vol. 78, 1972, pp.564-70.

Series 5, cards 1, 8 & 11 high Z_{in}
4 & 5 low Z_{in}
5 fixed Z_{in}

Economy i.c. audio circuits

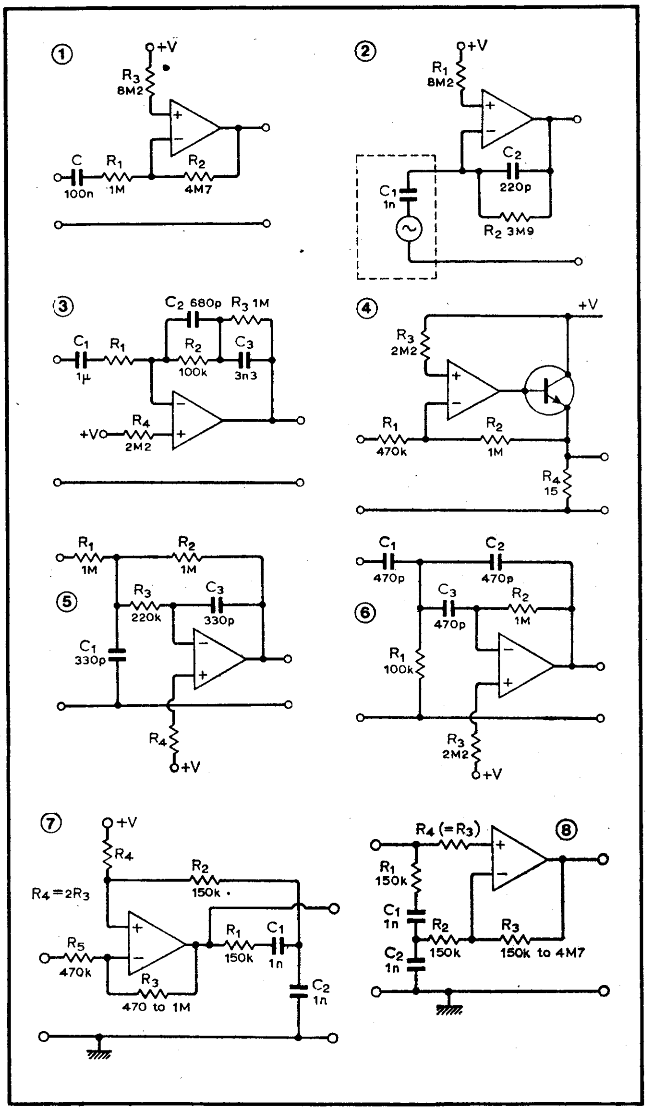
Circuit description
The i.c. contains four identical amplifiers each of which is similar to the d.c. feedback pair of a common-emitter stage followed by a common-collector stage. Feedback is applied externally, and the novel feature of the circuit is the second input forming part of a current mirror such that the effective current seen by the amplifier is the difference between the currents fed to the two inputs. The non-inverting input normally receives a separate bias from the positive supply or from any other convenient positive potential including the output of some preceding stage. Signal and feedback are most often applied to the inverting input with the system having properties associated with the see-saw amplifier familiar in op amp circuits. The difference is that the virtual earth is so only for a.c., having a d.c. potential equal to the input transistor V_{be} i.e. ≈ 0.6V. This configuration allows most functions normally provided by op amps to be achieved with a single-ended supply over a wide voltage range.

Circuit 1. A basic buffer amplifier with a voltage gain of $-R_2/R_1$ and an input impedance of R_1. If $R_3 = 2R_2$ the output voltage is approximately $V/2$. This is because the negative feedback through R_2 causes the current flowing in the inverting input to be an approximate match to operation by the d.c. at the output. Although large capacitances are then required ($>500\mu$F) if good low-frequency response is the aim. The low voltage rating keeps capacitor cost low. Heavy feedback keeps distortion low, assisted by class A operation. Bandwidth of 10Hz to 100kHz possible with distortion $<1\%$ in range 100Hz to 10kHz. Ripple induced hum may be minimized by centre tapping R_3 and decoupling to ground.

Circuit 2. Alternative capacitance feedback for inputs such as piezoelectric pick-ups (see card 11). Resistor R_2 necessary to provide bias, reduces low-frequency gain. Possibly centre-tap and decouple to ground using capacitance of 1nF to avoid peak in audio band.

Circuit 3. Magnetic pickup may also be accommodated using feedback network of type described on card 1. Correct bias obtained for $R_4 = 2(R_2 + R_3)$ with R_1 chosen to give correct input impedance. These amplifiers are not specified for low-noise performance but are still worth considering where economy is predominant aim. It is possible to combine the properties of 3 & 4 to give direct headphone drive with one stage from such an input, but at increased distortion and reduced bandwidth. Separate circuits for 3 & 4 give full stereo operation using a single i.c. package together with two low-cost output transistors. Major cost in such output stages must be that at the non-inverting input, and this can only be so when the output voltage is related to the supply voltage by the appropriate ratio of the resistors. An alternative bias method for an input with a d.c. resistance to ground $\ll R_1$, is to dispense with C and R_3 and choose the ratio R_2/R_1 to set the output d.c. conditions. This follows from the 0.6V at the inverting input which then defines the current in R_1. The same current flows in R_2 as that drawn by the inverting terminal is very small. Hence $V_O(\text{d.c.}) \approx [(R_2/R_1) + 1]0.6$V. The gain option is thus restricted by the d.c. setting as the ratio R_2/R_1 is involved in both.

Circuit 4. Where higher output currents are required, a simple class A stage can be added. Output current of the amplifier is up to 10mA, allowing a load current of 100 to 200mA to be achieved without difficulty. Operating with a supply voltage of ~ 5V dissipation in the output transistors is $\ll 1$W and low-cost plastic/epoxy units are satisfactory. Current levels are such that this is a very convenient output stage for low-resistance headphones—up to quadraphonic operation using identical configurations for each of the four amplifiers in the i.c. package. Capacitance coupling to the load is then indicated since the headphones would be driven into non-linear region (power supply etc.)

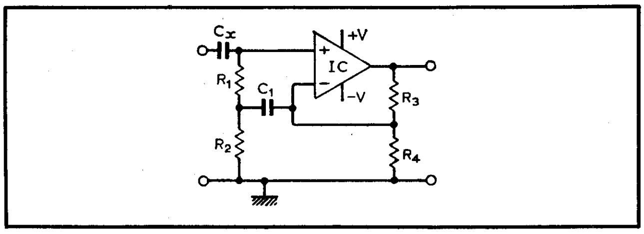


Circuits 5 & 6. Rumble and scratch filters may be constructed (cards 3 and 7) but as the voltage follower mode is not obtainable directly, different passive networks are used. Those shown are both for cut-off frequencies of around 1kHz. The damping may be adjusted without changing this frequency by varying the ratios C_2/C_1 for the scratch filter, 5, and the ratio R_2/R_1 in the rumble filter, 6. Increasing all capacitors in each circuit by a factor n reduces the cut-off frequency by the same factor.

Circuits 7 & 8. If more specialized filters are required in audio systems such as boosting response over a limited band e.g. to help overcome losses in tape heads at high frequencies, then active filters offer an alternative to LC passive sections. Circuit 7 is a low-Q bandpass filter of centre frequency $1/2\pi R_1C_1R_2C_2$ while 8 represents a simple notch filter of the same frequency, adjusting R_4 to obtain the notch. R_3 sets the gain and the bias is obtained from the d.c. level of the preceding stage.

Wireless World Circard

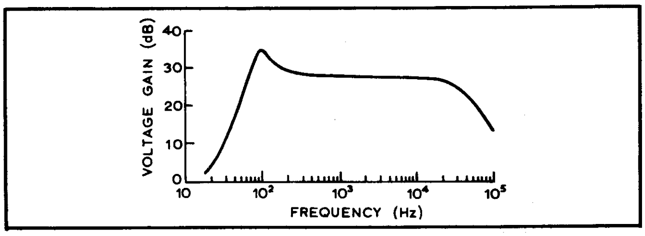
Ceramic cartridge preamplifier



Typical performance

IC: 741
Supplies: ± 15V
R_1, R_2: 10kΩ; R_3: 220kΩ; R_4: 22kΩ;
C_1: 22μF (tantalum).
C_x: 1.5nF (see circuit description)

Input: 100mV r.m.s.
Source resistance: 60Ω; (see curve opposite)
Note. Components chosen to show resonant peak.
Gain $1 + \dfrac{R_3(R_2 + R_4)}{R_2R_4}$



Circuit description

Unlike magnetic cartridges which produce an output proportional to the velocity of the stylus movement, a piezo-electric cartridge provides an output proportional to the amplitude of the stylus deflection and appears as an almost pure capacitance at the preamplifier input. The RIAA recording characteristic is essentially one of constant amplitude except for a small mid-band range of constant velocity recording. Piezo-electric cartridges are normally designed to provide mechanical equalization that caters for imperfections in the almost constant-amplitude recording characteristic. Thus, the preamplifier should provide a flat amplitude response and a high input impedance, not less than about 1MΩ. As the input impedance falls so also does the response at the bass end of the spectrum, leading to an approximation to an unequalized magnetic cartridge response. In the circuit shown, C_x represents the capacitance of a crystal or ceramic cartridge, normally in the range 400pF to 1nF. Bootstrap capacitor C_1 allows the cartridge to be lightly loaded by the high input impedance at the non-inverting input of the operational amplifier. To achieve this C_1 must be a high-value, high-quality capacitor such as a tantalum bead type. Then, the p.d. across R_1 is essentially zero as the high open-loop gain of the operational amplifier forces its differential input p.d. to be zero. Thus, no current flows in R_1 and the cartridge sees it as an open circuit. At low frequencies, the bootstrapping becomes imperfect and the inductive input impedance results in the resonant peak shown above. In the circuit shown below (left) $e = V/(1 + jC_1R_2)$ and $i = e/R_1$, hence $Z_{in} = V/i = R_1(1 + jC_1R_2)$ which contains an equivalent inductance $L = C_1R_1R_2$. This inductance resonates with C_x at $f_0 = 1/2\pi C_1R_1R_2C_x$ (88Hz for above component values). For good transient performance f_0 must be well below the audio band. See below for suitable values.

Component changes

Useful range of supplies: ± 3 to ± 18V.
To make resonant peak occur below 10Hz with C_1=22μF use $R_1 = R_2 =$ 180kΩ for $C_x =$ 400pF and $R_1 = R_2 =$ 120kΩ for $C_x =$ 1nF.
Ratio R_3/R_4 may be adjusted to give desired gain.
Lower values of R_3 and R_4 make gain less dependent on C_1, R_1 and R_2.
Screened wire between cartridge and amplifier reduces gain and resonant peak frequency e.g. 1 metre of separately-screened pair (136pF/m) reduces mid-band gain by about 1.6dB and f_0 from 88Hz to 78Hz.

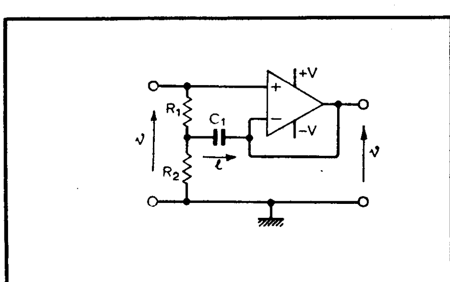
Circuit modifications

The circuit shown centre demonstrates a method of obtaining a flat amplitude response by making the time constant C_xR_5 and C_2R_6 equal, where C_x is the capacitance of the piezo-electric cartridge. With $R_6 = kR_5$ and $C_2 = C_x/k$ the gain of the amplifier is given by: $G = (kR_5 - jk/\omega C_x)/(R_5 - j/\omega C_x) = k$. Resistive loading of the cartridge is determined by R_5 which should not be less than about 1MΩ. The d.c. negative feedback provided by R_7 and R_8 will not significantly affect the response of the amplifier provided that R_7 and R_8 are both very much greater than R_6 and that C_3 adequately decouples them throughout the audio band. Incorrect choice of component values can again lead to a resonant peak at some low audio frequency, but as its Q is not high it could be used to give a degree of bass boost if desired.

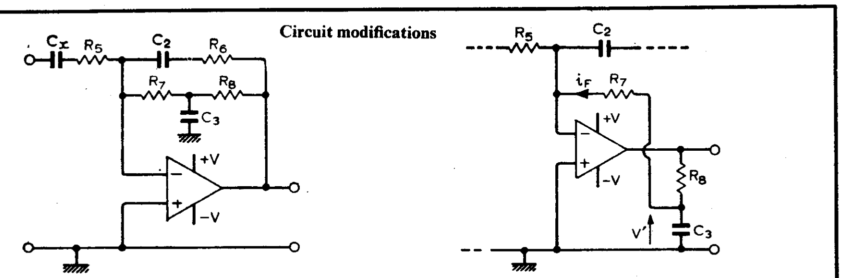
To see this effect, refer to circuit on right where $V' = V_{out}/(1 + j\omega C_3R_8)$ and $i_F = V'/R_7$. Hence the equivalent impedance (Z_{eq}) of the d.c. feedback network is given by $Z_{eq} = V_{out}/i_F = R_7(1 + j\omega C_3R_8)$. Thus, Z_{eq} contains an equivalent inductance $L' = C_3R_7R_8$. This inductance will resonate with C_2 (with which it is in parallel) at a frequency $f'_0 = 1/2\pi C_2C_3R_7R_8$ Hz. To demonstrate this effect, component values were chosen as: A: 741; C_x: 1.5nF; R_5: 1kΩ; R_6: 1.5kΩ; C_2: 1nF; R_7, R_8: 10kΩ; C_3: 22μF and the resonant peak found to occur at 108Hz compared with a predicted f'_0 of 107Hz.

Further reading

Evans, J. H. and Williams, P., Modular integrated circuit audio mixer, *Wireless World*, Dec. 1972, pp.564-70.
Burrows, B. J. C., Ceramic pickups and transistor pre-amplifiers, *Wireless World*, 1970. pp.56-60
Earl, J., How to Choose Pickups and Loudspeakers, Fountain Press, 1971.
A preamplifier for use with crystal and ceramic pickups, Design Note 14, SGS-Fairchild, 1965.



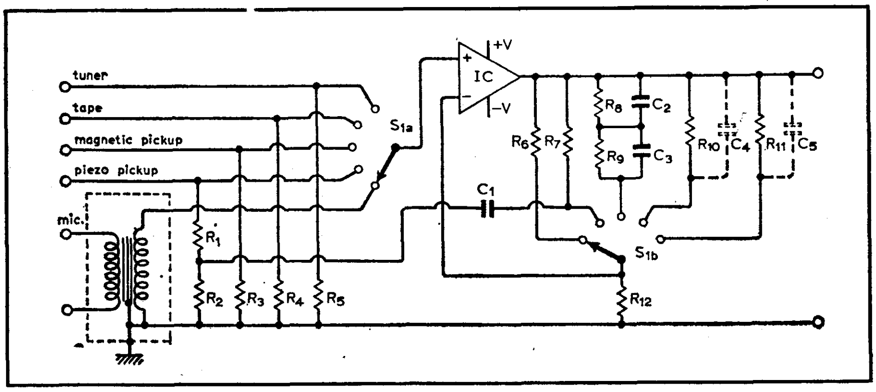
Circuit modifications



Wireless World Circard

Series 5: Audio circuits 12

Multi-input preamplifier



Typical performance

IC: 741
Supplies: $\pm 15V$
Transformer: AKG 0204
R_1, R_2: 150kΩ; R_3: 47kΩ
R_4, R_5: 470kΩ;
R_6: 8.2kΩ; R_7: 120Ω
R_8: 560kΩ; R_9: 47kΩ
R_{10}: 120Ω; R_{11}: 1.2kΩ
R_{12}: 820Ω
C_1: 22μF tantalum
C_2: 5.6nF; C_3: 1.8μF
Sensitivities:

For 350mV output at 1kHz
mic: 3.5mV
piezo: 308mV
mag: 4.82mV
tape: 310mV
tuner: 141mV

Circuit description

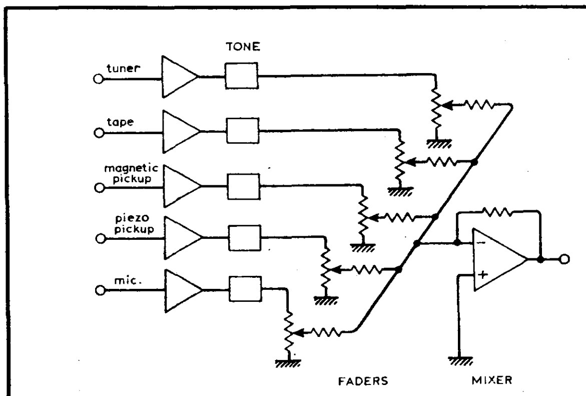
When designing a complete preamplifier circuit to accommodate a large range of signals from different sources, the output voltage to be fed to the main amplifier must normally be of the same value irrespective of which source is being used. The preamplifier must therefore provide different sensitivities for different sources as well as providing any equalization appropriate to a particular input. Many preamplifier circuits therefore use a single amplifier with passive input and feedback networks switched into circuit to meet the requirements of a given source. The circuit shown is of this type and uses a single integrated-circuit operational amplifier as the active block although a discrete transistor version could also be used. The operational amplifier is connected as a voltage follower with gain where the gain is defined by the feedback network components. Connected in this manner, the amplifier offers a basically high input impedance at its non-inverting input which can be reduced as far as the source is concerned by inserting suitable passive input networks where necessary. The amplifier shown accepts input signals from any one of five sources, viz magnetic microphone, piezo pickup, magnetic pickup, tape replay amplifier and a radio tuner. The amplifier was designed to give an output of 350mV at 1kHz giving, at this frequency, gains for the above inputs of approximately: 100; 1.15; 86.5; 1.15 and 2.47 respectively, with R_{12} chosen as a fixed value to simplify the switching arrangement. Components R_8, R_9, C_2 and C_3 provide an RIAA equalization when using the magnetic pickup input and R_3 was chosen to load the cartridge with 47kΩ.

Circuit description

Useful range of supplies: ± 3 to $\pm 18V$.
741 operational amplifier may be replaced by a 748 or 301 using a 30–pF compensation capacitor. Break-before-make switches could be replaced by make-before-break types to prevent sudden changes causing annoying clicks in the loudspeaker.
C_4 and C_5 may be included to roll off the flat response at high frequencies. The C_4 to R_{10} network could be changed to provide additional gain and the correct equalization when the preamplifier is fed directly from a tape head instead of from a tape replay amplifier.

Circuit description

- A resistor of the order of 10kΩ may be taken from the output of the preamplifier to supply a tape recorder.
- A disadvantage of the switched-network type of preamplifier is that only one signal source may be used at any given time. If each source is provided with a separate and appropriate preamplifier, any or all the inputs can be fed to a following mixer stage via faders making the system more flexible. Further versatility can be obtained by inserting a separate tone control circuit between the preamplifier and fader of each channel as shown in diagram left.
- As the various input sockets must be physically separated to some extent, the provision of a separate preamplifier for each source is useful as it can be placed in close proximity to its input socket in an attempt to reduce noise and hum pick-up.
- Diagram right shows a preamplifier using the see-saw configuration where networks X and Y are designed to provide the desired gain and equalization for a particular source, the networks again being introduced by switches.

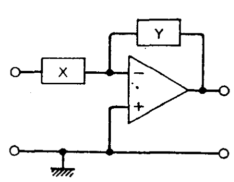


Further reading

Evans, J. H. & Williams, P., Modular integrated circuit audio mixer, *Wireless World*, Dec. 1972, pp.564-70.
Linsley Hood, J. L., Modular pre-amplifier design, *Wireless World*, vol. 75 1969, pp.306-10.
High Fidelity Audio Designs, Ferranti Ltd 1967.
Thorig, J., High-quality pre-amplifier, *Mullard Technical Communications*, vol. 11, 1970, pp.153-5.

Cross references

Series 5, cards 1, 4, 5, 8 & 11.



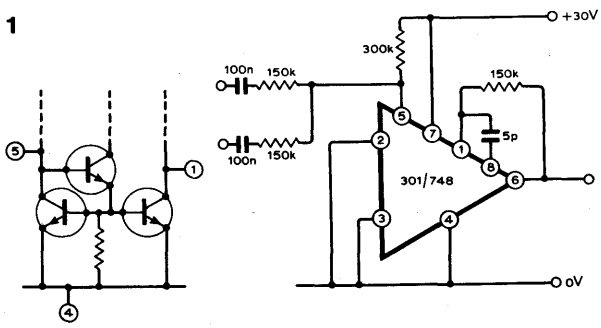
Wireless World Circard

Series 5: Up-date

Audio circuits

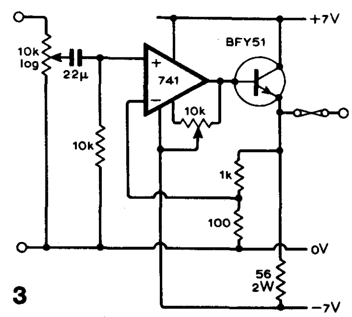
1. Standard operational amplifiers have terminals other than the usual inverting/non-inverting inputs into which signals may be injected, though the terminals are designed to allow balance compensation etc. An example is shown where the op-amp is made to behave as a current-differencing amplifier (equivalent to LM3900). The input and feedback paths have no direct connection and single-ended supply operation is shown. An audio mixer is given as an example, but the technique can be extended to produce astable/bistable circuits, RC oscillators etc with the added option that the normal inputs are available for the injection of synchronizing pulses if required.
June, W. G. Op-amp in current differencing mode becomes a noninverting audio mixer, *Electronic Design*, vol. 22, 1974, p.130 (May 10).

1



2. Analogue multipliers fed with a signal at one input and a single-polarity direct control voltage at the other give an output duplicating the signal in form but with a magnitude that can be reduced to zero for zero control voltage. If that control voltage is derived from a precision half-wave rectifier its value can be restricted to a given polarity, while depending on some combination of bias and control signals (see reference article for details). The offset network shown is based on that of the i.c. manufacturer, and allows the output to be set as a true product of the two inputs. A similar circuit is used in the voltage-controlled filter.
Orr, T. and Thomas, D. W. Electronic sound synthesizer: part 2, *Wireless World*, vol. 79, 1973, pp.429–34.

3



3. A headphone amplifier has to accommodate a wide range of load resistances (e.g. 8 to 100 Ω), have a low output impedance for good damping, and should meet high standards in terms of frequency response noise etc since the headphones now available are themselves of very high quality. A simple circuit meeting these requirements takes a standard operational amplifier and adds a class A emitter-follower stage to enable the voltage gain to be sustained into loads as low as 8 Ω. A further reduction in cross-over distortion can be achieved by adding a 1-k Ω bias resistor from the op-amp output to the negative supply rail (i.e. operating output stage of op-amp in class A). The transistor operates at ≈ 800mW dissipation with a quiescent collector current of 125mA and requires a heat sink thermal resistance of 50deg C/W.
Wherritt, P. Headphone amplifier, *Hi-Fi News*, vol. 19, 1974, pp.63-5 (March).

2