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INSULATED GATE BIPOLAR TRANSISTORS (IGBTs)

By our Editorial Staff

Insulated Gate Bipolar Transistors are being used more and more widely. Modern drive system engineering would not be the same without these devices.

IGBTs are ideally suited to applications where high voltages coupled with high currents must be switched. They are found in voltage converters, heavy-duty control circuits and audio power amplifiers. The first project that used IGBTs was published in this magazine in December 1994 (1-to-3 phase converter).

Noteworthy properties of IGBTs are the ease of voltage control and the low losses at high voltages. These characteristics make one think of power MOSFETs. However, the effective on-resistance of IGBTs is significantly lower than that of MOSFETs. The relevant symbols and equivalent circuits of n-doped and p-doped IGBTs are shown in Fig. 1. Note, however, that the symbols are (not yet) world-standardized, so that others may be encountered.

When the gate-emitter potential exceeds the threshold, $V_{GE(th)}$, a collector current flows. The current amplification, I_C/I_G, of an IGBT, is very high: 10^9, which is possible because the gate current only needs to charge the effective input capacitance (ignoring the tiny leakage currents through the gate oxide). Perhaps more appropriate is the transconductance, $\Delta I_C/\Delta V_{GE}$ (see Fig. 2). The definitions of saturation and linear range are the same as for standard bipolar transistors.

Construction of IGBT

An IGBT consists of a heavily boron-doped substrate (p+), which is fused on to the collector, on which a phosphorus-doped n-eptiaxial layer is deposited. The gate and emitter are epitaxial layers formed by a high-resolution n-channel DMOS process. Since the transistor therefore obtains an n-p-n-p structure, like a silicon-controlled rectifier (SCR), a p+ diffusion layer is inserted into the centre of the device. This layer reduces the current amplification of the upper n-p-n transistor and prevents any latch-up effects of the SCR. Without it, the IGBT would cut off at high currents owing to the breakdown of the gate control.

A separate p-base region permits independent control of the threshold voltage at the gate during the onset of conduction.

The maximum permitted reverse voltage is determined by the thickness and resistance of the n-epitaxial layer, which is optimized for a minimum forward voltage.

Switching characteristics

IGBTs (and power MOSFETs) have a gate-emitter threshold potential, $V_{GE(th)}$, and a capacitive reactance. To make these devices conduct, that is, before a collector current can flow, it is necessary for the input capacitance to be charged to a voltage that exceeds $V_{GE(th)}$. To switch off an IGBT, a resistor, R_{GE}, between gate and emitter is required, via which the input capacitance can be discharged. The minimum value of R_{GE} is specified in the data sheet of the relevant IGBT.

An IGBT has a peak controllable collector current that is dependent on the gate-emitter dv/dt transient. The higher this ratio, the lower the controllable collector current.

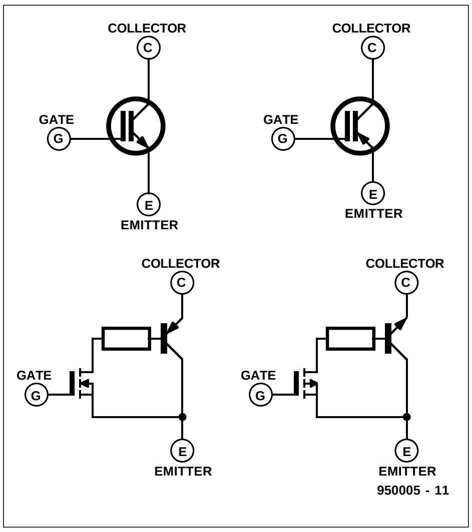

Fig. 1. Graphical symbol and equivalent circuit of n-IGBTs and p-IGBTs.

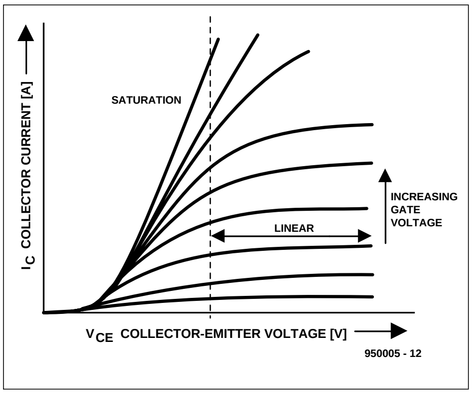

Fig. 2. Output characteristics of an IGBT.

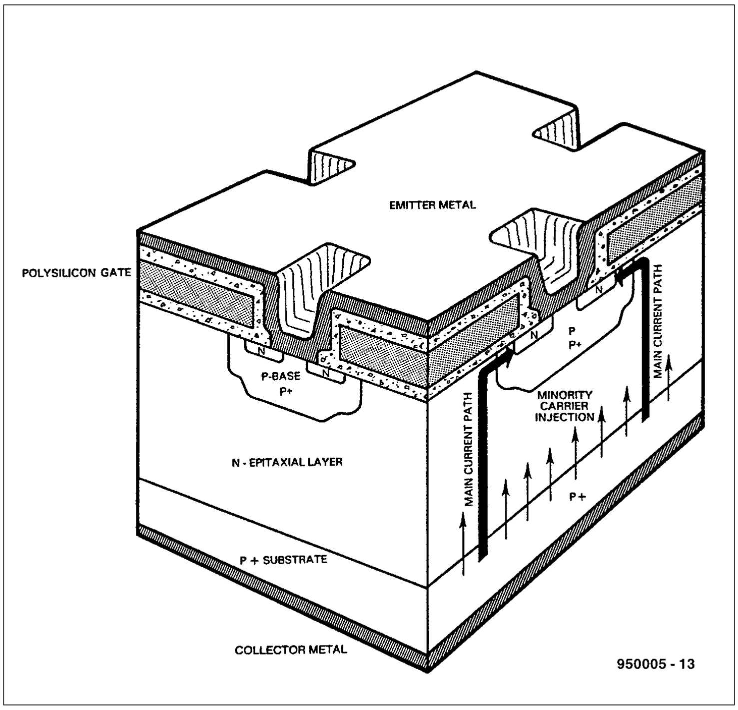

Fig. 3. Construction of an Insulated Gate Bipolar Transistor.

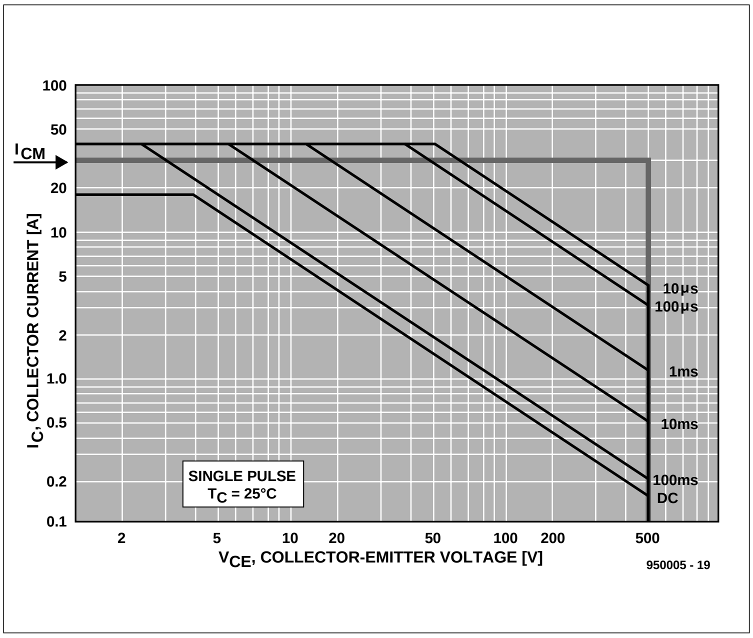

Fig. 4. Safe Operating Area (SOA) during normal operation and when the IGBT gets switched off. (peak values of I_C and V_{CE} shown).

Because of its construction, the switch on and switch off times of an IGBT are influenced by the gate-emitter impedance.

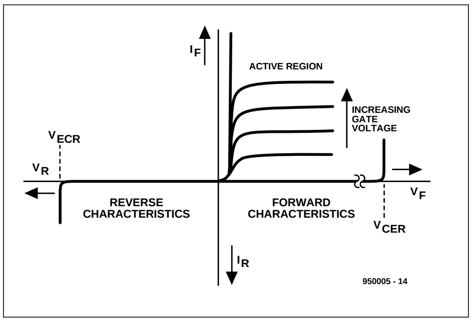

Fig. 5. Parametersof the collector characteristics.

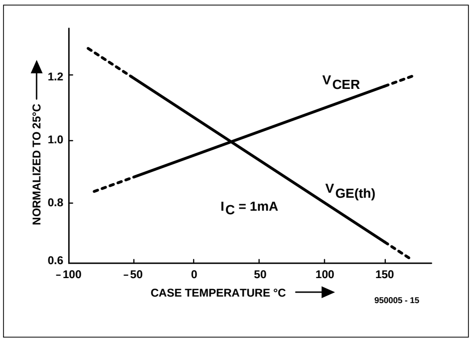

Fig. 6. Temperature dependence of V_{CE} and $V_{GE(th)}$.

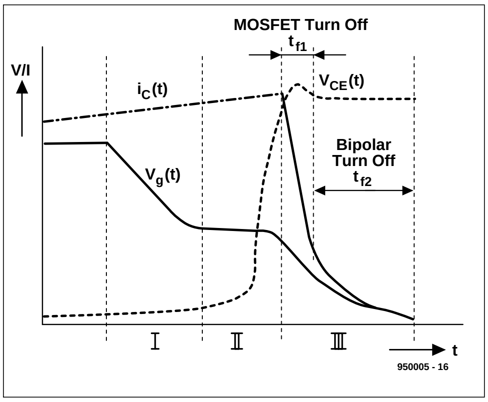

Fig. 7. Various phases in the switching off of an IGBT.

This impedance is much lower as that of a power MOSFET handling comparable voltages and currents. An IGBT is switched on by a positive potential on the gate and emitter terminals. When V_{GE} is greater than $V_{GE(th)}$, (which in switching applications is always the case), a collector current flows.

The switch-off behaviour of an IGBT is a mixture of that of a standard bipolar transistor and that of a power MOSFET. The switch-off time is determined by three different stages: I, II, and II, in Fig. 7. During the first phase, the gate-emitter voltage drops until the onset of the Miller effect (gate-collector capacitance) and V_{CE} rises. The second phase is typified by a constant gate potential (Miller effect). During this phase, the rising collector-emitter voltage causes a diminishing gate capacitance and a reversal of the gate polarity. Thereupon, the emitter voltage rises to a peak value, which is determined by the drive circuit. The final phase consists itself of stages: (a) the (very short) switch-off time, t_{f1}, of the MOSFET, and (b) the rather longer switch-off time, t_{f2}, of the bipolar transistor. The latter time does not commence until the MOS channel is off and the base of the p-n-p transistor is open.

Owing to the various phases of the switch-off time, it is difficult to put a value on the losses on the basis of the 10–90% switch-off time given in the data sheet. Instead, the equivalent switch-off time, $t_{F(eq)}$ is taken which assumes a linear decay of the collector current; however, the following expression gives the actual decay time:

$$t_f = 2/I\int_{t_o}^{\infty} i(t)\mathrm{d}t$$

If inductive loads are switched, the switch-off losses are given by

$$\tfrac{1}{2}\cdot V_{CE}\cdot I_C\cdot f\cdot t_{f(eq)}.$$

Since the switch-off period of the bipolar transistor part of the IGBT is really constant, the duration of the switch-off period of the MOSFET part can be influenced by the correct choice of R_{GE}. The higher the value of this, the longer the switch-off period. In case of an inductive load, the period may be protracted to such an extent that operation without a snubber network becomes possible.

There are slow and fast IGBTs. For applications of slow types (d.c. and a.f.), the minimum switch-off current is of prime importance, whereas with fast types the switch-off characteristics should be as linear as possible (Fig. 9). For high-frequency applications, fast IBGTs with small R_{GE} should be used; this ensures that switching losses are kept to a minimum.

The typical Safe Operating Area (SOA) of an IBGT is shown in Fig. 4. The device can handle peak currents exceeding the maximum collector direct current; this current is limited only by the thermal threshold and the thickness of the connecting leads.

Sources: 'IGBT Driver'; Toshiba, March 1993.
SGS Thomson: Technical Note 1/5.

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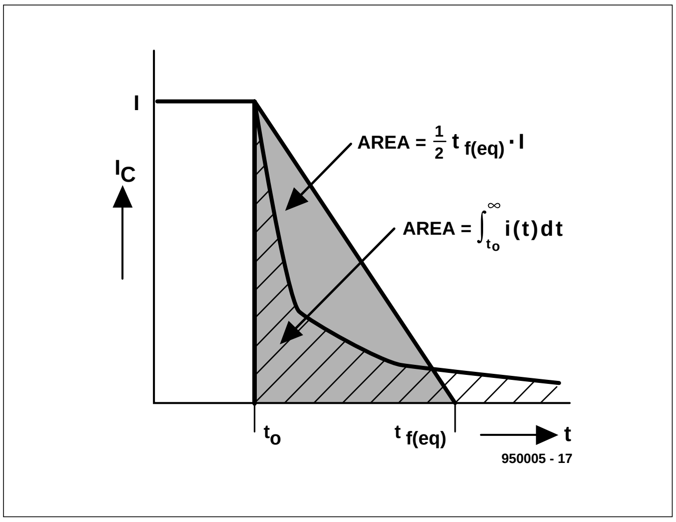

Fig. 8. Power losses at switch-off can be assessed from the equivalent decay times.

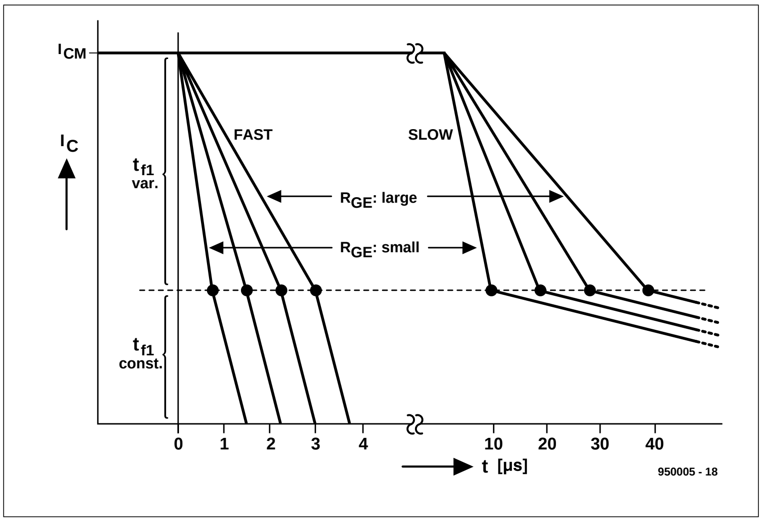

Fig. 9. Switching off of fast and slow IGBTs with various values of gate-emitter resistance.

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SURROUND SOUND PROCESSOR

Based on a design by D. Laues

The processor described in this article expands the sound of an existing stereo TV receiver or audio installation with a centre channel and a surround channel. It does not use special Dolby ICs. Additional output amplifiers are not needed, because they are provided in the processor.

Creating surround sound in a domestic room can be approached in two ways: by a processor that generates the four signals required: left-hand, right-hand, centre and surround, or by one that adds the two missing channels, that is, centre and surround, to the existing stereo sound. The first is the most elegant, but also the most complicated and most expensive. Moreover, it requires an additional line to return the left-hand and right-hand signals to the input of the TV/audio equipment. The second way is much more straightforward and has proved in practice to give an excellent spatial effect. Moreover, it can be accomplished in a compact and fairly inexpensive unit. The design in this article is of the second kind.

The basic setup in a domestic room is shown in Fig. 1. The left-hand and righthand channels are reproduced as before, that is, via the loudspeakers in the TV receiver, as shown, or by those of the audio installation to which the TV receiver is connected. The extra items are the processor and three loudspeakers. The inputs of the processor are linked, possibly via the SCART connector, with the line out terminal of the TV receiver, or audio amplifier, while the extra loudspeakers are connected to the outputs of the processor. The processor contains two integral amplifiers each of which provides 20 W output into 4 Ω: quite sufficient for the centre and surround loudspeakers.

As briefly discussed in last month's article on Dolby surround sound, the additional loudspeakers need be no more than compact (bookcase type) hi-fi types that are not too expensive. If possible, however, choose types whose efficiency is about the same as that of the main loudspeakers: this affords rather more freedom when the system is set up as a whole.

Basic design

A block diagram of the basic design is given in Fig. 2. The design of the processor is similar to that of the active decoder discussed in last month's article on Dolby surround sound. There are some differences between the two, since the present processor not only functions as a decoder, but also provides the signals for the centre and surround channels and contains two output amplifiers.

As explained last month, the two additional channels are processed (whence the name of the unit) from the sum and difference of the two stereo channels. The centre channel is the simpler to produce, because it suffices to add the left-hand and right-hand channels together and apply the sum to a suitable output amplifier via a voltage-controlled amplifier (VCA).

To produce the surround signal, the right-hand signal is deducted from the left-hand signal (or vice versa) and the resulting signal is applied to a delay network via a low-pass filter. The delay can be preset between 10 ms and 30 ms. The signal is then applied to an expander via a band-pass filter. The expander is essential because the surround signal is compressed during recording. The output of the expander is applied to a second output amplifier via a VCA.

The dynamic compensating network, in conjunction with the VCAs, reflects the difference between an active and a passive (matrix) design. In this network, the correlation between the two stereo channels is analysed continuously. The results of the analysis are converted into control signals for the VCAs which constantly adjust the levels of the centre and surround signals. This arrangement ensures a much larger channel separation than possible with a passive design.

Centre channel

From the inputs of the left-hand and right-hand channels, K_1 and K_2 in Fig. 5, the signals are taken via buffer amplifiers IC_{1a} and IC_{1b} to R_{51} and R_{52}, and then summed in preset P_6. From there, the signal is applied to pin 5 of IC_9. This IC contains four electronic potentiometer circuits (of which only two are used) which function as VCAs; their amplification is governed by a control voltage at pins 9 and 10. The output of one of the circuits is available at pin 7, from where it is applied to output amplifier IC_{10}. This circuit provides an output of up to 20 W into 4 Ω.

The output of IC_{10} is applied to the centre channel loudspeaker via relay contact Re_{1b}. The relay is controlled by a simple delay circuit, T_1, and obviates any clicks and plops in the speaker caused by the switching on and off of the processor.

Surround channel

The signals at the outputs of IC_{1a} and IC_{1b} are also applied to the inverting and non-inverting inputs of IC_{2b} respectively, so that the output of the op amp is the difference of the two stereo signals (L–R). The difference signal is applied to a 4th-order low-pass filter based on IC_{2a}, which limits its upper bandwidth to 7 kHz. This anti-aliasing filtering serves to obviate the formation of spurious mixing products of the signal and the clock of the following delay line, which is based on IC_4. This IC is a 2048-stage bucket brigade device. The rate at which the internal electronic switches are operated is

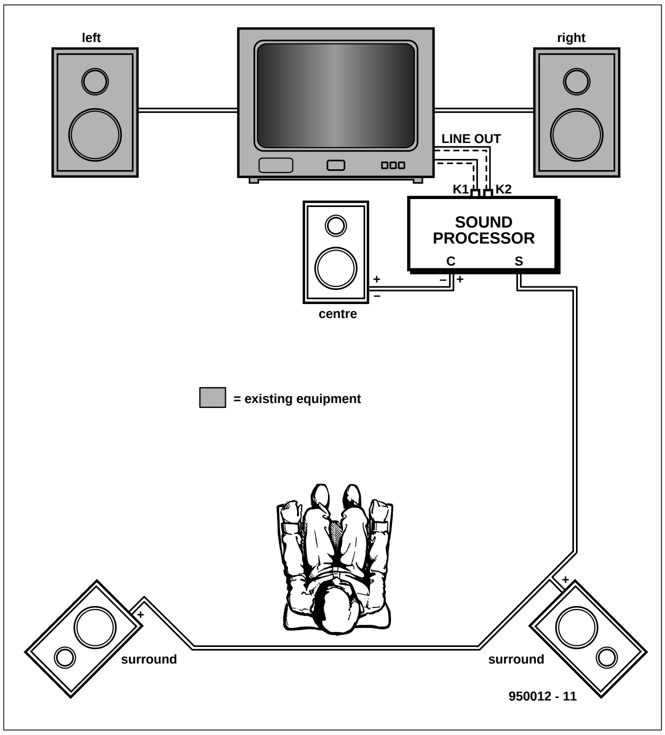

Fig. 1. Basic setup of a surround sound system in a living room.

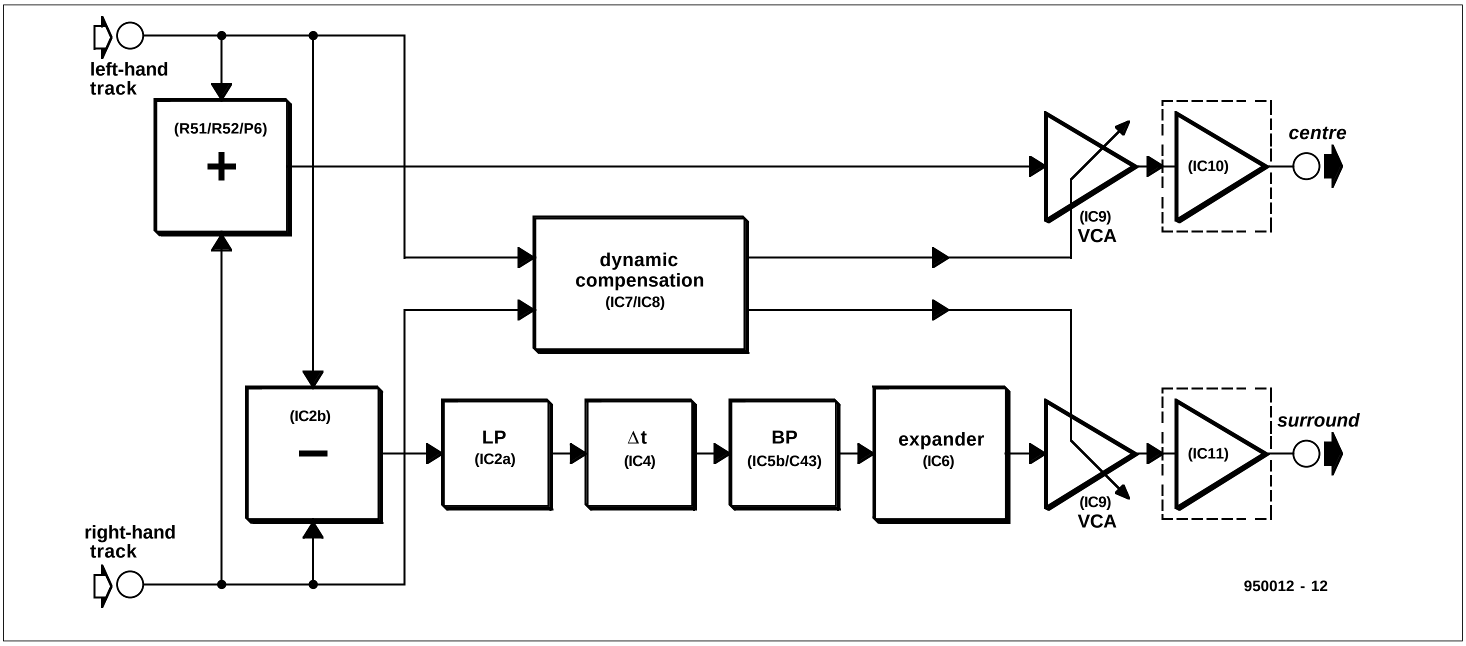

Fig. 2. Block diagram of the surround sound processor.

determined by IC_3. This CMOS-IC is designed especially to generate a low-impedance, double-phase clock. The specified values of its frequency-determining components, R_{15}, R_{16}, C_9 and P_1 allow a delay between 10 ms and 30 ms to be set with P_1.

The outputs of IC_4 are applied via buffer IC_{5a} to a low-pass filter based on IC_{5b} (identical to that based on IC_{2a}) which filters out any residue of the clock signal. The cut-off frequency is 7 kHz. The signal is subsequently fed to compander IC_6, whose input network contains a high-pass filter, R_{55}-C_{43}, the specified values of which give a lower cut-off frequency of about 50 Hz. The overall effect of the low-pass and high-pass filters is, of course, that of a band-pass filter as shown in Fig. 2.

The compander IC contains two circuits each consisting of a rectifier, a variable gain cell and an op amp. In the present processor only one of these circuit is used and that as an expander. The values of external components R_{27}–R_{32} and C_{23}–C_{29} allow for an expansion factor of 1:1.3.

The surround signal is then applied to the second electronic potentiometer circuit in IC_9, whose output is available at pin 17. From there, the signal is fed to output amplifier IC_{11}, whose amplification is identical to that of IC_{10}. The output of IC_{11} is applied to the surround loudspeaker(s) via a second contact on Re_1.

Dynamic compensation

The outputs of buffers IC_{1a} and IC_{1b} are also applied to twin comparators IC_{7b} and IC_{7c} via C_{31} and C_{32}. The output of each of these comparators is a rectangular voltage the frequency of which is a measure of the variation in the relevant input signal. Both outputs are applied to XOR gate IC_{8c}. (Remember that an XOR gate has an output only when its inputs are dissimilar). Integration of the output pulses of the gate by R_{37}-C_{36} results in a direct voltage whose amplitude is a measure of the phase difference between the two stereo signals.

This direct voltage is applied via IC_{7a} (inverted) and IC_{7d} (non-inverted) to the control inputs (pins 9 and 10) of IC_9. This arrangement ensures that when a mono signal is present at the inputs (no or hardly any phase difference), the amplification of the VCA controlling the centre channel is raised. Conversely, when a surround signal is present (large phase difference), the amplification of the VCA controlling the surround channel is increased.

The degree to which the amplification of the VCAs is influenced by the control signals is preset by P_2 and P_3. When the wipers of these controls are at earth potential, the amplification is fixed; when they are at the opposite end of their travel, control is maximum.

Presets P_4 and P_5 serve to shift the operating point of the VCAs to some extent. They thus make the preset range wider and, in fact, support the operation of P_6 and P_7. If, for instance,

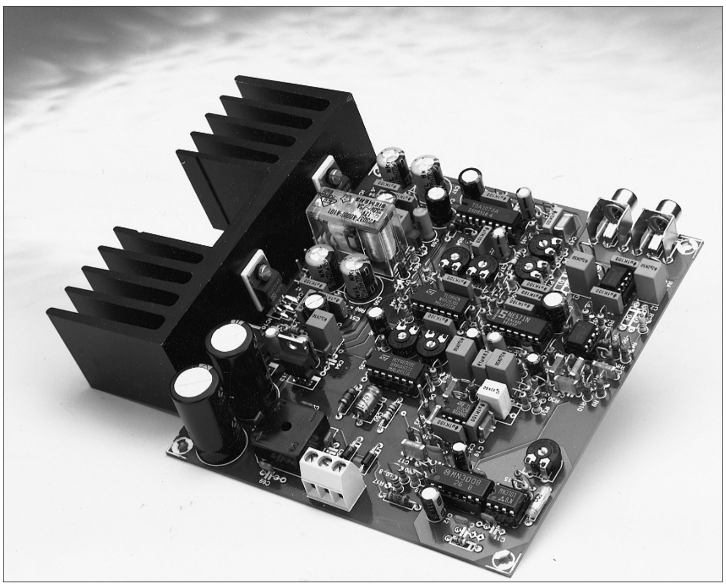

Fig. 3. Completed prototype board.

Fig. 4. Completed prototype processor.

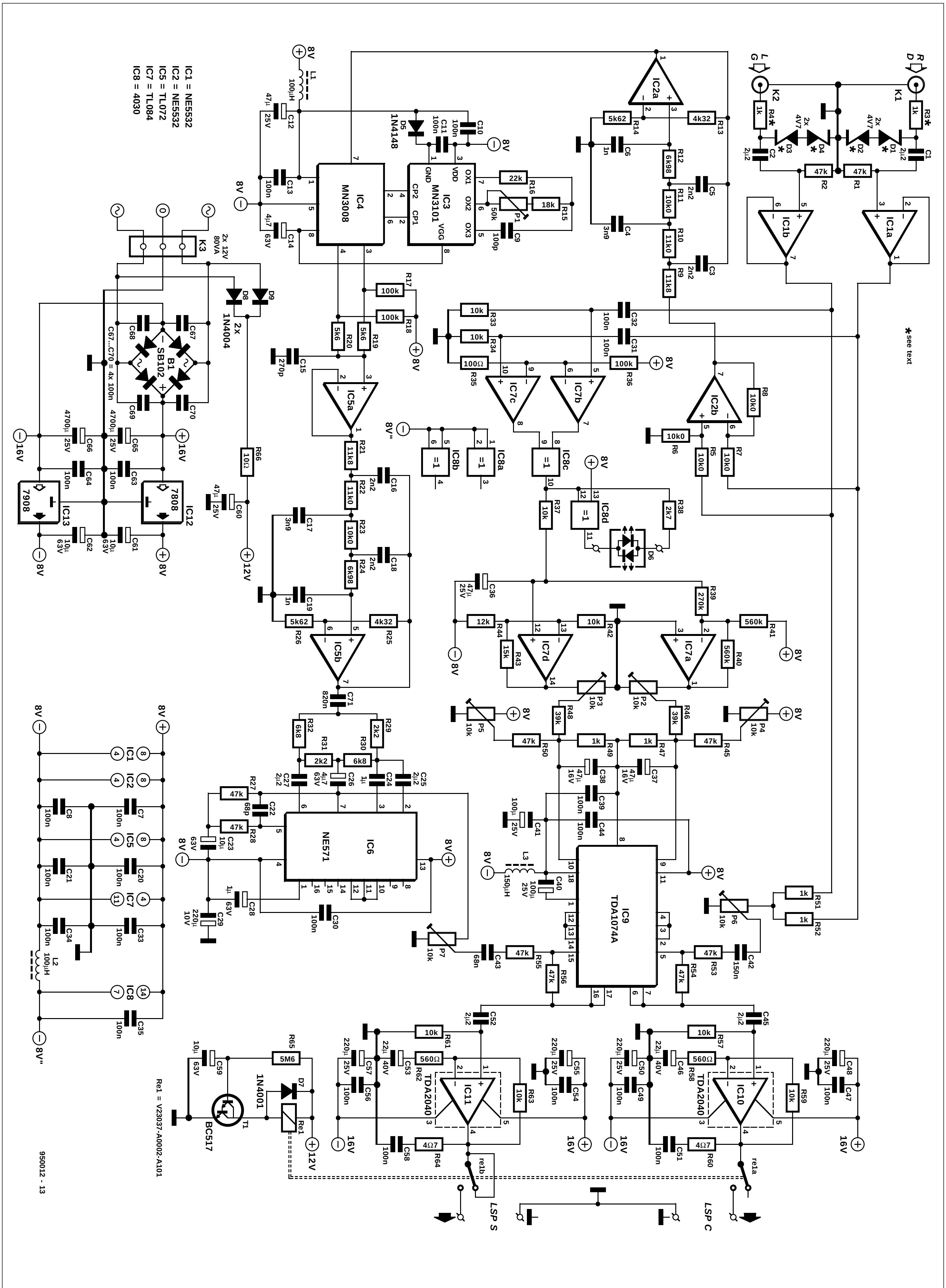

Fig. 5. Circuit diagram of the surround sound processor.

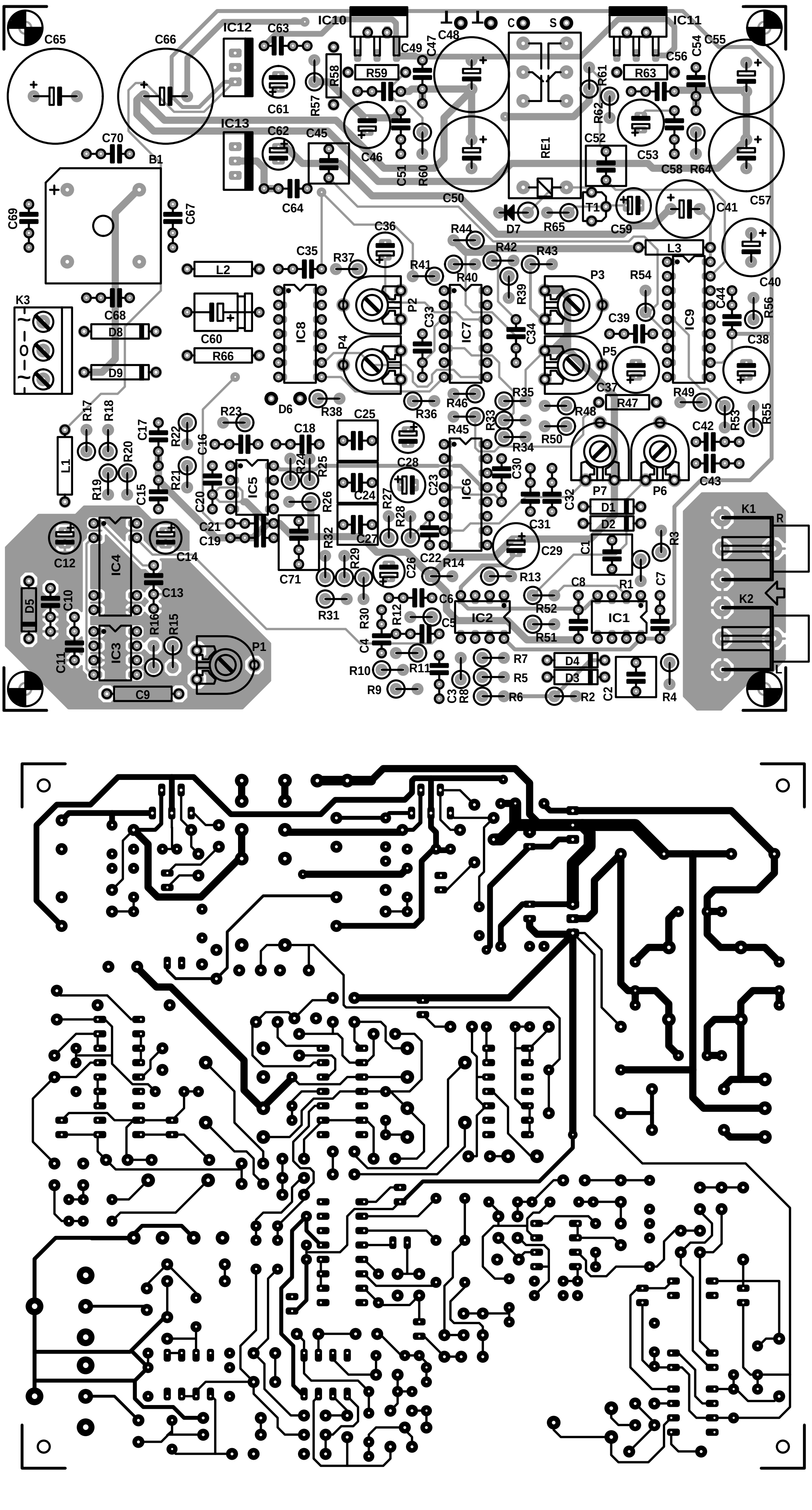

Fig. 6. Printed-circuit board for the surround sound processor. (See also next page)

P_7 has already set the surround level to maximum, P_5 enables this to be increased slightly. The same applies to P_4 insofar as the level of the centre channel preset with P_6 is concerned.

The currents through R_{45} and R_{46} and those through R_{48} and R_{50} are simply added together: there is, therefore, no interaction between P_2 and P_4 nor between P_3 and P_5.

Further circuit details

Resistors R_3 and R_4 and diodes D_1–D_4 limit the level of the input signal to a safe value and are imperative if the stereo signals are taken from the loudspeaker outputs of the TV receiver. Note that even line out terminals sometimes provide a signal at a level well above 1 V. If it is absolutely certain that the line output level is 1 V, and this is the only input, the resistors can be replaced by a wire bridge and the diodes may be omitted.

Bi-colour LED D_6 functions as a kind of signal monitor that shows the change from surround channel to centre channel and vice versa. In the case of a surround signal, the output of IC_{8c} is high. Since one output of IC_{8d} is at +8 V, both inputs of this XOR are then high, so that its output is low. This results in the red segment of D_6 lighting. In the case of a centre signal, pin 12 of IC_{8d} is low, so that its output is high, resulting in the green segment of D_6 lighting. In practice, the changes between the signal are so rapid that the LED shows a fluent transition from red to green and back to red again.

The power supply provides three different voltages. The secondary of the mains transformer is connected to K_3. The 12 V input, after rectification and smoothing, results in a symmetrical supply of ±16 V, which is used to power output amplifiers IC_{10} and IC_{11}.

From the ±16 V lines, voltage regulators IC_{12} and IC_{13} derive a supply of ±8 V, which is used to power the remainder of the circuit.

The 12 V line for the relay is taken directly from K_3 and rectified by D_8-D_9.

Construction

The processor is best constructed on the PCB illustrated in Fig. 6. Since this board is double sided and through-plated, it is not possible to make it without special tools and equipment.

Populating the board is straight-

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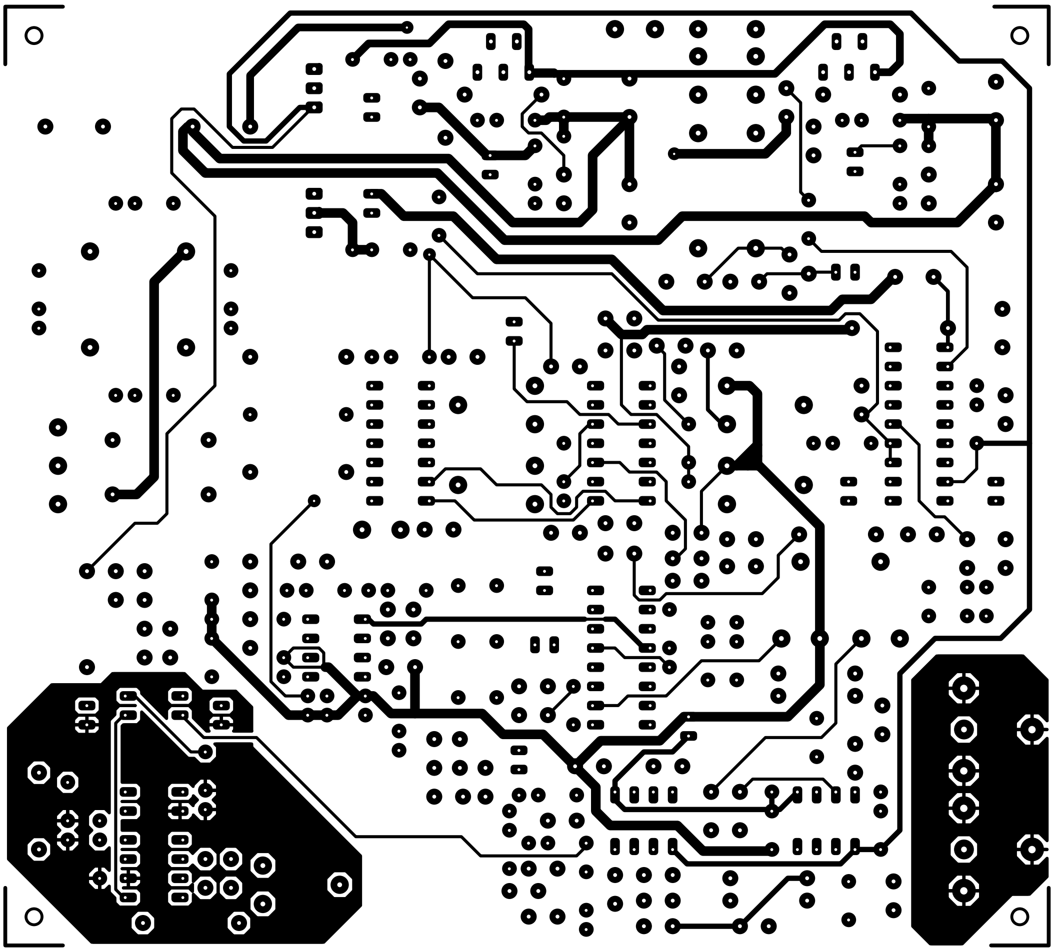


forward and should not present undue difficulties. Note that when the specified enclosure is used, capacitors C_{65} and C_{66} must be not higher than 38–40 mm.

Use gold-plated connectors for K_1 and K_2 to ensure good, lasting connections. The positions for these connectors are at the edge of the board so that all that is necessary when fitting the board into the enclosure is drilling a few holes in the back panel through which these sockets can protrude. Note that they must not touch the enclosure.

Output amplifiers IC_{10} and IC_{11} are located at the edge of the board so that they can be fitted readily to a heat sink. The ICs must be electrically isolated from the heat sink by ceramic washers and heat conducting paste.

The photograph in Fig. 3 shows the completed prototype board.

The main requirements of the enclosure are that it is made of metal and that it provides adequate space for the finished board.

Apart from K_1 and K_2, fit suitable sockets or spring-loaded terminals at the back of the enclosure for connecting the centre and surround loudspeakers. Link these terminals with heavy-duty insulated wire to points 'C' and 'S' and the adjacent earthing points on the board. The specified mains entry with integral fuse holder should also be fitted at the back of the enclosure.

The mains on/off switch and D_6 should be fitted at the front panel (for which a ready-made foil is not available). The diode should be connected to the relevant points on the board via lengths of flexible stranded wire.

Finally, link the centre pin of K_3 to the mass of the enclosure with the aid of a solder tag.

The completed prototype is shown in the photograph of Fig. 4.

The output amplifiers are suitable for operation with load impedances ≥ 4 Ω, but not lower ones. Thus, for the surround channel, two 8 Ω loudspeakers may be connected in parallel only if it is absolutely certain that the impedance is 8 Ω. If it is not, connect the speakers in series: this is safer. It is essential that the two speakers are in phase: the + terminals must go to the same terminal on the board: whether this is earth or 'S' does not matter.

The centre loudspeaker must be in phase with the main speakers. Since the relevant VCA functions as an inverter, the + terminal of this speaker must be connected to the earth point on the board; the –terminal to point 'C'.

Calibration

Start by setting presets P_2–P_7 to the centre of their travel, and P_1 to maximum (fully clockwise).

Inject a (mono) speech signal and adjust P_6 until the sound appears to come from the centre speaker. Wait for a surround signal (indicated by the red segment of D_6 lighting) and turn P_7 till sound emanates from the surround speakers. Do not set the level too high, because this leads quickly to an exaggerated effect. If, however, it is felt that the desired level can not be obtained with P_6 or P_7, as the case may be, adjust P_4 or P_5, or both, as required.

Next, create a spatial effect by slowly turning P_2 and P_3 until the centre channel and the

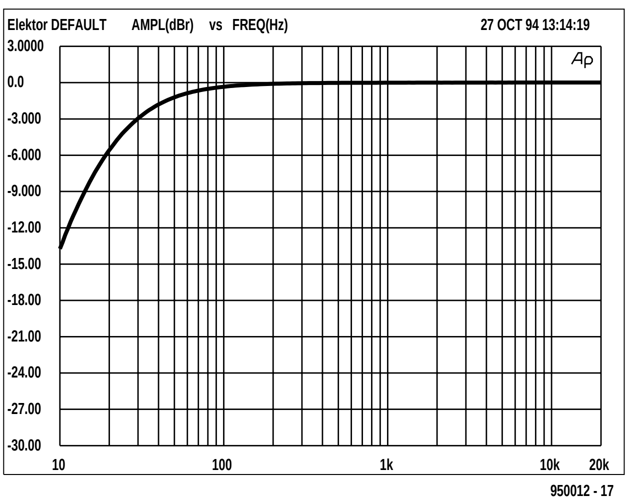


Fig. 7. Frequency response of the centre channel.

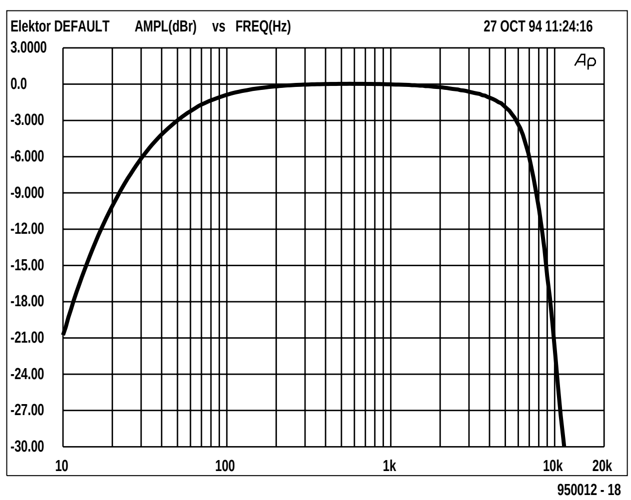


Fig. 8. Frequency response of the surround channel.

surround channel seem well 'separated'. It is more than likely that P_6 and/or P_7 must then be readjusted. Note that these controls give an instinctive 'wrong' feel: turning them clockwise *reduces* the level.

If the surround speakers give exaggerated reverberation, reduce the delay with P_1. In the average living room, a delay of 25 ms appears correct: this corresponds to P_1 being almost at its maximum setting.

Do not be surprised if after watching and listening to a number of films, some readjustment of the controls are found desirable.

Characteristics

The curves in Fig. 7 and 8 show the amplitude vs frequency characteristics of the centre channel and surround channel respectively. The curves were obtained with an audio analyser. It is evident that they correspond closely with the descriptions. The –3 dB point of the centre channel is at 30 Hz. The –3 dB points of the surround channel are at 50 Hz and 6 kHz; the –6 dB bandwidth is roughly 30 Hz to 7 kHz.

Figure 8 shows that it does not make sense to use tweeters with a linear characteristic up to 20 kHz for the surround channels.

Both curves make it clear that a subwoofer must be connected to the main channels and not to the centre channel or surround channel.

Parts list

Resistors:
R_1, R_2, R_{27}, R_{28}, R_{45}, R_{50}, R_{53}–R_{56} = 47 kΩ
R_3, R_4 = 1 kΩ (see text)
R_5–R_8, R_{11}, R_{23} = 10.0 kΩ, 1%
R_9, R_{21} = 11.8 kΩ, 1%
R_{10}, R_{22} = 11.0 kΩ, 1%
R_{12}, R_{24} = 6.98 kΩ, 1%
R_{13}, R_{25} = 4.32 kΩ, 1%
R_{14}, R_{26} = 5.62 kΩ, 1%
R_{15} = 18 kΩ
R_{16} = 22 kΩ
R_{17}, R_{18}, R_{36} = 100 kΩ
R_{19}, R_{20} = 5.6 kΩ
R_{29}, R_{31} = 2.2 kΩ
R_{30}, R_{32} = 6.8 kΩ
R_{33}, R_{34}, R_{37}, R_{42}, R_{57}, R_{59}, R_{61}, R_{63} = 10 kΩ
R_{35} = 100 Ω
R_{38} = 2.7 kΩ
R_{39} = 270 kΩ
R_{40}, R_{41} = 560 kΩ
R_{43} = 15 kΩ
R_{44} = 12 kΩ
R_{46}, R_{48} = 39 kΩ
R_{47}, R_{49}, R_{51}, R_{52} = 1 kΩ
R_{58}, R_{62} = 560 Ω
R_{60}, R_{64} = 4.7 Ω
R_{65} = 5.6 MΩ
R_{66} = 10 Ω
P_1 = 50 kΩ preset
P_2–P_7 = 10 kΩ preset

Capacitors:
C_1, C_2, C_{25}, C_{27}, C_{45}, C_{52} = 2.2 µF, polypropylene, pitch 5 mm
C_3, C_5, C_{16}, C_{18} = 2.2 nF
C_4, C_{17} = 3.9 nF
C_6, C_{19} = 1 nF
C_7, C_8, C_{20}, C_{21}, C_{30}–C_{34}, C_{39}, C_{44}, C_{47}, C_{49}, C_{51}, C_{54}, C_{56}, C_{58} = 100 nF
C_9 = 100 pF polystyrene, axial
C_{10}, C_{11}, C_{13}, C_{35}, C_{63}, C_{64}, C_{67}–C_{70} = 100 nF, ceramic
C_{12}, C_{36}, C_{60} = 47 µF, 25 V, radial
C_{14}, C_{26} = 4.7 µF, 63 V, radial
C_{15} = 270 pF
C_{22} = 68 pF
C_{23}, C_{59}, C_{61}, C_{62} = 10 µF, 63 V, radial
C_{24} = 1 µF, polypropylene, pitch 5 mm
C_{28} = 1 µF, 63 V, radial
C_{29} = 220 µF, 10 V, radial
C_{37}, C_{38} = 47 µF, 16 V, radial
C_{40}, C_{41} = 100 µF, 25 V, radial
C_{42} = 150 nF
C_{43} = 68 nF
C_{46}, C_{53} = 22 µF, 40 V, radial
C_{48}, C_{50}, C_{55}, C_{57} = 220 µF, 25 V, radial
C_{65}, C_{66} = 4700 µF, 25 V, radial
C_{71} = 820 nF

Inductors:
L_1, L_2 = 100 µH
L_3 = 150 µH

Semiconductors:
D_1–D_4 = zener diode, 4.7 V
D_5 = 1N4148
D_6 = bi-colour LED (green/red)
D_8, D_9 = 1N4004
B_1 = SB102, 10 A, 100 V, for PCB mounting
T_1 = BC517

Integrated circuits:
IC_1, IC_2 = NE5532
IC_3 = MN3101
IC_4 = MN3008
IC_5 = TL072
IC_6 = NE571
IC_7 = TL084
IC_8 = 4030
IC_9 = TDA1074A
IC_{10}, IC_{11} = TDA2040
IC_{12} = 7808
IC_{13} = 7908

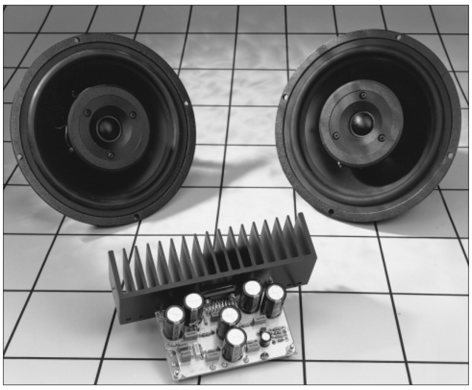
Miscellaneous:
K_1, K_2 = audio socket for PCB mounting
K_3 = 3-way terminal block, pitch 5 mm
Re_1 = relay 12 V, 5 A, 270 Ω
Heat sink SK57, 37.5 mm high*
Ceramic washers Type AOS220*
Enclosure 300×45×210 mm (W×H×D) ($11^7/_8$×$1^3/_4$×$8^1/_4$ in)
Mains transformer with 2×12 V, 80 VA secondary
Mains entry with integral fuse holder and 500 mA slow fuse
Loudspeaker terminals (spring loaded) or suitable audio sockets
Mains on/off switch
PCB Order No. 950012-1

[950012]

* Available from Dau (UK) Ltd, 70–75 Barnham Road, Barnham PO22 0ES, telephone (01243) 553 031

Contents

30 W AF AMPLIFIER FOR CARS



Design by T. Giesberts

The power that can be obtained from a standard car radio amplifier operating from a 12 V car battery is 5–6 W, which (for many listeners) is not really enough for satisfactory hi-fi reproduction. It is, of course, possible to boost the 12 V supply with a power inverter, but that is fairly expensive and not always acceptable. Now, a Philips IC enables audio power of about 30 W to be obtained from a 12 V car battery.

Until not so long ago, the Class B output stages of a standard car radio could not deliver more than 2× 5–6 W_{rms} into 4 Ω loudspeakers. More was not possible with a single supply line of 12 V. Most modern car radios use bridge amplifiers to boost the output to 12–16 W. Often, each of the four loudspeakers has its own dedicated amplifier. Many car manufacturers do not like the use of power inverters to raise the on-board voltage out of fear that these can cause (embarrassing or even dangerous) interference with the remainder of the electronic systems in the car (of which there can be many). There is also the problem of heat generation in the output stages, which may necessitate forced cooling.

Class H

Electronics manufacturers have been researching ways and means of obtaining adequate output power without the use of a power inverter, and Philips have come up with the TDA1560Q.

Output amplifiers can be arranged in a number of different configurations, of which most audio enthusiasts only know Class A and Class B. A different one that provides fairly high output power with relatively low dissipation is Class G. In this configuration, use is made of two supply voltages: a fairly low one that is constantly available and a much higher one that becomes available only when the the voltage swing of the output stages can not be sustained by the low supply voltage.

Since in cars only one supply voltage is available, Philips engineers have devised a pseudo Class G configuration in which a number of electrolytic capacitors are charged by the battery voltage. During brief voltage peaks in the output signal, semiconductor switches connect these capacitors in series with the 12 V line so that the supply voltage to the amplifier is temporarily doubled. Since this is a further development of the Class G technique, it is named Class H. The (temporary) 24 V supply to the amplifier enables (theoretically) a power of 80 W to be delivered into 4 Ω or 40 W into 8 Ω.

A simplified diagram of a Class H output amplifier is shown in Fig. 1. It contains two principal circuits: the first is a Class B amplifier, T_1–T_4, which is loaded by R_1, and the second raises the internal

Brief Technical Data

Class H operation	
Low dissipation with music signals	
Extensive protection circuits (output current; temperature; load impedance)	
Supply voltage	12 V nominal
Quiescent current	100 mA
Output power	
(1 kHz sinusoidal, THD = 0.5%)	30 W r.m.s. into 8 Ω
(music signal)	40 W into 8 Ω
THD + noise (1 W into 8 Ω)	<<0.01% (1 kHz)
	<<0.05% (20 Hz to 20 kHz)
THD + noise (20 W into 8 Ω)	<<0.06% (1 kHz)
	<<0.2% (20 Hz to 20 kHz)
Power bandwidth (–3 dB)	5 Hz to 100 kHz

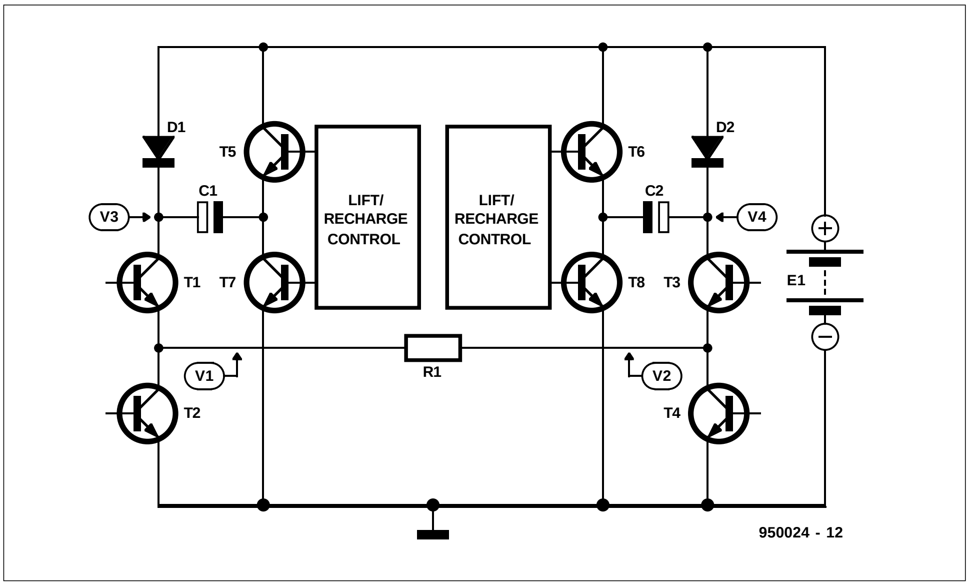


Fig. 1. Diagram of a basic Class H amplifier.

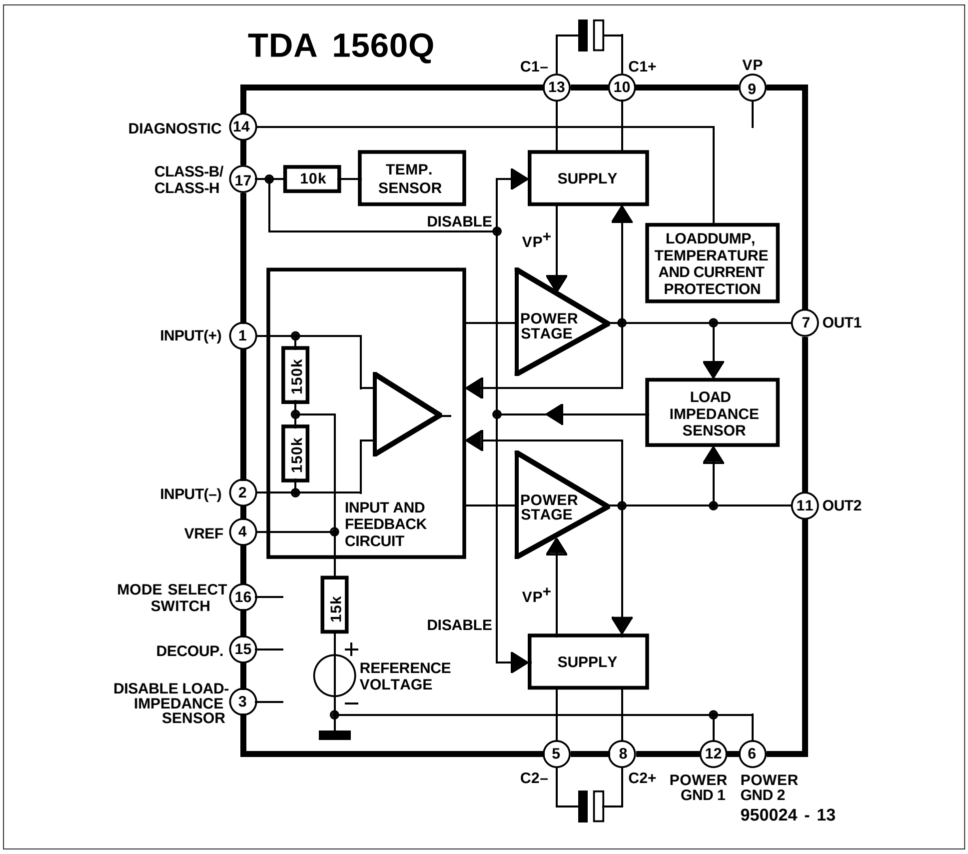


Fig. 2. Block diagram of the TDA1560Q

supply voltage. The second circuit uses two external capacitors, C_1 and C_2, which serve as supply buffers.

Since a music signal consists only for a small part of high level components, the supply voltage needs to be raised for a small part of the time only.

Because the supply voltage is raised for short periods of time only, the average dissipated power will be only slightly higher than that of an amplifier without a voltage-raising circuit, in spite of the fact that the peak output power is appreciably higher.

Capacitors C_1 and C_2 are charged through current sources T_7 and T_8 to a voltage which is nearly equal to the supply voltage, E_1. When either voltage V_1 (or V_2) rises and T_1 (or T_3) approaches the saturation voltage, the lift control circuit detects this. Lift transistors T_5 and T_6 then conduct, so that the charged capacitors are switched between the collector of T_1 (or T_3) and supply voltage E_1. Diodes D_1 and D_2 prevent the capacitors being discharged via the battery. Voltage V_1 (or V_2) can increase to nearly twice the supply voltage. The lift/recharge control circuit ensures that T_5 and T_7, and T_6 and T_8, can not conduct simultaneously.

Inside the TDA1560Q

A block diagram of the TDA1560Q is shown in Fig. 2. A differential input stage in the input and feedback circuit is connected to pins 1 and 2. Because of this stage, the IC is highly insensitive to common-mode interference. The input impedance is 300 kΩ, so that for a good low-frequency response even small input capacitors are sufficient.

The input and feedback circuit contains circuitry that controls the supply circuits and the power stages.

The control circuitry monitors the input signal and anticipates saturation of the output transistors. As soon as this happens, the supply voltage is raised. Because the input signal is monitored, it is possible to control the lift voltage. To keep the dissipation at a minimum, the supply voltage is raised only to a level where the output will stay below the clipping level.

A current limiter protects the output stages against being short-circuited to ground or to the supply line. In general, whenever the current drawn exceeds a level of 5.5 A, the output and the power stages are switched off. The protection circuit monitors at short intervals whether the short-circuit has been removed. If so, the output stages are reactuated. This arrangement limits the dissipation in the power stages during short-circuits to a minimum.

There is a dual temperature protection. The first switches off the voltage doublers when the temperature of their cases rises above 120 °C. The amplifier can then operate in Class B only. The second protection uses sensors located close to the output and switching transistors. If these sensors measure a temperature higher than 165 °C, the base current of the associated transistor is lowered.

There is also a circuit that monitors the load impedance. After the amplifier has been switched on, the d.c. resistance of the loudspeaker(s) is determined by passing a current through the speaker coil(s) and measuring the consequent voltage drop across it.. Because of the peak current that the power stages can handle, the Class H section is switched off when a 4 Ω loudspeaker is detected. The IC then operates as a Class B amplifier. When the load impedance drops below 0.5 Ω, it is considered a short circuit and the entire IC is disabled. The impedance sensor is very sensitive and may be actuated by spurious pulses (for instance, when a car door gets closed during switch-on: the loudspeakers then act as microphones). The sensor may be disabled by shorting pin 3 to ground.

Finally, the IC contains an internal reference voltage source for the input circuits. This reference is decoupled by a capacitor at pin 4.

Circuit description

The circuit diagram of the 30 W AF amplifier is shown in Fig. 3. The input is connected to the differential input stage (pins 1 and 2) via coupling coupling capacitors C_1 and C_2. Although the –ve input is linked to earth via wire bridge A-B, the arrangement gives good common-mode rejection in spite of the assymetrical input. A true symmetrical input is obtained when wire bridge A-B is omitted from the board. Network R_1-C_3 forms a low-pass filter that suppresses HF interference at the input. The input impedance is determined largely by the value of R_2.

The circuit around darlington T_1 provides a delay at switch-on to suppress switching noise reaching the output stages. This is essential since the impedance measuring sensor (which normally provides the suppression) has been disabled here (pin 3 to ground).

As soon as the supply voltage is on, a potential of 3 V is applied to pin 16 via R_3 and D_1 to place the IC in the mute mode. Initially, since C_{10} is discharged, T_1 remains off. However, the capacitor then gets charged via R_4 and within a few seconds the voltage across it has risen to a level at which T_1 begins to conduct. This results in a voltage of 12 V being applied to pin 16, whereupon the power stages operate normally.

Various facilities are open to the user via pin 14. For instance, it may be used to detect whether one of the protection circuits is working. Normally, this pin is at the level of the supply voltage. If its potential is only half that value, one of the protection circuits is actuated. When the pin is linked to ground via jumper JP_1, the amplifier is muted. When the jumper is connected to +12 V, all protection circuits are disabled. For normal operation, therefore, no jumper should be placed at JP_1.

Capacitors C_{15} and C_{16} buffer the 12 V battery voltage. Capacitors C_{11}-C_{12} and C_{13}-C_{14} are required by the lift/control circuit. Two parallel-connected 4700 µF electrolytic capacitors are used in each case, because these take up less space on the board than a single 10,000 µF component. Networks R_6-C_6 and R_7-C_7 are interference suppressors.

A Boucherot network, R_8-C_8 and R_9-C_9, is provided at each loudspeaker terminal to maintain a 'normal' load at high frequencies (at which the loudpeaker impedance rises appreciably owing to its inductive behaviour).

Construction

The construction is limited to populating the printed-circuit board shown in Fig. 4. The IC must protrude slightly from the board so that it can be fitted flush against the heat sink. An insulating washer between IC and heat sink is a must: apply heat conducting paste to both the base of the IC and the heat sink. The insulating washer may have to be cut to size from a mica T03 washer.

Most constructors will use two amplifiers in one enclosure for a compact stereo setup. It is, however, also possible to use four amplifiers in one case: one for each of the two front and rear speakers. Make sure that none of the loudspeaker cables or terminals can touch the chassis, because that does not do the bridge amplifier any good in the long term in spite of the protection circuits.

Parts list

Resistors:

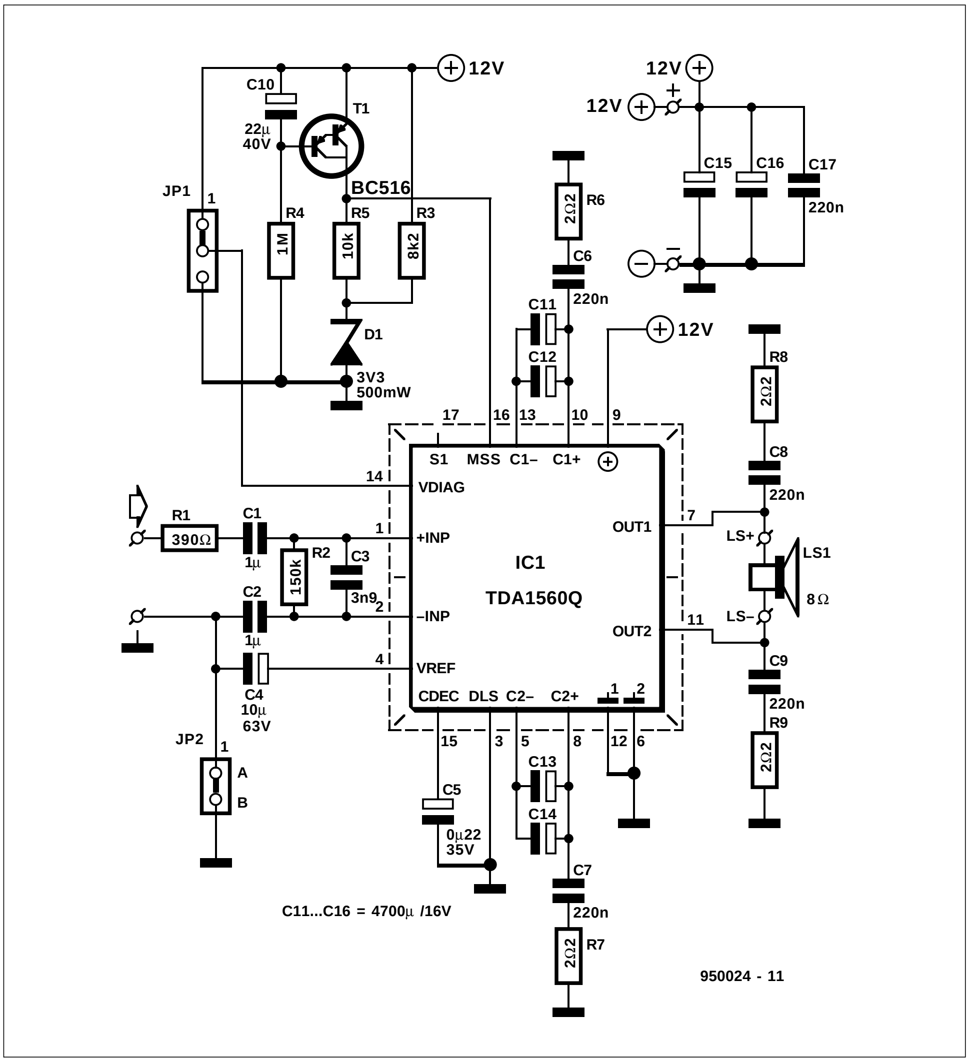


Fig. 3. Circuit diagram of the 30 W AF amplifier for cars.

Contents

R_1 = 390 Ω
R_2 = 150 kΩ
R_3 = 8.2 kΩ
R_4 = 1 MΩ
R_5 = 10 kΩ
R_6–R_9 = 2.2 Ω

Capacitors:
C_1, C_2 = 1 µF, pitch 5 mm
C_3 = 3.9 nF
C_4 = 10 µF, 63 V, radial
C_5 = 220 nF, 35 V, tantalum
C_6–C_9, C_{17} = 220 nF
C_{10} = 22 µF, 40 V, radial
C_{11}–C_{16} = 4700 µF, 16 V, radial

Semiconductors:
D_1 = zener, 3.3 V, 500 mW
T_1 = BC516

Integrated circuits:
IC_1 = TDA1560Q

Miscellaneous:
Heat sink for IC_1; R_{th} << 2.5 K W^{-1}
PCB Order No. 950024-1

[950024]

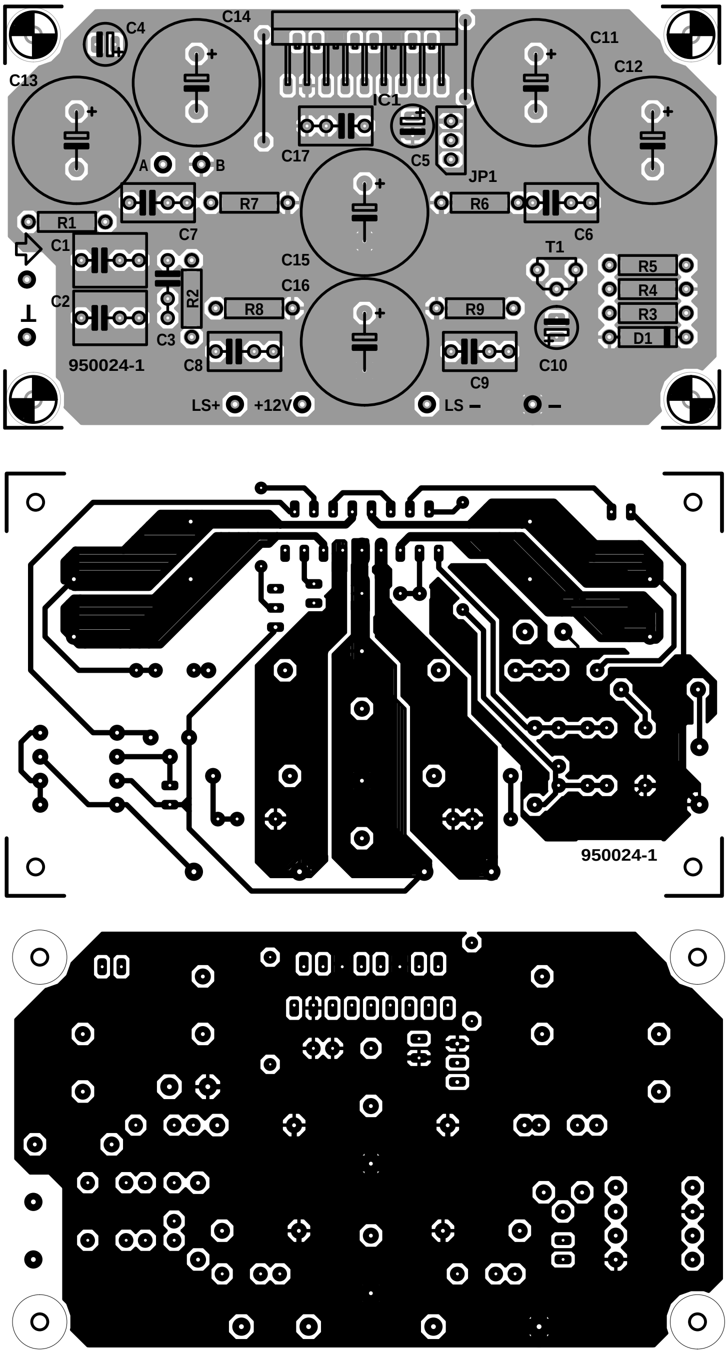


Fig. 4. Printed circuit board for the 30 W AF amplifier for cars.

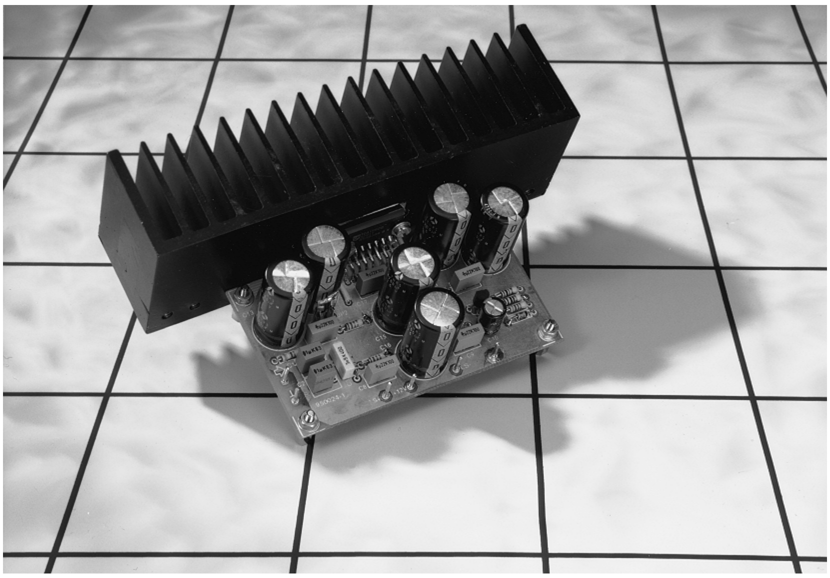


Fig.5. The completed (prototype) 30 W AF amplifier for cars.

HEADPHONE AMPLIFIER

Design by T. Giesberts

On much audio equipment, the headphone output is simply derived from the loudspeaker output via a series resistor: not a very elegant design! The present circuit describes a 'real' headphone amplifier that can be added to most equipment, but may also be used as a stand-alone unit.

On much audio equipment, the headphone output is simply derived from the loudspeaker output via a series resistor: not a very elegant design! The present circuit describes a 'real' headphone amplifier that can be added to most equipment, but may also be used as a stand-alone unit.

Although many audiophiles still believe discrete components are best, the relentless progress of integrated circuits can not be stopped: these devices get better and better. Even top-quality commercial equipment is now loaded with ICs and no one can doubt their quality and reliability. In many modern CD players, preamplifiers and digital-to-analogue converters (DACs) there is hardly a transistor to be found. Only the design of power amplifiers often still relies on discrete devices. The present amplifier is based on an IC: a surface mount device (SMD) Type TDA1308T from Philips Components.

The IC was developed specifically for use as a headphone driver: the enthusiastic claims of the manufacturer as to its qualities appear to be rather less exaggerated than is often the case (see opposite page: Performance). A signal-to-noise ratio of 110 dB and a distortion factor of <0.009% (with a 5 kΩ load) are undeniably good.

The IC can be used to good effect in CD players, DCC players, keyboards, laser disc systems, multimedia amplifiers, and more. It draws a quiescent current of only 3 mA and can work from supplies of 3–7 V. The latter makes it suitable for use in either battery-powered circuits or in standard mains-operated equipment. Its dynamic range is good, its bandwidth is 5.5 MHz and its slew rate is 5 V μs^{-1}.

The (simplified) internal design of the IC is shown in Fig. 1. The differential input stage uses MOSFETs, M_1, M_2, is provided with current mirrors and is powered by a current source, J_1. The input stage is followed by two operational amplifiers, A_1 and A_2, that drive output stages M_3 and M_6, which are also MOSFETs. The advantage of MOSFETs is that the necessary input bias current is very small: typically 10 pA; moreover, the swing of the amplifier with high impedance loads is nearly equal to the supply voltage.

The inverting and non-inverting inputs of the op amps have an excellent common mode suppression that ranges from the negative supply voltage to 1.5 V under the positive supply voltage. The IC can be fed from single as well as bipolar supplies. The closed-loop gain can be set with two external resistors.

The outputs are short-circuit-proof and totally free of switching noise. The hum suppression is 90 dB.

Circuit description

The circuit diagram is given in Fig. 2. Values of components are generally those recommended by the manufacturer. Power is derived from a single, standard mains adaptor, which should output at least 9 V. The adaptor output is smoothed by C_8 and regulated by IC_2. Diode D_1 protects the circuit against wrong polarity.

The input impedance is determined largely by R_2 (R_6). The value of 3.9 kΩ presents no problems to any preamplifier. The amplification factor is set by the ratio R_2:R_3 (R_6:R_7). As is seen, the factor here is unity, so that the name 'amplifier' is, strictly, a misnomer; 'driver' would have been more appropriate. There is no need for amplification, because the usual line level of 1 V (nominal) is more than enough to drive any headphone. However, a standard line output can not provide sufficient current for driving low-impedance headphones. The present amplifier remedies this.

Resistors R_9 and R_{10}, and capacitor C_5, set the IC for operation from half the supply voltage. Capacitor C_6 provides additional decoupling of the amplifier. Since the supply is asymmetrical, input capacitors C_1 and C_2 (C_3, C_4) are essential. Some audiophiles will raise their eyebrows at this, but in this application no adverse effects of these capacitors have been detected. Resistors R_1 and R_4 (R_5, R_8) make sure that these capacitors are charged even when the input and output are open.

Construction

The design has been kept as compact as feasible as can be seen from the drawings of the printed-circuit board in Fig. 4. In spite of the small dimensions, the construction is no more testing than most, at least not as far as the standard components are concerned. Soldering the surface-mount IC into place (not at the component side of the board, but at the track side as is usual with SMDs) is a tedious job. But, with patience and a small soldering iron with a fine tip, even relatively inexperienced constructors should be able to make a good job of it. Lightly tin the pads on the board and the pins of the IC and place the device in position. Take great care with the positioning of the IC: it is so small that mistakes are easily made. The side of the device where pins 1–4 are located is marked by a chamfered edge on the case. If you can not see this properly, use a magnifying glass. The chamfered edge should point in the direction of R_2 and R_3. Since it is very small, pressing

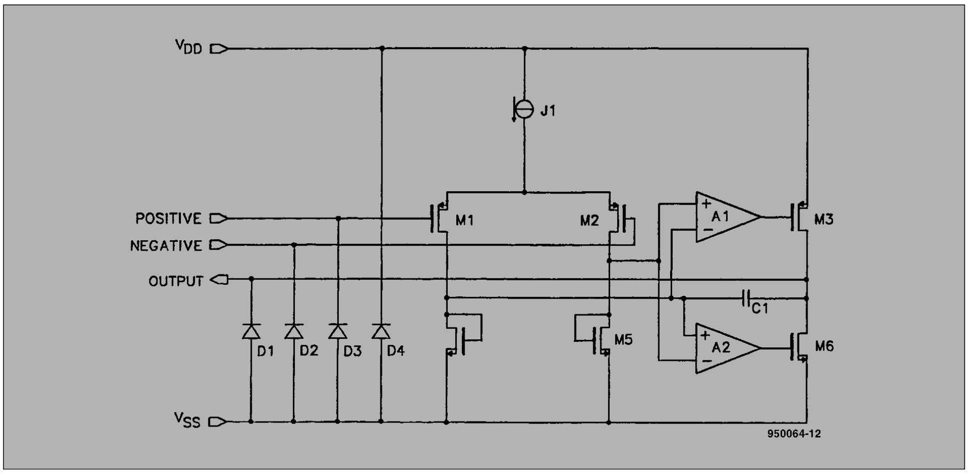


Fig. 1. Diagram of internal circuitry of the TDA1308T.

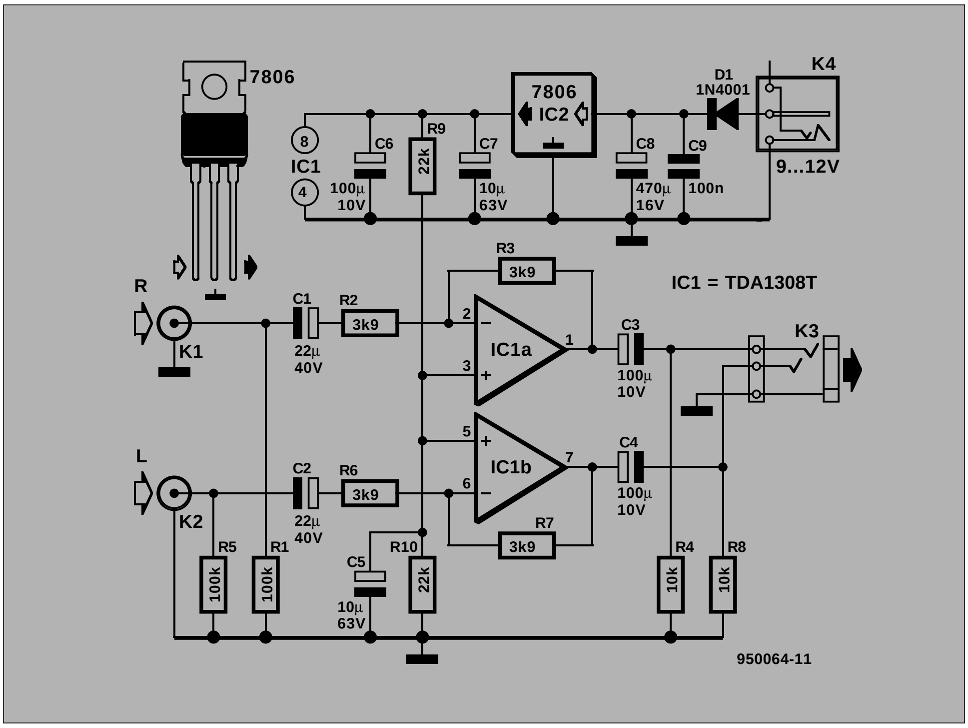


Fig. 2. Circuit diagram of the headphone amplifier.

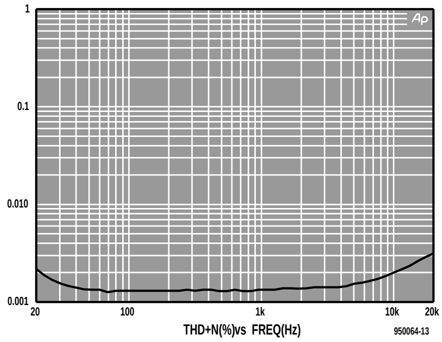


Fig. 3. The THD + N characteristic for 1 V input and a load of 600 Ω.

with a fingernail will do. Gently solder one of the pins into place and make sure that everything is as it should be. If so, carefully solder the other pins on to the board.

The top of the finished board is shown in the photograph in Fig. 5. Note that the various connectors are soldered directly to the board: two phono plugs for inputs K_1 and K_2; an adaptor socket for K_4 and a 6.3 mm stereo jack for K_3. These connectors are, of course, required only if the amplifier is to be used as a stand-alone unit. If the amplifier is built into an equipment, the connectors can all be replaced by soldering pins from where the various connections are made. The signal lines should be in screened cable. Moreover, a mains adaptor will normally not be required, since power can invariably be derived from the main equipment: the amplifier draws only a very small current. If the voltage in he main equipment is too high, it can be dropped by a series resistor and zener diode (9 V or 12 V).

Performance

The performance of the Philips chip is typified by the distortion characteristic in Fig. 3. This shows that the THD + N is, as claimed, low: with a 1 kHz input signal at a level of 1 V and an output load of 600 Ω, the measured value was about 0.0015%. With a load of 32 Ω (Walkman-type headphones), it rose to 0.028%, which is still impressive for such a simple IC.

The channel separation measured at K_3 hovered around 90 dB with a 600 Ω load ad 70 dB with a 32 Ω load (frequency range 20 Hz to 20 kHz). These values depend largely on the internal wiring of the headphones: a common earth wire leads to worse channel separation, but this can not really be attributed to the amplifier.

The maximum output voltage is 2 V r.m.s. across 560 Ω and 1.5 V r.m.s. across 32 Ω.

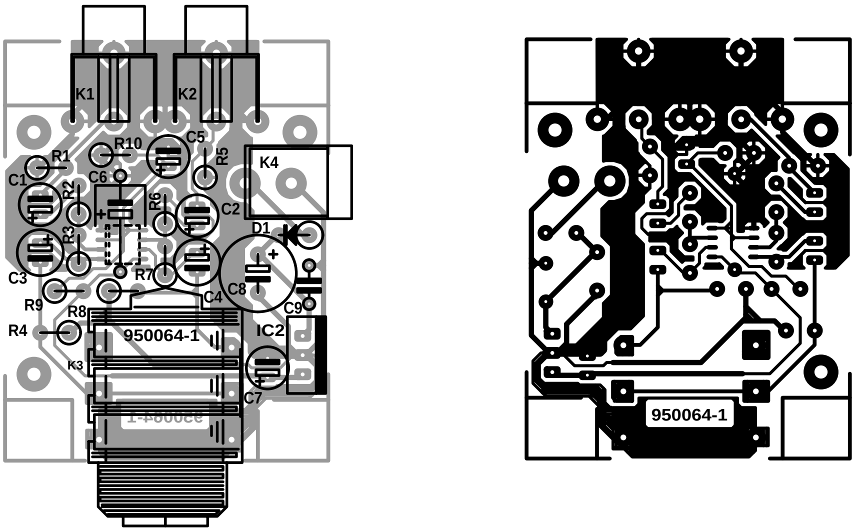


Fig. 4. Printed-circuit board for the headphone amplifier.

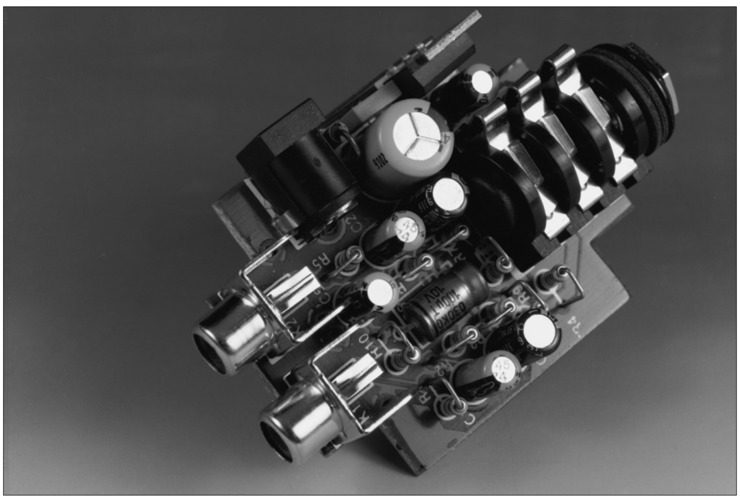


Fig. 5. The completed prototype headphone amplifier board.

Parts list

Resistors:
R_1, R_5 = 100 kΩ
R_2, R_3, R_6, R_7 = 3.9 kΩ
R_4, R_8 = 10 kΩ
R_9, R_{10} = 22 kΩ

Capacitors:
C_1, C_2 = 22 µF, 40 V, radial
C_3, C_4 = 100 µF, 10 V, radial
C_5, C_7 = 10 µF, 63 V, radial
C_6 = 100 µF, 10 V
C_8 = 470 µF, 16 V, radial
C_9 = 100 nF, pitch 5 mm

Semiconductors:
D_1 = 1N4001

Integrated circuits:
IC_1 = TDA1308T (SMD)
IC_2 = 7806

Miscellaneous:
K_1, K_2 = audio socket for PCB mounting
K_3 = 6.3 mm stereo jack for PCB mounting
K_4 = Inlet for mains adaptor (for PCB mounting)
Enclosure (optional): 65×50×30 mm (e.g., Bopla E406 from Phoenix Mecano Ltd, 6–7 Faraday Road, Aylesbury HP19 3RY, Great Britain. Telephone +44 (0)1296 398855)
PCB Order no. 950064 (see p. 70)

[950064]

Contents

HEXFET AMPLIFIER UPGRADE

Design by T. Giesberts

The Medium Power HEXFET amplifier published in this magazine in December 1993 has one small drawback: it delivers 'only' 60 W into 8 Ω (or 120 W into 4 Ω). Otherwise, it is a first class amplifier that provides excellent music reproduction, which is evidenced not so much by measurement as by audition. To some listeners, it has a quality not unlike that of a valve amplifier. Because of its popularity and the many requests for a version with higher output power, it has been upgraded to provide around 90 W into 8 Ω (about 160 W into 4 Ω).

By a stroke of good fortune, a pair of IGBTs (Insulated Gate Bipolar Transistors – see our June 1995 issue) proved ideal replacements for the HEXFETs used in the original design. Apart from the figures for power output, the technical specification remains virtually the same (see box).

Modification

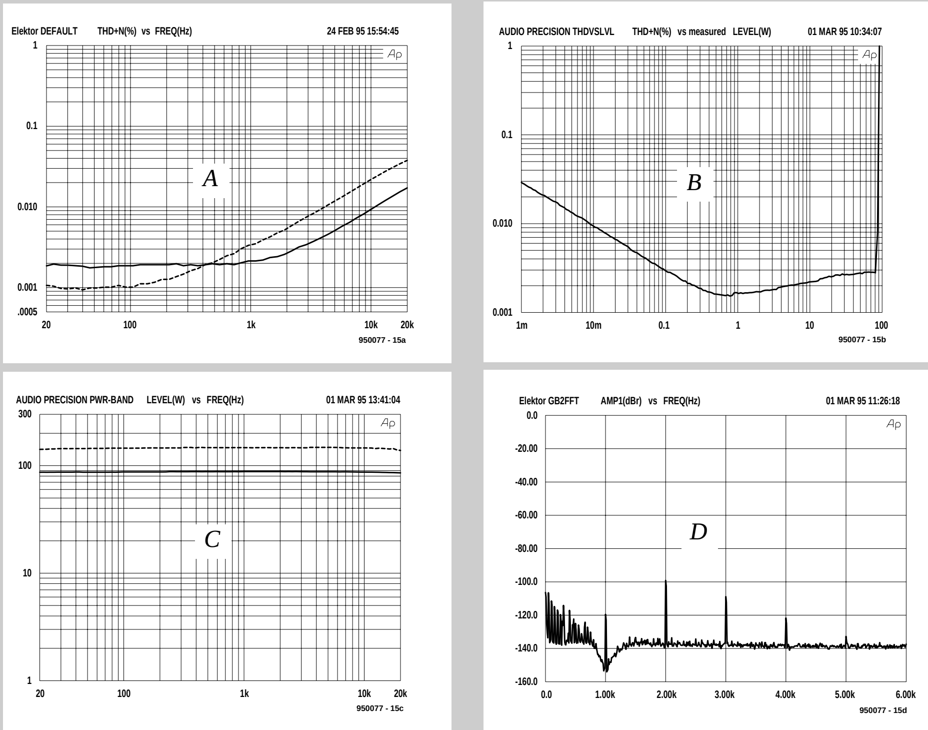
The original design already allowed for a higher-output version, whence the duplicated holes for the output transistors on the printed-circuit board. At that time, advance information on the IGBTs was already available, but samples were not.

Although IGBTs are quite different from HEXFETs, the board for the original design can be used without any modification. In fact, the circuit has hardly changed. The most noticeable alteration is the replacement of the fuses in the source lines of the power FETs by emitter resistors for the IGBTs. The only other changes are in the value of two resistors in the compensating circuit of the input stage, of one in the quiescent-current circuit, and of one resistor and two capacitors in the protection circuit. This means that anyone who has built the original HEXFET amplifier can quickly modify it to the upgraded version.

One item needs to be replaced, however: the mains transformer. After all, more power can not be obtained from the same supply voltage/current. The original transformer with 2×25 V secondaries must be replaced by one that provides 2×30 V at 3.75 A. This will result in a direct voltage of ±43 V.

Circuit description

The circuit diagram of the upgraded amplifier is given in Fig. 1. Changed with respect to the earlier version are T_{12}, T_{13}, RF_1, RF_2, R_3, R_4, R_{21}, R_{35}, C_{13} and C_{14}. Also, to improve performance at high frequencies, a damping resistor has been added to, or rather in, inductor L_1. Finally, to improve the noise figure, the impedance



Brief technical data

Input sensitivity	1.1 V r.m.s.
Input impedance	47.7 kΩ
Power output (1 kHz, 0.1% THD)	88 W into 8 Ω
	146 W into 4 Ω
Music power (1 kHz burst, 5 cycles on, 5 cycles off)	94 W into 8 Ω
	167 W into 4 Ω
Power bandwidth (40 W into 8 Ω)	1.5 Hz – 115 kHz
Slew rate	>35 V μs^{-1}
Signal-to-noise ratio (1 W into 8 Ω)	105 dB (A-weighted)
	101 dB (linear 22 Hz – 22 kHz)
Harmonic distortion (1 W into 8 Ω)	0.002% (1 kHz)
(80 W into 8 Ω)	0.003% (1 kHz)
	<0.05% (20 Hz – 20 kHz)
Intermodulation distortion (50 Hz:7 kHz; 4:1)	0.002% (1 W into 8 Ω)
	0.003% (40 W into 8 Ω)
Dynamic intermodulation distortion (rectangular 3.15 kHz + sine wave 15 kHz)	0.0025% (1 W into 8 Ω)
	0.002% (80 W into 8 Ω)
Damping factor (at 8 Ω)	>600 (1 kHz)
	>400 (20 kHz)

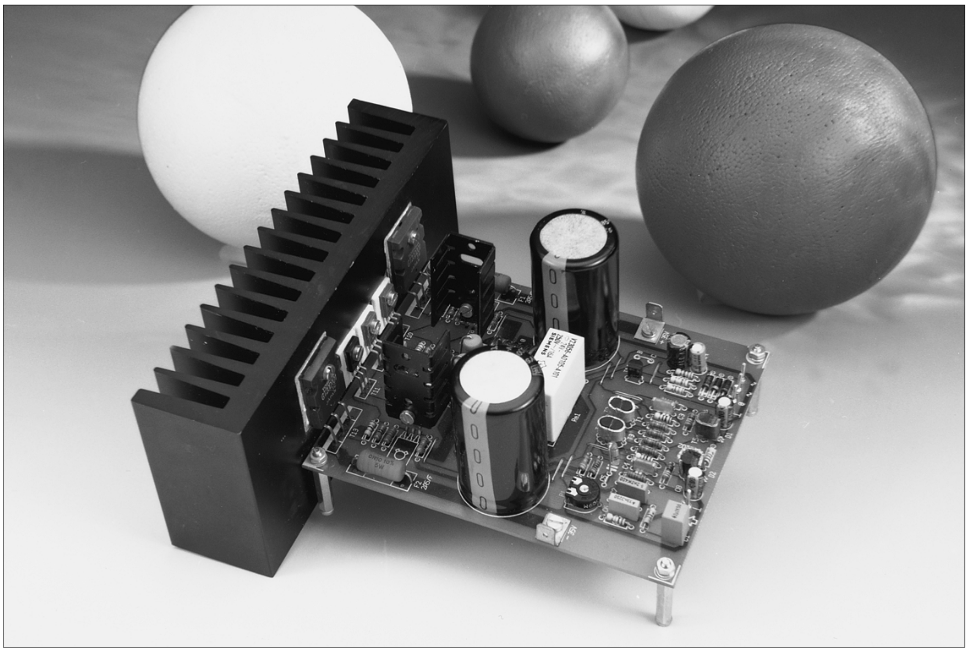
Measurements for the characteristics shown were made with an Audio Precision analyser.

A shows the total harmonic distortion (THD+N) from 20 Hz to 20 kHz. The solid curve refers to 1 W into 8 Ω and the dashed one to 75 W into 8 Ω.

B shows the distortion at 1 kHz as a function of drive (bandwidth 22 Hz – 22 kHz; load 8 Ω). The sharp bend at the end of the curve is the clipping point.

C shows the maximum power output when the distortion is 0.1%. It shows that the power is independent of frequency, whether the load is 8 Ω (solid curve) or 4 Ω (dashed curve).

D shows a Fourier analysis of a 1 kHz signal (1 W into 8 Ω) with the fundamental suppressed. The 2nd, 3rd and 4th harmonics can be seen, but they are attenuated, respectively, by 100 dB, 110 dB and 120 dB with respect to the fundamental frequency.



of input filter R_1-C_2 has been lowered.

A symmetrical design has the advantage that it minimizes problems with distortion, particularly that associated with even harmonics. Therefore, the input stages consist of two differential amplifiers, T_1-T_2 and T_3-T_4. These use discrete transistors, not expensive dual devices, to keep the cost down. Performance is excellent, particularly if the transistors are matched.

A differential amplifier is one of the best means of combining two electrical signals: here, the input signal and the feedback signal. The amplification of the stage is determined mainly by the ratio of the collector and emitter resistances (in the case of T_1-T_2 these are R_9, R_{10}, R_{11} and R_{12}). These provide a form of local feedback: limiting the amplification reduces the distortion.

Two *RC* networks (R_3-C_3 and R_4-C_4) limit the bandwidth of the differential amplifiers and these determine, to a degree, the open-loop bandwidth of the entire amplifier.

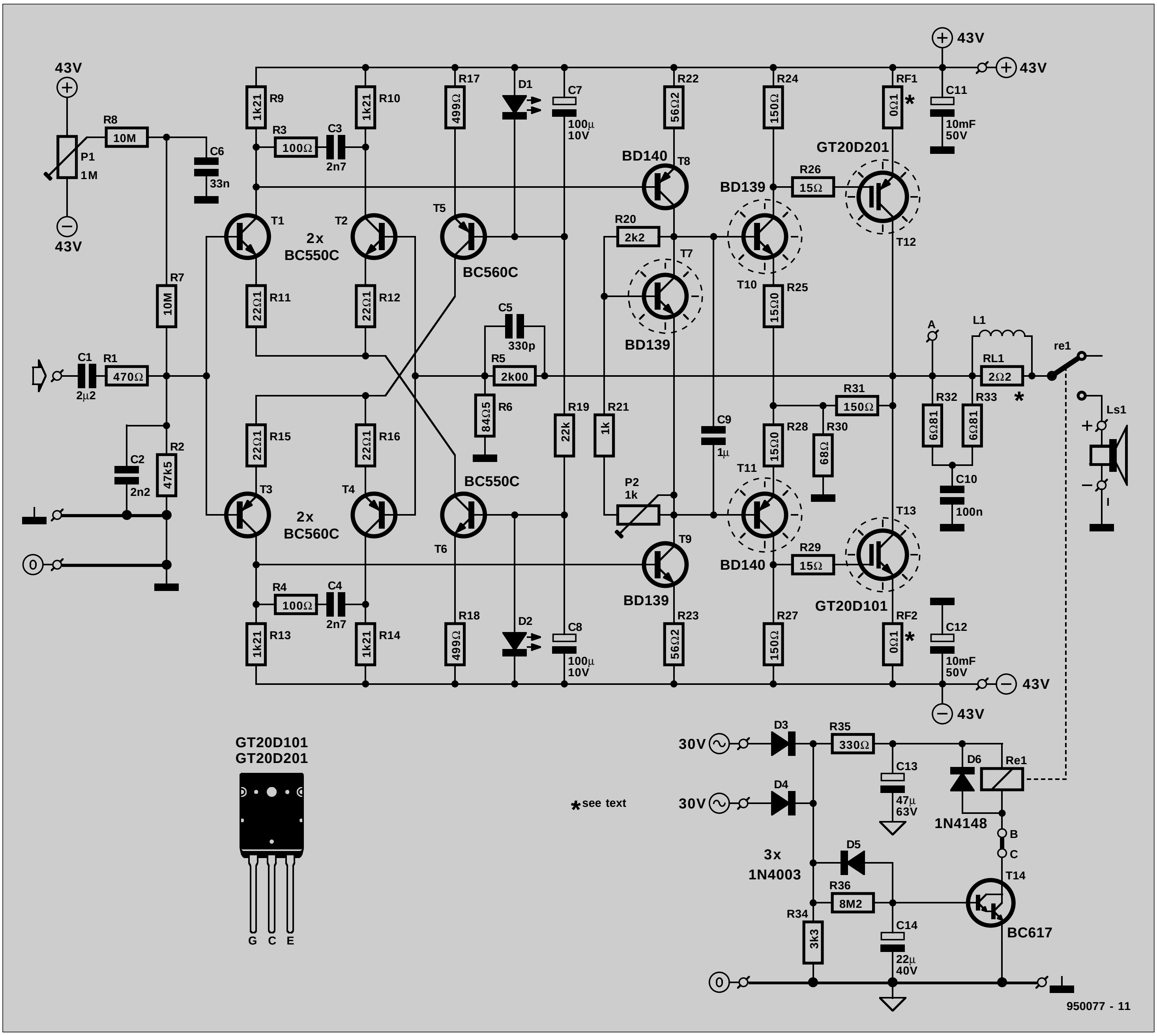


Fig. 1. Circuit diagram of the upgraded (IGBT) amplifier.

Contents

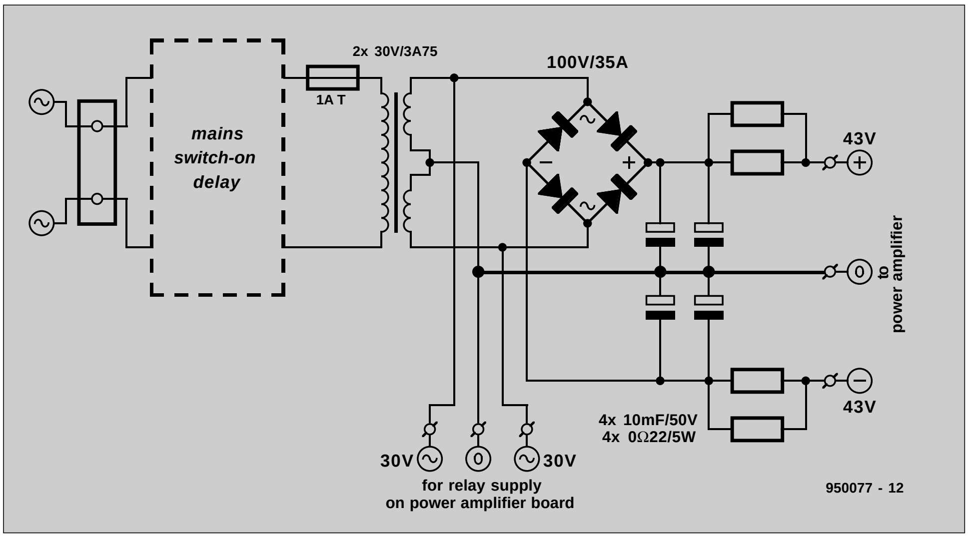


Fig. 2. Circuit diagram of the power supply for one mono IGBT amplifier.

The d.c. operating point of the differential amplifiers is provided by two current sources. Transistor T_6, in conjunction with R_{18} and D_2, provides a constant current of about 2 mA for T_1-T_2. Transistor T_5, with R_{17} and D_1, provides a similar current for T_3-T_4. The combination of a transistor and an LED creates a current source that is largely independent of temperature, since the temperature coefficients of the LED and the transistor are virtually the same. It is, however, necessary that these two components are thermally coupled (or nearly so) and they are, therefore, located side by side on the printed-circuit board.

In the input stage, C_1 is followed by a low-pass section, R_1-C_2, which limits the bandwidth of the input to a value that the amplifier can handle. Resistor R_2 is the base resistor of T_1 and T_3. So far, this is all pretty normal. Network P_1-R_7-R_8 is somewhat out of the ordinary, however. It forms an offset control to adjust the direct voltage at the output of the amplifier to zero. Such a control is normally found *after* the input stage. The advantage of putting it before that stage is that the inputs of the differential amplifiers are exactly at earth potential, which means that the noise contribution of their base resistors is negligible.

The signals at the collectors of T_1 and T_3 are fed to pre-drivers T_8 and T_9. Between these transistors is a 'variable zener' formed by T_7 which, in conjunction with P_2, serves to set the quiescent current of the output transistors.

The output of the pre-drivers is applied to T_{10} and T_{11}, which drive IGBTs T_{12} and T_{13}. This power section has local feedback (R_{30}-R_{31}).

The design of T_{10}–T_{13} is a kind of compound output stage, since the collector of the power transistors is connected to the output terminal. The voltage amplification is limited to ×3 by the local feedback resistors (R_{30}-R_{31}). Here again, this feedback serves to reduce the distortion. The overall feedback of the amplifier is provided by R_5-R-C_5.

Electrolytic capacitors C_{11} and C_{12} (10,000 µF each and part of the power supply) are located close to the IGBTs, so that the heavy currents have only a short path to follow.

At the output is a Boucherot network, R_{32}-R_{33}-C_{10}, that ensures an adequate load on the amplifier at high frequencies, since the impedance of the loudspeaker, because of its inductive character, is fairly high at high frequencies.

Inductor L_1 limits any current peaks that may arise with capacitive loads.

The signal is finally applied to the loudspeaker, LS_1, via relay contact Re_1. The relay is not energized for a few seconds after the power is switched on to obviate any plops from the loudspeaker. Such plops are caused by brief variations in the direct supply voltage arising in the short period that the amplifier needs to reach its correct operating level.

The supply voltage for the relay is derived directly from the mains transformer via D_3 and D_4. This has the advantage that the relay is deactuated, by virtue of the low value of C_{13}, immediately the supply voltage fails. The delay in energizing the relay is provided by T_{14} in conjunction with R_{36} and C_{14}. It takes a few seconds before the potential across C_{14} has risen to a value at which T_{14} switches on. This darlington transistor requires a base voltage of not less than 1.2 V before it can conduct.

The power supply—see Fig. 2—is traditional, apart from the resistors, R_5–R_8, in the power lines. These limit, to some degree, the very large peak charging currents drawn by electrolytic capacitors C_{11} and C_{12}. Moreover, together with these capacitors, they form a filter that prevents most spurious voltages from reaching the amplifier. Measure-ments on the prototype showed that this was particularly evident at frequencies below 500 Hz.

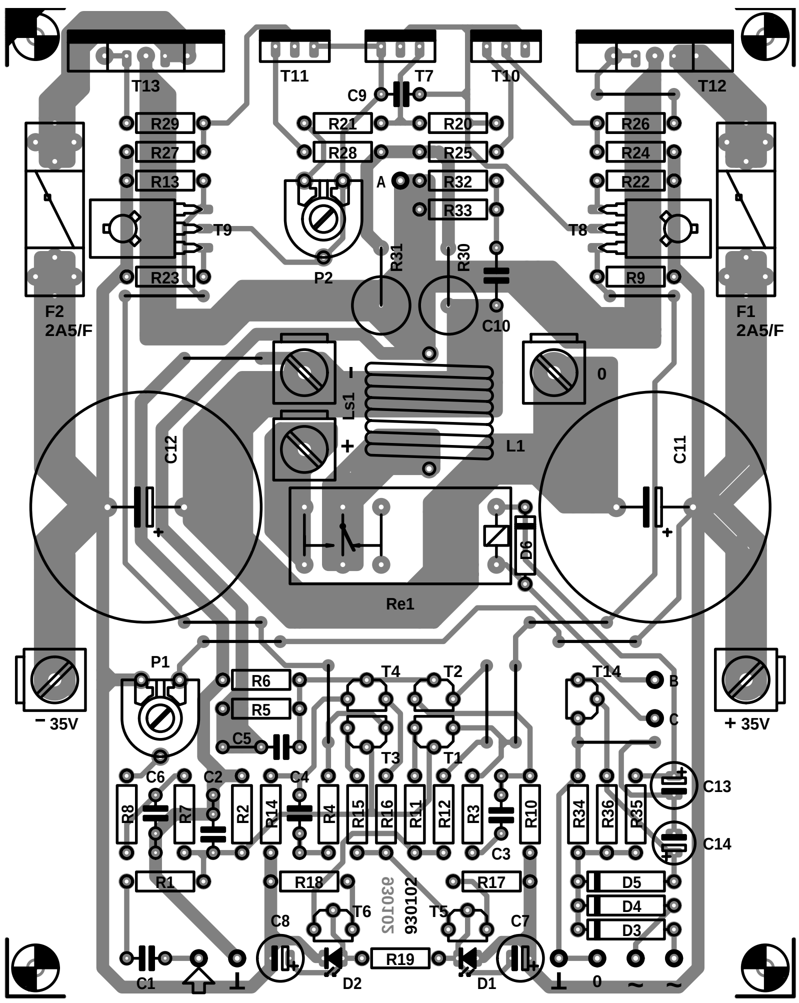


Fig. 3. Printed-circuit board

Construction

The design of the printed-circuit board for the amplifier (Fig. 3) takes good account of the large currents that flow in the amplifier. This has given rise to a couple of tracks being paralleled instead of combined, so that the effect of currents in the power section on the input stages is minimal.

Populating the board is straightforward. Although not strictly necessary, it is advisable to match the transistors used in the differential amplifiers. This may be done conveniently on an h_{fe} tester by measuring the amplification at a collector current of about 1 mA. If such a tester is not available, use a base resistor that results in a collector current of about 1 mA measured with a multimeter. With the same resistor, test a number of other transistors and note the collector currents. Mount the selected pairs on the board and pack them closely together with a 5 mm wide copper ring (made from a piece of 12 mm copper water pipe).

Inductor L_1 consists of six turns, inner diameter 16 mm ($^5/_8$ in), of insulated copper wire 1.5 mm ($^1/_{16}$ in) thick. Mount RL_1 inside the coil.

The large transistors are located on one side of the board, so that they can be fixed directly to the heat sink. They must be insulated from the heat sink with the aid of ceramic washers.

The two sizes indicated on the board for T_{12} and T_{13} were explained earlier.

Connections from the power supply and to the loudspeaker are by means of terminal blocks that can be screwed on to the board.

Mount the two amplifier boards, mains transformers and electrolytic capacitors in a suitable enclosure. The wiring diagram for one channel is given in Fig. 5.

It is advisable to measure the supply voltages before they are applied to the amplifiers. Also, turn P_2 to maximum (wiper towards R_{33}) before connecting the power supply to the amplifiers. Set input presets P_1 to the centre of their travel. A few seconds after the supply has been switched on, the relay should come on. Connect a multimeter (1 V direct voltage range) and adjust P_1 until the meter reads zero (both channels!).

Switch the supply off again and connect a multimeter (100 mV d.c. range) across RF_1 or RF_2. Switch on the supply and adjust P_2 for a meter reading of about 10 mV: this corresponds to a quiescent current of 100 mA through T_{12} and T_{13}. After about half an hour, the current will have stabilized at about 200 mA (meter reading of about 20 mV). Readjust P_2 slightly if required. Note that owing to the positive temperature coefficient of IGBTs, the quiescent current does not increase but drop with rising dissipation.

Finally, recheck the direct voltages at the outputs of the amplifiers and, if necessary, readjust P_1 slightly.

The loudspeakers must be 4-ohm or 8-ohm types, whose impedance must not drop below 3 Ω. It is not permissible to connect two 4-ohm units in parallel to the amplifier, because that would give problems when large drive signals are applied to the IGBTs.

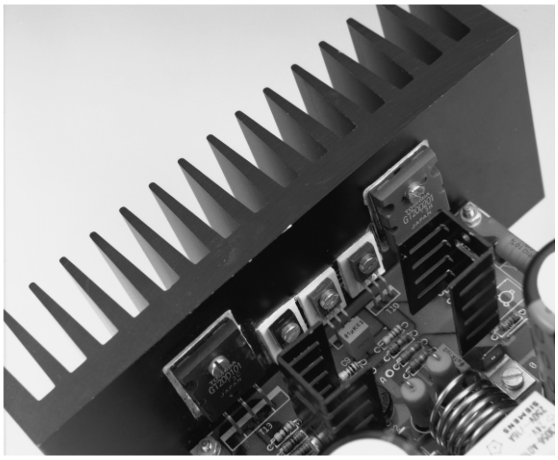
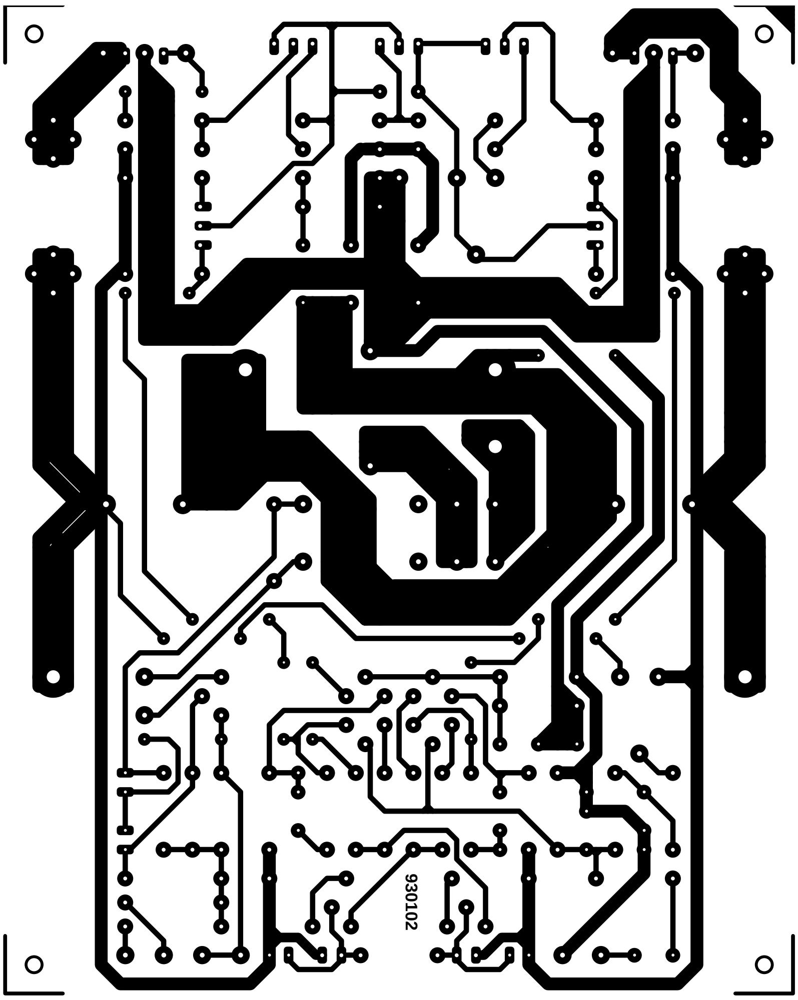


Fig. 4. Good thermal coupling between the transistors and the heat sink ensures a long life of the devices.

Parts list (one channel)

Resistors:
R_1 = 470 Ω
R_2 = 47.5 kΩ, 1%
R_3, R_4 = 100 Ω
R_5 = 2.0 kΩ, 1%
R_6 = 84.5 Ω, 1%
R_7, R_8 = 10 MΩ
R_9, R_{10}, R_{13}, R_{14} = 1.21 kΩ, 1%
R_{11}, R_{12}, R_{15}, R_{16} = 22.1 Ω, 1%
R_{17}, R_{18} = 499 Ω, 1%
R_{19} = 22 kΩ
R_{20} = 2.2 kΩ
R_{21} = 1 kΩ
R_{22}, R_{23} = 56.2 Ω, 1%
R_{24}, R_{27} = 150 Ω, 1%
R_{25}, R_{28} = 15.0 Ω, 1%
R_{26}, R_{29} = 15 Ω
R_{30} = 68 Ω, 5 W
R_{31} = 150 Ω, 5 W
R_{32}, R_{33} = 6.81 Ω, 0.6 W, 1%
R_{34} = 3.3 kΩ
R_{35} = 330 Ω
R_{36} = 8.2 MΩ
RF_1, RF_2 = 0.1 Ω, 5 W
RL_1 = 2.2 Ω, 5 W (fit inside L_1)
P_1 = 1 MΩ preset
P_2 = 1 kΩ preset

Capacitors:
C_1 = 2.2 μF, 50 V, polypropylene
C_2 = 2.2 nF
C_3, C_4 = 2.7 nF



for the IGBT amplifier.

Contents

C_5 = 330 pF, polystyrene, axial
C_6 = 33 nF
C_7, C_8 = 100 µF, 10 V, radial
C_9 = 1 µF, polypropylene, pitch 5 mm
C_{10} = 100 nF
C_{11}, C_{12} = 10,000 µF, 50 V, radial, for PCB mounting
C_{13} = 47 µF, 63 V, radial
C_{14} = 22 µF, 40 V, radial

Inductors:
L_1 = air-core, 0.1 mH (see text)

Semiconductors:
D_1, D_2 = 3 mm LED, red (1.6 V drop at 3 mA)
D_3–D_5 = 1N4003
D_6 = 1N4148
T_1, T_2, T_6 = BC550C
T_3–T_5 = BC560C
T_7, T_9, T_{10} = BD139
T_8, T_{11} = BD140
T_{12} = GT20D201
T_{13} = GT20D101
T_{14} = BC617

Miscellaneous;
Re_1 = relay, 24 V, 1 make contact (e.g., Siemens V23056-A0105-A101*)
F_1, F_2 = fuse, 2.5 A, fast, with holder for PCB mounting
Ceramic washers (5) for T_7, T_{10}, T_{11}
Mica washers (2) for T_{12}, T_{13}
Terminal block (5) (see text)
Heat sink, 0.6 K W^{-1} (e.g., Fischer SK85**)
PCB Order No. 930102 (see p. 70)

Power supply:
Mains transformer, 2 × 30 V, 375 VA
Mains on-off switch with indicator
Fuse 1 A, slow with holder
Bridge rectifier Type B200C35000
Electrolytic capacitor (4), 10,000 µF, 50 V
Resistor (4) 0.22 Ω, 5 W
PCB (for mains-on delay) Order no. 924055 (see p. 70)

* ElectroValue, 3 Central Trading Estate, Staines, TW18 4UX, ☎ (01784) 442 253. Private customers welcome.

** Dau (UK) Ltd, 7075 Barnham Road, Barnham, West Sussex PO22 0ES. Tel. (01243) 553 031. Trade only, but information as to your nearest dealer will be given by telephone.

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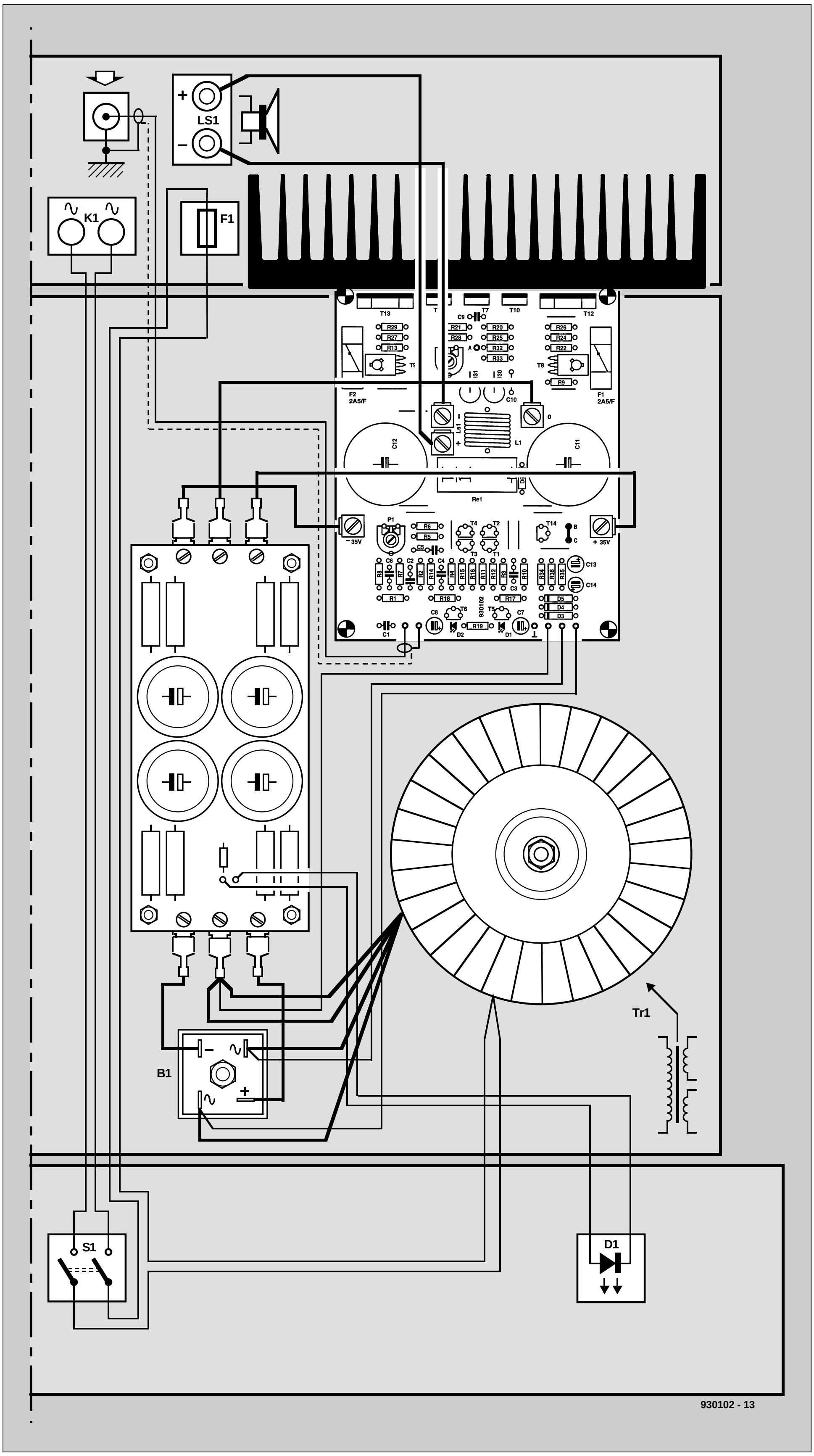


Fig. 5. Wiring diagram of the IGBT amplifier.

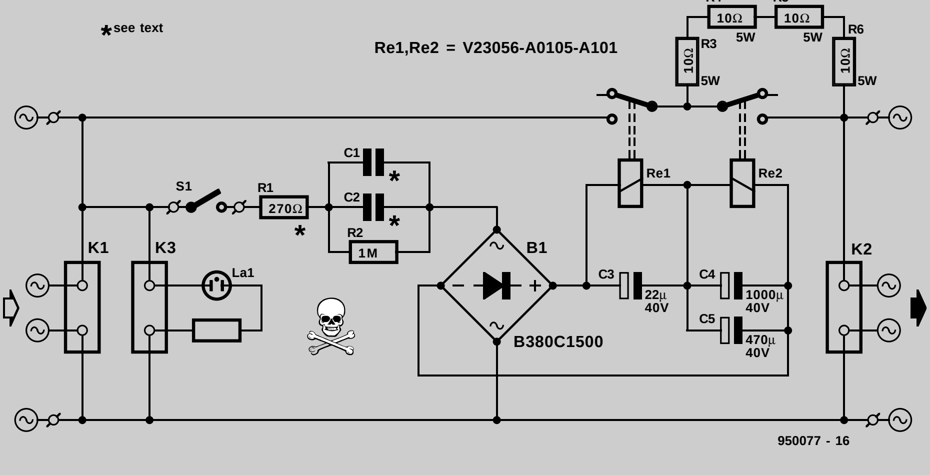
Mains switch-on delay

The circuitry of the 'black box' (in dashed lines) in Fig. 2 is shown above. It may be asked what the function of it is, since there is already a power-on delay in the amplifier itself. That delay serves to obviate plops and clicks caused by switching; it connects the loudspeakers to the amplifier only after this has had a short period of 'settling down'.

The mains switch-on delay is intended to switch on the mains gradually with heavy loads, so that the fuses do not blow

In the circuit, a number of power resistors, R_3–R_6, are connected in series with the mains supply lines to limit the current at switch-on to 5 A. When the mains is switched on with S_1, only relay Re_1 is energized in the first instance, so that the current must flow through the power resistors. When after a few seconds capacitors C_4 and C_5 have been charged, relay Re_2 is also energized, whereupon the power resistors are short-circuited by the relay contacts. These few seconds allow the buffer capacitors in the power supply to be charged at a reasonable rate, so that high currents are prevented.

The relay coils are connected in series and are energized directly by the mains via bridge rectifier B_1, impedances Z_{C1} and Z_{C2}, and R_1. The value of C_1 and C_2 depends on the current required by the relay coils and the level of the mains voltage. The relays specified are rated at 30 mA. If the mains voltage is 240 V, C_2 can be omitted and C_1 should have a value of 470 nF and a rating of 630 V d.c.



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COPYBIT ELIMINATOR REVISITED

WARNING. The information in this article is intended solely for the recording, processing and copying of private musical work. The Editor and Publishers disclaim all responsibility for its use that infringes any copyright vested in commercial compacts discs and (digital) tape cassettes.

Design by H. Schaake

Our February 1994 issue carried an article describing an inexpensive and straightforward circuit for eliminating the copybit from a digital S/PDIF* audio signal to enable users to copy (digitally) their own musical work many times without degradation by the SCMS**. The present article describes an updated version of that circuit, which can be used with the latest DAT, DCC and MD players.

A comparison

The integrated circuits used in digital recorders fulfil more and more functions. Even the until recently discrete S/PDIF buffer/amplifier which converts the S/PDIF signal to TTL level is integrated in modern equipment. This is a good thing, of course, since fewer ICs bring the cost, and thus the price to the consumer, down.

The original eliminator needed a TTL signal at its input, but the updated version contains a separate S/PDIF buffer amplifier. However, the design allows S/PDIF signals already at TTL level to be processed without any difficulty.

The ICs used in modern recorders are faster and run at a higher master-clock frequency than those produced only a few years ago. Nowadays, a $256f_s$ master clock is quite normal, and even $384f_s$ and $512f_s$ models are in production. Since the eliminator needs a $128f_s$ clock, a binary scaler is provided in the updated version. If a $128f_s$ clock is available in the recorder and this can be used, it should be preferred: experience shows that this enables some recorders to lock on more readily.

The present version provides a choice of using either a LOCK or an UNLOCK signal and the facility of inverting the clock if required. These options may be useful for modifying types of DAT, DCC or MD recorder that have not been considered by the designer.

The copybit indication outputs, COPYIN and COPYOU, can not only drive LEDs directly, but may be used, if a copybit is present, as a block (=192 frames) trigger. This allows the S/PDIF signal to be inspected on an oscilloscope at or near the timing-slot position of the copybit frame.

The circuit

The circuit of the original version has been complemented with an IC (by using a MACH210 in the IC_1 position), a couple of resistors, some jumpers, and a capacitor—see Fig. 1. The MACH210 is equivalent to two MACH110s (used in the earlier version), These chips are identical as far as housing is concerned.

The operation of the updated eliminator is identical to that of the earlier version: the reader is therefore referred to the earlier article for background information, block schematic and timing diagram.

Construction

The printed-circuit board in Fig. 3 has been kept small to facilitate its incorporation into a recorder.

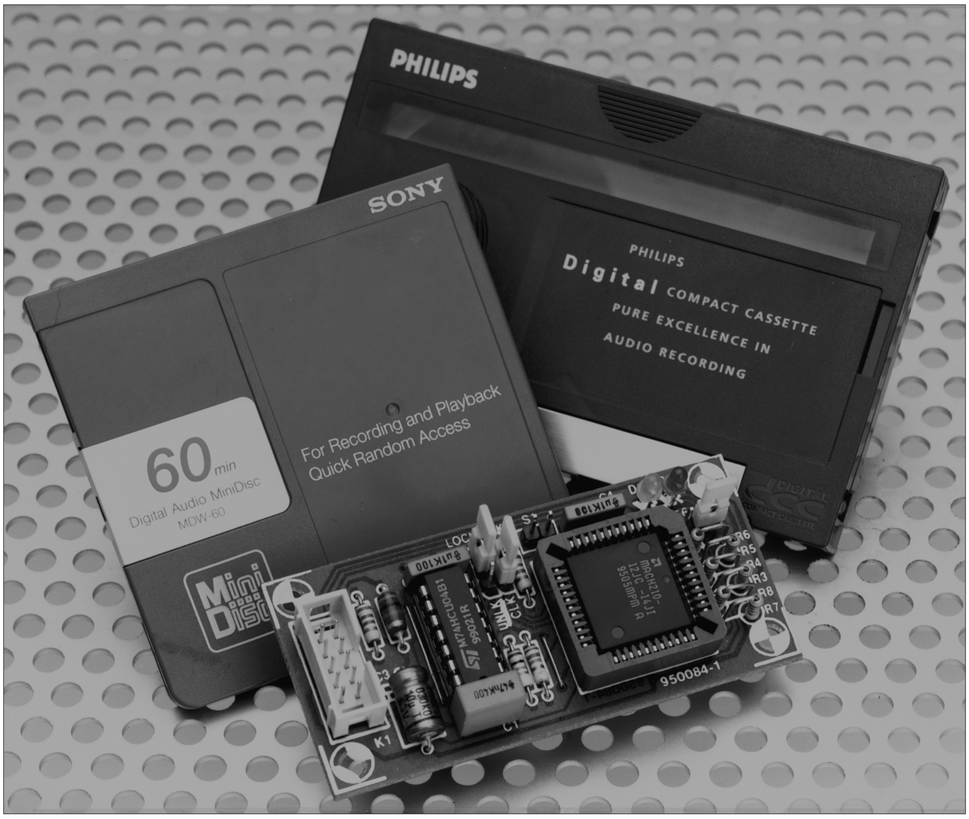
Population of the board is straightforward. The MACH210 is housed in a PLCC case. One of the four corners has been chamfered to indicate how the chip should be located in the relevant IC socket.

Header K_1 is signal-compatible with the earlier version.

A length of flatcable (keep this as short as possible) terminated into a 10-pin connector links the circuit with the appropriate points in the recorder. Switch S_1 enables the eliminator to be switched on and off as required. Thus, if the eliminator is intended to be in circuit at all times, the switch may be omitted.

Building into a recorder

Selecting between CLK and CLK, UNLK and LOCK, and $128f_s$ and $256f_s$ is by means of jumpers. The table on the next page gives connection data for a number of modern recorders; these data may also prove useful with recorders not specified. The selection between CLK and CLK is not given; this must be determined empirically, since



* Sony/Philips Digital Interface Format – the consumer version of the AES/EBU standard. This standard was devised by the American Audio Engineering Society and the European Broadcasting Union to define the signal format, electrical characteristics and connectors to be used for digital interfaces between professional audio products.
** Serial Copy Management System.

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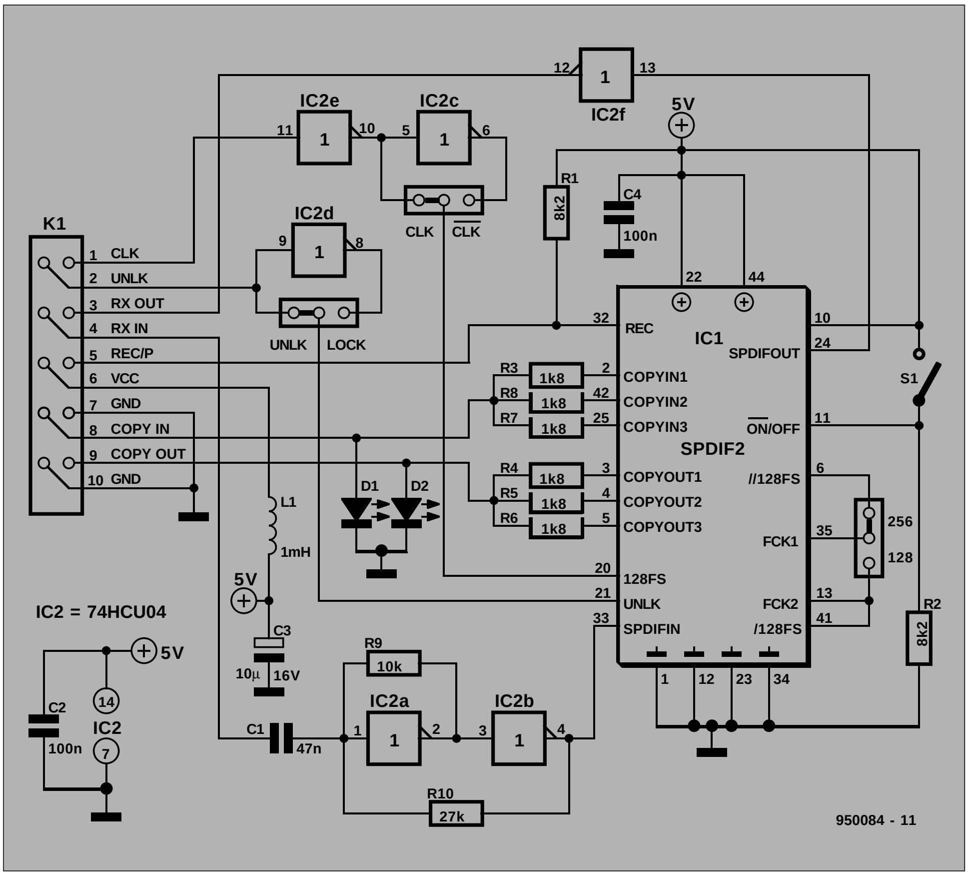

Fig. 1. Circuit diagram of the updated copybit eliminator.

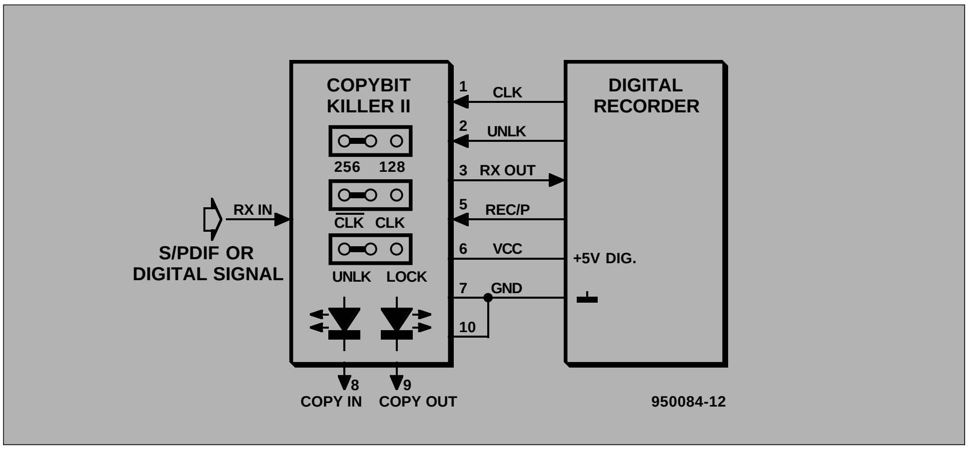

Fig. 2. Interconnection diagram of the signal source, the updated copybit eliminator and the recorder.

it depends on other connections. This involves nothing more than reversing the jumper and seeing whether the recorder locks or does not lock to the digital audio data of the eliminator.

If the connections of a particular recorder are not given, order a service manual in which these can normally be found.

A typical interconnection diagram of the signal source, the updated copybit eliminator and a digital recorder is given in Fig. 2.

Pin signals

Pin 1 (CLK). This is either 128f_s or 256f_s, depending on the position of jumper 128/256. A jumper at CLK gives an inverted clock signal; at CLK, it gives a direct clock signal.

Pin 2 (UNLK). This gives the PLL lock indication. The signal must be low when the PLL is locked (jumper at UNLK). If the opposite is the case, set the jumper in position LOCK.

Pin 3 (RXOUT). The relevant track on the board is broken before or after the coaxial/optical S/PDIF input buffer, depending on the accessibility of these points in the recorder. The part of the track from the input bus or the output of the buffer is linked to RXIN and the other part to RXOUT.

Pin 4 (RXIN). See text for pin 3.

Pin 5 (REC/P). The record indication signal is connected to this pin (high level when the equipment is recording).

Pin 6. (VCC). Supply voltage (+5 V) tapped from the recorder.

Pin 7 (GND). Link to ground of recorder.

Pins 8, 9, 10 are for connecting remotely sited LEDs. The signals may also be used as block trigger for inspecting the S/PDIF signal on an oscilloscope (only possible when the copybit is present in the input signal).

Parts list

Resistors:

R_1, R_2 = 8.2 kΩ

R_3–R_8 = 1.8 kΩ

R_5 = 27 kΩ

R_9 = 10 kΩ

Capacitors:

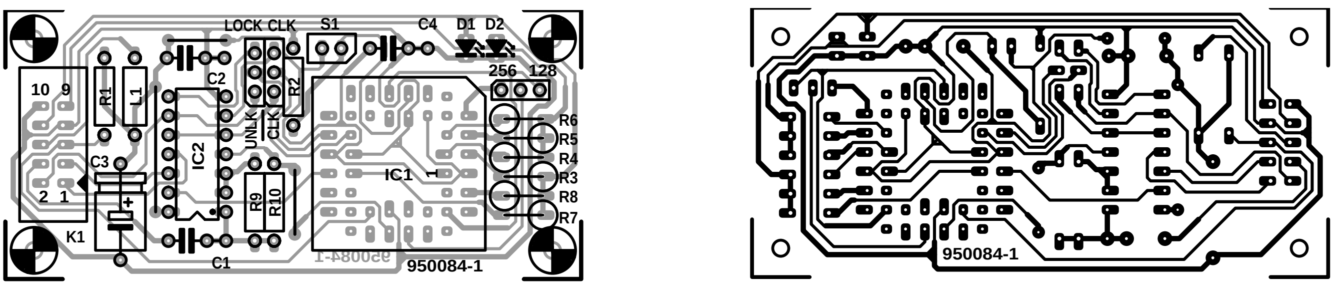

Fig. 3. Printed-circuit board for the updated copybit eliminator.

Contents

C_1 = 47 nF
C_2, C_4 = 100 nF
C_3 = 10 µF, 16 V

Semiconductors:
D_1 = LED, 3 mm, yellow
D_2 = LED, 3 mm, red

Integrated circuits:
IC_1 = MACH210 (Order no. 956504-1† – see p. 70)
IC_2 = 74HCU04

Inductors:
L_1 = 1 mH

Miscellaneous:
K_1 = 10-way right-angle box header
S_1 = switch with single make contact
PCB Order no. 950084† (see p. 70)

† The board and IC_1 may be ordered as a package: Order no.950084-C

[950084]

K_1 pin no.	Connect in recorder to	Remarks	Jumpers on PCB
Sony MDS-101 (MD recorder)			
1 (CLK)	IC_{510} pin 21 (128f_s)		128
	or IC_{505} pin 6 (256f_s)		256
2 (UNLK)	IC_{103} pin 5		LOCK
3 (RXOUT)	IC_{103} pin 65		
4 (RXIN)	CNP_{103} (connector) pin 4	break track between through metallization and connector	
5 (REC/P)	IC_{111} pin 70		
6 (VCC)	CNP_{103} pin 7		
7 (GND)	CNP_{103} pin 6		
Philips DCC-900 (DCC recorder)			
1 (CLK)	Q_{441} pin 26 (256f_s)		256
2 (UNLK)	Q_{441} pin 9		UNLK
3 (RXOUT)	coaxial: J_{421} (connector) pin 7	at mother board side	
	optical: J_{421} (connector) pin 3	at mother board side	
4 (RXIN)	signal side of C_{457} (150 pF)	break track to J_{421}	
5 (REC/P)	not known	link to VCC	
6 (VCC)	J_{421} (connector) pin 25		
7 (GND)	J_{421} (connector) pin 4		
Sony DTC-59ES (DAT recorder)			
1 (CLK)	IC_{307} pin 58 (128f_s)		128
2 (UNLK)	IC_{307} pin 31		UNLK
3 (RXOUT)	IC_{302} pin 6		
4 (RXIN)	IC_{301} pin 8	break track to this pin	
5 (REC/P)	IC_{309} pin 9		
6 (VCC)	IC_{322} pin 3	fit heat sink on to IC	
7 (GND)	chassis		
Sony DTC-690 (DAT recorder)			
1 (CLK)	R_{320} (256f_s)		256
2 (UNLK)	eponymous wire bridge		UNLK
3 (RXOUT)	IC_{302} pin 6		
4 (RXIN)	R_{316} at side of IC_{302}	disconnect resistor at IC_{302} side	
5 (REC/P)	not known	link to VCC	
6 (VCC)	eponymous wire bridge		
7 (GND)	eponymous wire bridge		
Sony DTC-750ES (DAT recorder)			
1 (CLK)	IC_{307} pin 58 (128f_s)		128
2 (UNLK)	IC_{307} pin 31		UNLK
3 (RXOUT)	IC_{307} pin 52		
4 (RXIN)	IC_{301} pin 8	break track to this pin	
5 (REC/P)	IC_{309} pin 8		
6 (VCC)	IC_{322} pin 3	fit heat sink on to IC	
7 (GND)	chassis		

Connections between K_1 and relevant points in various recorders.

Contents

SYMMETRICAL SUPPLY IN CARS

There is a problem if you want to use a car battery to power a circuit which requires a symmetrical supply voltage. Having abandoned the idea of mounting a second battery in your car for that purpose, take a serious look at the circuit discussed here. It converts the car battery voltage into symmetrical ±12 V or even ±15 V rails. Since the converter is able to supply a continuous output current of about 0.5 A, and a peak current of up to 1 A, it is eminently suited to powering control amplifiers and small power amplifiers.

If you want to obtain a symmetrical supply voltage of ± 12 V from a car battery, it is always necessary to first double the battery voltage. The only way to 'step up' a direct voltage is by using a switch-mode power supply. Such a DC-DC converter operates on one of two principles. Low-power supplies usually rely on a clever switching system by which capacitor charges are 'stacked' rapidly to give the higher voltage. If more output current is demanded, however, DC-DC converters almost invariably use an inductor for energy storage. By periodically interrupting the current through an inductor, a relatively high induction voltage is generated. This voltage may be rectified and stabilized, but it may also be 'stepped up' beforehand.

Flyback principle

Two concepts are available for the design of a DC-DC converter based on the 'switched-inductor' principle. The best known are the forward converter, the boost converter and the flyback converter. Only the latter is considered here because it is applied in the practical circuit discussed further on.

The principle is illustrated in Fig. 1. While the switch is closed, the current flowing through the inductor causes a magnetic field to be built up in the inductor core. During that period, no current flows through load resistor *R*. When the switch is opened, the inductor starts to act as an energy source. The diode then ensures that the current supplied by the inductor flows through the load. Although the output voltage may be taken directly from the inductor, it is also possible to replace the inductor by a transformer.

As regards the current through the inductor, there are two different operating conditions. The first is illustrated by the voltage and current graph in Fig. 2. After the switch is closed, energy is stored in the inductor core during period 'A' of the voltage graph. This energy is released again a little later during the period marked 'B'. Nothing happens for a while after all anergy has been released, which can be seen by the ringing of the signal during period 'C'. The lower graph shows the consequences for the current, which rises linearly during period 'A', decreases linearly again during period 'B', and drops to nought during period 'C'. Because the current drops to nought intermittently, this mode is called 'discontinuous current mode'. The advantage of this operating mode is that it allows the control system to give good response to changes in the input voltage as well as to output load variations. The disadvantage is the relatively high peak current carried by the switch.

Another possible condition is 'continuous current operation'. As illustrated by the upper curve in Fig. 3, all available time is then used to store energy and release it again. The period so created is even a little too short, because the inductor is still busy supplying energy at the end of the interval. This is indicated by the section marked 'D' in the current graph printed at the bottom. It shows that the current jumps immediately to a certain value when the switch is closed, and then rises linearly. Because the current does not drop to nought again at the end of the period, this is called 'continuous current

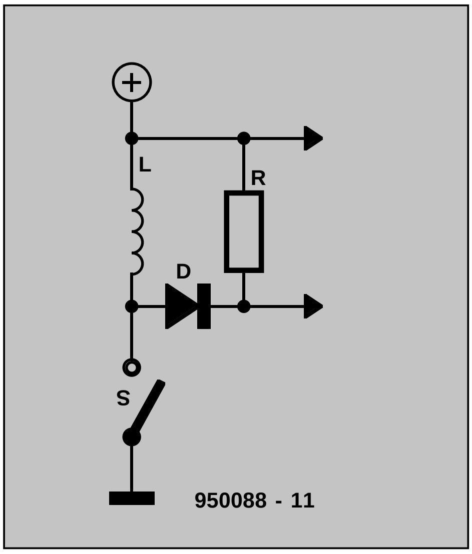

Fig. 1. As long as the switch is closed, a certain amount of energy is stored in the inductor. This energy is released when the switch is opened.

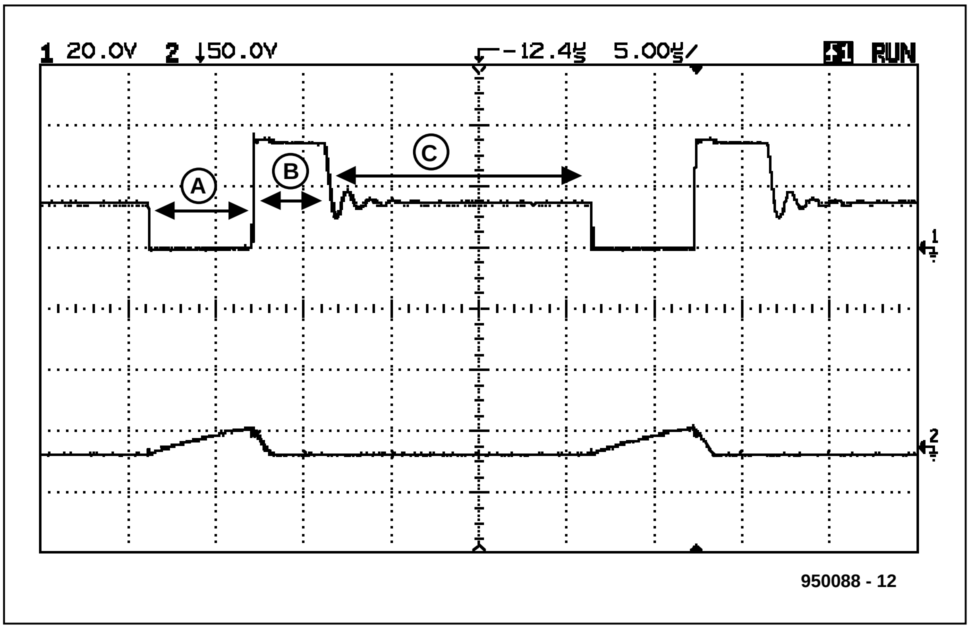

Fig. 2. Voltage and current graph of a converter in non-continuous current mode. The current graph (below) clearly shows that the current actually drops to nought between the 'down' and 'up' slopes.

mode'. An advantage of this mode is that the ripple current through the inductor is small with respect to the load current. On the down side, this mode offers not so good response to load variations.

It goes without saying that non-continuous operation is the best mode for lighter loads, while larger output currents are best handled in continuous mode. A system capable of switching between these modes depending on the current demand is, therefore, the best of both worlds. Such a control system is far from easy to stabilize, however, and that is why we have resorted to an integrated circuit which is 'easy going' and specially designed for the purpose.

The LT1070

The heart of the circuit is formed by an integrated switching regulator type LT1070. The manufacturer, Linear Technology, calls this IC a 'current mode switcher'. The internal structure of this IC is shown in Fig. 4. On board the LT1070 are all the standard ingredients of a switch-mode power supply. In fact, the LT1070 requires only a handful of external components. The most important elements are a robust high-efficiency switch, an oscillator and a measurement and control section. All of this is contained in a compact 5-pin TO-220 style case, so that the LT1070 is almost as easy to use as any of the familiar three-pin fixed voltage regulators. The LT1070 accepts input voltages between 3 V and 60 V, and works happily with a quiescent current of only 6 mA. Despite its 'modest' appearance, the regulator is capable of supplying a maximum output power of about 100 W without the help of an external power transistor.

The designation 'current mode switcher' means that the duty factor of the switch is controlled directly by the output current rather than the voltage. Referring to the block diagram, the switch starts to conduct at the start of every oscillator cycle. Regulation of the output voltage is achieved by changing the toggle level of the output current on the basis of the output voltage supplied by an error amplifier.

An internal low-drop regulator supplies a reference potential of 2.3 V for all circuits. The reference source allows input voltages between 3 V and 60 V to be used without any adverse effects on the performance of the IC. A 1.24-V bandgap voltage source is used as a reference for the error amplifier. The – input of the error amplifier is bonded out to a pin ('FB') and acts an the output voltage sensor. This feedback connection has a second function: when pulled low by an external resistor, the output of the error amplifier is decoupled from the comparator, and the latter is connected to the flyback amplifier. Because the output voltage is directly proportional with the flyback pulse, it may be adjusted without a direct link between input and output.

The error signal at the input of the comparator is also fed to a pin, 'V_c', which has four different functions. It is used for frequency compensation, soft-start, current limiting, and for complete shut-down of the regulator. The level at this pin is normally between 0.9 V (low output current) and 2.0 V (high output current). This voltage may be limited externally to set the maximum current. Soft-start may be implemented with the aid of a capacitor-coupled external clamp circuit. If V_c is pulled to ground by means of a diode, the regulator enters a kind of stand-by state, while pulling the voltage under 0.15 V causes the regulator to be switched off completely. In the latter state, the quiescent current is reduced to a mere 50 µA.

Practical circuit

As already mentioned, the LT1070 works happily with a minimum of external components. Consequently, Fig. 5 shows a very simple circuit diagram. Indeed, had we limited ourselves to the parts required for the function of DC-DC converter only, the circuit would have been even simpler, because a fair number of components serves for buffering and cleaning of the input and out-

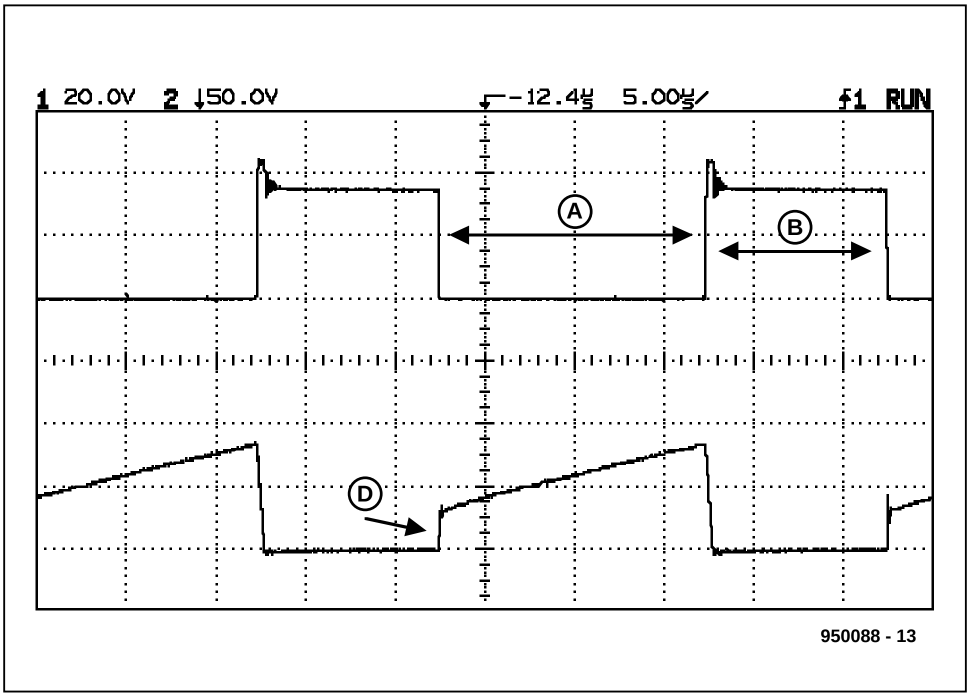

Fig. 3. The same graphs as in Fig. 2, but measured on a converter operating in continuous current mode. All available time is used to store and release energy.

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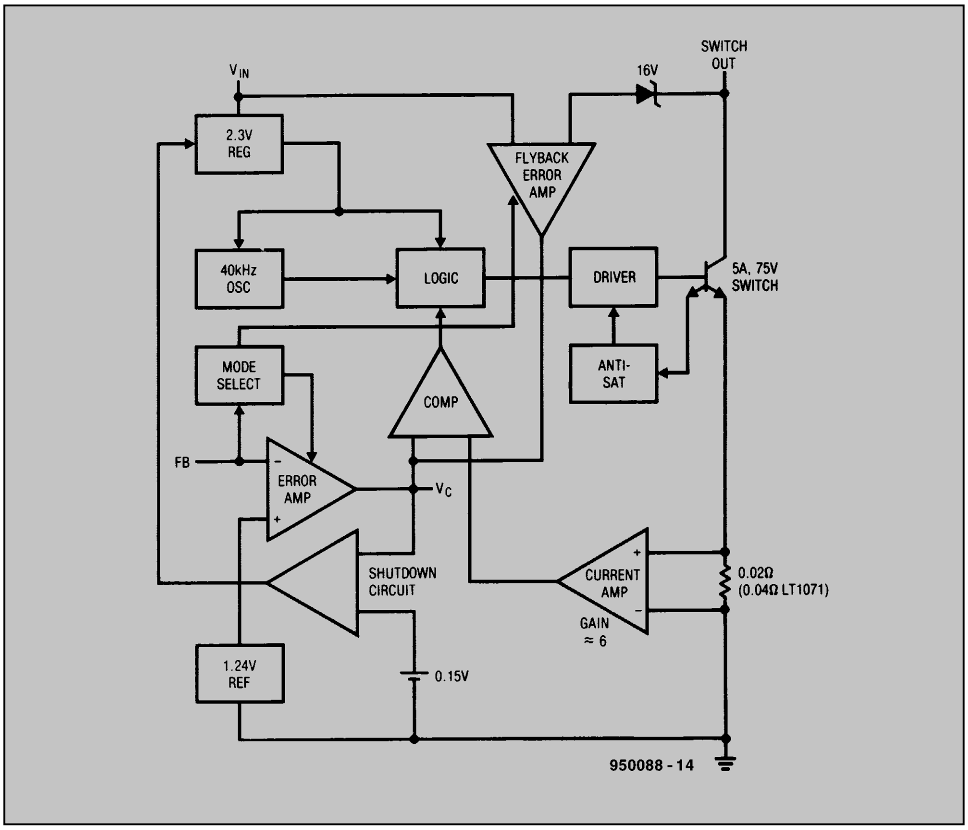

Fig. 4. The internal structure of the LT1070 is fairly complex, and contains just about everything you need to build a DC-DC converter.

put voltages.

The core of the converter is actually restricted to IC_1, transformer Tr_1, the rectifiers connected to secondary windings of Tr_1, and resistor network R_3-R_4. As stated earlier, a flyback-type regulator allows the output voltage to be set by feeding (a part of) the output voltage back to the feedback input. Here, that is done with the aid of R_3 and R_4. The ratio between these two resistors therefore enables the output voltage to be set to ±12 V or ±15 V without the need of changing the turns ratio of the transformer. The resistor values indicated in the circuit diagram gave an output voltage of ±13.8 V on the prototype of the converter.

The LT1070 keeps the output voltage constant within a few millivolts (which can be measured only when changing from no load to full load). Note, however, that only the positive output voltage is used as a reference for the regulator. As long as the converter is symmetrically loaded (i.e., roughly equal current consumption on the positive and negative output rails), there will be no undue unbalance, and the regulation on the negative rail will be hardly worse than that on the positive rail. However, when only the negative rail is loaded, that output voltage may drop appreciably. So, if you want to load the negative rail only, be sure to provide a continuous load on the positive rail, for instance, with the aid of a resistor. If you are after a 'rock-steady' supply, connect a linear regulator to each of the converter outputs.

Components D_1, R_2 and C_7 form a so-called snubber network, which serves to prevent the LT1070's output voltage from exceeding its maximum value (65 V). Because of the stray inductance in the circuit, a large voltage surge may occur the moment the chopper switch opens. This surge is diverted via D_1 and C_7, while the capacitor, in turn, is discharged slowly by R_2.

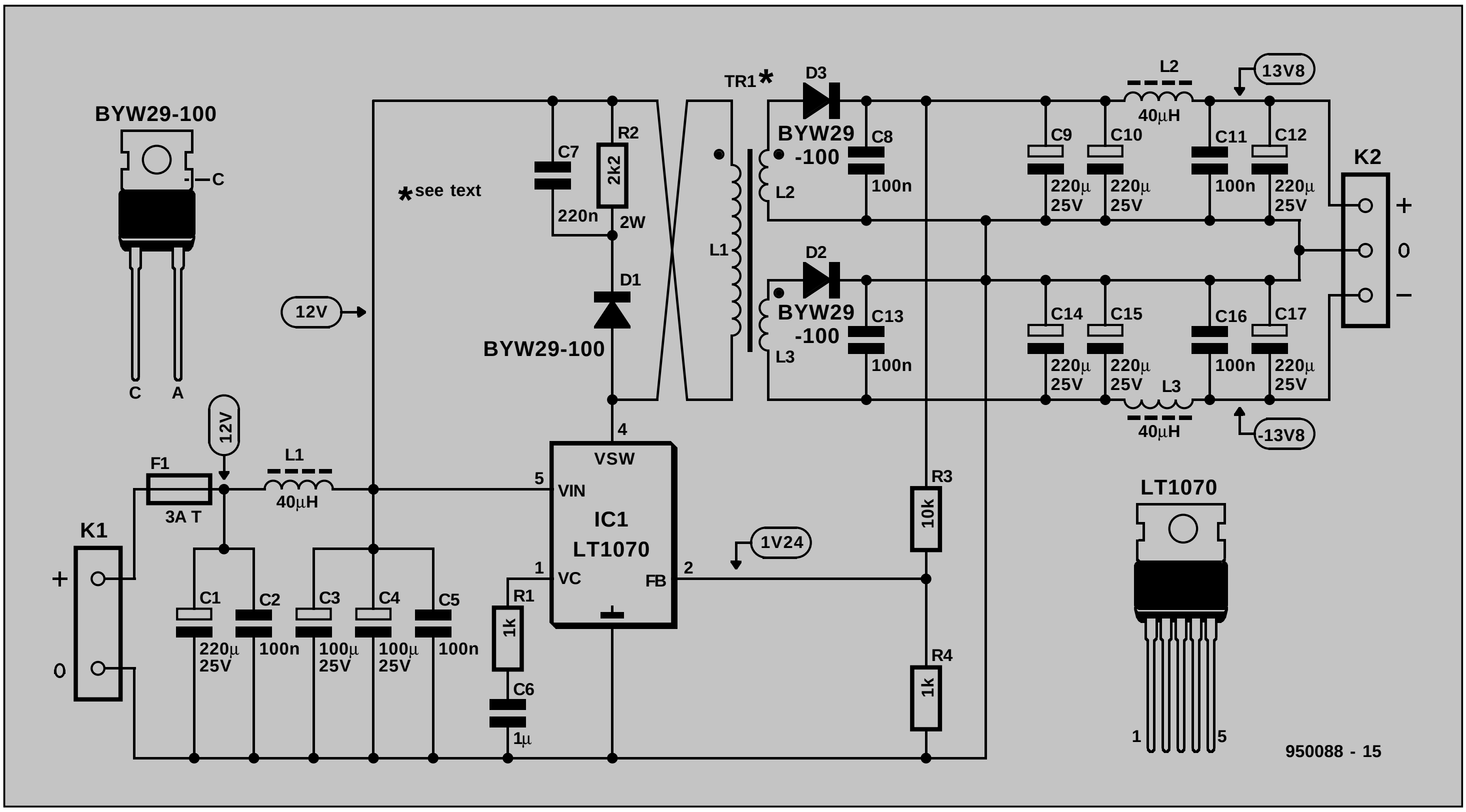

Fig. 5. Apart from the IC and the transformer, the circuit diagram contains only a double rectifier, a number of reservoir capacitors and noise suppression filters.

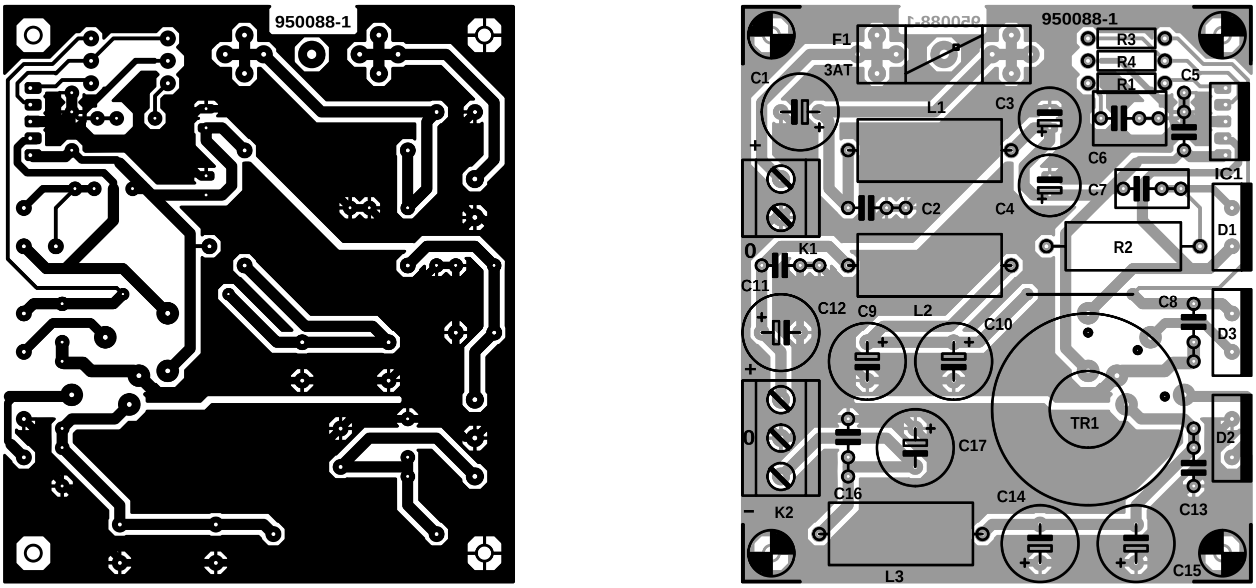

Fig. 6. Populating the board is easy. Its moderate size allows the board to be fitted in many different types of metal enclosure (board not available ready-made).

To keep unwanted emissions to a minimum, the circuit has no fewer than three *LC* filters: one for the input voltage (L_1-C_2), and one on each of the output rails (L_2-C_{11} and L_3-C_{16}). For L_1, L_2 and L_3 it is best to use those well-known triac suppressor coils with a minimum current rating of 1 A (usually 2.5 A). If your demands are not so high, the coils may be replaced with 6-hole ferrite beads with a few turns of wire through them. Although the suppression of the 40-kHz fundamental component is then slightly less effective, the effect on higher harmonics above 500 kHz is nearly identical.

The relatively high pulse currents in the circuit cause the electrolytic reservoir capacitors to run fairly hot (up to about 50 °C). To keep their heat dissipation within reasonable limits, these capacitors are therefore fitted in pairs. To ensure the highest possible efficiency and life expectancy of the converter, it would be better to use electrolytic capacitors specially designed for use in switch-mode power supplies. However, these costly and difficult to obtain parts are not strictly required in the present circuit. Our prototype gave satisfactory results with 'ordinary' caps fitted.

Do not use ordinary diodes (like 1N4002 etc.) in positions D_1, D_2 and D_3, because they are too slow in this application. Almost any real switching diode may be used, as long as it is capable of passing a current of at least 3 A. It is best to use fast diodes with soft recovery — the latter feature is important to keep spurious emission to a minimum. The diode indicated in the circuit diagram is a Schottky type rated at 100 V, 8 A, which has an additional advantage in not dropping too much voltage.

Construction

The printed circuit board designed for the DC-DC converter is shown in Fig. 6. The size is modest, while the tracks are laid out generously. Populating the PCB with step-by-step reference to the component overlay and the parts list is not expected to cause undue problems. The IC and the diodes are purposely located at the edge of the board, so that they are easily secured to a heat-sink (using washers and plastic bushes). Alternatively, if you use a metal (die-cast or aluminium) case for the converter, that may be used as a heat-sink as well. However, we haven't got as far as fitting the board into a case. First, concentrate on making the transformer, Tr_1. The core is that of an ordinary triac suppressor coil with an inductance between 25 µH and 100 µH, and a current rating between 3 A and 5 A. The parts list states a few types that may be used. Usually, such coils have between 30 and 50 turns of enamelled copper wire on the core. This winding may be left in place, and becomes the primary winding of the transformer. Count the number of turns. Next, apply two secondary windings over the primary, each having the same number of turns as the primary. Use 0.5-mm dia (24SWG) enamelled copper wire, and wind the two secondaries simultaneously, that is, with two wires at the same time. Be sure to observe the same winding direction as the primary. It does not matter whether you start winding the wire from the left or the right — the thing to keep an eye on is whether the wire is inserted into the core from below or from the top. So, look carefully at how the primary is wound. Distribute the two new windings as evenly as possible across the circumference of the core. The final construction is illustrated in Fig. 7.

A final remark on the transformer. If a

COMPONENTS LIST

Resistors:
R1;R4 = 1kΩ
R2 = 2kΩ2 2W
R3 = 10kΩ

Capacitors:
C1;C9;C10;C12;C14;C15;C17 = 220µF 25V radial
C2;C5;C8;C11;C13;C16 = 100nF
C3;C4 = 100µF 25V radial
C6 = 1µF MKT
C7 = 220nF

Inductors:
L1;L2;L3 = SFT10-30 or SFT1030 (40µH) (TDK)
Tr1 = SFT12-50 or SFT1240 (see text)

Semiconductors:
D1;D2;D3 = BYW29-100
IC1 = LT1070 (Linear Technology)

Miscellaneous:
K1 = 2-way PCB terminal block.
K2 = 3-way PCB terminal block.
F1 = fuse 3A (slow) w. PCB mount holder.
Metal case, e.g., Hammond 1590S, 110x82x44mm.
Insulation set (washer and bush) for IC1, D2, D3.

Most components for this project are available from C-I Electronics, P.O. Box 22089, NL-6360-AB, Nuth, The Netherlands. Fax (+31) 45 5241877

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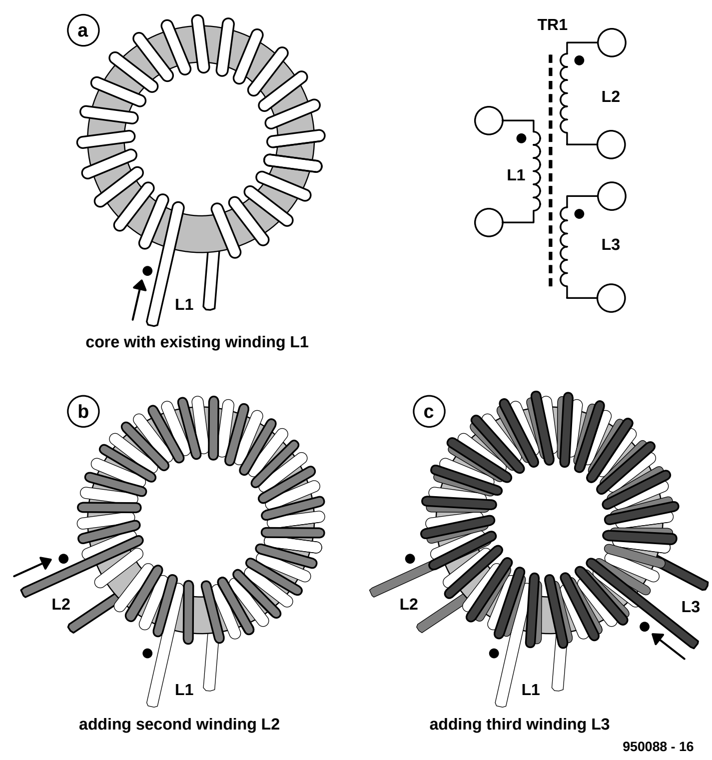

Fig. 7. Transformer Tr1 consists of a normal ring core suppressor choke, whose existing winding is used as the primary winding. A bifilar secondary is wound over the primary, observing that the winding direction is the same as that of the primary.

higher output voltage is desired (more than ±15 V), the efficiency of the converter may be improved by making the secondary windings a little larger — for example, 60 turns instead of 50. Fortunately, the exact number of turns is not critical, so gross errors are hard to make.

Testing and setting up

The photograph in Fig. 8 shows the completed prototype of the converter. To be able to test the circuit properly, diode D_2 should not be fitted for the moment. Connect two 1-kΩ load resistors to the output rails of the converter (one to the positive rail and one to the negative rail). In the interest of safety, it is best at this stage to use an adjustable power supply instead of a car battery. Connect it to the converter, and increase the voltage slowly. The converter should start to work at an input voltage of between 3 V and 5 V. If you do not have an adjustable supply, connect a 12-V, 5-W lamp in series with the battery. The lamp will limit the current in case something goes wrong.

Use an oscilloscope to check the voltage at pin 4 of the LT1070 against the oscillogram in Fig. 2. It is essential to be able to discern three levels: zero volts, the supply voltage and the doubled supply voltage. If you find that there are only two levels, the winding direction of the secondaries on Tr_1 is wrong. Remove the transformer and wind it again.

If the waveform is okay, the input voltage may be increased to 12 V, or the lamp may be removed. The positive output voltage should then be around 13.8 V. If that is the case, D_2 may be fitted on the board. Next, run another check on the waveform shape at pin 4 of IC_1. Also check the level of the negative output voltage at the – clamp of K_2.

If the circuit behaves properly so far, it is time to increase the load current. Secure the heat-sink to the IC and the diodes, and exchange the 1-kΩ resistors with 5-watt car lamps. The waveform at both outputs should be as shown in Fig. 3. If you use an adjustable supply for this test, be sure to turn it up to 12 V first, and not connect the lamps until the output voltages of +12 V and –12 V are present. If you do it the other way around, the power supply may actuate its current limiter. That may happen because the converter always tries to supply the necessary output current. Where a power of, say, 12 W requires a current of 1 A at an input voltage of 12 V, that goes up to 2 A at 6 V, and 4 A at 3 V. Such output currents are beyond the capacity of most (hobby) power supplies, hence you have to start without the load connected to the converter.

For the sake of completeness, the circuit diagram shows a number of measurement values. Although the input voltage is given as 12 V nominally, the actually allowed level here is up to 15 V. Using the indicated component values and an input voltage of 12 V, the output voltage should be about ±13.8 V. This voltage may be changed within certain limits by changing the ratio between R_3 and R_4. In all cases, however, a voltage of 1.24 V (±5%) should be present at pin 2 of IC_1. If not, the regulation of the IC does not function properly. This voltage should correspond to the internal reference voltage. At a too high voltage, the output of the converter is probably not loaded, while a too low voltage at pin 2 indicates a too heavy load or a too low input voltage.

The internal regulation voltage of the IC may be measured at pin 1. This voltage depends on the output current, and changes between 1.1 V at no load to about 2 V at full load.

Final remarks

Having passed all tests and experiments without serious mishaps, the circuit is ready for the finishing touches, and the test gear and car lamps may be put away. To keep unwanted electromagnetic radiation as low as possible, the circuit should be built into a sturdy metal case. The prototype was housed in a die-cast case from Hammond. The type number is 1590S and the outside dimensions are approximately 110×82×44 mm. The circuit board fits exactly in this case, and is easily secured to the bottom with the aid of four PCB pillars. The IC and the three diodes may be secured to the side panel using insulating washers. This arrangement will afford sufficient cooling for not too heavy use. The input and output cables to be connected to K_1 and K_2 enter the case through suitable grommets, and should be fitted with a heavy-duty strain relief at the inside.

As already mentioned, the converter is capable of supplying a peak output current of up to 1 A, and a continuous current of up to 0.5 A. Note, however, that these ratings are achieved at normal room temperatures only (up to about 25 °C). In the blazing sun, however, the temperature inside a car may easily reach about 60 °C, which means that the converter is not able to supply its maximum power because the thermal shut-down will be actuated much earlier than under normal circumstances. Fortunately, the same thermal protection ensures that the IC can not be damaged or destroyed by high temperatures.

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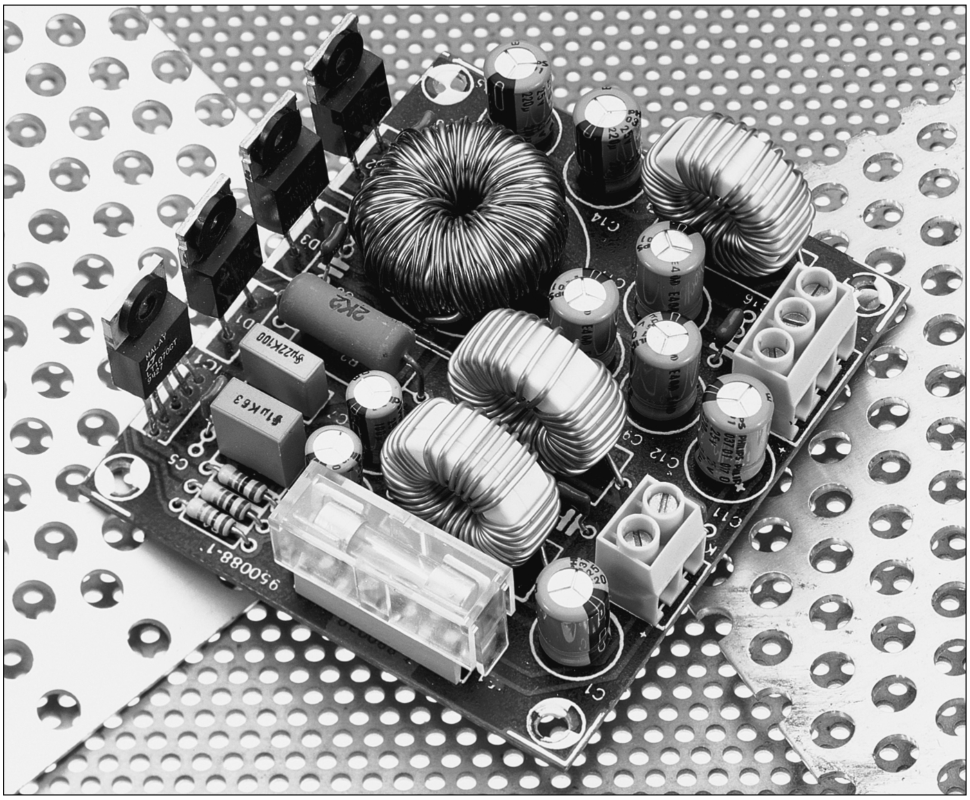

Fig. 8. Completed prototype board. Diode D2 should be omitted for the purpose of testing the circuit.

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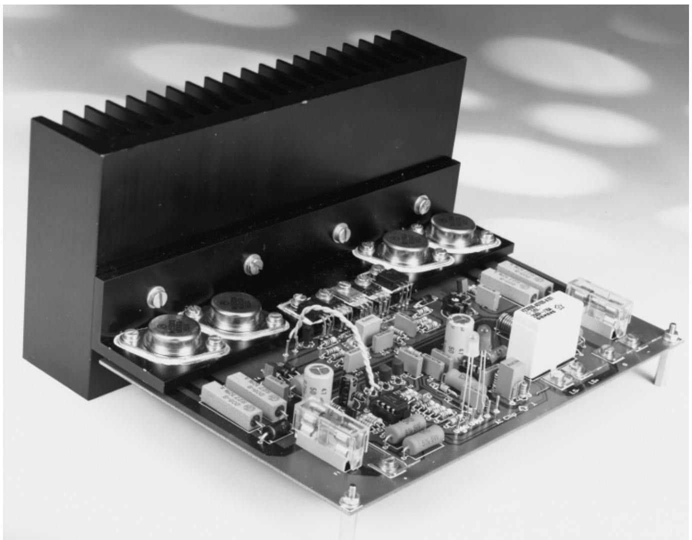
PA300 POWER AMPLIFIER



Design by A. Riedl

Taken by themselves, the properties of the PA300 amplifier are not revolutionary. But taken in combination, they show something special: a robust 300 watt hi-fi power amplifier that is not too difficult to build.

There are several starting points to the design of a power amplifier: pure hi-fi without any compromise; simplicity and reliability; high output power. The design of the present amplifier is a mixture of these. The result is a unit that does not use esoteric components, is not too complex, and is fairly easily reproduced. In fact, it could well be named a 'Hi-fi public address amplifier'.

There will be a few eyebrows raised at the power output of 300 watts (into 4Ω); it is true, of course, that in the average living room 30–40 W per channel is more than sufficient. However, peaks in the reproduced music may have a power of 10–20 times the average level. This means that some reserve power is desirable. Also, there are loudspeakers around with such a low efficiency that a lot more than 30–40W is needed. And, last but not least, there are many people who want an amplifier for rooms much larger than the average living room, such as an amateur music hall.

Straightforward design

Since every amplifier contains a certain number of standard components, the circuit of Fig. 1 will look pretty familiar to most audio enthusiasts. Two aspects may hit the eye: the higher than usual supply voltage and the presence of a couple of ICs. The first is to be expected in view of the power output. One of the ICs is not in the signal path and this immediately points to it being part of a protection circuit. What is unconventional is an IC in the input stage. Normally, this stage consists of a differential amplifier followed by a voltage amplifier of sorts, often also a differential amplifier, to drive the predriver stages. In the PA300, the entire input stage is contained in one IC, a Type NE5534 (IC_1).

The internal circuit of IC_1 is shown in the box on further on in this article. It may also be of interest to note that the NE5534 is found in nine out of every ten CD players (as amplifier in the analogue section). This is reflected in its price which is low. Its only drawback is that its supply voltage is far below that of the remainder of the amplifier. This means an additional symmetrical supply of ±15 V. Moreover, it restricts the drive capability of the input stage. The supply requirement is easily met with the aid of a couple of zener diodes and resistors. The drive restriction means that the amplifier must provide a measure of voltage amplification after the input stage.

Circuit description

The input contains a high-pass filter, C_5-R_3 and a low-pass filter, R_2-C_6. The combination of these filters limits the bandwidth of the input stage to a realistic value: it is not necessary for signals well outside the audio range to be amplified – in fact, this may well give rise to difficulties.

Opamp IC_1 is arranged as a differential amplifier; its non-inverting (+) input functions as the meeting point for the overall feedback. The feedback voltage, taken from junction D_7-D_8, is applied to junction R_4-R_5 via R_9. Any necessary compensation is provided by C_9, C_{12} and C_{14}. The voltage amplification is determined by the ratio R_9:R_5, which in the present circuit is ×40.

The output of IC_1 is applied to drive stages T_1 and T_3 via R_6. These transistors operate in Class A: the current drawn by them is set to 10 mA by voltage divider R_{10}-R_{13} and their respective emitter resistors. Their voltage and current amplification is appreciable, which is as required for the link between the input and output stages.

The output amplifier proper consists of drive stages T_6 and T_7 and power transistors T_8, T_9, T_{14}, T_{15}. which have been arranged as symmetrical power darlingtons. Because of the high power, the output transistors are connected in parallel. The types used can handle a collector current of 20 A and have a maximum dissipation of 250 W.

The output stages operate in Class AB to ensure a smooth transition between the n-p-n and p-n-p transistors, which prevents cross-over distortion. This requires a small current through the power transistors, even in the absence of an input signal. This current is provided by 'zener' transistor T_2,

Technical data (measured with power supply shown in Fig. 2)

Input sensitivity	1 V_{rms}
Input impedance	17.8 kΩ
Power output (0.1% THD)	164 W into 8 Ω
	275 W into 4 Ω
Music power(500 Hz burst 5 cycles on, 5 cycles off)	176 W into 8 Ω
	306 W into 4 Ω
Power bandwidth (90 W into 8 Ω)	7 Hz–67 kHz
Slew rate	20 V/μs
Signal-to-noise ratio (referred to 1 W into 8 Ω)	>96 dB (A-weighted)
Harmonic distortion (THD+N) (bandwidth 80 kHz)	at 1 W into 8 Ω: <0.004% (1 kHz)
	at 150 W into 8 Ω: <0.001% (1 kHz)
	<0.05% (20 Hz–20 kHz)
Intermodulation distortion (50 Hz:1 kHz; 4:1)	at 1 W into 8 Ω: <0.003%
	at 100 W into 8 Ω: <0.0035%
Dynamic IM (rectangular wave + 15 kHz sine wave)	1 W into 8 Ω: <0.004%
	150 W into 8 Ω: <0.06%
Damping factor (at 8 Ω)	<345 at 1 kHz
	<275 at 20 kHz

which puts a small voltage on the bases of T_6 and T_7 so that these transistors just conduct in quiescent operation. The level of the quiescent current is set accurately with P_1.

To ensure maximum thermal stability, transistors T_1–T_3 and T_6–T_7 are mounted on and the same heat sink. This keeps the quiescent current within certain limits. With high drive signals, this current can reach a high level, but when the input signal level drops, the current will diminish only slowly until it has reached its nominal value.

Diodes D_7, D_8 protect the output stages against possible counter voltages generated by the complex load. Resistor R_{30} and capacitor C_{17} form a Boucherot network to enhance the stability at high frequencies. Inductor L_1 prevents any problems with capacitive loads (electrostatic loudspeakers). Resistor R_{29} ensures that the transfer of rectangular signals are not adversely affected by the inductor.

Protection circuits

As any reliable amplifier, the PA300 is provided with adequate protection measures. These start with fuses F_1 and F_2, which guard against high currents in case of overload or short-circuits. Since even fast fuses are often not fast enough to prevent the

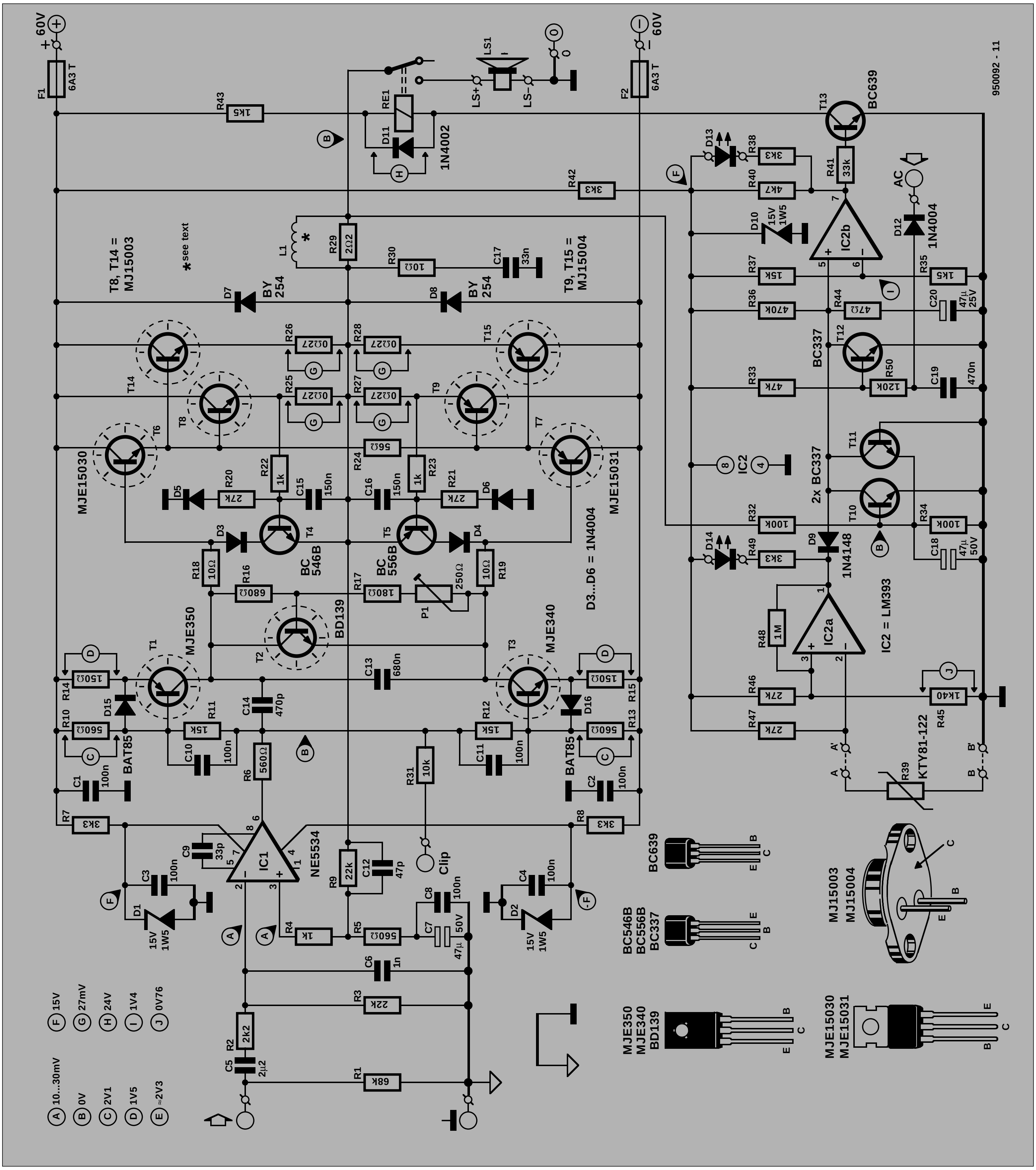


Fig. 1. With the exception of an IC at the input, the circuit of the PA300 amplifier is conventional.

power transistors giving up the ghost in such circumstances, an electronic short-circuit protection circuit, based on T_4 and T_5, has been provided. When, owing to an overload or short-circuit, very high currents begin to flow through resistors R_{25} and R_{27}, the potential drop across these resistors will exceed the base-emitter threshold voltage of T_4 and T_5. These transistors then conduct and short-circuit or reduce drive signal at their bases. The output current then drops to zero.

If a direct voltage appears at the output terminals, or the temperature of the heat sink rises unduly, relay Re_1 removes the load from the output. The loudspeakers are also disconnected by the relay when the mains is switched on (power-on delay) to prevent annoying clicks and plops.

The circuits that make all this possible consist of dual comparator IC_2, transistors T_{10}–T_{13}, and indicator diodes D_{13} and D_{14}. They are powered by the 15 V line provided by zener diode D_{10} and resistor R_{42}.

The 'AC' terminal on the PCB is linked to one of the secondary outputs on the mains transformer. As soon as the mains is switched on, an alternating voltage appears at that terminal, which is rectified by D_{12} and applied as a negative potential to T_{12} via R_{50}. The transistor will then be cut off, so that C_{20} is charged via R_{36} and R_{44}. As long as charging takes place, the inverting (+) input of comparator IC_{2b} is low w.r.t. the non-inverting (–) input. The output of IC_{2b} is also low, so that T_{13} is cut off and the relay is not energized. This state is indicated by the lighting of D_{13}. When C_{20} has been charged fully, the comparator changes state, the relay is energized (whereupon D_{13} goes out) and the loudspeakers are connected to the output. When the mains is switched off, the relay is deenergized instantly, whereupon the loudspeakers are disconnected so that any switch-off noise is not audible.

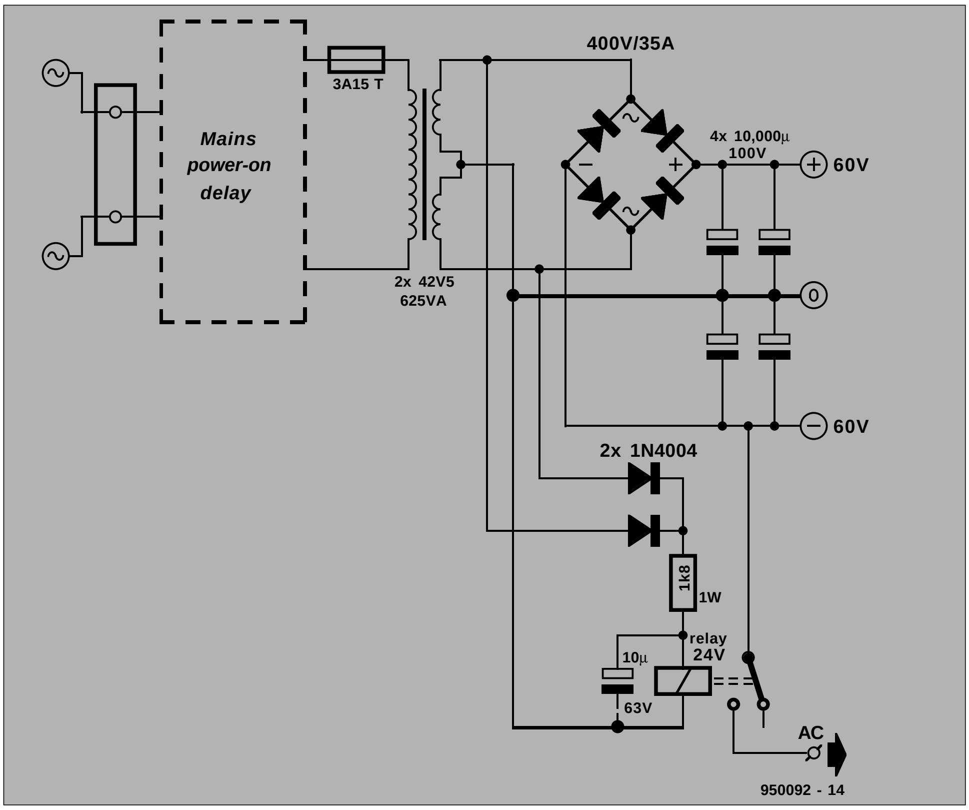


Fig. 2. The power supply is straightforward, but can handle a large current. Voltage 'AC' serves as drive for the power-on delay circuit.

The direct-voltage protection operates as follows. The output voltage is applied to T_{10} and T_{11} via potential divider R_{32}-R_{34}. Alternating voltages are short-circuited to ground by C_{18}. However, direct voltages greater than +1.7 V or more negative than –4.8 V switch on T_{10} or T_{11} immediately. This causes the +ve input of IC_{2a} to be pulled down, whereupon this comparator changes state, T_{13} is cut off, and the relay is deenergized. This state is again indicated by the lighting of D_{13}.

Strictly speaking, temperature protection is not necessary, but it offers that little bit extra security. The temperature sensor is R_{39}, a PTC (positive temperature coefficient) type, which is located on the board in a position where it rests against the rectangular bracket. Owing to a rising temperature, the value of R_{39} increases until the potential at the –ve input of IC_{2a} rises above the level at the +ve input set by divider R_{45}-R_{46}, whereupon the output of IC_{2a} goes low. This causes IC_{2b} to change state, whereupon T_{13} is cut off and the relay is deenergized. This time, the situation is indicated by the lighting of D_{14}. The circuit has been designed to operate when the temperature of the heat sink rises above 70 °C. Any relay clatter may be obviated by reducing the value of R_{48}.

The terminal marked 'CLIP' on the PCB is connected to the output of IC_1 via R_{31}. It serves to obtain an external overdrive indication, which may be a simple combination of a comparator and LED. Normally, this terminal is left open.

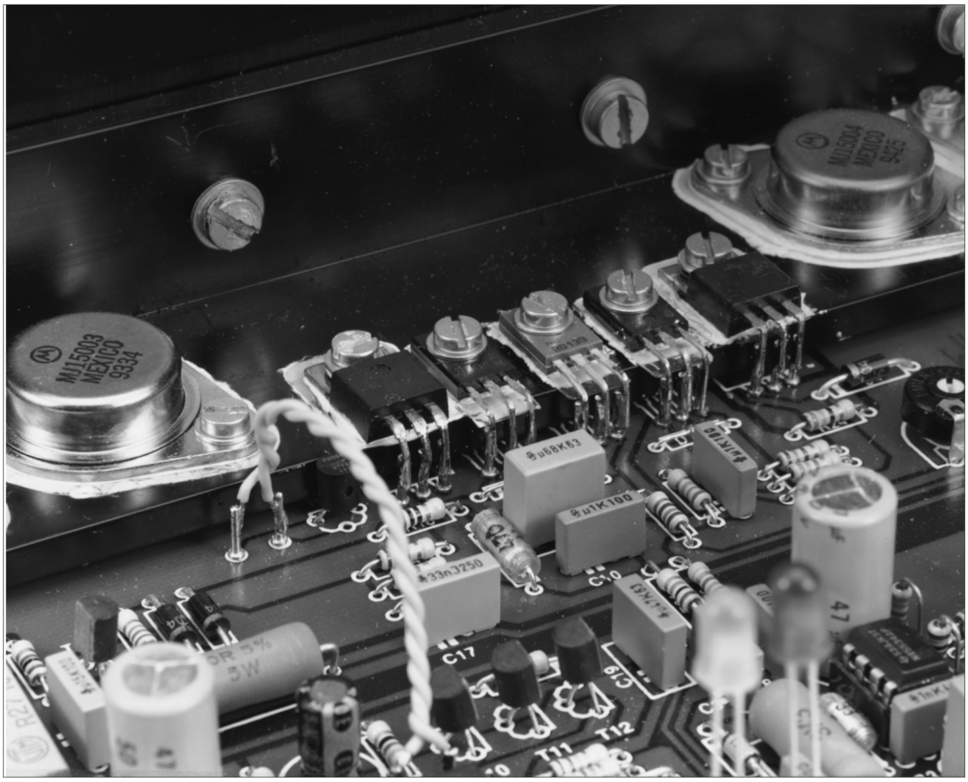


Fig. 3. This close-up photograph shows clearly how the transistors are fitted to the heat sink via a rectangular bracket.

Power supply

As with most power amplifiers, the ±60 V power supply need not be regulated. Owing to the relatively high power output, the supply needs a fairly large mains transformer and corresponding smoothing capacitors—see Fig. 2. Note that the supply shown is for a mono amplifier; a stereo outfit needs two supplies.

The transformer is a 625 VA type, and the smoothing capacitors are 10 000 µF, 100 V electrolytic types. The bridge rectifier needs to be mounted on a suitable heat sink or be mounted directly on the bottom cover of the metal enclosure.. The transformer needs two secondary windings, providing 42.5 V each. The prototype used a toroidal transformer with 2×40 V secondaries. The secondary winding of this type of transformer is easily extended: in the prototype 4 turns were added and this gave secondaries of 2×42.5 V.

The box ‘Mains power-on delay’ provides a gradual build-up of the mains voltage, which in a high-power amplifier is highly advisable. A suitable design was published in *305 Circuits* (page 115).

The relay and associated drive circuit is intended to be connected to terminal ‘AC’ on the board, where it serves to power the power-on circuit. If a slight degradation of the amplifier performance is acceptable, this relay and circuit may be omitted and the PCB terminal connected directly to one of the transformer secondaries.

Construction

Building the amplifier is surprisingly simple. The printed-circuit board in Fig. 4 is well laid out and provides ample room. Populating the board is as usual best started with the passive components, then the electrolytic capacitors, fuses and relay. There are no ‘difficult’ parts.

Circuits IC_1 and IC_2 are best mounted in appropriate sockets.

Diodes D_{13} and D_{14} will, of course, have to be fitted on the front panel of the enclosure and are connected to the board by lengths of flexible circuit wire.

Inductor L_1 is a DIY component; i consists of 15 turns of 1 mm. dia. enamelled copper wire around R_{29} (not too tight!).

Since most of the transistors are to be mounted on and the same heat sink, they are all located at one side of the board. However, they should first be fitted on a rectangular bracket, which is secured to the heat sink and the board—see Fig. 3. Note that the heat sink shown in this photograph proved too small when 4 Ω loudspeakers were used. With 8 Ω speakers, it was just about all right, but with full drive over sustained periods, the temperature protection circuits were actuated. If such situations are likely to be encountered, forced cooling must be used.

As already stated, temperature sensor R_{39} should rest (with its flat surface) against the rectangular bracket. On the board, terminals ‘A’ and ‘B’ terminals to the left of R_{39} must be connected to ‘A’ and ‘B’ above IC_2 with a twisted pair of lengths of insulated circuit wire as shown in Fig. 3.

The points where to connect the loudspeaker leads and power lines are clearly marked on the board. Use the special flat AMP connectors for this purpose: these have large-surface contacts that can handle large currents. The loudspeaker cable should have a cross-sectional area of not less than 2.5 mm^2.

Finally

How the amplifier and power supply are assembled is largely a question of individual taste and requirement. The two may be combined into a mono amplifier, or two each may be built into a stereo amplifier unit. Our preference is for mono amplifiers, since these run the least risk of earth loops and the difficulties associated with those. It is advisable to make the ‘0’ of the supply the centre of the earth connections of the electrolytic capacitors and the centre tap of the transformer.

The single earthing point on the supply and the board must be connected to the enclosure earth by a short, heavy-duty cable. This means that the input socket must be

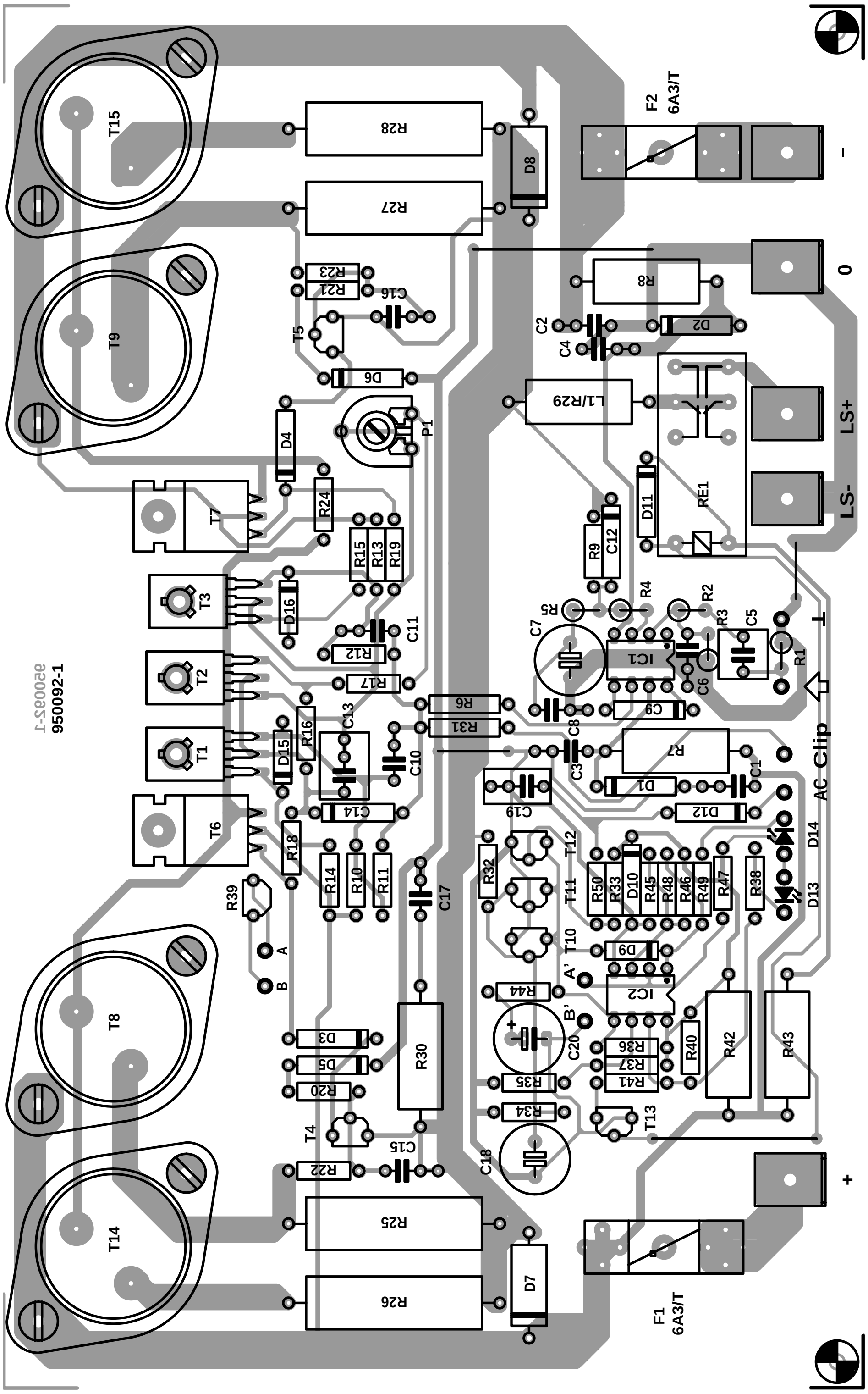


Fig. 4a. Component layout of the printed-circuit board for the 300 W power amplifier.

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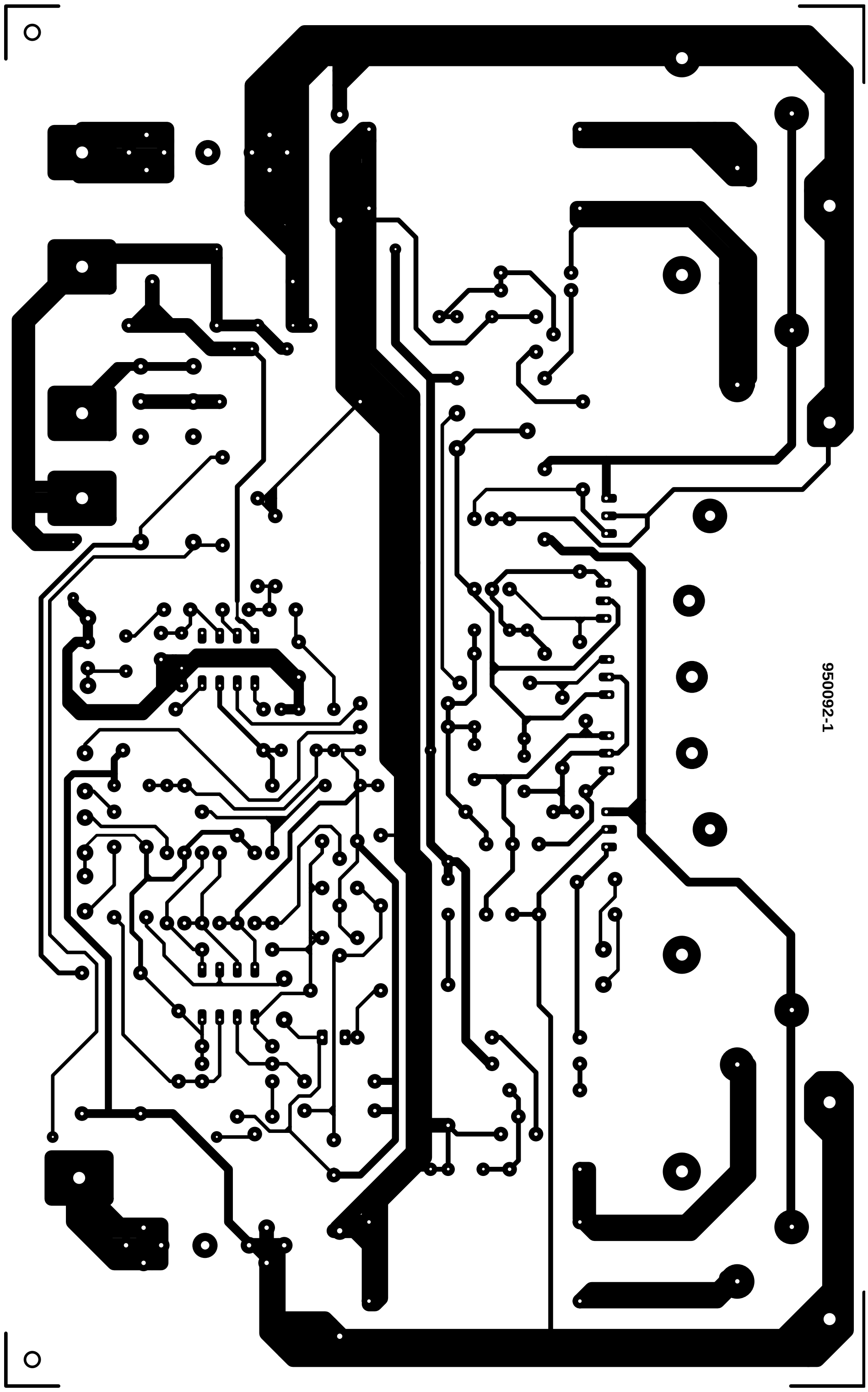


Fig. 4b. Track layout of the printed-circuit board for the 300 W power amplifier.

an insulated type. This socket must be linked to the input on the board via screened cable.

To test the amplifier, turn P_1 fully anticlockwise and switch on the mains. After the output relay has been energized, set the quiescent current. This is done by connecting a multimeter (direct mV range) across one of resistors R_{25}–R_{28} and adjusting P_1 until the meter reads 27 mV (which corresponds to a current of 100 mA through each of the four power transistors). Leave the amplifier on for an hour or so and then check the voltage again: adjust P_1 as required.

Test results

The technical data given on page0 0 were verified or obtained with a power supply as shown in Fig. 2. They show that in spite (or because?) of its simple design, the amplifier offers excellent performance. The distortion figures are particularly good.

Measurements with the Audio Precision analyser are illustrated in Fig. 5.

Figure 5a shows the total harmonic distortion (THD+N) over a frequency range of 20 Hz to 20 kHz.with a bandwidth of 80 kHz and a power output of 150 W into 8 Ω. Up to 1 kHz, the distortion is very low and then increases, which is usual and caused by the inertia of the semiconductors.

Figure 5b shows the distortion at 1 kHz as a function of the output level at a bandwidth of 22 Hz to 22 kHz. The dashed curve refers to a load of 4 Ω and the solid curve to a load of 8 Ω. The irregularities between 10 W and 100 W are not caused by the amplifier but by the limits of the measuring range of the analyser. From the clipping points, the curves rise almost vertically.

Figure 5c shows the maximum for a distortion of 0.1%. The dashed curve (4 Ω load) is very close to the 300 W line. The small reduction at low frequencies is caused by the imperfectness of the electrolytic buffer capacitors in the power supply.

Figure 5d shows the Fourier analysis of a 1 kHz signal for a power output of 1 W into 8 Ω. The fundamental frequency is suppressed. The 2nd and 3rd harmonics are down by 110 dB and 120 dB respectively referred to the fundamental frequency. The THD+N figure at this measurement was 0.0009%.

Parts list

Resistors:
R_1 = 68 kΩ
R_2 = 2.2 kΩ
R_3, R_9 = 22 kΩ
R_4, R_{22}, R_{23} = 1 kΩ
R_5, R_6, R_{10}, R_{13} = 560 Ω
R_7, R_8, R_{42} = 3.3 kΩ, 5 W
R_{11}, R_{12}, R_{37} = 15 kΩ
R_{14}, R_{15} = 150 Ω
R_{16} = 680 Ω
R_{17} = 180 Ω
R_{18}, R_{19} = 10 Ω
R_{20}, R_{21}, R_{46}, R_{47} = 27 kΩ
R_{24} = 56 Ω
R_{25}–R_{28} = 0.27 Ω, 5 W
R_{29} = 2.2 Ω, 5 W
R_{30} = 10 Ω, 5 W
R_{31} = 10 kΩ
R_{32}, R_{34} = 100 kΩ
R_{33} = 47 kΩ
R_{35} = 1.5 kΩ
R_{36} = 470 kΩ
R_{38}, R_{49} = 3.3 kΩ
R_{39} = sensor Type KTY81-122
R_{40} = 4.7 kΩ
R_{41} = 33 kΩ
R_{43} = 1.5 kΩ, 5 W
R_{44} = 47 Ω
R_{45} = 1.40 kΩ, 1%
R_{48} = 1 MΩ
R_{50} = 120 kΩ
P_1 = 250 Ω preset

Capacitors:
C_1–C_4, C_8, C_{10}, C_{11} = 100 nF
C_5 = 2.2 µF polypropylene, pitch 5 mm
C_6 = 1 nF
C_7, C_{18} = 47 µF, 50 V, bipolar, radial;
C_9 = 33 pF, 160 V, polystyrene
C_{12} = 47 pF, 160 V, polystyrene

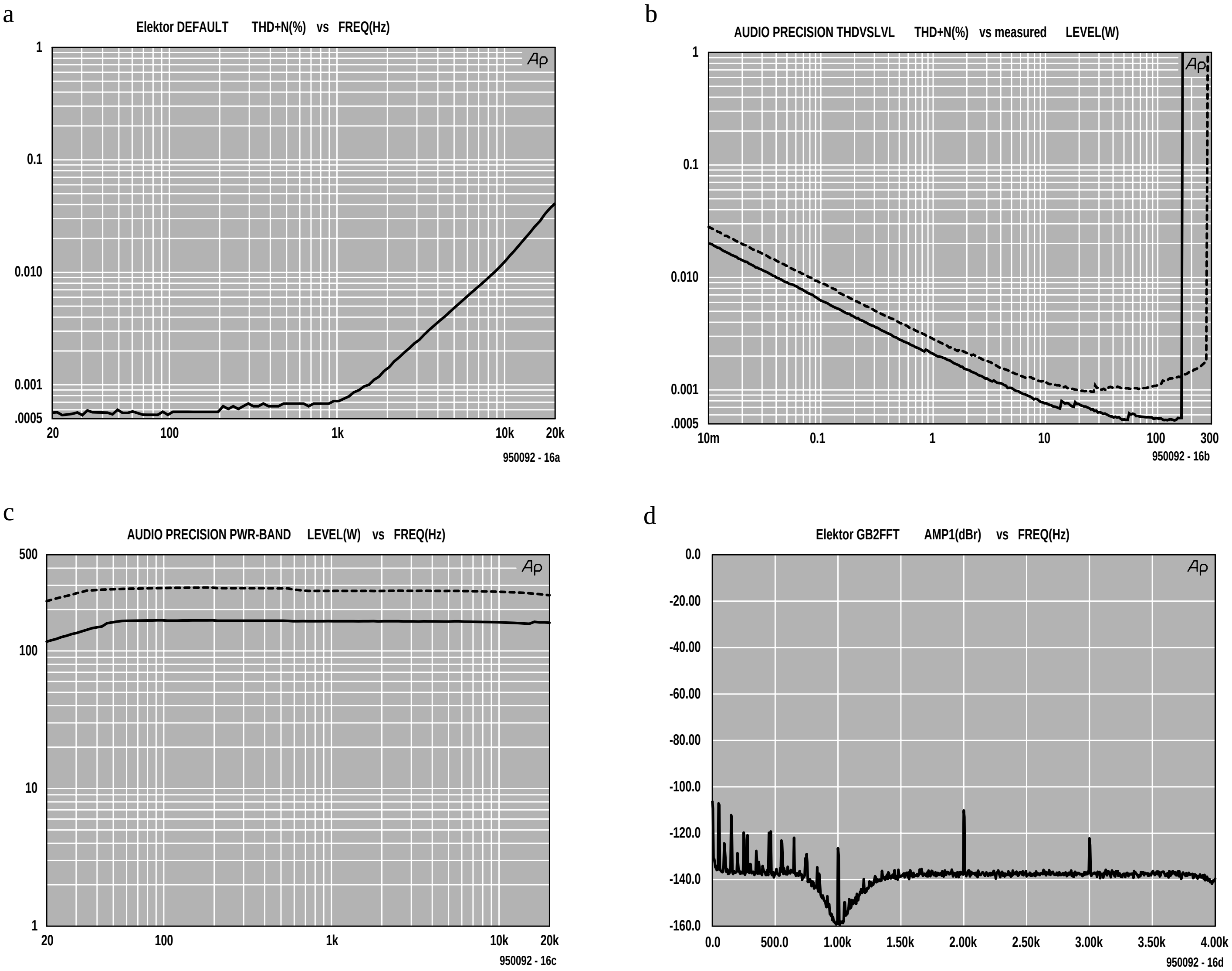


Fig. 5. Curves obtained during measurements on the amplifier with an Audio Precision Analyser (see text).

C_{13} = 680 nF
C_{14} = 470 pF, 160 V, polystyrene
C_{15}, C_{16} = 150 nF
C_{17} = 33 nF
C_{19} = 470 nF
C_{20} = 47 µF, 25 V, radial

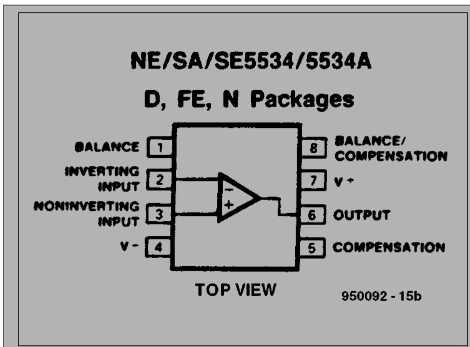
Semiconductors:
D_1, D_2, D_{10} = zener, 15 V, 1.5 W
D_3, D_6, D_{12} = 1N4004
D_7, D_8 = BY254
D_9 = 1N4148
D_{11} = 1N4002
D_{13}, D_{14} = LED
D_{15}, D_{16} = BAT85
T_1 = MJE350
T_2 = BD139
T_3 = MJE340
T_4 = BC546B
T_5 = BC556B
T_6 = MJE15030
T_7 = MJE15031
T_8, T_{14} = MJ15003
T_9, T_{15} = MJ15004
T_{10}, T_{12} = BC337
T_{13} = BC639

Integrated circuits:
IC_1 = NE5534
IC_2 = LM393

Miscellaneous:
L_1 = see text
Re_1 = 16 A, 24 V, 875 Ω relay (e.g. Siemens V23056-AO105-A101*)
F_1, F_2 = glass fuse, 6.3 A, slow complete with PCB type holder
Loudspeaker and mains connectors for board mounting (AMP - see text)
Mica washers for T_1–T_3, T_6–T_9, T_{14} and T_{15}
Rectangular bracket e.g. SWP40, 20 cm long (Fischer 40×30×5**)
Heat sink <0.4 K W^{-1}
PCB Order no. 950092
Mains transformer, 2×42.5 V, 625 VA (see text)
Fuse (power supply) 3.15 A, slow, $I^2t \geq 400$
Bridge rectifier 400 V, 35 A
4 off electrolytic capacitors, 10,000 µF, 100 V
PCB Order No. 924055

[950092]

* ElectroValue 01784 33603 or 0161 432 4945
** Dau 01243 553 031; trade only, but information on your nearest dealer will be given by telephone.



The NE5534

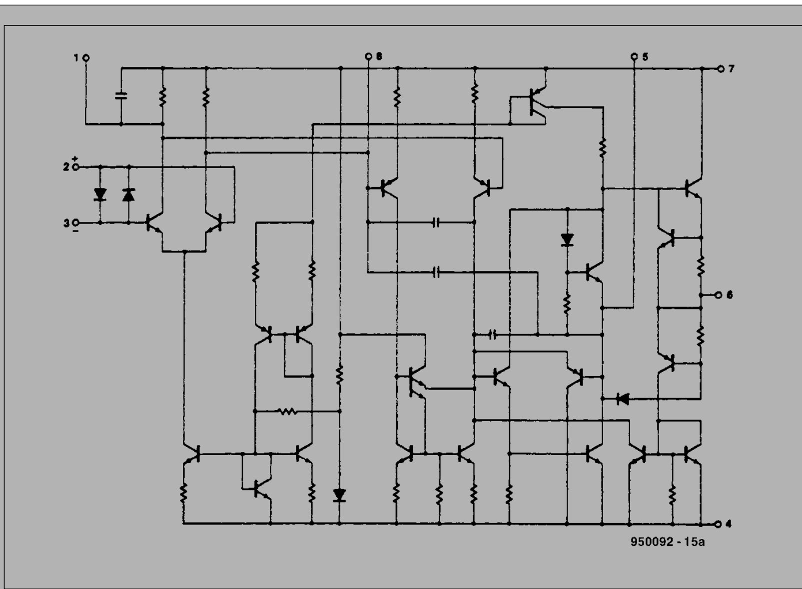
The NE5534 is a good quality, versatile, low-noise operational amplifier which is excellent value for money.

Compared with older types, it has better noise figures, small signal performance, power bandwidth, and output drive capability.

These characteristics make it ideally suited to high-end audio applications. It is found even in the most expensive CD players.

The adjacent simplified diagram gives an idea of the internal structure of this versatile device. It consists of a number of differential amplifiers that are set with the aid of current sources and current mirrors. Well-designed compensation circuits result in excellent linearity and very low distortion.

The standard design gives an amplification of ×3. The frequency response can be optimized for various applications with the aid of an external capacitor. It may be adjusted for a capacitive load, high slew rate, low overshoot or for application as a unity amplifier.



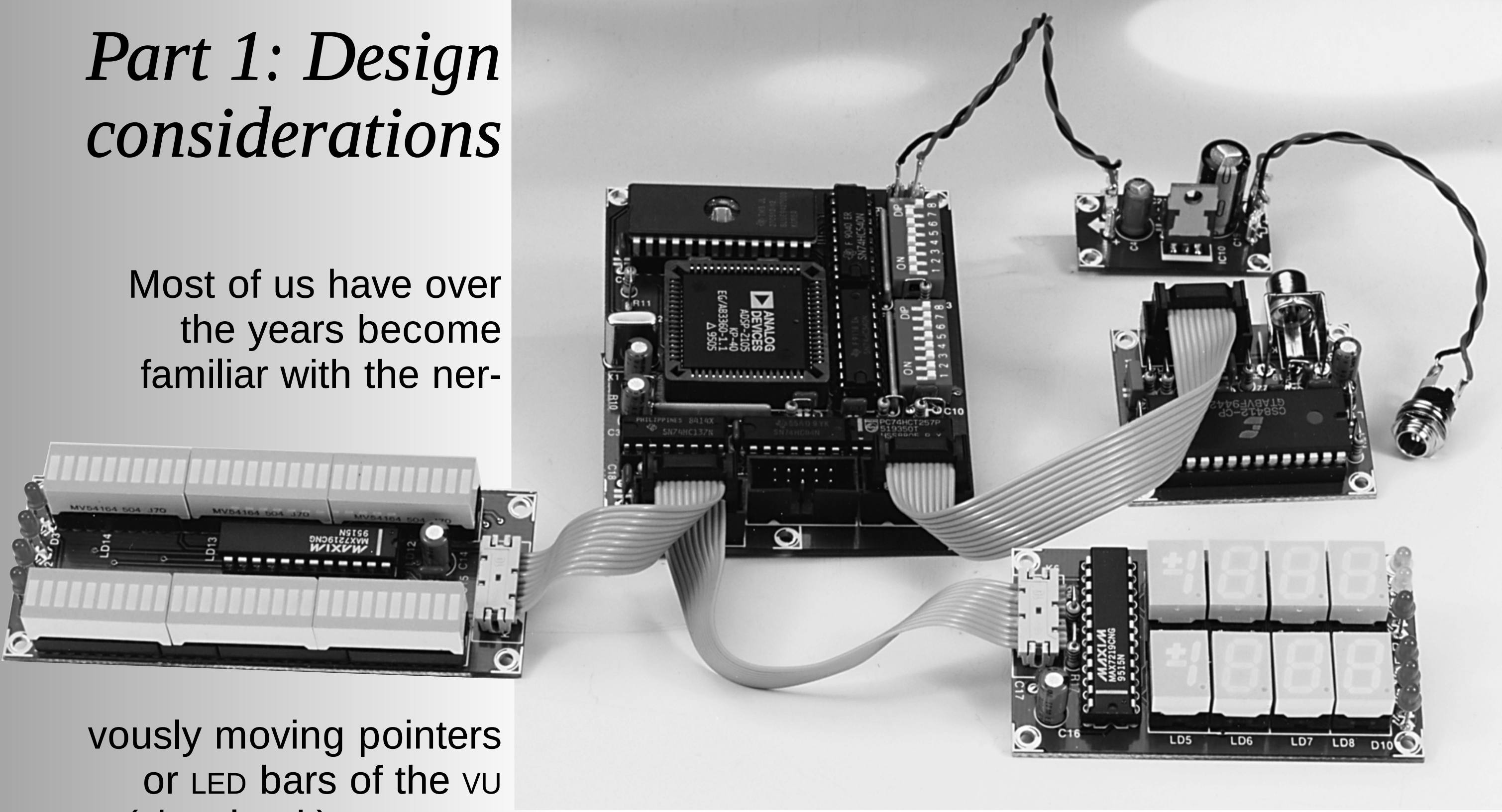
Some technical data

Small-signal bandwidth	10 MHz
Output voltage (at U_b = ±18 V)	10 V_{rms} across 600 Ω
Input noise	4 nV Hz^{-1}
DC voltage amplification	10^5
AC voltage amplification	6×10^3 at 10 kHz
Power bandwidth	200 kHz
Slew rate	13 V μs^{-1}
Supply voltage range	±3 V to ±20 V

Contents

digital VU meter

Part 1: Design considerations



Most of us have over the years become familiar with the nervously moving pointers or LED bars of the VU (visual unit) meter on the front panel of a cassette tape recorder or mixing panel that indicate the level of the a.f. signal. The circuit presented in this two-part article is a variant of this meter that can be inserted directly in series with the a.f. signal line. Its specifications are reminiscent of professional equipment. We are not entirely certain, but think that this is the first DIY VU meter ever published in a magazine.

Design by T. Giesberts

for direct measurements of digital audio signals

The introduction of digital audio (CD, DCC, DAT, MiniDisc) in the 1980s has drastically changed the world of audio and hi-fi. Many analogue circuits have been replaced by black boxes like digital filters and signal processors. The a.f. data has been changed from a series of waveforms to a train of binary digits (bits).

Brief technical data

Display	double alphanumeric 31/2-digit
	double 30-segment led bar with peak indication
Brightness	individually presettable
Measuring range	138.5 dB (with 24-bit input)
Resolution	0.1 dB (±0.005 dB)
Input	S/PDIF 16–24 bit; i2s 16–24 bit
Sampling frequencies	32 kHz, 44.1 kHz, 48 kHz
Measurement	peak, PPM, and RMS
Status indication	by 10 LEDS on front panel

The present VU meter is geared to the new technology. Where in earlier times a net-

** Sony/Philips Digital Interface Format – the consumer version of the AES/EBU standard. This standard was devised by the American Audio Engineering Society (AES) and the European Broadcasting Union (EBU) to define the signal format, electrical characteristics and connectors to be used for digital interfaces between professional audio products.*

work consisting of a capacitor, a resistor, a diode and a mini moving-coil meter was used for level indication, in modern equipment this network is replaced by a digital signal processor—DSP. This results in a rather more compact instrument that gives excellent performance.

The VU meter is based on a Type 2105 DSP from Analog Devices. This 16-bit device is designed and programmed to enable data to be processed with a 64-bit resolution. This means that 24-bit wide data are processed with an arithmetical error that, in the end result, is smaller than 0.025 per cent. The arithmetic is carried out fast and accurately. The speed of it is provided by an integral multiply accumulator (MAC). For example,the multiplication of two 16-bit numbers which must be retrieved from the memory and the adding of the result to an existing number or storage into a memory location takes rather less than 100 ns.

Since the VU meter is intended for measuring digital a.f. signals, it itself is designed on digital lines.

Also, the processing is controlled by software wherever possible, which obviates the use of special components (to keep any errors down). This arrangement also keeps the cost down and results in a compact, flexible meter.

MEASURING: WHAT AND HOW

The basic design of the meter is shown in the block diagram in **Figure 1**. A large part of the meter is taken up by the displays. Apart from the two 30-LED bars, there are also two 3½-digit-wide alphanumeric displays. Both display groups are controlled by a dedicated controller from Maxim, the Type MAX7219.

The LED bars give a a good visual indication of the signal level: they simulate the moving pointer of VU meters of yesteryear.

The alphanumeric display shows the peak level measured during a recording session.

There are also several LEDS that indicate which functions of the meter have been selected.

The input of the meter is formed by an S/PDIF* receiver connected to a multiplexer. Several inputs of the multiplexer are (as yet) unused, but are intended for connecting an analogue-to-digital (A/D) converter which we hope to publish in a future issue.

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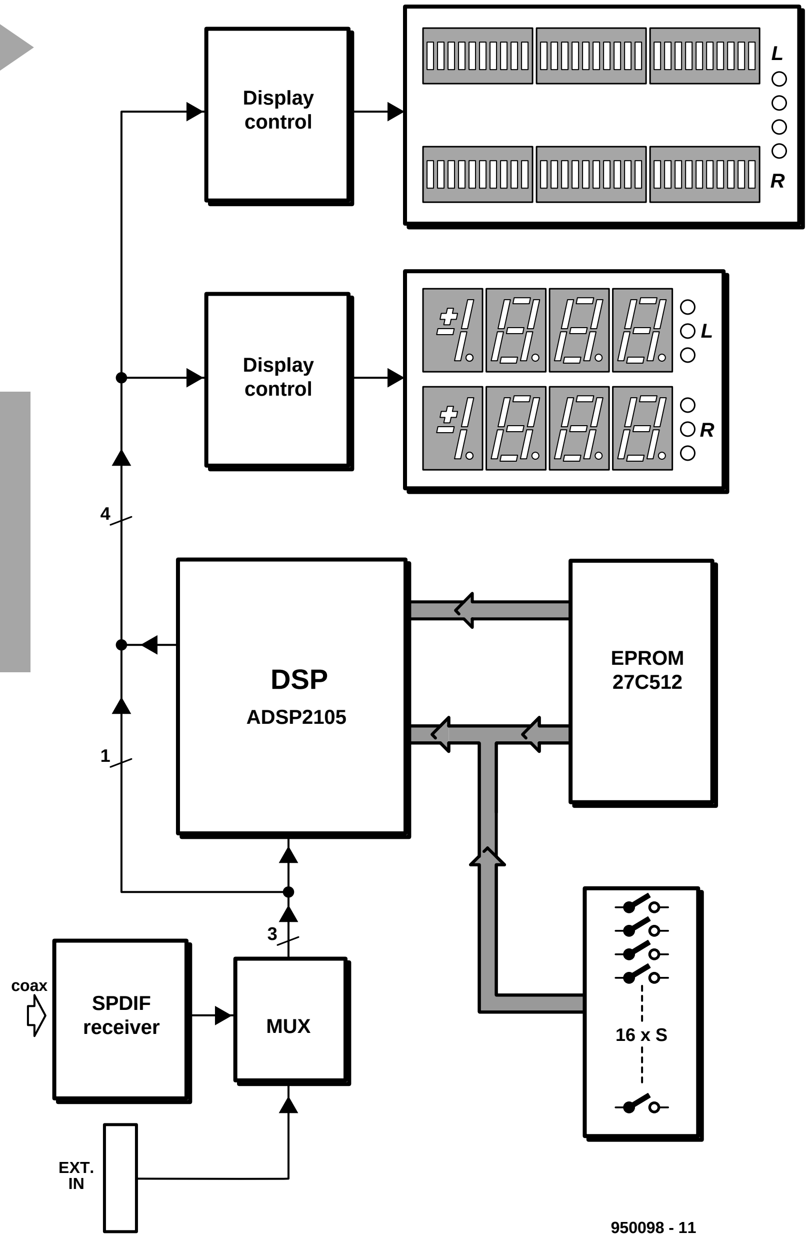


Fig. 1. Block diagram of the VU meter. The meter is based on a digital signal processor—DSP—which carries out all the arithmetic necessary for displaying the digital a.f. level.

Table 1. Positions of dip switches

1.1	*peak hold time*	*hold/update (1.95 s)*
1.2	*spot hold time*	*hold/update (1.3 s)*
1.3	*spot mode*	*peak/PPM*
1.4	*led bar mode*	*peak/PPM*
1.5	*mode*	*RMS/PPM or peak*
1.6	*scale*	*dBu/dBfs*
1.7	*0 dB ref. left*	*set*
1.8	*0 dB ref right*	*set*
2.1	*B0 current led bar*	*1/0*
2.2	*B1 current led bar*	*1/0*
2.3	*B2 current led bar*	*1/0*
2.4	*M0 current margin display*	*1/0*
2.5	*M1 current margin display*	*1/0*
2.6	*M2 current margin display*	*1/0*
2.7	*Input selection*	*S/PDIF/i2s*
2.8	*not used*	

The output signals of the multiplexer are applied to the digital signal processor. Note, by the way, that this DSP has been used in an earlier article in this magazine.

Since a DSP is designed for a specific application, that is, the fast processing of digitized analogue signals, it needs software to perform as needed. The software is stored in a Type 27512 EPROM. The RAM required by the dsp to work properly is integral: thus, there is no external ram.

The writing of the state of the dip switches is enabled by the addition of a 16-bit wide input gate.

The various modes and functions of the meter are selected with a switch on the front panel and some DIP switches on the board. They are summarized in **Table 1**. Note that they can be selected semi-permanently with the dip switches, because it is assumed that modes and functions are chosen only once. If more flexibility is required, for instance, if the meter is to be used as a laboratory measuring instrument, the DIP switches can be replaced by standard switches on the front panel.

Circuit description

The complete circuit of the VU meter is shown in the diagram in **Fig. 2**. It may be split into five parts: the processor, the S/PDIF receiver, the LED bars, the alphanumeric display and the power supply.

The power supply is a straightforward regulated design. Since digital signals are processed at high switching speeds, the supply decoupling has been given more than usual attention. This has resulted in a number of ICS being provided with their own decoupling capacitor.

The S/PDIF, a Type CS8412 from Crystal (IC_{11}), has no

2

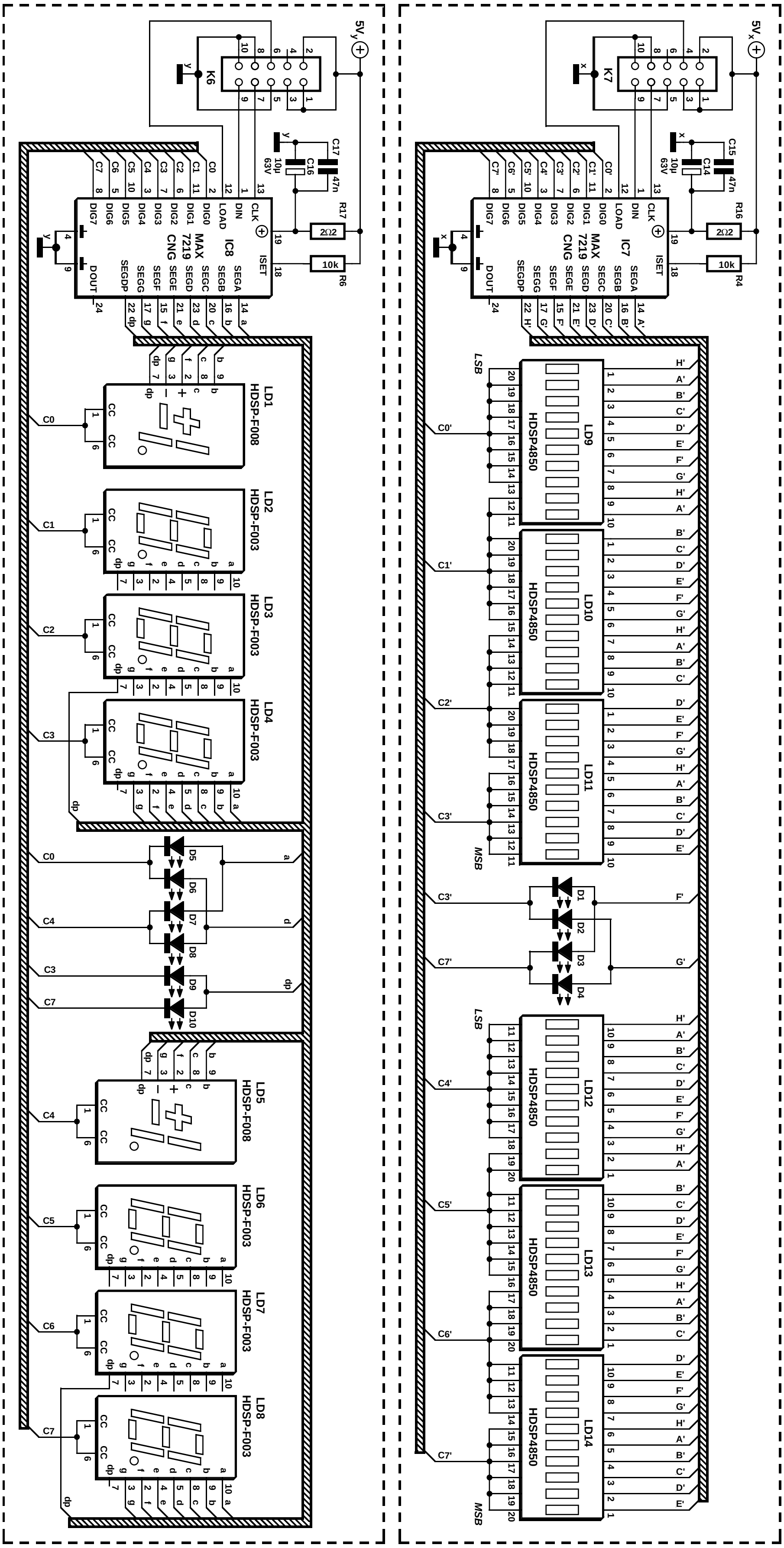
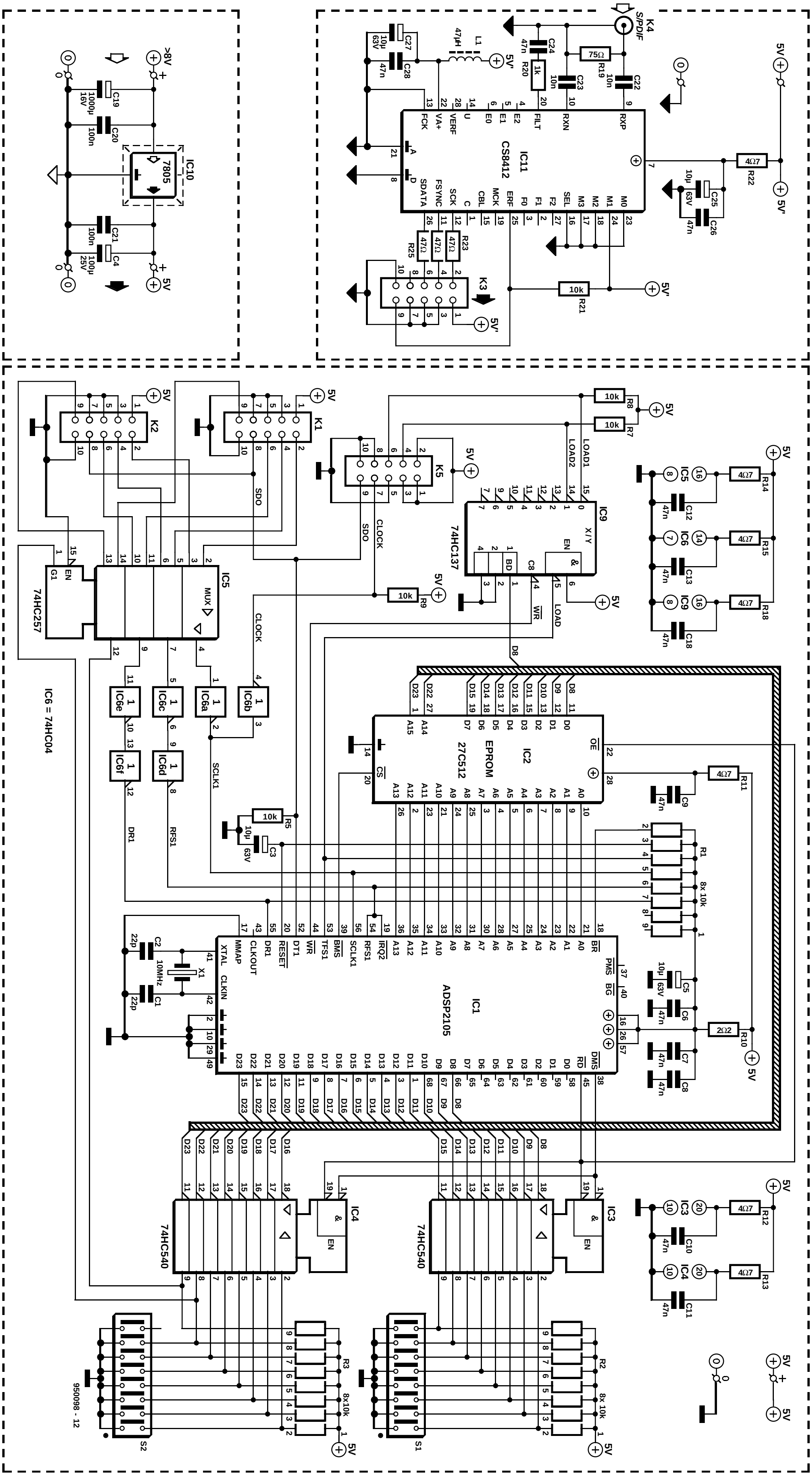


Fig. 2. The use of intelligent components keeps the circuit fairly small. Most of it consists of LEDS, displays and associated drivers.



surprises: it offers a good and ready integrated solution to the design requirement. It is discussed in more detail in the box on page 39. Note, however, that it has only a coaxial input; optical signals must be connected via an optical receiver, such as the TORX173 from Toshiba.

The digital a.f. data produced by the interface is output via K_3. This connector is linked to K_1 on the mother board via a short length of flatcable. Connectors K_1 and K_2 are arranged in an identical manner; K_2 is intended for a future extension, such as a modern analogue-to-digital converter—ADC.

The signals at K_1 are applied to the DSP, IC_1, via multiplexer IC_5.

The clock retrieved from the digital data is applied to the SLCK1 input. In this way, the digital signal determines the digital serial clock frequency.

The serial a.f. data derived from the S/PDIF signal are applied to input DR1.

Finally, the synchronization signal contained in the S/PDIF data is applied to input RFS1.

The circuits on the display boards operate in like manner. Display controllers IC_7 and IC_8, both Type 7219 from Maxim, communicate serially with the mother board. This data link uses signals Load, Clk and Data. To give the user maximum freedom in the building of the meter, the displays are linked to the mother board via a short length of flatcable (K_5–K_7).

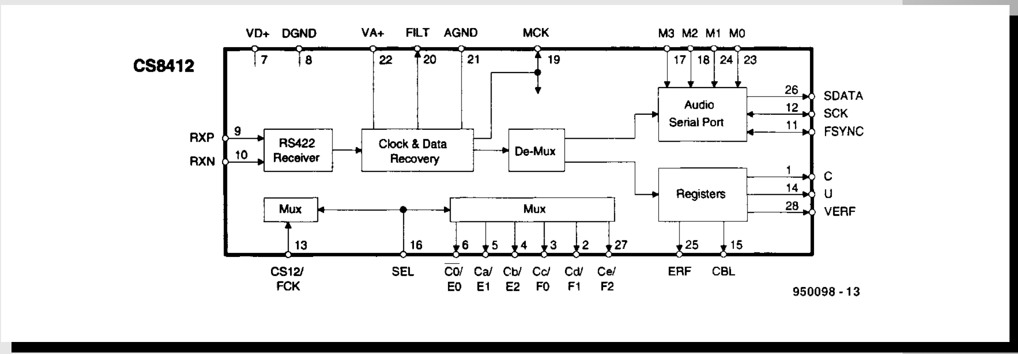
Connectors K_5–K_7 also carry the supply lines (+5 V and earth).

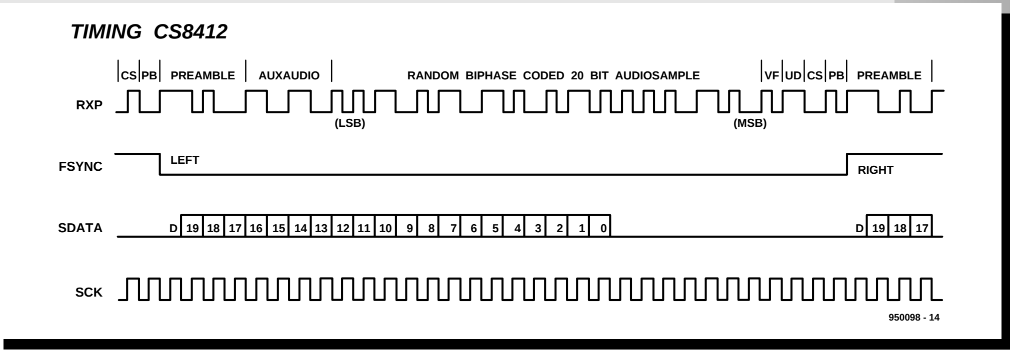
The brightness of the displays may be

S/PDIF signal decoder

Decoding digital a.f. signals is fairly straightforward with the use of a special ic, here the Type CS8412 from Crystal Semiconductor Corporation. It is a monolithic CMOS device that receives and decodes a.f. data according to the AES/EBU, IEC958, S/PDIF and EIAJ CP-340 interfaces standards. It receives the data from a transmission line, recovers the clock and synchronization signals, and demultiplexes the a.f. and digital data. The timing diagram shows how this is done.

The chip can accommodate ×256 oversampling since the clock is 256 times the sampling frequency.





altered as required by changing the values of resistors R_4 and R_6. Which of the displays is accessed depends on the level of the Load line. The Load signal is generated by the DSP and split into Load1 and Load2 by multiplexer IC_9.

After the LED controller has written a complete data word, it begins to control the display. One controller can control 64 LEDS or eight 7-segment displays. Since the present meter uses 3½-digit displays, some capacity is left to control LEDS from the display. The needed multiplex signals are generated by the display controller.

Writing the state of the DIP switches is effected by IC_3 and IC_4, both Type 74HC540 devices, which are used as 16-bit input. Since this gate is the only I/O hardware available to the DSP, the DMS (Data Memory Select) line can be used to select either of the ICS. The state of all switches is written in one go. The gate is linked to data lines D_8–D_{23}.

The system software is stored in IC_2. This EPROM is enabled by the BMS (Boot Memory Select) signal. It will be noted that only eight data lines are used, whereas internally a 24-bit wide bus is provided. Analog Devices has chosen a system whereby in the EPROM three bytes are sequential, which in the DSP are placed in parallel again. A slight disadvantage of this arrangement is that writing the boot software takes just a little longer. The benefit is, however, that the circuit is simpler and cheaper.

VARIOUS POSSIBILITIES

The VU meter is a flexible instrument: the desired function may be selected with DIP switches. A summary of these possibilities follows.

One LED of the LED bar is used to retain the peak level. The function of the peak indicator is set to hold or update with DIP switch 1.2. In the update mode, the measured value is adapted every 1.3 seconds.

Both the LED bar and the spot measurement may operate as required in the Peak Program Meter (PPM) mode or the Peak mode. The PPM mode is a standard used for the registration of the average a.f. level. This standard also specifies the attack and decay times. All this means that in practice the LED bar reacts very rapidly to the applied a.f. signal.

DIP switch 1.6 enables either of two scale units to be selected: dBfs (decibel full scale) or dBu, which is an analogue reference, in which 0 dB corresponds to 775 mV, that is, 1 mW into 600 Ω. This scale is intended for measurements of digital signals where the 0 dB level is determined by the largest figure that can be generated by a given number of digits. An externally connected A/D converter is usually set so that full-scale deflection occurs at a level of +12 dBu.

The 3½-digit display shows the peak value in dB, just as the spot of the LED bar. However, it shows the level in figures with an accuracy of 0.5 per cent (0.1 dB). To improve legibility, 'hold' or 'update' may be selected with DIP switch 1.1. In the update mode, the measured value is updated every 1.95 second.

The r.m.s. mode is selected with DIP switch 1.5. Once selected, this mode applies to both the LED bar and the alphanumeric display.

The standard with r.m.s. measurements is the 0 dB full scale. In the present meter, this can be set manually for either the left-hand or right-hand channel. This arrangement allows a dB measurement to be carried out from any random signal level.

Calibration is straightforward: apply a test signal to the input and briefly close left-hand r.m.s. switch 1.7 and right-hand r.m.s. switch 1.8. This sets the 0 dB reference to the level of the applied test signal.

Finally, an important aspect is setting the brightness of the display. The brightnesses of the LED bars and alphanumeric display can be set independently of each other: DIP switches 2.1–2.3 are for the LED bars and DIP switches 2.4–2.6 are for the alphanumeric displays. This arrangement makes it possible for the brightness to be adapted to the ambient brightness.

Apart from by these digital settings, the setting current, I_{set}, can be altered by changing the values of reference resistors R_4 and R_6.

This ends the discussion on the design and setting of the VU meter. Next month the practical aspects and application of the meter in an existing system will be discussed.

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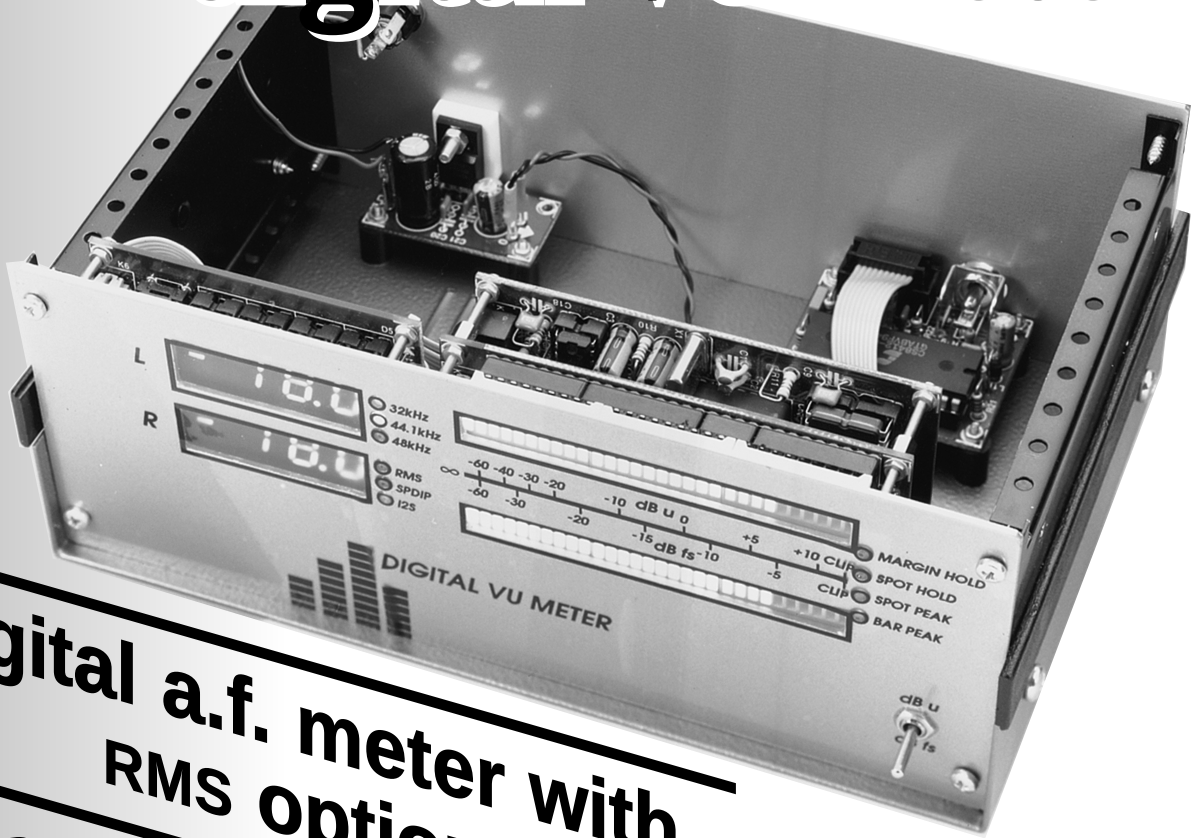
digital VU-meter



digital a.f. meter with RMS option

Part 2: Construction and Operation

After last month's detailed description of the design and facilities of the accurate, multifaceted digital vu meter, this month's final instalment deals with the construction and operation of the instrument.

design: H. Schaake/R. Smeding

As discussed in Part 1, the circuit consists of five functional blocks, which are readily recognized in the components layout of the double-sided through-plated printed-circuit board in **Figure 3**. The board is available ready made—see the Readers services list elsewhere in this issue. The board should be cut along the indicated lines into five discrete parts, so that each functional block gets its own board.

This arrangement makes the construction very flexible: not only can a variety of enclosures be used to house the meter, but matching the function to a specific application becomes straightforward.

Figure 3 forms an excellent guide for populating the five boards. That of the supply board is straightforward, although it should be noted that the voltage regulator needs cooling. The simplest way of achieving this is to screw the device to the rear panel of the enclosure if this is a metal one; if a plastic case is used, the regulator should be fitted to a heat sink.

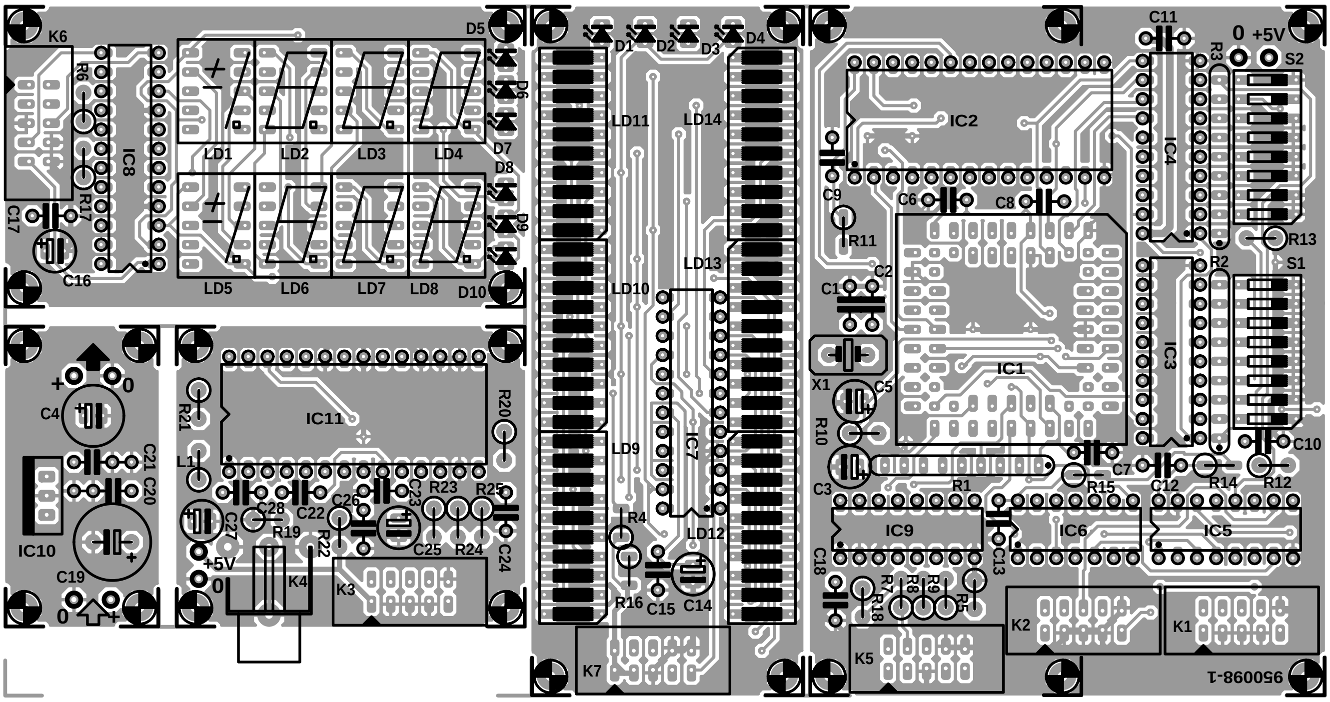
The S/PDIF interface is provided with a coaxial connector, K_4, to which the serial data is applied via a coaxial cable. It is advisable to use gold-plated connectors to ensure good contact at all times. Fit the device in a suitable socket to minimize the risk of damage through soldering. Mount all resistors upright. The present design does not include an optical input, but it may be seen from **Figure 4** that a relevant interface can be added without any difficulties. An added benefit of such an addition is that an extra coaxial output is provided as well.

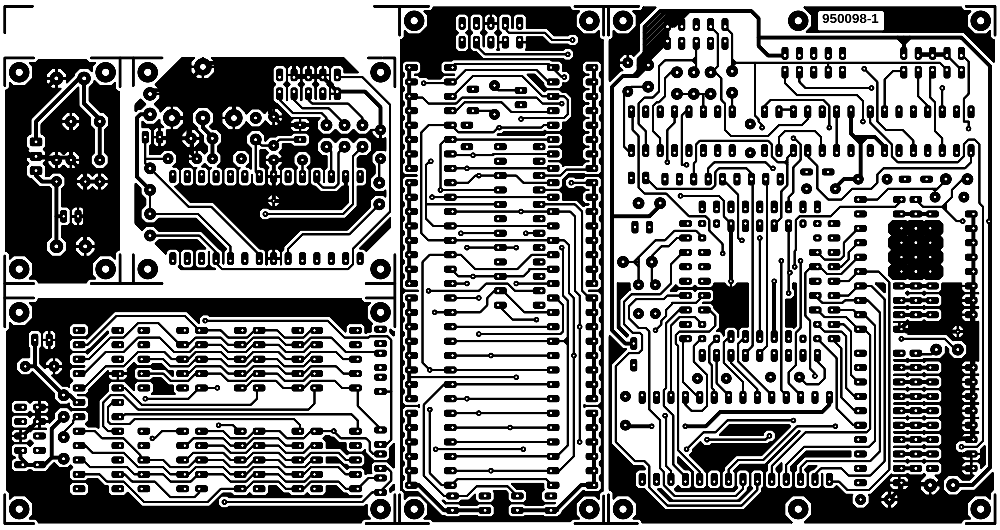
Populating the two display boards should not present any difficulties. Mount all LED displays and the Maxim controllers, IC_7 and IC_8, in suitable sockets. Connectors K_6 and K_7 are flatcable types for board mounting. Provided the polarity of the LEDs and the electrolytic capacitors is observed, nothing can really go wrong.

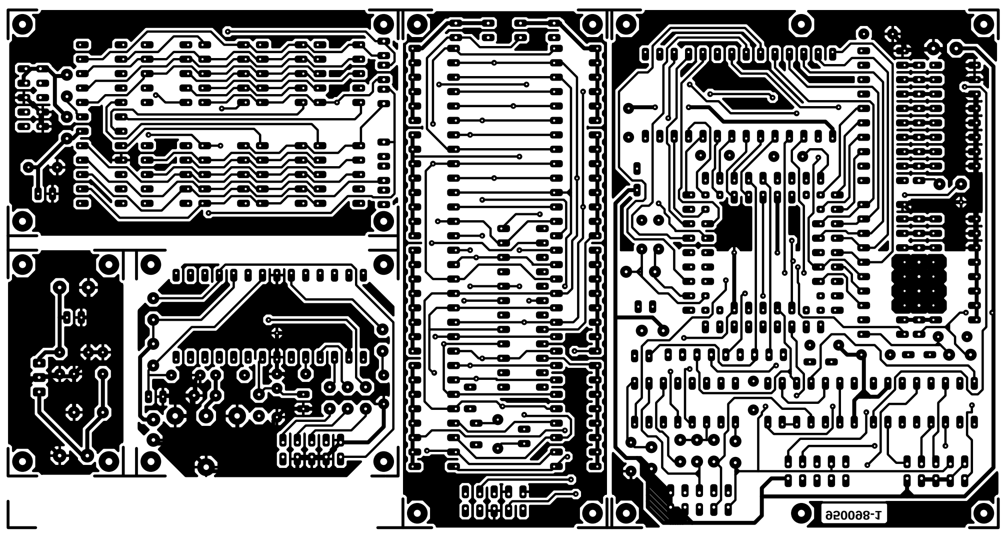
The central processing unit, IC_1, must be mounted in a suitable

Figure 3. The printed-circuit board for the digital VU meter.

3







Parts list

Resistors:
R_1–R_3 = 8×10 kΩ array
R_4–R_9, R_{21} = 10 kΩ
R_{10}, R_{16}, R_{17} = 2.2 Ω
R_{11}–R_{15}, R_{18}, R_{22} = 4.7 Ω
R_{19} = 75 Ω
R_{20} = 1 kΩ
R_{23}–R_{25} = 47 Ω

Capacitors:
C_1, C_2 = 22 pF
C_3, C_5, C_{14}, C_{16}, C_{25}, C_{27} = 10 μF, 63 V, radial
C_4 = 100 μF, 25 V, radial
C_6–C_{13}, C_{15}, C_{17}, C_{18}, C_{26}, C_{28} = 47 nF, ceramic
C_{19} = 1000 μF, 16 V, radial
C_{20}, C_{21} = 100 nF, ceramic
C_{22}, C_{23} = 10 nF, ceramic
C_{24} = 47 nF, MKT (metallized polyester), pitch 5 mm

Inductors:
L_1 = 47 μH

Semiconductors:
D_1–D_4, D_7–D_{10} = LED, 3 mm, red
D_5 = LED, 3 mm, green
D_6 = LED, 3 mm, yellow

Integrated circuits:
IC_1 = ADSP-2105 KP-40 (Analog Devices)
IC_2 = 27C512 (programmed: Order no. 946646-1)*
IC_3, IC_4 = 74HC540
IC_5 = 74HC257
IC_6 = 74HC04
IC_7, IC_8 = MAX7219CNG (Maxim)
IC_9 = 74HC137
IC_{10} = 7805
IC_{11} = CS8412CP (Crystal)

Miscellaneous:
K_1–K_3, K_5 = 10-way header
K_4 = audio connector for board mounting
K_6, K_7 = 10-way flatcable connector for board mounting
S_1, S_2 = 8-pole DIP switch
X_1 = crystal, 10 MHz
LD_1, LD_5 = HDSP-F008 (Hewlett Packard)
LD_2–LD_4, LD_6–LD_8 = HDSP-F003 (Hewlett Packard)
LD_9–LD_{14} = HDSP-4820 (red) or HDSP-4850 (green) from Hewlett Packard; or MV54164 (green) or MV57614 (red) from Quality Technology
Enclosure 80×200×132 mm (3³⁄₁₆×7¹⁵⁄₁₆×5¼ in) (h×w×d); e.g. Telet (Italy) Type LC750
PCB Order no. 950098*

* The PCB and EPROM are available as a package Order no. 950098-C

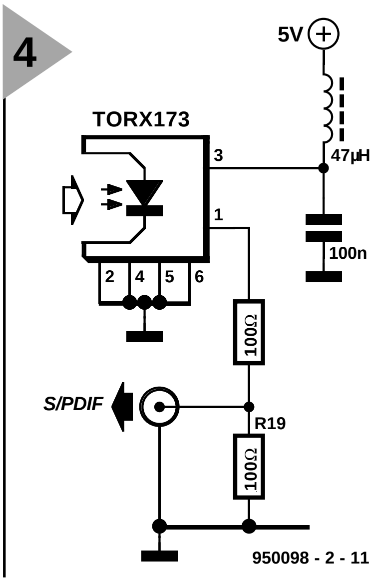


Figure 4. The meter may be provided in a simple way with an optical input and a coaxial output.

socket. Pay good attention to the position of pin 1, because once the socket has been fitted, it is very difficult to change its position. The IC fits into the socket only one way, so nothing can go amiss there. Fitting the remainder of the components and parts should not present any undue difficulties.

Set the various DIP switches as required—see Table 1 in Part 1. In case of doubt, the following is a good default configuration: S1.3, S1.4, S2.1–S2.6 all 'on' and the remainder of the switches 'off'.

Figure 5 shows how the five boards are interconnected, but do not make any connections yet, because each of the boards has to be tested independently.

Initial switch-on

When the boards are completed and a visual inspection has not thrown up any errors, each of them should be tested, starting with the supply board. Connect a 9–12 V mains adaptor to this board. Its output should then carry a stable potential of 5 V. When this requirement has been met, connect the board to the mother board.

To this end, some cables need to be made up: (1) a flatcable terminated at one end into a 10-pin crimp-on connector and at the other end into two board connectors, and (2) a ribbon cable terminated at each end into a crimp-on connector. The length of these cable depends on the type of enclosure used for housing the meter.

The first flatcable interconnects K_5, K_6 and K_7, and the second links the S/PDIF interface to the mother board. If use is made of the possibility of mounting the LED bar directly above the mother board (for which suitable holes have been provided in the board), the distance between K_5 and K_7 is ≤ 2 cm. Keep the relevant cable as short as possible to minimize the spurious radiation it emanates.

After the supply voltage has been switched on, the centre segment of the alphanumeric displays should light. If they do, it is virtually certain that the mother board functions correctly.

Connect the *digital* output of a sound source, for instance a DCC recorder, DAT recorder, or CD player to the input of the meter via a coaxial cable. When the first sounds emanate from the audio system, the LEDs will indicate the (relative) sound level. If there is no free digital output available on the recorder or CD player, use a digital output on the audio amplifier or digital-to-analogue converter (DAC) to which the recorder or player is connected. If all that is not possible, build the 'Splitter for S/PDIF coax/optical output' (*Elektor Electronics*, July/August 1995, page 78). An alternative is building the circuit in Figure 4, which gives the meter an additional output.

Faultfinding

In the unlikely event that the meter does not work (correctly) when it is first switched on, the circuit may actually help in faultfinding.

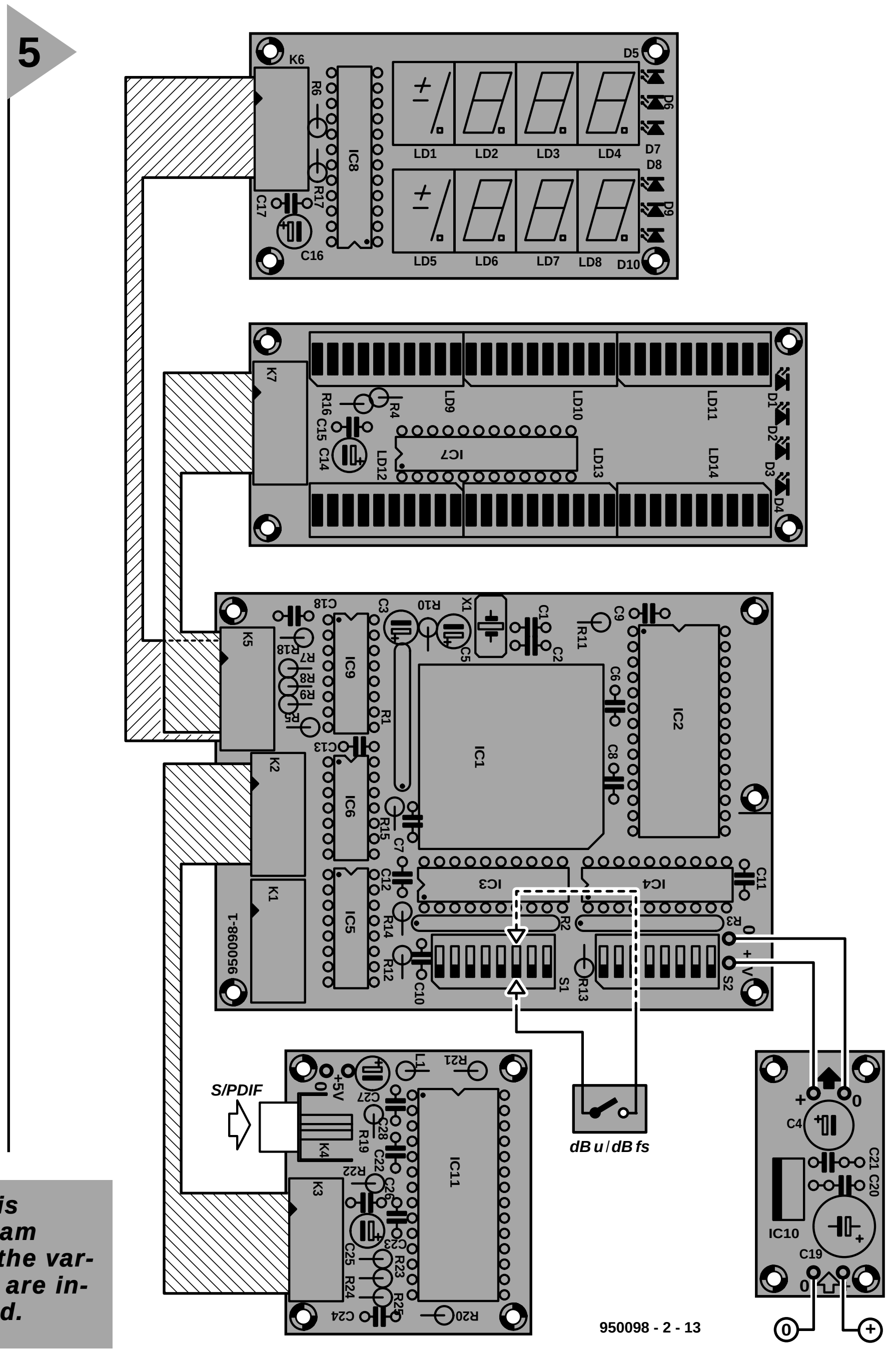


Figure 5. This wiring diagram shows how the various boards are interconnected.

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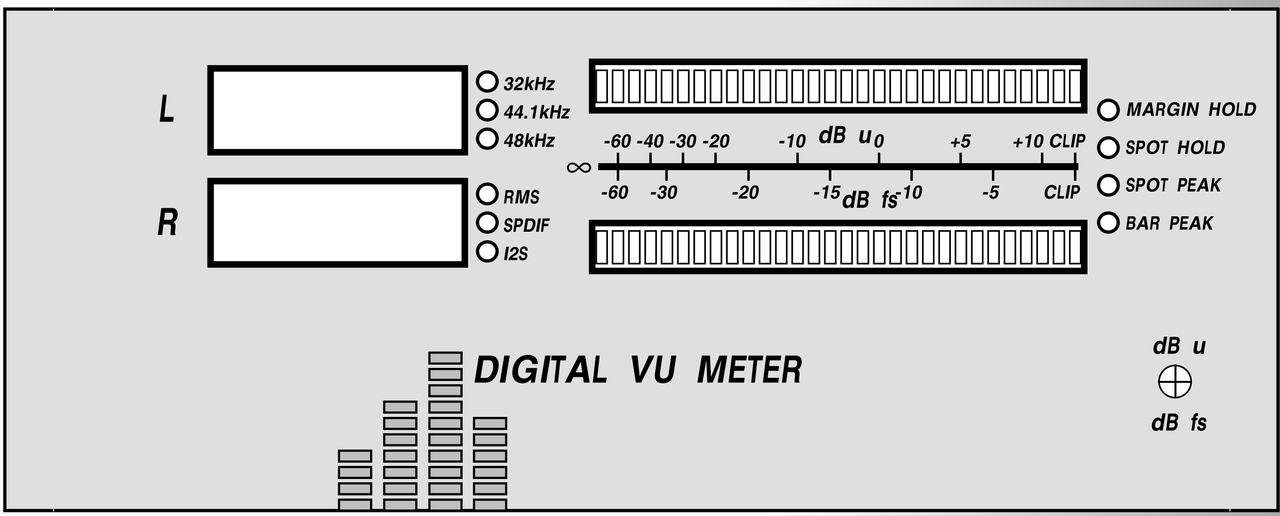

Figure 6. This illustration is a good basis for a design-it-yourself front panel for the meter.

First, check the supply voltage across C_4; if this is 5 V, the power supply is working correctly. In that case, check that the supplies to all ICs are all right: measure these carefully at the IC pins, not at the sockets. If they are all right, the S/PDIF decoder needs to be checked.

It will be remembered that the decoder retrieves the audio information from the digital data stream and also generates the clock that controls the displays. With the supply switched on and a digital S/PDIF audio signal applied to K_4, the clock must be present at pin 12 of IC_{11}. At the same time, the digital data should appear at pin 26, and the synchronization signal at pin 11. The clock should also be present at pin 13 of display controller IC_7 and IC_8.

If up to this stage everything works correctly, the operation of the DSP should be checked with an oscilloscope, starting with the oscillator. The operation of the IC can be checked only if there is a clock signal. Immediately upon power-on, the DSP reads the content of the boot ROM, IC_2. Short-circuit R_3 briefly and check whether within a few milliseconds of the short-circuit being removed there is activity on the address bus and data bus. If there is not, IC_1 is almost certainly defect and must be replaced. After the program has been written into the IC, there must be activity on lines Load (pin 53) and SDO (pin 52).

When the circuit works satisfactorily, it should be built into a suitable enclosure. If this has the same dimensions as that of the prototype, a front panel as shown in **Figure 6** can be used: a ready-made foil for this available—see the Readers Services column elsewhere in this issue.

Using the meter

The meter may be used in a specific audio system or in a digital audio workroom. Because of its digital design, it does not need to be calibrated. **Figure 7** summarizes the (relative) audio level associated with a lighted LED in the bar. An optimum setting for a specific application may be achieved with the DIP switches. **Table 2** shows how the brightness of the displays can be adapted with these switches.

(960098-2)

Table 2. Brightness of LED displays w.r.t. maximum brightness for various settings of DIP switches

Bar-display			Brightness (relative) %
B2	B1	B0	
0	0	0	56.25
0	0	1	62.50
0	1	0	68.75
0	1	1	75.00
1	0	0	81.25
1	0	1	87.50
1	1	0	93.75
1	1	1	100
Margin-display			
M2	M1	M0	
0	0	0	56.25
0	0	1	62.50
0	1	0	68.75
0	1	1	75.00
1	0	0	81.25
1	0	1	87.50
1	1	0	93.75
1	1	1	100

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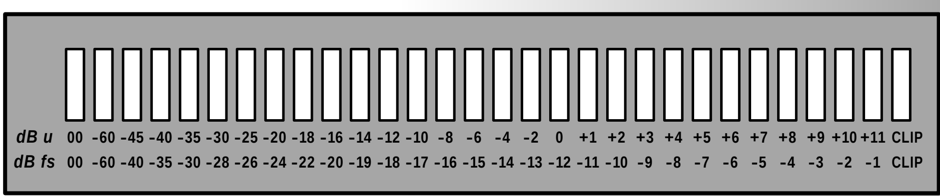

Figure 7. Each LED of the bar display represents a (relative) audio signal level. The scale illustrated shows which level is associated with which LED.

Surround speakers

Visaton minis for surround sound

To convert a standard stereo installation into a surround sound setup, at least three additional speakers are needed, of which the centre one must be magnetically shielded. The miniature speakers described in this article have been designed solely for this purpose by the German loudspeaker specialist Visaton.

The surround-sound decoder published last year proved very popular with readers from all over the world. Many of them wrote, however, to ask about suitable loudspeakers.

Since many of these readers expressed a wish to build the boxes themselves, it was a godsend when the German firm of Visaton offered a set of miniature loudspeakers for review. These speakers have been designed especially for surround-sound applications and are available as kits at very reasonable prices.

The centre loudspeaker, which should be placed immediately underneath the television set, uses magnetically shielded drive units. This is a must, otherwise the TV sound and vision will be adversely affected (and how!). Shielded drive units for home construction are still few and far between.

The back speakers have a novel facility. From their own tests, Visaton knew that the directional sensitivity of these speakers had an important effect on the optimum location of the listener. The pair discussed in this article beam upwards, which gives them a more diffuse character, and this in turn enlarges the listening area of optimum surround effect. A simple, but effective solution.

CINEMA-LIKE SOUND WITH LIMITED BANDWIDTH

A complete surround-sound installation consists of the power amplifier(s), surround-sound decoder, the standard loudspeakers for the left-hand and right-hand channels, a centre speaker that is placed between these two, and two back speakers that provide the spatial information. The standard speakers must be of hi-fi quality, since they largely determine the overall sound quality; they are normally driven by the extant a.f. amplifier.

The centre channel is used primarily for speech and its information consists mainly of the sum of the left-hand and right-hand signals from which the low frequencies have been filtered. This is why the frequency range of this speaker need not extend into the very low frequencies. In other words, neither the box nor the drive units need to be large.

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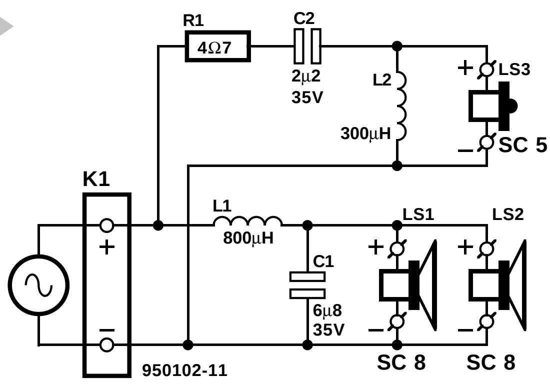

Figure 1. Circuit diagram of the cross-over filter for the Center 80. The filter has a cross-over frequency of about 5 kHz.

The frequency range of the back speakers needs to be only about 100 Hz to 7 kHz, since this is the range of decoder output. Also, the sound level from these speakers is relatively low compared with that from the other three. This means that the drive units for these speakers can be small, good-quality wide-band types.

The power rating of the back speakers need not be high either, primarily because they are not required to reproduce low frequency signals. A rating of 20 W is sufficient in almost all cases.

Center 80

The central speaker uses two 80 mm wideband drive units Type SC8 and a 10 mm tweeter Type SC5. These units are, as stated earlier, magnetically shielded. The associated filter, whose circuit diagram is shown in **Figure 1**, has a cross-over frequency of 5 kHz and roll-offs of 12 dB per octave.

Because of the shielding, the speaker

may be placed in close proximity to a television receiver or computer monitor.

The two 80 mm drive units are located in the bottom half of the enclosure with the tweeter above them. In many other speakers, the tweeter is placed between the two wideband units (the so-called d'Appolito configuration), but this has the drawback that the radiation pattern of frequencies around the cross-over frequency varies appreciably in the vertical direction (assuming that the speaker is upright). Normally, this does not matter much, but since the speaker in surround-sound applications is frequently used lying down, it would mean that sound reproduction varies when the listener moves his/her head slightly to the right or left and this is, of course, not the idea. With the present configuration this effect is virtually non-existent, so that the sound remains homogeneous outside the listening axis. The performance of the Center 80 can be assessed from the frequency characteristic in **Figure 2**. Note that the slight hump at 150 Hz ensures that the speaker, in spite of its modest dimensions, produces a voluminous sound.

Effekt 80

The effect described in the previous paragraph fortunately does not occur in the back speakers, since these use only a single 80 mm drive unit Type FRS8. In spite of their compactness, the speakers produce a an excellent spatial sound effect. As mentioned earlier, they radiate the sound upwards. This produces good scattering of the sound, and obviates the hot spot so often encountered with other surround-sound systems. (A hot spot is a single location in a room where the sound appears concentrated, although it should, of course, be evenly distributed).

The performance of the speakers was measured in a practical setup: the frequency response at a distance of 1 m from each of them, suspended from a wall at ear-height, is shown in **Figure 3**. The roll-off at higher frequencies is caused by the fact that only reflections are measured there. The 'normal' frequency curve, measured with the speaker lying down and radiating into the direction of the test microphone, is shown in **Figure 4**.

Enclosures

The enclosures of all three speakers are very easily constructed. Each consists of six rectangular pieces of medium density chipboard, which many DIY retailers will saw to size for you. The boards are glued together with the aid of suitable clamps. The construction diagrams are given in **Figure 5**. The drive units may be protected by grilles or covers.

Apart from those of the apertures for the drive units, taking into account the grilles or covers, the dimensions of the enclosures are not sacrosanct.

The rear of each enclosure should have holes for the ter-

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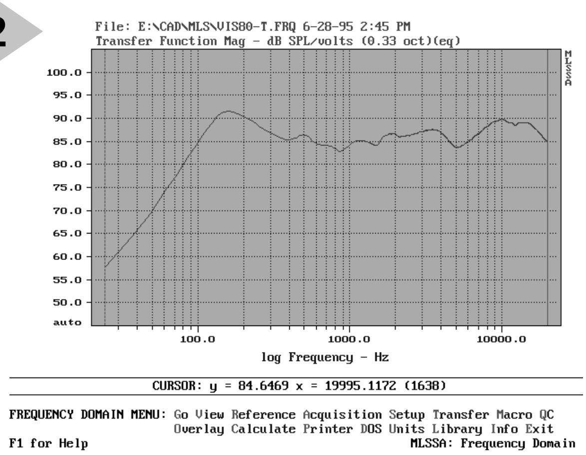

Figure 2. Frequency response of the Center 80. The slight hump at about 150 Hz ensures a voluminous sound reproduction.

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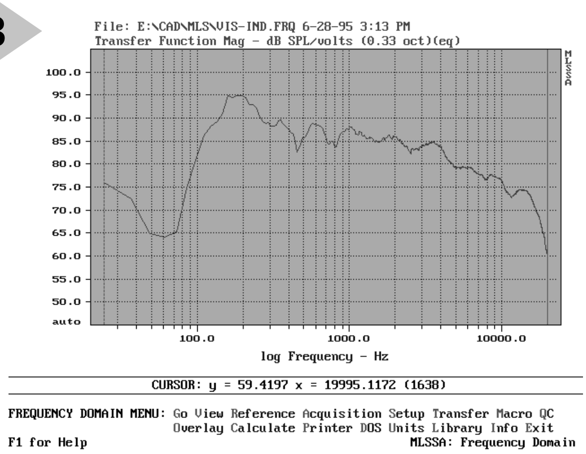

Figure 3. Frequency characteristic of the Effekt 80 when it is hanging from a wall at ear-height with the drive unit pointing upward.

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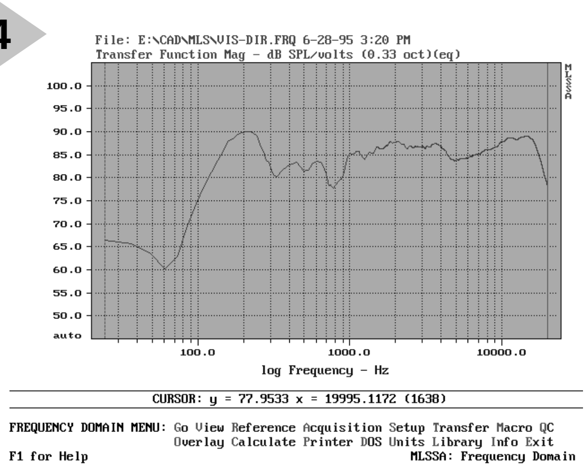


Figure 4. Frequency characteristic of the Effekt 80 with the drive unit radiating in the direction of the test microphone.

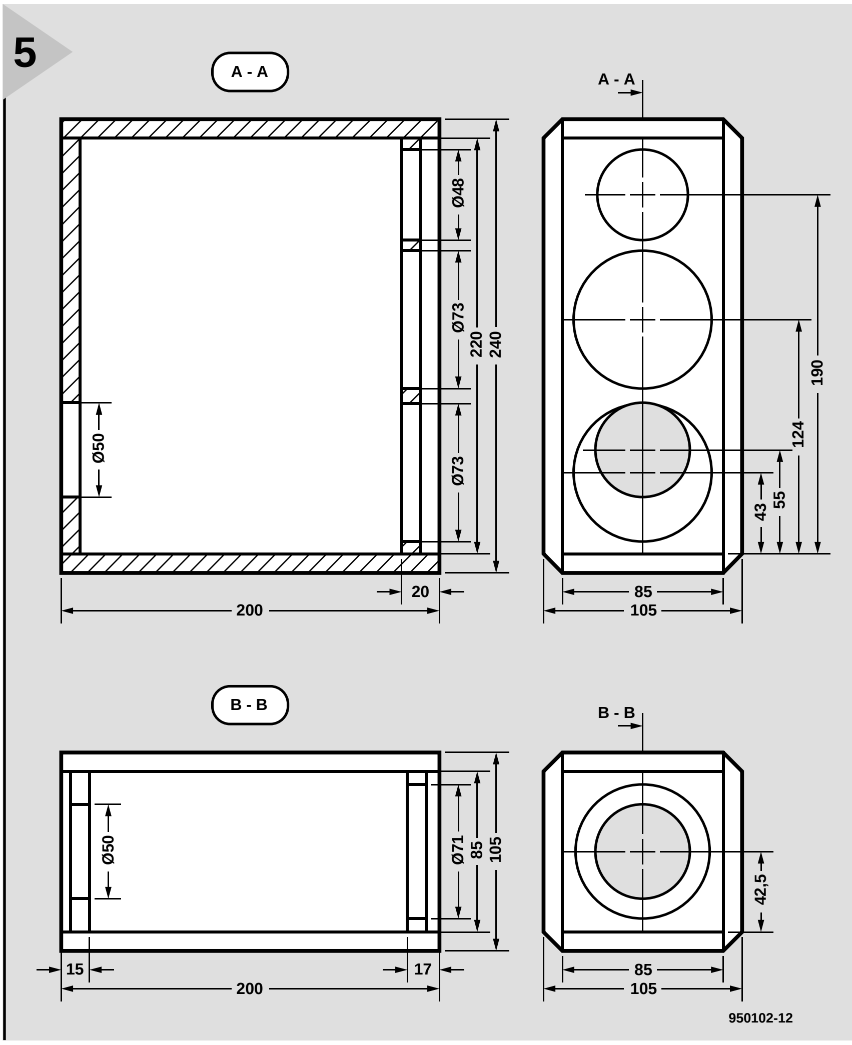


Figure 5. Construction diagram of the enclosures for the Center 80 and the Effekt

minals. Some constructors may find it more convenient to cut these holes before the enclosure is glued together.

Make sure to keep the cross-over filter as compact as possible, because there is not much room in the central speaker box.

A 15 mm hole may be cut in one of the sides of the back speakers. This hole is covered at the inside of the enclosure by a small piece of chipboard glued in place. This provides a convenient way of hanging the speakers from a wall (with the drive unit pointing to the ceiling).

The enclosures may be finished to individual taste.

When all that is done, the boxes can be wired up and the terminals and drive units screwed into place. Each enclosure should then be filled with a suitable damping material, such as polyester wool. [950102]

PARTS LIST

CENTER **80**
Cross-over filter:
R_1 = 4.7 Ω, 5 W
C_1 = 6.8 μF, 35 V, bipolar or polystyrene
C_2 = 2.2 μF, 35 V, bipolar or polystyrene
L_1 = 0.3 mH air-cored inductor wound from 0.6 mm dia. enamelled copper wire
L_2 = 0.8 mH air-cored inductor, wound from 1 mm dia. enamelled copper wire.

Drive units:
Ls_1, Ls_2 = 8 Ω magnetically shielded wideband unit Type SC8 (Visaton)
Ls_3 = 8 Ω magnetically shielded polycarbonate dome tweeter Type SC5 (Visaton)

Chipboard
(10 mm thick medium density)

2 sheets 85×220 mm (fore and aft)
2 sheets 200×240 mm (sides)
2 sheets 85×200 mm (top and bottom)
Damping material
Front grille for Center 80 (Visaton)
Loudspeaker terminal

EFFECT 80 (two required)
Drive unit
Ls_1 = 8 Ω full-range driver Type FRS8 (Visaton)

Chipboard
(100 mm thick medium density)

2 sheets 105×200 mm (sides)
2 sheets 85×200 mm (sides)
2 sheets 85×85 mm (top and bottom)
Damping material
Front grille for Effect 80 (Visaton)
Loudspeaker terminal

VISATON

Although Visaton is located in Germany, the firm is an associate member of the American Audio Engineering Society (AES).

The company has recently published a new colour catalogue, which is brimful of drive units and all sorts of accessory. All parts for the speakers described in this article are available from them: drive units, all components for the cross-over filter (or fully constructed filter board), grilles and covers. Complete construction kits are also available, as are fully constructed speakers in beautifully finished enclosures.
The address is
VISATON
Peter Schukat
Postfach 1652
D–42760 Haan
Germany
Telephone 0049 2129 552-0
Fax 0049 2129 55210

Contents

COPYBIT INVERTER

WARNING. The information in this article is intended solely for the recording, processing and copying of private musical work. The Editor and Publishers disclaim all responsibility for any use that may infringe any copyright vested in commercial compact discs and (digital) tape cassettes.

The copybit eliminator published in the February 1994 issue of this magazine has two drawbacks. The first of these is that it cannot be used without modifying the digital audio equipment. The second is clear from the revisit to the eliminator in the September 1995 issue: from time to time, the eliminator needs updating – it is not 'future-proof'. The copy-permit converter described in this article does not have these drawbacks

Design by W. Foede

Like the copybit eliminator published in this magazine in the February 1994 and September 1995 issues, the copybit inverter is an inexpensive and simple-to-build circuit for inverting the copybit in a digital S/PDIF* audio signal to enable users to copy (digitally) their own musical work many times without degradation by the SCMS**.

The inverter can be included in the S/PDIF link between any digital a.f. signal source (such as a DCC recorder, a CD player, a DSR receiver) and a (second) DCC recorder without the need of opening or modifying any of the equipment. During the copying, any copybit is inverted and at the same time the category code is altered. This means that the S/PDIF signal so modified is accepted by the recorder as if it comes from a CD player (so that unlimited copying of the signal becomes possible). The inverter also offers a number of other facilities, such as S/PDIF detection.

Category code

The coding of the S/PDIF has been described in detail in the article on the 'Copybit eliminator'. The following description is therefore limited to the most important aspects of it.

In all domestic audio equipment, the format uses sample frequencies of 32 kHz, 44.1 kHz and 48 kHz. These data contain, among others, information as to the copybit. The format reacts to the content of the S/PDIF. The SCMS, which inhibits multiple (digital) copying of the source signal, can be bypassed as shown in the flow diagram in Fig. 1. It is not sufficient to invert only copybit 1. As the diagram shows, when copying with category code 00 000 000 (general) takes place, for instance, the copybit is not sampled. This means that the recording has to be passed through a copybit eliminator a second time. It is, therefore, much safer to set the copybit to (or hold it at) 1 and assign to it the category code of an apparatus whose copybit is always sampled.

The present inverter always outputs the category code of a DAT or a CD, depending on the input signal. The code changes automatically, so that the subdata in the USER channel of the relevant equipment are retained.

The unit can be used as a converter with either optical or coaxial inputs and outputs without any change in the input signal.

The left-hand and right-hand channels

Brief specification

- Opening or modification of the audio equipment not required
- 'Future-proof', since it is independent of the category code
- Unlimited (digital) copying of source material
- Optical or coaxial inputs and outputs as required
- S/PDIF detector
- Indication of the position of the copybit and automatic setting to 1 (= digital copying permitted)
- Indication of the category code and automatic setting to 10 000 000 (CD) or 11 000 000 (DAT)
- AES/EBU format can be copied
- Transparent USER-Subcode channel
- Indication as to whether operation is as converter or as inverter
- Minimal number of components
- Non-critical setting up without test instruments

* Sony/Philips Digital Interface Format – the consumer version of the AES/EBU standard. This standard was devised by the American Audio Engineering Society and the European Broadcasting Union to define the signal format, electrical characteristics and connectors to be used for digital interfaces between professional audio products.
** Serial Copy Management System.

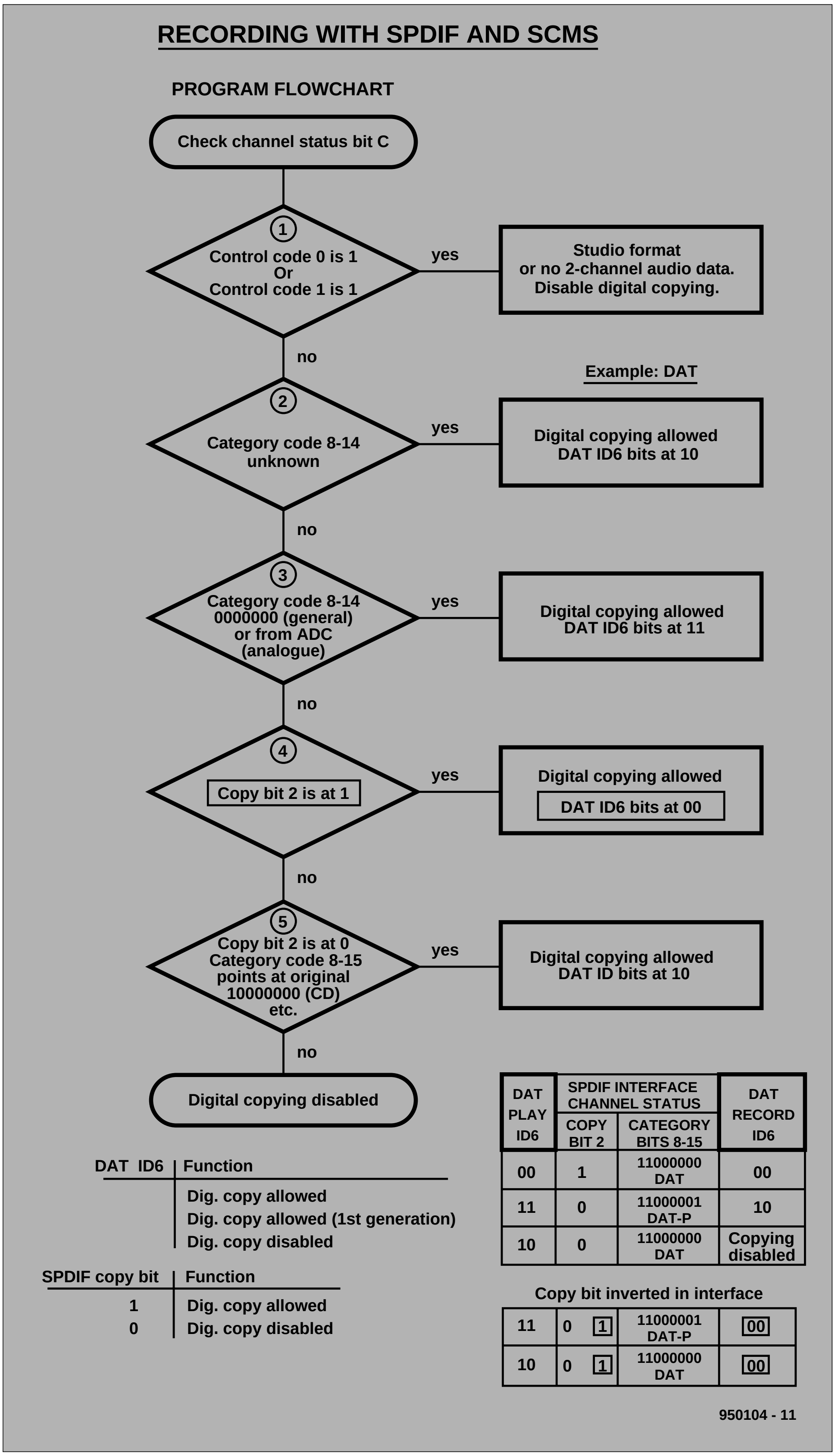


Fig. 1. Flow diagram of the evaluation process of the copybit in an S/PDIF signal.

each build a subframe of 32 bits, which together form a frame of 64 bits with a sampling frequency of, say, 44.1 kHz. A block contains a total of 192 frames (or 384 subframes) as shown in Fig. 2. The data are transmitted in biphase mark code, in which a bit is split into two bit-cells. In case of a digital 0, both cells have the same level, that is, 00 or 11. In case of a logic 1, the levels of the cells are inequal, that is, 01 or 10, which means that there is a level change at the centre of the bit. The longest a level can last is thus 1 bit—see Fig. 3.

This process also means that the clock is included in the transmitted information.

So as to identify the subframes and the start of the block, eight bit-cells have a bit sequence that does not occur in the biphase code—see Fig. 4. They are the block preamble *B*, which also identifies the left-hand channel in subframe 0, subframe preamble *M* (left-hand channel) and subframe preamble *W* (right-hand channel).

For the inversion of the SCMS, only bit 30, the channel status bit *C* of each subframe, is of importance. A complete channel status is repeated in each block of 384 subframes. The assignment of copy and category is the same in both channels.

The copybit is contained in subframes 4 and 5, and the category code in subframes 16–31.

Bits 30 and 31 in frame 15 have a special meaning. The so-called generation bit indicates whether the signal is an original or a copy. The assignment of a level for the original signal depends o the equipment. When the copybit is 0, bits C_{15} are sampled. An original signal permits one copy, and this must be an analogue copy. When the copybit is 1, no sampling takes place. To avoid any interference in the biphase code, two successive bits must be altered in the left-hand and right-hand channels respectively.

Reverting to the data contained in the user bit. only parity bit *P* needs to be considered as adjacent second bit. This thus determines the parity.

Conversion with PLD chip

The block diagram of the copybit inverter is shown in Fig. 7, and the circuit diagram in Fig. 8.

The digital a.f. signal – 0.5 V_{pp} into 75 Ω – is coupled capacitively to inverter IC_{1a}, which is arranged as an amplifier. The standard circuit is an inverter with feedback, but this has the disadvantage that the circuit tends to oscillate with open input. In the present circuit, the operating point is set permanently with P_1.

Inverter IC_{1a} is followed by a delay circuit with a delay time of 120 ns.

To ensure that both inputs (optical and coaxial) provide equal signal levels, the output of the opto-receiver, which is about 1.5 V_{pp}, is applied to the coaxial input via R_8. A change-over switch is not needed, since R_8 decouples both inputs adequately. In optical operation, the signal can thus be taken straight from the coaxial socket.

The direct and delayed signals are XOR-gated in IC_2 This makes the signal independent of the polarity at the input, since all subsequent steps are related to the XOR signal. The spacing of the positive edge in case of a logic 0 is 354 ns and in case of a logic 1, 177 ns—see Fig. 5.

Normally, the clock is retrieved by a phase-locked loop, PLL, which, as far as time and phase ratios are concerned, is not easily kept stable. Moreover, the voltage-controlled oscillator, VCO, remains operational in the absence of an input signal, which makes decoding of the block and subframe clocks more complicated.

The XOR signal starts non-retriggerable monostable IC_{3a}, which has a dwell time of

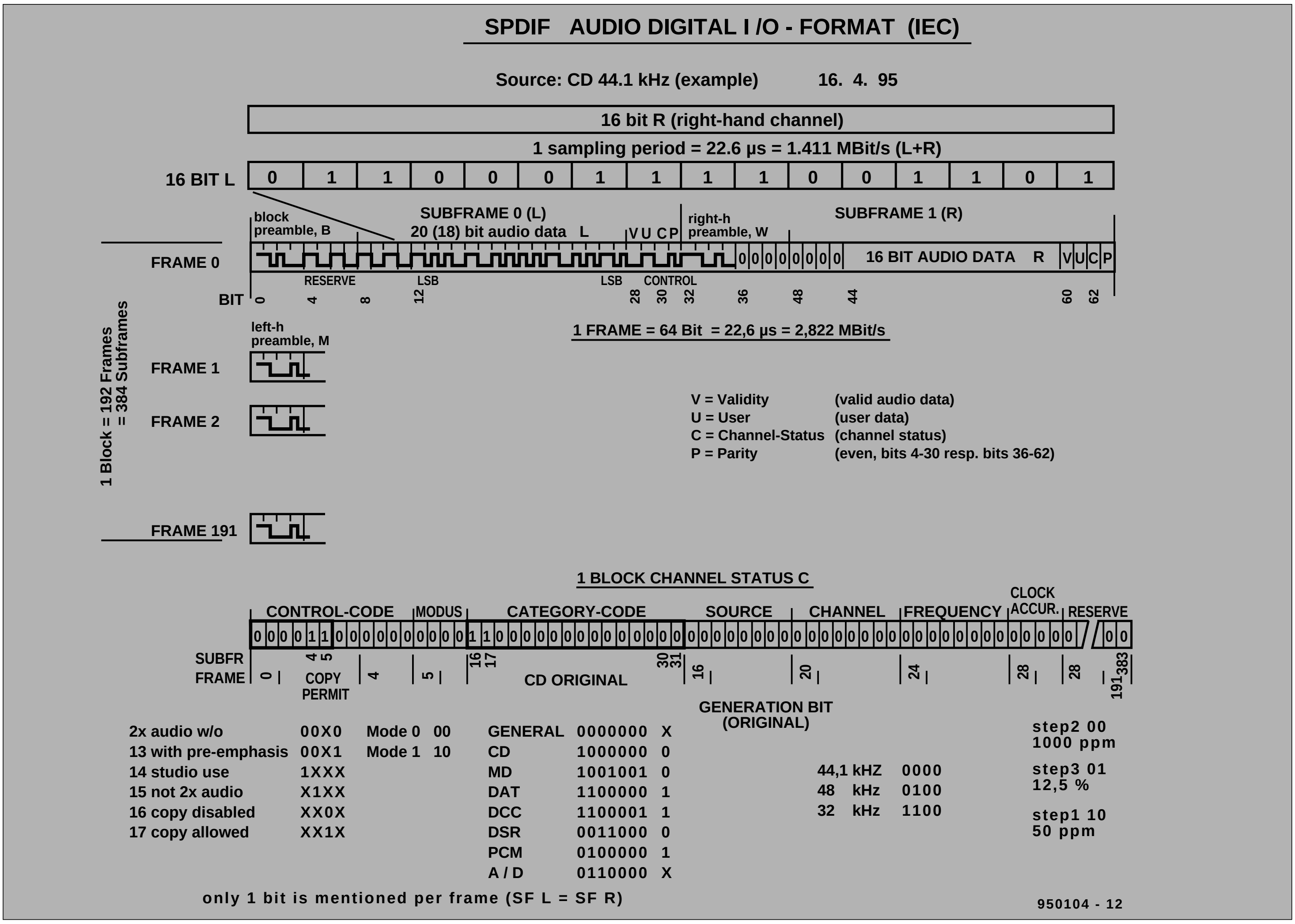


Fig. 2. Composition of a digital audio signal.

about 240 ns, to retrieve the bit clock. In the range of the preambles, the start spacings are >350 ns. This is made use of by retriggerable monostable IC_{3b}, which has a mono time of around 420 ns, to generate the subframe clock. The X-coded XOR pulse occurs only with the block preamble at the first pulse of the subframe clock. This enables the block clock to be decoded.

The block clock is generated regularly when the circuit operates as specified, that is, when there is a digital input signal in S/PDIF format, and the dwell times of the monostables are in accord. The block clock is stretched to a constant-1 signal (NOINV) by IC_{1d} and IC_{1e}; its presence is indicated by D_{10}. If, for instance, the signal has an incorrect frequency, NOINV prevents it being modified and this is indicated by D_{10} lighting less brightly. The LED does not light at all when the input signal is not of the S/PDIF format.

To count the subframes, a 9-bit counter is timed by the subframe clock and reset by the block clock. The 5-bit counter for the subframe bits is in synchrony with the bit clock and is reset by the subframe clock.

Filtering the desired bits (bit 30 in subframes 4, 5, and 16–31) is effected by programmable IC_2. The INVERT pulse has the correct position when signal IN_1 is delayed by about 60 ns (IN_2). A logic 1 is indicated when the relevant LED is driven by the level detector signal output by IC_2. This signal is generated in a manner similar to that of the block clock. Each D-bistable associated with a given LED is reset by the block clock and set with the 1-signal. The period to the next reset is long enough to enable the LED indicating the 1 in a stable way (without flickering).

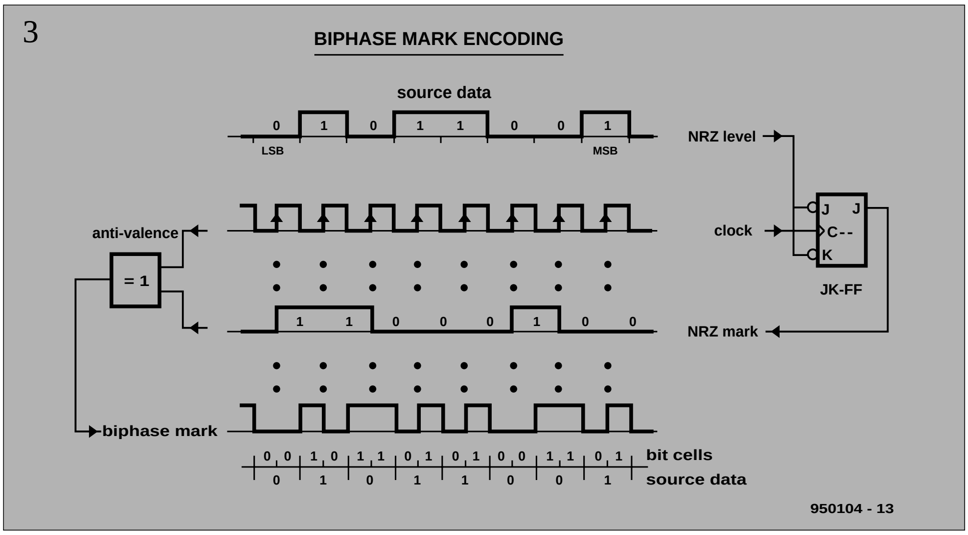
Inversion of the bit is accomplished by an XOR gate and the 354 ns long INVERT pulse that is located half-way between the C-bit and P-bit—see Fig. 6. The change from 0 to 1 presents no difficulties, since the edges of the INVERT pulse in signal IN_{12} meet at the centre of the bit at equal levels. Short spurious pulses at the centre of the bit can, however, not be avoided entirely. This does not matter, however, since the biphase-modulated signal is always sampled at the centre of a bit-cell, that is, at 1/3 and 2/3 of the bit.

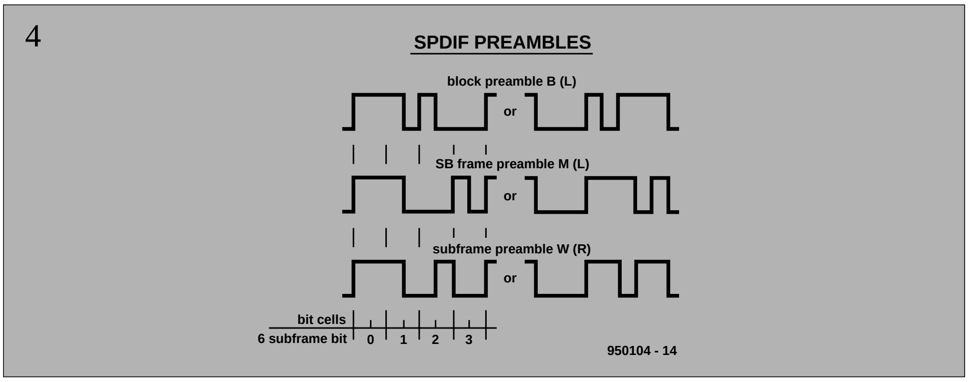
When bit C_8 and C_9 in the input signal are logic 1 (magnetic tape drive), they will not be affected. All other signal sources are assigned the code of a CD player.

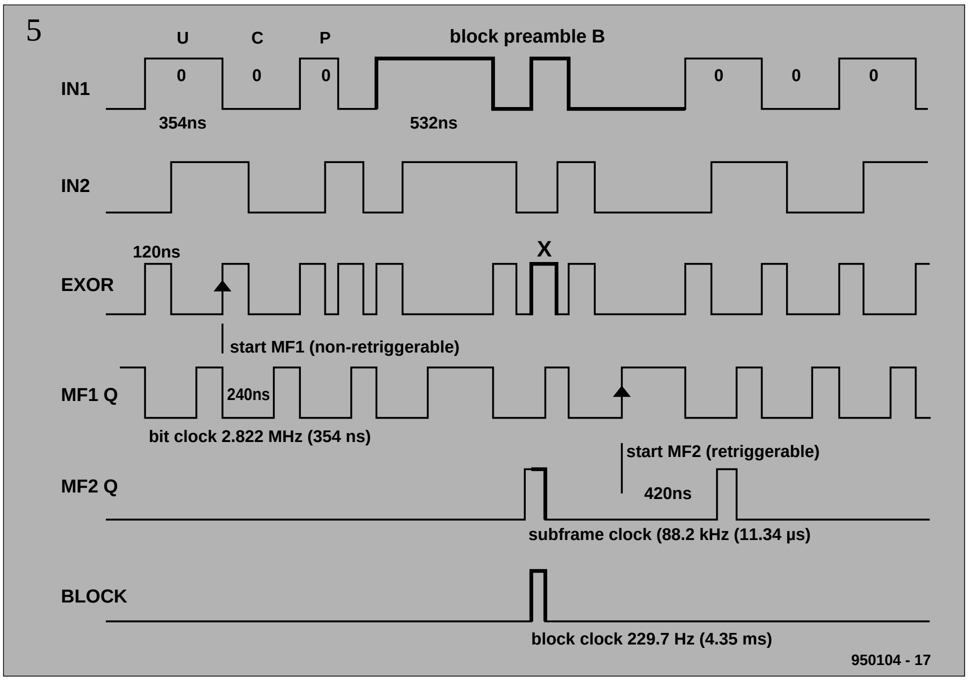
Moreover, bits 0 and 1 of the channel status are held at logic 0. Although this is

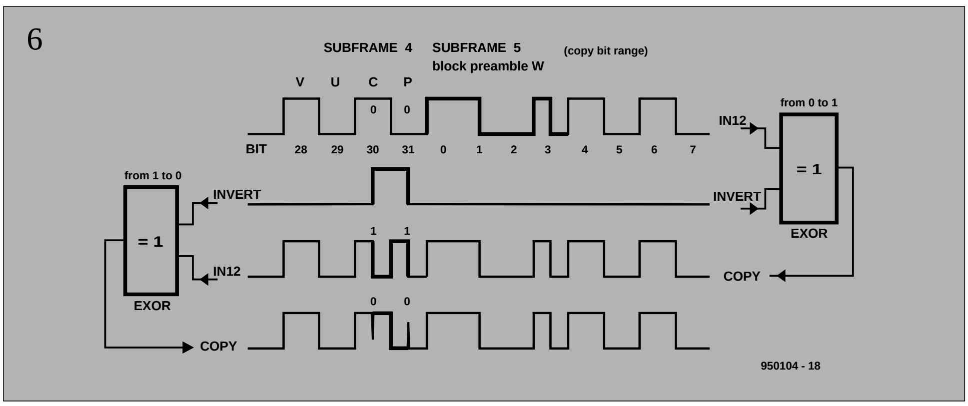
LED indications

- Only D_9 lights in both switch positions: ready for use; no S/PDIF
- LEDs flicker: unit is not operating correctly. It may be that both inputs are used simultaneously, or that the input signal is not of the correct format, or that the setup is incorrect, or that the optical input receives spurious signals.
- D_9 and D_{10} do not light: unit functions as converter; S/PDIF signal is transferred unchanged.
- D_{10} lights brightly: the S/PDIF input signal, with copybit and category, is indicated The output signal is DAT when the input comes from a magnetic tape drive, and CD when comes from any other source. In both cases, copying is permitted.









not really necessary in domestic equipment (since the bits then are always logic 0), it makes it possible for professional recordings or other recordings marked by these bits (which are inhibited) to be copied—but see warning at beginning of this article.

With switch S_1 open, the inverter accepts sampling frequencies of 44.1 kHz and 48 kHz, but with 32 kHz it must be closed to alter the time constants of the monostables. If this switch is in the wrong position, D_9 and dimly lit D_{10} indicate that the signal is unchanged: the unit functions as a converter. A no-signal condition is indicated by D_9 lighting.

It is highly improbable that only the generation bit, which does not count in the equipment coding, is encoded. Anyway, there is always D_{10} as a controlling element.

The output is buffered by inverter IC_{1f}. Resistors R_4 and R_5 lower the signal level to about 0.5 V_{pp} into 75 Ω. Capacitor C_1 blocks any direct voltage.

Timing the monostables

Construction of the inverter on the printed-circuit board in Fig. 9 should not present any undue difficulties. All ICs, except IC_6, should be seated in sockets. Be careful with inserting IC_2 into its PLCC socket. Do not forget the wire bridge underneath IC_2.

After the board has been finished and thoroughly checked, set the presets to the centre of their travel.

Apply an audio signal, preferably from a CD player set to PAUSE (which ensures a very stable signal) to the coaxial input socket. Set switch S_1 to 44.1/48 kHz, whereupon D_3 (category code CD) should light. If an oscilloscope or logic analyser is not available, adjust presets P_1, P_2 and P_3 (in that order) on to the centre of the stable LED indication.

With S_1 in position 32 kHz, the signal source must be a DAT recorder, set to the long-play analogue recording mode, or a DSR tuner. Carefully readjust P_3 (which should not be much) and recheck the settings with a signal from a CD player. For most practical purposes, these settings are fine.

If more accurate settings of the presets are required, an oscilloscope is needed. Apply an a.f. signal at a level of 0.5 V_{pp} to the coaxial, not to the optical, input. Set the oscilloscope time base to 100 ns cm^{-1}

Fig. 3. Biphase coding enables the simultaneous transmission of the audio signal and the clock.

Fig. 4. Various waveforms of a non-biphase coded preamble.

Fig. 5. Extraction of the bit-clock, subframe clock and block clock.

Fig. 6. Principle of bit inversion.

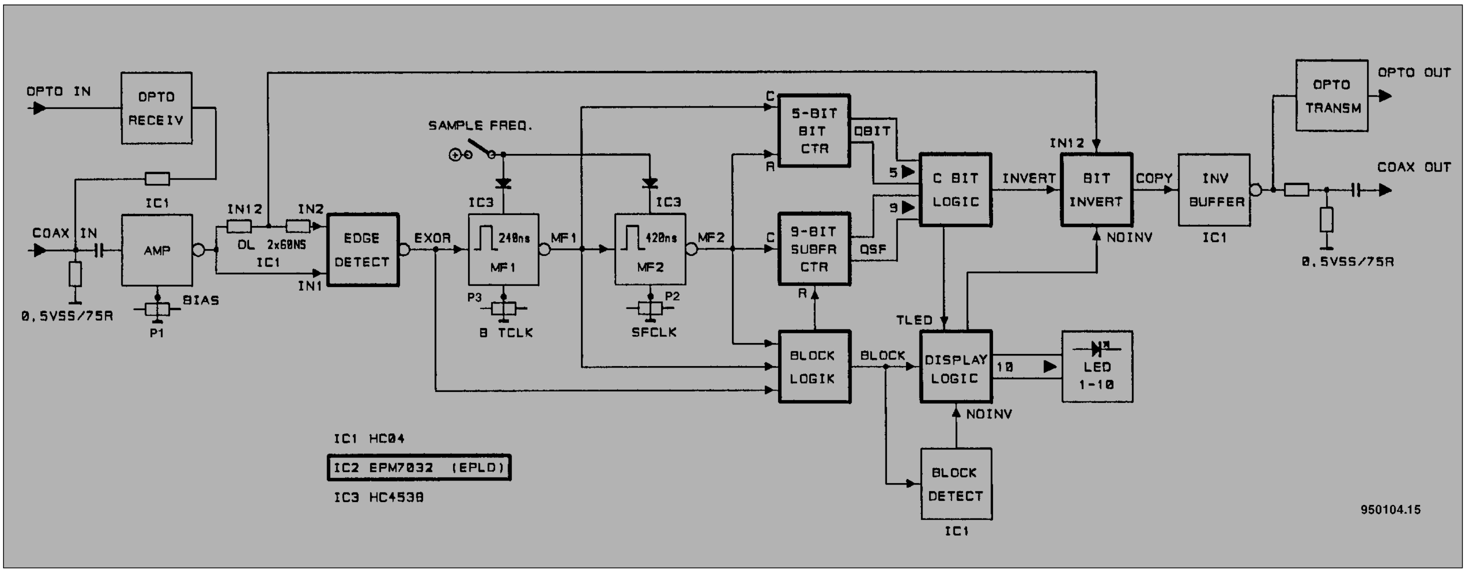


Fig. 7. Block diagram of the copybit inverter.

and connect the instrument to pin 9 of IC_3. Adjust P_1 so that all edges cover one another as well as possible. This ensures that the operating point of the unit is centralized and that the delay of the rising edge of signal IN_2 is equal to that of the trailing edge.

With P_2, set the pulse width of the subframe clock at pin 7 of IC_3 to 100–150 ns. If there is an appreciable difference in dwell times between the standard and long-play settings, the value of R_{12} may be adapted accordingly.

If the oscilloscope has a second time base or *x*-multiplier ×10, the copybit (bit 30) corrected to logic 1 in subframe 4 or 5

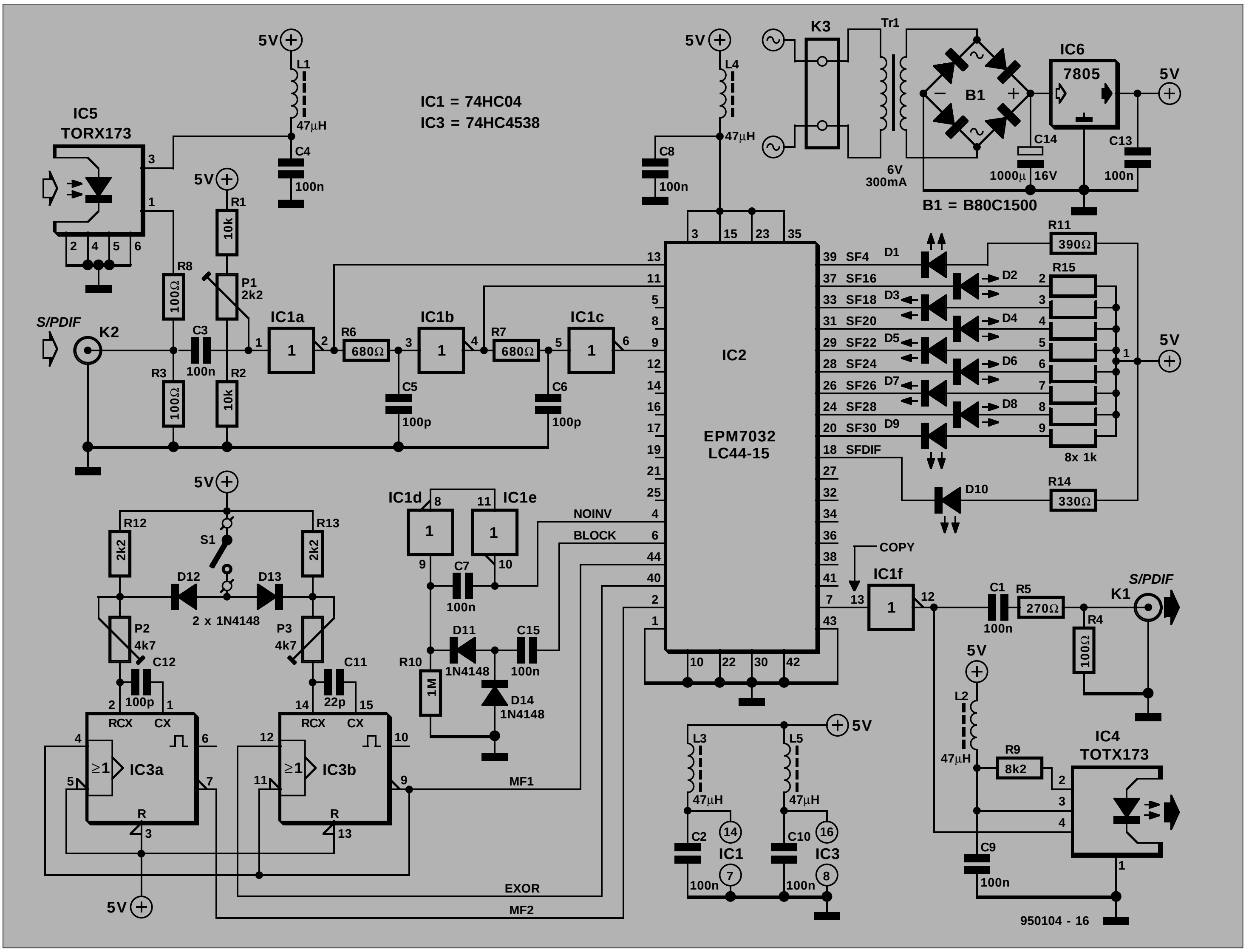


Fig. 8. Circuit diagram of the copybit inverter.

can be timed in the output signal with P_3 [time base set to 10 (1) μs cm^{-1} and triggering to start the block at the leading edge of the cathode signal of a lighted LED (D_1–D_9)].

The high and low level portions of the C-bit set to logic 1 should be equal or very nearly so. If the P-bit is inverted from logic 0 to 1, it should be virtually undistorted.

If the bits away from the block start are to be checked, the LED voltage (trigger at trailing edge) associated with the P-bit can be used for marking them.

During the setting up, make sure that the LEDs light over a fairly wide range. Appreciable differences can be negated by adapting the value of R_{13}.

As a final check, record the output of the copybit inverter on a DAT or DCC recorder. Some DAT recorders show the ID6: this must be 00 both during recording and subsequent playback of the recording—see Fig. 1. Make sure that D_1 lights as an indication that the unit has correctly inverted and processed the input signal.

Parts list

Resistors:
R_1, R_2 = 10 kΩ
R_3, R_4, R_8 = 100 Ω
R_5 = 270 Ω
R_6, R_7 = 680 Ω
R_9 = 8.2 kΩ
R_{10} = 1 MΩ
R_{11} = 390 Ω
R_{12}, R_{13} = 2.2 kΩ
R_{14} = 330 Ω
R_{15} = resistor array, 8.1 kΩ
P_1 = preset, 2.2 kΩ
P_2, P_3 = preset, 4.7 kΩ

Capacitors:
C_1–C_4, C_7–C_{10}, C_{13}, C_{15} = 100 nF, ceramic
C_5, C_6, C_{12} = 100 pF
C_{11} = 22 pF
C_{14} = 1000 μF, 16 V, vertical

Semiconductors:
D_1, D_{10} = LED, low-current, 3 mm, green
D_2–D_9 = LED, low-current, 3 mm, red
D_{11}–D_{14} = 1N4148

Integrated circuits:
IC_1 = 74HC04
IC_2 = EPM7032LC44-15 (Altera), programmed with software 956513-1*
IC_3 = 74HC4538
IC_4 = TOTX173 (Toshiba)
IC_5 = TORX173 (Toshiba)
IC_6 = 7805

Miscellaneous:
K_1, K_2 = audio socket for board mounting
K_3 = 2-way spring-loaded terminals for board mounting, pitch 7.5 mm
S_1 = toggle switch with on contact
B_1 = B80C1500, round
Tr_1 = mains transformer, 6 V, 300 mA
44-pin PLCC socket for IC_2
Enclosure 120×40×70 mm

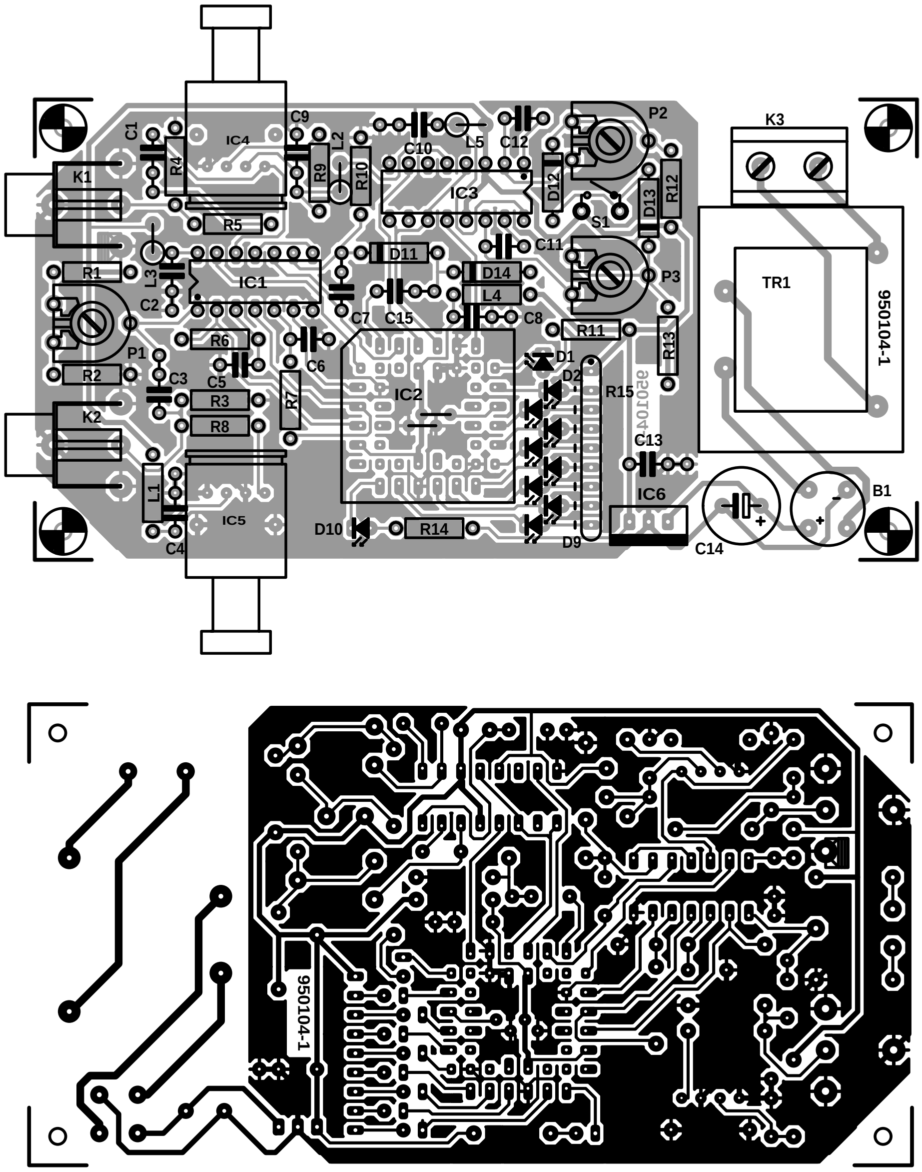


Fig. 9. Printed-circuit board for the copybit inverter (scale 1:1).

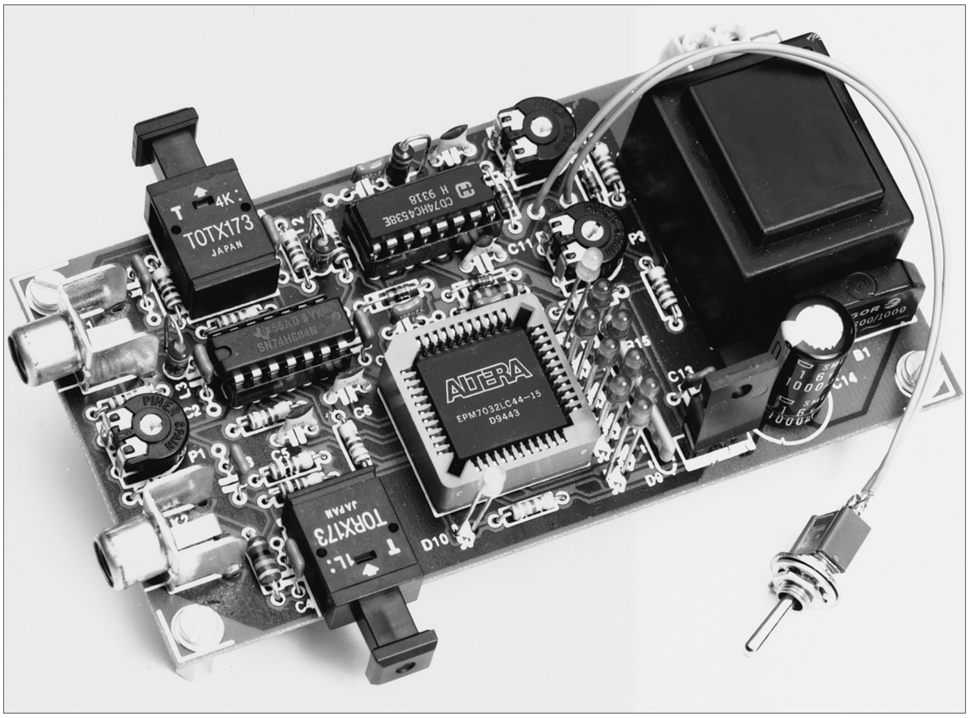


Fig. 10. Finished prototype board.

When a digital-to-analogue converter (DAC) is used in conjunction with a CD player, their clocks must be in synchrony to make sure that the DAC can process the data error-free. In practice, this means that the clock of the CD player has to be applied to the DAC.

If the DAC is built into the CD player, the CD player clock can be applied as shown in the upper diagram. The clock signal is available at TTL level at the output of IC_{1b}. The DAC clock, IC_{2a}, is synchronized with IC_{1a} via P_1 and C_6. In practice, P_1 is set just past the point where synchrony commences: this ensures that oscillator IC_2 continues to work when IC_{1a} is disabled for whatever reason.

An important advantage of the design is that the circuit does not influence the operation of the electronics in the CD player (which thus retains its original functionality).

If the DAC is used as a stand-alone unit, a transmission line for the data and clock signals is required. As usual, this is a 75 Ω coaxial cable. The lower diagram shows how this setup can be arranged.

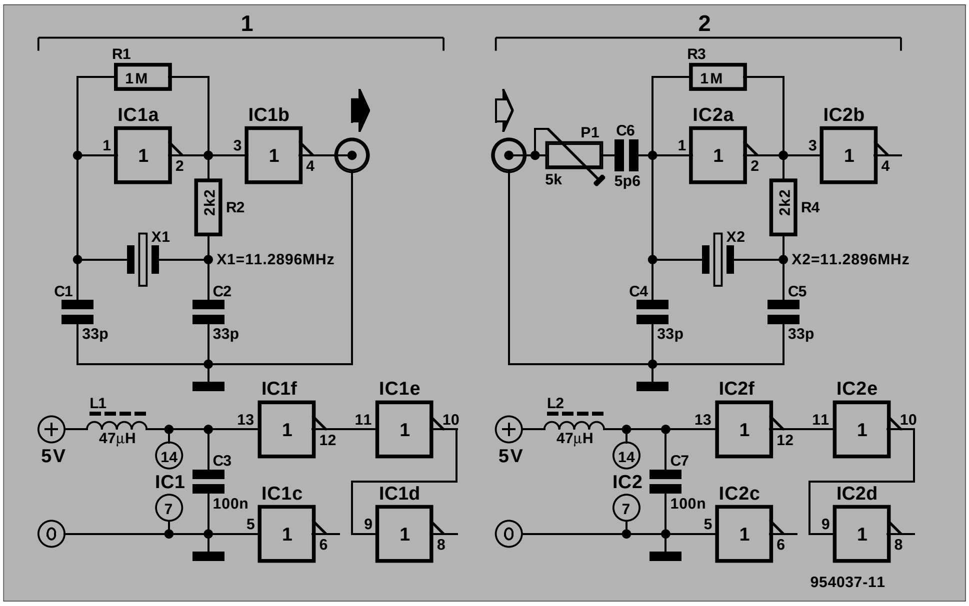
The chosen signal level of 1.5 V_{pp} is more than sufficient to ensure synchrony. The values of coupling components P_2 and C_{10} are, however, different from those of P_1 and C_6 in the upper diagram.

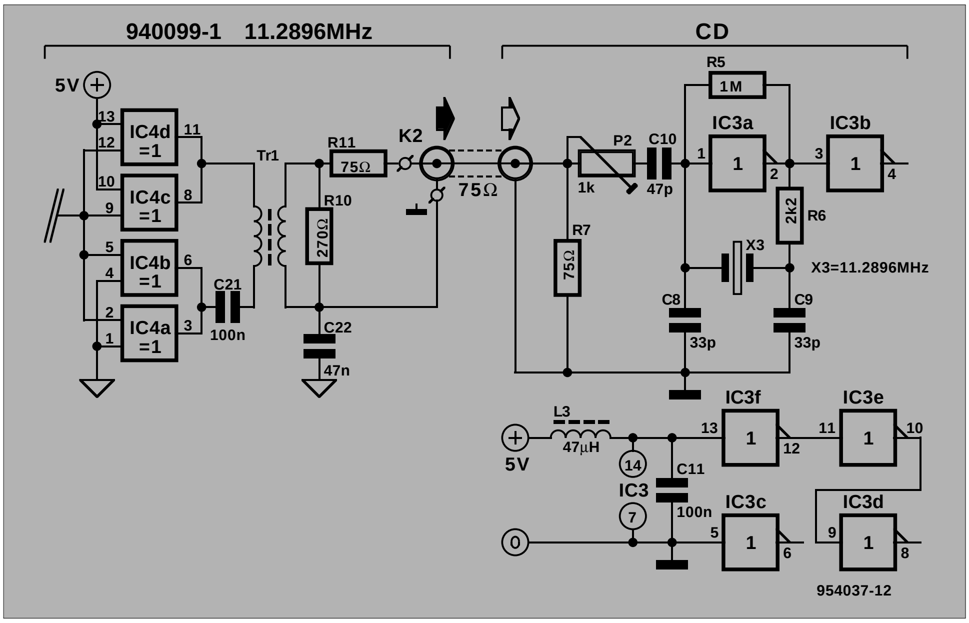
A drawback of this setup is that the oscillator no longer starts spontaneously owing to the increased damping. Fortunately, this can be remedied readily. Resistor R_6 limits the energy transferred from the buffer amplifier to the crystal. If the value of this resistor is greatly reduced, even down to 0 (wire bridge), it will be found that the oscillator starts spontaneously again. Note that in some CD players R_6 is replaced by a wire bridge.

Another remedy is reducing the turns ratio of the transformer, which increases the level of the clock signal. If this is done, the value of P_2 can be increased an that of C_{10} reduced. The load on the oscillator is then smaller, so that it starts spontaneously,

The oscillator circuit draws a current of about 10 mA.

Design by T. Giesberts





[954037]

PCB Order no. 950104*

* Combination packet Order no. 950104C

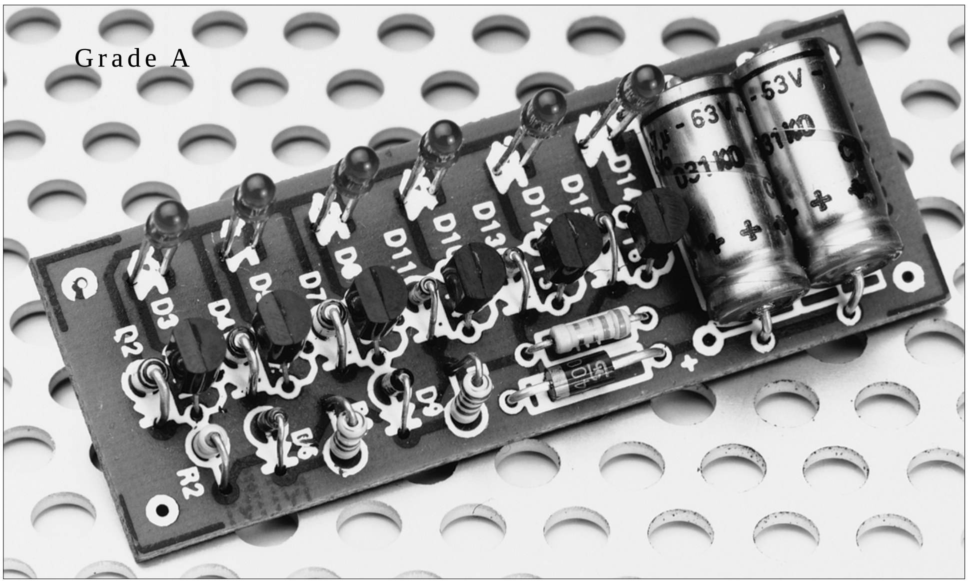
Sources:
Sony SCMS Handbook DTC-55ES

Valvo TI871011

DIN EN 60 968

[950104]

PASSIVE VU METER



Design by T. Giesberts

If you want to fit your loudspeaker enclosure(s) with a drive indicator, it is best to use a unit that does not need a power supply. The September 1995 issue of this magazine (p. 75) described an AF POWER INDICATOR for PA (public address) loudspeakers, that is, for fairly high power outputs. This article describes a passive VU unit for smaller amplifiers and loudnesses—see Fig. 1. It consists of a rectifier and six (may be fewer or more) identical stages that each comprises a current source, a zener diode and a light emitting diode, LED. The current sources are built from JFETs with interconnected gate-source terminals—here, Type BF256A. The saturation current, I_{DSS}, of these devices with a drain-source voltage, U_{DS} = 15 V is about 5 mA. This current is not exactly constant, but is perfectly all right for driving a low-current LED and will not exceed the permissible value of 7 mA. Networks R_2-D_2, R_3-D_4, and R_4-D_7 are protection circuits; they prevent the drain voltage of the relevant JFET rising above 30 V, which normally destroys the transistor.

The rectifying circuit is formed by D_1 and capacitors C_1, C_2. Resistor R_1 limits the peak current to about 1.5 A at a source voltage, U_S, of 50 V. Since it is in series with C_1 and C_2, and thus in parallel with the amplifier output, it has no effect on the level of the input voltage. The peak output voltage of the rectifier is applied directly to the single LED stages. The potential across C_1 and C_2 is not exactly equal to the instantaneous peak voltage, but, because of time constant R_1-C_1-C_2, is a good average of it. Consequently, the unit indicates briefly the instantaneous peak voltage, and then the mean of it.

The coworking of the three parts of a stage is easily understood by considering the following. If the rectified and smoothed voltage rises a few volts over the level set by the zener diode, the current source comes into action and causes the LED to light. Since the (input) voltage to the meter is directly proportional to the amplifier output and the (assumed constant) impedance of the loudspeaker, the indicated threshold level (in watts) can be converted into a zener voltage:

$$P = U_{RMS}/R = (U_s/\sqrt{2})^2/R = U_s^2/2R$$

$\therefore$

$$U_s = U_{ZENER} = \sqrt{2PR} - U_{LED}$$

where U_{LED} is the starting voltage of the LED (and the voltage drop across the current source), which is equal to 2 V. Thus, for an indication of 100 W into an 8 Ω loudspeaker, the zener voltage is

$$U_{ZENER} = \sqrt{(100 \times 2 \times 8)} - 2 = 38 \text{ V}.$$

The zener to be used should have the next lower rating in the table (36 V), so that it lights brightly when the output is 100 W. In this way, the stages may be designed more or less to individual requirements.

In the most sensitive stage, T_1-D_2-D_3, the zener diode is, strictly speaking, superfluous since the indicated power is determined entirely by the threshold values of D_1, T_1 and D_3.

The input current to the circuit has a peak level of U_{in}/R_1 = 50/33 = 1.5 A. With a constant 1 kHz signal and an output level of 150 W into 8 Ω, the current drops to 280 mA. However, if this signal is pulsed with a duty factor of 1:99, the current rises to 1.3 A owing to the then low average potential across capacitor C_1. It is noteworthy

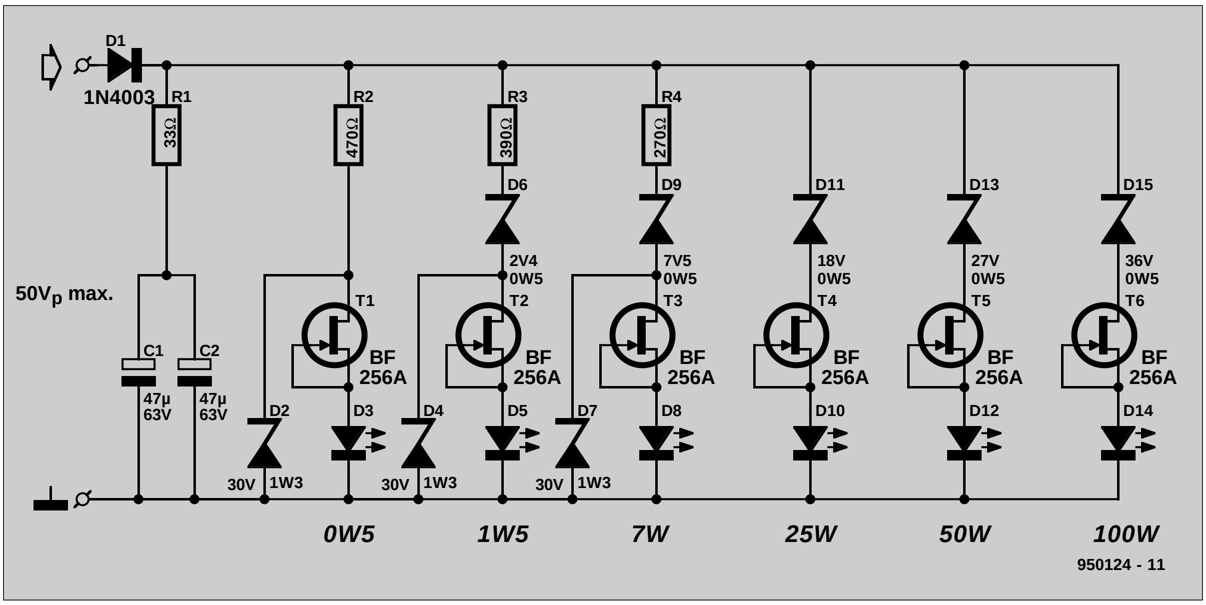


Fig. 1. The circuit of the passive VU meter is based on six identical stages.

that the circuit is not truly passive, because it draws its energy from the audio signal. This means, by the way, that there is a (very) slight rise in the distortion!

The circuit may be built quickly and without any undue difficulties on the printed-circuit board shown in Fig. 2. The finished board is best fitted in a small plastic enclosure, which is then fitted on to the loudspeaker box(es).

Standard zener diode voltages (V)

1.0*	10	100
	11	110
	12	120
1.4*	13	130
1.5*	15	150
	16	160
	18	180
2.0*	20	200
	22	
2.4	24	
2.7	27	
3.0	30	
3.3	33	
3.6	36	
3.9	39	
4.3	43	
4.7	47	
5.1	51	
5.6	56	
6.2	62	
6.8	68	
7.5	75	
8.2	82	
9.1	91	

* rare

Parts list

Resistors:
R_1 = 33 Ω
R_2 = 470 Ω
R_3 = 390 Ω
R_4 = 270 Ω

Capacitors:
C_1, C_2 = 47 µF, 63 V

Semiconductors:
D_1 = 1N4003
D_2, D_4, D_7 = zener diode 30 V, 1.3 W
D_3, D_5, D_8, D_{10}, D_{12}, D_{14} = low-current LED
D_6 = zener diode 2.4 V, 500 mW
D_9 = zener diode 7.5 V, 500 mW
D_{11} = zener diode 18 V, 500 mW
D_{13} = zener diode 27 V, 500 mW
D_{15} = zener diode 36 V, 500 mW
T_1–T_6 = BF256A

Miscellaneous:
PCB Order no. 950124

[950124]

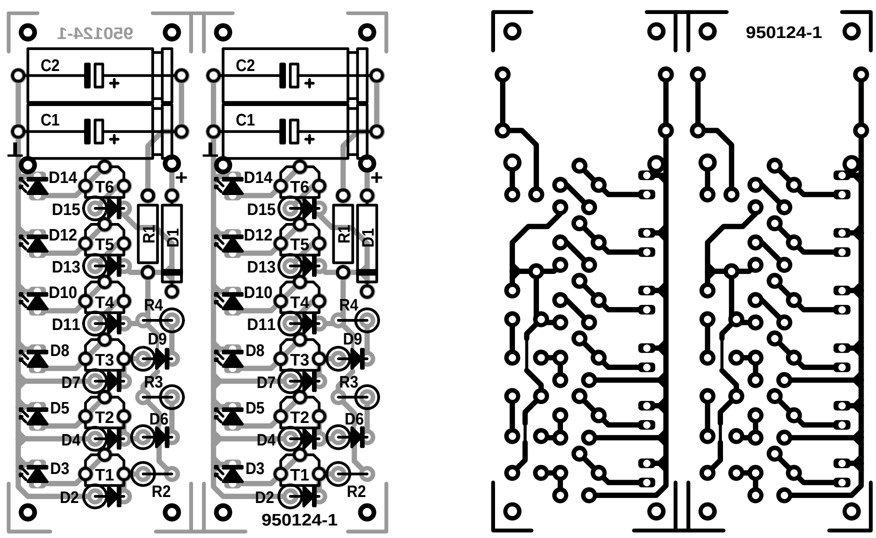


Fig. 2. The printed-circuit board for the passive VU meter must be cut into two before any assembly work is begun.

SPLITTER FOR S/PDIF COAX/OPTICAL OUTPUT

THIS circuit was originally designed to enable one S/PDIF output on a CD player to drive several inputs. The circuit acts as a 3-way splitter or format converter for S/PDIF ('digital') or optical signals. The function of the circuit is determined with the aid of two jumpers, as follows:

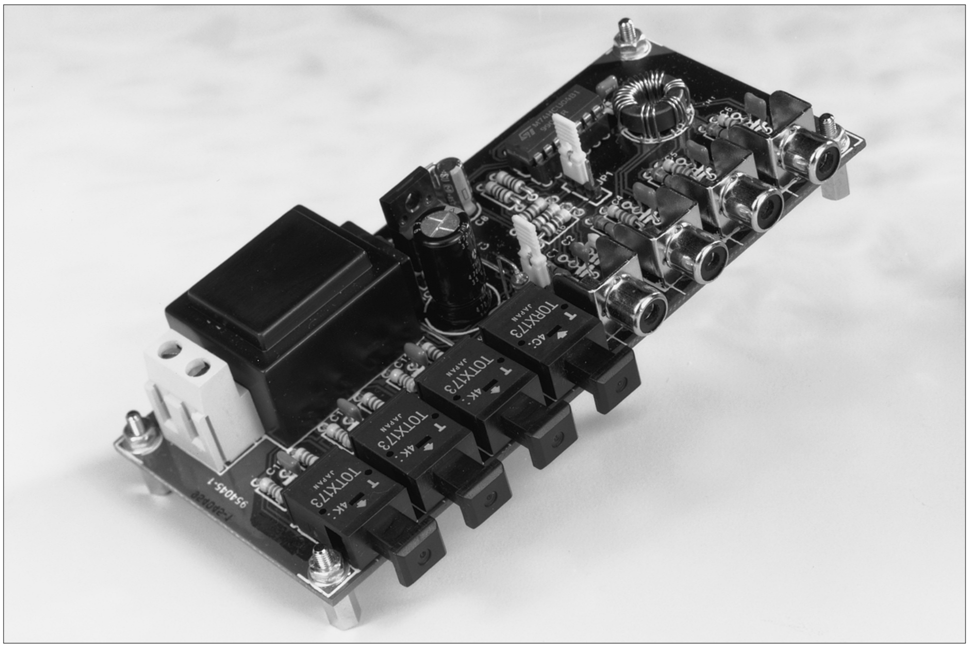
JP_1	JP_2	Function
O	O	optical input to optical and coax output
O	A	optical input to optical (buffered) and coax output
C	O	optical input to optical output, coax input to coax output
C	A	coax input to optical and coax output

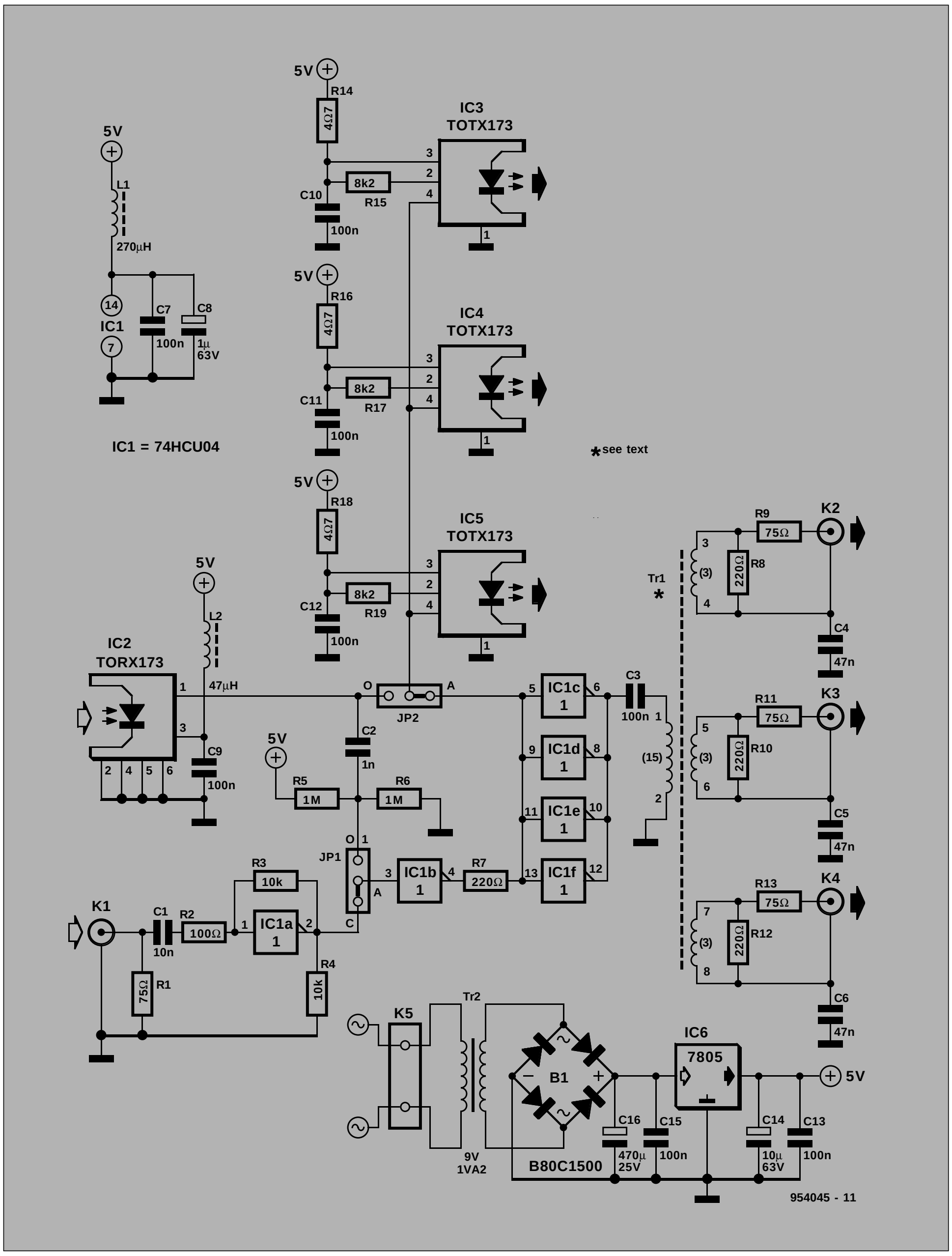
In the selections available, JP_1 has priority. The third option indicates that the circuit is capable of splitting two sources separately, while retaining their formats. It is not possible to make cross connections because it is not usually required to convert from coax to optical. The other way around, though, is far more usual, and available with the other three options.

The optical inputs and outputs consist of Toshiba 'Toslink' modules, which are TTL compatible. The coax inputs and outputs are formed by RCA-style 'phono' sockets. Both the Toslink modules and the phono sockets are accommodated on the printed circuit board.

A digital (coax) input signal is first amplified by IC_{1a}. Resistor R_4 lightly loads the gate to prevent it from oscillating if no input signal is connected. In some cases, a resistor with a slightly lower value may be required in this position. Next, jumper JP_1 allows the output of the amplifier (position 'C'), or that of Toslink receiver IC_2 (position 'O'), to be selected. The Toslink signal is superimposed on a half-supply bias created by R_5 and R_6, and then applied to IC_{1b}. The half-supply bias corresponds to the switching threshold of IC_{1b}. This allows Toslink transmitters IC_3, IC_4 and IC_5, to be driven from the output of receiver IC_2, or from the output of IC_{1b}. Remember, the signal supplied by the latter is cleaned and amplified. Simply select the option which gives the best results.

The buffer formed by the four parallel-connected inverters has sufficient drive capacity to match the low impedance of the primary of transformer Tr_1. The transformer actually forms the heart of the circuit. The three digital outputs are electrically isolated from the input, although the r.f. signals 'see' the ground level through capacitors C_4, C_5 and C_6. Matching resistors R_9, R_{11} and R_{13} prevent reflections at the relevant output from exceeding the ×1 level. Resistors R_8, R_{10} and R_{12} similarly damp the ringing effects which occur on the non-used secondary windings. The drawing shows the practical construction of the transformer, which is based on a type G2.3-FT12 ferrite ring core. This type was chosen because of its high bandwidth and coupling factor, factors which allow the primary winding of 15 turns to occupy only





about half the core, while the secondary windings of three turns each are distributed over the remainder. All windings are made from 0.5-mm dia (24 SWG) enamelled copper wire.

The power supply is conventional and 'on board', consisting mainly of a 1.2-VA mains transformer, a bridge rectifier and a 7805 three-pin voltage regulator. All ICs on the board have individual supply decoupling parts. Current consumption of the circuit is of the order of 70 mA. We regret that the printed circuit board shown is not available ready-made.

Parts list

Resistors:
R_1;R_9;R_{11};R_{13} = 75Ω
R_2 = 100Ω
R_3;R_4 = 10kΩ
R_5;R_6 = 1MΩ
R_7;R_8;R_{10};R_{12} = 220Ω
R_{14};R_{16};R_{18} = 4Ω7
R_{15};R_{17};R_{19} = 8kΩ2

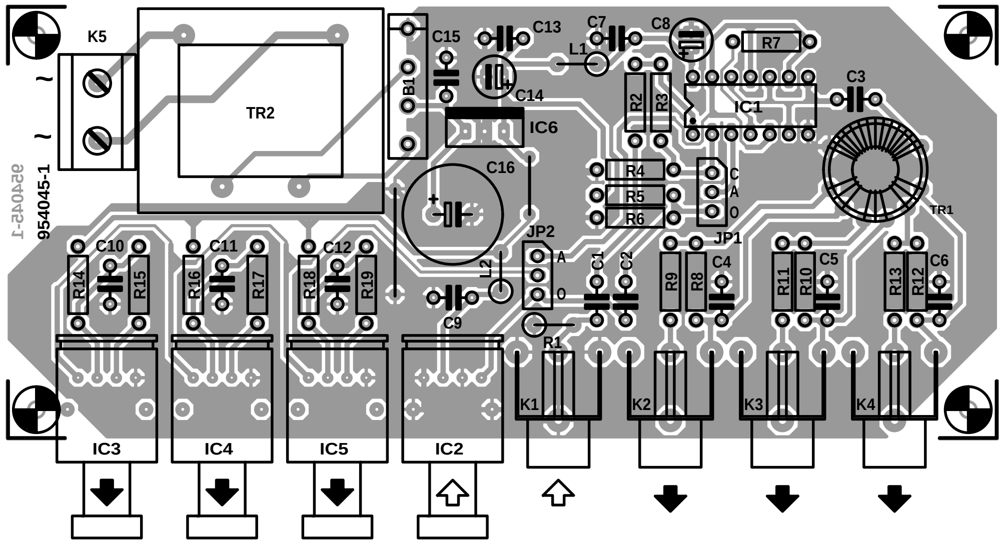
Capacitors:
C_1 = 10nF ceramic
C_2 = 1nF ceramic
C_3;C_7;C_9-C_{13};C_{15} = 100nF ceramic
C_4;C_5;C_6 = 47nF ceramic
C_8 = 1µF 63V radial
C_{14} = 10µF 63V radial
C_{16} = 470µF 25V radial

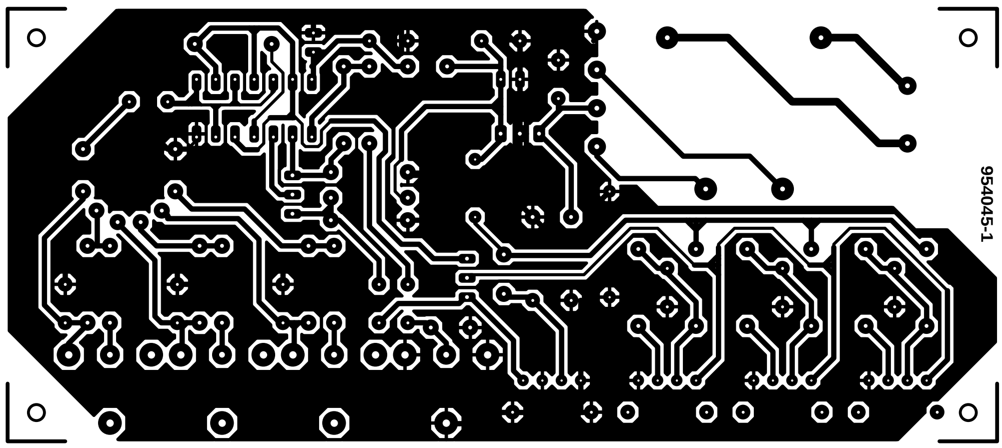
Inductors:
L_1 = 270µH choke
L_2 = 47µH choke

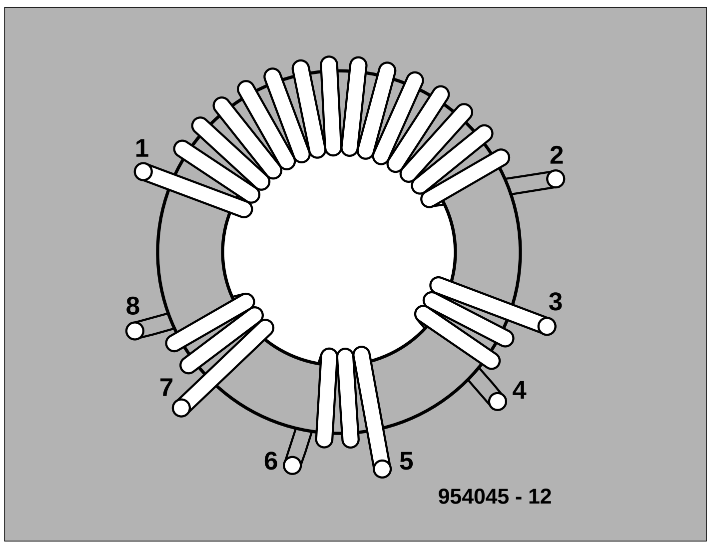
Semiconductors:
IC_1 = 74HCU04
IC_2 = TORX173 (Toshiba)
IC_3;IC_4;IC_5 = TOTX173 (Toshiba)
IC_6 = 7805

Miscellaneous:
JP_1;JP_2 = 3-way pin header, w. jumper.
K_1-K_4 = RCA style PCB mount socket, Monacor T709G.
K_5 = 2-way PCB terminal block, pitch 7.5mm.
B_1 = B80C1500
Tr_1 = G2.3-FT12 ferrite ring core; primary 15 turns 0.5mm dia. ecw, secondary 3×3 turns 0.5mm dia. ecw.
Tr_2 = 9V/1.2VA transformer, e.g, Hahn BV EI 302 0376; Velleman 1090012M; Monacor VTR1109 (1.5VA); Block VR1109 (1.5VA).

Design by T. Giesberts
[954045]







PICK-UP INPUT BECOMES LINE INPUT

Many audio amplifiers are still fitted with one or two inputs for a dynamic pick-up. Since record players are used less and less frequently, these inputs remain unused in many cases. At the same time, many audio systems lack a good line input.

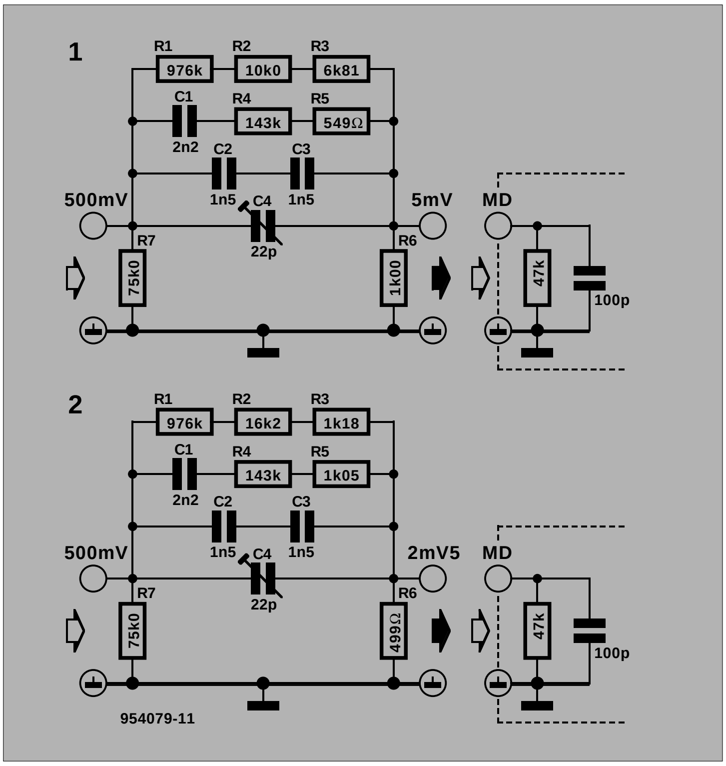
The present circuit enables the (largely) unused dynamic input to be used as a line input. A simple passive network arranges the required level matching.

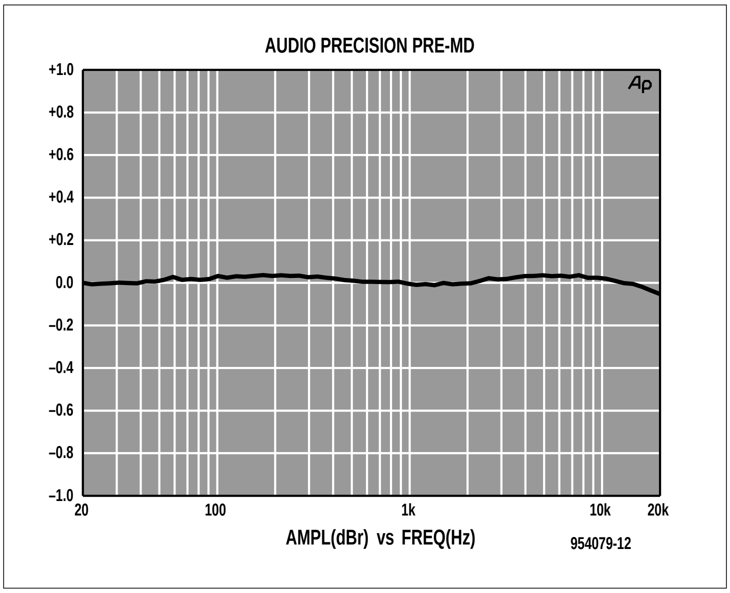
The diagram shows two versions of the circuit: version 1 provides higher amplification than version 2. Which version is to be used depends on the sensitivity of the dynamic input, which is normally 5 mV or 2.5 mV at 1 kHz. Both versions have a line input sensitivity of 500 mV.

The design is based on highly accurate components: the resistors have a tolerance of 0.1%, while the capacitors must be manually selected with the aid of an accurate capacitance meter. The resulting circuit is much more accurate than the usual RIAA correction network in the amplifier. It is, therefore, highly suitable for testing a pre-amplifier for compliance with the RIAA correction.

Nevertheless, standard 1% components may be used; the resulting frequency characteristic is shown in the second diagram. The theoretical deviation of circuit 1 from the ideal curve –0.05 dB at 20 kHz and that of circuit 2 at the same frequency is –0.012 dB.

Design by T. Giesberts
[954079]





Contents

AF LEVEL MATCHING

The matching circuit consists of a variable passive attenuator and an amplifier with variable gain: 0–20 dB. The attenuator reduces the signal by 0, $^1/_4$, $^1/_2$ or $^3/_4$. If required, the circuit may be adapted for other reduction factors.

When the upper section of DIP switches S_1 and S_2 is closed, the attenuation is 0 dB. The input impedance of the circuit, 40 kΩ, can then be changed to 30 kΩ, 20 kΩ or 10 kΩ by closing one of the other switches.

The buffer/amplifier is formed by IC_1. Potentiometers P_1 and P_2 serve to set the amplification factor. Make sure that they are both set to exactly the same value since presets have a tolerance of 20%; if they are not, the amplification in the left-hand and right-hand channels is not the same. If the circuit is used primarily as an attenuator, set both presets to their minimum value; the op amp then functions as a voltage follower.

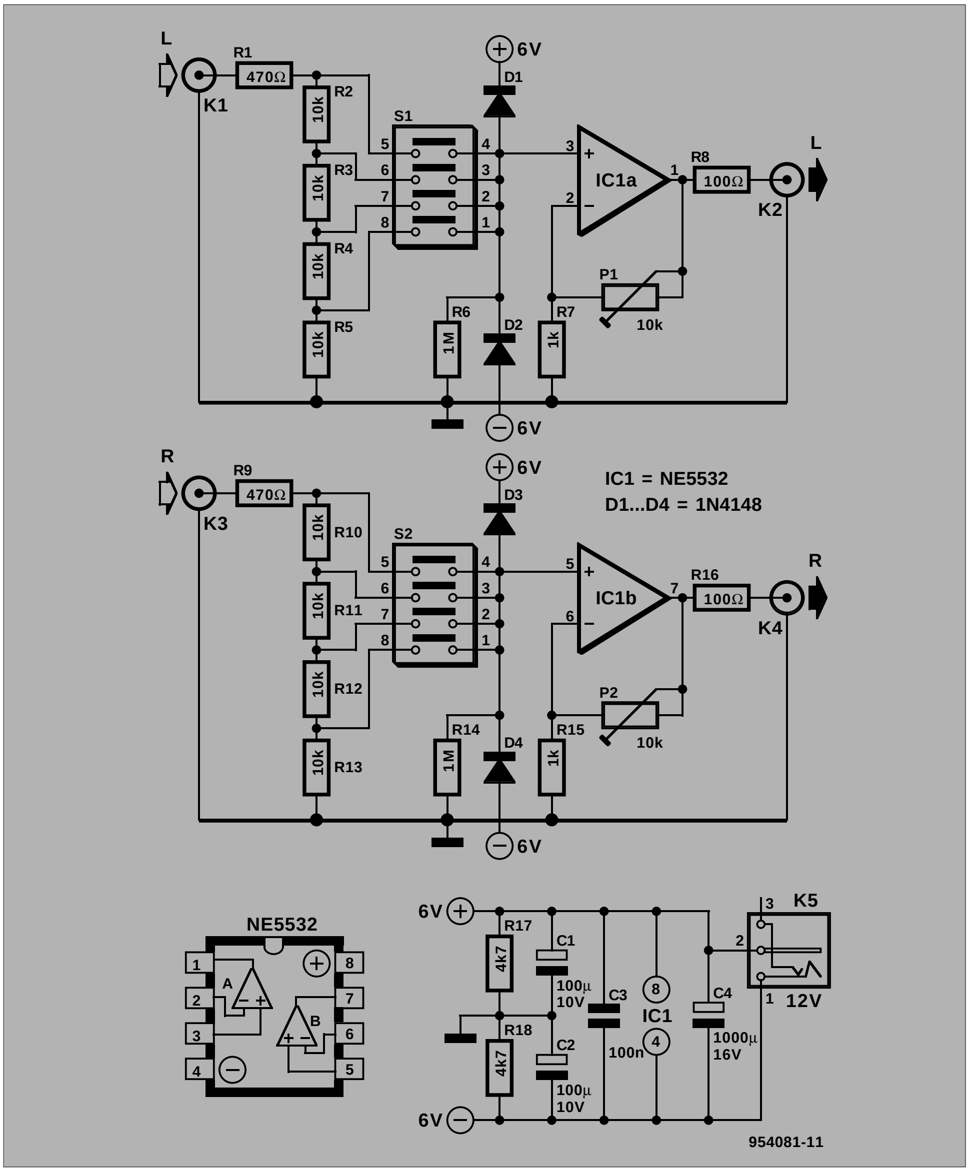
Power may be derived from a mains adaptor. Since the matching circuit should work with a symmetrical supply (when coupling capacitors may be omitted), a virtual earth is provided with the aid of R_{17}, R_{18}, C_1 and C_2. The specified values of these components apply to a load impedance of 50 kΩ. For lower load impedances, the values of the resistors must be reduced and that of the capacitors increased accordingly. The mains adaptor is decoupled by C_4.

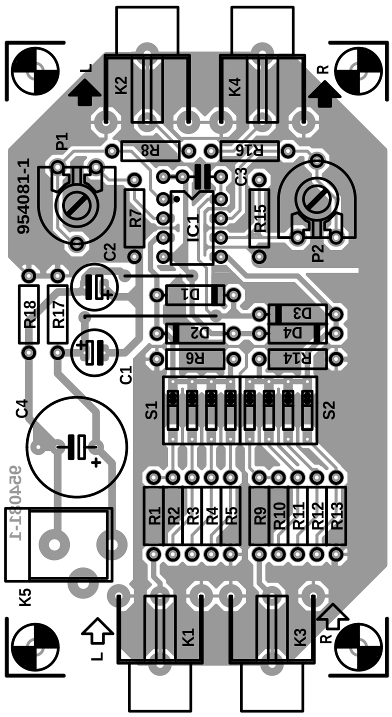
Diodes D_1–D_4 protect the inputs of IC_{1a} and IC_{1b} against too large input signals.

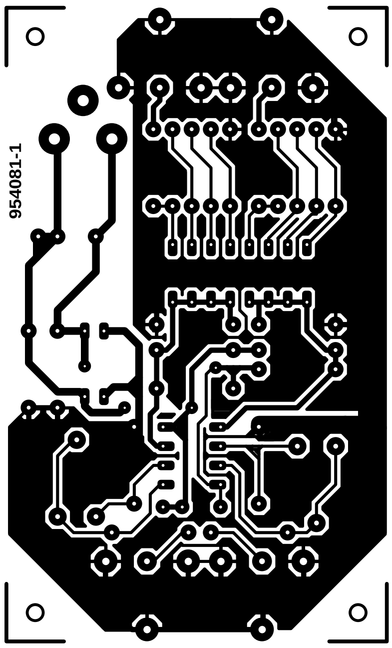
Resistors R_6 and R_{14} provide a bias current for the op amps when all switches are open.

The total harmonic distortion plus noise measured in the prototype working with a gain of 0 dB, a frequency of 1 kHz, an output voltage of 2 V, and a load of 50 kΩ was smaller than 0.0004% (at a bandwidth of 80 kHz). When the gain is raised to 20 dB and the input signal is 200 mV, the distortion rises to 0.0012%. Channel separation is >100 dB at 1 kHz and >80 dB at 20 kHz.

The circuit draws a current of not more than 10 mA.



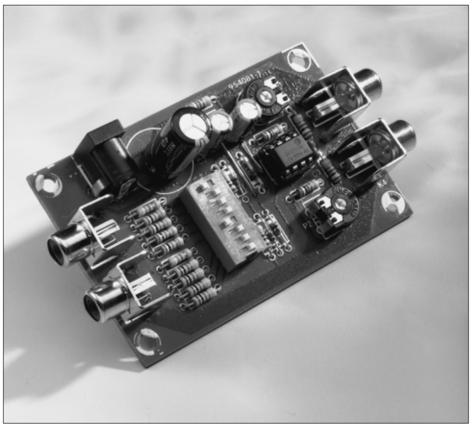




Parts list

Resistors:
R_1, R_9 = 470 Ω
R_2–R_5, R_{10}–R_{13} = 10 kΩ
R_6, R_{14} = 1 MΩ
R_7, R_{15} = 1 kΩ
R_8, R_{16} = 100 Ω
R_{17}, R_{18} = 4.7 kΩ
P_1, P_2 = 10 kΩ preset

Capacitors:
C_1, C_2 = 100 µF, 10 V, radial
C_3 = 100 nF
C_4 = 1000 µF, 16 V, radial



Semiconductors:
D_1–D_4 = 1N4148

Integrated circuits:
IC_1 = NE5532A

Miscellaneous:
K_1–K_4 = Audio socket for board mounting
K_5 = Plug for accepting mains adaptor socket
S_1, S_2 = 8-position DIP switch
PCB not available ready made

Design by T. Giesberts
[954081]

Contents

SINGLE-CHIP 50 W AF AMPLIFIER

The chip on which the amplifier is based, a Type LM3876, is a member of the Overture family from National Semiconductor, All members of this family are pin-compatible and mutually interchangeable. They are typified by an internal protection (called SPIKE). In practice, the difference between them is the power output. The series was described on the basis of the LM3886 in an earlier issue*.

The PCB has been designed so that it can accommodate the LM3876 (50 W) as well as the LM3886 (150 W). Because of this, pin 5 of the IC on the board is connected to the positive supply line. This connection is not needed for the LM3876, since its pin 5 is not (internally) connected (NC).

The IC is located at the side of the board to facilitate fitting it to a heat sink as shown in the photograph.

An important aspect for optimum performance is the decoupling of the unregulated supply lines by C_7–C_{10}. All earth connections go to a single terminal on the board.

Air-cored inductor L_1 consists of 13 turns of 1 mm dia. enamelled copper wire with an inner diameter of 10 mm. The completed inductor is pushed over R_7 and its terminals soldered to those of the resistor.

All electrolytic capacitors must be mounted upright.

The amplifier can be muted with a single-pole switch connected to the MUTE input (pin 8). This function is enabled when the switch is open. If muting is not required, solder a wire bridge across the mute terminals on the board.

Boucherot network R_6-C_6 is not normally required in this application, but provision has been made for it for use in other applications.

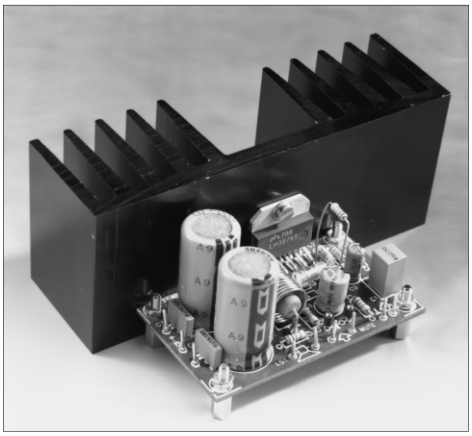
According to the manufacturers, both chips are optimalized for a load of 8 Ω. The output power is lower when a 4 Ω load is used or when the supply voltage is reduced. When a 4 Ω load is used, the SPIKE protection becomes active when the supply voltage is about 27 V, resulting a in a reduction of the power output to 10 W. This means that it is not advisable to use loudspeaker with an impedance <8 Ω.

Parts list

R_1, R_3 = 1 kΩ
R_3, R_4, R_5 = 18 kΩ
R_6 = see text
R_7 = 10 Ω, 5 W
R_8, R_9 = 22 kΩ

Capacitors:
C_1 = 2.2 µF, polypropylene, pitch 5 mm
C_2 = 220 pF, 160 V, polyester
C_3 = 22 µF, 40 V, radial
C_4 = 47 pF, 160 V, polyester
C_5 = 100 µF, 40 V, radial
C_6 = see text
C_7, C_8 = 100 nF
C_9, C_{10} = 1000 µF, 40 V, radial

Inductors:



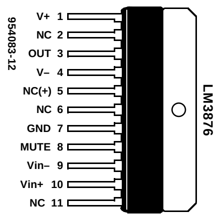
L_1 = 0.7 µH – see text

Integrated circuits:
IC_1 = LM3876T

Miscellaneous:
Heat sink for IC_1 <1.5 K W^{-1},
e.g. SK71/50 (Dau 01243 553 031)
Single-pole switch – see text
PCB not available ready made

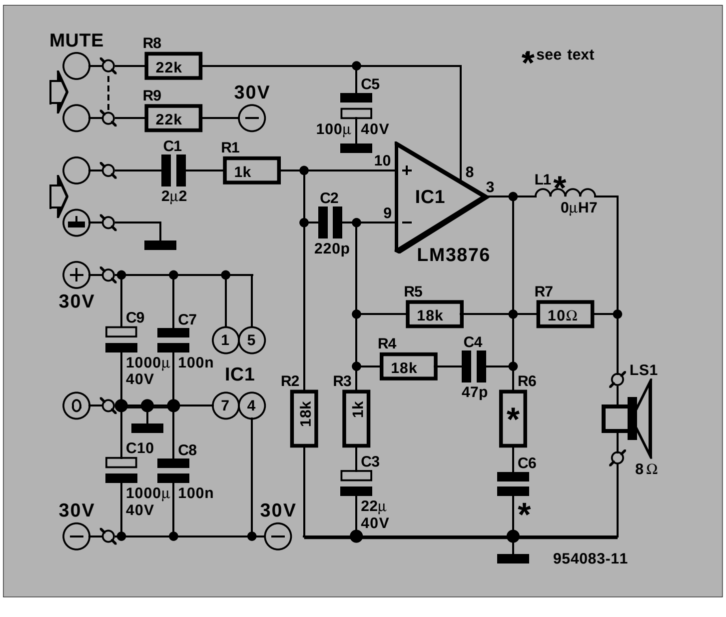
Design by T. Giesberts
[954083]

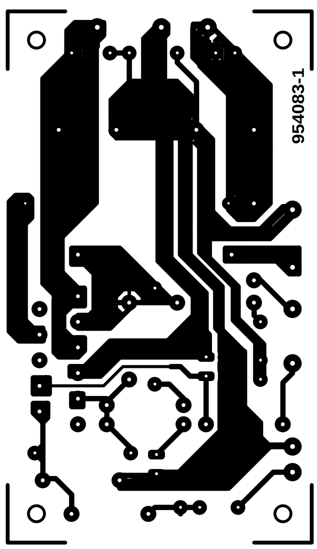
*May 1995.

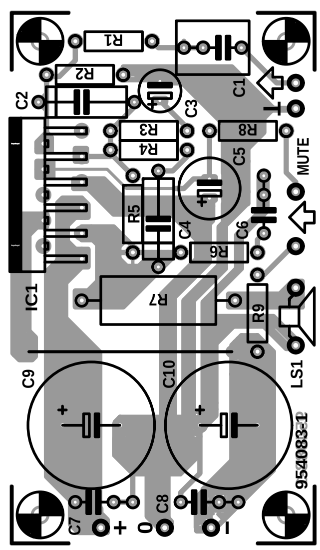


Main parameters

Input sensitivity	1 V r.m.s.
Output power	43 W into 8 Ω (THD + N = 0.1%)
Damping factor (8 Ω)	350 at 1 kHz; 220 at 20 kHz
Slew rate	11 V μs^{-1}
Power bandwidth	8.5 Hz – 117 kHz
Signal-to-noise ratio	>95 dB (linear 22 Hz – 22 kHz)
	>98 dBA







FRONT/REAR CAR RADIO FADER

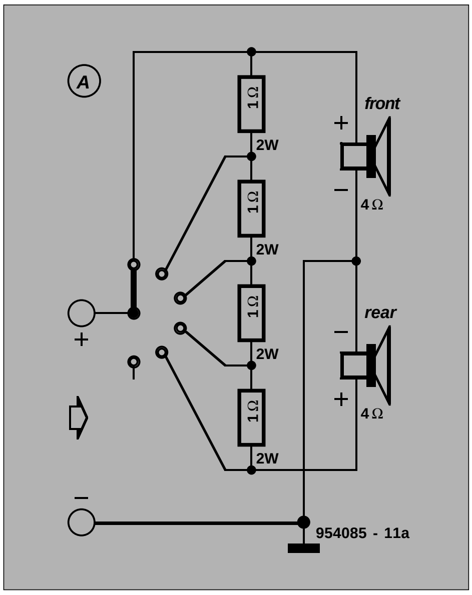
Although most car radio/cassette players produced in the past five years or so are fitted with a front/rear speaker volume control, there are still many about that have no such provision. A simple way of adding this facility is shown in diagram *A*, which is the principle on which many car radio faders operate. This arrangement gives a stepped front/rear speaker volume control with step ratios between 4:1 and 1:4. (Many modern car radios have a continuously variable control, however. Editor).

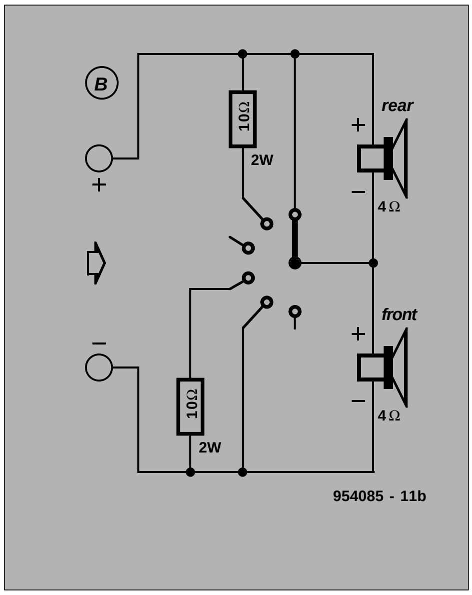
The arrangement has a few drawbacks in that power is dissipated (unnecessarily) in the resistors, which lowers the overall efficiency (down to 83% in positions 1 and 5, but as low as 67% in position 3), and the load impedance is lowered to 3 Ω in position 3 and to 2.67 Ω in positions 1 and 5.

A better arrangement is shown in diagram *B*. Since this is a series circuit, the load impedance can not drop to the dangerously low values met in *A*: it varies from 4 Ω in positions 1 and 5 to 8 Ω in position 3. Since the dissipation in the resistors is smaller, the efficiency is higher than in *A*: only in positions 2 and 4 does this drop to 88% – in the other positions it is 100%.

The volume control proper is similar to that in *A*, but in the extreme positions one speaker is on full, while the other is short-circuited.

Make sure that the switch can handle the power of the car radio.





A:

position	front/rear ratio
1	4:1
2	2:1
3	1:1
4	1:2
5	1:4

B:

position	front/rear ratio
1	∞
2	2:1
3	1:1
4	1:2
5	∞

Design by J. Seyler
[954085]

ACTIVE POTENTIOMETER

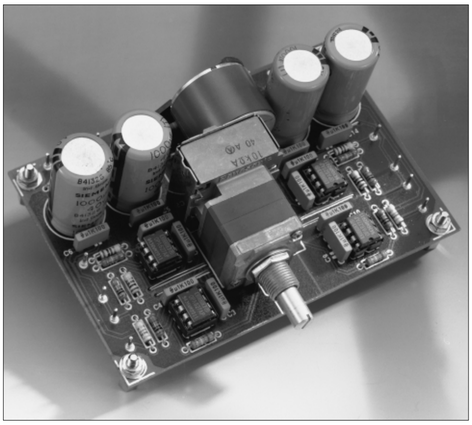
The potentiometer introduced by Panasonic a little while ago is of a quality exceeded only by the likes of the Penny & Giles potentiometer (which cost in excess of £ 100). The Panasonic devices have multilayer tracks made from conductive plastics and carbon, which are linked to the terminals by silver electrodes. The five-fold wiper is also made of silver and guarantees high accuracy (tracking within 0.8 dB) and smooth operation. In other words, this is an attractive, reasonably priced, high-quality volume control.

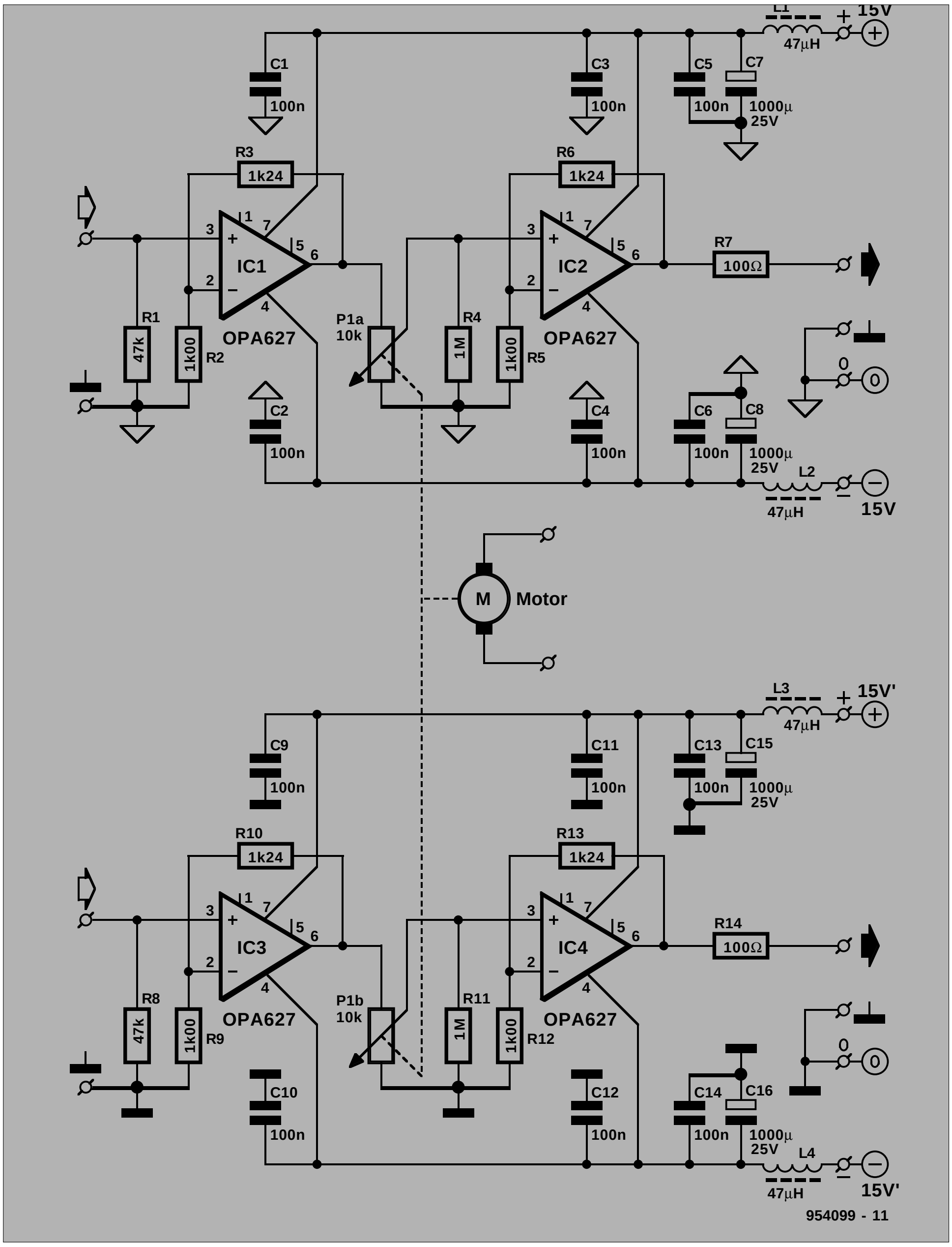
The potentiometer is a standard device which is preceded by an input amplifier and followed by an output buffer. It can be inserted into a line connection, so that appliances that have no volume control can be expanded to complete control amplifiers.

With the component values specified in the diagram, each op amp amplifies ×2.24 to give a total amplification per channel of ×5. This is sufficient to raise the line level of 200 mV to the standard output amplifier input level of 1 V. It is possible to alter the amplification to some extent, but it is advisable to carry any changes only to the buffer stages (IC_2 and IC_4). For example, the amplification of IC_2 is $1+R_6/R_5$. In most applications, this will do fine. With an input signal of 2 V (for instance, from a CD player), there is still a headroom of 6 dB.

If there is a need to add a selector switch at the input, R_1 and R_8 may be omitted. Bear in mind, however, that it must be possible for a bias current to flow.

The PCB allows the use of the Panasonic potentiometers and models from Alps, motor-driven as well as manually operated types. The board provides complete electrical isolation of the two channels. Moreover, signal earth and the negative supply line have been kept as far apart as feasible: they are linked only at the buffer capacitors. These arrangements prevent any effect of decoupling currents on the signal quality.





Moreover, r.f. decoupling capacitors and chokes (L_1–L_4) in the supply lines prevent any spurious products entering the signal processing circuits.

The circuit is highly suitable for being combined with the IR volume control published earlier*.

Parts list

Resistors:
R_1, R_8 = 47 kΩ
R_2, R_5, R_9, R_{12} = 1.00 kΩ, 1%
R_3, R_6, R_{10}, R_{13} = 1.24 kΩ, 1%
R_4, R_{11} = 1 MΩ
R_7, R_{14} = 100 Ω
P_1 = 10 kΩ logarithmic stereo (motor-driven) potentiometer

Capacitors:
C_1–C_6, C_9–C_{14} = 100 nF
C_7, C_8, C_{15}, C_{16} = 1000 µF, 25 V, radial

Inductors:
L_1–L_4 = 47 µH

Integrated circuits:
IC_1–IC_4 = OPA627AP

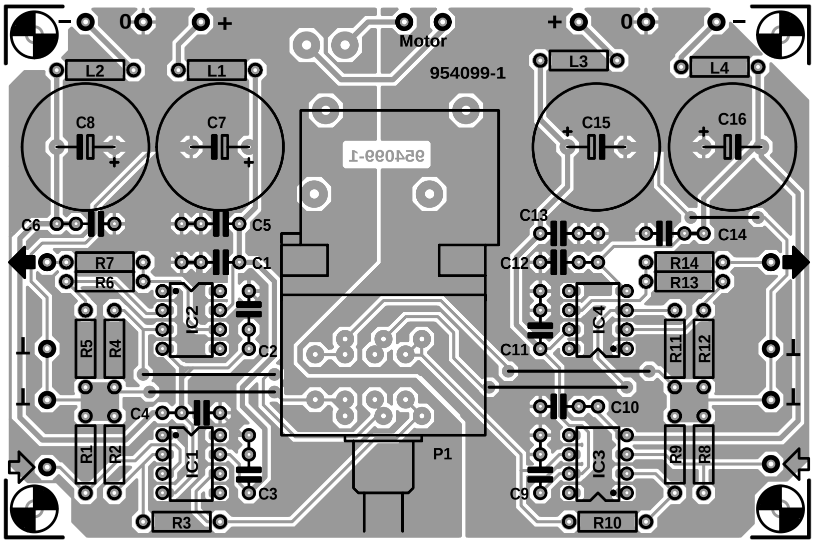
Miscellaneous:
PCB order no. 954009 (see p. 70)

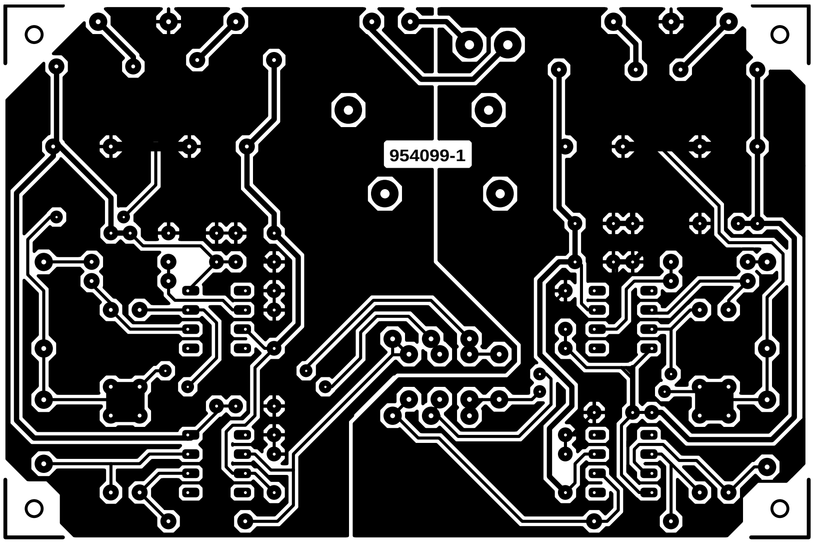
* July/August 1994

Design by T. Giesberts
[954099]

Parameters
(measured with U_{in} = 200 mV and U_B = 15 V)

Nominal output voltage	1 V r.m.s.
Maximum input voltage	4 V r.m.s.
Maximum output voltage	9 V r.m.s.
THD+N	
(bandwidth 80 kHz, 1 kHz, 1 V out)	0.0011%
(bandwidth 80 kHz, 20 kHz, 1 V out)	) 0.0012%
THD	
(2^{nd} + 3^{rd} harmonic, 1 kHz, 1 V out)	0.00012%
(2^{nd} + 3^{rd} harmonic, 20 kHz, 1 V out)	0.00054%
Signal-to-noise ratio	
P_1 at max (22 Hz – 22 kHz)>	>106 dB (108 dBA)
0.5 V out (22 Hz – 22 kHz)	>94 dB (95 dBA)
Crosstalk	
(20 Hz, 1 V out)	–140 dB
(20 kHz, 1 V out)	–115 dB
(20 Hz – 20 kHz, 0.5 V out)	–75 dB
Tracking error P_1	
(up to –60 dB)	<0.8 dB
(–60 dB to –80 dB)	<1–3 dB
Bandwidth	
(0.5 V out	2.7 MHz
(1 V out)	9 MHz
Slew rate	19 V µs^{-1}
Current drawn per channel	
(4 V in)	15.5 mA





surround-sound subwoofer part 1

Most surround-sound installations use loudspeaker boxes of modest dimensions so as to avoid making them too obtrusive for the usual living room. The consequence of this is a limited bass response, whereas especially the low frequencies can provide impressive effects with a good surround-sound system. To counter this drawback, the bass response can be enhanced with the subwoofer described in this three-part article.

Design by T. Giesberts

Surround sound, the popular audio craze of the past few years, can provide an impressive combination of sound and picture when it is used in conjunction with a TV set. Good-quality spatial sound is provided by a number of loudspeakers (usually five) located in front of and behind the listener(s). Five loudspeakers present a problem, of course, in that they take up a lot of space in the average living room. To keep the space occupied by them to a minimum, the loudspeakers are often fairly small. Moreover, in economy-price systems, cost is important, too, and this also tends to keep the boxes small.

Unfortunately, small loudspeaker boxes are detrimental to good bass reproduction. On the surface, this may not seem such a terrible thing in an audio-visual system until it is realized that the low frequencies contain spatial information. Moreover, we perceive low frequencies not only via our ears, but also through our entire body and this causes good low-frequency reproduction to give that added feeling of reality to the sound. All this makes it clear that the importance of low frequencies must not be underestimated.

The reproduction of low frequencies requires the displacement of large volumes of air. This in turn means that a large low-frequency drive unit (woofer) should be used. But such a unit must be contained in a large enclosure to enable it to reproduce low frequencies effectively. And this is where the crux of the matter is: most living rooms just do not have the space for such a large box.

In the subwoofer described in this article an attempt has been made to find a compromise between the contradictory requirements just outlined. It uses a large (300 mm) drive unit

Technical data

Drive unit	*300 mm (8 in), e.g. Monacor (SPH-300TC); KEF; Radio Shack (40-1024); Parts Express (295-240)*
Dimensions of box	*660×406×420 mm (26×16×16 9/16 in) incl. legs*
Volume of box	*about 65 l net*
Type of box	*bass reflex*
Nominal impedance	*8 Ω per channel*
Efficiency	*88 dB W^{-1} m^{-1}*
Frequency range	*45–105 Hz*
Loading	*max 250 W per channel*

housed in a modestly-sized enclosure of 65 *l*. The enclosure is designed in the form of a side table with the drive unit fitted between the legs so as to make it (virtually) invisible. The volume of the enclosure is not really large enough for very low frequency reproduction, but a solution for this will be published in next month's instalment. This consists of an active correction network and associated amplifier that bring the –3 dB point down to 20 Hz. This article describes the passive version of the subwoofer which can be used without any difficulties with existing apparatus. Its frequency range extends from about 45 Hz to 105 Hz. The upper frequency and the efficiency of the unit provide a good match with the (smaller) front loudspeakers.

Although so far reference has been made only to a surround-sound system, the subwoofer may, of course, also be used with a standard stereo sound system.

THE (PASSIVE) DESIGN

The design is based on a 300 mm (8 in) Monacor SPH-300TC drive unit, but other makes, such as KEF, Radio Shack (40-1024), or Parts Express (295-240) should give good performance as well. The SPH-300TC is a relatively inexpensive unit with a fairly large magnet that displaces a volume of around 0.2 *l*. Its parameters make it suitable for use in a bass reflex enclosure.

If the loudspeaker is to be used with a stereo system, it should have connections for both channels. This means that either two drive units or a drive unit with dual voice coil should be used. Each voice coil is connected to one of the channels via a suitable filter. The present design uses the latter solution, since the use of two drive units would make the box unnecessarily large.

The alignment of the enclosure is determined with the simulation program Boxcalc, and aims to arrive at a compromise between a (relatively) small volume and a low –3 dB point. This results in a 65 *l* box with the pipe (acoustical resonator) tuned to 23 Hz. The overall frequency response is shown in **Fig. 1**. The –3 dB point is at 45 Hz, which, considering the small box volume, is pretty good. The –3 dB point is low enough to allow the subwoofer to be used as a passive unit with most existing systems.

1

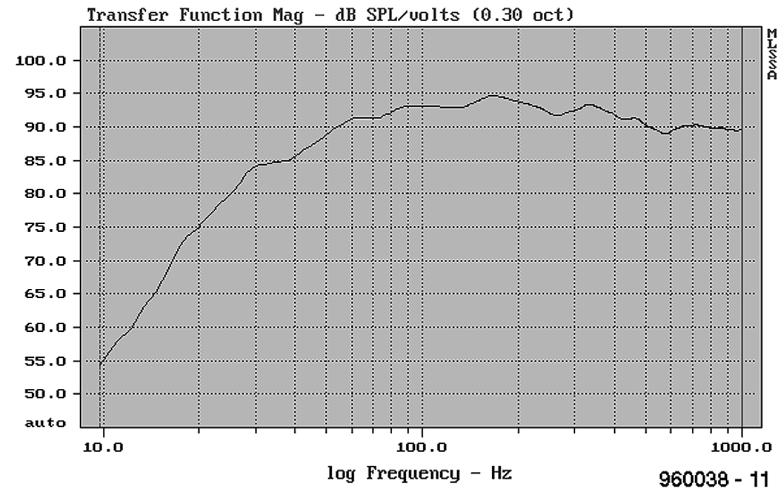


Fig. 1. The frequency response curve of the SPH-300TC in a 65 l bass reflex enclosure tuned to 23 Hz.

THE FILTER

Since the design aims at keeping the costs as low as feasible, the (passive) filter has been kept as simple as possible, which, in the case of a subwoofer, is not as easy as it may seem.

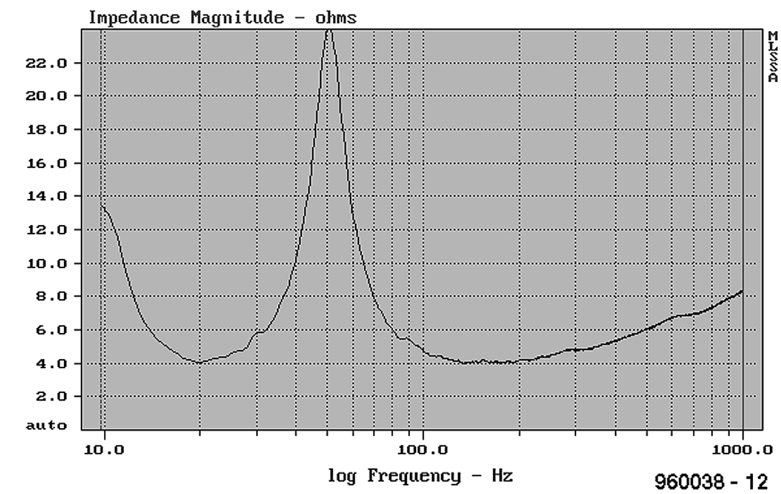
The impedance characteristic of the drive unit is shown in **Fig. 2**. The two voice coils are connected in parallel to obtain a reliable curve (which means that for each coil double the impedance value must be taken). The curve shows two peaks. The lower one at about 10 Hz results from the bass reflex alignment (which, by the way, is exactly in line with the 23 Hz resonator). The second peak, just above 50 Hz, is caused by the resonance frequency of the drive unit in the box.

Normally, filtering of a subwoofer starts at around 100 Hz or slightly lower to ensure good matching with the standard stereo loudspeakers. A passive filter, however, has the drawback that it functions properly only if it is terminated into a pure resistance. If the cut-off point were chosen at 100 Hz, the 52 Hz peak would create a problem: the resulting overall curve of a theoretically computed filter would not be usable. To solve this problem, the impedance curve of the drive unit has to be corrected. This is often effected by connecting parallel across its terminals (for each channel) an *RLC* network with the same resonant frequency. Unfortunately, at such low frequencies, the values of the necessary inductors and capacitors are such that they result in physically large (and expensive) components.

The solution in the present design consists simply of shunting the voice coil with a resistor. This does not totally eradicate the peaks, but

Fig. 2. The impedance curve when the two voice coils are in parallel. The high peak poses a problem for the passive filter.

2



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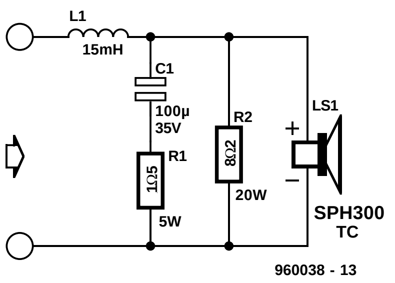


Fig. 3. The filter has been kept simple. Resistor R2 corrects the impedance curve. Inductor L1 and capacitor C1 provide a slope of 12 dB per octave and a high cut-off point at about 105 Hz.

flattens them sufficiently to enable a simulation program—Calsod—correcting the filter such that its frequency response is close to requirements.

To keep the number of components small, the filter is a second-order type consisting of conductor L_1 and capacitor C_1 (see **Fig. 3**). The resistor in series with the capacitor damps the *LC* circuit to some extent. The effect of the filter is shown in **Fig. 4**. Although the high cut-off point is about 105 Hz, the response will ensure a good match to most small loudspeakers.

BUILDING THE BOX

The prototype box is made from 28 mm thick medium-density chipboard (MDF), but, as in some cases it may not be possible to obtain this, 22 mm thick chipboard may be used (note that the dimensions in **Fig. 5** must then be adapted as appropriate). The box consists of six rectangular sheets and a stiffening crosspiece which are firmly fixed together with a suitable heavy-duty glue.

At one side are the apertures for the drive unit and acoustical resonator. The resonator consists of a 365 mm long piece of 80 mm dia. PVC pipe available from a builders merchant.

The four banana sockets for connecting the cables from the amplifier are fitted at the bottom of the one of the side panels.

The box is designed to rest on four 50 mm high legs with the drive unit fitted at the bottom facing the floor of the living room.

After the glue has dried thoroughly and the material has been sand-papered, the box can be given a final coat to individual taste.

The box is half filled (up to the cross piece) with suitable loudspeaker wadding, but take care that the opening of the pipe remains reasonably free of it.

The filter components are available from a specialist audio/hi-fi retailer or a good electronics shop. The inductor is a 15 mH type with a 56 mm ferrite core, preferably an HQ56 from IT. The capacitor is a bipolar type with smooth terminals.

The filter components may be glued to a small sheet of wood,chipboard, or prototyping board and then wired together. Note that some retailers stock general-purpose filter boards.

Screw the completed filter into the box and wire it up as shown. Take care not to interchange the plus and minus connections to the two channels. The cables to the drive unit must be terminated into cable clips to avoid the necessity of having to solder to the drive unit terminals.

Place the resonator in position, make the connections to the drive unit (make sure that the connections to the + terminals match, otherwise the drive unit does not work). Finally, place a strip of draught-excluding tape under the rim of the drive unit and screw the unit to the box.

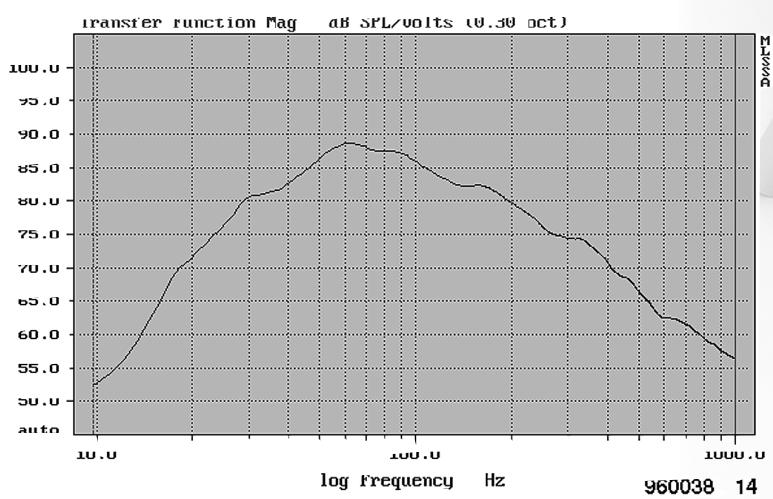
Some constructors (or their wives) may find it aesthetically pleasing to place a sheet of glass, marble or similar material on top of the box to give it the look of a side table.

The passive subwoofer is then ready for use. It may be connected in parallel with the existing stereo speakers. It will work most satisfactorily when its efficiency of about 88 dB W^{-1} m^{-1} corresponds roughly to that of the existing loudspeakers and it is placed in close proximity to these. Note that if you want the active version, which will be described next month, you do not need the passive filter; the box remains the same.

(960038)

Fig. 4. The frequency response of the loudspeaker and filter combination. It ensures correct matching to most smaller stereo loudspeakers.

4



Drive unit revisited

During the design of the loudspeaker, a thorough search was made for a 300 mm drive unit at a reasonable price (to keep total costs down to not more than £ 80–90). Of course, such an economy-price unit cannot be expected to be perfect. And, indeed, in the testing of the SPH-300TC unit, it appeared that the parameters stated by the manufacturer did not agree with our own measurements. Fortunately, the deviations were beneficial to the box dimensions. Also, there was a kind of rustling noise at large cone movements. This was suspected in the first instance to be caused by a loose cone or air leak, but a second example exhibited exactly the same noise. A detailed investigation showed that the dust hood in the cone (the convex cap that closes the upper side of the cone) was the culprit. Its material is fairly soft, so that at large cone movements it begins to vibrate at its own (higher) frequency and thus causes the rustling noise. This deficiency is easily negated by spraying the dust hood a couple of times with a suitable plastic spray or applying a few layers of a suitable cone impregnator. This makes the cap more rigid so that it is not set into vibration at large cone movements.
The Parts Express unit appears to be rather more rugged than the Radio Shack and is rather cheaper.

5

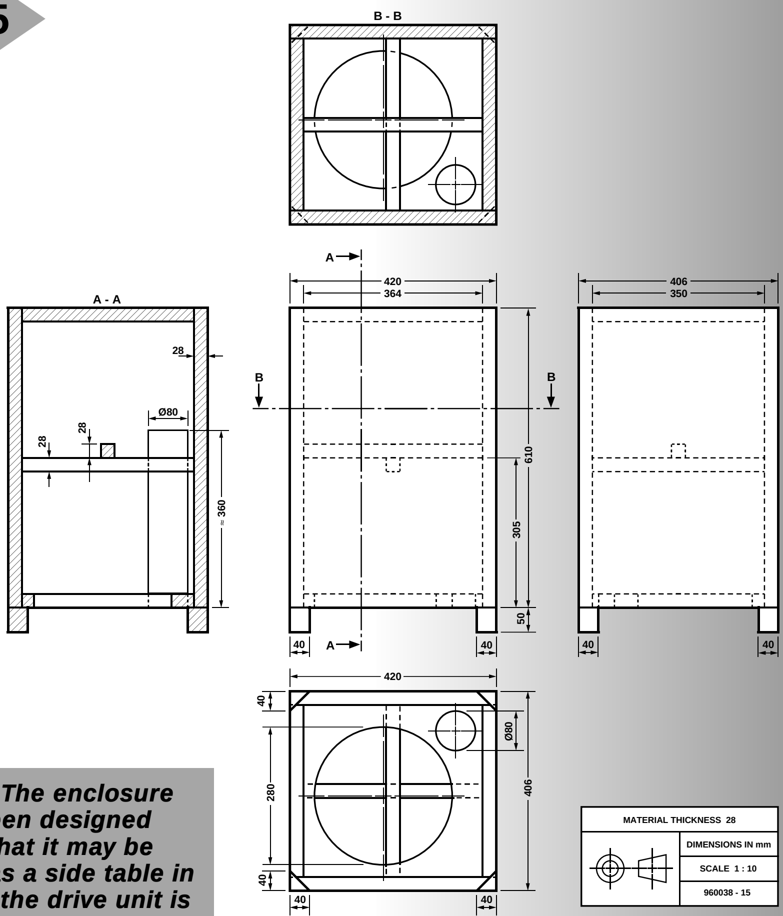


Fig. 5. The enclosure has been designed such that it may be used as a side table in which the drive unit is not (or hardly) visible.

Contents

surround-sound subwoofer Part 2

This active version of the subwoofer described in last month's instalment is a plus for virtually any hi-fi system. Where the low cut-off frequency of the passive version is around 40 Hz, it is down to about 20 Hz in the active subwoofer. With its integral 240 W amplifier, it is the answer for those seeking a realistic bass foundation for their system. The more so, since its building cost is very reasonable.

Design by T. Giesberts

In Part 1 of this article, the benefit was explained that a subwoofer may have for realistic reproduction of hi-fi sound, particularly in audio-visual systems with surround sound. So, what is the value added of this active version, it may be asked. Is a cut-off frequency of around 40 Hz not sufficient for good (bass) sound reproduction?

The answer is yes and no: it depends what you want. For most music reproduction 40 Hz is a good figure: it corresponds roughly with the lowest tone of a double bass. Loudspeakers that can reproduce this frequency with good sound pressure are few and far between. Nevertheless, there are a.f. signals where 40 Hz is not sufficient. This is the case, for instance, when the canon fire in Tchaikovsky's '1812' is to be reproduced faithfully, or when thunder claps are to sound realistic. Also, the sound tracks of films like *Jurassic Park* and *Top Gun* gain in reality if the audio range goes down well below 40 Hz.

Technical data

✔ Drive unit	300 mm (8 in), e.g. Monacor (SPH-300TC), KEF, Radio Shack (40-1024); Parts Express (295-240)
✔ Type of enclosure	Bass reflex
✔ Box dimensions	660×406×420mm (incl.legs) 26×16×16 9/16 in
✔ Volume of box	65 l
✔ Frequency range	20 Hz to 40 Hz, 50 Hz, 60 Hz or 70 Hz (as selected)
✔ Cross-over frequency	40 Hz, 50 Hz, 60 Hz or 70 Hz (as selected)
✔ Power output	245 W into 4 Ω (thd = 0.1%) 130 W into 8 Ω (thd = 0.1%)
✔ THD + N at 100 Hz	at 1 W into 8 Ω: 0.0046% at 50 W into 8 Ω: 0.001% at 1 W into 4 Ω: 0.007% at 100 W into 4 Ω: 0.0016%
✔ Signal to noise ratio	90 dB linear (93 dBA) at 1 W into 8 Ω
✔ Damping factor	>400 (with 4 Ω load)

Although the question may be asked how far down to go, to which the answer is ‘the further the better’, a sensible, practical limit appears to be about 20 Hz. This is because the threshold of human hearing is at around that figure. Lower frequencies are ‘felt’ rather than heard (to hear them would require a battery of loudspeakers that could not be accommodated in the average home. Moreover, even if it could, the possibility of damage to the building at the required volume is not imaginary).

If the cut-off frequency is set at 20 Hz, very good low-frequency reproduction is possible, while the required air displacement can be achieved with normal means. However, with a passive system, this would required an enclosure of a couple of hundred litres, and that again would be unacceptable in the average home. Therefore, what is required is an …

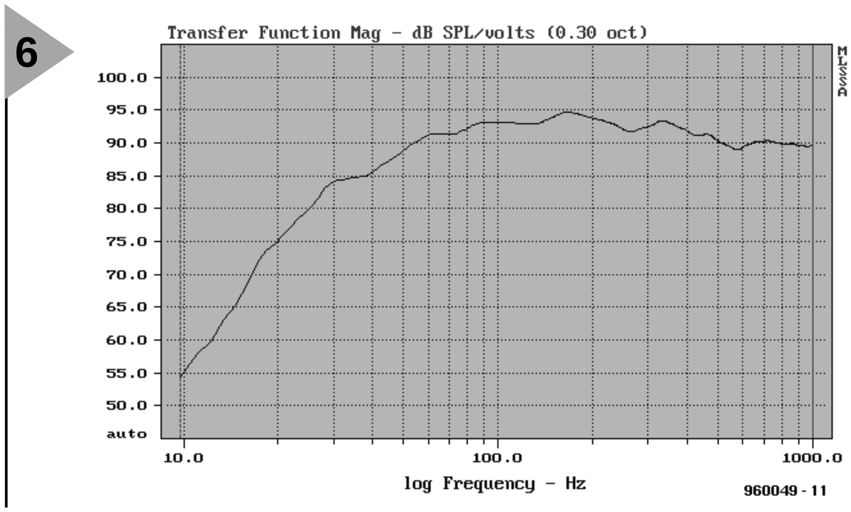


Figure 6. Frequency characteristic of the 300 mm drive unit in its base reflex box, without filter and without correction.

ACTIVE DESIGN

The most notable difference between an active and a passive loudspeaker is the amplifier in the former. In a multiple system, two or more would be needed, but fortunately only one in a subwoofer. The fact that an active design has its own amplifier makes it easily brought into line with the loudspeakers in the system into which it is being introduced.

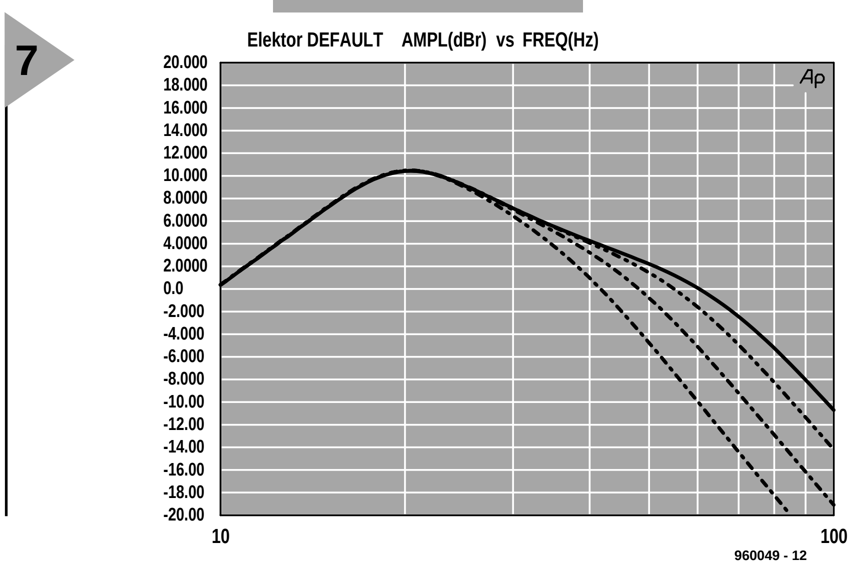
Another beneficial aspect of an active design is that the necessary filtering can take place before the power amplifier. This filtering is carried out electronically, which has the advantage of offering virtually limitless opportunities for correcting or manipulating the frequency response of the drive unit.

In the present design, these opportunities are taken gratefully, since they allow a relatively small enclosure to reproduce frequencies down to 20 Hz. This is done by measuring the response of the drive unit in its (too small) enclosure and creating a filter with a mirror image of that response. This results in the filter compensating the irregular response of the box until a straight response curve is obtained.

The response of the passive loudspeaker described in Part 1 (using the Monacor drive unit) is shown in **Figure 6**. It will be recalled that the volume of the enclosure is 65 l. The cut-off frequency is about 45 Hz, but a close look at the curve shows that the response begins to roll off at around 85 Hz already. The curve becomes slightly steeper at about 60 Hz and even more so at 30 Hz. The latter is a direct result of the bass reflex vent: above the vent frequency, the roll off occurs at 12 dB/octave (second order) and below it, at 18 dB/octave (third order).

The active design has a basic frequency range of 20–70 Hz. To straighten the response curve, the electronic filter must have a response as shown in **Figure 7**. This curve peaks at 20 Hz (note that it begins to straighten out between 30 Hz and 20 Hz). There are four curves in the figure, because the filter is designed with four switched (upper) cut-off frequencies. This makes it easier for the loudspeaker to be combined with existing systems. Actually, we have jumped ahead slightly, because Figure 7 shows the responses of the active part of the subwoofer.

Figure 7. Response of the combined correction filter and cross-over filter. The four curves refer to (upper) roll-off frequencies of 40 Hz, 50 Hz, 60 Hz and 70 Hz.



The acoustic end result of the design is shown in **Figure 8**, which shows that the response is virtually straight between 20 Hz and 70 Hz. The solid curve is that of the actively corrected subwoofer and is obtained with a standard microphone and spectrum analyser. Comparing this curve with that of Figure 6 shows immediately the enhancement provided by the added electronics. The dotted curve is obtained when the (upper) cut-off frequency is set to its lowest value of 40 Hz.

DESIGN CONSIDERATIONS

The active subwoofer is based on the 30 cm drive unit specified in Part 1, and is housed in the same bass reflex box described in that instalment. The traditional cross-over filter is not used in the active design: it is re-

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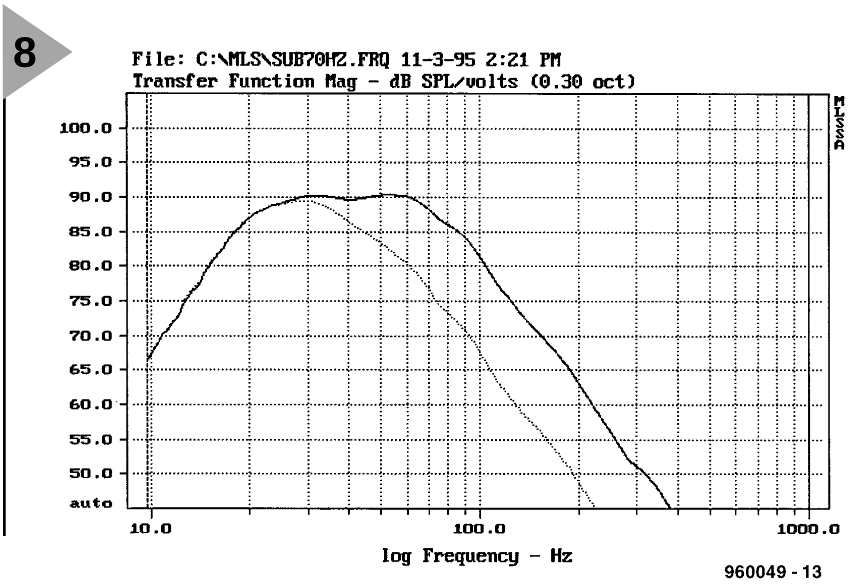

Figure 8. After correction, the frequency response curve of the active subwoofer is straight from 20 Hz to 70 Hz. The dotted curve is measured with a roll-off frequency of 40 Hz.

placed by an electronic filter and a power amplifier.

The electronic filter is a combination of a correction filter and a cross-over filter. It straightens the response curve of the drive unit and can be switched to give one of four different (upper) roll-off frequencies.

Since the filter correction is no less than 10 dB at 20 Hz, the amplifier must provide a reasonable output power: in the present design, 240 W. The amplifier drives both voice coils of the drive unit, which are connected in parallel.

Since the electronic filter has line inputs as well as high-level inputs, the active subwoofer may be driven by a preamplifier (or via the pre-out terminals of an integrated amplifier) or via the loudspeaker terminals—see Figure 9.

The existing a.f. amplifier and the subwoofers must be linked by screened audio cable, not by loudspeaker cable.

The filter, output amplifier and the necessary power supply are housed in a common enclosure that is placed close to the loudspeaker or even fastened on to it.

THE FILTER

The circuit of the filter is shown in the diagram in **Figure 10**. It consists of four distinct parts: correction filter IC_{2d}, IC_{2c}; cross-over filter IC_{2b}, IC_{2a}; drive level indicator IC_3, T_1; and symmetrical power supply IC_4, IC_5.

Operational amplifier IC_{1a} functions as an up-counter for the left-hand and right-hand channels. Its amplification is varied with P_1. High-value resistors R_1 and R_2 ensure that loudspeaker signals can be processed without any difficulty.

The op amp is followed by the correction filter. This is a second-order low-pass type based on IC_{2d}. Its output is added to the unfiltered signal in IC_{2c}. Capacitors C_3 and C_4 limit the bandwidth (as does capacitor C_1 at the input of the amplifier—see **Figure 11**). The correction is enhanced by output buffer R_{28}-R_{29}-C_8, which enables the response of the loudspeaker to change gradually from second order to third order.

The cross-over filter is based on IC_{2b}. It is an active third-order low-pass Butterworth filter that can be set

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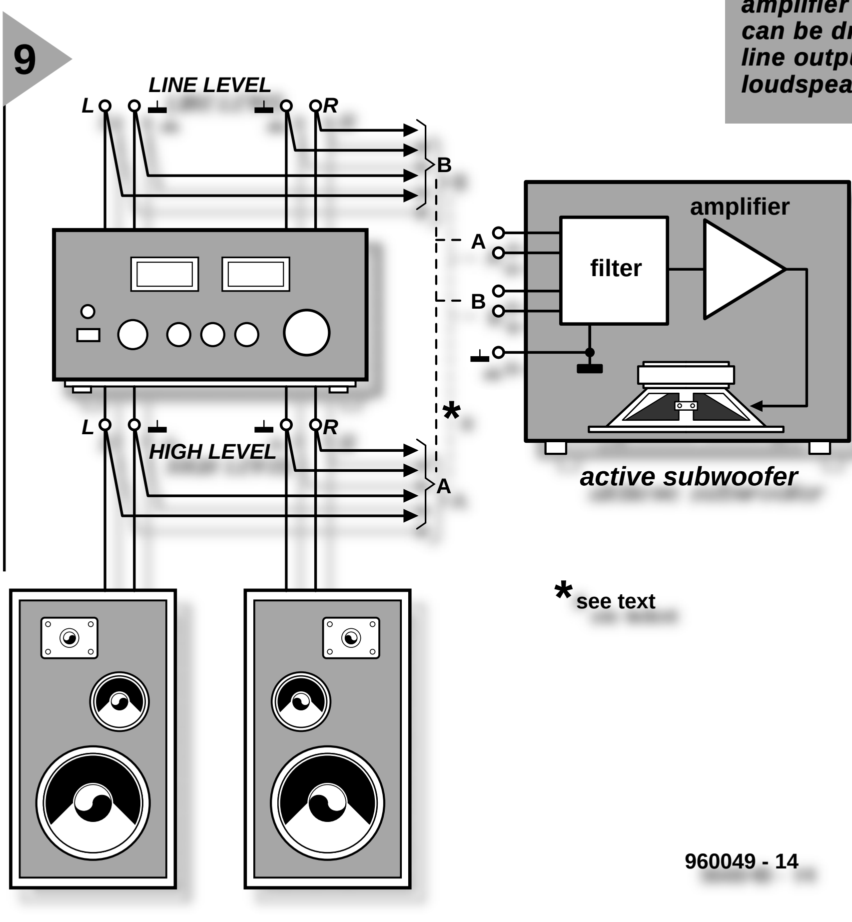
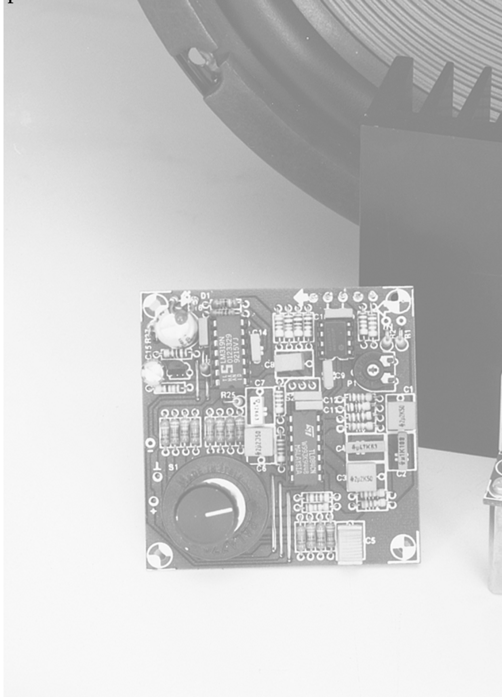


Figure 9. The active subwoofer is a combination of loudspeaker, amplifier and filter. It can be driven via a line output or via the loudspeaker output.



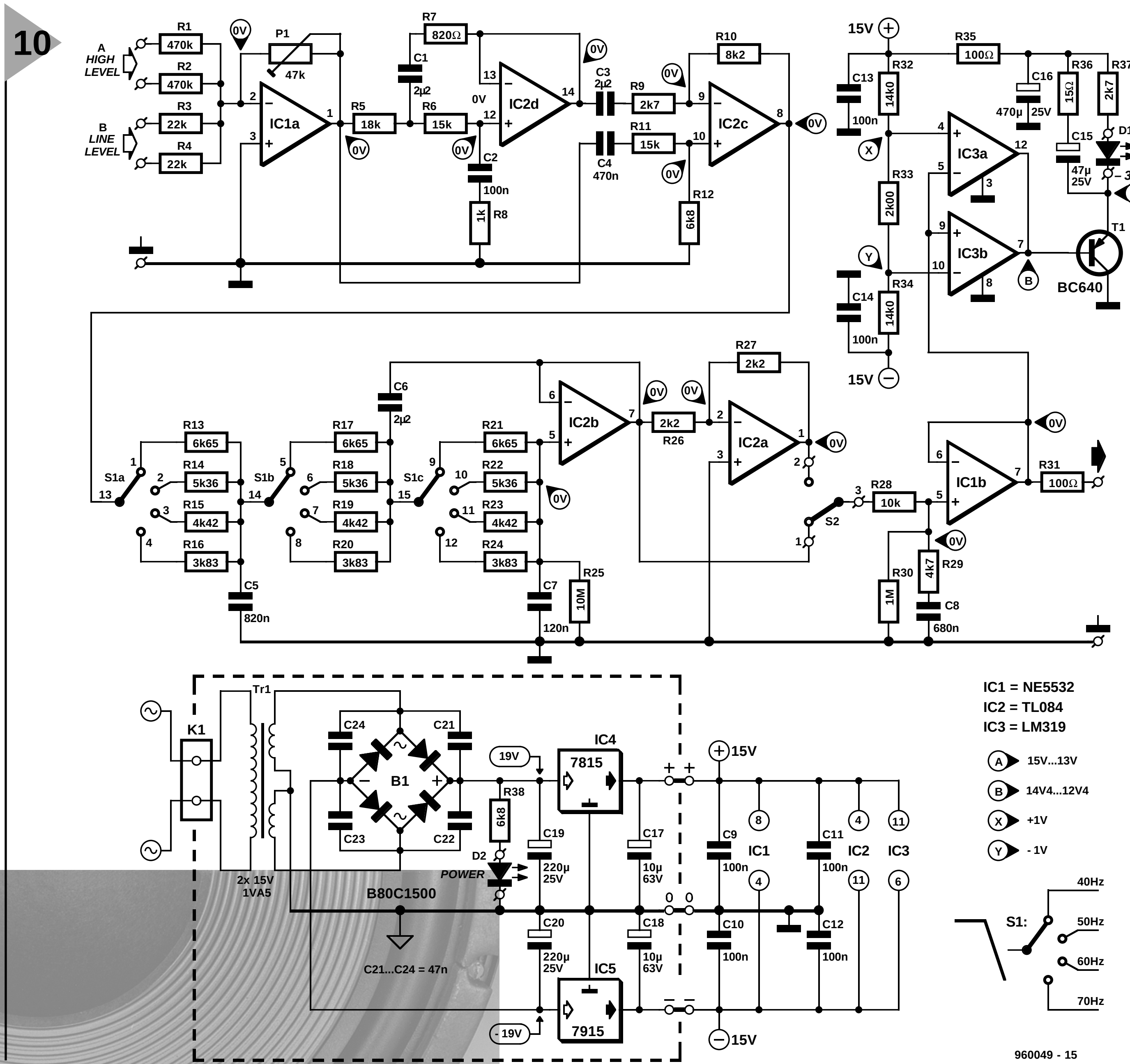
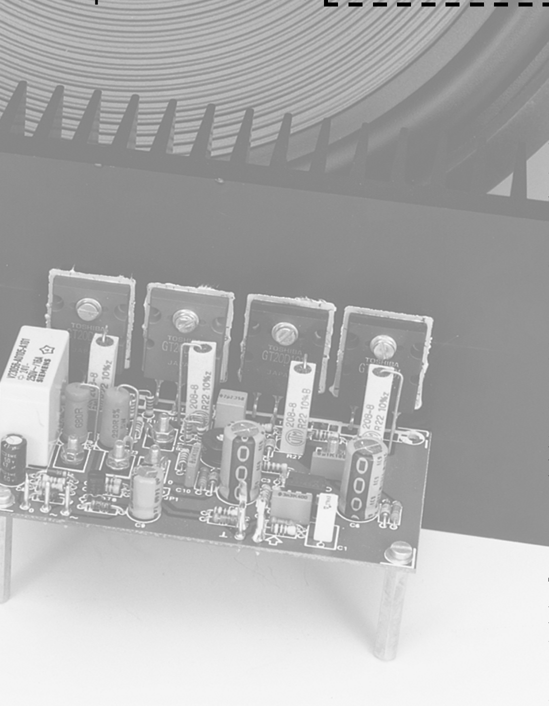


Figure 10. The circuit of the active filter is simplicity itself. The drive level indicator based on IC3 is a boon.



to any one of four different (upper) roll-off frequencies with S_1. With component values as specified on the circuit diagram, these frequencies are: 40 Hz, 50 Hz, 60 Hz, 70 Hz.

The filter is followed by inverter IC_{2a}, so that it is possible to select (with S_2) either the original signal or one that is 180° out of phase with it. This is often an advantage with certain loudspeaker systems.

The filtered signal is applied to the output via buffer IC_{1b}.

The drive level indicator, IC_3 and T_1, is intended as a protection for the loudspeaker: when the amplifier is driven to half its maximum output, diode D_1 lights. This optical signal is a warning to turn down the volume to some extent.

The indicator is based on IC_{3a} and IC_{3b}, which form a window comparator, which is designed such that the led lights when the output voltage of IC_{1b} exceeds a level of 1 Vpeak. Since the output amplifier has an input sensitivity of 1 Vrms, its drive remains about 3 dB below maximum (provided that the warning signal has been responded to).

The brightness of D_1 is enhanced by the high charging current (1 A) through C_{15} delivered by T_1. This also results in a certain amount of afterglow once the peak has passed. Network R_{35}-C_{16} decouples the power line, so that charging pulses do not cause any interference in the filter.

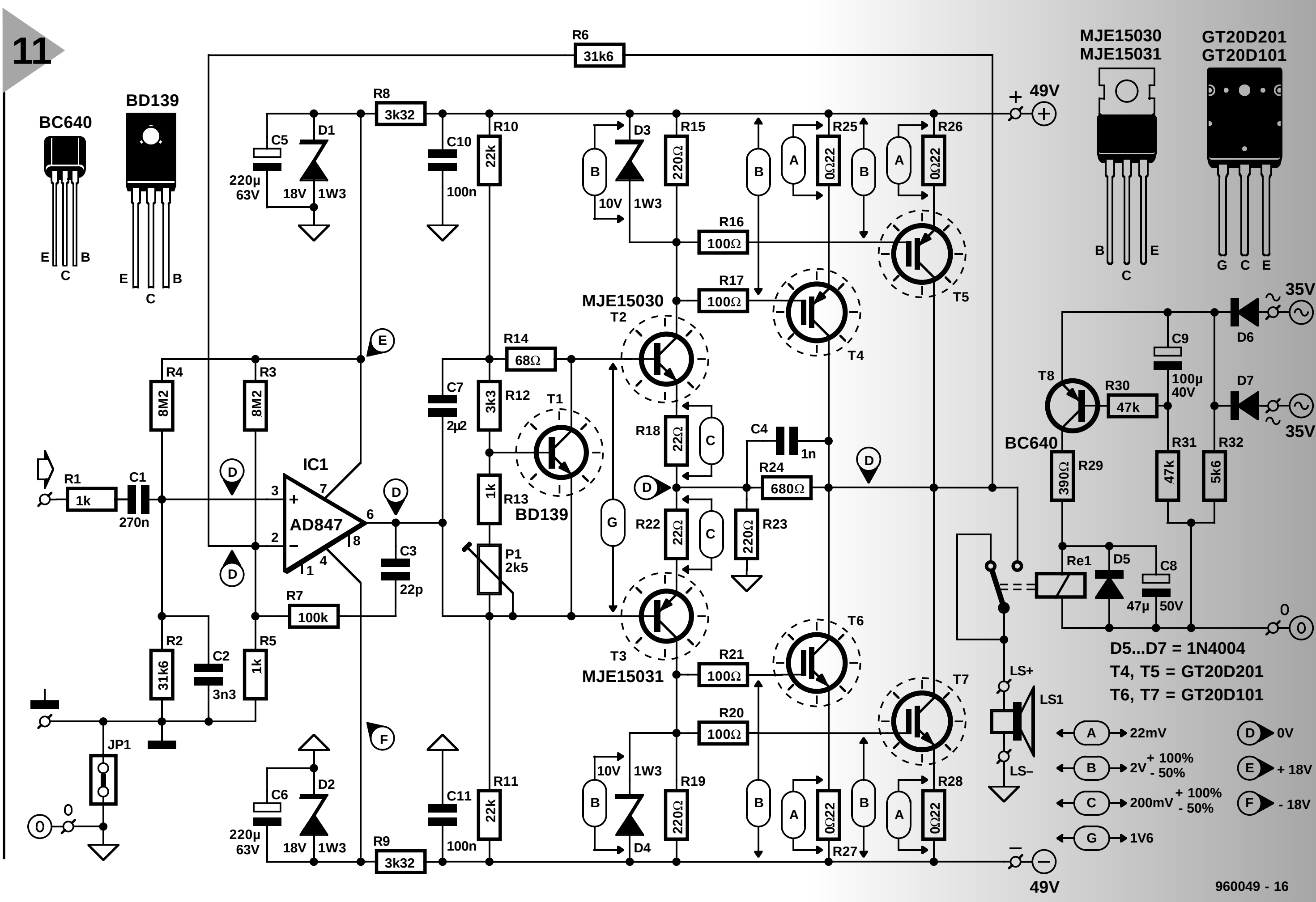


Figure 11. Since the output amplifier does not have to process frequencies above about 100 Hz, its design is spartan. In spit of this, its performance is excellent and its power is sufficient to drive the subwoofer to the very limits of its loadability.

The symmetrical 15 V power supply is a traditional design: mains transformer, bridge rectifier, smoothing capacitors and two voltage regulators, IC_4 and IC_5. Diode D_2 is the on/off indicator.

The power amplifier

The output of the filter is coupled directly to the power amplifier whose circuit is shown in the diagram in **Figure 11**. Considering its output power, the amplifier is fairly compact and straightforward. The compactness is a conscious part of the design, while the simplicity is brought about by the fact that the amplifier needs to perform well only up to about 100 Hz.

The amplifier is a combination of an integrated voltage amplifier and a discrete current amplifier. Since the voltage amplifier needs to meet certain strict requirements, it is based on a very fast op amp (IC_1), the Type AD847 from Analog Devices. Its supply voltage has been made as high as feasible (±18 V) with the aid of zener diodes D_1 and D_2 to minimize the risk of overdriving.

The current amplifier is formed by two 'darlington-like' configurations, each consisting of a medium power driver, T_3/T_4, followed by two parallel-connected Insulated Gate Bipolar Transistors (igbts), T_4-T_5 and T_6-T_7. Network R_{23}-R_{24} ensures that the power stages not only provide current amplification, but also voltage amplification of ×4. This is necessary because IC_1 works from a supply of only ±18 V, whereas the output stages need to be driven to about ±45 V.

'Zener' transistor T_1 enables the correct setting of the quiescent current. For good quiescent-current stability, it is necessary that T_1 is fitted on to the same heat sink as the drivers and power transistors. The stage is designed so that it has a slightly negative temperature coefficient. This means that when the heat sink warms up, the quiescent current, set with P_1, drops a little so that the amplifier cools more quickly

Annoying and possibly damaging switch-on plops are avoided by the traditional relay, controlled by a delay circuit, in series with the loudspeaker. Transistor T_8 conducts only when C_9 has been charged to a certain level via R_{31}: that is, a few seconds after the supply has been switched on.

The delay circuit is powered directly by the secondary winding of the mains transformer. This has the advantage of the relay being deenergized immediately the supply is switched off and not after the reservoir capacitors in the power supply have been discharged.

Next month's instalment will deal with the construction.

(960049)

surround-sound subwoofer

Part 3 (final)

After the detailed descriptions of the passive and active versions of the subwoofer in the previous two instalments, this third and final part deals with the complete construction. If the enclosure has already been built on the basis of Figure 5, only the printed-circuit boards need to be completed and built into the enclosure together with the drive unit.

The circuits shown in Figures 10 and 11 (Part 2), including the power supply for the cross-over filter in Figure 10, are intended to be built on the three printed-circuit boards shown in **Figure 12.** The three sections shown in the illustration are easily separated from one another by cutting or snapping off along the indicated lines. The power supply for the power amplifier will be described later in this instalment.

Building the filter board is straightforward, as one would expect with only three ICs, a handful of resistors, and some capacitors on a good-sized board. Make sure to fit the six wire bridges first. Rotary switch S_1 may be mounted directly on to the board. Connect the two leds and switch S_2 to the board via short lengths of stranded circuit wire (7/029).

The ±15 V power supply for the filter does not take much more space than a box of matches. Since the filter draws only a modest current, the mains transformer can be kept small (1.5 VA). The voltage regulators need not be cooled. Connect terminals '+', '0' and '–' to the corresponding terminals on the filter board via short lengths of stranded circuit wire. The mains voltage from the central mains entry is connected to the supply via terminal block K_1.

Populating the power amplifier

Parts list:

FILTER

Resistors:

R_1, R_2 = 470 kΩ
R_3, R_4 = 22 kΩ
R_5 = 18 kΩ
R_6, R_{11} = 15 kΩ
R_7 = 820 Ω
R_8 = 1 kΩ
R_9, R_{37} = 2.7 kΩ
R_{10} = 8.2 kΩ
R_{12}, R_{38} = 6.8 kΩ
R_{13}, R_{17}, R_{21} = 6.65 kΩ, 1%
R_{14}, R_{18}, R_{22} = 5.36 kΩ, 1%
R_{15}, R_{19}, R_{23} = 4.42 kΩ

board is straightforward, but a few aspects need to be watched. Firstly, transistors T_1–T_3 must be fitted at the track side of the board. These devices, together with T_4–T_7, must be screwed to one and the same heat sink, in each and every case with the aid of insulating washers and bushes. To ensure maximum heat conduction, apply heat conducting paste to both sides of the washers for the IGBTs.

Secondly, as described in the previous instalment, the collectors of T_4–T_7 provide the output current in unison. To ensure that the transfer resistances are kept small and at the same time that the board does not become unduly warm, the connection between the collectors and the output relay is rather unusual. Immediately adjacent to the collector terminal is an additional hole on the board into which a solder pin is to be inserted (at the track side). The four pins must be interconnected by heavy-duty single-strand copper wire (≥ 1.5 mm²) or

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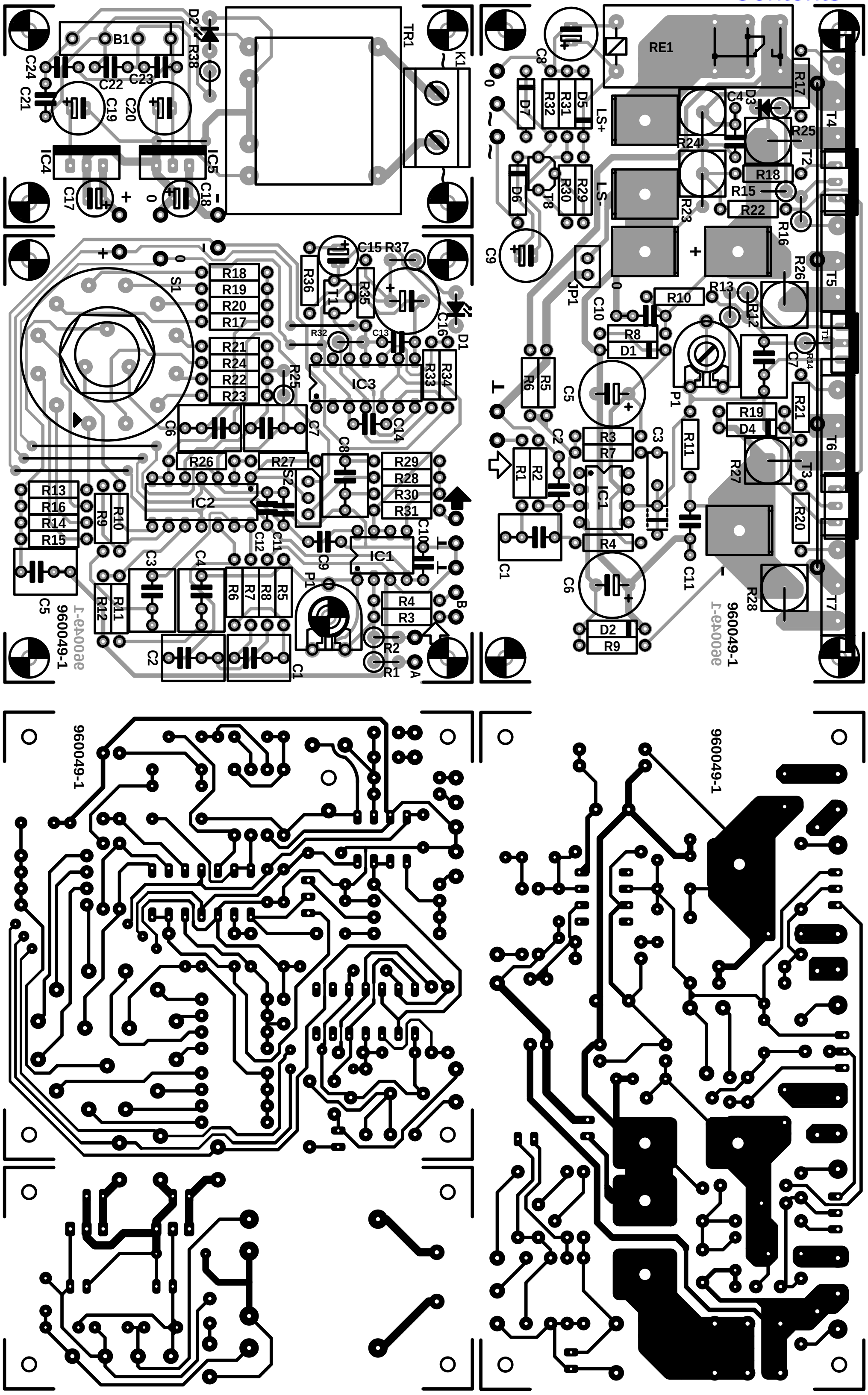

Figure 12. Because of the relative simplicity of the circuits, the amplifier and filter can be accommodated on one board; it is, however, advisable to separate the boards.

R_{16}, R_{20}, R_{24} = 3.83 kΩ
R_{25} = 10 MΩ
R_{26}, R_{27} = 2.2 kΩ
R_{28} = 10 kΩ
R_{29} = 4.7 kΩ
R_{30} = 1 MΩ
R_{31}, R_{35} = 100 Ω
R_{32}, R_{34} = 14.0 kΩ, 1%
R_{33} = 2.00 kΩ, 1%
R_{36} = 15 Ω
P_1 = 47 kΩ preset potentiometer

Capacitors:
C_1, C_3, C_6 = 2.2 µF, metallized polyester, pitch 5 mm
C_2, C_9–C_{14} = 100 nF
C_4 = 470 nF, metallized polyester, pitch 5 mm
C_5 = 820 nF, pitch ≤ 7.5 mm
C_7 = 120 nF
C_8 = 680 nF
C_{15} = 47 µF, 25 V, radial
C_{16} = 470 µF, 25 V, radial
C_{17}, C_{18} = 10 µF, 63 V, radial
C_{19}, C_{20} = 220 µF, 25 V, radial
C_{21}–C_{24} = 47 nF, ceramic

Semiconductors:
D_1, D_2 = LED, low current
B_1 = B80C1500
T_1 = BC640

Integrated circuits:
IC1 = NE5532
IC2 = TL084
IC3 = LM319
IC4 = 7815
IC5 = 7915

Miscellaneous:
K_1 = 2-way terminal block, pitch 7.5 mm
S_1 = rotary switch, 3-pole, 4-position, for board mounting
S_2 = single-pole change-over switch
Tr_1 = mains transformer, 2×15 V, 1.5 VA
PCB Order no. 960049

Parts list:

AMPLIFIER

Resistors:
R_1, R_5, R_{13} = 1 kΩ
R_2, R_6 = 31.6 kΩ, 1%
R_3, R_4 = 8.2 MΩ (see text)
R_7 = 100 kΩ
R_8, R_9 = 3.3 kΩ, 0.5 W
R_{10}, R_{11} = 22 kΩ
R_{12} = 3.3 kΩ
R_{14} = 68 Ω
R_{15}, R_{19} = 220 Ω
R_{16}, R_{17}, R_{20}, R_{21} = 100 Ω
R_{18}, R_{22} = 22 Ω
R_{23} = 220 Ω, 5 W
R_{24} = 680 Ω, 5 W
R_{25}–R_{28} = 0.22 Ω, 5 W
R_{29} = 390 Ω
R_{30}, R_{31} = 47 kΩ
R_{32} = 5.6 kΩ

P_1 = 2.5 kΩ preset potentiometer

Capacitors:
C_1 = 270 nF
C_2 = 3.3 nF
C_3 = 22 pF, 160 V, polyester
C_4 = 1 nF
C_5, C_6 = 220 µF, 63 V, radial
C_7 = 2.2 µF, metallized polyester, pitch 5 mm
C_8 = 47 µF, 50 V, radial
C_9 = 100 µF, 40 V, radial
C_{10}, C_{11} = 100 nF

Semiconductors:
D_1, D_2 = 18 V, 1.3 W zener
D_3, D_4 = 10 V, 1.3 W zener
D_5–D_7 = 1N4004
T_1 = BD139
T_2 = MJE15030 (Motorola)
T_3 = MJE15031 (Motorola)
T_4, T_5 = GT20D201 (Toshiba)
T_6, T_7 = GT20D101 (Toshiba)
T_8 = BC640

Integrated circuits:
IC_1 = AD847JN (Analog Devices)

Miscellaneous:
JP_1 = 2-way contact row and jumper, or wire bridge
Re_1 = relay, 16 A, 24 V, 875 Ω (e.g., Siemens V23056-A0105-A101)
LS_1 = drive unit, see text
5 off flatcable (car-type) connectors with screw fitting
1 off heat sink for T_1–T_7, <0.55 K W^{-1}, e.g. SK47/100 SA (Fischer, available from Dau Ltd 01243 553 031)
Insulating kits (washers; bushes) for T_1–T_7
PCB as for filter (see text)

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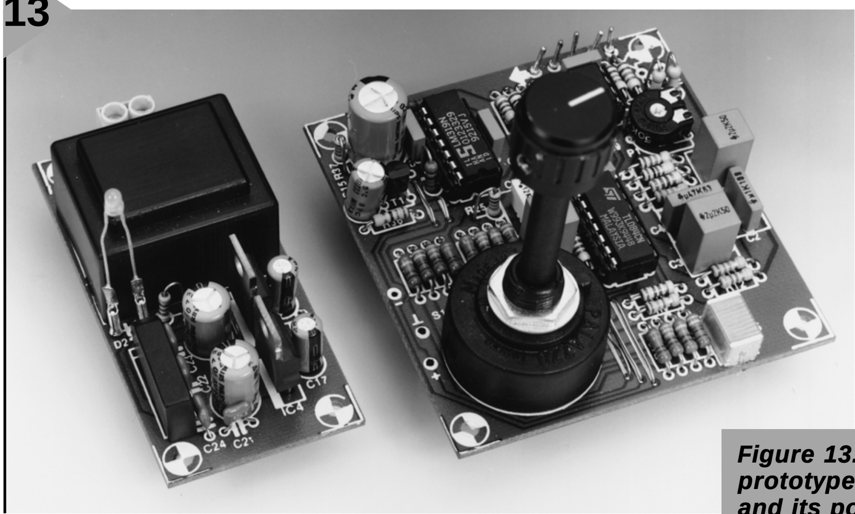

Figure 13. Completed prototype of the filter and its power supply.

a narrow strip of thin copper plate. It hardly needs emphasizing that thorough soldering of these parallel links is a prime requirement.

Furthermore, in order to ensure that the copper tracks for the supply lines and loudspeaker connections are not longer than strictly necessary, they are not taken to the edge of the board as is usual. Instead, the relevant solder pads are at the centre of the board and have been designed to allow flatcable (car-type) sockets to be screwed to them, preferably at the track side. Solder tags may be used, but these are nowhere near as robust as car connectors, which must, therefore, be preferred.

Three final remarks. (1) When populating the amplifier board, do not (yet) fit resistors R_3 and R_4; why will be discussed shortly. (2) Wire bridge JP_1 provides the necessary link between the negative power supply line and earth; it should, however, not be fitted if such a link is already present in the power supply itself. (3) Turn preset potentiometer P_1 fully anticlockwise.

Figures 13 and **14** show the completed prototype boards. Moreover, **Figure 15** shows the underside of the power amplifier module, which gives a good view of T_1, T_2 and T_3, the flatcable connectors for the supply lines and the loudspeaker connections, and the voltage rail that links the collectors of T_4–T_7. Note that this rail in the prototype consists of a folded strip of copper plate.

AMPLIFIER POWER SUPPLY AND CALIBRATION

The power amplifier needs a symmetrical supply of ±49 V. The design of the supply may be simple, but it should ensure the provision of sufficient current. The design shown in **Figure 16** is advised, since a transformer of 2×35 V at 300 VA, a 35-A bridge rectifier and four 10 000 µF, 63 V

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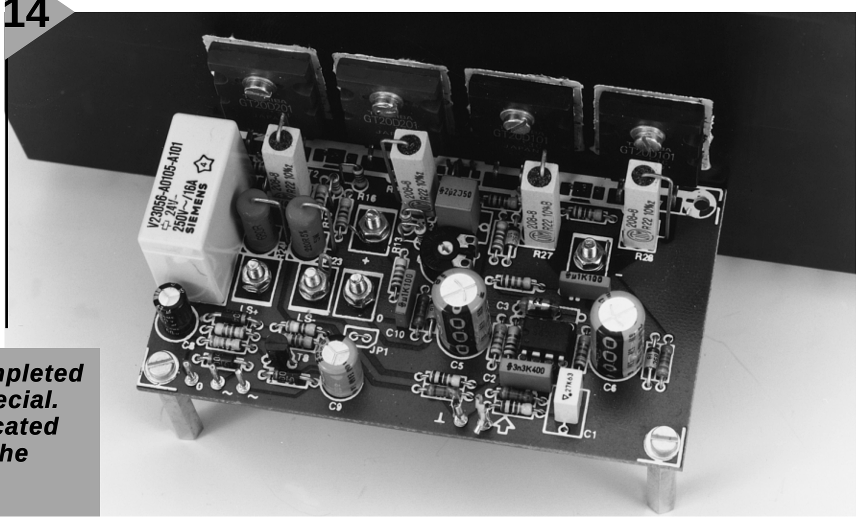

Figure 14. The top view of the completed amplifier board shows nothing special. The four output transistors are located together and well insulated from the heat sink.

electrolytic capacitors are fully up to the required task. The series resistors, in conjunction with the electrolytic capacitors, effectively decouple the supply lines to the amplifier.

As shown in Figure 16, the power-on delay circuit for the loudspeaker is supplied directly from the secondary windings of the mains transformer. If desired, a mains switch-on delay circuit, similar to that described on page 19 of our September 1995 issue, may be included at the primary side of the mains transformer.

Thoroughly check the amplifier board before commencing with the calibration. Provisionally connect the power supply and check that the potential across zener diodes D_1 and D_2 is about 18 V. If this is so, the setting of IC_1 is correct and it may be assumed that the remainder of the amplifier is also all right.

Reverting to R_3 and R_4: these resistors are intended to compensate the bias current of IC_1 to ensure that there is no direct voltage at the amplifier output. Check this by measuring the voltage, U_C, at pin 3 of IC_1 with a high-impedance voltmeter (multimeter) set to a millivolt range. The value of the resistors is calculated by

$R_3 = R_4 = (18/U_C) \times 31.6 \times 10^3$. [Ω]

In the prototype, the bias current through R_2 is 2.2 µA, which results in a voltage at the non-inverting input of IC_1 of 70 mV. Substituting this in the foregoing formula gives a value of the compensating resistors of 8.2 MΩ as specified in Figure 11. Note that the current levels may differ appreciably from one AD847 to another.

Set the quiescent current, which should be 100 mA through each of the output transistors. This is done by connecting a voltmeter or multimeter set to the 100 mV range across one of resistors R_{25}–R_{28} and gently turning P_1 clockwise until the meter reads 22 mV. Leave the amplifier on for about an hour and measure the voltage again; adjust P_1 as required. Finally, check that the voltage drop across the other three resistors is the same.

FINALLY …

The construction of the wooden loudspeaker enclosure has already been described in Part 1. It is now time to select a metal case for the filter, amplifier and amplifier power supply. Note that it is important that the heat sink of the power amplifier remains on the outside so that it is in cooling ambient air.

It is best to mount the filter on spacers directly at the back of the front panel of the case so that rotary switch S_1 is readily accessible. Phase selector S_2, overdrive indicator D_1 and on/off indicator D_2 must also be mounted on the front panel. Drill a small hole in the relevant position to make the subwoofer's sound level control P_1 accessible for adjustment. It is, of course, possible to use a standard potentiometer for P_1, mount this on the front panel, and connect it to the board with two short lengths of stranded circuit wire.

Construction of the amplifier power supply should not present undue difficulties, but it should be robust. The transformer may be

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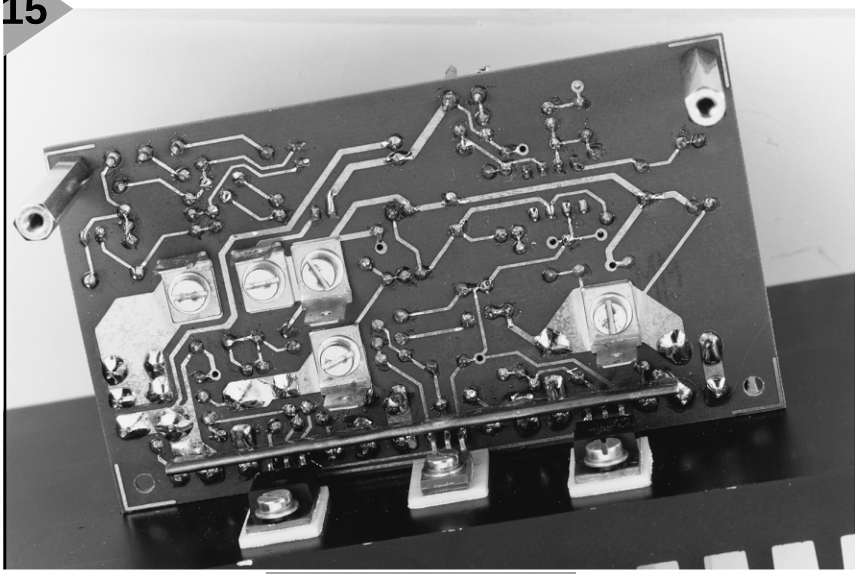
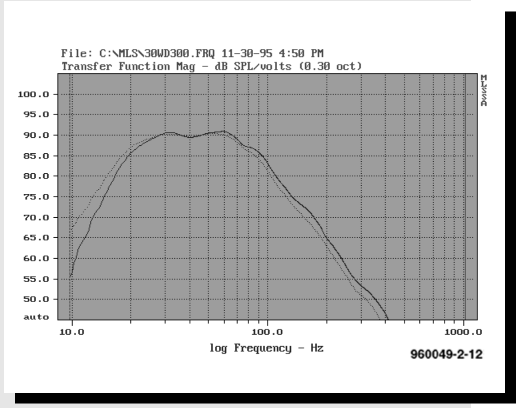


Figure 15. The underside of the completed amplifier board shows T1, T2 and T3, as well as the flatcable (car-type) connectors. The collectors of T4–T7 are interconnected with the aid of solder pins and a narrow strip of thin copper plate.

Different drive unit?

Although in the previous instalments reference has been made to a number of drive units, the preferred one remains the Monacor, since it gives excellent performance and offers very good cost-to-quality ratio. However, as mentioned in Part 2, even this otherwise excellent unit suffers from a deficiency. Although this deficiency is not serious, some readers may, none the less, be interested to know that we have found yet another drive unit that is suitable for the subwoofer. Finding alternative drive units is not easy since their parameters tend to differ to such an extent that changes to the dimensions of the enclosure and to the bass reflex tuning are often required.
The alternative is the Type 30WD300 from Vifa. This is also a 300 mm woofer that fits readily into the enclosure described in Part 1 and which can use the same filter as the other drive units. The characteristics in the illustration show that there are some differences between it and the Monacor unit, but that these are of not much consequence. The solid curve refers to the Vifa unit, and the dashed one to the Monacor unit.



The 30WD300 produces no spurious sounds and has an excellent overall performance. Unfortunately, like the Radio Shack and Parts Express units, it does not have a double voice coil, which makes it unsuitable for use in the passive version of the subwoofer. Used in the active version, it has a small benefit in that, because of its higher impedance, it requires a lower current from the power amplifier. This in turn means that the mains transformer may be a (less expensive) 225 VA type (and the fuse rating can be reduced to 1 A). On the other hand, the Vifa unit is dearer than the Monacor unit, so the choice is yours.

screwed to the bottom of the case and the remainder on a small sheet of prototyping board. The connections between the secondary windings of the transformer, the bridge rectifier, the capacitors and resistors should be in heavy-duty, single-strand copper wire. The negative supply line shown bold in Figure 16 must be kept as short as possible. It is best to take the capacitor terminals and the centre tap of the transformer to a common point (star connection). The bridge rectifier must be mounted on a small heat sink or be screwed directly to the bottom of the case.

As far as the interconnections are concerned, those between the filter supply board and the filter board have already been discussed, as have the operating controls of the filter. Connect the high-level and line-level input pins to a couple of audio sockets via single screened audio cable. The output of the filter should be taken also via single screened audio cable to the input of the power amplifier (if the length of this cable is only a few centimetres, stranded circuit wire may be used instead of screened cable).

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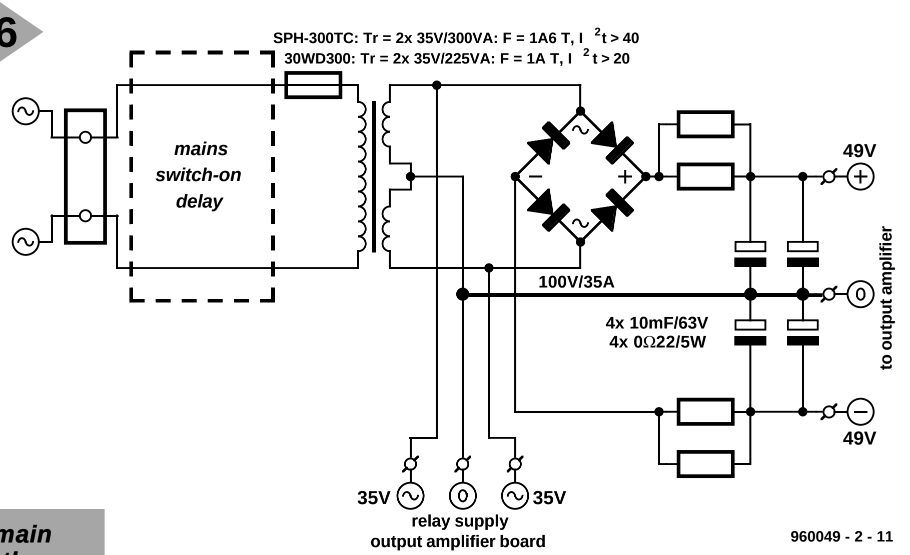


Figure 16. The main requirement of the amplifier power supply is the ability to deliver sufficient current.

Alteration of range of S_1

Cross-over frequency	Values of resistors (kΩ)			
	R_{13},R_{17},R_{21}	R_{14},R_{18},R_{22}	R_{15},R_{19},R_{23}	R_{16},R_{20},R_{24}
80 Hz	3.32			
90 Hz		2.94		
100 Hz			2.67	
110 Hz				2.43

The connections that need most attention are those between amplifier and amplifier power supply, and between amplifier and loudspeaker. The amplifier delivers an appreciable power output, so that the connections must be able to handle fairly large currents, while the transfer resistances must be kept to a minimum. It is, therefore, recommended to use cable with a cross-sectional diameter of ≥ 2.5 mm². Connect the amplifier power supply via three lengths of this cable, preferably terminated into car-type flat connectors to terminals '+', '0', and '–' on the amplifier board. Connect the loudspeaker (with the two voice coils connected in parallel) via two lengths of the same cable to terminals '+LS' and '–LS' on the amplifier board.

Connect the primary of both mains transformers via heavy-duty, well-insulated cable to the mains entry at the rear of the case. Use a good-quality mains entry with integral fuse holder and on/off switch.

Which cross-over frequency?

The setting of P_1 and S_1 can be found empirically only: there are no definite rules, since it is really a question of personal preference. Knowing the lower frequency limit of the existing loudspeaker system, either though personal measurement or from manufacturers' data, is a help, however, since the cross-over frequency of the subwoofer can then be chosen close to this limit.

The cross-over frequency chosen in the prototype is different from that found in many popular satellite or subwoofer systems. This is because such systems normally use small drive units that can just about handle 100–150 Hz, which means that the central bass unit must take over at that frequency. The present subwoofer is meant really as an addition to normal full-range loudspeaker systems, that is, compact to medium large loudspeaker enclosures that give a reasonable bass performance, but do not perform so well at the lower bass range. Position '70 Hz' will prove fine in use with most bookcase-type loudspeakers, while with compact loudspeakers '60 Hz' will normally the best value. Cross-over frequencies 40 Hz and 50 Hz are intended for combination with medium to large loudspeaker enclosures. The two lower frequencies, 30 Hz and 20 Hz, are usually best for use with electrostatic loudspeakers to give the 'thin' bass of these units rather more 'punch' at the lower end.

If your stereo loudspeakers are rather small, we can imagine that you will find the cross-over frequency chosen for the prototype rather low and would prefer to use the standard frequency for satellite/subwoofer systems. This is possible by altering the values of resistors R_{13}–R_{24} as shown in **Table 1** for frequencies 80 Hz, 90 Hz, 100 Hz and 100 Hz. These alterations do not affect the amplifier or the loudspeaker. [960049]

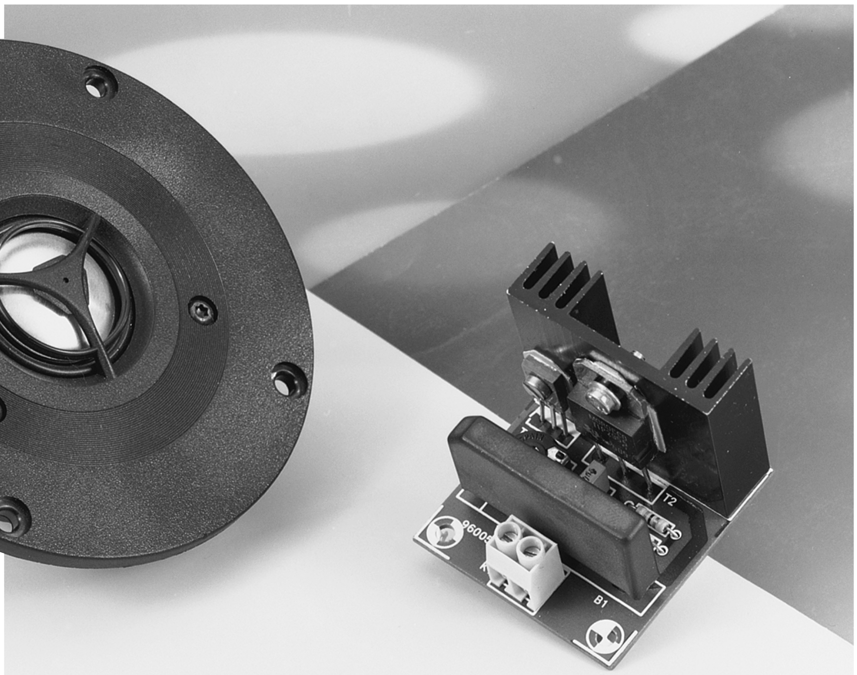
Contents

a.f. power limiter

protects tweeter

Owing to their relatively low rating, tweeters (high-frequency loudspeakers) form the weak link in an audio system. If the volume is suddenly turned up too high, the risk of the tweeters being damaged irreparably is high. Such an impetuous and costly mistake, can, however, be avoided in two different ways. The first is to curb your desire to turn up the volume to levels that the loudspeakers cannot handle. The second is to build the power limiter presented in this article…it's much safer than controlling yourself when you're adjusting the volume of your beloved audio system.

Design by T. Giesberts



There will be many readers who, after reading the introduction to this article, will say that this does not concern them. They have a 100 W amplifier and the loudspeakers are also rated at 100 W. So, nothing can go wrong. Really?

Unfortunately, things can go wrong, since the rating given by the loudspeaker manufacturers is true only for *average* music signals. In arriving at this rating, account is taken of the fact that the energy contained in music signals is strongly dependent on frequency. Of the power delivered by the output amplifiers roughly 75 per cent is applied to the woofers (low-frequency loudspeakers), 25 per cent to the mid-frequency loudspeakers, and only 5 per cent to the tweeters. This means that of the power output of 100 W only about 5 W is applied to the tweeters.

Equally unfortunately, not all music signals are average. For in-stance, in the case of synthesizer music it can happen that a sudden burst of high-frequency music is produced, which at that instant contains more than half the total emitted energy. This means in this example that some 50–60 W of music power is applied to the tweeters instead of the *average* 5 W. Many tweeters just cannot cope with this sort of power.

There is yet another aspect concerning the specified rating of tweeters. Although in the case of woofers and mid-frequency speakers the 'true' rating is given by the manufacturers, this is not so in the case of tweeters. For these units, the specified rating applies only if they are used with a cross-over filter! On close examination, it appears that a rating of, say, 50 W applies only if the speaker is used with a 2nd-order high-pass filter with a cut-off frequency of 4000 Hz. If, however, the cut-off frequency is, say, 2000 Hz, the rating is lowered to 20 W. Without a filter, the rating appears to be only 5 W!

All this is, of course, reasonable, since, at lower frequencies, a diaphragm has to move over a larger distance and tweeters just are not designed for this. Nevertheless, it goes to show that loudspeaker constructors should be well aware of how ratings are specified.

FUSE OR ZENERDIODE?

The question that arises in view of the foregoing is how the tweeters can be protected effectively.

The simplest way is merely to connect a fuse in series with the tweeters. However, this gives only a limited degree of protection, and also introduces a few drawbacks. If a fast fuse is used, chances are that it will blow at the first peak in the music signal. A slow fuse on the other hand does not guarantee that it will always be faster than the tweeters. In other words, the tweeters might still give up the ghost before the fuse blows. Add to this that any fuse introduces a certain resistance, which may vary from some tenths of an ohm to more than an ohm. This should undoubtedly be borne in mind, since, unless compensating measures are taken, it will inevitably lead to some attenuation of the high-frequency sound.

A variation of the standard fuse is a special device with positive temperature coefficient (PTC), which is available from many loudspeaker dealers. It is a semiconductor element that reacts just like a slow fuse when the current through it becomes too high. Unlike a fuse, however, it recovers when the danger is past: it need not be replaced, therefore. Unfortunately, its resistance is slightly higher than that of a fuse.

It is clear that series current limiting by a fuse or PTC device has its drawbacks. What other means are there?

One is a voltage limiter *across* the tweeter. In its simplest form, this could consist of two anti-series connected zener (power) diodes, assuming that the necessary series resistor is already present in the cross-over filter (damping resistor). A possible arrangement is shown in **Figure 1**, in which the zener diodes are at the right. Resistor R_1 is the series (damping) resistor mentioned earlier. If the zener ratings are 5.6 V, the power applied to the tweeter is restricted to about 5 W.

It may be asked whether such a simple protection is sufficiently effective, to which the answer is yes and no. The difficulty is that this sort of protection is too effective. This is because the zener action normally commences at fairly small currents when the zener voltage is nowhere near its nominal value. This results in untimely limiting, which causes a compression effect even at fairly small signals. Another, practical, problem is that power zener diodes are not easy to come by.

SIMULATED ZENER DIODE

What we need is a protection that is faster and more reliable than a series element and does not have the disadvantages of a pair of zener diodes in parallel. This requirement could be met by a sort of simulated power zener diode that has a sharply defined starting point.

The simple circuit in **Figure 2** is such a zener diode, consisting of two discrete darlington transistors. Connector K_1 is simply connected in parallel with the tweeter terminals. There is no need of a supply voltage, because this is drawn from the loudspeaker signal.

The alternating signal across the loudspeaker is rectified by B_1, so that a pulsating direct voltage exists across network R_1-R_2-P_1, which is averaged (to a degree) by capacitor C_1. When the alternating signal increases, transistor T_1 begins to conduct at a given value determined by the setting of P_1. Transistor T_1 turns on the power transistor, T_2, which consequently short-circuits part of the alternating signal. A part only, of course, because if the signal were short-circuited completely, T_1 would be cut off, leaving T_2 without drive. All this means that there is a limiting effect which stabilizes itself at a certain signal level, just as a zener diode does. The difference is that the simulated zener diode has a defined starting voltage, so that signals below that level are not affected. Thus, compression effects do not occur. In Figure 2, the values of R_1, R_2, and the preset have been chosen to ensure that the zener voltage can be set with P_1 between 5 V and 9 V, roughly corresponding to powers between 3 W and 10 W into 8 Ω.

In the design stages, it was considered to expand the circuit with an indicator LED, but in practice it was found that the music peaks are too short to make an LED light visibly. It is, of course, possible to lengthen the signal peaks electronically, but bear in mind that the required energy must be drawn from the loudspeaker signal and this may lead to an increase in distortion.

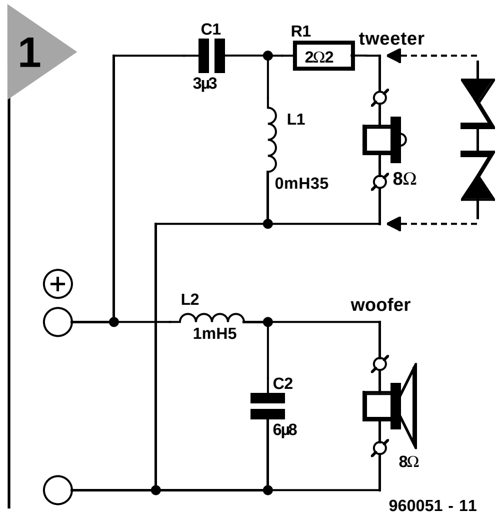


Figure 1. If the cross-over filter contains a damping resistor, R1, for the tweeter, the voltage may be limited by two anti-series-connected zener diodes.

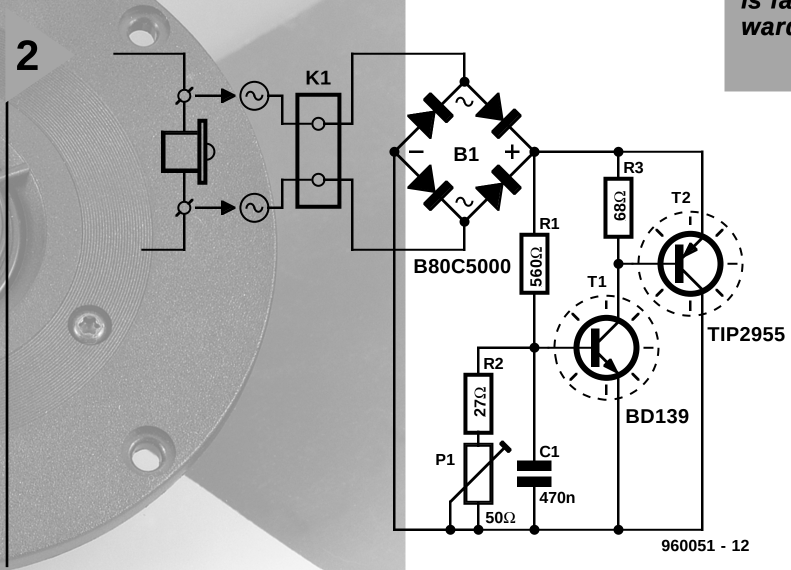


Figure 2. The design of a power diode with variable zener voltage is fairly straightforward.

CONSTRUCTION

The limiter is best built on the printed-circuit board shown in **Figure 3**, but it is, of course, just as easily built on a small prototyping board. The only aspect that needs attention is that the two transistors are to be fitted on to a common heat sink of about 6.5 K W^{-1}. This is necessary, because when the twee-ter is overloaded, there is quite a heat dissipation. The transistors must be electrically isolated from the heat sink with the aid of insulating washers

Setting up

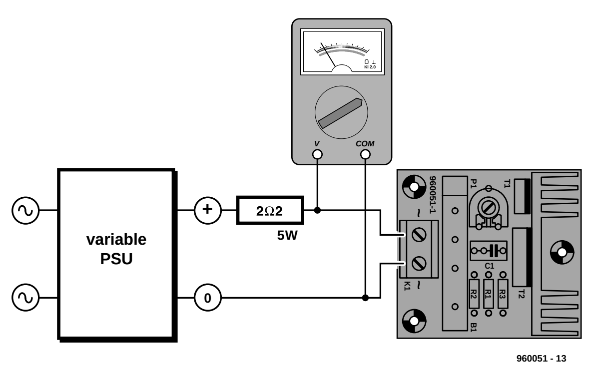
Much thought was given to the control range of P_1. The data books of a number of loudspeaker manufacturers showed that the majority of tweeters are normally rated at 3–5 W, with some as high as 8 W. This led to the decision to make the range 3–10 W into 8 Ω, corresponding to a signal voltage range of 5–9 V. If you have no suitable measuring equipment available, take it as a rule of thumb that the power restriction is about 3 W with P_1 fully anticlockwise, about 5 W with the preset turned one third of its travel clockwise, and about 10 W with it fully clockwise.

Setting the zener voltage of the limiter requires a variable power supply and a multimeter. Connect the power supply across K_1 via a 2.2 Ω, 5 W resistor, and the multimeter in parallel with this as shown in the diagram.

If we assume a power limit, P, of 5 W into a (loudspeaker) impedance, R, of 8 Ω, the signal voltage, u, is:

$$u = \sqrt{PR} = 6.3\ \text{V}$$

The power supply provides a direct voltage, which, as far as level is concerned, is equal to √2 = 1.414 times the r.m.s. value of an alternating voltage. Thus, for the above values of 5 W into 8 Ω, the preset must be set to give a meter reading of 1.414×6.3 = 8.9 V. Make sure, of course, that the power supply output level is sufficiently high.



and non-metallic screws and nuts. The photograph shows what the finished limiter looks like. A good place for it is on or close to the cross-over filter board. A good alternative is beside the tweeter on the inside of the front panel of the enclosure. Connect K_1 to the tweeter terminals with medium-duty, flexible, insulated circuit wire.

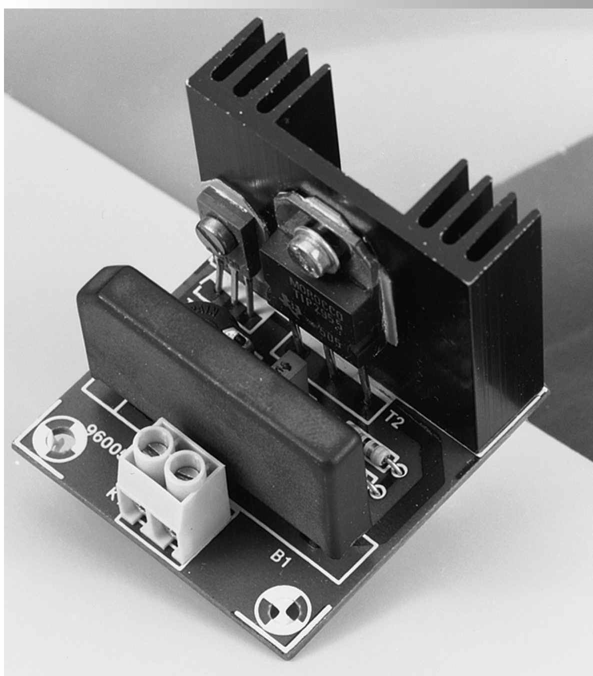
USAGE

As stated earlier, in the design it is assumed that the cross-over filter contains a resistor in series with the tweeter. This resistor is essential, because the surplus voltage when the limiter is active is dropped across it.

If you are worried by the thought that when the limiter is active the amplifier is virtually short-circuited as far as high frequencies are concerned and that it may not be able to cope with this, connect a 500 mA fuse in series with the tweeter. This protects the amplifier against a full short-circuit: its resistance of 0.3 Ω is, in this case, negligible.

The setting of P_1 is described in the box.

[960051]



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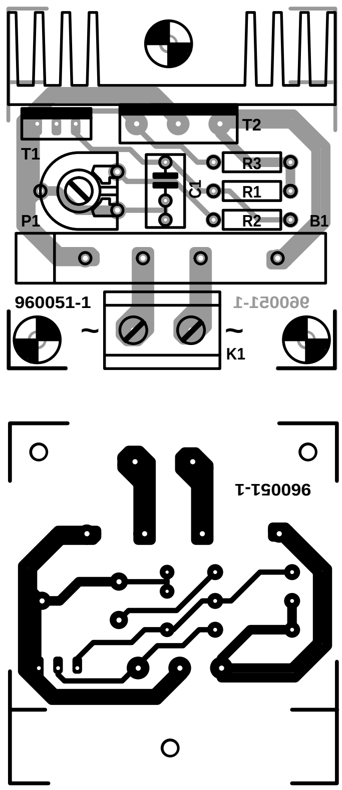


PARTS LIST

Resistors:
R_1 = 560 Ω
R_2 = 27 Ω
R_3 = 68 Ω
P_1 = 50 Ω preset

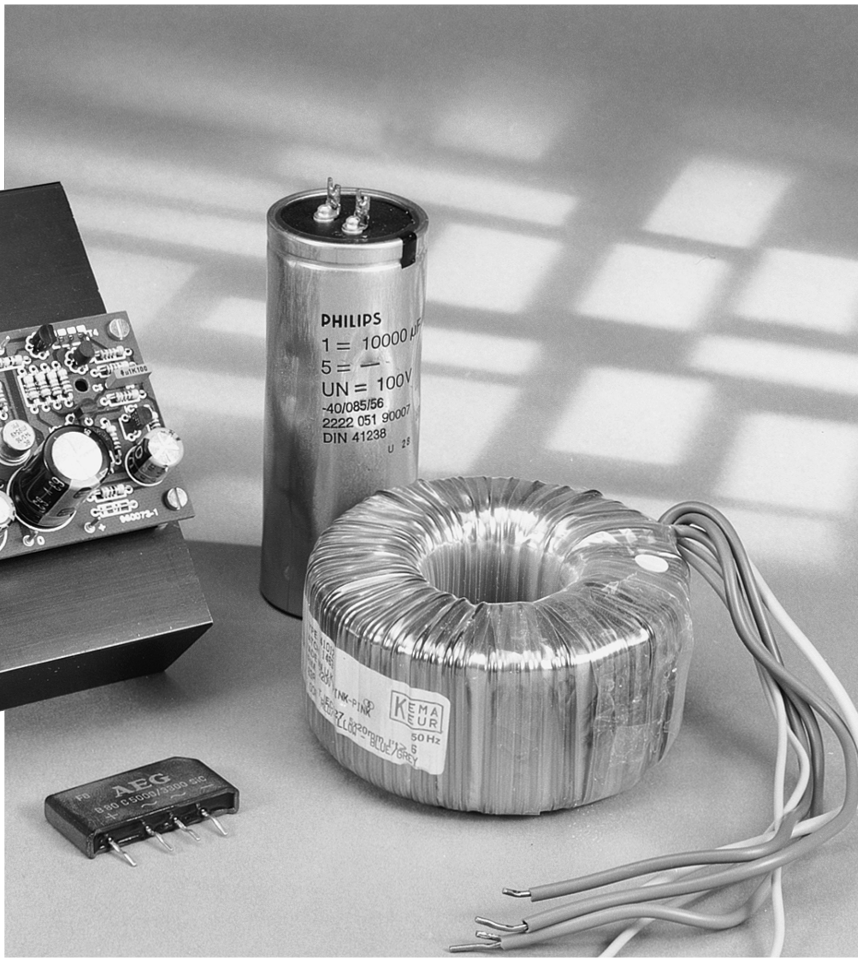
Capacitor:
C_1 = 470 nF

Semiconductors:
T_1 = BD139
T_2 = TIP2955

Miscellaneous:
K_1 = 2-way terminal block for board mounting, pitch 7.5 mm
B_1 = B80C5000
Heat sink, 6.5 K W^{-1}, e.g., Fischer SK59 (37.5 mm) (available from Dau, telephone 01243 553 031)
Insulating washers, and non-metallic screws and nuts for T_1 and T_2

Figure 3. The printed-circuit board for the limiter shows how simple the construction is.

active power buffer



with temperature monitor

Frequently, a direct voltage is required in a circuit that must be free of hum and noise, but not regulated. For instance, the supply to an audio output stage must be able to vary with the mains voltage and the load. Another example is when in the workshop a variable-ratio transformer (variac) is to produce a well-filtered direct voltage with good loading capacity for general purposes.

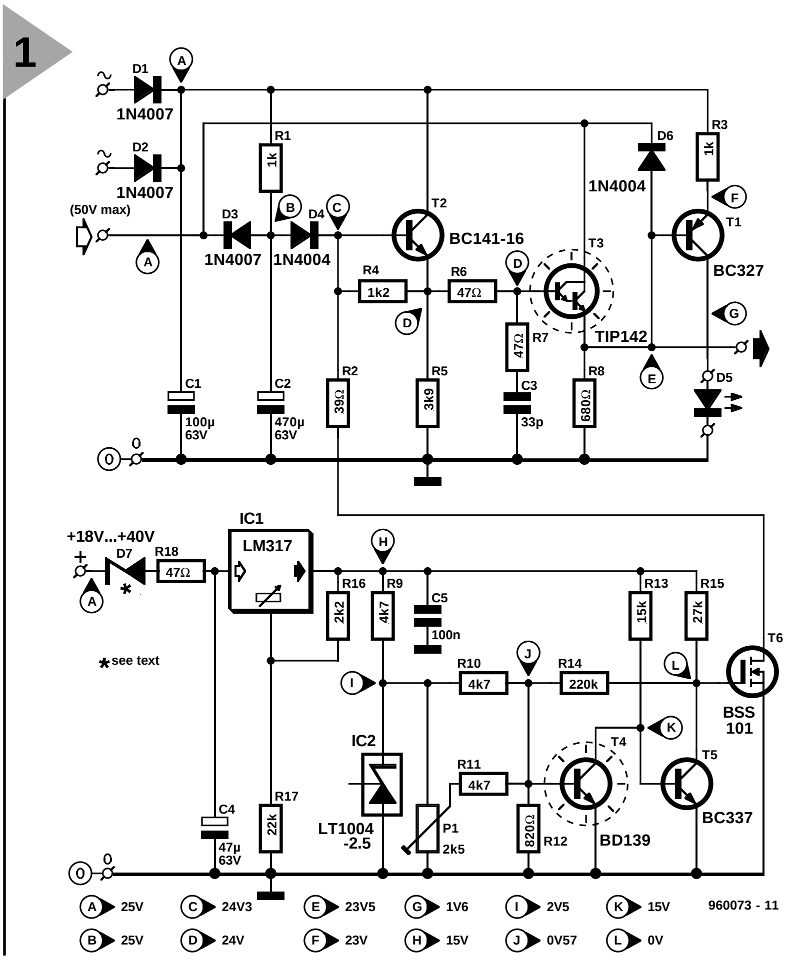
Design by W. Steimle

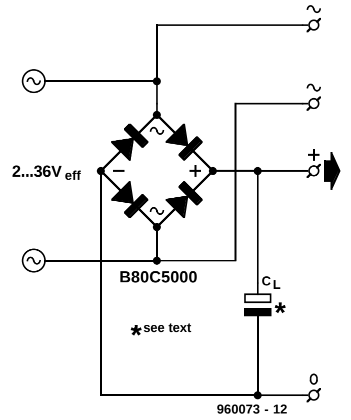
The hum emanating from an unregulated power supply is normally caused by too high a current drain or too small a reservoir capacitor. Enlarging the capacitor is often the simplest, but not always the most effective, way of dealing with the problem. Hum is the manifestation of a ripple on the output voltage and it is best to suppress this in an active manner.

The circuit diagram in **Figure 1** looks like that of a conventional series regulator, but has no regulating amplifier with control comparator. Therefore, the output voltage automatically adapts itself in accordance with the input alternating voltage and the current through the load. Consequently, the entire hum voltage is applied to the collector-emitter junction of darlington transistor T_3. The advantage of such an arrangement lies in a drastic reduction of the maximum dissipated power (at the highest mains voltage).

The lower part of the figure is a temperature monitor in which T_4 is the sensor. If this transistor detects an over-temperature, the monitor circuit pulls the base of driver T_2, and thus that of T_3, to ground. This effectively cuts off the output current, so that no more power is dissipated.

1





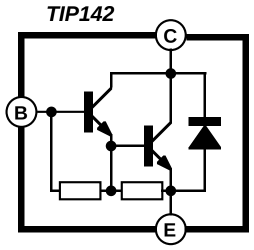


Figure 1. Circuit diagram of the active power buffer, including temperature monitor.

CIRCUIT DESCRIPTION

The circuit is arranged so that the power section (upper part of the figure) and the temperature monitor have their own rectifier. That for the monitor is the bridge rectifier, shunted by reservoir capacitor C_L, while D_1 and D_2 serve the power section.

Whereas the potential across C_L has a ripple whose level depends on the load current, the voltages across C_1 and C_2 are virtually free of ripple, since the load is small. The potential across C_1 is about equal to the peak value of the alternating voltage applied to the bridge rectifier. The voltage across C_2 depends on the minimum level across C_L, $U_{o(min)}$, because it is pulled down by D_3 twice in each mains period to $U_{o(min)}$ + U_{D3}. The potential across C_2, U_{C2}, determines the output voltage, which is about U_{C2}–$4U_{D3}$.

Diode D_4 is necessary to ensure that U_{CE} of emitter follower T_2 retains a nominal value when the potential across C_1 is a minimum. Resistors R_4 and R_5 ensure that the direct current through T_4 is not too dependent on the current amplification of the transistor and the load current.

Resistor R_1 provides the charging current for C_2, which must, of course, always be greater than the current through D_4 and the base current of T_2. Since the drop across R_1 increases the dissipation of T_3, the resistor should be kept as small as possible. It should be not too small, however, because at large load currents it determines the ripple on U_{C2}. Its specified value is a compromise between these conflicting requirements.

Since the base of T_3 is fed from the inductive output of T_2, resistor R_6 is necessary to obviate any tendency of the darlington to oscillate. Note that R_7 and C_3 already tend to make the reactance more resistive.

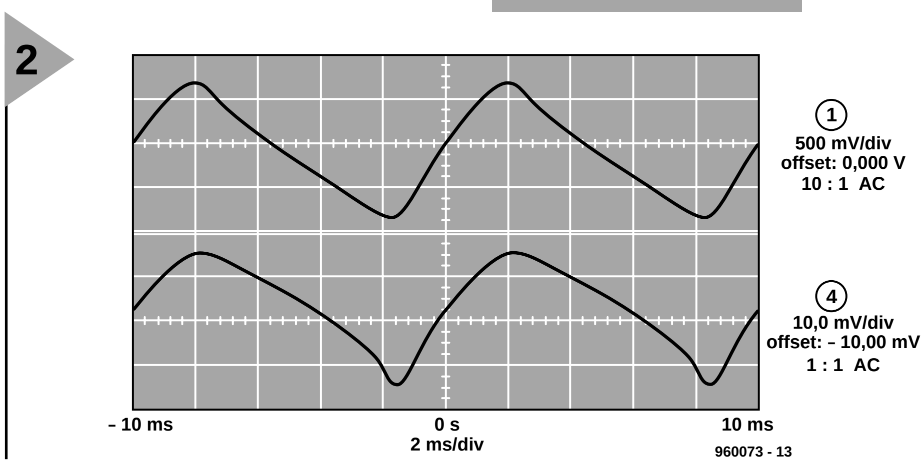
Resistor R_8 provides a minimum load for the circuit.

Resistor R_5 prevents driver T_2 from switching off during the quiescent state when T_3 draws only a tiny base current.

Transistor T_1 draws a current through D_5, which is more or less directly proportional to the peak hum voltage and inversely proportional to the value of R_3. This means that the

Figure 2. A 100 Hz hum voltage and the direct voltage at the output that depends on the load current; input alternating voltage is 20.7 V.

2



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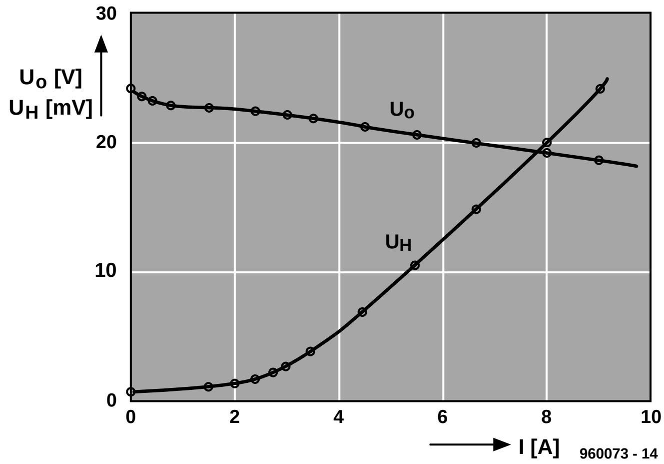

Figure 3. Hum voltage vs time at the input and at the output of the circuit.

brightness of the LED increases with rising load current.

Diode D_6 comes into action only when, for instance, the circuit is being used to charge a battery and the mains fails. The diode then prevents too high a reverse-bias voltage at the base-emitter junctions of T_2 and T_3.

The output resistance, R_o, of the circuit depends in the first instance on the value of C_L and the peak-to-peak value of the hum voltage, $U_{H(pp)}$, across this capacitor:

$$R_o \approx U_{H(pp)}/I_o \approx 1/2fC_L,$$

where f is the mains frequency. Thus, if C_L = 10 mF, R_o = 1 Ω.

The residual hum voltage at the output depends on load current I_o. This is shown diagrammatically in **Figure 2**, which assumes an output voltage of 24 V. When the load current is 10 A, hum suppression is about 30 dB. Note that the frequency of U_H is 100 Hz. Figure 2 also shows the relationship between U_o and I_o. From this it will be seen that R_o = 0.3 Ω.

These measurements were carried out with the input alternating voltage held constant at 20.7 V. In practice, this voltage drops somewhat owing to the internal resistance of the transformer.

Figure 3 shows the hum voltage at the input and output. Note that the hum suppression at peak-to-peak values is about 50, that is, 34 dB, or a little better than previously.

If the average direct voltage at T_3 is made equal to half the hum voltage, the dissipated power, P, is

$$P = U_a \cdot I_o = I_o^2/4fC_L.$$

Thus, when I_o = 10 A and C_L = 10 mF, the power dissipation is 50 W. Note that this value is independent of the output voltage.

Temperature monitor

For input voltages greater than about 6 V, the circuit is not proof against sustained short-circuits.

It has, however, a temperature monitor which arranges for the load to be disconnected when the temperature rises above a preset level. This arrangement is particularly sensible for occasions when, for instance, the present circuit is used in conjunction with a variac for various purposes when an unwanted overload can happen all too easily.

The sensor is a Type BD139 transistor, which should be mounted on a suitable heat sink. The sensor action is effected by the temperature dependence of the base-emitter voltage, which is easily computed. The sensor and T_5 form a Schmitt trigger with temperature-dependent threshold and hysteresis.

The base-emitter voltage of T_4 is held constant by IC_2 at a value set with P_1, at which the monitor comes into action when the temperature of the heat sink reaches 85–90 °C. When that happens, T_6 pulls the base of T_2 to ground via R_2, whereupon T_3 is cut off and the output is open-circuited.

Power for the monitor circuit is derived from a mains adaptor that must provide at least 18 V. It may also be taken from the + terminal of C_L if the potential across this capacitor lies between 18 V and 40 V. If it is higher than 40 V, it is pulled to this value by D_7. The rating of this zener diode must be such that the voltage at the input of IC_1 remains below 40 V when U_{CL} is a maximum and above 18 V when U_{CL} is a minimum.

Construction

The circuit complete with temperature monitor is best built on the printed-circuit board shown in **Figure 4** (which unfortunately is not available ready made). The two sections of the circuit are linked on the board by the common earth track at the centre and the connection from T_6 to the base of T_2 via R_2. If this resistor is omitted, the two sections are completely isolated from one another (apart from the common earth track).

If D_7 is not used, it should be replaced by a wire bridge.

Note that T_3 and T_4 are fitted at the track side of the board. When the board is fitted on to the heat sink as in **Figure 5**, these transistors must be fitted on to the heat sink with the aid of insulating washers, screws and nuts. Two holes are provided in the board near T_2 and T_5 to allow access to the relevant screws.

Connect the ~ terminals from the bridge rectifier to the relevant terminals on the board via light-duty insulated circuit wire, and the + and 0 terminals to the relevant terminals on the board by medium-duty insulated circuit wire.

The buffered output is available at the emitter terminal of T_3.

Initial test

The circuit may be tested by checking whether the voltages measured at various points indicated in Figure 1 coincide with the values shown in that figure. The values were measured in the prototype with an input voltage of 25 V and an open-circuit output.

Setting up

The temperature at which the monitor switches off the output is set with P_1. This temperature is determined by the heat resistance of the heat sink on which T_3 is fitted. If this is 1.8 K W^{-1}, for instance, and the output power is 40 W, the temperature of the heat sink is 1.8 × 40 = 72 K with respect to the ambient temperature. The temperature of the power transistor itself is another 40 K above this, so that, when the ambient temperature is 25 °C, it reaches 137 °C. The maximum permissible temperature of the TIP142 is 150 °C.

Set P_1 fully anticlockwise and load the circuit with a power resistor of 0.5 Ω, rated at ≥ 32 W and set the input voltage derived from a variac to such a value that the output current is 8 A (maximum current through the TIP142 is 10 A). If only a fixed alternating voltage is available as input, the load resistance should be chosen to cause a current of about 8 A to flow through it.

The voltage at the power transistor depends on the value of C_L, and it is well known that the tolerances of electrolytic capacitors vary widely. Therefore, it is better to measure the collector-emitter voltage and multiply this by the current to arrive at the dissipated power, which should be about 40 W. At this power, the LED should light.

Parts list

Resistors:
R_1, R_3 = 1 kΩ, 5 W
R_2 = 39 Ω
R_4 = 1.2 kΩ
R_5 = 3.9 kΩ
R_6, R_7, R_{18} = 47 Ω
R_8 = 680 Ω, 5 W
R_9–R_{11} = 4.7 kΩ
R_{12} = 820 Ω
R_{13} = 15 kΩ
R_{14} = 220 kΩ
R_{15} = 27 kΩ
R_{16} = 2.2 kΩ
R_{17} = 22 kΩ
P_1 = 2.5 kΩ preset

Capacitors:
C_1 = 100 µF, 63 V, upright
C_2 = 470 µF, 63 V, upright
C_3 = 33 pF
C_4 = 47 µF, 63 V, upright
C_5 = 100 nF

Semiconductors:
D_1–D_3 = 1N4007
D_4, D_6 = 1N4004
D_5 = LED, high-efficiency, red
D_7 = see text
T_1 = BC327
T_2 = BC141-16
T_3 = TIP142
T_4 = BD139
T_5 = BC337
T_6 = BSS101 (Siemens)

Integrated circuits:
IC_1 = LM317LZ (National Semiconductor)
IC_2 = LT1004-2.5 (Linear Technology) or LM336-2.5 (National Semiconductor)

Miscellaneous:
Heat sink 1.8 K W^{-1}
Insulating washers, screws and nuts for T_3 and T_4

Wait about 30 minutes, after which the heat sink temperature should be about 72 K higher than the ambient temperature (as computed earlier). It is, of course, better to measure the temperature, but this presupposes that a suitable thermometer is available.

When the stated temperature has been reached, turn the wiper of P_1 carefully clockwise until the LED lights brightest. The monitor then comes into action: the current is cut off and the temperature drops.

Wait a while for the temperature to reach a lower value, when the monitor should be deactivated, and an output current flows again.

Increase the power dissipation slightly and repeat the foregoing. In this way, average cut-off and switch-on temperatures will be set.

[960073]

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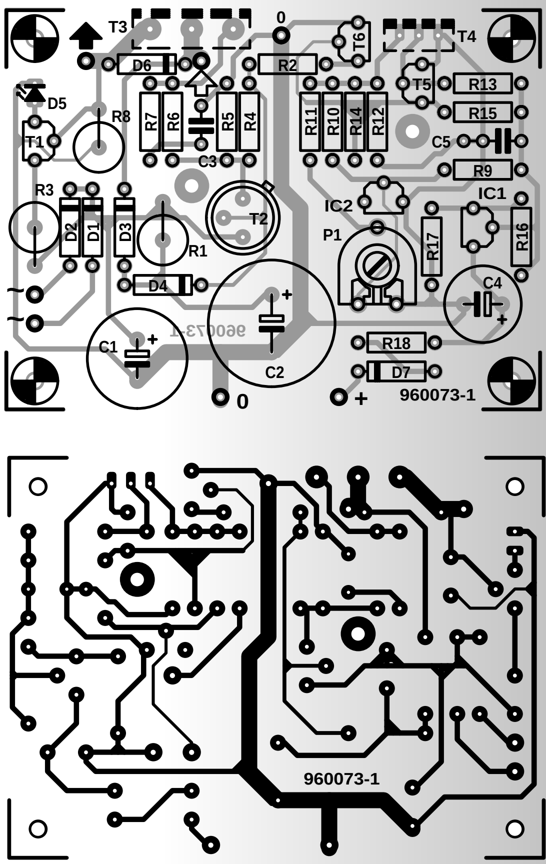

Figure 4. The printed-circuit board for the active power buffer, including temperature monitor.

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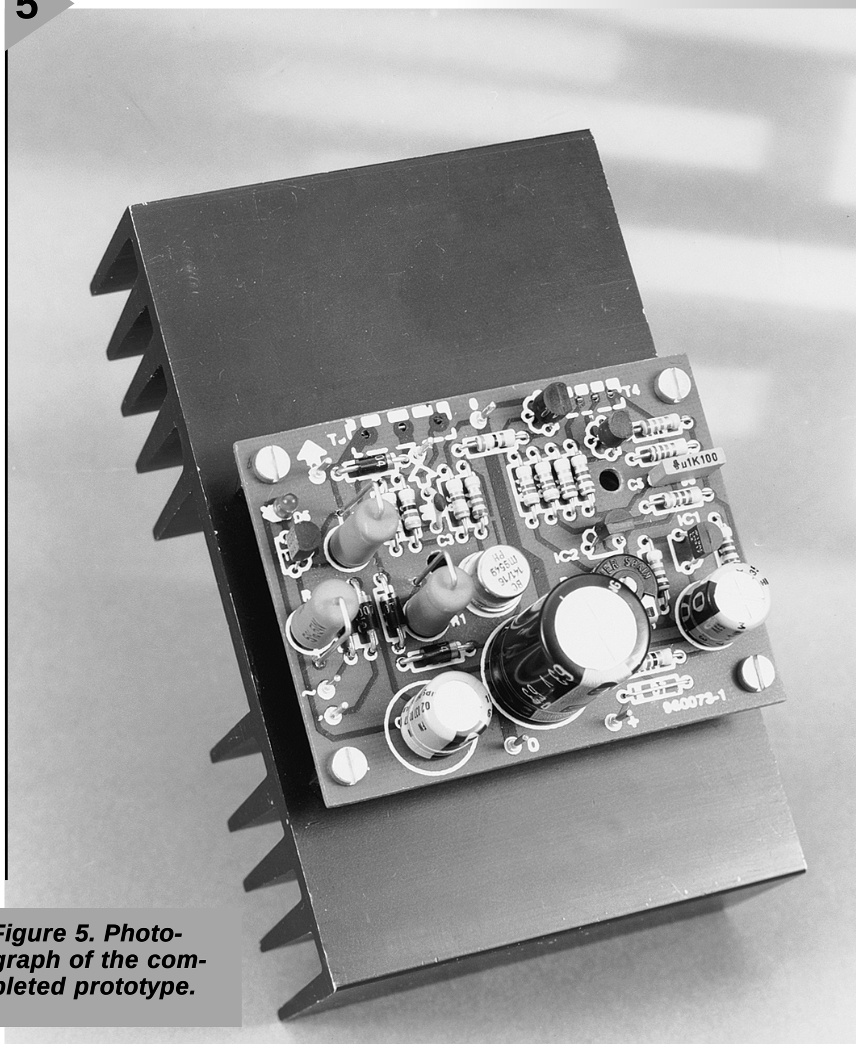
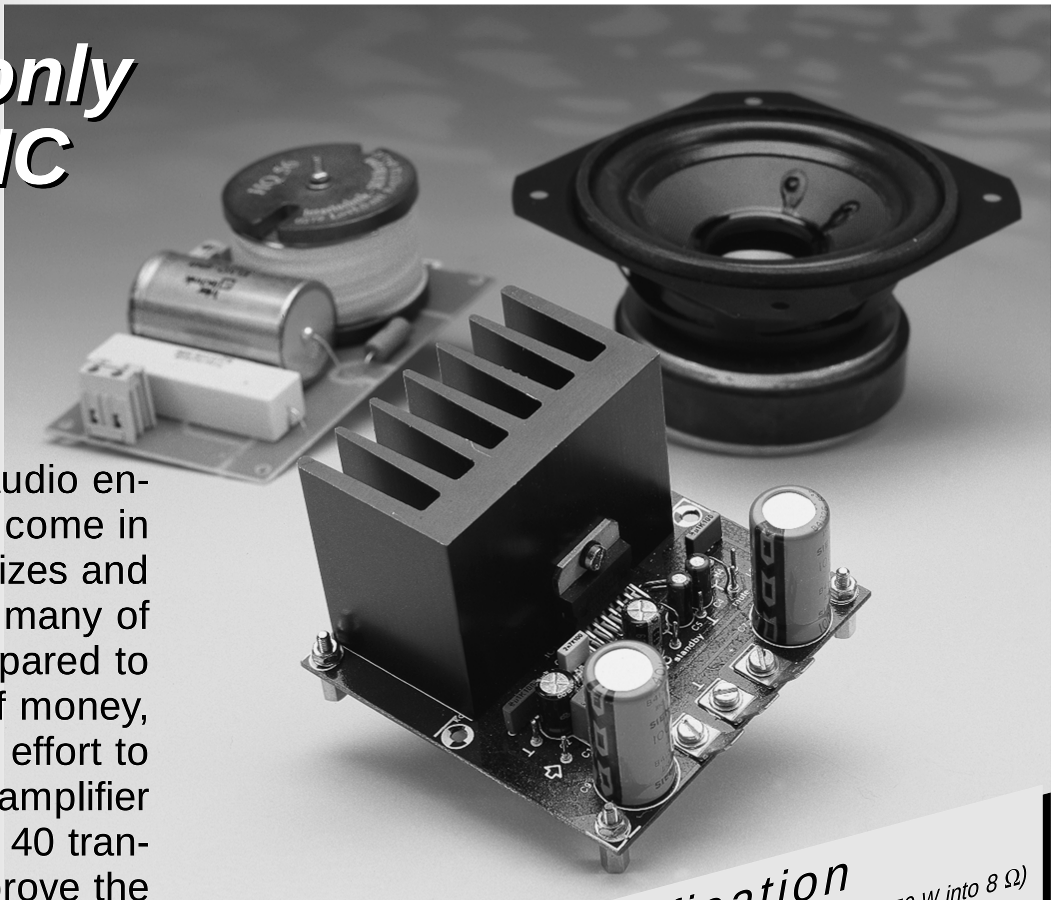


Figure 5. Photograph of the completed prototype.

50 W a.f. amplifier

uses only one IC



Although audio enthusiasts come in many sizes and colours, not many of them are prepared to spend a lot of money, time and effort to build an a.f. amplifier using up to 40 transistors to improve the distortion by a fraction of a per cent. Therefore, the amplifier described here should appeal to those enthusiasts. It is compact, presents no problems and yet has properties that make it fully suitable for all but the most demanding audio applications. In short, an amplifier that is geared to the practical audio buff.

Source: SGS Thomson

Specification

Input sensitivity:	1.3 V (50 W into 8 Ω)
Input impedance:	10 kΩ
Bandwidth:	16 Hz – 100 kHz
Slew rate:	10 V µs–1
Output power:	50 W into 8 Ω (0.1% THD)
	82 W into 4 Ω (0.1% THD)
Signal-to-noise ratio:	105 dBa (1 W/8 Ω)
	0.002% (1 kHz)
THD+N with 40 W into 8 Ω:	< 0.04% (20 Hz – 20 kHz)

The Type TDA7294 IC from SGS-Thomson is an integrated a.f. amplifier intended for use in all sorts of hi-fi application. Its circuit diagram is shown in **Figure 1**. Its most prominent feature is the much higher power output than is usual with this kind of integrated amplifier. According to the manufacturer's data sheets, the special DMOS output stage of the 15-pin chip can deliver outputs of up to 100 watt. Considering other properties, such as low noise, low distortion and reliable short-circuit and thermal protection circuits as well, the chip is indeed an interesting one.

Having said that, power output specifications are often rather optimistic. In this instance, the 100 W appears to refer to the IEC norm for music power with 10 per cent distortion, which, as far as hi-fi applications are concerned, is not the correct way of specifying output power. Moreover, with peak supply voltages of ±40 V and a load impedance of 4 Ω, the maximum dissipation of the IC will easily be exceeded. For these reasons, the supply in the present amplifier has been kept down to a safe ±30 V. At these voltages, the chip delivers, without any difficulty, 50 W into an 8 Ω load and 80 W into a 4 Ω load. These are still very respectable figures, particularly in view of the reasonable price of the chip.

CIRCUIT DESCRIPTION

The circuit diagram of the amplifier in **Figure 2** shows that the IC needs only a small number of external components. To keep the harmonic distortion low, the amplifier has a large feedback ratio and its closed-loop gain has been restricted to only 24 dB.

The input signal is applied to pin 3 via capacitor C_1 and low-pass filter R_6-C_{10}. The filter improves the pulse response and flattens the frequency response. For minimum output offset, the values of R_1 and R_3 should be equal, so that the input impedance is 10 kΩ. The roll-off frequencies of R_1-C_1 and R_2-C_2 determine the lower bandwidth limit of the amplifier: with values a specified, this is about 16 Hz. The upper –3 dB point is at about 100 kHz.

The amplifier is muted by a relevant input to pin 10 and placed in the stand-by mode by a relevant signal at pin 9. Muting should always take place before the stand-by mode is selected. Connecting the mute and standby pins permanently to the supply line ensures that the amplifier comes on immediately the power is switched on. Any switch-on clicks may be eliminated by increasing time constants R_3-C_4 and R_5-C_5.

If large-value electrolytic capacitors are used in the power supply, switching off will be rather slow. If that is considered a nuisance, an external mains detection network may be added. This can consist of, say, two diodes and two small smoothing capacitors for rectifying the secondary voltage of the mains transformer. The board has provision for this in the form of additional soldering pins adjacent to the mute and stand-by inputs: an earth pin in case use is made of an external protection circuit and a plus pin if such protection is not foreseen.

CONSTRUCTION

It is best to build the amplifier on the printed-circuit board shown in **Figure 3**. The illustration proves what a compact unit this amplifier is. In view of the scarcity of components, populating the board is very simple.

The back surface of the IC is linked internally to the negative supply rail. Consequently, to preclude electrical contact between the heat sink and the enclosure, the heat sink is mounted on the board. Insulating material between the heat sink and the IC is, therefore, not needed, although the use of some heat conducting paste is advisable.

In the selection of a suitable heat sink, a continuous output of 50 W into 8 Ω was assumed. The selected heat sink is also all right for music outputs of 80 W into 4 Ω. Problems caused by high temperatures are very unlikely, since the IC has internal thermal protection that causes the mute to come into operation at 145 °C and switches the amplifier to stand by at 150 °C.

Provision for connecting the power lines to the board is by three PCB terminal blocks (clamping-screw type). These ensure loss-free passage of the supply current.

The symmetrical power supply is

1

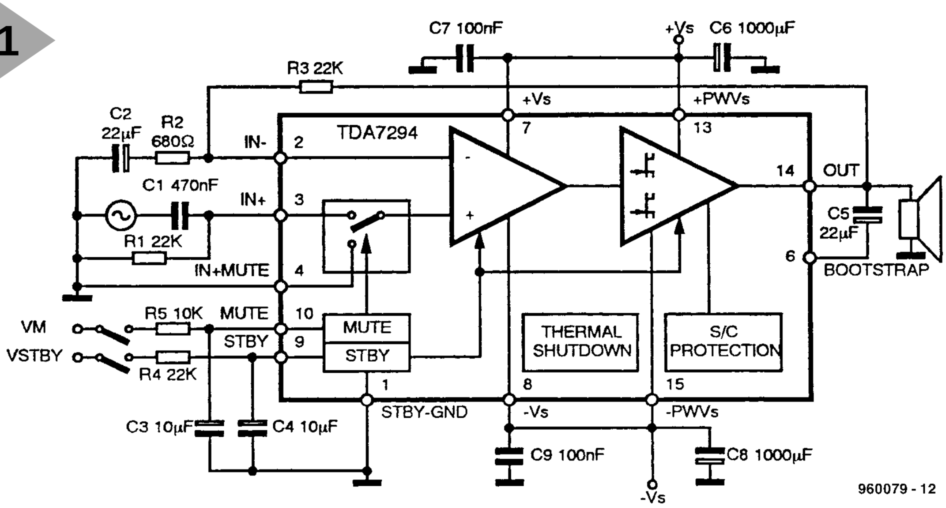
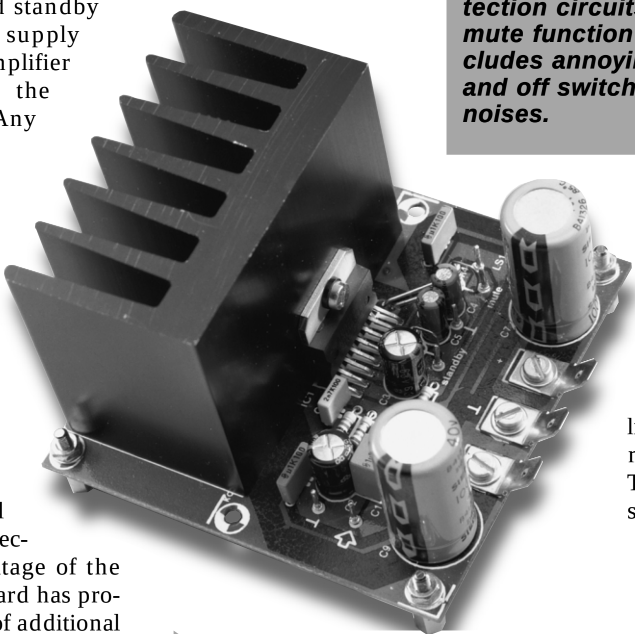


Figure 1. The TDA7294 has standard thermal and short-circuit protection circuits. The mute function precludes annoying on and off switching noises.



2

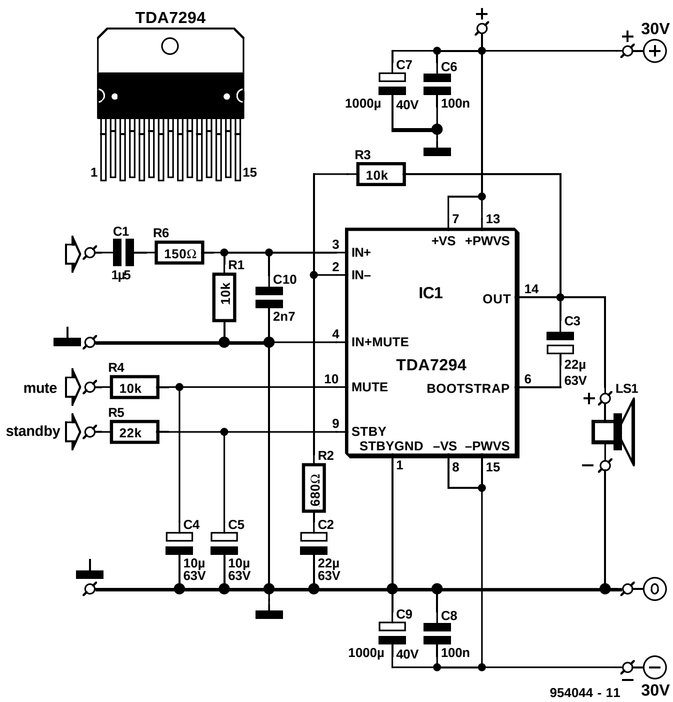


Figure 2. In the final design of the amplifier, supply voltages of ±30 V were decided upon; these are more than sufficient for a power output of 50 W into 8 Ω.

3

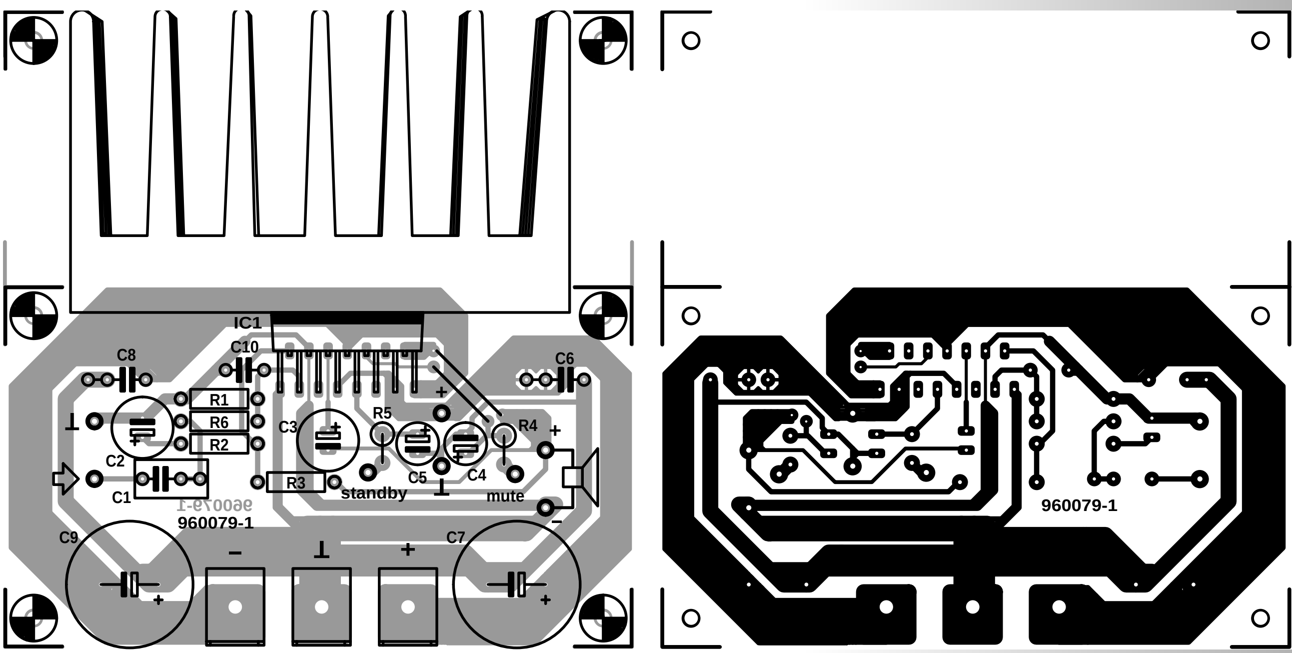


Figure 3. The printed-circuit board is very compact and even houses the requisite heat sink.

best constructed from a toroidal mains transformer, a 25 A bridge rectifier and two 10,000 µF, 50 V electrolytic capacitors.

FINALLY

As mentioned before, thanks to its good performance and high power output, the amplifier is in principle usable in virtually any hi-fi set-up. Owing to its compactness, it is particularly suitable for use in combination with a preamplifier as an integrated amplifier or as part of an active loudspeaker system where space is almost always at a premium.

For those who would like some proof of the figures given in the specification table, **Figure 4** shows the distortion characteristic of the amplifier obtained with a spectrum analyser. The measurements were carried out at an output power of 40 W into 8 Ω and a bandwidth of 80 kHz. As usual, the characteristic slopes upward at higher frequencies, but the distortion does not exceed 0.04 per cent. In a large part of the a.f. range (up to about 1 kHz), the total-harmonic-distortion-plus-noise (THD+N) does not even rise above 0.02 per cent. This sort of performance is excellent for all but the most demanding applications. [960079]

Parts list

Resistors:
R_1, R_3, R_4 = 10 kΩ
R_2 = 680 Ω
R_5 = 22 kΩ
R_6 = 150 Ω

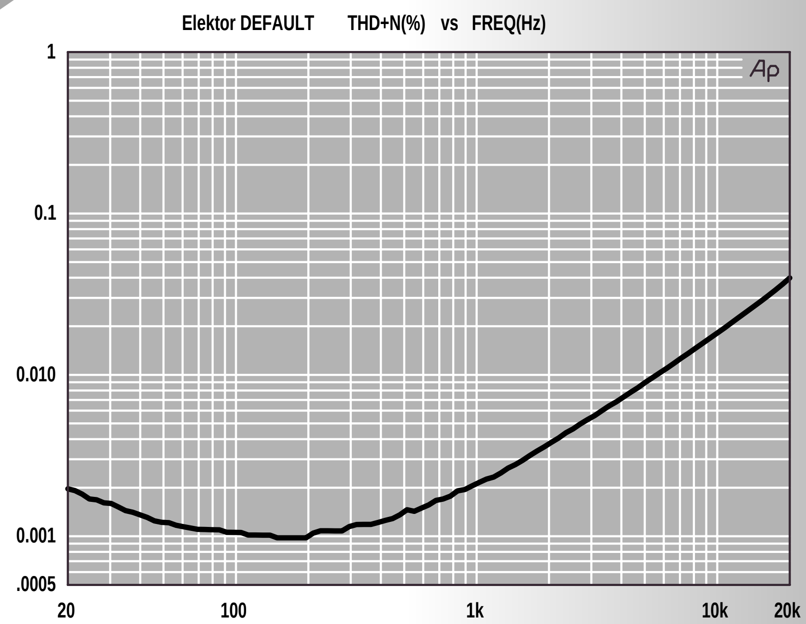
Capacitors:
C_1 = 1.5 µF, 63 V*
C_2, C_3 = 22 µF, 63 V, radial
C_4, C_5 = 10 µF, 63 V, radial
C_6, C_8 = 100 nF
C_7, C_9 = 1000 µF, 40 V, radial
C_{10} = 2.7 nF*, pitch 5 mm
* metallized polyester

Integrated circuits:
IC_1 = TDA7294V

Miscellaneous:
3 off PCB terminal block with clamping screws
1 off heat sink, 2.5 K W^{-1} (e.g. Fischer Type SK100, available from Dau – telephone 01243 553031))
for power supply:
1 off mains transformer, 2×22 V, 80 VA
2 off electrolytic capacitor, 10,000 µF, 50 V
1 off 25 A bridge rectifier
PCB Order no 960079-1 (see Readers' Services towards the end of this issue)

Figure 4. The distortion characteristic, measured with an output power of 40 W into 8 Ω, is excellent for this type of amplifier.

4



Contents

sampling rate converter

Although there are still dyed-in-the-wool sound technicians who swear by the good old analogue recording technology, most others, bitten by the digital audio recording bug, do not want to revert to analogue recording. They have experienced the pleasures of loss-free processing and copying of recordings at digital level. These pleasures turn to frustration, however, when they want to convert a DAT recording on to a CD. This cannot be done just like that because the two have different sampling frequencies: DAT 48 kHz, and CD 44.1 kHz. To overcome this difficulty, a converter as described in this article is required.

Design by T. Giesberts

It is a regrettable fact, with which we will have to learn to live, that different audio techniques use incongruous sampling frequencies (CDI – 18.9 kHz; 8 mm VCR – 31.5 kHz; NICAM – 32 kHz; CDI – 37.8 kHz; VCR – 44.056 kHz; CD – 44.1 kHz; DAT – 48 kHz; and others).

The growing popularity of digital audio is creating an increasing need of some means of coupling equipment using such different techniques – without loss of quality, of course. This can be done by altering the sampling rate in one of the two units to be coupled, while ensuring that the two sampled signals are adequately synchronized. Clearly, this requires a well-designed intelligent converter.

The design of the present converter is based on a dedicated IC: the Type TDA1373H from Philips. This circuit is very versatile and may be used for almost any imaginable conversion (but not quite – see later). Thus, it can be used for converting a DAT recording into a CD recording. Also, it enables CD data to be recorded on a DAT machine with a sampling rate of only 32 kHz, which, of course, results in a much longer playing time. Another possibility is converting the consumer standard S/PDIF* to the professional AES/EBU* format. True, the converter has no AES/EBU connectors, but the conversion is possible.

Apart from as a converter, the circuit may be used as a copybit eliminator. In that case, the two sampling rates are made equal (as in the converter), but the category code, the copybit and the generation-status bit are set. The sampling rate used must correspond with the code set in most DAT machines to ensure that the signal is accepted.

Finally, the circuit may also serve as jitter killer since the first-in-first-out (FIFO) and gain stage in the TDA1373H suppresses any jitter.

DESIGN

The circuit of the converter is shown in the diagram of **Figure 1**. Circuit IC_1 is the integrated digital converter, IC_2 is the controller, and IC_3 is the output interface.

The most important property of the TDA373H is the integrated Audio Digital Input Circuit (ADIC), which enables the chip to decode IEC958 signals (S/PDIF or AES/EBU). The circuit can work on a stand-alone basis or be controlled by a microprocessor. In the present circuit, it is controlled by IC_2, since this gives a wider choice of output formats. The circuit can process up to 20 bits and afterwards provides the con-

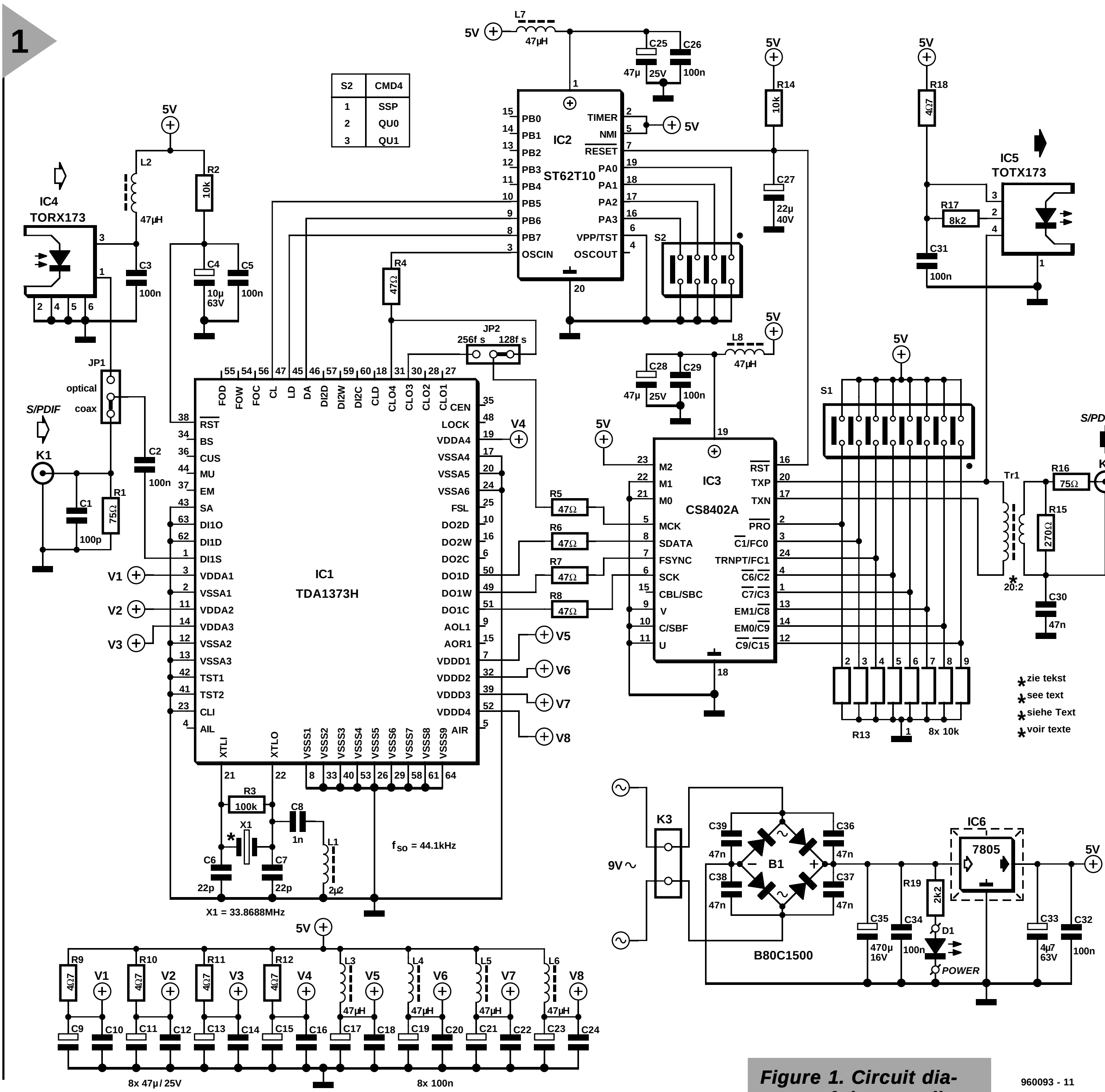


Figure 1. Circuit diagram of the sampling rate converter in which IC_1 is the actual converter, IC_2 is the controller, and IC_3 is the output interface.

verted data in 16-, 18-, or 20-bit format.

The TDA373H is designed to provide up to four different applications, but since the present circuit is geared to being used as a sampling rate converter, the circuit is limited to this application.

The input of the converter may be optical (via IC_4) or coaxial (via K_1). The selection between these two is by jumper JP_1, since it is assumed that the converter will be used invariably in a fixed setup in which there is seldom or never a need for changing from one to the other. The remaining two input pins of IC_1 are linked to earth.

Correct operation of IC_1 requires the setting of six command registers, which is effected by controller IC_2.

After a brief power-up reset (by R_{14}-C_{27}), IC_2 sends twelve 8-bit words (six addresses and data) to IC_1 via a serial connection.

The in-band noise shaper and the stop-band suppression of the ×64

2

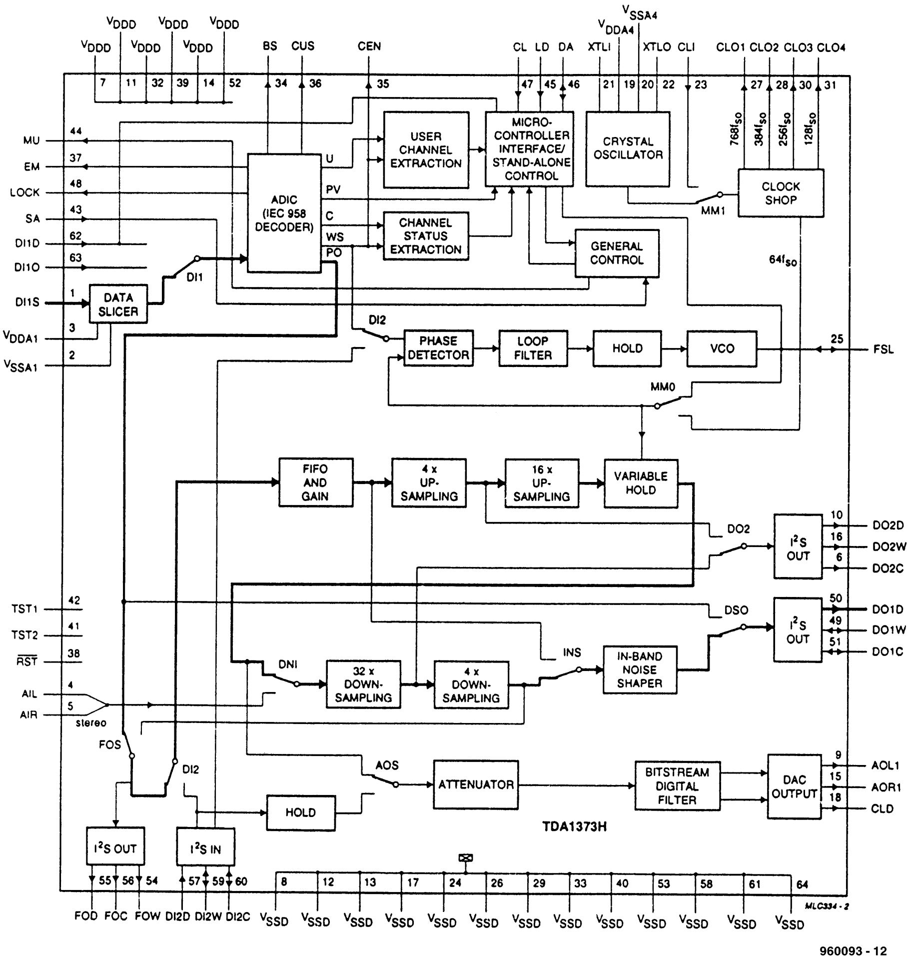


Figure 2. IC_1 affords various functions. The relevant signal paths for the sampling rate conversion are indicated by bold lines.

oversampling filter are set by, respectively, sections 2, 3 and 1 of quadruple DIP switch S_2. The switches are debounced by a delay of 1–1½ seconds between their being operated and the relevant function being actuated. This delay is effected by the controller.

The sampling rate of the output signal, $f_{s(o)}$, of IC_1 is determined by the crystal between pins 21 and 22, according to equation

$$f_{x1} = 768 f_{s(o)}.$$

Thus, the crystal frequency should be 33.8688 MHz for a sampling rate of 44.1 kHz, and 24.576 MHz for a sampling rate of 32 kHz.

The sampling rate of the input, $f_{s(i)}$, must be not lower than 0.35 $f_{s(o)}$, nor higher than 1.45 $f_{s(o)}$. Thus, if the circuit is used as IEC598 decoder only, the input sampling rate should not exceed 45 kHz if the output sampling rate is 32 kHz. If conversion from 48 kHz to 32 kHz is required, consideration should be given to using two converters in cascade.

The converted data are available at serial digital audio output 1 and applied to output interface IC_3 via resistors R_6–R_8, which provide di/dt limiting.

Circuit IC_3 is a digital audio interface transmitter Type CS8402. This IC can also process various formats, but in the present circuit the serial input (pins, 6, 7 and 8) is fixed for I²S by the levels at inputs M0, M1 and M2. Virtually all functions of IC_3 may be obtained by appropriate setting of the various sections of DIP switch S_1.

The symmetrical output at TXP, TXN. is converted into a standard S/PDIF output (0.5 V_{pp} into 75 Ω). The electrical isolation provided by the transformer has the benefit of preventing earth loops.

An optical output is provided by optoisolator IC_5.

The power supply may be based on a standard 9 V mains transformer or mains adaptor rated at not less than 300 mA. The supply lines are stabilized by regulator IC_6 and lavishly decoupled as shown in the diagram.

THE TDA1373H

The TDA1373H, called general digital input, is a circuit that provides four different modes of operation. However, the present application is that of sampling rate converter, SRC, and, therefore, only the parts relevant to this will be discussed in this section.

In the block diagram in **Figure 2** the relevant signal paths are shown in bold lines.

The input signal is applied to the data slicer via pin DI1S. The slicer can handle signals at levels from 200 mV_{pp} to 5 V_{pp}.

The output of the slicer is applied to the audio digital input circuit, **ADIC**, which decodes the stereo audio samples, the word clock, the bit clock and various data (**V**, **U**, **C** and **P**) bits. The last function is not used here.

The **ADIC** locks to a 44.1 kHz signal in not more than 1 ms. Until it has locked, there is no word clock, and the audio bits are muted.

The output of the **ADIC** is applied to

Table 1. Two different response curves of the x64 oversampling filter can be selected with section 1 of DIP switch S_2.

S2-1:				
bit SS	pass band		stop band	
0	0–0.45351$f_{s(i)}$	±0.004 dB	0.54648–1$f_{s(i)}$	–70 dB
1	0–0.46875$f_{s(i)}$	±0.004 dB	0.53125–1$f_{s(i)}$	–50 dB

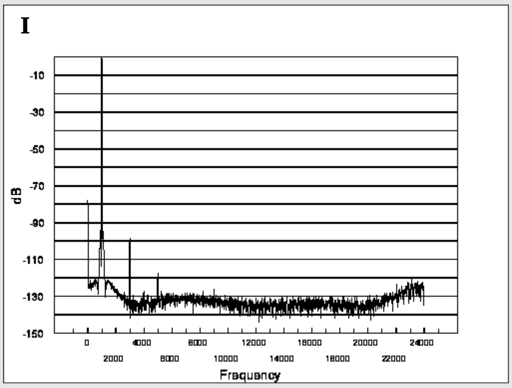
Test results

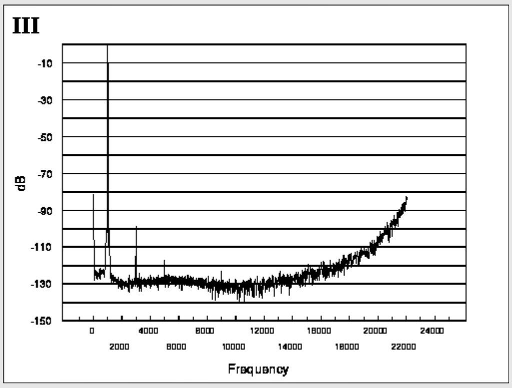
The quality and properties of the sampling rate converter were tested in the digital domain at the various output formats. The (analogue) generator of an Audio Precision Analyser was used as the digital signal source, and this was coupled to a 20-bit analogue-to-digital converter (ADC). Of the four characteristics shown, I represents the output spectrum of the ADC. The other three characteristics may be compared with this.

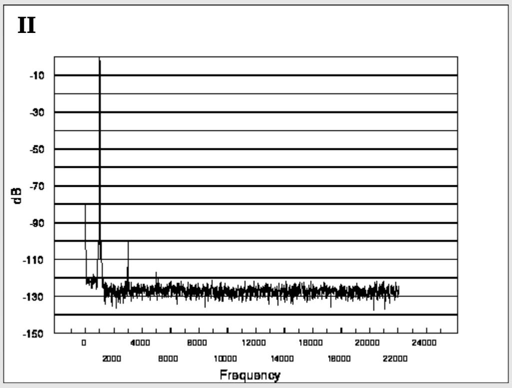
Note how well the action of the in-band noise shaper is seen in Characteristic IV; a clear rise is discernible only in the (virtually inaudible) range above 18 kHz.

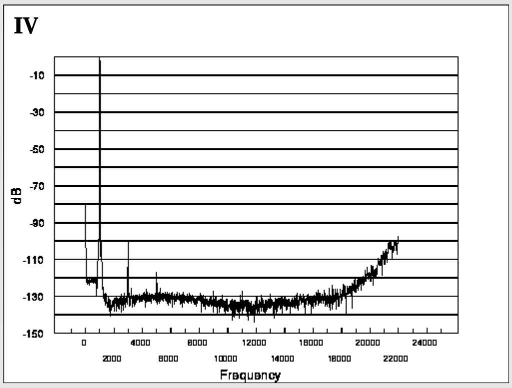
At the same time, and with the same setup, the signal-to-noise ratio for the various output formats was determined:

- *16 bit rounded off: –94.5 dB;*
- *16 bit INS: –89.7 dB;*
- *18 bit INS: –95.5 dB;*









- *II: 16 bit rounded off (section 2 of S_2 on; section 3 of S_2 on).*
- *III: 16 bit INS (section 2 of S_2 on; section 3 of S_2 off).*
- *IV: 18 bit INS (section 2 of S_2 off; section 3 of S_2 off).*

No characteristic is shown for the 20-bit mode (section 2 of S_2 off; section 3 of S_2 on), since, at least up to 20 kHz, this is all but identical to the input signal of the converter (Characteristic I).

- *20 bit: –97 dB.*

Note that the signal-to-distortion ratio of the ADC was 97.5 dB (measured without noise).

The signal-to-noise of the ADC plus sampling rate converter (without distortion) in the same test setup was about –107 dB (dynamic range of the ADC). The dynamic range, measured at a digital-to-analogue converter, DAC, was about 5 dB better with 16 bit INS than with 16 bit rounded off.

the first-in-first-out, **FIFO**, and gain stage. The **FIFO** section equalizes any speed variations of the incoming samples. Its size is eight samples and it ensures a tracking speed of 4 kHz ms^{-1}.

The gain section enables the signal to be amplified or attenuated. In the present application, the signal is attenuated by 0.068 dB to prevent clipping in the digital filters.

The samples are fed for interpolation to a ×64 oversampling filter. This filter consists of a ×4 section and a ×16 section. There is a choice of two filter characteristics: one with

Table 2. The word-lenght of the samples may be adapted with section 2 and 3 of DIP switch S_2.

S_2-2:	S_2-3:	
bit QU0	bit QU1	lenght of word
0	0	16 bit (rounded off)
1	0	20 bit
0	1	16 bit INS
1	1	18 bit INS

a stop-band suppression of 70 dB and the other with a stop-band suppression of 50 dB but with steeper skirts. The second is intended especially for use with signals with a sampling rate of 32 kHz, and a pass-band of 0–15 kHz as, for instance, in digital satellite radio. The characteristics are selected by section 1 of DIP switch S_2 – see also **Table 1**.

The samples are applied from the filter to the variable hold stage in which the actual sampling rate conversion takes place. Depending on the ratio of the input sampling rate and the output sampling rate, the sample is used once or twice, whence the name variable hold. When the ratio is 1:1, all samples are used twice, since the down sampling is ×128.

The variable hold function is controlled by a digital phase-locked loop, **PLL**, formed by the phase detector, loop filter, hold, and **VCO** (voltage-controlled oscillator) stages.

The loop filter ensures that the **PLL** locks rapidly. This is necessary because, after power-up, the bandwidth of the filter is reduced in two steps from 500 Hz to 50 Hz and then from 50 Hz to 0.5 Hz. The frequency difference is only 1 Hz for 512 input samples (10 ms for a sampling rate of 44.1 kHz).

After the **PLL** has locked in, the audio signals are demuted and the conversion commences. To prevent any errors, the **FIFO** is monitored continuously in the variable-hold phase. As soon as the slightest tracking error is detected, the bandwidth of the loop filter is enlarged.

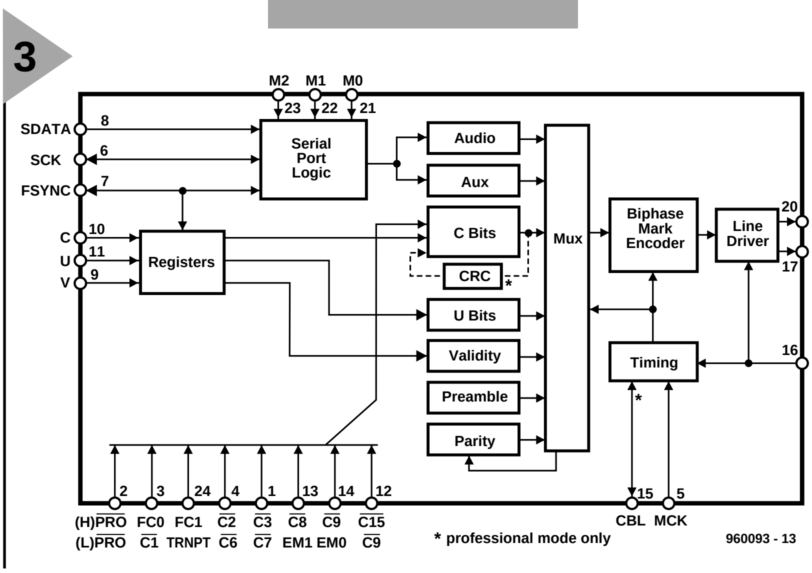
To convert the sampling rate to the requisite output value, the samples are passed through a ×128 down-sampling filter, which consists of a ×32 section and a ×4 section. The overall filter provides a stop-band suppression of 80 dB from 0.54648 of the output sampling rate.

Finally, the samples are applied to an in-band noise shaper, **INS**, which adapts the word-length of the samples to specific requirements. The standard length of 20 bits may be reduced to 16 bits or 18 bits by the relevant sections of **DIP** switch S_2. There are four possibilities as enumerated in **Table 2**, the first of which is 20 bits.

The **INS** has a facility for adapting the digitization noise in a psycho-acoustical manner, whereby the noise to which human hearing is most sensitive is shifted upwards in frequency. This facility gives a subjective improvement of two bits with respect to the real quantization level.

Finally, the 20 bit length may be reduced to a 16-bit length by a simple rounding off action.

Figure 3. IC_3 is a digital audio interface transmitter intended to encode and send audio data in accordance with the usual standards.



CODING & CONTROL

The digital audio interface transmitter, IC_3, is intended primarily for coding and sending audio data according to usual interface standards. The circuit provides the possibility of setting the most important channel status bits via seven inputs: pins 3, 24, 4, 1, 13, 14 and 12 in **Figure 3**. These inputs are controlled by octal **DIP** switch S_1 (see Figure 1). All seven inputs have a double function, which depends on the level at pin 2. This level is set with section 8 of S_1 and determines whether the IC works in the professional (**AES/EBU**) or in the consumer (**S/PDIF**) mode. The audio data are coded to the standard associated with the selected mode.

In the professional mode, a **CRC** code may be generated (channel status byte 23) as shown in dashed lines.

The serial input, pins 6, 7 and 8, can handle seven different formats and audio samples of 16–24 bits. In the present circuit, the format is fixed for I^2S with M0, M1 and M2.

The serial inputs for channel status, **C**, user data, **U**, and validity, **V**, are not used and linked to earth. The **V** bit must be low to indicate that audio data are being processed which can be converted to analogue signals.

Pin 15, channel block start, **CBL**, is not used in the present application either. Normally, it is an output that may be used for writing **C**, **U** and **V** bits. CBL is an input only when in the professional mode the transparent option is chosen (in which the **C**, **U** and **V** bits can be looped in via a receiver). In this way, synchronization of signals coming from separate equipment is possible. Normally, the master clock, MCK, is $128 f_s$, where f_s is the signal frequency, but in the transparent mode, MCK = $256 f_s$. The multiplier is set with jumper JP_2.

Table 3a. Sampling rates in professional mode.

S_1-8:	S_1-5:	S_1-4:	
PRO	C6	C7	
0	0	0	not defined
0	0	1	48 kHz
0	1	0	44.1 kHz
0	1	1	32 kHz

PROFESSIONAL MODE

When pin 2, $\overline{\text{PRO}}$, is low, that is, when section 8 of S_1 is closed, the digital audio interface transmitter, IC_3, is in the professional mode. In this mode, bits 1, 2, 3, 4, 6, 7 and 9 may be set after a 1 has been sent for channel status bit 0.

C0 indicates whether the channel status block applies to the professional (1) or the consumer (0) mode.

C1 determines whether the data are audio (0 – section 6 of S_1 closed) or

Table 3b. Sampling rates in consumer mode.			
S_1-8:	*S_1-6:*	*S_1-7:*	
PRO	*FC1*	*FC0*	
1	*0*	*0*	*44.1 kHz*
1	*0*	*1*	*48 kHz*
1	*1*	*0*	*32 kHz*
1	*1*	*1*	*44.1 kHz, CD-mode*

not audio (1 – section 6 of S_1 open).

C2, C3 and C4 are coded by EM0 (section 2 of S_1) and EM1 (section 3 of S_1) and determine the emphasis to be used: for instance, 110 is 50/15 µs.

C6 and C7 determine the sampling rate. The requisite setting of the relevant sections of S_1 is given in **Table 3a**.

Table 4. Setting the category code.			
S_1-8:	*S_1-3:*	*S_1-2:*	
PRO	*C8*	*C9*	
1	*0*	*0*	*general format*
1	*0*	*1*	*PCM-encoder/decoder*
1	*1*	*0*	*CD*
1	*1*	*1*	*DAT*

A 1 at C9 (section 2 of S_1 open) indicates a stereo signal; a 0 means that the mode is indeterminate.

In the transparent mode, none of the stated pins is used: the channel code is read at the C input only.

CONSUMER MODE

When pin 2, $\overline{\text{PRO}}$, is high (section 8 of S_1 open), the digital audio interface transmitter is in the S/PDIF (consumer) mode. In this mode, bits 2, 3, 8, 9, 15, 24 and 25 may be set after a 0 has been sent for channel status bit 0.

C0 – 0 – indicates that the channel status block applies to the consumer mode.

FC0 and FC1 determine the sampling rate. The requisite setting of the relevant sections of S_1 is given in **Table 3b**.

C2 gives a choice between copy prohibit (0 – section 5 of S_1 closed) or copy permit (1 – section 5 of S_1 open).

C3 determines whether emphasis (50/15 µs) will be applied (1 – section 4 of S_1 open) or not (0 – section 4 of S_1 closed).

C8 and C9 determine the category code: the requisite setting of the relevant sections of S_1 is shown in **Table 4**.

C15 is the generation status bit. Depending on the category code, the function of this bit, determined by the setting of section 1 of S_1, is inverted. When the category code is 001xxxx, 0111xxx or 100xxxx, a 0 indicates that the bit is an original and a 1 that it is a copy. With all other category codes, the reverse is true.

When the copy bit is 1, there is no copy protection, so that copying is possible. If the copy bit is 0, the generation status bit, in combination with the category code, determines whether copying may be carried out or not.

The outputs of IC_3 are RS422 compatible line drivers that go low when the IC is being reset.

Make sure that the channel status bits are active low and that a 0 is set by closing the relevant section of the DIP switch.

CONSTRUCTION

The converter is best built on the printed-circuit board shown in **Figure 4**. Although the board is compact, it affords ample space for all parts and components, including audio sockets K_1 and K_2 and optoisolators IC_4 and IC_5. It does not have space for the mains transformer, however. Note that controller IC_2 is available ready programmed through our Readers services (see towards the end of this issue).

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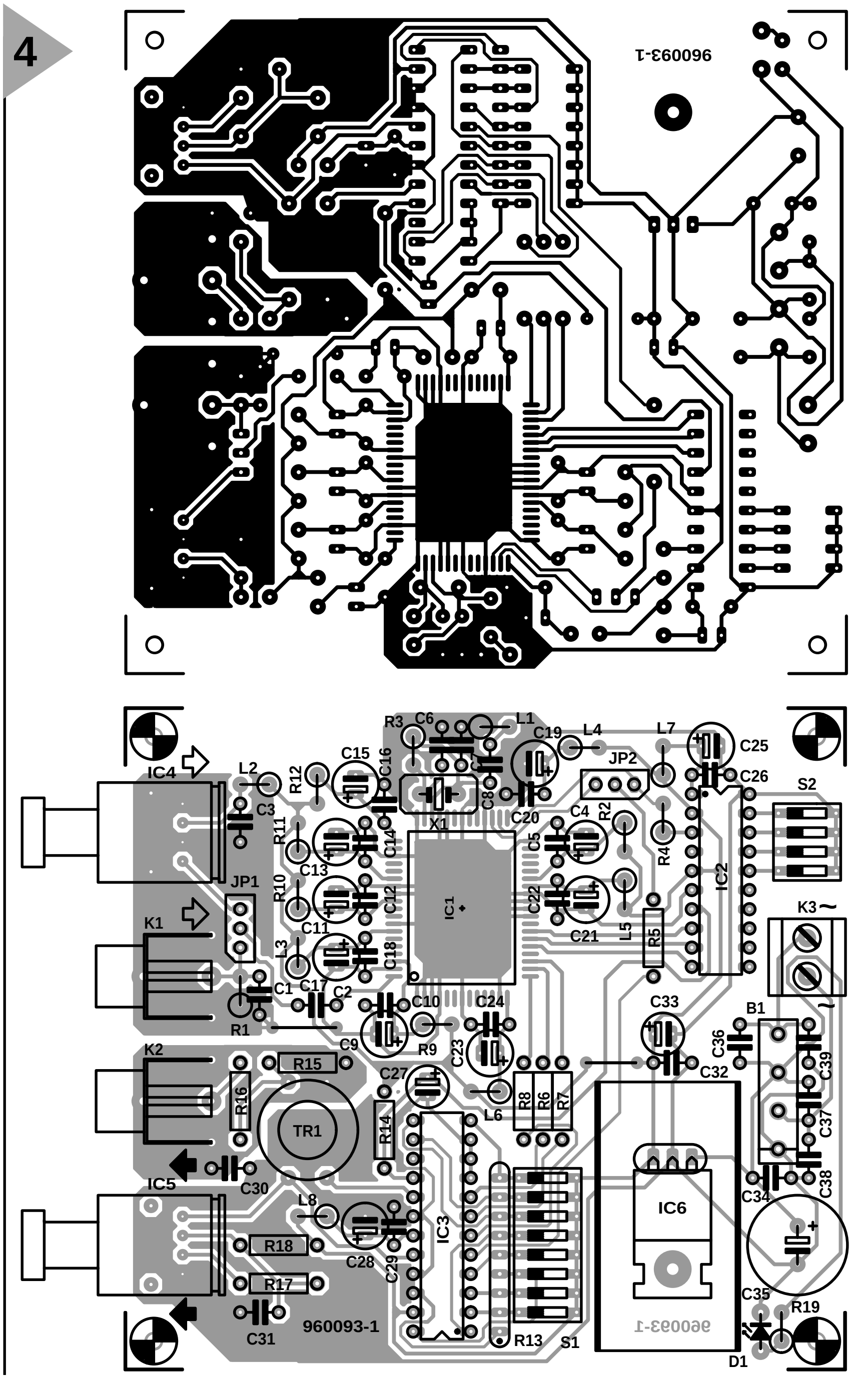

Figure 4. Printed-circuit board for the sampling rate converter. Note that IC_1, a surface-mount device, must be soldered at the track side.

Contents

PARTS LIST

Resistors:
R_1, R_{16} = 75 Ω
R_2, R_{14} = 10 kΩ
R_3 = 100 kΩ
R_4–R_8 = 47 Ω
R_9–R_{12}, R_{18} = 4.7 Ω
R_{13} = 8×10 kΩ array
R_{15} = 270 Ω
R_{17} = 8.2 kΩ
R_{19} = 2.2 kΩ

Capacitors:
C_1 = 100 pF
C_2, C_3, C_5, C_{10}, C_{12}, C_{14}, C_{16}, C_{18}, C_{20}, C_{22}, C_{24}, C_{26}, C_{29}, C_{31}, C_{32}, C_{34} = 100 nF ceramic
C_4 = 10 µF, 63 V, radial
C_6, C_7 = 22 pF
C_8 = 1 nF ceramic
C_9, C_{11}, C_{13}, C_{15}, C_{17}, C_{19}, C_{21}, C_{23}, C_{25}, C_{28} = 47 µF, 25 V, radial
C_{27} = 22 µF, 40 V, radial
C_{30}, C_{36}–C_{39} = 47 nF, ceramic
C_{33} = 4.7 µF, 63 V, radial
C_{35} = 470 µF, 16 V, radial

Inductors:
L_1 = 2.2 µH
L_2–L_8 = 47 µH

Semiconductors:
D_1 = LED, low current

Integrated circuits:
IC_1 = TDA1373H (Philips)
IC_2 = ST6210 (see Readers services)
IC_3 = CS8402A (Crystal)
IC_4 = TORX173 (Toshiba)
IC_5 = TOTX173 (Toshiba)
IC_6 = 7805

Miscellaneous:
JP_1, JP_2 = 3-way pin header and jumper
K_1, K_2 = audio socket for board mounting
K_3 = terminal block, pitch 5 mm
S_1 = octal DIP switch
S_2 = quadruple DIP switch
Tr_1 = wound on G2/3FT12 core - see text
B_1 = rectifier Type B80C1500
X_1 = see text
Heat sink for IC_6: 29 k W^{-1}, for instance, Fischer ICK35/SA, available from Dau, telephone 01243 553 031)
PCB Order no. 960093 (see Readers services)
Mains transformer or mains adaptor, 9 V, 300 mA

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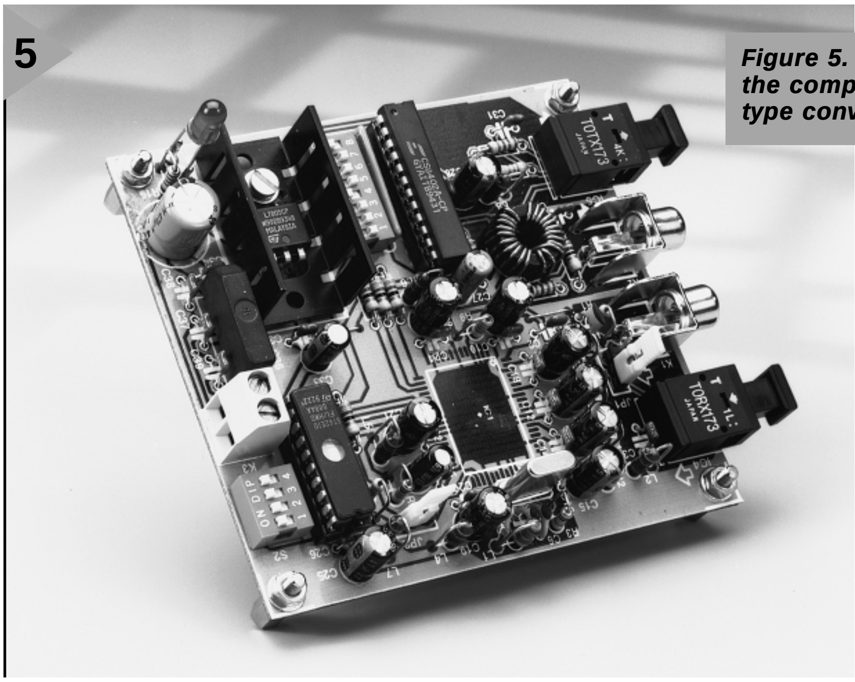

Figure 5. Top view of the completed prototype converter board.

The completion of the board should not present any undue difficulties – see **Figure 5** for a view of the top of a completed board. Note that IC_6 is mounted on an appropriate heat sink.

Less experienced constructors may find the mounting of IC_1 and the construction of output transformer Tr_1 not so straightforward.

Circuit IC_1 is a surface-mount device, SMD, which should be soldered at the track side (underside) of the board using an iron with a very fine tip – see **Figure 6**. Note the correct way of fitting: pin 1 is identified by a small disk on the case: this side must point towards the connectors on the board.

Transformer Tr_1 is a DIY device which is wound on a G2/3FT12 core. The primary winding consists of 20 turns, and the secondary of two turns, of enamelled copper wire with a diameter of 0.7 mm.(SWG22). Spread the primary evenly across the core, but leave some space at the centre for the two secondary turns that have to be wound subsequently.

6

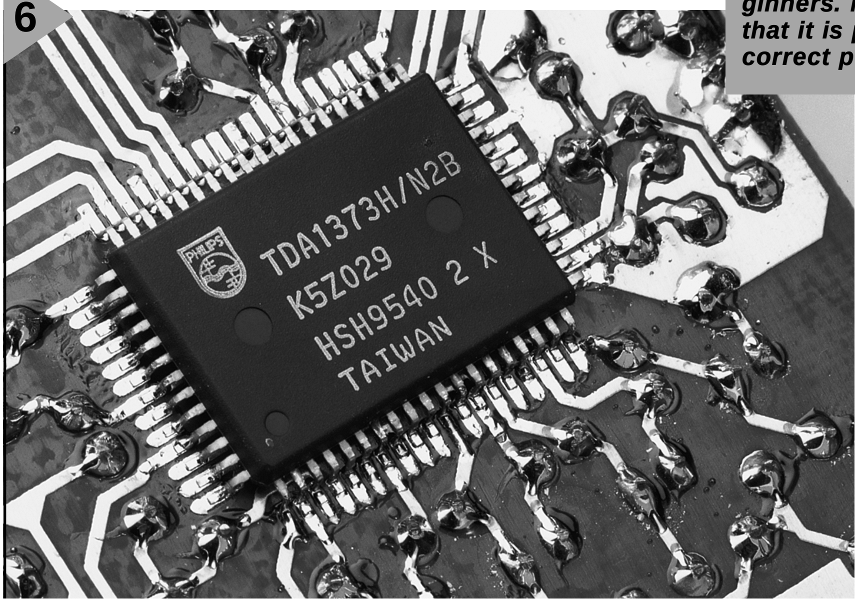

Figure 6. Mounting IC_1 may be a somewhat difficult task for beginners. Make sure that it is placed in the correct position.

When the board has been completed and checked thoroughly, a suitable mains transformer (9 V, 300 mA) or a 9 V mains adaptor may be connected to K_3. Indicator D_1 should then light.

Using a multimeter, check whether a stable potential of 5 V exists across C_{32} and C_{33}. If so, it is virtually certain that the converter will work satisfactorily. If it does not, recheck the board thoroughly. It is not possible to give suitable test points, since all that can really be checked are the supply lines.

(960093-I)

battery-operated pre-amplifier - part 1

a unit for purists

This preamplifier is intended for those audio enthusiasts who are interested in only the best. Its discrete design produces distortion figures that are measured in tenthousandths of a per cent (the remainder of its specification looks very good, too!). Moreover, bearing in mind that in certain areas the mains supply voltage is not as pure as it should be, the preamplifier is powered by rechargeable NiMH batteries. In this first of a two-part article, the pre-amplifier proper will be described. The second part, to be published next month, will deal with the specially designed battery charger.

Design by T. Giesberts

The design of high-end audio equipment is typified by the continual striving for the best possible quality. In the high-end design world, bass boost networks, equalizers and impressive displays are not contemplated. In fact, such ancillary facilities are considered undesirable. The only aspect that is of importance is the quality of the signal processing. This quality is the be all and end all as far as real audio enthusiasts are concerned: no concessions are given or asked for. Only true quality is worthy of the name hi-fi.

A difficulty arises when the specifications of an amplifier are so good that the designer is testing at the limits of his measuring equipment. Real improvements are then not possible or, in any case, cannot be measured. Dyed-in-the-wool audio enthusiasts then seek enhancement in psychoacoustic matters such as solid silver connecting cables, gold-plated connectors and bizarre units to treat CDs. The sense of such matters is a topic of argument among themselves and questionable to outsiders. Certainly, they cannot be measured or made audible.

A more tangible means of enhancing the performance of an amplifier is the eradication of all aspects that, outside the normal signal path, may affect the performance. One of those aspects is the mains supply. In many countries, this is far from stable, while interfering spikes and other undesirable facets are the rule rather than the exception. This justifies the assumption that it must be possible to improve the amplifier's performance by powering it not from the mains but from a battery. This assumption has been put into practice in the present preamplifier. The mains voltage is used solely to operate the battery charger and this only when the amplifier is switched off.

THREE SECTIONS

The design of the preamplifier proper is not really spectacular. True, the fact that no integrated circuits are used in the signal path is conspicuous but nothing special. The exemplary good specifications are nothing new because they have been achieved in earlier preamplifiers. The property that differentiates the present preamplifier from its predecessors is that these specifications are allied with battery operation.

The block diagram of the preamplifier in **Figure 1** shows that it may be divided into three sections, each of which is housed on a discrete printed-circuit board.

The input selector is a passive stage which needs little comment. After all, a six-position selector is a familiar sight in many amplifiers.

The preamplifier proper consists of an input buffer, balance control, volume control, amplifier stage, and a relay that obviates annoying on/off clicks and acts as a detector of the bat-

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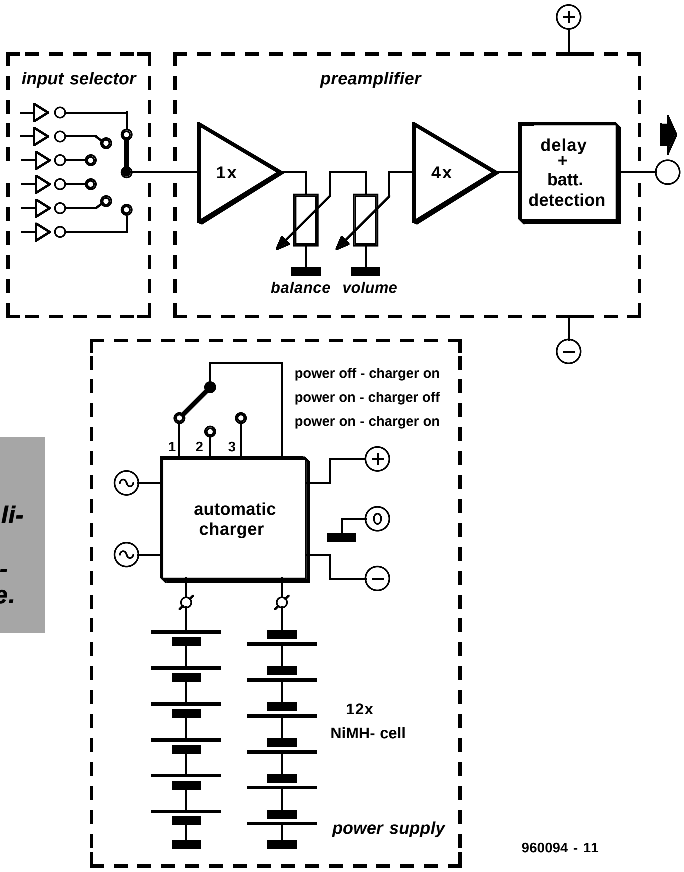


Figure 1. Block diagram of the battery-operated AF preamplifier. Only the upper two sections are discussed in this article.

Test results

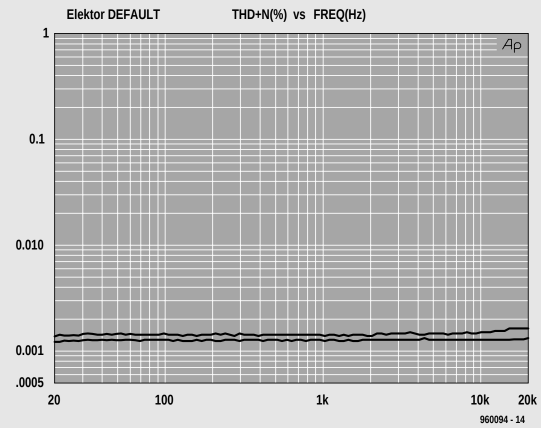
Signal-to-noise ratio	*P2 max*	*108 dBA*
(at 1 V r.m.s. output)	*P_2 min*	*106 dBA*
THD+N at 1 kHz and 1 V r.m.s. output		*0.00070%*
		(bandwidth 22 kHz)
Channel separation	*1 kHz*	*>92 dB*
	20 kHz	*>67 dB*
Crosstalk	*1 kHz*	*<–107 dB*
	20 kHz	*<–84 dB*
Input impedance		*23.6 kΩ*
Output impedance		*100 Ω*
Tape-out impedance		*500 Ω*
Sensitivity for 1 V r.m.s. output		*260 mV*
Bandwidth		*DC to 200 kHz*
Maximum input voltage		*4.5 V r.m.s.*
Maximum output voltage		*5.5 V r.m.s.*
Current drain		*21 mA*

All tests were carried out with a supply voltage of ±8 V (batteries fully charged) and the relevant inputs terminated by a 560 Ω impedance.

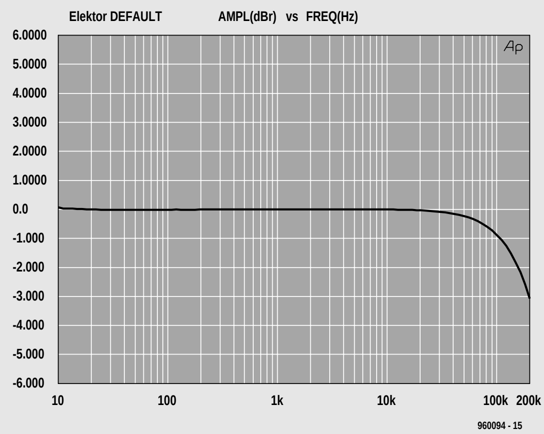
***Figure A** shows the THD+N characteristics for two dissimilar inputs. The upper curve pertains to an input of 2 V r.m.s., and the lower one to an input of 260 mV r.m.s. In both cases, the output was 1 V r.m.s. and the bandwidth was 80 kHz. It is clear that the differences between the two are very small; the respective distortion figures were 0.0014% and 0.0012%.*

***Figure B** shows the frequency response of the preamplifier. Note that this is entirely in accord with the specified bandwidth: the upper cut-off frequency is exactly 200 kHz.*

A



B



tery voltage.

The circuits are designed to operate from a symmetrical supply of ±7.2 V, which is provided by twelve NiCd batteries or twelve NiMH batteries.

The third section contains the power supply, consisting of the twelve batteries and a microprocessor-controlled charger. A three-position switch selects one of two operating positions or an emergency position.

In position 1, the power to the preamplifier is switched off and the batteries are being charged. In position 2, the charger is switched off and the preamplifier is powered by the batteries. In position 3 – an emergency position for use only when the batteries are flat – power is supplied to the preamplifier and the batteries are being charged. In practice, position 3 may never be used, since the preamplifier can operate from the batteries for up to 50 hours, while charging flat batteries takes only 2–3 hours.

When the preamplifier is not being used, the batteries are kept fully charged. When the fast charging process is over, the charger automatically goes on to trickle-charging. More about this in Part 2.

2

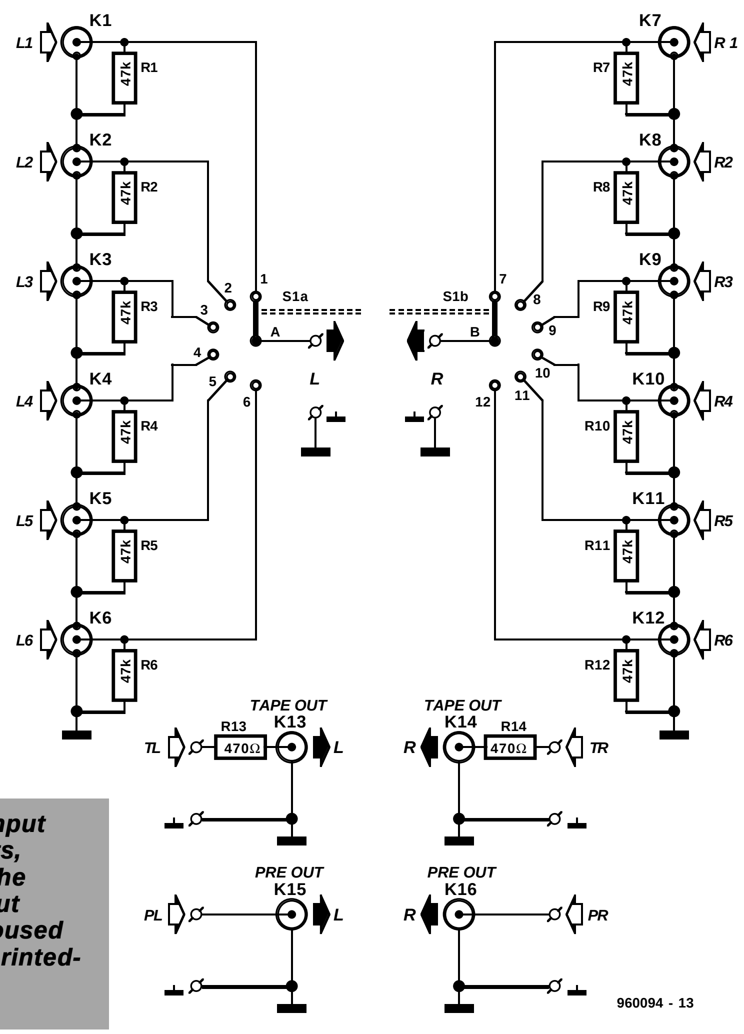


Figure 2. The input selector circuits, together with the input and output sockets, are housed on a discrete printed-circuit board.

THE CIRCUIT DIAGRAM

Although some ICs are used for special functions of the preamplifier, the signal processing circuits are all discrete designs. Moreover, virtually all components used are standard parts, so that availability of them should not cause undue difficulties.

The circuit of the input selector is shown in **Figure 2**. There is not much that needs to be said about this, because it is limited to a rotary switch, a number of resistors, and 12 input sockets. Switch S_1 provides a choice between six inputs at line levels which in practice is more than sufficient.

The input selector is housed on its own printed-circuit board together with output sockets K_{15} and K_{16}, and tape out buses K_{13} and K_{14}.

The diagram of the electronic circuits of the preamplifier is shown in **Figure 3**. This contains the complete stereo circuits: those for the right-hand channel in the top half and those for the left-hand channel in the lower half. In the following description, reference will be made to the right-hand channel only.

The input buffer is formed by stages T_1–T_7. Potentiometers P_1 and P_2 are the balance and volume controls respectively.

Stages T_8–T_{22} are amplifiers. Circuit IC_1 is a servo control for offset compensation. The combination of IC_3 and relay Re_1 provides switch-on delay (to obviate on/off clicks) and supply voltage detection.

BUFFER

Since there is a practical limitation to the number of batteries that can or should be used, a battery supply necessarily means that the supply lines are low. In the present circuit, the level is ±7.2 V. Because of this low level, amplification of the signal takes place only after the balance and volume controls. This makes it possible for input signals of a few volts to be processed.

It is clear from the foregoing that the buffer stage preceding the balance and volume controls is little more than a sophisticated emitter-follower. Cascading two emitter-followers, T_1-T_3 and T_2-T_4, renders the input and output of the buffer stage theoretically offset-free to make direct coupling possible. In practice, it is necessary for the pairs of emitter-followers to be preselected on the basis of identical amplification factor and U_{BE}, and to be thermally coupled.

The direct current operating point of the buffer, which is operating in Class A, is provided by current sources T_5 and T_6. The references for these sources are D_1 and D_2, the current through which is held constant by T_{17}. To further minimize any drift, the current sources and the diodes are thermally coupled.

High frequency interference is suppressed by a low-pass filter, R_1-C_1, at the input of the buffer stage. The output of the buffer stage is not only applied to the balance and volume controls, but is also used as the tape-out signal.

AMPLIFIER SECTION

As for other parts of the circuit, a low supply voltage and minimum current drain were important design parameters for the amplifier stages.

Obviously, the design is totally symmetrical. The input circuits are formed by two complementary differential amplifiers, T_8–T_{11}, each of which is provide with a discrete current source, T_{12} and T_{13}. In view of the required stability, transistors pairs T_8-T_9 and T_{10}-T_{11} are thermally coupled. To keep the offset low, it is desirable for the transistor pairs to be preselected on the basis of (near)identical specifications.

The outputs of the differential amplifiers drive push-pull stage T_{15}-T_{16}, which in turn actuates the output stage.

Because of the low supply voltage,

the output stage is a compound one, a kind of darlington, which produces an amplification of about ×1.7. This gain gives the other stages rather more room to operate, and ensures that the peak output voltage is all but equal to the supply voltage.

The quiescent current of the output stage is held constant by intercoupling T_{21} and T_{22} via 'zener transistor' T_{17}-T_{18}. The zener voltage, and consequently the quiescent current (about 2 mA), is set with P_3.

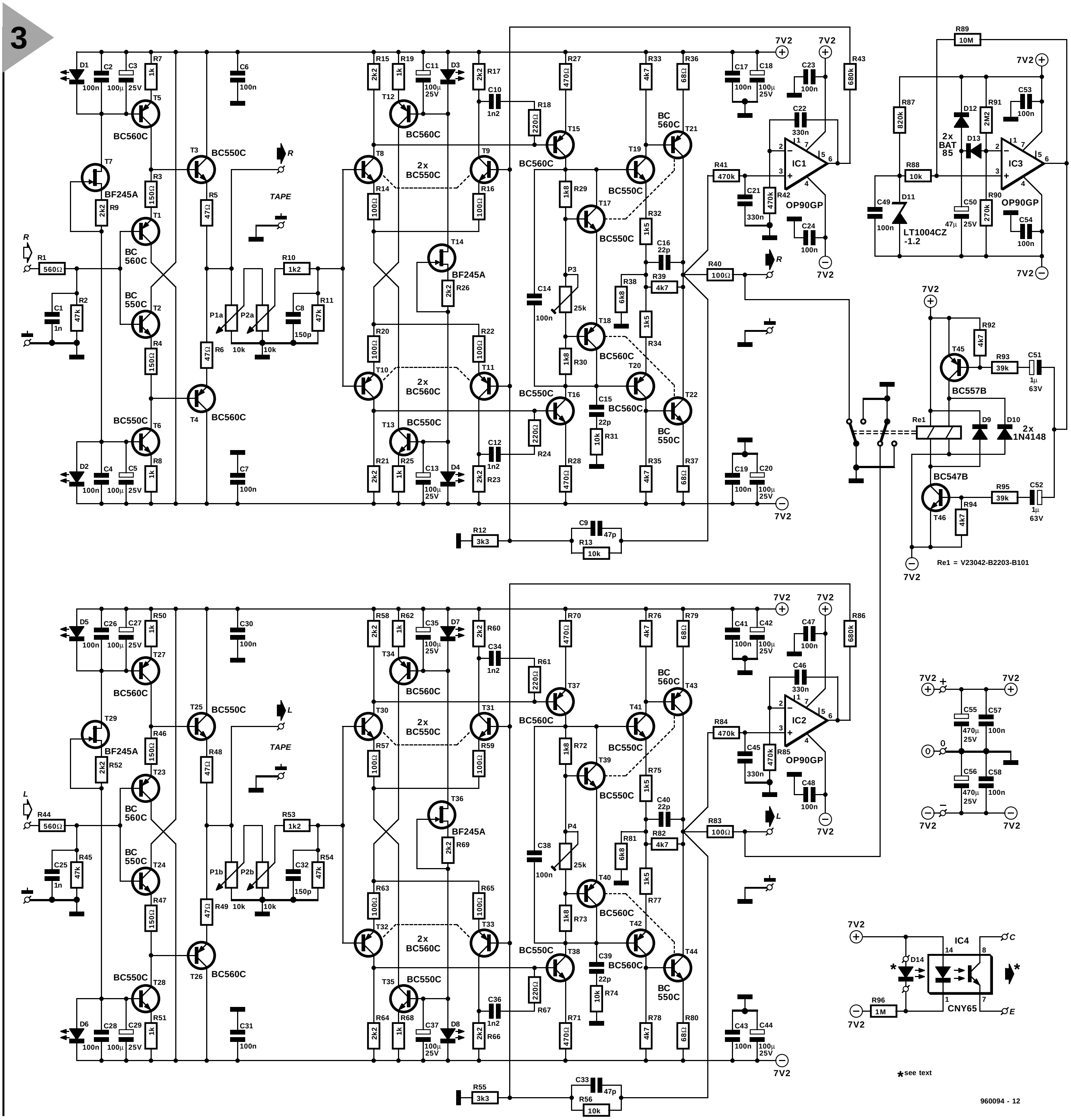
Combinations R_{18}-C_{10}, R_{24}-C_{12}, R_{31}-C_{15}, and capacitor C_{16}, are compensating networks. The open-loop amplification is ×1700, and the open-loop bandwidth is about 2 kHz.

Since even with matched transistors the symmetry of the amplifier will never be 100 per cent, offset compensation is provided by a servo-control, which is based around precision low-voltage micropower op amp IC_1. The low current drain of 20 µA and the low input offset of typically 0.13 mV make this op amp particularly suitable for this application.

Any offset voltage present at the output is amplified and fed back via R_{43} to the base of T_9-T_{11}. The actual offset at the output of the preampli-

Figure 3. The AF pre-amplifier consists of a buffer and an amplifier separated by the balance and volume controls. Circuits IC_1 and IC_2 form a servo control circuit for offset compensation. The circuit based on IC_3 and Re_1 is a switch-on delay and supply voltage detector.

3



fier is, therefore, minimal in all circumstances.

RELAY STAGE

As mentioned before, any on/off switching clicks and other noises are suppressed through the use of a relay, Re_1. It is noteworthy that the relay contacts are not, as usual, in series with the output signal; here suppression is obtained by short-circuiting the signal to ground. This has the advantage that the relay contacts are not in the signal path during normal operation and, therefore, cannot adversely affect the signal quality. Resistor R_{40} ensures that the short-circuiting has no detrimental effects on the amplifier.

It is also worth noting that the relay is not of the usual kind, but a bistable type. Such a relay needs a pulse of only a few milliseconds to be actuated: this prevents unnecessary current drain.

The relay has two isolated windings, of which one is used to actuate the make contact, and the other to reset the relay.

Transistors T_{45} and T_{46} function as pulse-shapers. The switching pulse originates in the very brief charging current of C_{51} or C_{52}. Which of these capacitors will be charged depends on the output state of IC_3.

Comparator IC has a threefold function: (1) it provides a delay at switch-on; (2) it acts as a detector of the supply voltage, and (3) it provides rapid suppression of the output voltage when the preamplifier is switched off.

In essence, functions (2) and (3) are identical since in both the relay switches over as soon as the supply voltage drops below about 12 V. To this end, the potential at junction R_{90}-R_{91}, which is derived from the supply line, is compared with the reference voltage provided by D_{11}. This diode is a special micropower reference type that needs a current of only 10 µA. The feedback provided by R_8 and R_{89} ensures smooth operation of the comparator around the toggle voltage.

Delayed switch-on is provided by C_{50}. At switch-on, this capacitor must be charged via D_{13} before IC_3 toggles. Diode D_{13} ensures that the potential at junction R_{90}-R_{91} drops immediately the supply voltage decreases. The capacitor is discharged via diode D_{12} when the supply voltage approaches 0 V.

The value of C_{50} gives a delay time of 10–15 seconds, depending on the supply voltage. This time is inversely proportional to the supply voltage. A very long delay therefore indicates that something is amiss with one of the batteries or that the charger is faulty. The delay time may be shortened by lowering the value of C_{50} to some extent.

DISPLAY

To obtain an indication that the preamplifier is switched on, low-current diode D_{14} may be energized via R_{96}. If this is desired, the resistor should have a value of about 6.8 kΩ. It is also possible to obtain an indication via optoisolator IC_4, with which the state of the batteries can be displayed in two colours. An advantage of the optoisolator is that it draws a current of only 15 µA from the batteries. In this case, the value of R_{96} should be 1 MΩ; the LED is energized directly by the charger supply line.

The on/off indication, as well as the construction and the operation of the charger, will be further discussed in Part 2 which is intended to be published in the February issue.

[960094-1]

CONSTRUCTION GUIDELINES

Elektor Electronics (Publishing) does not provide parts and components other than PCBs, front panel foils and software on diskette or IC (not necessarily for all projects). Components are usually available form a number of retailers – see the adverts in the magazine.

Large and small values of components are indicated by means of one of the following prefixes :

E (exa)	= 10^{18}	a (atto)	= 10^{-18}
P (peta)	= 10^{15}	f (femto)	= 10^{-15}
T (tera)	= 10^{12}	p (pico)	= 10^{-12}
G (giga)	= 10^{9}	n (nano)	= 10^{-9}
M (mega)	= 10^{6}	µ (micro)	= 10^{-6}
k (kilo)	= 10^{3}	m (milli)	= 10^{-3}
h (hecto)	= 10^{2}	c (centi)	= 10^{-2}
da (deca)	= 10^{1}	d (deci)	= 10^{-1}

In some circuit diagrams, to avoid confusion, but contrary to IEC and BS recommandations, the value of components is given by substituting the relevant prefix for the decimal point. For example,

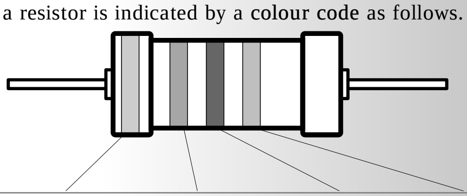
3k9 = 3.9 kΩ   4µ7 = 4.7 µF

Unless otherwise indicated, the tolerance of resistors is ±5% and their rating is ⅓–½ watt. The working voltage of capacitors is ≥ 50 V.

In populating a PCB, always start with the smallest passive components, that is, wire bridges, resistors and small capacitors; and then IC sockets, relays, electrolytic and other large capacitors, and connectors. Vulnerable semiconductors and ICS should be done last.

Soldering. Use a 15–30 W soldering iron with a fine tip and tin with a resin core (60/40) Insert the terminals of components in the board, bend them slightly, cut them short, and solder: wait 1–2 seconds for the tin to flow smoothly and remove the iron. Do not overheat, particularly when soldering ICS and semiconductors. Unsoldering is best done with a suction iron or special unsoldering braid.

The value of a resistor is indicated by a colour code as follows.



color	1st digit	2nd digit	mult. factor	tolerance
black	–	0	–	–
brown	1	1	$\times10^{1}$	±1%
red	2	2	$\times10^{2}$	±2%
orange	3	3	$\times10^{3}$	–
yellow	4	4	$\times10^{4}$	–
green	5	5	$\times10^{5}$	±0,5%
blue	6	6	$\times10^{6}$	–
violet	7	7	–	–
grey	8	8	–	–
white	9	9	–	–
gold	–	–	$\times10^{-1}$	±5%
silver	–	–	$\times10^{-2}$	±10%
none	–	–	–	±20%

Examples:
brown-red-brown-gold = 120 Ω, 5%
yellow-violet-orange-gold = 47 kΩ, 5%

Faultfinding. If the circuit does not work, carefully compare the populated board with the published component layout and parts list. Are all the components in the correct position? Has correct polarity been observed? Have the powerlines been reversed? Are all solder joints sound? Have any wire bridges been forgotten?

If voltage levels have been given on the circuit diagram, do those measured on the board match them – note that deviations up to ±10% from the specified values are acceptable.

Possible corrections to published projects are published from time to time in this magazine. Also, the readers letters column often contains useful comments/additions to the published projects.

battery-operated AF pre-amplifier - part 2

power supply and construction

Design by T. Giesberts

Following the detailed description of the preamplifier in last month's instalment, this second and final part of the article deals with the power supply and the construction of the complete unit. More so than usual, the power supply forms an important part of the overall design. It consists of 12 NiMH batteries, size AA (RG/HP7) with a nominal capacity of 1.2 Ah, and a moderately fast charger that takes about three hours to fully charge a set of flat batteries (note that it is not a good idea to discharge these batteries completely – the nominal voltage level is maintained at 1.2 V during 80% of the discharge cycle).

Since the current drain of the preamplifier is not more than 21 mA, a set of fully charged 12 NiMH batteries will enable the preamplifier to operate about 50 hours continuously. As this type of operation is seldom, if ever, required, there is normally plenty of time for the batteries to be recharged. Note, by the way, that NiMH batteries are free of the undesirable memory effect. Nevertheless, NiCd batteries may be used if for one reason or another NiMH batteries cannot be obtained. Unlike NiMH batteries, NiCd cells will show some degradation of capacity in the long term.

The charger is based on the well-known Type TEA1101 IC, which is eminently suitable for this purpose since it uses the ΔU method for controlling the charging current. With this method, the terminal voltage of the battery rises gradually when it is being charged. When the battery is fully charged, its temperature rises, which causes a

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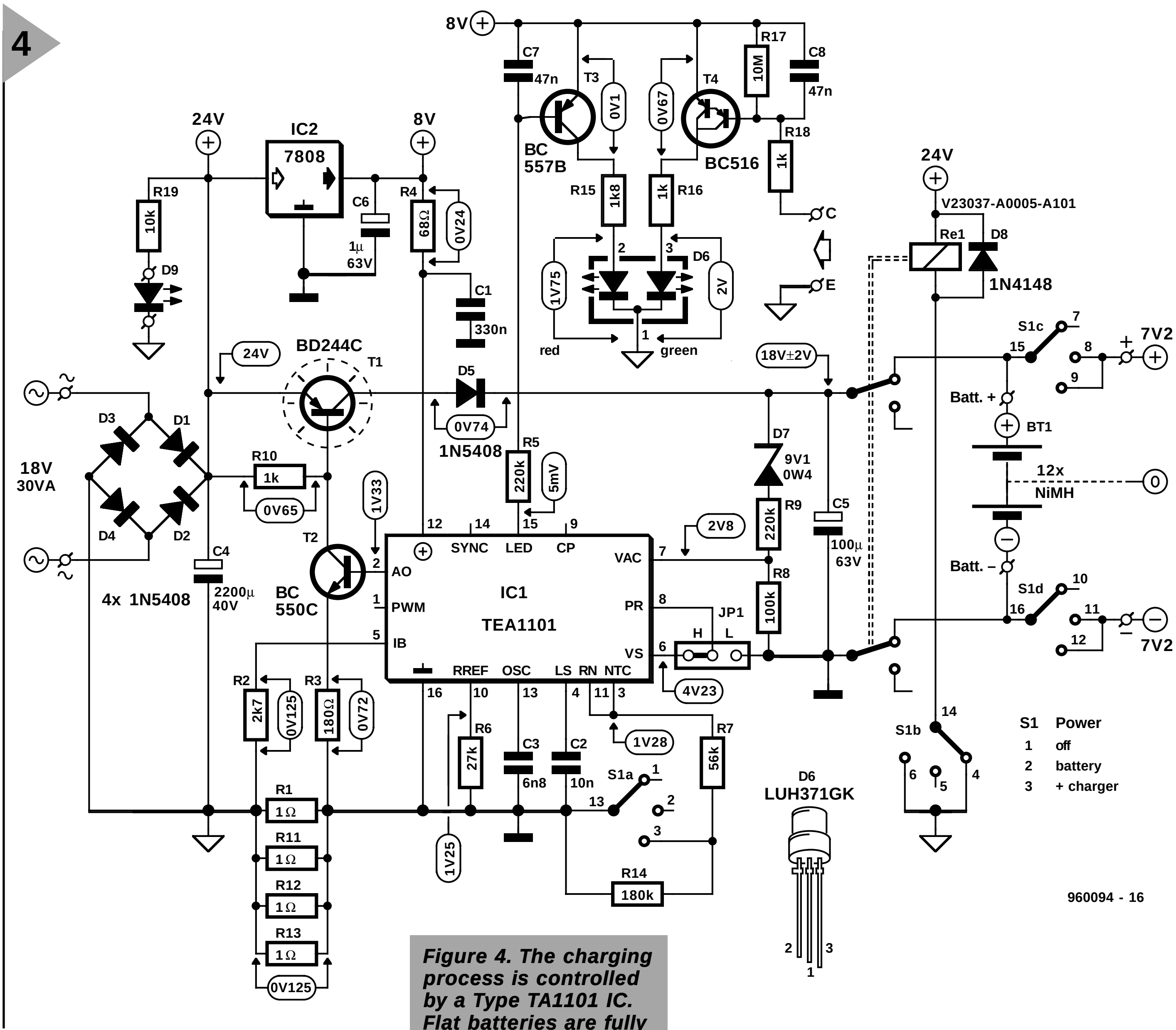

Figure 4. The charging process is controlled by a Type TA1101 IC. Flat batteries are fully charged in about 3 hours. After this period, trickle-charging begins.

slight drop in terminal voltage (ΔU). The TEA1101 monitors this process continuously and regularly cuts off the charging current whereupon a precise measurement of the terminal voltage takes place. When the circuit registers the drop in terminal voltage, the full charging current is switched off and replaced by a trickle-charging current.

The level of the full charging current is 500 mA, which is high enough to charge the batteries in a fairly short time, but not so high as to require temperature monitoring of the cells. The trickle-charging current is 5 mA, but this can be altered slightly if desired.

CIRCUIT DESCRIPTION

In the circuit diagram in **Figure 4**, switch S_1 is a quadruple three-position switch. In position 1 (off), the link between the batteries and the preamplifier is broken by S_{1C} and S_{1D} – the preamplifier is then off. At the same time, double-pole relay Re_1 is energized via S_{1B}. The relay contacts connect the charger to the batteries, whereupon the batteries are being charged. Note that this happens only when the charger is linked to the mains supply, since otherwise the relay cannot be energized.

In position 2 (battery), the preamplifier is connected to the batteries via S_{1C} and S_{1D}, so that it is switched on. At the same time, the relay is deactuated via S_{1B}, so that the charger is disconnected from the batteries.

In position 3 (+ charger), which should not often be used, the preamplifier is switched on and the batteries are being charged. This position is really only for those situations where the batteries are flat and the preamplifier is to be used.

The various modes of operation are indicated by a number of LEDs. One of these, D_9, is the on/off indicator for the charger. The red half of D_6, a dual diode, is controlled by the LED output of IC_1 and lights up continuously during normal charging and flashes during trickle-charging.

If terminals C and E are linked to the corresponding terminals on the preamplifier, lighting of the green half of D_6 shows that the preamplifier is switched on.

Both halves of D_6 may light to give a mixture of colours. A continuous orange colour shows that the preamplifier is on and that the batteries are being charged. Green combined with flashing orange indicates that the preamplifier is on and that the batteries are being trickle-charged.

Some people may find it odd that the batteries are drawn as one unit, whereas in reality they are arranged in two sets of six, each of which provides one half of the symmetrical supply voltage. They are charged in series. The common line is taken from the junction between the 6th and 7th battery.

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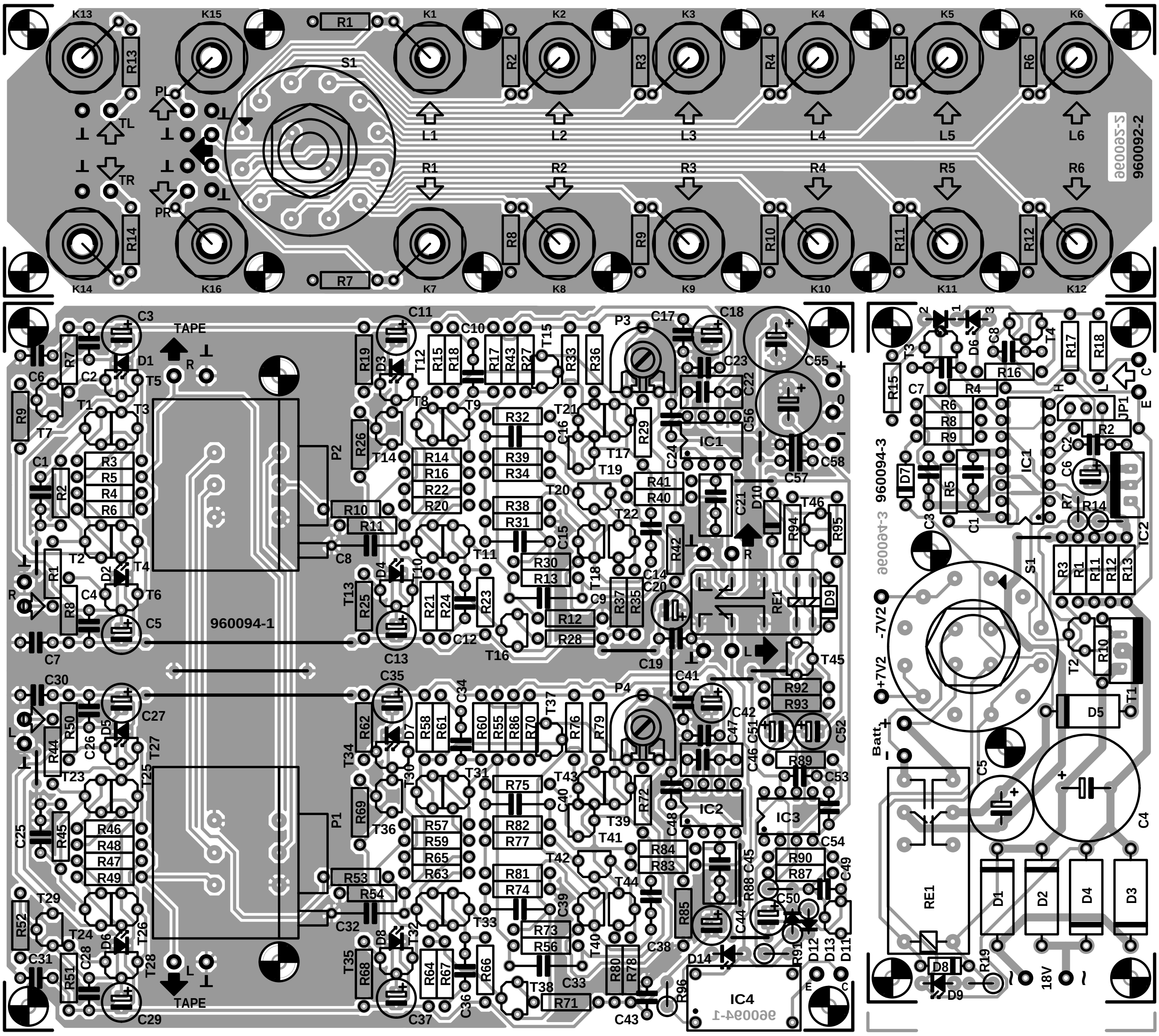


Parts list PREAMPLIFIER

Resistors:
R_1, R_{44} = 560 Ω
R_2, R_{11}, R_{45}, R_{54} = 47 kΩ
R_3, R_4, R_{46}, R_{47} = 150 Ω
R_5, R_6, R_{48}, R_{49} = 47 Ω
R_7, R_8, R_{19}, R_{25}, R_{50}, R_{51}, R_{62}, R_{68} = 1 kΩ
R_9, R_{15}, R_{17}, R_{21}, R_{23}, R_{26}, R_{52}, R_{58}, R_{60}, R_{64}, R_{66}, R_{69} = 2.2 kΩ
R_{10}, R_{53} = 1.2 kΩ
R_{12}, R_{55} = 3.3 kΩ
R_{13}, R_{31}, R_{56}, R_{74}, R_{88} = 10 kΩ
R_{14}, R_{16}, R_{20}, R_{22}, R_{40}, R_{57}, R_{59}, R_{63} R_{65}, R_{83} = 1-- Ω
R_{18}, R_{24}, R_{61}, R_{67} = 220 Ω
R_{27}, R_{28}, R_{70}, R_{71} = 470 Ω
R_{29}, R_{30}, R_{72}, R_{73} = 1.8 kΩ
R_{32}, R_{34}, R_{75}, R_{77} = 1.5 kΩ
R_{33}, R_{35}, R_{39}, R_{76}, R_{78}, R_{82}, R_{92}, R_{94} = 4.7 kΩ
R_{36}, R_{37}, R_{79}, R_{80} = 68 Ω
R_{38}, R_{81} = 6.8 kΩ
R_{41}, R_{42}, R_{84}, R_{85} = 470 kΩ
R_{43}, R_{86} = 680 kΩ
R_{87} = 820 kΩ
R_{89} = 10 MΩ
R_{90} = 270 kΩ
R_{91} = 2.2 MΩ
R_{93}, R_{95} = 39 kΩ
R_{96} = 1 MΩ
P_1 = 10 kΩ stereo linear, special balance (Alps)
P_2 = 10 kΩ stereo log (Alps)
P_3, P_4 = 25 kΩ preset

Capacitors:
C_1, C_{25} = 1 nF
C_2, C_4, C_6, C_7, C_{14}, C_{17}, C_{19}, C_{23}, C_{24}, C_{26}, C_{28}, C_{30}, C_{31}, C_{38}, C_{41}, C_{43}, C_{47}, C_{48}, C_{49}, C_{53}, C_{54}, C_{57}, C_{58} = 100 nF ceramic
C_3, C_5, C_{11}, C_{13}, C_{18}, C_{20}, C_{27}, C_{29}, C_{35}, C_{37}, C_{42}, C_{44} = 100 μF, 25 V, radial
C_8, C_{32} = 150 pF, 160 V, polyester
C_9, C_{33} = 47 pF, 160 V, polyester
C_{10}, C_{12}, C_{34}, C_{36} = 1.2 nF
C_{15}, C_{16}, C_{39}, C_{40} = 22 pF, 160 V, polyester
C_{21}, C_{22}, C_{45}, C_{46} = 330 nF
C_{50} = 47 μF, 25 V, radial
C_{51}, C_{52} = 1 μF, 63 V, radial
C_{55}, C_{56} = 470 μF, 25 V, radial;

Semiconductors:
D_1–D_8 = LED, red, rectangular, 5 mm
D_9, D_{10} = 1N4148
D_{11} = LT1004CZ-1.2 (Linear Technology)
D_{12}, D_{13} = BAT85
D_{14} = low-current, green, 5 mm
T_1, T_4, T_5, T_{10}, T_{11}, T_{12}, T_{15}, T_{18}, T_{20}, T_{21}, T_{23}, T_{26}, T_{27}, T_{32}, T_{33}, T_{34}, T_{37}, T_{40}, T_{42}, T_{43} = BC560C
T_2, T_3, T_6, T_8, T_9, T_{13}, T_{16}, T_{17}, T_{19}, T_{22}, T_{24}, T_{25}, T_{28}, T_{30}, T_{31}, T_{35}, T_{38}, T_{39}, T_{41}, T_{44} = BC550C
T_7, T_{14}, T_{29}, T_{36} = BF245A
T_{45} = BC557B
T_{46} = BC547B

Integrated circuits:
IC_1–IC_3 = OP90GP (Analog Devices)
IC_4 = CNY65 (Temic/Telefunken)

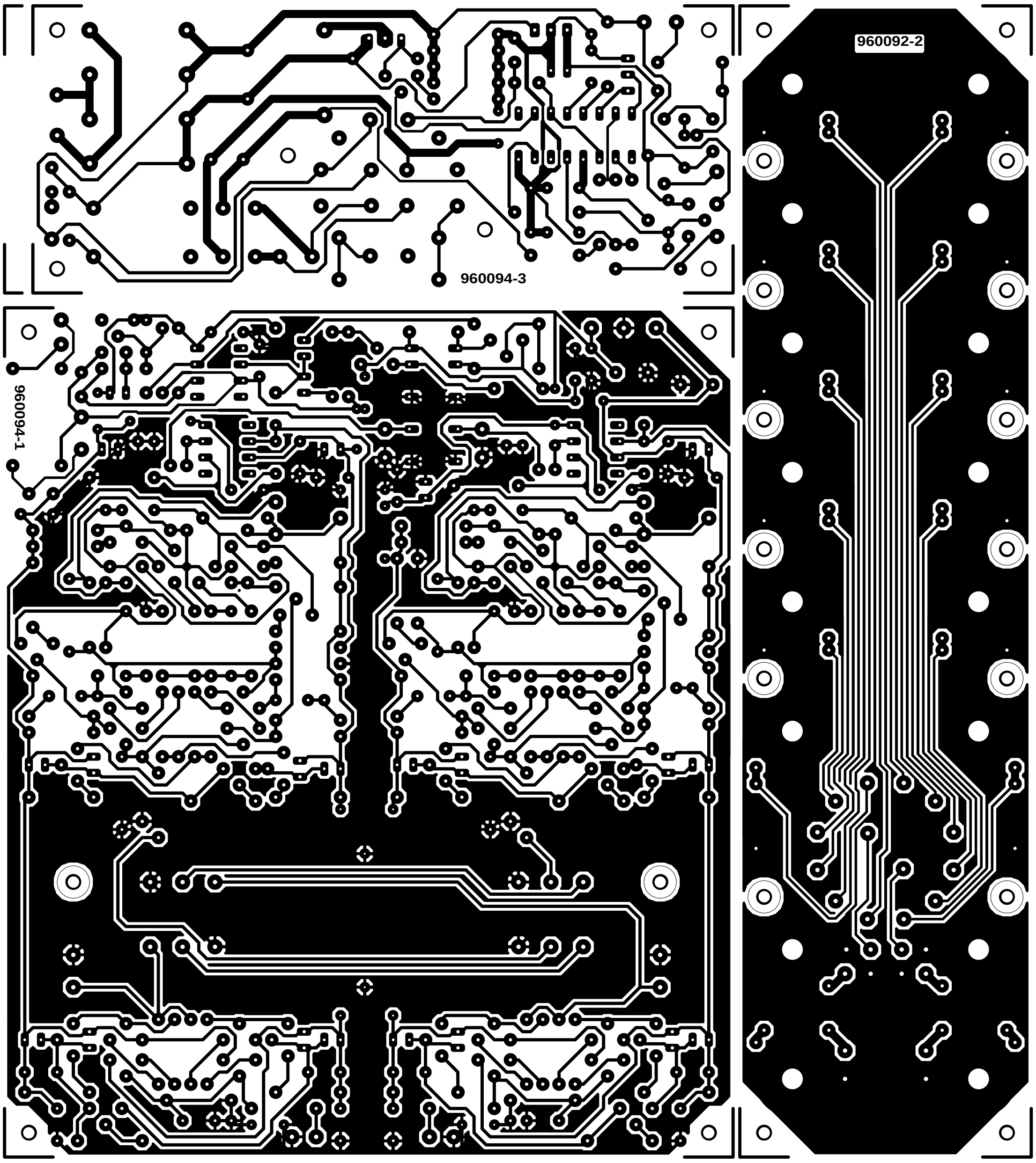
Miscellaneous:
Re_1 = bistable relay, 2 change-over contacts
Case 300×57×235 mm (12×2¼×9¼ in), e.g. Monacor UC-202H/SW

Figure 5. The printed-circuit board for the input selector, preamplifier and charger must be cut into three as indicated before any further work is done.

Parts list INPUT SELECTOR

Resistors:
R_1–R_{12} = 47 kΩ
R_{13}, R_{14} = 470 Ω

Miscellaneous:
K_1–K_{16} = audio socket (preferably gold-plated) for chassis mounting
S_1 = rotary switch, 2-pole, 6-position, for board mounting



Parts list CHARGER

Resistors :
R_1, R_{11}, R_{12}, R_{13} = 1 Ω
R_2 = 2.7 kΩ
R_3 = 180 Ω
R_4 = 68 Ω
R_5, R_9 = 220 kΩ
R_6 = 27 kΩ
R_7 = 56 kΩ
R_8 = 100 kΩ
R_{10}, R_{16}, R_{18} = 1 kΩ
R_{14} = 180 kΩ
R_{15} = 1.8 kΩ
R_{17} = 10 MΩ
R_{19} = 10 kΩ

Capacitors:
C_1 = 330 nF
C_2 = 10 nF
C_3 = 6.8 nF
C_4 = 2200 µF, 40 V, radial
C_5 = 100 µF, 63 V, radial
C_6 = 1 µF, 63 V, radial
C_7, C_8 = 47 nF

Semiconductors:
D_1–D_5 = 1N5408
D_6 = dual LED, 4 mm, common cathode
D_7 = zener diode 9.1 V, 400 mW
D_8 = 1N4148
D_9 = LED, low current, 3 mm
T_1 = BD244C
T_2 = BC550C
T_3 = BC557B
T_4 = BC516

Integrated circuits:
IC_1 = TEA1101 (Philips)
IC_2 = 7808

Miscellaneous:
JP_1 = 3-way header and jump lead
S_1 = rotary switch, 4-pole, 3-position, for board mounting
Re_1 = relay, 24 V, 1100 Ω coil, 2 change-over contacts
Mains transformer 18 V, 30 VA secondary, preferably toroidal
Mains fuse = 160 mA, slow-acting
Single-pole mains on /off switch

Contents

6

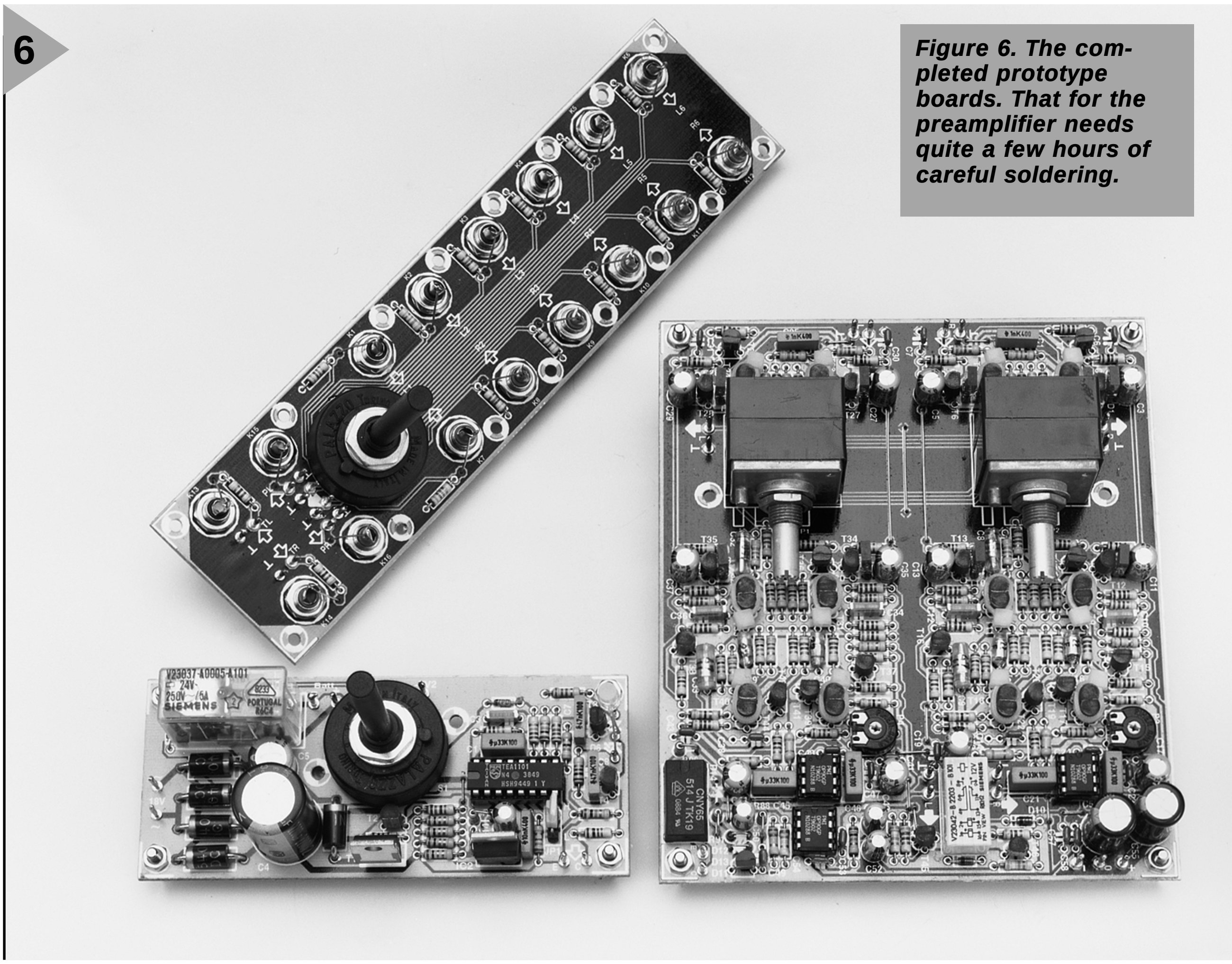


Figure 6. The completed prototype boards. That for the preamplifier needs quite a few hours of careful soldering.

BATTERY MONITOR

The remainder of the circuit in Figure 4 is a linear regulator arranged in the standard application of the TEA1101 suggested by the manufacturers.

Voltage regulator IC_2 provides the 8 V supply for IC_1.

The secondary voltage of the mains transformer is rectified by D_1–D_4 and smoothed by C_4. The value of this electrolytic capacitor is purposely taken higher than strictly required to enable the circuit to be used with higher charging currents.

The control loop is formed by transistor T_1, the batteries, current sensor R_1-R_{11}-R_{12}-R_{13}, and driver T_2. The latter transistor is controlled via pin 2 (A0) of IC_1. Diode D_5 prevents the batteries being discharged via T_1.

The prime function of IC_1 is performed by a monitor and control network whose input is pin 7 (VAC). The input to this pin is part of the battery voltage taken from potential divider D_7-R_8-R_9. The internal monitor regularly checks the potential at pin 7. During these periodic tests, the charging current is discontinued via pin 2. Each measured voltage is quantized and compared with the previous one. If the new value is higher, it is stored; if it is lower, a check is made whether the difference (ΔU) is greater than 0.25% (owing to the presence of D_7, this comes down to 0.125%). If the latter is the case, normal charging is replaced by trickle-charging.

When at the onset of the charging cycle the value at pin 7 is below the reference value of 380 mV, IC_1 arranges for the cycle to start with trickle-charging. Only when the potential at pin 7 rises above the reference value does normal charging begin. If no difference is measured, IC_1 acts as if the battery is defect and disconnects the charging current.

When at the start of the charging cycle the potential at pin 7 is high, IC_1 acts as if no battery is connected and resets the circuit.

The value of the charging current is determined by resistor R_6 (connected to pin 10): with the resistor value as specified, it is 500 mA.

The level of the trickle-charging current is determined normally by R_7-R_{14}: with values of these components as specified, the current is 5 mA. With S_1 in position 3, charging takes place with the preamplifier on and the trickle-charging current must then be increased to compensate for the current drain of the preamplifier. This is achieved by S_{1A} short-circuiting R_{14}, whereupon the trickle-charging current rises to about 25 mA.

The duty factor of the trickle-charging pulses may be lowered with jumper JP_1. This may be handy in case good-quality batteries are used which have no measurable self-discharge. Normally, JP_1 should be in position H; when it is set to position L, the duty factor is lowered by 75%, which means an effective drop of the trickle-charging current to 1.25 mA.

It should be borne in mind that for both the quality of the preamplifier and the correct functioning of the charger it is vital that the transfer resistance in the battery holders is kept as low as possible. The total transfer resistance should not exceed 4.5 Ω. So, it is advisable to use high-quality holders, preferably with sintered terminals.

CONSTRUCTION

The entire unit, that is, the input selector, preamplifier proper and charger, is best built on the printed-circuit board shown in **Figure 5**. This consists of three sections which, before any further work is done, should be cut apart.

Completing the input selector board is simplicity itself, since it contains only the input and output sockets, the selector switch and terminal resistors.

The preamplifier board is densely populated and needs to be completed with great care. Accurate soldering is of prime importance, since a number of terminals and copper tracks are very close together.

It is of vital importance that transistor sets T_1-T_3, T_2-T_4, T_8-T_9, T_{10}-T_{11}, T_{17}-T_{21}, and T_{18}-T_{22} are in good thermal contact. Therefore, their cases

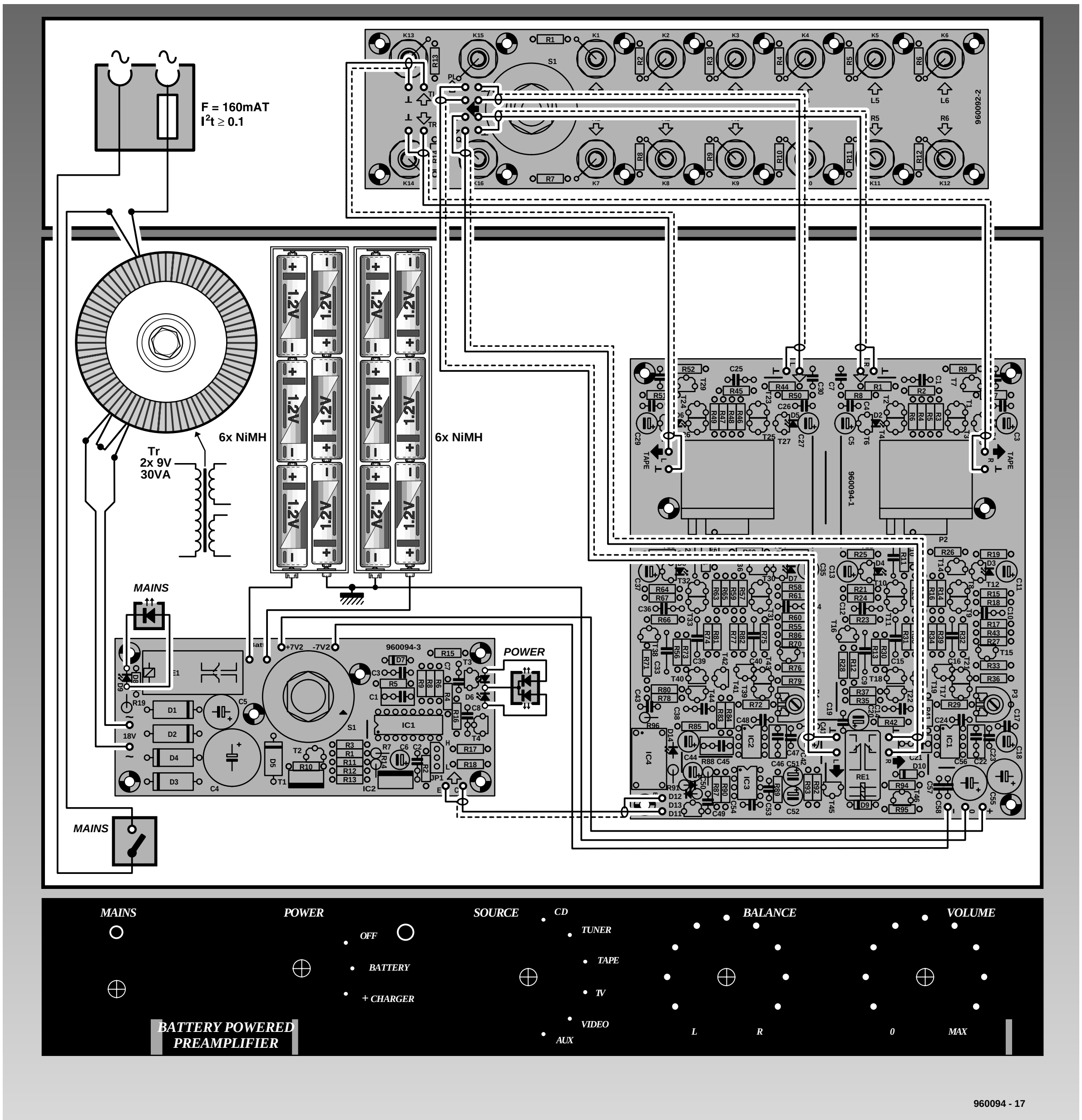


Figure 7. The wiring diagram shows that the link between the preamplifier common and case earth is connected to the junction of the 6th and 7th battery.

should be linked firmly with the aid of, for instance, a nylon cable strap pulled as tight as possible. Good thermal coupling is also necessary between D_1–D_8 and the associated transistors. So, use rectangular diodes and fit them tightly against the flat side of the adjacent transistor: cable straps are not necessary here.

There is ample room on the board for the balance control and volume control. This has the advantage that the distances between the various components are as short as feasible and that no cables are needed between the controls and the board. The spindles of the controls are fitted with special extensions to the front panels (as in the case of the input selector).

Both the balance control and the volume control are closed Alps types. In the case of the balance control, this is a special model with half-silvered tracks to prevent attenuation when the control is at the centre of its travel.

Finally, it is imperative that axial lead 160 V polystyrene capacitors be used in the C_8, C_9, C_{15}, and C_{16} (C_{32}, C_{33}, C_{39}, and C_{40}) positions. Do not use ceramic types, which are not suitable and do not fit on the board.

Completion of the charger board is, like that of the input selector board, a piece of cake. Again, the rotary switch is mounted directly on the board. Bear in mind that D_6 and D_9 have to be visible at the front panel.

Cooling of transistor T_1 is best effected by fitting the device isolated on a small sheet of aluminium to the bottom the enclosure.

Do not forget the jumper!

The completed prototype boards are shown in **Figure 6**.

CHECKING & TESTING

When the boards have been completed, it is good practice to carry out a thorough check of the work. Are all diodes and electrolytic capacitors fitted with correct polarity? Are all ICs fitted correctly? This kind of check may save a lot of time later on if there is a malfunction.

When the boards appear to be all right, connect the mains transformer in a provisional, but safe way to the ~ terminals on the charger board. Do not yet connect the batteries. Measure the potential across C_4, which should be about 24 V, and across C_5, which should be 1–2 V lower. Check that the output voltage of IC_2 is about 18 V and, if so, connect the batteries. The operation of D_6 shows whether the charger works correctly. If so, charge the batteries fully. In the unlikely case that the charger does not function correctly, check that the potentials at the test points shown in Figure 4 are as shown.

Next, link the batteries to the supply terminals on the preamplifier board. The subsequent faint lighting of diodes D_1–D_8 indicates that by and large the preamplifier works correctly. The quiescent current is set to 2 mA by

Table 1 Test voltages *(preamplifier - Figure 1)*	
measured across:	***Potential***
D_1–D_8	1V6
R_7–R_9, R_{19}, R_{25}, R_{26}, R_{50}, R_{51}, R_{52}, R_{62}, R_{68}, R_{69}	1 V
R_3, R_4, R_{46}, R_{47}	0V15
R_5, R_6, R_{48}, R_{49}	0V13
R_{14}, R_{16}, R_{20}, R_{22}, R_{57}, R_{59}, R_{63}, R_{65}	0V05
R_{15}, R_{17}, R_{21}, R_{23}, R_{58}, R_{60}, R_{64}, R_{66}	1V1
R_{27}, R_{28}, R_{70}, R_{71}	0V5
C_{14}, C_{38}	1V7
R_{32}, R_{34}, R_{75}, R_{77}	0V25
R_{33}, R_{35}, R_{76}, R_{78}	0V78
R_{36}, R_{37}, R_{79}, R_{80}	0V14 (set with P_3, P_4)
pins 4 and 6 of IC_3	0V4
D_{11}	1V23
R_{90}	1V72
D_9, D_{10}, R_{92}, R_{94}	0 V
measured between earth and:	***Potential***
base T_1; junction R_5-R_6; base T_8; junction R_{38}-R_{39}; junction R_{39}-R_{40}; pins 2 and 3 of IC_1	0 V
base T_{23}; junction R_{48}-R_{49}; base T_{30}; junction R_{81}-R_{82}; junction R_{82}-R_{83}; pins 2 and 3 of IC_2	0 V
pin 6 of IC_1; pin 6 of IC_2	0 V (if not, select devices)

connecting a multimeter (1 V d.c. range) across R_{36} or R_{37} (R_{79} or R_{80}) and adjusting P_3 (P_4) to obtain a meter reading of 140 mV.

For clarity's sake, the voltages at the various test points in Figure 1 have been omitted and are shown in **Table 1**. Note that all potentials were measured with a digital multimeter (high input impedance). If the measured voltage deviate no more than 10% from the specified ones, it may be assumed that all is well.

Be careful when using an oscilloscope for testing the preamplifier not to link the common lines of the charger board and the preamplifier board as this would short-circuit the negative supply line.

ASSEMBLY & WIRING

Any type of metal enclosure may be used as long as the boards fit into it easily. It is best to mount the charger board directly behind the front panel, the input selector board to the rear panel and the preamplifier board in between. The spindles of the various controls and switches must, of course, be provides with suitable extensions.

Mind that D_6 and D_9 can be seen at the front panel.

Fit a mains switch on the front panel and the mains entry with integral fuse at the rear. The fuse should be a slow type rated at 160 mA (I^2t>>0.1). Use well-insulated wire for the connections between mains entry, transformer and on/off switch.

With the boards positioned as indicated, the (toroidal) mains transformer and battery holders fit nicely in the space behind the charger board. Keep the transformer away as far as possible from the input selector board and as close as feasible to the mains entry.

Although the wiring has been kept to a minimum, a wiring diagram is shown in **Figure 7**. The links between the input selector board and the preamplifier board need to be single screened audio cable. This type of cable is also used for interlinking terminals C and E on the charger board and the preamplifier board.

The supply lines may be normal flexible stranded circuit wire. Note that the common of the preamplifier board functions as the earth: as the wiring diagram shows, it is connected directly to the junction of the 6th and 7th battery. This junction should also be strapped to the earthing point of the metal case. Do not connect the earth of the charger board to the case earth since that would short-circuit the supply line. It is for this reason also that T_1 must be isolated from the case.

Figure 8 (scale 8:10) shows a suggested front panel layout and marking for the case: this is not available ready made.

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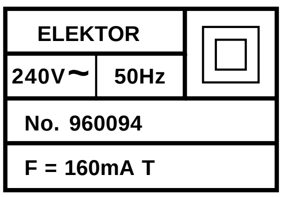


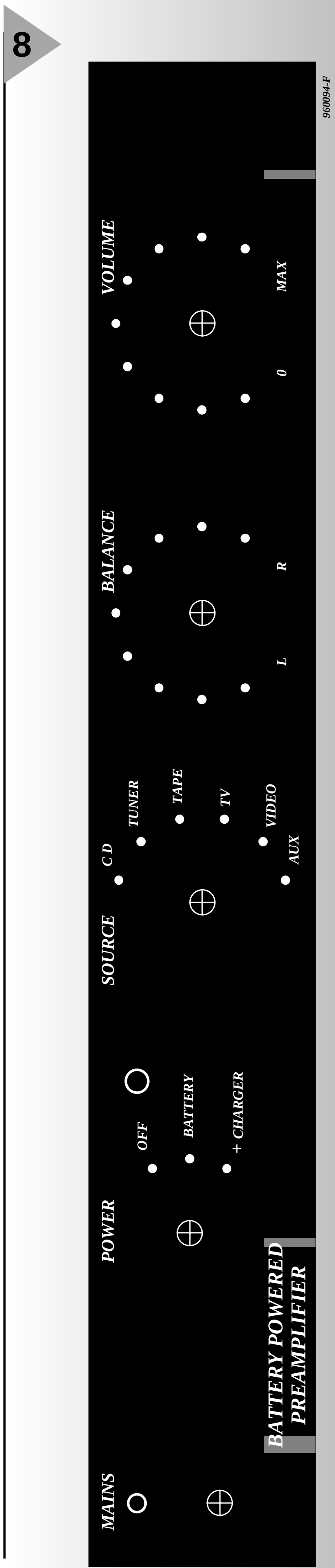
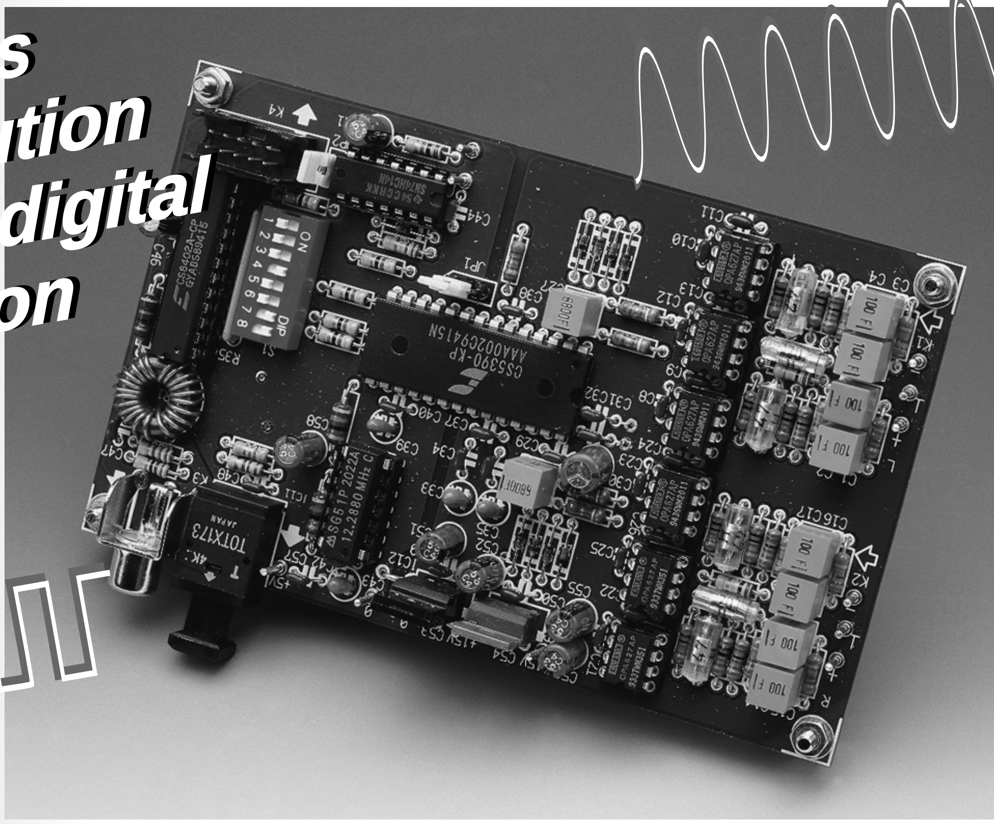


Figure 8. Suggested front panel layout and marking for use with the Monacor case.

20-bit A-D converter

provides high-resolution analogue-to-digital conversion



The converter described in this article is, technically speaking, state-of-the-art as far as analogue-to-digital conversion is concerned. It offers a true resolution of 20 bits, is of superb quality and has symmetrical inputs. In short: it is difficult to imagine a better converter for quality-conscious sound engineers than the present one – and certainly not at the price.

Design by T. Giesberts

The circuit is based on a Type CS5390 integrated stereo analogue-to-digital converter (ADC) from Crystal, which is currently taken as the quality standard and is used in much professional equipment. It uses ×64 oversampling, contains a phase-linear digital anti-aliasing filter, and has a dynamic range of 110 dB. It is pin-compatible with the 18-bit Type CS5389 IC.

The converter stage is preceded by an elaborate analogue input amplifier, whose outstanding property is its ability to accept symmetric as well as asymmetric signals. The quality of this amplifier can be improved even more by the use of rare (expensive) operational amplifiers.

The coding and sending of the converted audio data is effected with the aid of a Type CS8402A integrated digital audio interface. This IC is used in the same configuration as in the ‘sampling rate converter’ published in the October issue of this magazine.

The amplifier has an electrically isolated S/PDIF output, but there is also provision for an optical output.

SYMMETRIC INPUTS

Attaining a true resolution of 20 bits places a heavy demand on the circuit designer, whose ingenuity is taxed from the inputs onwards.

If optimum use is to be made of the symmetric inputs of the CS5390 (IC_7), the input amplifier and source signals must also be symmetric. Since, however, many users do not work with symmetric signals, the converter must also be capable of processing asymmetric signals.

Such considerations have led to the converter being given a fairly complex input circuit, whose special property is its capability of processing differential as well as single-ended signal without the need of switching. In the circuit diagram in **Figure 1**, the circuit for the left-hand channel consists of op amps IC_1–IC_3 and that for the right-hand channel of op amps IC_4–IC_6. Otherwise, the channels are identical.

The following discussion is based on the left-hand channel; components in the right-hand channel are stated in brackets.

Input socket K_1 is followed by attenuator R_1–R_4 (R_{12}–R_{15}), which is necessary for presetting the input stage, particularly in case of asymmetric signals. Op amps IC_1 and IC_2 (IC_4 and IC_5) are arranged as ×10 amplifiers

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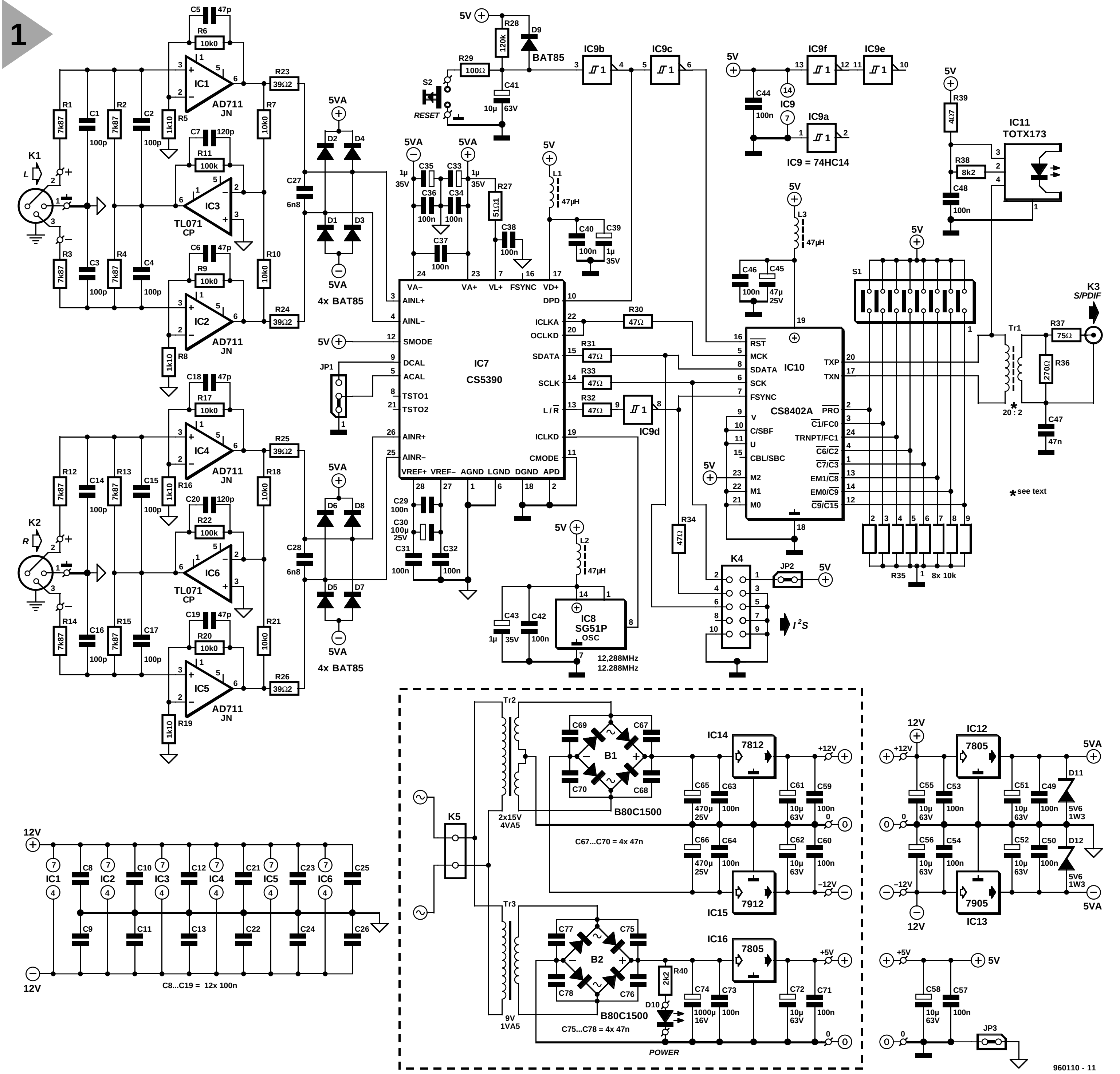


Figure 1. The circuit consists essentially of four sections: the analogue input amplifier, IC_1–IC_6; the ADC proper, IC_7; the output interface, IC_{10}; and the power supply, IC_{12}–IC_{16}.

with an input sensitivity of 1 V_{RMS}. When an asymmetric signal is input, the inverting (–) input of the op amps is grounded, which disturbs their setting. However, the consequent imbalance of the signals is corrected immediately by IC_3 (IC_6) with an accuracy of no less than ±0.08 dB.

The quality of the op amps used is, of course, of paramount importance: the Type AD711 from Analog Devices in the IC_1 and IC_2 (IC_4 and IC_5) positions was found to give very good performance. In the IC_3 (IC_6) position, a standard TL071 (which is far less expensive) was found to be perfectly satisfactory. In any case, this latter IC has no role to play in the case of symmetric signals, while with asymmetric signals it is just part of the feedback loop where it has far less effect on the signal quality than the earlier mentioned op amps.

If superlative performance is required, use Type OPA627 chips in the IC_1–IC_3 (IC_4–IC_6) positions. Bear in mind, however, that these ICs are about ten times as costly as an AD711.

CONVERSION

Full data of the Type CS5390 A-D converter (IC_7), as well as its internal block diagram, are given in the data sheets elsewhere in this issue.

Schottky diodes D_1–D_8 and zener diodes D_{11} and D_{12} protect the IC against excessively high input signals. The Schottky diodes protect against

test results

The prototype was, of course, tested thoroughly, including FFT analyses of the output data of the converter chip for which the total harmonic distortion (THD+N) at 1 kHz was calculated in the digital domain. This gave the following results:

Input level	*THD+N*
0 dB	*<–97 dB*
– 3 dB	*<–100 dB*
– 6 dB	*<–99 dB*
– 9 dB	*<–96 dB*
– 12 dB	*<–93 dB*
– 15 dB	*<–90 dB*

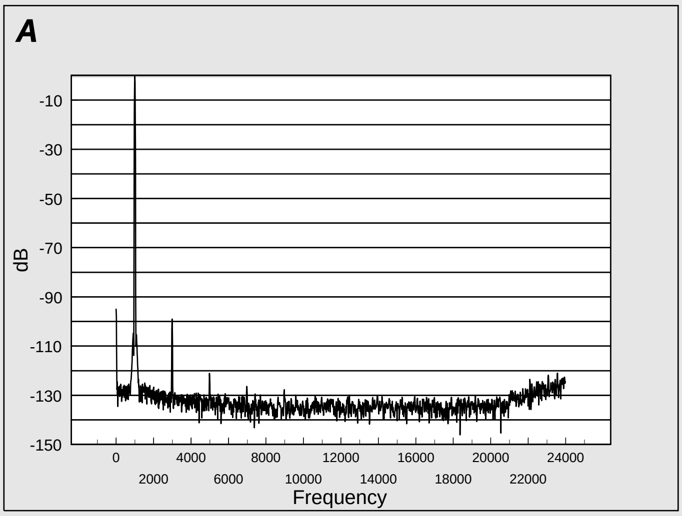
The same calculation was carried out at 7 kHz, since the 3rd harmonic of this falls just within the range of the digital filter. The differences with the results obtained at 1 kHz were negligible.

As the test results show, the manufacturers' stated dynamic range of 110 dB was, in practice, just not attainable. This is partly because of the ×10 amplification in the input amplifiers and the consequent higher sensitivity. When the ratio of the input level (in dB) is added to the THD+N value (in dB), the noise level is effectively 105 dB below full scale. This changes only when the input signals exceed –6 dB. However, from that point on, there is a noticeable increase in distortion. Note that distortion here means deviations of the order of not more than 5×10–6!

In any case, since in digital recording the normal headroom is 12 dB, it may be assumed that for average signal levels the distortion produced by the A-D converter is negligible.

The frequency spectrum of two FFT analyses is shown. Figure A is measured with a signal of 0 dB and Figure B with one of –12 dB. To recover the 7th and 9th harmonics from the noise, four measurements were carried out in each case. At an input level of –12 dB, there was no trace of any harmonics.

To satisfy the curiosity of some readers, the frequency spectrum with shorted input was measured: the result is shown in Figure C. This shows a virtually Gaussian distribution without any remains or effects of clock interference, supply ripple or other possible sources of interference.



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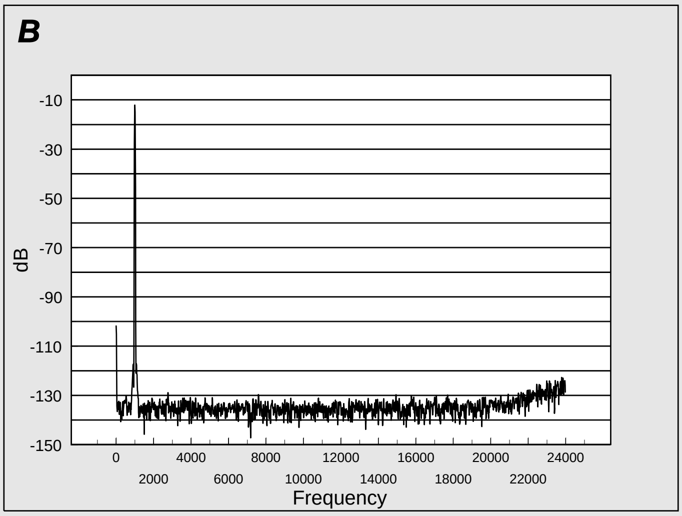

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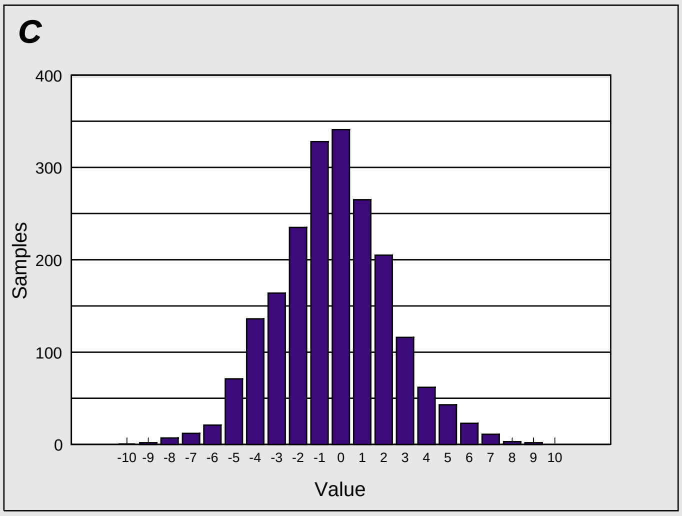

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latch-ups, while the zener diodes limit an increase in the supply voltage in case of overdrive. The latter is vital, because the maximum permissible supply voltage for the CS5390 is ±6 V.

The value of series resistors R_{23} and R_{24} (R_{25} and R_{26}) ensures optimum source impedance for IC_7.

Parallel capacitor C_{27} (C_{28}) must be a close-tolerance type, since it suppresses r.f. noise caused by the oversampling. It is, as it were, complementary to the digital filter, since the suppression range of this does not extend into the r.f. bands.

As far as the operation of IC_7 is concerned, it is best to treat it as a black box, into which analogue signals are pumped and from which digital signals are extracted. According to the manufacturers' data, the chip uses delta-sigma modulators with ×64 oversampling for the conversions. The modulators are followed by a digital filter, whereupon the sampling frequency is reduced to 48 kHz. Because of the very high oversampling rate, a separate anti-aliasing filter is not required.

Most of the remaining circuitry, except the reset circuit connected to the digital power down (DPD) input, and jumper JP_1, serves to decouple the power supply and reference voltage lines.

When switch S_2 is pressed, network R_{28}-C_{41} and Schmitt trigger IC_{9b} provide a 1 s long positive pulse at the DPD input. This results in an overall reset of the entire digital section of IC_7. After the pulse has decayed, an offset calibration cycle is started automatically. During this cycle, the offset in either channel is measured and deducted from the sampling values. It is thus possible to start this calibration at any desired moment, without having to switch the entire converter off and then on again, simply by pressing S_2. In other words, this design makes possible a conversion without having to worry about any offset voltages, which for certain applications is of vital importance. It should be borne in mind that each and every cold start results in a (small) drift of any offset voltage in both the converter chip and the input amplifiers.

The level at the ACAL input determines whether the offset is measured in the input amplifiers or not. If this input is grounded via JP_1, the offset in the input amplifiers will be measured. In this case, there must be no input signal to the amplifiers, since that may lead to measurement errors. If ACAL is linked to DCAL via JP_1, the offset in the input amplifiers is ignored. Note that the output of DCAL remains high for 4096 clock pulses after DPD has gone low.

By means of the logic level at CMODE, the serial output interface of IC_7 enables the value of the requisite ICLKD clock, which has been fixed for an output word rate – OWR – of 48 kHz, to be determined. In the present design, CMODE is permanently low, which results in a clock frequency of 256 × OWR = 12.288 MHz.

The level at SMODE determines whether the IC operates in a slave mode or a master mode. A high level at this terminal results in SCLK, FSYNC, and L/R functioning as outputs whose level is derived from ICLKD by internal dividers.

The 20-bit samples in 2-complement code are available at output SDATA.

The clock generator is a module containing both the crystal and the requisite active circuitry. This arrangement saves space and also minimizes the risk of r.f. interference.

The converted data at the output of IC_7 are applied to output socket K_4 via stoppers R_{30}–R_{34} and the output circuit. The timing of the data is nearly, but not quite, I^2S compatible: it lacks an inverted L/R signal, but this is provided by IC_{9d}.

OUTPUT

The general I^2S output is taken from K_4. A miscellany of signal-processing equipment (for volume control, tone control, interface for test purposes, and so on) may be connected to this socket. The output is also well suited for receiving the 'digital VU meter' (a special test instrument for digital audio signals, fitted with a double 30-segment LED bar and 3.5-digit displays) described in the April/May 1996 issues of this magazine.

Short-circuiting JP_2 links the +5 V supply line to K_4, which is usable if the equipment connected to the socket draws a current of not more than a few milliamperes. In all other cases, the equipment must have its own power supply.

There is, of course, an S/PDIF output. For this, use is made of the same digital audio interface transmitter (IC_{10}) as used in the 'sampling rate converter' described in the October 1996 issue of this magazine. The most significant channel-status bits of this IC are set with octal DIP switch S_1.

Apart from coaxial output K_3, an optical output is provided by IC_{11}.

POWER SUPPLY

Since the power lines to the digital and analogue sections of the circuit must be isolated, two separate mains transformers are used: Tr_2 for the analogue section, and Tr_3 for the digital section.

The output of Tr_2 is rectified by B_1, smoothed by a number of electrolytic capacitors, and regulated by IC_{14} and IC_{15}. The resulting ±12 V output is used to power the input amplifiers, and also to provide the ±5 V lines for the analogue circuits in IC_7. The ±5 V lines are regulated by IC_{12} and IC_{13}.

The output of Tr_3 is rectified by B_2, smoothed by several electrolytic capacitors, and regulated by IC_{16}, resulting in a single-ended supply line of +5 V.

The only link between the two supply lines is JP_3, which commons their earth returns.

The supply lines are copiously decoupled for r.f. Also, the arms of the bridge rectifiers are shunted by antirattle capacitors, while the electrolytic buffer capacitors are without exception shunted by 100 nF ceramic capacitors.

PRINTED-CIRCUIT BOARD

The converter is best built on the printed-circuit board shown in **Figure 2**. This board is through-plated an double-sided and has, at the component side, two large earth planes, one each for the analogue and digital circuits.

The board consists of two distinct parts: that for the power supply (shown in dashed lines in Figure 1) must be cut off and built up independently. Not all mains transformers may fit neatly on the supply board, which must then be adapted accordingly. The ±12 V and +5 V lines must be linked to the corresponding terminals on the converter board via flexible circuit wire. None of the voltage regulators needs a heat sink.

Populating the converter board is fairly straightforward but, in a quality unit such as the converter, the utmost care must, of course, be observed in the work. Whether or not IC sockets should be used is immaterial: from a technical point of view it is better to solder the ICs directly to the board, but some less-experienced constructors may prefer to use sockets.

All constructors are advised not to scrimp on the components. A fair number of 1% resistors, as well as quite a few close-tolerance polystyrene (not cheap) capacitors, are used. Furthermore, high-quality types of socket must be used for K_1–K_3.

Output transformer Tr_1 is a DIY component, wound on a toroidal core Type G2/3FT12. Wind the primary, consisting of 20 turns of 0.7 mm dia. enamelled copper wire, evenly spread over the core, first. Leave some space at the centre of the winding, where the two-turn secondary (same wire) is laid. This is exactly the same transformer as used in the 'sampling rate converter' published in the October 1996 issue of this magazine.

Using heavy-duty wire, short-circuit

2

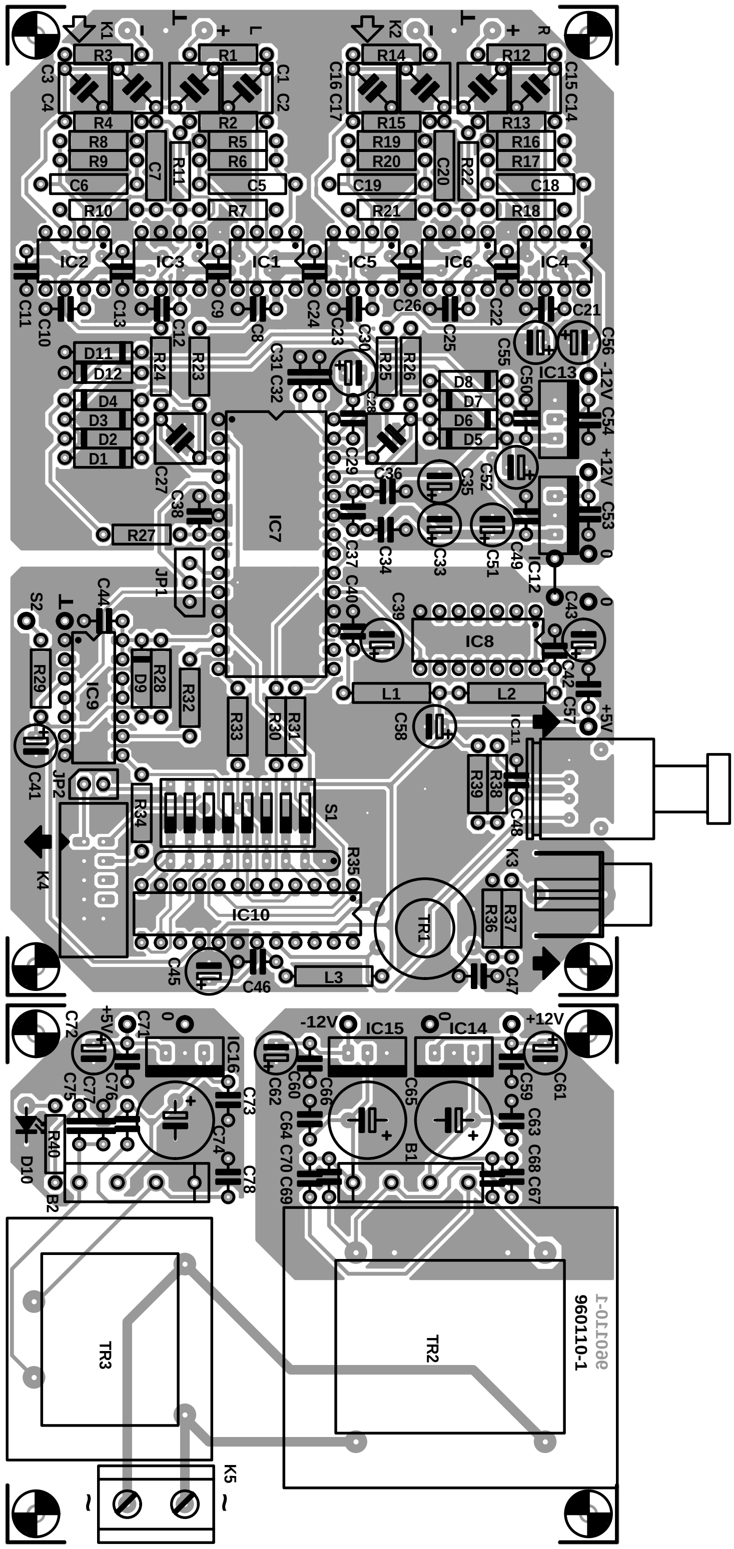


Figure 2. The double-sided, through-plated printed-circuit board consists of two sections: one for the power supply and the other for the digital and analogue circuits. The board is intended to be cut into two.

the two terminals of wire bridge JP_3.

With JP_1 in the right-hand position (pin 5 of IC_7 to ground), the offset of the input amplifiers is measured during the automatic calibration; with JP_1 in the left-hand position (pin 5 linked to pin 9), this is not done.

If the equipment connected to K_4 can be powered by the converter (current drawn by it no more than a few mA), the +5 V line is linked to K_4 via wire bridge JP_2.

When both boards have been completed, a thorough check (consulting the circuit as well as the parts list) is recommended. Also, compare the boards with the photograph of the completed prototype in **Figure 3**.

INITIAL TEST

Connect 2.2 kΩ resistors to each of the outputs of the supply board.

Connect the mains to terminal block K_5 via a suitable mains cable, whereupon on/off indicator D_{10} should light.

Using a good multimeter, check the output voltages of the power supply. If all potentials are what they should be, the 2.2 kΩ resistors must be removed from the supply board outputs, which should then be linked to the relevant terminals on the converter board.

ENCLOSURE

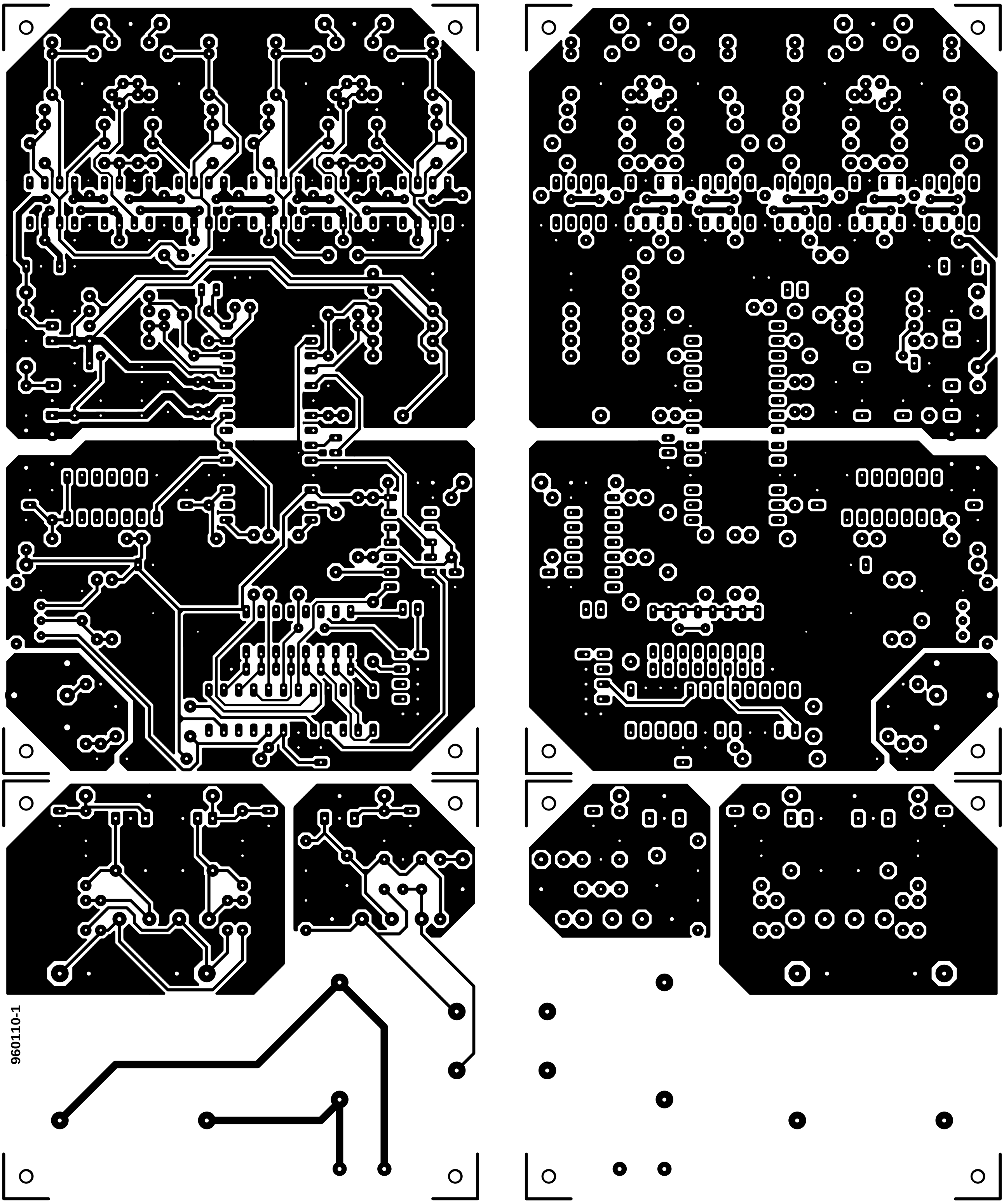
When all has been found in order, the converter should be fitted in a suitable enclosure. The only restrictions on the choice of this are that it must be of metal and large enough to house the converter and its power supplies.

Parts list

Resistors:
R_1–R_4, R_{12}–R_{15} = 7.87 kΩ, 1%
R_5, R_8, R_{16}, R_{19} = 1.10 kΩ, 1%
R_6, R_7, R_9, R_{10}, R_{17}, R_{18}, R_{20}, R_{21} = 10.0 kΩ, 1%
R_{11}, R_{22} = 100 kΩ
R_{23}–R_{26} = 39.2 Ω, 1%
R_{27} = 51.1 Ω
R_{28} = 120 kΩ
R_{29} = 100 Ω
R_{30}–R_{34} = 47 Ω
R_{35} = 10 kΩ octal array
R_{36} = 270 Ω
R_{37} = 75 Ω
R_{38} = 8.2 kΩ
R_{39} = 4.7 Ω
R_{40} = 2.2 kΩ

nductors:
L_1–L_3 = 47 µH

Capacitors:
C_1–C_4, C_{14}–C_{17} = 100 pF, 63 V, polystyrene, radial, pitch 7.5 mm
C_5, C_6, C_{18}, C_{19} = 47 pF, 160 V, polystyrene
C_7, C_{20} = 120 pF, 160 V, polystyrene
C_8–C_{13}, C_{21}–C_{26}, C_{29}, C_{31}, C_{32}, C_{34}, C_{36}–C_{38}, C_{40}, C_{452}, C_{44}, C_{46}, C_{48}–C_{50}, C_{53}, C_{54}, C_{57}, C_{59}, C_{60}, C_{63}, C_{64}, C_{71}, C_{73} = 100 nF, ceramic
C_{27}, C_{28} = 6.8 nF, 63 V, 1%, polystyrene, radial, pitch 7.5 mm
C_{30} = 100 µF, 25 V, radial
C_{33}, C_{35}, C_{39}, C_{43} = 1 µF, 35 V, tantalum
C_{41}, C_{51}, C_{52}, C_{55}, C_{56}, C_{58}, C_{61}, C_{62}, C_{72} = 10 µF, 63 V, radial
C_{45} = 47 µF, 25 V, radial
C_{47}, C_{67}–C_{70}, C_{75}–C_{78} = 47 nF, ceramic
C_{65}, C_{66} = 470 µF, 25 V, radial



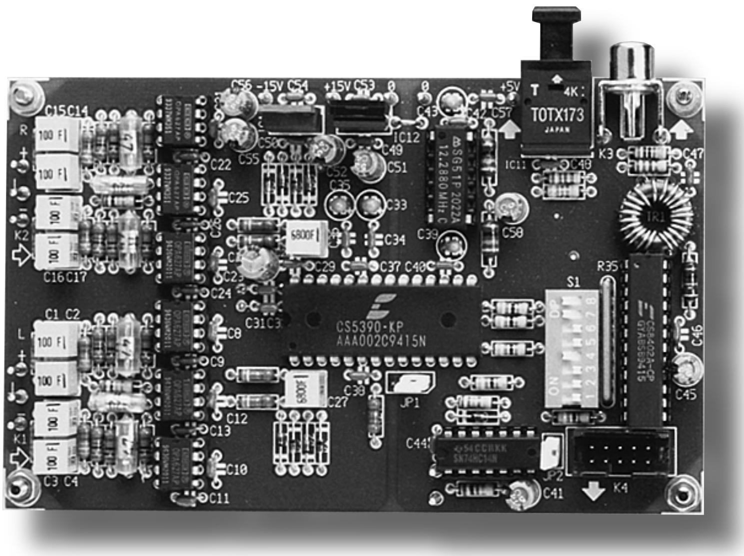
C_{74} = 1000 µF, 16 V, radial

ISemiconductors:
D_1–D_9 = BAT85
D_{10} = LED, low current
D_{11}, D_{12} = zener diode 5.6 V, 1.3 W
B_1, B_2 = B80C1500 right-angled bridge rectifier

Integrated Circuits:
IC_1, IC_2, IC_4, IC_5 = AD711JN (Analog Devices)
IC_3, IC_6 = TL071CP (see text)
IC_7 = CS5390-KP (Crystal)
IC_8 = oscillator module, 12.288 MHz (Seiko Epson Type SG51P)
IC_9 = 74HC14
IC_{10} = CS8402A (Crystal)
IC_{11} = TOTX173 (Toshiba)
IC_{12}, IC_{16} = 7805
IC_{13} = 7905
IC_{14} = 7812
IC_{15} = 7912

Miscellaneous:
JP_1 = 3-way pin header and jumper
JP_2 = 2-way pin header and jumper
JP_3 = wire bridge
K_1, K_2, K_3 = phono socket for board mounting
K_4 = 10-way box header
K_5 = 3-way terminal block (for accepting 3-wire mains cable), pitch 7.5 mm
S_1 = octal DIP switch
S_2 = single-pole push-button switch
Tr_1 = see text
Tr_2 = mains transformer 2×15 V, 4.5 VA
Tr_3 = mains transformer 1×9 V, 1.5 VA
PCB, Order no. 960110

3



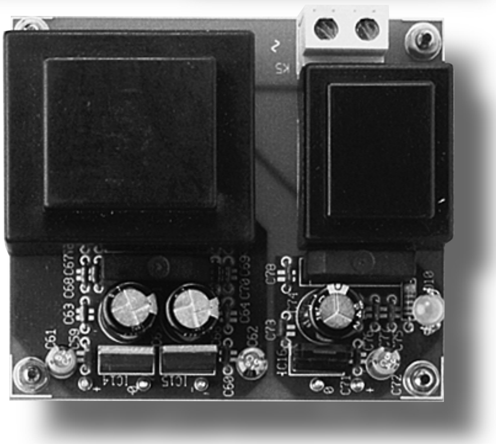


Figure 3. The completed prototype boards. Although the photograph is small, it may be seen that Type OPA627 circuits are used in the IC_1–IC_6 positions.

Wiring up the unit is simple, but the best way is to fix the converter in such a way that output sockets K_3 and IC_{11} protrude through the rear panel of the case.

Next, fit sockets K_1 and K_2 and link these with screened cable to the input terminals of the converter board.

Fit a mains entry, preferably with integral on/off switch, to the rear panel of the case and link this via a length of suitable mains cable to K_5. If the on/off switch is preferred on the front panel, insert it in series with the mains entry and K_5.

The only other operating controls on the front panel are on/off indicator D_{10} and reset push-button switch S_2.

To obtain maximum benefit of the screening of the metal enclosure, it should be linked at a single point to the ground of the converter and power supplies and to the mains earth. This is best done with a small bolt, nut, washer and solder tags fitted to a suitably drilled hole in the bottom of the enclosure. Wire bridge JP_3 may also be used for commoning the earths.

Finally, consult the Safety Guidelines elsewhere in this issue.

APPLICATIONS

The converter is suitable for use in a wide range of applications in which analogue-to-digital conversion of the highest quality is needed. One possible application is its use as an upgrade for a DAT recorder (no longer in production), which is straightforward thanks to its symmetric inputs. Where the necessary equipment exists for mixing at digital level, several converters may be used to make a master recording with only one DAT recorder.

Another use is in combination with the 'sampling rate converter' mentioned before. Such a combination would make it possible, for instance, to make analogue recordings with a sampling rate adapted to the CD standard. In principle, that may also be done without a sampling rate converter, but then the clock of IC_7 must be altered to 11.2896 MHz: this would give a sampling rate of 44.1 kHz. However, the published 'sampling rate converter' offers the possibility of converting the 20-bit data into a 16-bit format, whereby, through psychoacoustic noise shaping, a resolution of 18 bits is attained.

[960110]

In passing ...

What progress?

After much discussion, we had finally convinced our managing director that we urgently needed a new, faster computer for communicating over the Internet. The 386 unit to be replaced was, in spite of the Windows 95, a lame duck.

When the new 150 MHz pentium computer finally arrived, we coud not wait to get going on it. Alas, our joy did not last very long. After all connections were made and the mains switched on, all that worked was the internal fan. The computer itself remained dead. Now what? Of course, there is such a thing as a guarantee, but as engineers ourselves, we wanted to find out what was wrong. So, the case was opened and what did we find? The VGA card was suspended somewhere above the PCI connector it should mate with. A further check revealed that the cover of the card was not in the correct position, so that the card could not be pushed far enough into the PCI connector. When all this was righted, hurray! The PC worked.

The supplier had installed Windows 95 on to the hard disc, but he had probably never heard of cluster sizes. The entire 1.6 GB hard disc consisted of one partition! So, we arranged the disc in three partitions and reinstalled the software. After Windows had been reinstalled, the drivers for the VGA card also had to be reinstalled. Not so difficult, you might think, since there was an installation for Windows 95. This worked all right, but the utility for setting the image frequency (which was there originally) could not be found. But we had selected the correct video processor according to the manual and the floppy. So, we installed the DOS utilities, which, strangely enough, were on another floppy that was, according to the instructions, intended for OS/2. The utitlity we wanted could not be found. By now very suspicious, we looked again and found the Windows drivers for this card on a CD-ROM containing all sorts of demo for the VGA card. When these were tried, they proved to be for a different video processor than the present one. Oh, well, you can only try! And, lo and behold, they were the correct drivers with the associated Windows utilities. If you can understand all this, we cannot.

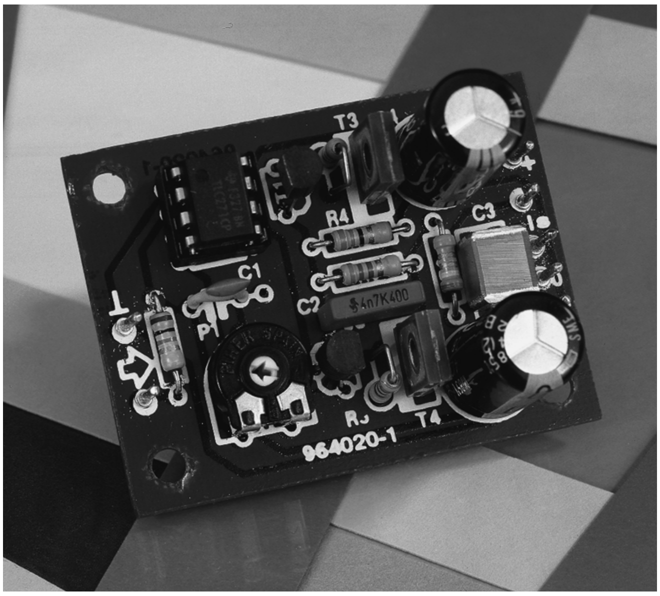
Browsing through the said CD-ROM, we had seen MPEG-player software, and this we had to try, of course. This did not prove to be a success, either. The program worked all right, but every time the film was run, the start menu for Windows disappeared after a second or so, so that there was no time to start the program. By then it was time to go home.

We shall sleep over it. Tomorrow, with fresh suspicions, we shall try the sound card.

Harry Baggen

Contents

miniature power amplifier



There are quite a few applications of an audio power amplifier in which power and hi-fi characteristics are of secondary importance. If, for instance, an active loudspeaker for a portable radio receiver is needed, compact dimensions and low current drain are far more important considerations.

These properties are the prime design basis for the present mini amplifier. It continues working satisfactorily with a battery voltage down to 1.5 V. Its quiescent current drain is about 1 mA, and its efficiency is a worthwhile 70 per cent. It provides an output power of 500 mW into 8 Ω (or 800 mW into 4 Ω), has a sensitivity of 400 mV, and its distortion is never higher than 1.2 per cent.

The low current drain is obtained through a combination of a low-power op amp followed by a discrete Class-B stage. The op amp is a Type TLC271 operating in its high-current mode (pin 8 to ground). Any complications arising from the common-mode range are prevented by using the amplifier as an inverting type. The voltage amplification is set with feedback resistance P_1.

The discrete power stage consists of two complementary darlingtons, each composed of a BC and a BD transistor. Resistors R_2–R_5 limit the internal amplification. Capacitors C_1, C_2, and C_6, are compensation devices.

Boucherot network R_6-C_3 ensures amplifier stability when the load is very low or very high.

Since the output transistors have no emitter resistor, the voltage is determined solely by the knee voltage of T_3 and T_4. With a load of 4–8 Ω, these voltages are limited to 0.2–0.3 V, so that the transistors can be driven virtually up to the supply voltage. This is the reason for the atypical high efficiency of the amplifier.

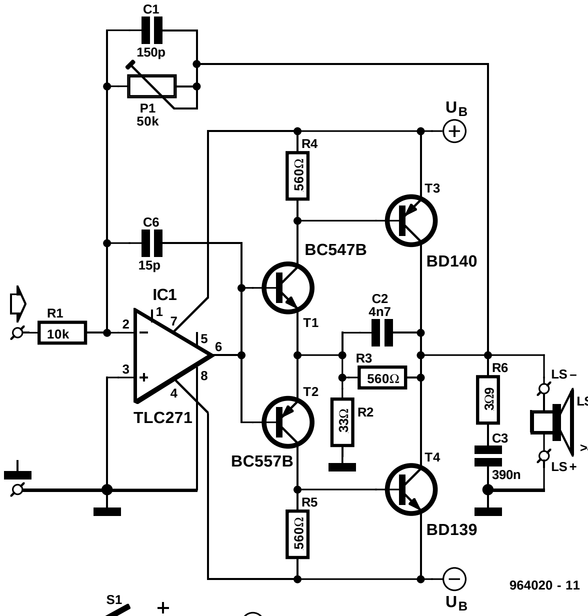
The overall bandwidth of the amplifier is limited to not less than 21 kHz at the maximum amplification of ×5.

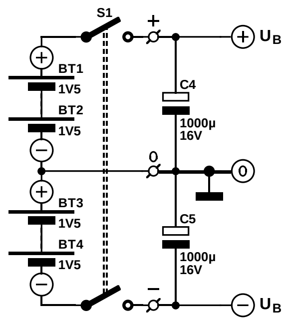
With a 4 Ω load, the peak output current is 700 mA. A 315 mA fuse in series with the output is, therefore, a simple, but effective short-circuit protection. At maximum drive with a music signal, the average current is only 50 mA. In practice the drive will never be continuously maximum, so that the actual current drain will be much lower. A set of four penlight batteries should last about 200 hours.

The amplifier is best built on the printed-circuit shown, which, unfortunately, is not available ready made.

On a final note: since the four batteries form a symmetrical supply, on-off switch S_1 needs to be a double-pole type.

[T. Giesberts - 964020]





PARTS LIST

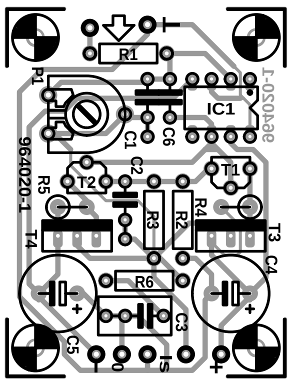
Resistors:
R_1 = 10 kΩ
R_2 = 33 Ω
R_3–R_5 = 560 Ω
R_6 = 3.9 Ω
P_1 = 47 kΩ preset

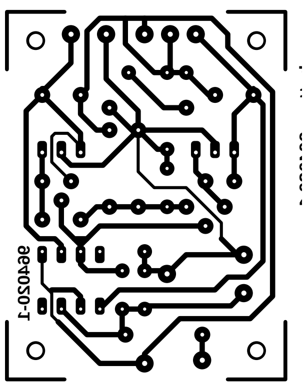
Capacitors:
C_1 = 150 pF
C_2 = 4.7 nF
C_3 = 390 nF
C_4, C_5 = 1000 µF, 16 V, radial
C_6 = 15 pF

Semiconductors:
T_1 = BC547B
T_2 = BC557B
T_3 = BD140
T_4 = BD139

Integrated circuit:
IC_1 = TLC271CP

Miscellaneous:
S_1 = double-pole on/off switch
Bt_1–Bt_4 = battery, 1.5 V





silent volume control

There is a growing tendency to fit audio equipment with electronic volume controls (virtually always remote controlled). This consists of a series of potential dividers actuated by electronic switches. Regrettably, this setup has a small drawback: faint click are sometimes heard when the volume is changed. These clicks are the result of the brief short-circuit of the multiplexers in 4000 series CMOS ICs to the negative supply line when they are switched over.

In the quad audio switch Type SSM2404 from Analog Devices this flaw is obviated by break-before-make logic drive circuits and by switching the output transistors gradually via a sawtooth generator. This switch has further benefits in that R_{DSon} is low and that it can operate from a high supply voltage (up to 13.5 V).

Since break-before-make electronic switches are used, it is not possible to construct a volume control from a potential divider with several branches followed by a buffer stage. This is because the switches in series with the buffer would briefly not have a bias setting when switching takes place. Therefore, the present design uses an inverting amplifier with a virtual ground. The total amplification is –1, but that may be changed by altering the value of R_{17}.

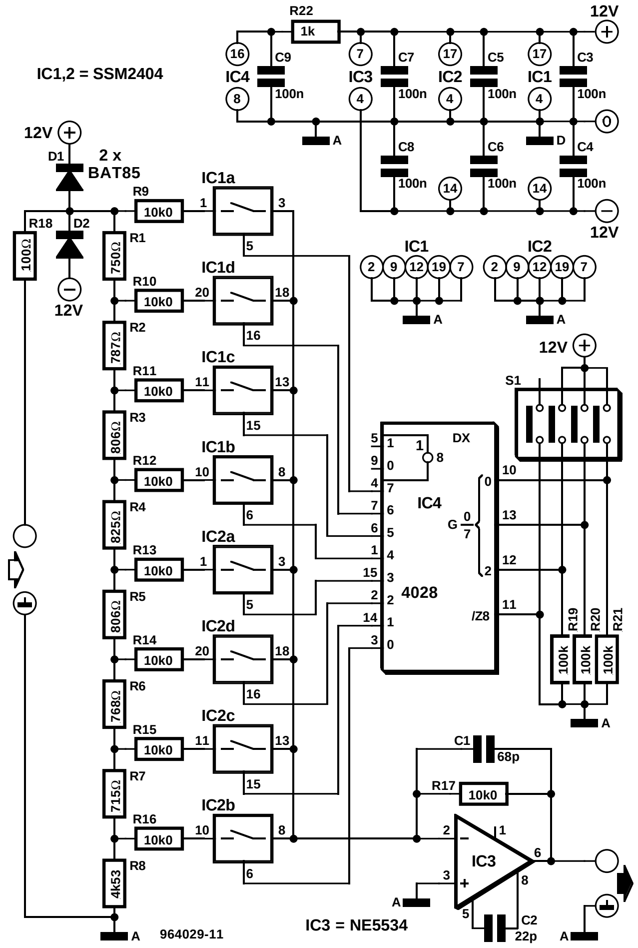
The voltage divider is designed for steps of 1.25 dB, so that an attenuation of 0–8.75 dB is possible. Resistor values for a total range of 70 dB are given in the table. Connecting two of these circuits (one with a range of 8.75 dB, and the other with a range of 70 dB) in series gives a volume control with a range of 0 dB to –78.5 dB. The volume control may be driven by a 6-bit counter or a microprocessor.

In the diagram, the switches of one chain of attenuators are driven by a BCD-to-decimal decoder Type 4028. The attenuation may be set manually with DIP switch S_1. Resistor R_{18} and diodes D_1, D_2 serve as overdrive protection for IC_1 and IC_2. These components may be omitted between two successive attenuator sections.

As far as the d.c. setting of the circuit is concerned, it suffices to state that the level at all points in the output condition (all DIP switches open) is 0 V, except at pin 6 of IC_2, which is at +12 V.

The maximum undistorted input voltage is 7.6 V r.m.s. In the prototype, the THD+noise with an input voltage of 2 V r.m.s. was 0.0007% at 1 kHz and 0.0009% at 20 kHz. At the largest attenuation (–8.75 dB) this figures rose to 0.001% and 0.0016% respectively. The overall bandwidth of the amplifier is 200 kHz. The current drain of a complete attenuator section is about 6 mA.

[T. Giesberts - 964029]



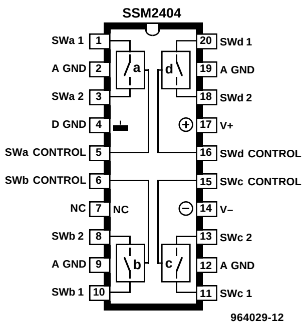


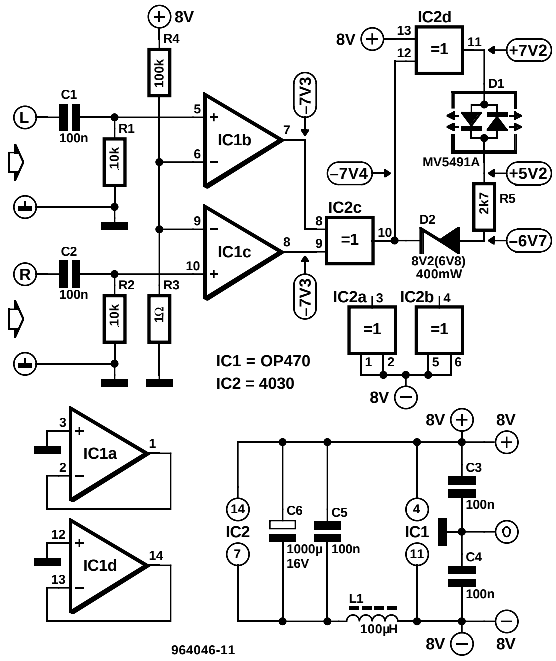
Table 1. Resistor values for a range of 0 dB to –70 dB.	
R1	6.04 kΩ
R2	2.80 Ω
R3	768 Ω
R4	226 Ω
R5	69.8 Ω
R6	21.5 Ω
R7	6.81 Ω
R8	3.16 Ω

surround sound indicator

The proposed circuit indicates with the aid of two LEDs whether or not the input signal contains surround data. The criterion for this is the phase difference between the two channels: if this is zero, there is no surround data.

In the circuit diagram, if there is a phase difference between the two channels, the output levels of comparators IC_{1b} and IC_{1c} will differ. These outputs are constantly compared by XOR gate IC_{2c}, and, in case of a difference, the output of the gate will go high. Depending on the output state, the red or green half of D_1 will be actuated via gate IC_{2d}, which is here connected as an inverter. In case of a pure surround signal, the red half will light brightly; in case of a mono signal, the green half will. If the input is a standard stereo signal, the rapid changes in the output of IC_{2c} will cause the diode to appear yellow-orange.

The circuit is an improved version of the design published in our January 1995 issue. There are two worthwhile improvements. The first is that the comparators are now Type OP470. To make sure that the comparators react satisfactorily with small input signals, the offset in the older version had to be greater than 15 mV. With the OP470, an offset of a few mV is sufficient, so that R_3 is now only 1 Ω. The value of this resistor is determined by the requirement that the outputs of the comparators must be low (≈ –7.3 V) in the absence of a signal. If this does not happen with 1 Ω, the value of R_3 may be increased by a couple of E12 values.



The second improvement is the addition of D_2. This is because brightness of the red and green halves of the LED used here differ from one another when the currents through them are equal. This results in the stereo indication being far too red. The additional diode acts as a zener for the red half and as a normal diode for the green half. The best rating for it appears to be 6.8 V or 8.2 V.

The circuit draws a current of about 15 mA.

Inductor L_1 and capacitors C_5 and C_6 have been added to prevent IC_2 affecting the operation of the comparators via the supply lines.

A final note. Internally, there is a diode limiter between the inputs of the comparators which clips input signals above about 1 V. If, therefore, input signals higher than, say, 700 mV are expected, it is advisable to connect a resistor of a few kilo-ohms in series with C_1 and C_2.

[T. Giesberts - 964046]

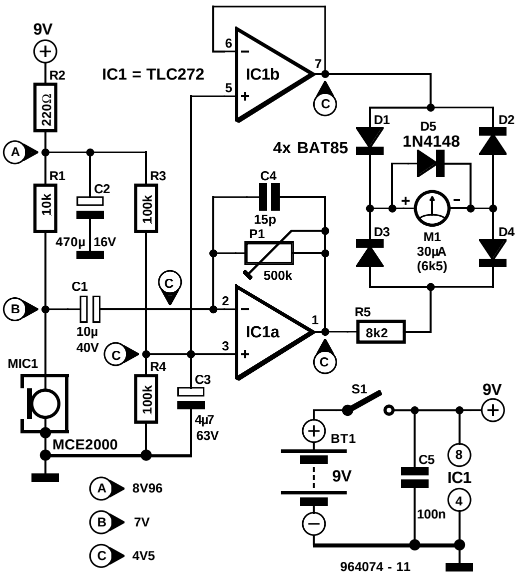
sound pressure meter

The design, consisting of an electret microphone, amplifier, and moving-coil meter, arranges for the meter to give a reading that is linearly proportional to the sound pressure. Sound pressure meters usually have a logarithmic scale, but this would make the design more complicated. Moreover, linear proportionality results in greater sensitivity to sound pressure differences. If a test CD with noise bands in 60-Hz steps is available, the meter may be used to obtain a frequency characteristic of a loudspeaker.

The microphone is a Type MCE2000 electret from Monacor. The operating point of the FET contained in this is set with R_1. Resistor R_2 and capacitor C_2 prevent noise on the supply lines from reaching the input.

The operating point of op amp IC_{1a} is set to half the supply voltage with R_3 and R_4.

The degree of amplification is determined by the ratio of P_1 to the output impedance of the microphone. With component values as specified in the circuit diagram, the amplification is ×60, which makes a 90 dB sound pressure result in full-scale deflection of the meter. In the prototype, P_1 is then roughly at the centre of its travel.



Capacitor C_1 is a bipolar electrolytic type, since the spread of the FET in the microphone may cause the voltage at metering point B to be lower than the 4.5 V at the input of the op amp. Note, by the way, that the voltages shown in the circuit diagram are taken from the prototype.

Half the supply voltage is buffered by IC_{1b}, so that the circuit can be driven by IC_{1a} without any difficulty. The buffering makes an additional potential divider in the metering circuit superfluous; this is a benefit, because such a divider affects the meter deflection and increases the current drain.

The meter, M, is a 30 µA moving-coil type with an internal resistance of 6.5 kΩ. The use of a meter with as low a current drain as possible keeps the voltage drop across bridge rectifier D_1–D_4 low. Resistor R_5 and diode D_5 limit the peak current through the meter.

The current drain of the complete circuit is only 1.5 mA.

The use of a wobbulator and the present meter enables, say, a subwoofer to be matched to an existing system. First measure the standard speakers at a frequency of 400–500 Hz, wobbulated over ⅓ octave. The level of the signal must be high enough to suppress ambient noises. Adjust P_1 for maximum meter reading. The, apply a 40–50 Hz signal to the subwoofer and adjust the active filter in the amplifier or subwoofer for an identical meter reading. These measurements should be taken at a distance of about 1 m (3 ft) from the loudspeakers.

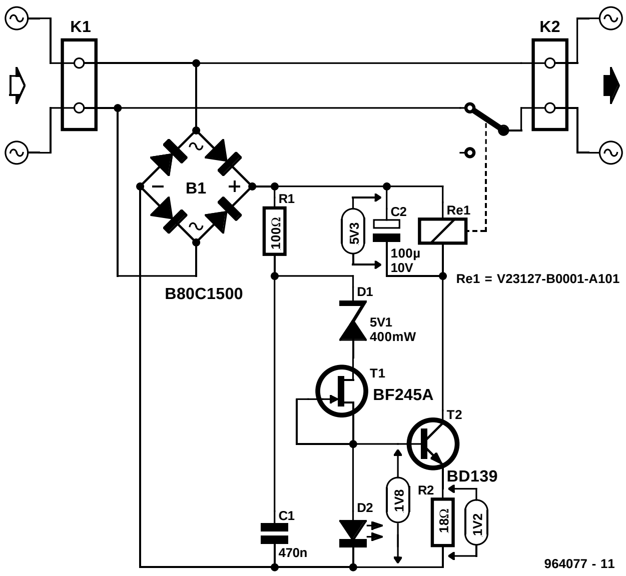
[T. Giesberts – 964074]

loudspeaker protector

This circuit automatically disconnects a mid-range loudspeaker or a tweeter from the output of an amplifier when too much power is applied. A relay with a normally closed (n/c) contact is used as the switching device.

The relay coil voltage is derived from the loudspeaker signal with the aid of bridge rectifier, B1. A current source, T2, is used to ensure a reasonably constant coil current. To ensure that the maximum coil current does not depend too much on level of the loudspeaker signal, the voltage across R2 is limited to about 1.2 V by an LED, D2. The maximum energizing current is then about 66 mA. The 6-V Siemens relay used here has a coil resistance of about 80 Ω. To keep the extra load on the amplifier to a minimum, the relay receives a coil voltage which is a little below the nominal value. Remember, the circuit draws current even when the relay is not energized! Fortunately, the current consumption is negligible at signal levels up to 5 V_{peak}.

The saturation current which flows through the J-FET current source, T1, is about 4 mA, enabling a low-current LED to act as a visual 'overload' indicator. Turn the volume down when the LED lights! A zener diode is connected in series with the current source and the LED to define the threshold at which the relay toggles. The zener value shown here results in an actuation level of about 7.5 V, or about 8.5 W into 8 Ω. The loudspeaker is connected again at a power level of about 3.5 W. Note that these are average power levels, so you may get quite different results when testing with music or white noise (13 W off, 5 W on). Network R1-C1 is a rudimentary peak detector which helps to improve the response of the circuit to noise.



The circuit works fine from a frequency of about 60 Hz. At lower frequencies, the relay may chatter. The protection is, therefore, mainly suitable for mid-range drive units and tweeters.

If you want to use the protection for higher signal levels, it is recommended to use a 12-V relay, and change the value of R2 according to

$R2 = 1.2V / I_{(nom)}$

where I_{nom} is the nominal coil current of the relay. The calculated value of R2 is then rounded off to the nearest (higher) E12 value.

FET T1 limits the maximum permissible input voltage to about 38 V. Current consumption of the protector is about 70 mA when actuated. The indicated test voltages apply to that state.

[Design by T. Giesberts - 964077]

Contents

single-chip AF power amplifier

A Burr-Brown Application

If you're looking for an audio amplifier that produces a lot of power using an absolute minimum number of components, this one is for you.

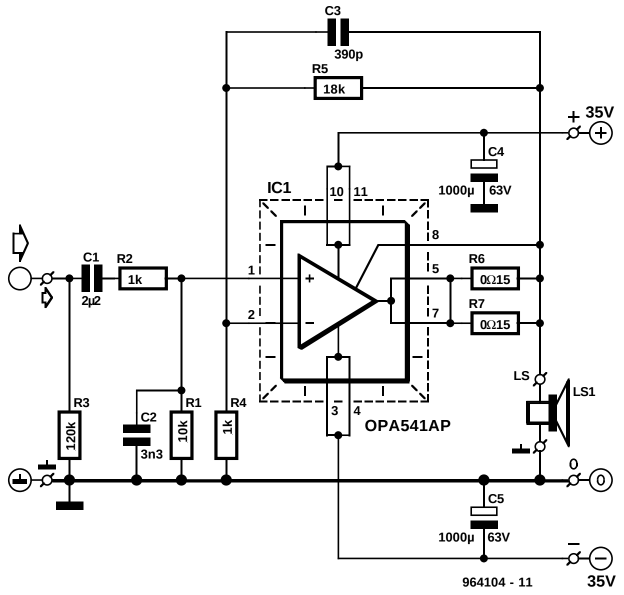
Burr-Brown's OPA541 is a power opamp capable of operation from power supplies up to ±40 V and delivering continuous output currents of up to 5 A. Internal current limit circuitry can be user-programmed with a single external resistor, protecting the amplifier and the load from fault conditions. The OPA541 is available in an 11-pin power plastic package and an industry-standard TO-3 hermetic package. The former is used here.

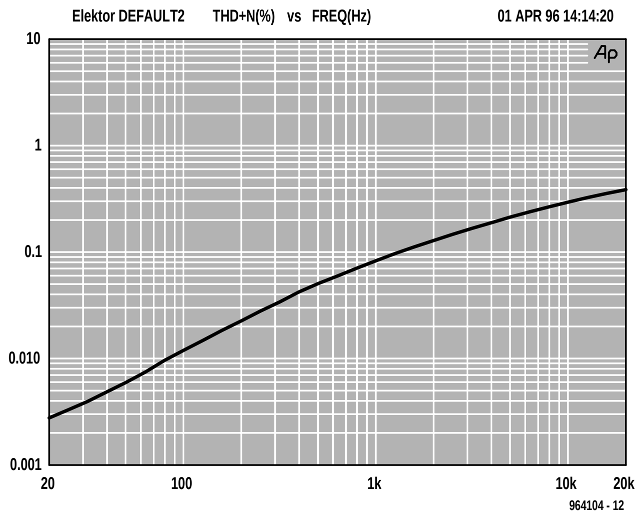
Although the OPA541 is primarily intended for applications like motor drivers, servo amplifiers and programable power supplies (says B-B), it is also fine for a medium-power AF amplifier with reasonable specifications. The design shown here is capable of supplying about 60 watts into an 8-Ω load. This is achieved with an audio drive level of 1.3V_{rms} and a symmetrical supply voltage of ±35 V. The on-chip current limiter is set to an actuation level of about 8.5 A by parallel-connected resistors R6/R7. This level ensures that the maximum drive margin can also be achieved with a load of 4 Ω. Note, however, that R6 and R7 do not make the amplifier short-circuit proof, because that would require a current limiter threshold of 1.8 A, assuming that the IC is operated within its SOA (safe operating area, for details consult the B-B datasheets). The value of the resistor, R_{cl}, that determines the current limiter actuation level is calculated from

$$R_{cl} = (0.813/I_{abs}) - 0.02\ [\Omega]$$

In practice, the positive half-cycle of the output current will be limited somewhat earlier, at about 10% below the calculated level. The opposite is true for the negative current, which will be about 10% higher than the calculated level.

The amplifier is not a bad performer as regards distortion. The graph shows that the THD level remains well below 0.5% over the full audio spectrum, assuming that a gain of ×6 is programmed (R5 approx. 5 kΩ) and a supply voltage of ±35 V. The curve applies to an output power of 50 watts into 8 Ω.





Because the IC operates at a quiescent current of only 20 mA, cross-over distortion occurs readily. The theoretical bandwidth is, therefore, limited to about 22 kHz by capacitor C3. Input filter R2-C2 serves to reduce IMD (intermodulation distortion), and reduces the actual bandwidth to about 16.6 kHz. The low-frequency roll-off is set to 6.6 Hz by R1-C1.

The IC must be fitted on to a fairly large heatsink with a thermal resistance of 1.2 K/W or better. A suggested type is Fischer's SK85SA/75mm, which will be just about sufficient for music into a 4-Ω load.

(964104)

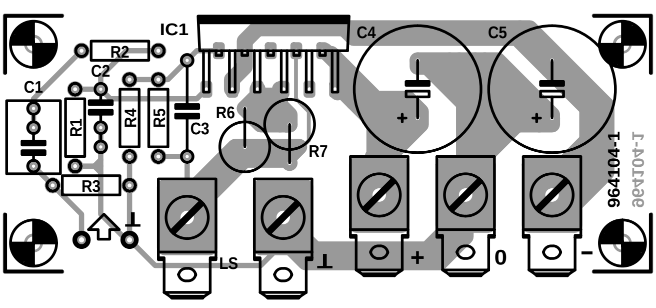
COMPONENTS LIST

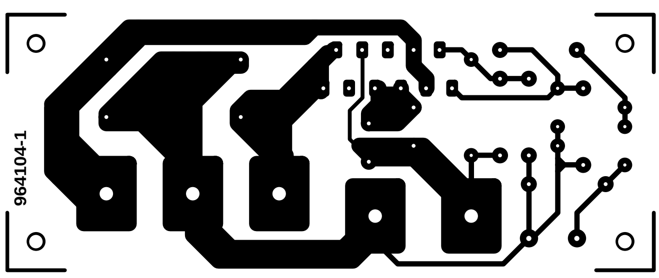
Resistors:
R1 = 10kΩ
R2,R4 = 1kΩ
R3 = 120kΩ
R5 = 18kΩ
R6,R7 = 0Ω15 5W

Capacitors:
C1 = 2µF2, MKT, 5mm
C2 = 3nF3
C3 = 390pF, 160V, polystyrene
C4,C5 = 1000µF 63V radial

Semiconductors:
IC1 = OPA541AP (Burr-Brown)

Miscellaneous:
5 spade terminals, screw mount
Heatsink, approx. 1°K/W
Printed circuit board, order code 964104-1 (see Readers Services page)





parametric equalizer

Straightening the frequency characteristic of an audio system is possible by either of two types of equalizer: graphic and parametric.

A graphic equalizer consists of a series of tuneable active filters, normally one for each frequency band. In spite of the complexity of this type of equalizer, it only provides adjustment of the amplification or attenuation in a given band. Also, however carefully the filters have been computed, in practice, the final characteristic never coincides with the positions of the potentiometers on the control panel.

A parametric equalizer normally consists of far fewer filters. Moreover, not only the amplification/attenuation of these filters is variable, but also their central frequency and *Q*-factor.

In the present circuit, the central frequency is set with P_2 in each of the three frequency bands (20–200 Hz, 200–2000 Hz, and 2–20 kHz), selected with S_1.

The *Q*-factor is set with P_1 in the range 0.25–2.5. This factor determines the slope of the skirts of the frequency characteristic.

The amplification/attenuation is set with P_3 over a range from –12 dB to +12 dB.

The curves in the lower diagram show various settings of the equalizer, starting with the upper curve: (a) maximum amplification, lowest *Q*; (b) maximum amplification, highest *Q*; (c) maximum attenuation, highest *Q*; (d) maximum attenuation, lowest *Q*.

The circuit in the diagram can provide only one dip or peak in each frequency curve. If this is not sufficient, a number of circuits per channel may be necessary. This is one of the reasons that it is advisable to use a simple buffer amplifier at the input. Another is that the input impedance of the filter varies appreciably during the setting-up process.

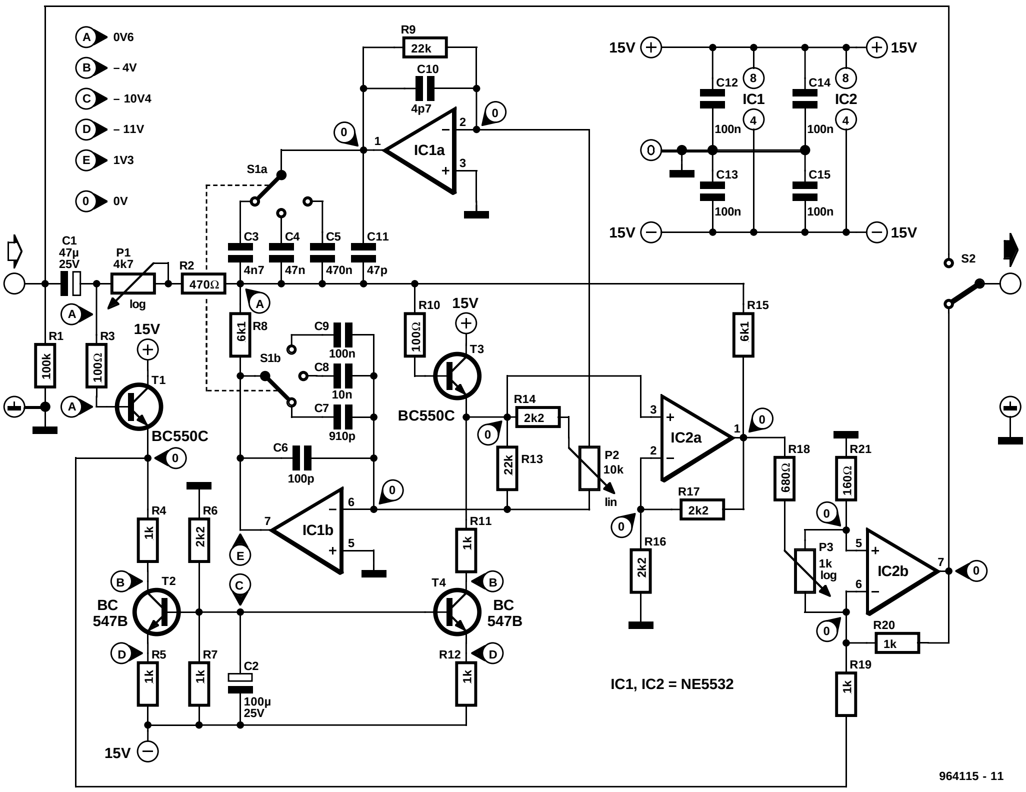
A few notes on the construction. Circuits IC_{1a} and IC_{1b} simulate a tuneable LC circuit in parallel with S_1 and P_2. The design is such that the reactances of the simulated inductance and capacitance hardly vary over the range of central frequencies that can be set with P_2. This makes independent setting of the *Q*-factor with P_1 possible.

The output voltage of the filter is buffered by IC_{2a}, after which it is mixed with the signal originating from the emitter of T_1 in a ratio determined by the setting of P_3.

The complete filter can be shorted out with S_2 to enable a quick comparison to be made with a guaranteed straight frequency characteristic.

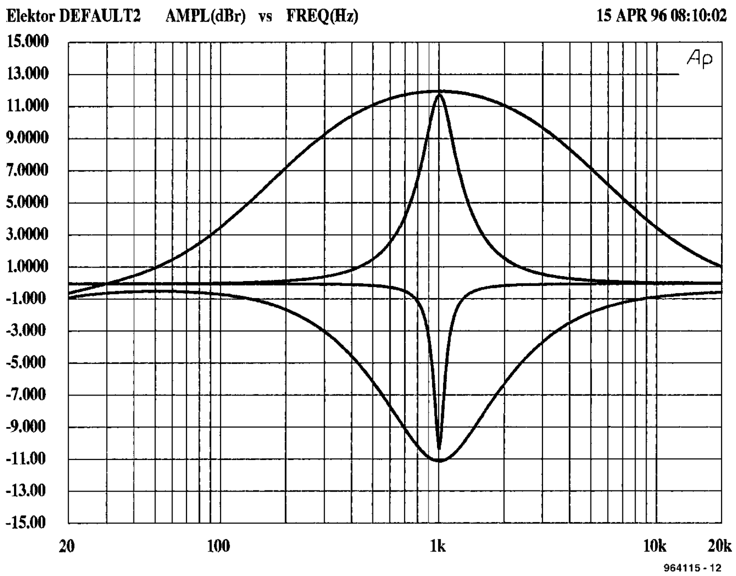
The circuit needs an input signal of 75–100 mV, which is below the standard line level.

Since the equalizer draws a current of only 25 mA, the supply voltage may be obtained from the preamplifier with which the equalizer is used.



The total harmonic distortion plus noise (THD+N) is ≤0.008 per cent at a frequency of 1 kHz (P_3 at the centre of its travel).

The value of R_{15} is not a standard one and the resistor must, therefore, be made from several others. A similar problem occurs with C_7, although its value is standard in the (often difficult to obtain) E24 series. This capacitor may, therefore, also have to be made from several others.



[R. Shankar - 964115]

delay line

The delay line makes possible an improvement of the surround-sound decoder published in this magazine in early 1995*.

The differential signal of the surround-sound channel is first passed through a low-pass filter and then split into two. Whereas IC_{1c} functions as a buffer, IC_{1b} inverts the signal. Both signals are then applied to identical memories of the bucket-brigade type.

Circuits IC_3 and IC_4 are controlled synchronously by IC_5. The delayed signals are buffered by IC_{2a} and IC_{2d}, after which they are applied to differential amplifier IC_{2c}. As the signals are in antiphase, the output of IC_{2c} is twice the level of each, so that the ripple caused by IC_5 is reduced appreciably (since the interfering signals are in phase).

The remaining interference signals at the output of IC_{2c} are inevitable, because they are related to the large tolerances of the memories.

According to data from the manufacturers, the distortion of an MN3008 is 0.5 per cent (average) and 2.5 per cent (maximum), while the amplification may vary up to ±4 dB from the nominal value. In the prototype, the use of one memory resulted in a distortion of 06–0.8 per cent at 1 kHz When two memories are used, the distortion drops to below 0.1 per cent. In both measurements, the clock frequency of IC_5 was 40 kHz (25 ms delay).

The improvement of the present circuit over the original is particularly noticeable with strong signals, because the signal-to-noise ratio increases to 63 dB.

The performance may improved even further by matching MN3008s. In the prototype, this reduced the distortion to 0.04 per cent. However, the price of the ICs may prove prohibitive for most constructors.

Another way of improving the performance is providing input buffers IC_{1b} and IC_{1c} with an offset compensation control. This requires the availability of a good distortion meter, however.

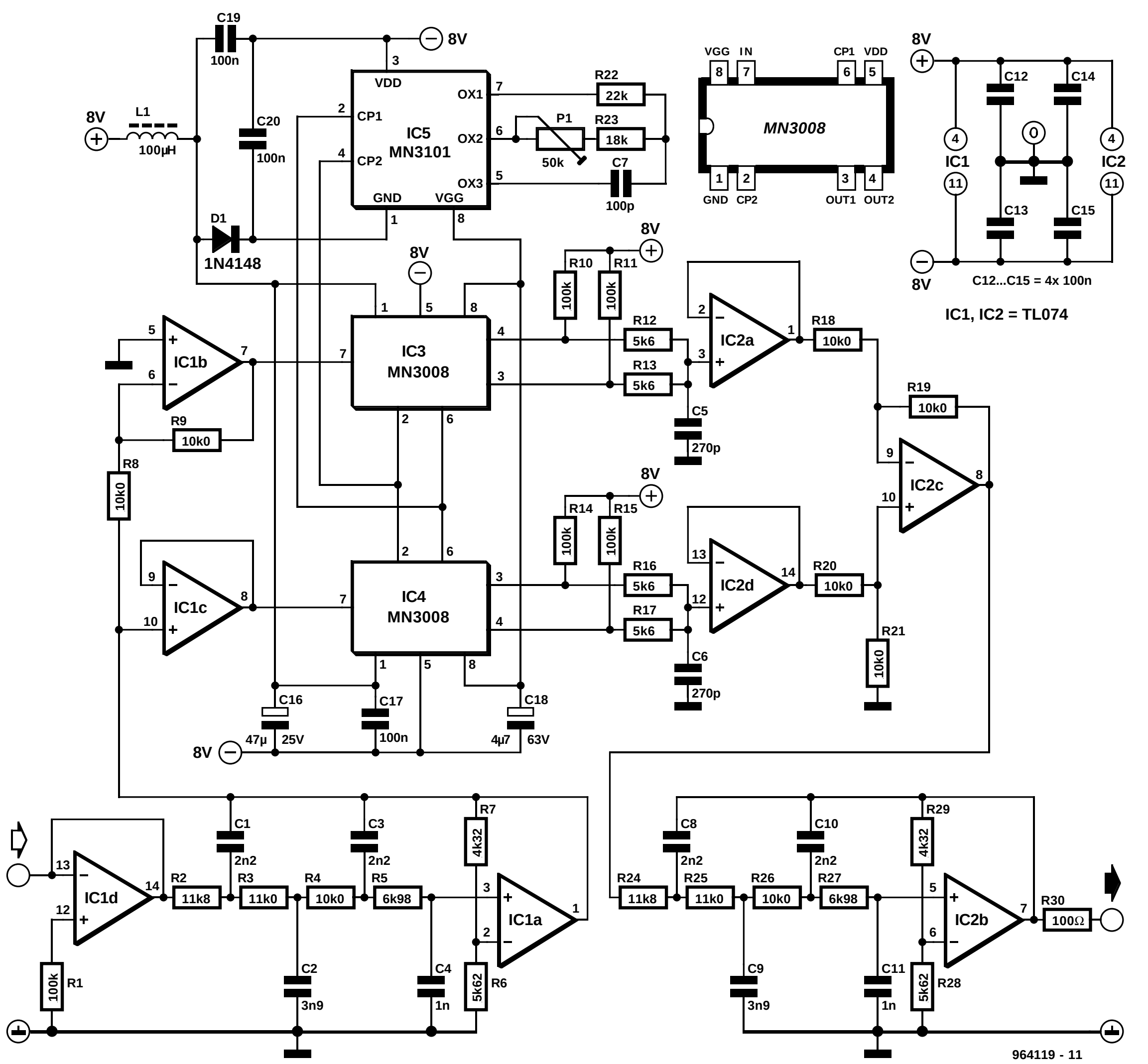
The bandwidth of the circuit is limited to about 7 kHz by input filter IC_{1a} and output filter IC_{2b}. Probably owing to tolerances of the capacitor values, the bandwidth in the prototype is 6.3 kHz. This is not terribly important, however. If desired, the bandwidth may be increased by giving R_2–R_5 and R_{24}–R_{27} proportionally lower values. Note that the bandwidth must not become larger than one quarter of the clock frequency, because the slope of the filter skirts does not allow this.

The clock frequency of IC_5 may be set between 30 kHz and 100 kHz with P_1. These values correspond to delays of 33 ms and 10 ms respectively.

The delay line draws a current of about 22 mA.

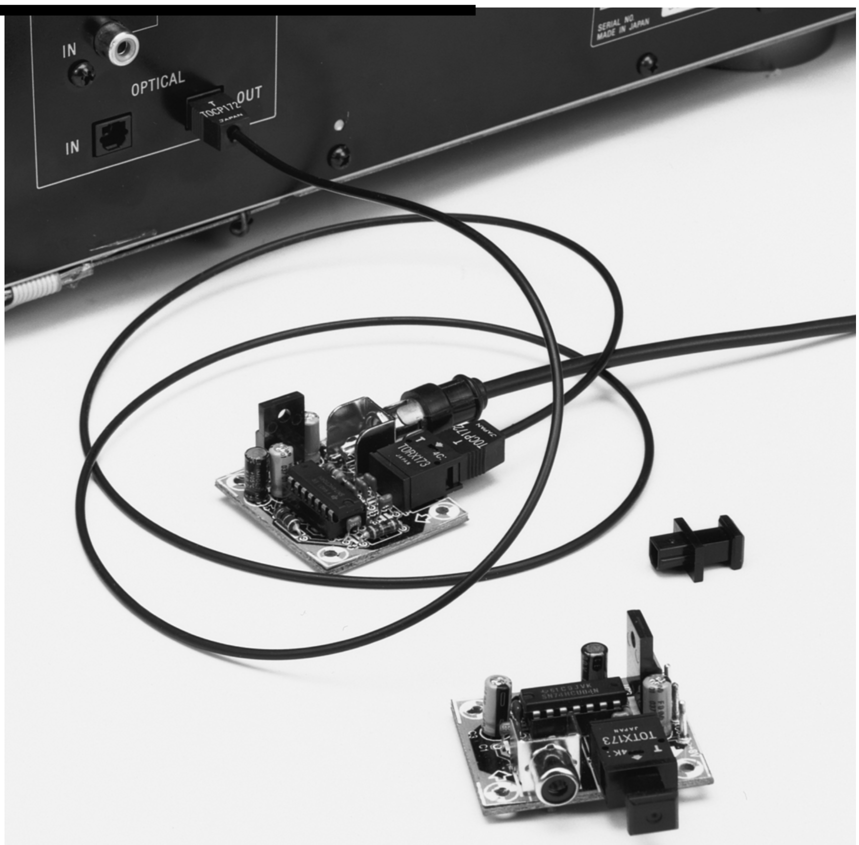
[T. Giesberts -964119]

* February 1995, p. 26



opto-to-coaxial audio converter

converts digital outputs



Most audio equipment produced in the past five to ten years has digital inputs and outputs. It is a pity, though, that there are two standards: optical and coaxial, and not all equipment has both. Fortunately, the connection difficulties that may ensue are easily resolved by the use of the adaptor described in this article. This enables any optical output to be converted into a coaxial one or vice versa.

Design by T. Giesberts

THE PROBLEM

Connection difficulties are frequent phenomena in the audio world. In the beginning there was the DIN (Deutsche Industrie Normen = German Industrial Standards) standard. Within a very short time, there were a number of variants on this: a five-pole plug does not fit into a three-way socket; 180° plugs and 240° sockets are not compatible either. And let's not even talk about the seven-way plugs and four-way square DIN connectors. The result of this was a thriving market in adaptors.

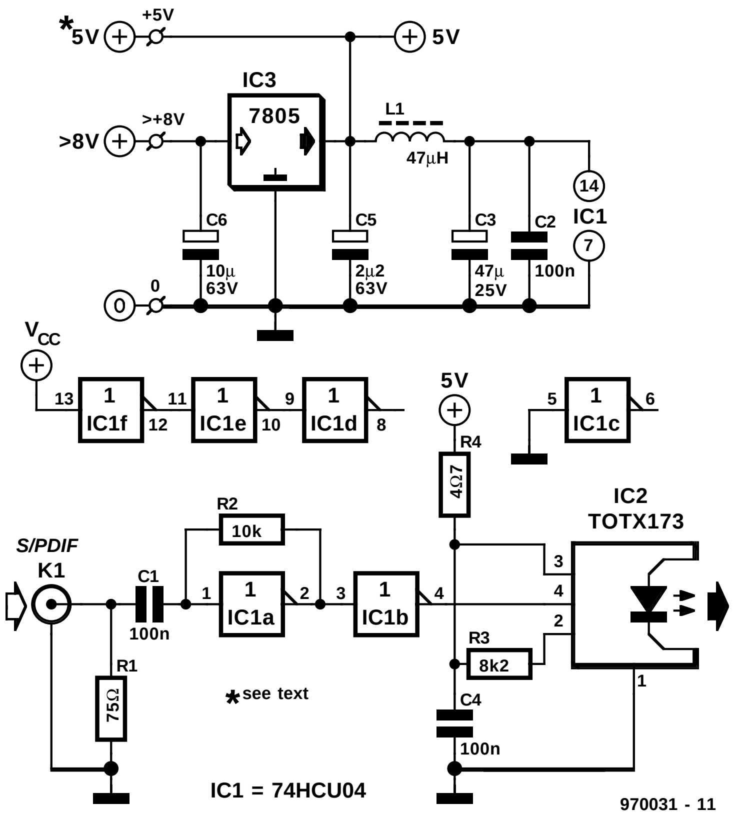
After the DIN era, there came the phono socket era. Once this had established itself, there was a period of relatively calm – until the advent of 'personal audio'. On this small equipment there was no space to fit standard phono sockets for the line outputs and so the 3 mm mini jack was born. And again, an adaptor was required if such a Walkman™ or Discman™ had to be connected to another piece of audio equipment.

It seems that history will repeat itself in respect of the connectors on modern audio equipment. From day one it was the case that where exchange of digital signals took place, there were two standards: one working with electrical pulses and the other with light pulses. In itself, that is, of course, no problem as long as both types of connector are fitted on all audio equipment. And that is mani-

** S/PDIF = Sony/Philips Digital Interface Format – the consumer version of the AES/EBU standard. This standard was devised by the American Audio Engineering Society and the European Broadcasting Union to define the signal format, electrical characteristics and connectors to be used for digital interfaces between professional audio products.*

Figure 1. A coaxial-to-optical converter basically needs only an amplifier, a buffer and a Toslink™ sender.

1



festly not happening. It is true that there is equipment on the market that has both, but the majority has not.

THE SOLUTION

Two converters are described that remedy the connection problem. They are easy to build and put paid to any interface problem between digital inputs and outputs.

One of the designs converts electrical signals into optical ones and the other does exactly the opposite. Both designs are based on the well-known Toslink™ modules. These compact converters may be built into the relevant audio equipment or they may be constructed as stand-alone units.

If the converter is built into the audio equipment, its power may be derived from that equipment. If used as stand-alone unit, the unit derives power from a mains adaptor.

FROM COAX TO OPTO

The circuit of the electrical-to-optical converter (EOC) is shown in the diagram of **Figure 1**. It is a straightforward design.

The S/PDIF* signal input to K_1 has a peak-to-peak value of about 0.5 V and is applied across 75 Ω terminating resistor R_1. This signal is amplified first by IC_{1a} and then by IC_{1b}, whereupon it has a peak-to-peak value of 5 V (HC level). The design of the amplifiers prevents any clipping taking place.

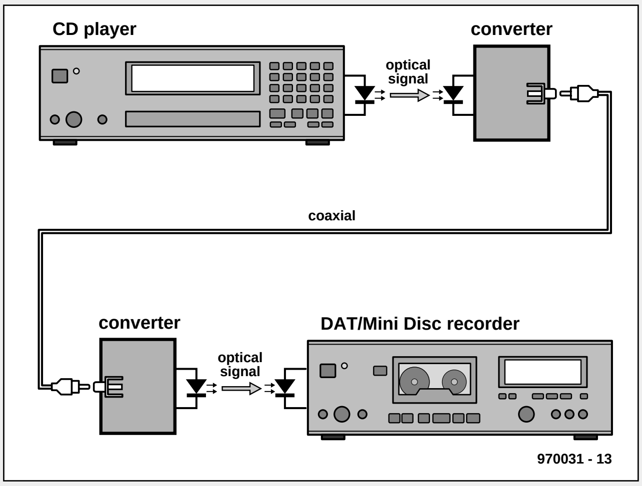
The supply to IC_{1a} is ½V_{cc} because of resistor R_2. Any input offset is blocked by C_1. This capacitor also prevents the operating point of IC_{1a} being affected by R_1.

The output of IC_{2b} is applied to IC_2, a Toslink™ sender, in which the actual conversion from electrical to optical signal takes place.

The circuit is powered by a supply line that is regulated by IC_3. Choke L_1, resistor R_4 and capacitors C_2–C_4 ensure adequate decoupling of the supply line. More about the supply later.

Long(ish) optical link

The converters described in this article may also be used for another purpose that has nothing to do with connection problems. It is a known fact that optical links suffer from a slight drawback: in practical use their length is very limited. This means that when it is desired to link two units that have only optical connectors over a distance of a few metres (10 feet or so) there is a little problem. However, the use of two converters makes the link possible. To do this, connect a converter to each of the two audio units via a short optical cable and link the converters via a standard, screened coaxial cable (see illustration). It may seem rather a long way around the problem, but it works well.



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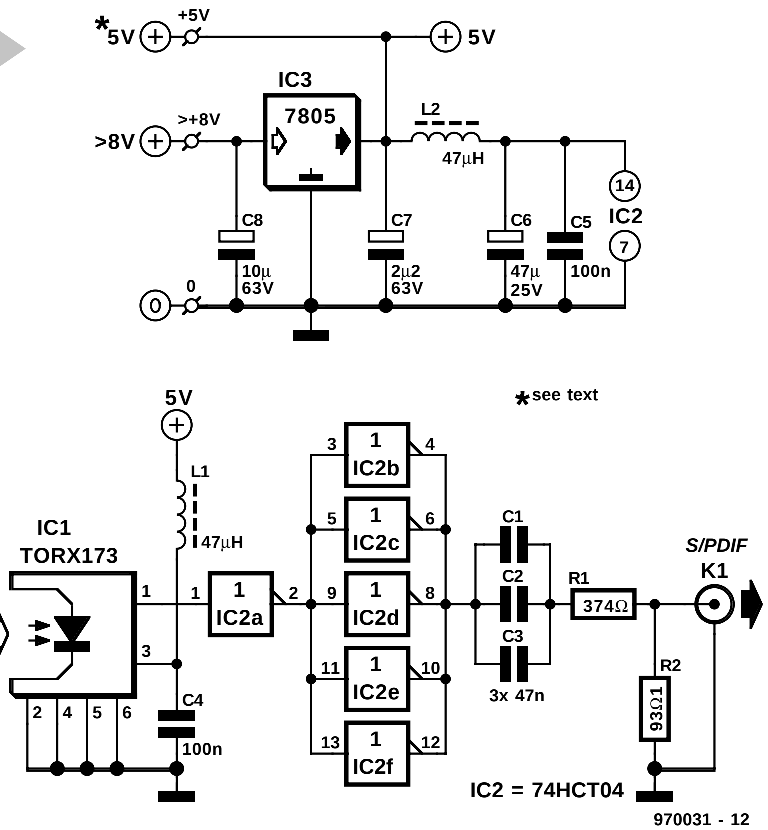

Figure 2. An optical-to-coaxial converter needs a Toslink™ receiver and not much more. A potential divider ensures the correct level of the output signal and the value of the output impedance.

FROM OPTO TO COAX

The circuit of the optical-to-electrical converter (OEC) is shown in the diagram of **Figure 2**. Like its sister circuit, it is a straightforward design.

The input signal is applied to IC_1, a Toslink™ receiver. Provided that the level of the optical signal is sufficient, the output of IC_1 is at TTL level (low = 0.5 V high = 3 V). This means that amplification is not necessary in this circuit, so that IC_1 is followed by buffer IC_{2a}. The five remaining inverters in IC_2 are paralleled to function as the output stage. The output signal is applied to K_1 via coupling capacitors C_1–C_3 and potential divider R_1-R_2.

The potential divider has two meet two criteria. The first is that, provided the output is correctly terminated into 75 Ω, the output voltage level does not exceed 0.5 V. The second is that the output impedance must be 75 Ω to prevent any mismatch. Consequently, the values of the two resistors are fairly low and this means that the output stage has to provide a relatively large current. This, in turn, is the reason that the five inverters in IC_2 are connected in parallel.

Asymmetric loading of the output stage is prevented by a.c. coupling.

Parts list

COAX-TO-OPTO CONVERTER

Resistors:
R_1 = 75 Ω
R_2 = 10 kΩ
R_3 = 8.2 kΩ
R_4 = 4.7 Ω

Capacitors:
C_1, C_2, C_4 = 100 nF ceramic, pitch 5 mm
C_3 = 47 µF, 25 V, radial
C_5 = 2.2 µF, 63 V, radial
C_6 = 10 µF, 63 V, radial

Integrated circuits:
IC_1 = 74HCU04
IC_2 = TOTX173 (Toshiba)
IC_3 = 7805

Miscellaneous:
L_1 = 47 µH (standard available)
K_1 = audio connector for board mounting
PCB Order no. 970031 (see Readers' services towards end of this issue)

OPTO-TO-COAX CONVERTER

Resistors:
R_1 = 374 Ω, 1%
R_2 = 93.1 Ω, 1%

Capacitors:
C_1–C_3 = 47 nF, ceramic, pitch 5 mm
C_4, C_5 = 100 nF, ceramic, pitch 5 mm
C_6 = 47 µF, 25 V, radial
C_7 = 2.2 µF, 63 V, radial
C_8 = 10 µF, 63 V, radial

Integrated circuits:
IC_1 = TORX173 (Toshiba)
IC_2 = 74HCT04
IC_3 = 7805

Miscellaneous:
L_1, L_2 = 47 µH (standard available)
K_1 = audio connector for board mounting
PCB Order no. 970031 (see Readers' services towards end of this issue

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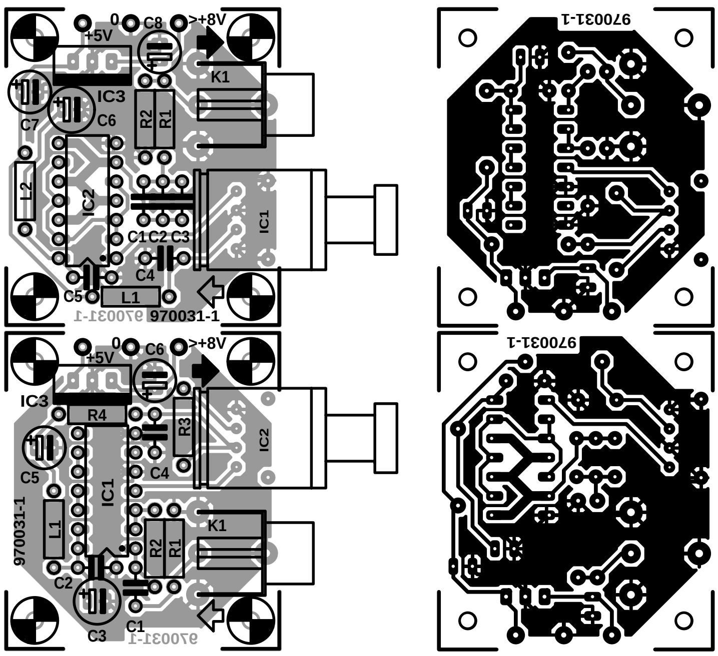

Figure 3. The printed-circuit boards for the two converters are manufactured as one.

This is effected by three capacitors connected in parallel so as to reduce the series resistance.

The power line is decoupled by chokes L_1 and L_2 and capacitors C_4–C_6.

POWER SUPPLY

As mentioned earlier, power may be derived from the relevant audio equipment into which the converters may be built, but it may also be provided by a discrete unit.

It is, of course, better to use a discrete power supply, since it being derived from the audio unit may give rise to hum and earth loops. In that case, it may even be necessary to use an output transformer to keep the various earths separated.

A discrete power supply is simply obtained from a mains adaptor whose output may be 8–35 V. Since the converters already have a regulator on board, and they draw a current of only a few milliamperes, the demands on the adaptor are small.

If the two converters are fitted in one and the same case, just one mains adaptor will do, otherwise each will need its own, of course. If only one mains adaptor is used, the regulator may be omitted from one of the boards and the +5 V and 0 terminals on the two boards interlinked. Again, take care that earth loops do not arise.

CONSTRUCTION

The converters are best built on the printed-circuit board shown in **Figure 3**. Note that this needs to be cut into two before any work is done.

Completing the boards is straightforward provided the usual sequence is followed: start with the resistors and capacitors, followed by the chokes and integrated circuits. Do not use sockets for the ICs. The Toslink™ modules and audio connectors may be soldered directly to the board.

When the board has been completed give it a good going-over to check that all connections are good, that the polarity of capacitors has been observed and that the ICs are fitted correctly. Note that the heat-conducting side of IC_3 is indicated by a small white square. **Figure 4** shows the completed prototype.

The enclosure(s) used depend on the application of the converters. If they are used permanently for one purpose only, the quality of the enclosure is not very important. If they are used for many purposes, a rather more robust case is necessary.

[970031]

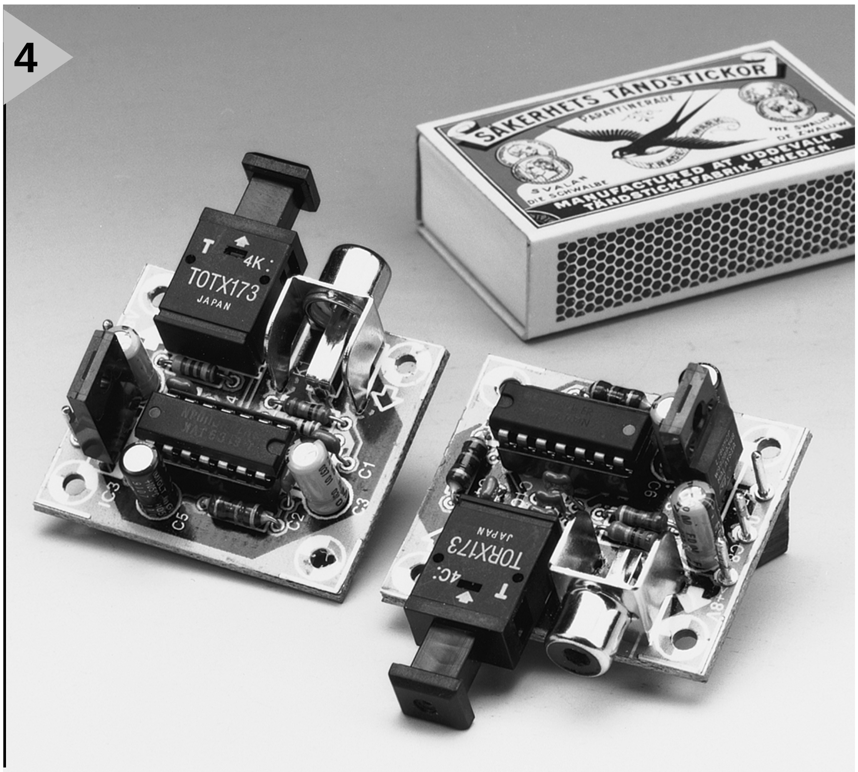

Figure 4. Completed prototype boards.

In passing ...

Since the late 1970s, many new consumer technologies have promised (threatened?) to transform our lives. Yet, the only really successful one has been that of the compact disk – CD – introduced by Sony and Philips in 1982; most of the other designer dreams have not come true. Why? There are several reasons: the new technology may not have filled a market need; or it may not have delivered what it promised; or maybe it was just too expensive.

Venture capitalists, of which the USA has many, but Europe, alas, has hardly any, have a rule of thumb, known as the ×10 rule, which helps them decide which new technology to back. Basically, the rule makes them ask themselves the question: "Will the consumer think the new device or equipment is so much better (ten times) than what it is replacing that it justifies the change?"

In retrospect it seems, therefore, that digital audio tape (DAT), introduced by Sony and Philips in 1987, in spite of its advantages over both the CD and standard audio cassettes, did not pass the rule. It is now only found in specialist niches. Production has been discontinued.

Then, in 1992, Philips introduced the digital compact cassette (DCC) and Sony the mini disc. The DCC has been a total flop and the mini disc is today only moderately successful in Japan (although Sony is still promoting it hard in Europe).

Another memorable flop is the videophone, first demonstrated in the USA in the 1960s: production models did not appear until the late 1980s (in Japan). By 1990, fewer than 100,000 had been sold world-wide; it is no longer in production anywhere. Perhaps it will be revived when the telephone and computer will begin to work together efficiently (which they do not as yet).

Robots, at least the general-purpose type, have disappointed, too. They have not taken over the routine chores of our lives as was promised (foreseen by many) all those years ago. Unlike the DAT and DCC, however, they will be reborne, but not until well into the 21st century.

[975032]

Contents

µP-controlled mixer board

No more noisy controls – ever!

The audio mixer board presented in this article cannot be compared with any mixer board previously published in this magazine. In a way, it is the prelude to a new era in audio engineering. All controls are operated by integrated digitally controlled amplifiers (DCAs). The position of the slide potentiometers is converted by a microprocessor into a control signal for the DCAs. It is even possible to do away with the potentiometers and microprocessor and leave the control of the mixer board to a PC, but this will be the subject of a future article.

Design by T. Giesberts

Parameters

- ▶ *8 controlled audio inputs*
- ▶ *all input signals can be directed to the left-hand or right-hand output*
- ▶ *maximum attenuation 63 dB in 1 dB steps*
- ▶ *mute function*
- ▶ *automatic muting at power up*
- ▶ *buffered line inputs and outputs; external amplifiers not required*
- ▶ *signal-to-noise ratio 82 dB with 10 dB headroom*
- ▶ *distorsion (THD+N) 0.007%*
- ▶ *standard 3-wire serial interface*
- ▶ *number of channels may be expanded*
- ▶ *wide range of applications: multimedia systems; PC sound cards; studio mixers; musical instruments, and many more*

Wherever there is a battle between digital and analogue technologies, the outcome is almost certain to be in favour of the digital. Analogue will live on, but only in some specialist uses. So, we have digital recording, digital control, digital television, digital mobile telephones, and more. Nevertheless, it may still come as a surprise to some to find a design for a digital mixer board, which many consider a traditional stronghold of analogue audio engineering. True, you will still find a number of slide potentiometers, but these operate with direct voltages only, not audio signals. The circuit consists of just two ICs, one of which is entirely digital and the other is a hybrid: part digital, part analogue. There are no transistors or op amps – nevertheless, it works like a dream!

INTRODUCTION

The circuit is based on an 8×2 digitally-controlled audio mixer Type SSM2163 from Analog Devices. Each of the eight inputs can be mixed under digital control to a stereo output A simplified block diagram of the device is shown in **Figure 1**. Each input channel can be attenuated by up to 63 dB in 1 dB intervals, and also fully muted. Furthermore, any input can be assigned to either or both outputs. A standard 3-wire serial interface is used as well as a Data Out terminal to facilitate daisy-chaining of multiple mixer ICs.

The control signal for the mixer board is provided by a Type ST62T25 microcontroller. This devices rapidly scans the potential at the wipers of the potentiometers and converts this information into an 8-bit code. The controller also monitors the position of two switches at each input so that the operator can arrange for the relevant input signal to be applied to the left-hand or right-hand channel or both. This information is contained in the control code.

THE SSM2163

The SSM2163 consists of an analogue (signal processing) section and a digital (control) section.The channel attenuation level and mixer functions are controlled by digital registers, which are loaded via a serial interface. A hardware mute input is included to asynchronously force all inputs into the muted state.

Analogue section

The analogue signal path is shown in **Figure 2**. Each input has a nominal impedance of 10 kΩ. Each input therefore appears as a digitally programmable 10 kΩ potentiometer. The SSM2163 input impedance remains constant as the attenuation level changes. So, the sources that drive the device do not have to drive complex and variable impedances.

The attenuated input is applied to the left-hand and right-hand channel inputs of the mixer. Each mixer channel consists of an analogue switch and a buffer amplifier. If the channel is selected (via the appropriate bit in the mixer control register), the analogue switch is turned on. The buffer amplifier is included after the analogue switch so that the gain of each channel will not be affected by the potentiometer setting or by the on-resistance [RDS(ON)] of the switch.

Each mixer channel that is ON is then summed into its respective (left-hand or right-hand) mixing summing amplifier. (If both the mixer channels are ON, the attenuated analogue input will be applied to both the Left and Right summing amplifiers). The buffered output of the summing amplifier will supply a current of ±500 µA to an external load.

Digital interface

The digital interface consists of two banks of eight data registers with a serial interface (**Figure 3**). One register bank holds the left/right mixer control bits, while the other register bank holds the 6-bit attenuator value.

To access the SSM2163, the controller writes a value to the serial shift register which selects the appropriate input channel register for subsequent attenuator-load operations. There are two ways: if bit 7 (MSB) is 1, the SSM2163 interprets the byte as an address; if bit 7 is 0, the SSM2163 interprets the byte as a data byte. Normally, the address byte is sent first. This indicates in which of the eight input channels the attenuation is to be altered and whether the signals are applied to the left-hand or right-hand channel or to both. Next, a data byte is sent that determines the degree of attenuation of the selected channel. Normally, this is followed by an address byte and a data byte for another channel. It is also possible, however, to write a sequence of data bytes. The selected channel remains the same, of course, but its attenuation changes with every data byte. This arrangement enables fading in and out without the necessity of writing the address every time. If, for instance, during the fading out of a channel a series of values is sent to the same address, a single write operation to the address register suffices.

When the mute input is made high, all channels are muted: this action does not affect the set attenuation values, which are retained. Switch-on noises are obviated by forcing all inputs into the muted state at that instant.

Serial data control inputs

The SSM2163 provides a simple 3- or 4-wire serial interface. Data is input on the DATA IN

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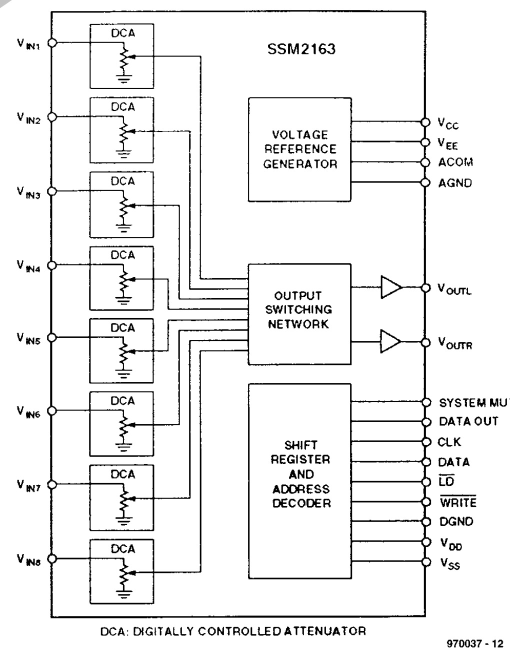

Figure 1. Simplified block diagram of digitally controlled audio mixer Type SSM2163.

Figure 2. The analogue signal path of the SSM2163. It contains the variable attenuator and the switches that determine whether the input signal is applied to the left-hand or right-hand channel.

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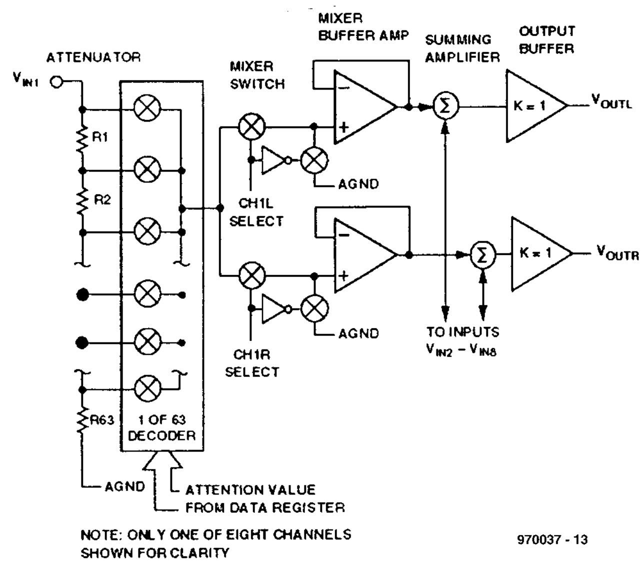

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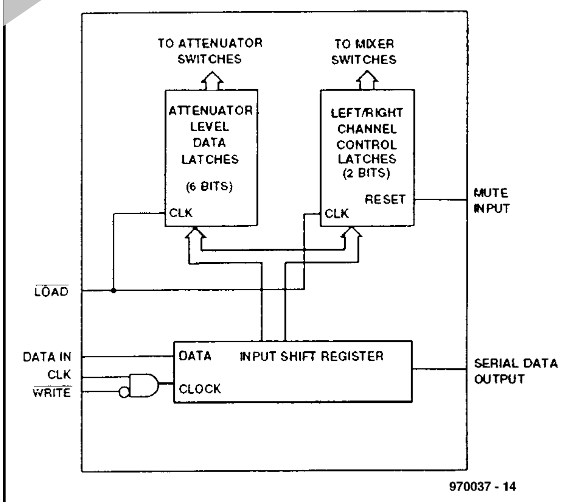

Figure 3. Block diagram of the digital serial data interface of the SSM2163.

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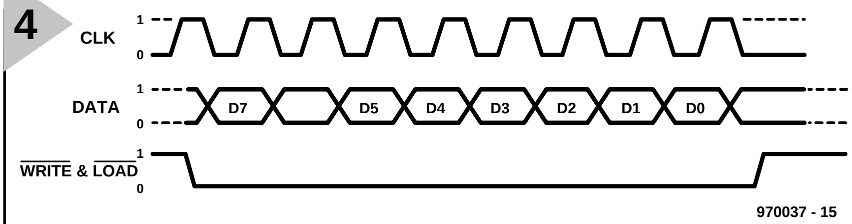

Figure 4. Timing diagram of the three-wire mode of operation.

pin, while CLK is the serial clock. Data can be shifted into the SSM2163 at clock rates up to 1 MHz.

The shift register clock, CLK, is enabled when the WRITE input is low. The WRITE pin can therefore be used as a chip select input. However, the shift register contents are not transferred to the register banks until the leading edge of LOAD. In most cases, WRITE and LOAD will be tied together, forming a traditional 3-wire serial interface. The process is clarified by the three-wire mode timing diagram in **Figure 4**.

To enable a data transfer, the WRITE and LOAD inputs are driven low. The 8-bit serial data, formatted MSB first, is input on the DATA IN input and clocked into the shift register on the leading edge of CLK. The data is latched on the leading edge of WRITE and LOAD. If the data is an address, the mixer control is updated. If the data is an attenuator value, the leading edge of WRITE and LOAD will update the appropriate attenuator value.

MUTE input

The MUTE pin, which is active high, provides a hardware input to force all the channels asynchronously into the muted state at any time. Most µC pins are in high impedance state or configured as inputs at power-up, so the SSM2163 will automatically be muted at switch-on. The mute function does not affect the attenuator values stored in the attenuator control registers.

Serial data input format

As mentioned previously, data is written to the SSM2163 in two 8-bit bytes. The serial data format is shown in **Figure 5**. The first byte sent contains the channel address and the Left/Right output mixer control bits. The address byte is identified by the MSB being high.

The second byte contains the data, that is, the attenuator value. The six LSBs of this byte set the attenuation level from 0 dB to –63 dB. The MSB of the data byte must be a logic zero.

The standard format for data sent to the SSM2163 is an address byte followed by a data (attenuator level) byte. In some cases, however, only one byte needs to be sent. For instance, attenuation levels are not affected by the MUTE input. To turn a muted channel on, simply send an address byte with the left-to-right mixer bit set. The addressed channel will immediately be enabled, using the previously set attenuation level. Furthermore, once a channel is addressed, the attenuation level can be varied by sending additional data bytes. For instance, fading a channel can be accomplished by simply increasing the data value sent to the SSM2163.

CIRCUIT DESCRIPTION

All connection between the inputs at the left and the outputs at the right of the circuit diagram in **Figure 6** are interconnected by screened cable. The screen of this cable is earthed.

Diode D_2 is arranged to make it light when data are being sent to IC_2. Controller IC_1 provides the requisite control signals, for which purpose I/O pins PA_0, PA_1, and PA_2 are the DATA, CLK and WRITE-LOAD outputs respectively. The 16 I/O pins at the left of the IC function as analogue inputs. These pins are sampled in turn; the input data so found is converted into a control signal for IC_2.

It will be seen that the inputs into IC_1 are all direct voltages. The attenuation level as well as the left/right information for each channel is set with the aid of potential dividers. The wipers of the slide potentiometers with which the attenuation of the input signals is set are linked to ‘P1’, ‘P2’, and so on, respectively. The other terminals of the potentiometers are connected to +5 V and 0 respectively. When the potentiometers are almost closed, the voltage is low and the attenuation great. When they are gradually opened, the potential at their wiper rises and the attenuation is reduced.

A potential divider consisting of three

Figure 5. Clarification of how two 8-bit bytes are constructed.

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SELECTION	ADDRESS MODE (MSB)	ADDRESS		LEFT	RIGHT			(LSB)	DATA MODE (MSB)	DATA						(LSB)
INPUT CHANNEL 1	1	X	X			0	0	0	0	X						
INPUT CHANNEL 2	1	X	X			0	0	1	0	X						
INPUT CHANNEL 3	1	X	X			0	1	0	0	X						
INPUT CHANNEL 4	1	X	X			0	1	1	0	X						
INPUT CHANNEL 5	1	X	X			1	0	0	0	X						
INPUT CHANNEL 6	1	X	X			1	0	1	0	X						
INPUT CHANNEL 7	1	X	X			1	1	0	0	X						
INPUT CHANNEL 8	1	X	X			1	1	1	0	X						

OUTPUT SELECT
1 = SELECTED, 0 = NOT SELECTED
INPUT SELECT
X = "DON'T CARE", SHADED AREA = DATA

DATA						ATTENUATION
0	0	0	0	0	0	0dB
0	0	0	0	0	1	-1dB
0	0	0	0	1	0	-2dB
·	·	·	·	·	·	·
1	1	1	1	0	1	-61dB
1	1	1	1	1	0	-62dB
1	1	1	1	1	1	-63dB

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2

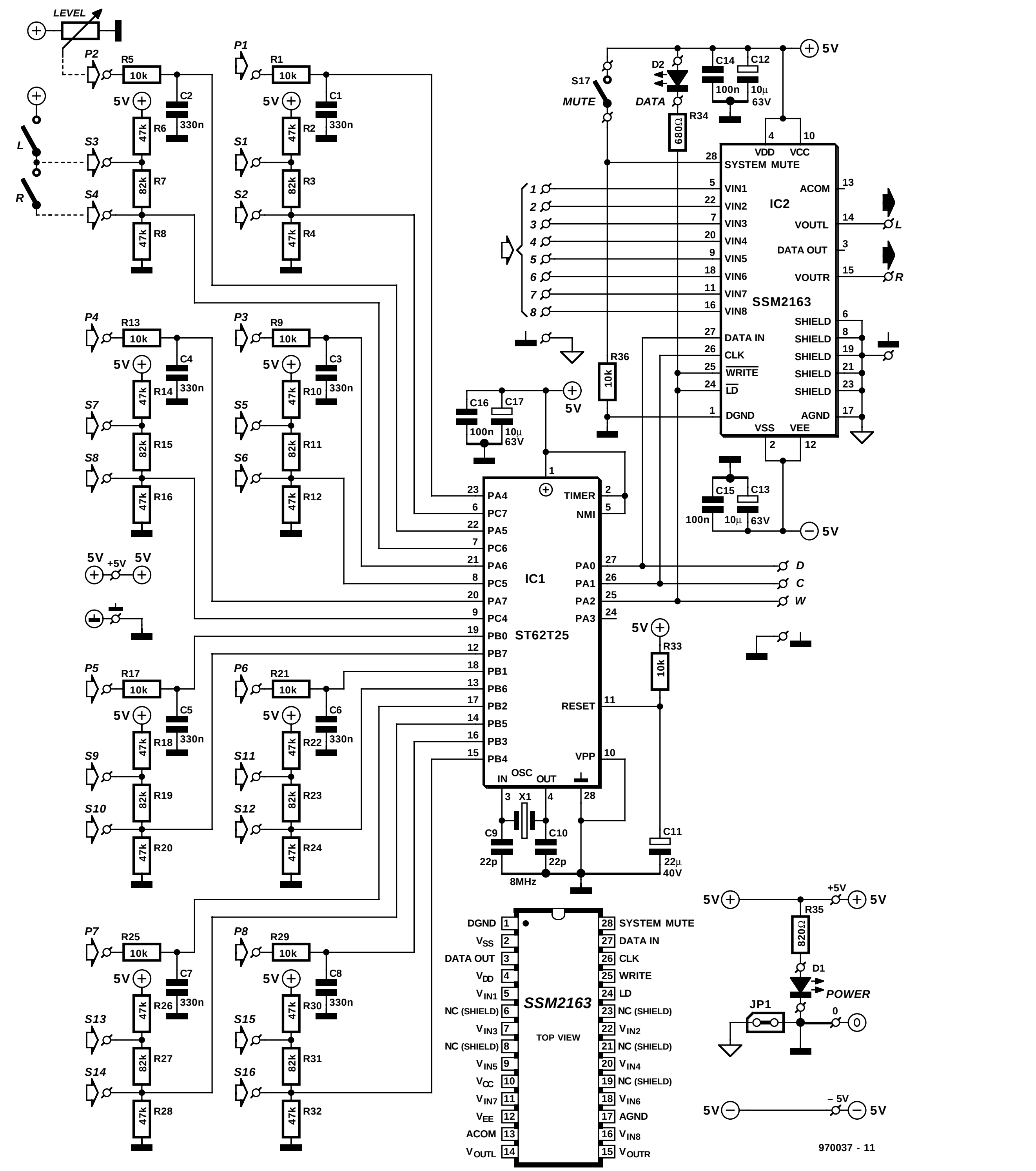


Figure 6. Circuit diagram of the mixer board. The dashed lines near Channel 2 indicate how the potentiometer and switches for the left-hand and right-hand channels must be connected.

resistors is provided for the left/right information for each channel. The individual resistors are switched with the aid of two switches. For instance, in the case of channel 2, it is easily calculated that when both switches are open, the voltage at the junction of R_7 and R_8 is about 1.3 V. The input signal is then not transferred to the output of the mixer board. When S_3 is closed and S_4 is open, the potential at junction R_7-R_8 is about 1.8 V and the input signal is

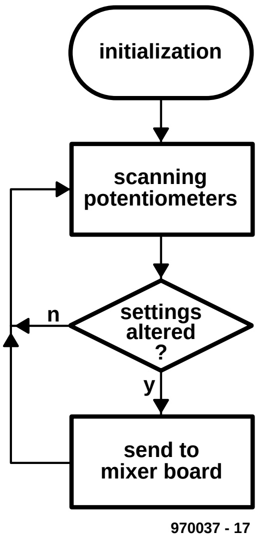
Contents

microcontroller control

The microcontroller used in the mixer board is a Type ST62T25 from SGS-Thomson. This device has 20 I/O lines which can be switched at will when the program is running to form an input or output in various configurations. Sixteen of them may be linked to the internal analogue-to-digital converter (ADC) to form the lines via which the positions of the slide potentiometers and the switches are written.

The positions of the slide potentiometers and the switches are continuously sampled by the software, which also processes the results of the analogue-to-digital conversions and stores them in the internal RAM. When even one of these settings differs from the previous one, the microcontroller passes all data to the mixer IC.

The allocation of the port lines to potentiometers and switches looks rather untidy on the circuit diagram; this results from the requirement of making the layout of the printed-circuit board as straightforward as possible. Note that PA_0–PA_3 cannot be linked to the ADC.



Analogue-to-digital

The simplest way of operating the ADC of the ST62 controllers is starting the conversion and waiting in a loop until the EOC (end-of-conversion) bit goes high, whereupon the normal process of the program can be continued. From a software point of view, this may be all right, but it appears that owing to internal and external noise it is impossible to obtain an accuracy of even 6 bits. The use of a WAIT command, which disables a large part of the processor, brings about a real improvement. The ADC generates an interrupt when the EOC goes high and this re-enables the processor.

The SSM2123 uses only six bits to set the attenuators and there is therefore nothing against ignoring the two LSBs of the ADC. Yet, this is not sufficient to achieve a stable setting of the mixer IC over the whole control range of the potentiometers: in some situations D_2 flashes to indicate that new data are being transferred. This is, of course, self-evident, because even if the two LSBs are ignored they continue to exert some influence. If bit 1 is high, a variation of just one LSB is sufficient to affect higher, relevant bits.

There are ways and means of solving this. For instance, it would be possible to sample a potentiometer a couple of times in succession and average the results. In the present design, a simple means is used: the entire 8-bit ADC value is retained, while the next sample is accepted only if its value differs by more than two LSBs from the previous one. This matter is less problematical when the state of the switches is written: in that case only four results are possible: some noise and a slight accuracy in the potential divider do not matter much

Digital to the mixer IC

Sending data to the SSM2163 is fairly straightforward. Three lines from port A of the ST6225 carry the CLK, WRITE and data signals. Unfortunately, with this configuration of the port, this cannot be effected with direct setting and resetting of the port bits. Because of space considerations, this cannot be further explained here.

applied to the left-hand channel. When S_4 is closed and S_3 open, the potential at junction R_7-R_8 is about 2.5 V and the input signal is applied to the right-hand channel. When both switches are closed, the potential across R_8 is 5 V and the input signal is applied to both outputs (mono).

To carry out these tasks, a special program is loaded in the ST62T25. The manner in which this device converts the direct voltages at the inputs into a control signal for IC_2 is described in the box on p. xx.

Circuit IC_1 is supplied by an asymmetrical voltage of 5 V, and IC_2 by a symmetrical line of ±5 V. Diode D_1 functions as on/off indicator. Jump lead JP_1 enables the screen of the cables to be linked to the negative supply line. This may be useful if hum occurs, unlikely though that may be. [970037]

(to be continued)

In passing ...

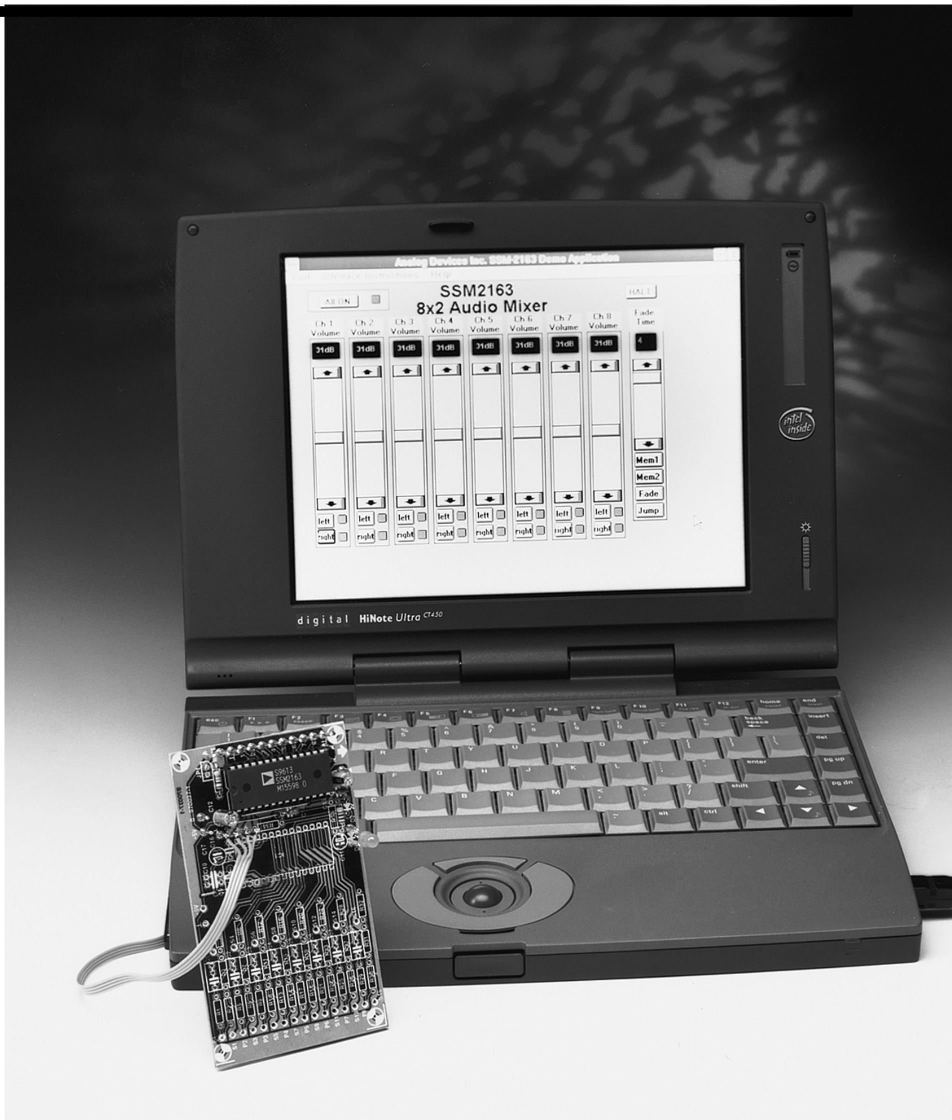
Sitting in a small restaurant the other day, I was intrigued to overhear a conversation between a small group of people at the next table about the relative merits or otherwise of digital cameras (apparently one of them had been given one as a Christmas present). It was argued that pictures from a digital camera are nowhere near as good as those from a 'real' camera. In a sense, this is true, of course, because a good-quality colour transparency requires about 80 million pixels (the tiny individual spots that all pictures are made up of), whereas a digital camera manages not much more than a few hundred thousand pixels (since its CCD – charge-coupled device – contains only that number of diodes). A simple 8-bit image allows each pixel to have 256 possible shades of grey or basic colours. With 24-bit colour (which most graphics programs use) each pixel can have more than 15 million different colour possibilities. Now, the memory in a digital camera just cannot cope with this: even a dozen 24-bit colour snapshots, each made up of only 200,000 pixels, will occupy 7.2 MB of memory (may be compressed by, for instance, JPEG to about 2 MB). A picture of the quality needed by a magazine such as ours would take up 60-80 MB. However, in publishing, a dark-room is a thing of the past. Colour-negative and transparency film is now normally machine-processed and scanned directly into digital files. The pictures are then edited on a computer screen and inserted directly into the electronic page make-up. Consequently, many professional photographers are already using digital cameras for situations where speed is important.

It is fairly certain that in the not too distant future there will be more digital cameras around than optical ones, in spite of the former's current imperfections. After all, the average person will only want to take snapshots, not become a photographic artist. If you are not convinced of this, look at the number of camcorders around which operate on the same principle as the digital still camera.

LS/975024

µP-controlled mixer board

Part 2 - Construction and computer control



This second and final part of the audio mixer board deals with the construction and the control of the board via a PC. To do this, all that needs to be done is to connect the board to the computer via a simple three-wire link and exchange some of the hardware by some intelligent, free software. This arranges for the mixer board to appear on the monitor, whereupon it can be controlled by the computer mouse. This part starts, however, with the construction of the board.

Design by T. Giesberts

Last month, the operation of digitally controlled audio mixer SSM2163, on which the 8-channel mixer board is based, the control circuits, and the complete circuit diagram, were discussed in detail. This second and final part starts with the construction and installation.

CONSTRUCTION

The mixer board is conveniently built on the compact printed-circuit board in **Figure 7**. In view of the small number of components, construction should not present anyone with undue difficulties. Sockets must be used for both ICs. The connections for the potentiometers and odd-numbered switches for the left-hand channel are at the left-hand side of the board; the even-numbered ones for the the right-hand channel are at the centre of the board; and the signal inputs and terminals for the mute switch are all at the right-hand side of the board. The supply line terminals and the two LEDs are located at the long sides of the board. **Figure 8** shows the completed prototype board.

INSTALLATION

While the construction of the board is simple, its fitting into a suitable case is rather more tedious: not difficult, but tedious since a great many short lengths of wire have to be soldered into place.

The audio (input) sockets and output sockets are linked to the relevant parts of the circuit with short lengths of screened audio cable. Flexible, stranded circuit wire may be used for all other connections. The +ve and –ve terminals of the potentiometers, as well as the +ve terminals of all left-hand channel switches, may be interlinked at the front panel. When all this is done, there will be quite a cable loom between the front panel and the printed-circuit board.

An illustration of the completed prototype is shown in **Figure 9**. The enclosure is specifically designed to house mixer boards. The one used in the prototype is a Retex Type ABOX RA.2, but similar cases (some without holes and cut-outs) are available from a number of manufacturers.

POWER SUPPLY

The mixer board requires a symmetrical ±5 V power supply. The current drawn from the positive line is about 20 mA, and that from the negative line, 8 mA. The supply in the prototype consists of a 2×9 V, 350 mVA transformer followed by a rectifier, a 7805 and a 7905 voltage regulator and the necessary smoothing capacitors. The design described in 'General-purpose power supply' elsewhere in this issue is very suitable.

POTENTIOMETERS

Slide potentiometers are available in a variety of models, sizes and prices. The ultimate choice is up to you and will depend on the demands you are likely to make on the mixer board. Whatever type is chosen, it cannot cause noise on the output, but at the same time its slide action must be smooth, and this is not always the case with the cheapest on the market.

If certain pairs of channels are used frequently, you may find it handy to couple the knobs of the relevant (adjacent!) potentiometers with a special clamp.

Note that it is possible in stereo use to do with one mono potentiometer for the two channels. If, for example, you arrange channels 1 and 2 permanently as the left- and right-hand channels of one stereo signal, channel 1 may be switched to 'left' with switch S_1, and channel 2 to 'right' with S_4 – the pins for P_1 and P_2 must be interlinked, of course.

ADDITIONAL CHANNELS

Although the board and software described in this article are not suitable for expansion, two or more SSM2163s can be paralleled to provide additional channels. This is done simply by interlinking inputs CLK, WRITE, LD and SYSMUTE. The DATA OUT pin of the first SSM2163 is linked to the DATA IN pin of the second. If outputs V_{OUTL} and V_{OUTR} of the two SSM2163s are summed by suitable buffers – a dual audio op amp such as the SSM2135 will do nicely – a 16-input, 2-output mixer is obtained. An advantage of the

Figure 7. The printed-circuit board for the 8-channel mixer console is surprisingly compact.

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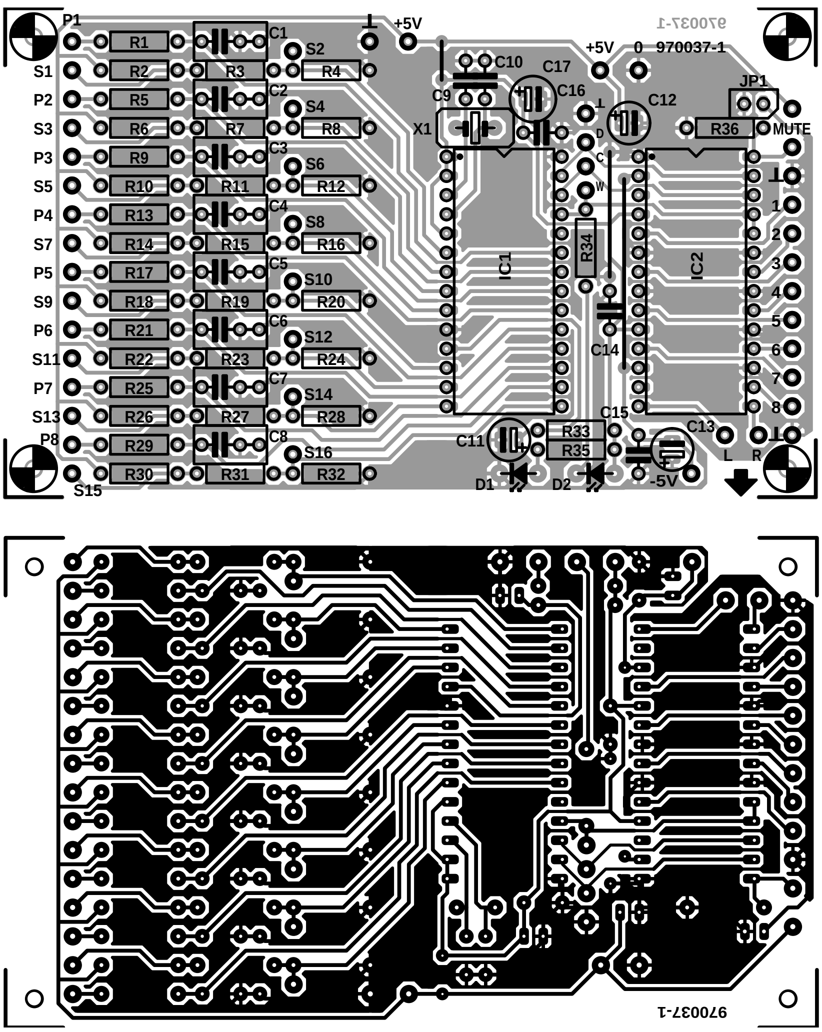

Parts list

Resistors:
R_1, R_5, R_9, R_{13}, R_{17}, R_{21}, R_{25}, R_{29}, R_{33}, R_{36} = 10 kΩ
R_2, R_4, R_6, R_8, R_{10}, R_{12}, R_{14}, R_{16}, R_{18}, R_{20}, R_{22}, R_{24}, R_{26}, R_{28}, R_{30}, R_{32} = 47 kΩ
R_3, R_7, R_{11}, R_{15}, R_{19}, R_{23}, R_{27}, R_{31} = 82 kΩ
R_{34} = 680 Ω
R_{35} = 820 Ω

Capacitors:
C_1–C_8 = 330 nF
C_9, C_{10} = 22 pF
C_{11} = 22 µF, 40 V, radial
C_{12}, C_{13}, C_{17} = 10 µF, 63 V, radial
C_{14}, C_{15}, C_{16} = 100 nF ceramic

Semiconductors:
D_1 = LED, low-current, green
D_2 = LED, low-current, red

Integrated circuits:
IC_1 = ST62T25B* – available programmed against Order no. 976502-1 (see Readers Services towards the end of this issue)
IC_2 = SSM2163 (Analog Devices)

Miscellaneous:
S_1–S_{17} = SPST switch
P_1–P_8 = slide potentiometer, 25 kΩ
JP_1 = jumper (wire bridge)
X_1 = crystal, 8 MHz
Enclosure = Retex ABOX RA.2 or similar
PCB Order no. 970037-1-AV*
* These are available as a combination against Order no. 970037-C – see Readers Services towards the end of this issue.

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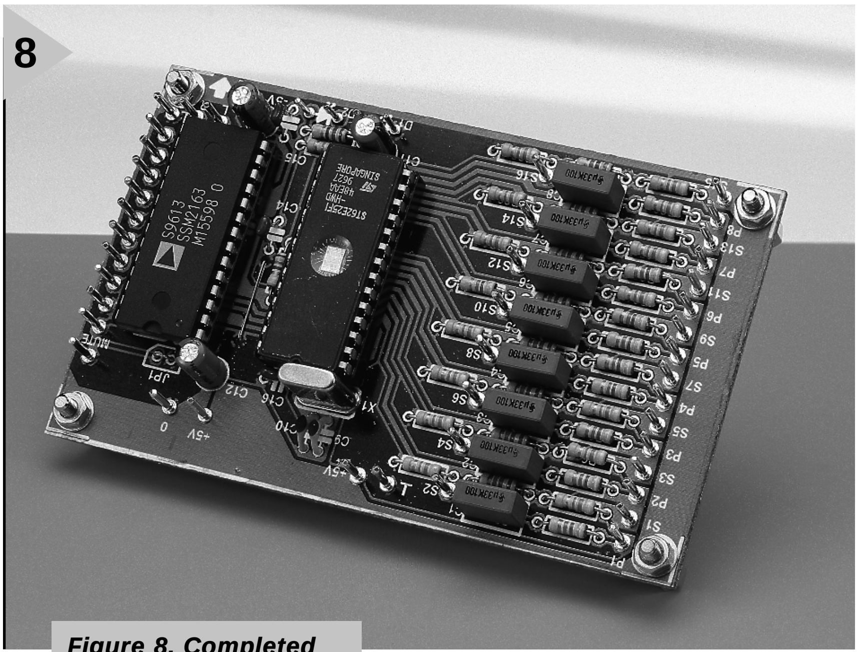

Figure 8. Completed prototype printed-circuit board. Note that sockets have been used for IC_1 and IC_2.

daisy-chaining is that it allows a 3-wire serial interface as used in the 8-input mixer to control two or more SSM2163s.

The serial data format for the daisy-chain circuit is similar to the 8-channel application, except that the SSM2163s are loaded in tandem. After setting WRITE and LOAD low, two bytes are clocked into the first SSM2163. When WRITE and LOAD return high, data will be latched into both SSM2163s simultaneously.

COMPUTER CONTROL

As mentioned earlier, the connections for the serial interface – DATA IN, CLK, and WRITE/LD – are accessible via PCB terminals. This enables control of the mixer board by a computer, for which Analog Devices, the manufacturer of the SSM2163, have special software available. Once you have this, the electronics can be reduced appreciably, and the operation of the slide potentiometers and the switches carried out via the monitor screen of the PC.

To achieve this, the D(ata in), C(lock), W(rite/LD) and ⊥ terminals on the PCB are linked to pins 2, 3, and 4 of the parallel (printer) port of the PC via a screened 3-wire cable. The screen is connected to the ⊥ terminal on the PCB and to pins 22, 23, and 24 of the parallel port. The manufacturers advise to include a 100 Ω resistor in series with each of the three links as protection against external potentials. Furthermore, IC_1, the potential dividers linked to the controller, the channel switches and the slide potentiometers on the mixer board are then superfluous and may be removed (but only if you are absolutely certain that you will always control the mixer board via the PC). What is left of the circuit in Figure 6 is shown in **Figure 10**, which also shows how the mixer board is connected to the PC.

SOFTWARE

Minimum requirements of the PC are 386/4 MB RAM/Windows 3.1. Since the software offered free by Analog Devices can be obtained only via the Internet, connection to this network (either your own or a friend's) is indispensable.

The software is installed as follows:

- Access Internet address www.analog.com/adibin/locate?SSM2163.
- Select 'SSM2163 Evaluation Board Software'.
- A file named 'SSM2163.ZIP' is then loaded. This can be unzipped with program 'PKUNZIP' or 'WINZIP'.
- Once the file is unzipped, a number of other files, among which 'Setup.exe' are shown. When this program is started, the software is installed automatically.
- After installation, a number of files have been added to Windows. Program 'SSM2163.exe' is found in a separate directory.
- When 'SSM2163.exe' is started, the software version of the mixer board appears on the monitor screen. Provided that the inputs and outputs of the SSM2163 are connected and that the mixer board is linked correctly to the PC, the program is then fully operational.

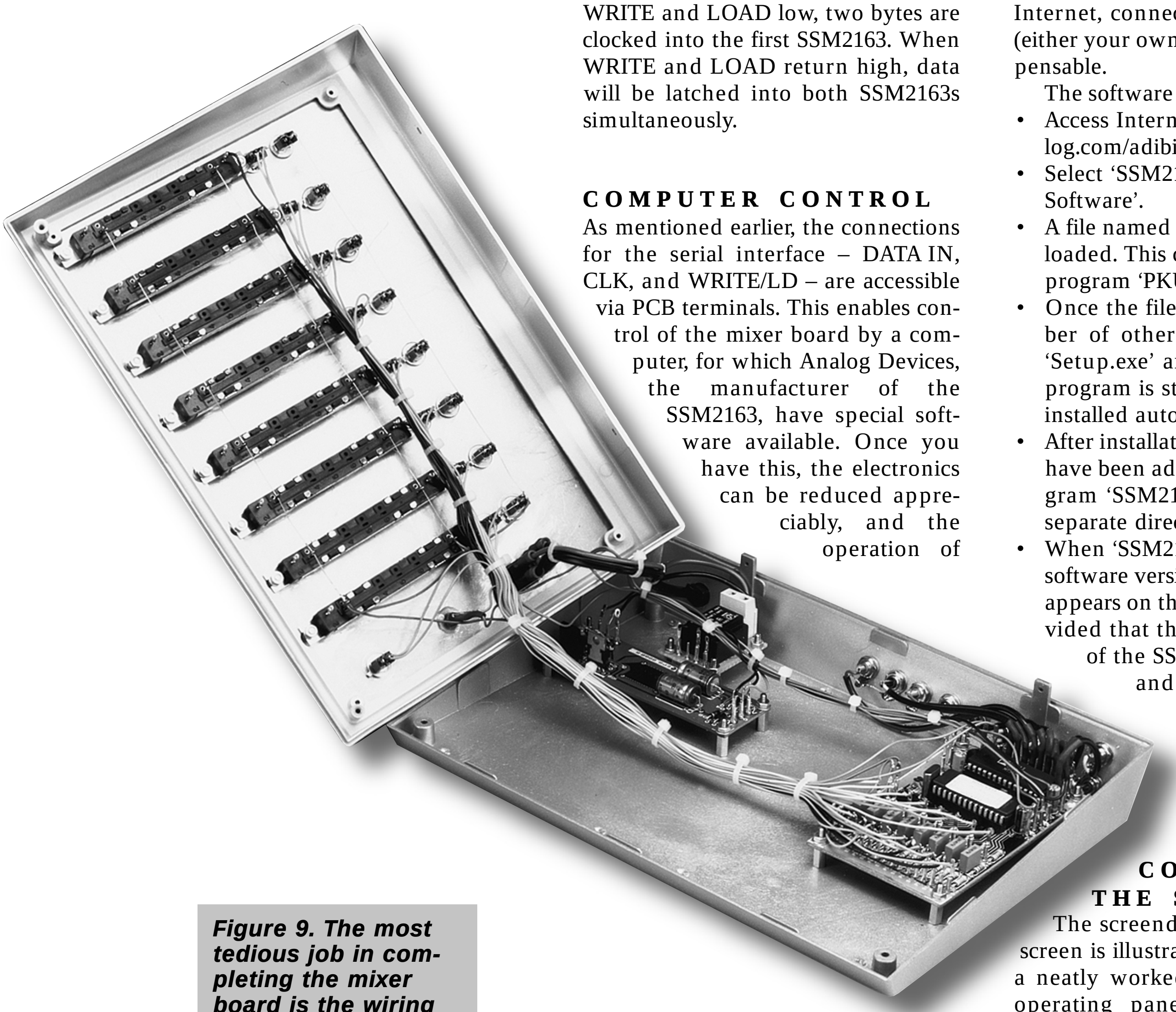

Figure 9. The most tedious job in completing the mixer board is the wiring up.

CONTROL ON THE SCREEN

The screendump on the monitor screen is illustrated in **Figure 11**. It is a neatly worked out version of the operating panel – the inscription 'SSM2163 8×2 Audio Mixer' leaves nothing to the imagination.

Just as on the actual board, there is a slide potentiometer for the volume control in each channel, as well as two switches with which the output channel is determined. Clicking on the knob of a slide potentiometer enables it to be moved to and fro with the mouse. If you are not used to working with a mouse, this operation may be slightly unusual, but you will soon get the hang of it, so that smooth in and out fading is achieved.

Note that there is a useful additional facility: in the window above each slide potentiometer, the exact attenuation in dB is shown.

Beside the channel switches marked left and right is an indication by means of a coloured square whether the switch is actuated or not: green is 'on' and red is 'off'. Clicking on the switch suffices to change the colour.

The switch at the top lefthand side, 'All ON' offers a kind of mute function, since it enables all channels to be switched on or off simultaneously. Here also: green is 'on' and red is 'off'.

The rectangular field at the righthand side, 'fade time', 'Mem 1', 'Mem 2', 'fade' and 'jump' offers some features that are not available on the original mixer board. These switches enable automatic fading between two previously selected positions of each of the eight slide potentiometers. If, with the mouse, you have set the controls to a certain position, these positions are stored in memory by a click on 'Mem 1'. If then the controls are set to different positions, a click on 'Mem 2' stores these positions in memory. If then 'fade' is clicked, the controls automatically move from one stored position to the other. The speed at which this occurs may be set between 1 and 10 seconds with the 'fade time' control – the set time can be read in the window underneath this control. A click on the 'jump' key moves the controls instantly from stored position 1 to stored position 2.

The 'HALT' key is for use if, for one reason or another, it is desired to interrupt the automatic fading process.

Reading this may make operation appear slightly tedious, but once you have the control panel on the monitor, you will see how smoothly it all works and how simple it is to operate.

[970037-2]

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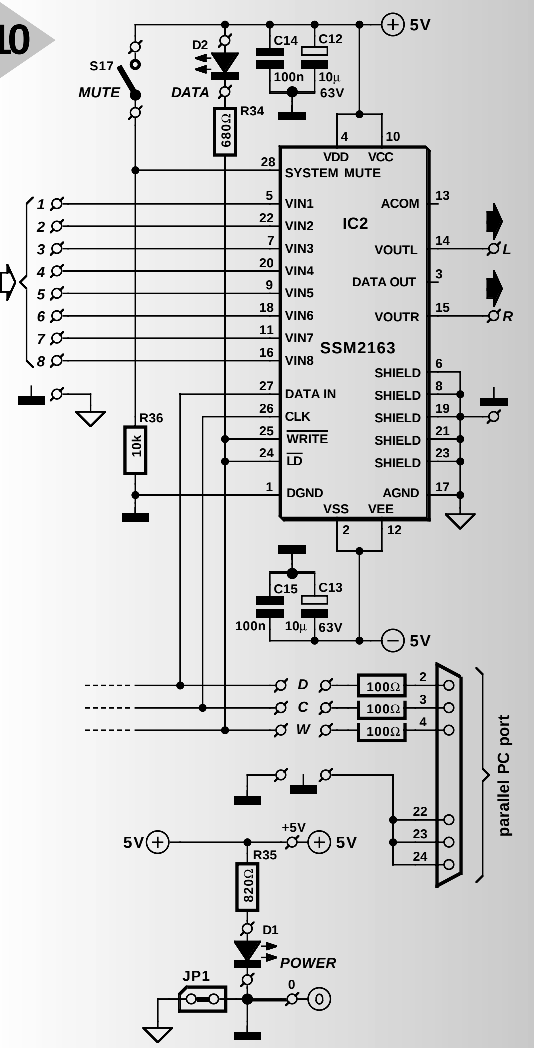

Figure 10. The mixer board may be controlled by a PC with the aid of this circuit. Pins D, C and W are linked via 100 Ω resistors to pins 2, 3 and 4 of the parallel port of the PC.

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Figure 11. Screen dump of the software developed by Analog Devices. Additional features are the indications of the attenuation in dB and the possibility of automatic fading between two previously stored settings of the controls.

Contents

compact power amplifier

Fast high-end design

The battery-powered preamplifier published in our February 1997 issue must have made many of our readers long for a matching power amplifier. Odd as it may seem, it is almost two years ago that we last published a high-end power amplifier. This realization prompted the present compact power amplifier whose quality and layout match those of the preamplifier. It offers medium power output, is not overly complex and has a very high slew rate.

Design by T. Giesberts

The introduction gives a good idea of the power amplifier described in this article.

The term 'compact' refers to both the power output and the dimensions of the amplifier. Of course, compactness is a relative term, because the amplifier contains no fewer than 19 transistors. Over the years, compactness has taken on a different meaning than, say, ten years ago. Then, a compact amplifier was understood to contain not more than eight to ten transistors; today this figure is nearer to twenty-five. After all, as everybody now realizes only too well, quality always has its price.

As far as output power is concerned, the present amplifier is definitely intended for normal domestic use. It produces some 50 watts into 8 Ω or 85 watts into 4 Ω, which is more than ample for its intended use. Even at relatively high volume, a domestic power amplifier seldom provides more than 1–2 power output. To high-volume freaks it may sound odd, but most loudspeakers produce a sound pressure of about 90 dB at a distance of 1 metre for an input of only 1–2 watts—enough to make the neighbours rush for the telephone.

It is clear, therefore, that the present amplifier has more than ample reserve to reproduce even the highest peaks in a piece of music without any discernible distortion.

A further benefit of the circuit is the use of such a relatively high quiescent current that with power outputs up to 2.5 W the amplifier operates in Class A, which enhances the overall quality of the reproduced sound.

There are some other aspects which will be reverted to in detail later. Firstly, the power output is produced by two insulated gate bipolar transistors (IGBTs). These bipolar MOSFETs or, if you prefer, gate-controlled transistors have been used in power amplifiers published by us before.

Secondly, the amplifier uses current feedback instead of the more usual voltage feedback. This has a beneficial effect on the open-loop bandwidth, the power bandwidth and the slew rate.

Thirdly, as may be seen from the technical data, the figures for harmonic distortion, intermodulation distortion and the damping factor are very good, indeed. They make it clear that although the amplifier is small in size, it is large in performance.

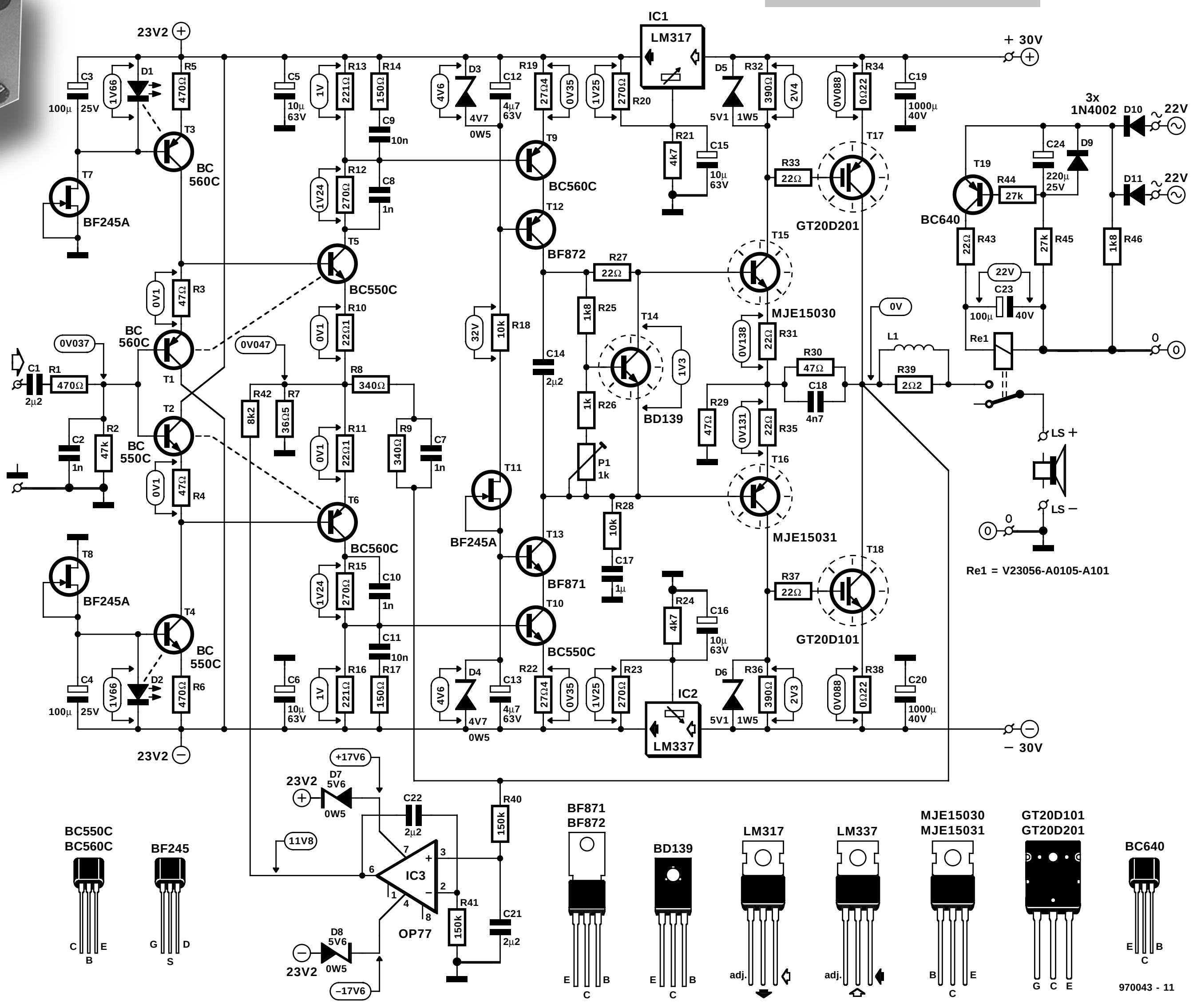
THE DESIGN

Experts in audio amplifier technology, as well as many other knowledgeable constructors, will see from **Figure 1** that in some respects the design of the power amplifier is different from the usual arrangement.

To start with, the almost obligatory differential input amplifier is not there. In its stead there is a symmetrical input stage that has some resemblance to the buffer stage used at the input of

Figure 1. The most conspicuous aspect of the circuit diagram is the omission of the traditional differential amplifier at the input. The output stage is based on insulated gate bipolar transistors.

1



Technical data

Input sensitivity	1 V r.m.s.
Output impedance	47.5 kΩ
Output power (0.1% THD)	50 W into 8 Ω; 85 W into 4 Ω
Power bandwidth (25 W into 8 Ω)	1.5 Hz–270 kHz
Slew rate	37 V µs^{-1}
Signal-to-noise ratio (1 W into 8 Ω)	107 dB (A-weighted)
	102 dB (B = 22 kHz linear)
Total harmonic distortion (B = 80 kHz)	
1 W into 8 Ω	0.0015% (1 kHz)
25 W into 8 Ω	0.0025% (1 kHz)
	0.008% (20 kHz)
Intermodulation distortion (50 Hz:7 kHz = 4:1)	
1 W into 8 Ω	0.0025%
25 W into 8 Ω	0.008%
Dynamic intermodulation distortion	
(Square wave 3.15 kHz, sine wave 15 kHz)	
1 W into 8 Ω	0.002%
25 W into 8 Ω	0.002%
Damping factor (8 Ω)	700 (1 kHz)
	450 (20 kHz)
Open-loop parameters (R_8 open)	
Amplification	×2500
Bandwidth	40 kHz
Output impedance	0.3 Ω

These data show the excellent performance of the power amplifier; exceptionally good are the power bandwidth and the slew rate. Some audio enthusiasts feel that performance data should be taken with a pinch of salt and that the important thing is a hearing test. This is not an opinion shared by this magazine. It is true that figures do not reveal everything, but it is surely admitted by all that poor data point to properties that cannot have a beneficial effect on the sound quality.

As usual, the data are accompanied by a number of curves obtained with an Audio Precision Analyser.

***Curves A** show the total harmonic distortion (THD + N) between 20 Hz and 20 kHz. The upper curve refers to 25 W into 8 Ω, and the lower to 1 W into 8 Ω. Both curves show an exemplary performance; note that the usual increase in distortion at high frequencies is small.*

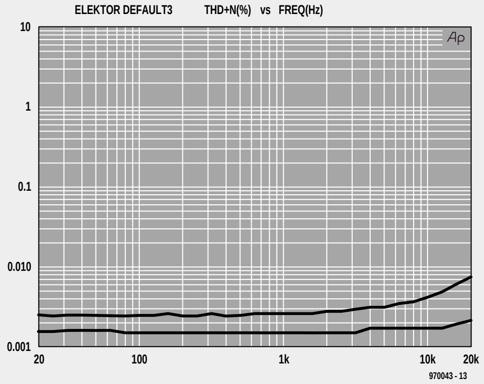
***Curve B** illustrates the distortion at 1 kHz as a function of the drive power over a bandwidth of 22 Hz to 22 kHz. The slight increase above about 2.5 W shows that from that point on the amplifier no longer operates in Class A. The clipping point is at 50 kHz.*

***Curves C** refer to the output power of the amplifier as a function of frequency with a load of 4 Ω and 8 Ω respectively.*

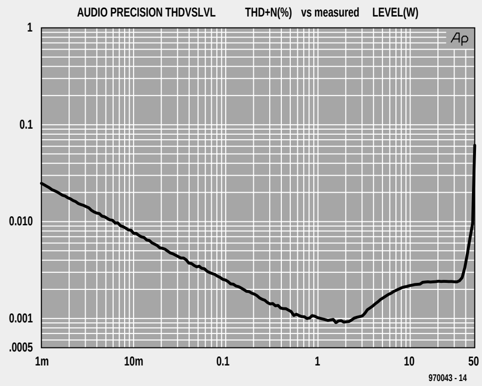
***Curve D** illustrates a Fourier analysis of a 1 kHz signal for an output of 1 W into 8 Ω. Although the 2nd and 3rd harmonics are clearly visible, their levels at –105 dB and –122 dB are well below that of the fundamental. All other harmonics remain below the noise floor.*

Listening tests showed the very good performance of the power amplifier even better than the technical data. The sound reproduction was very pleasant with good presence and transparent high-frequency performance that never became sharp. Low-frequency performance, both with classical and popular music, remained remarkably taut. The amplifier held the loudspeakers in an iron grip as it were and never relinquished control.

A



B



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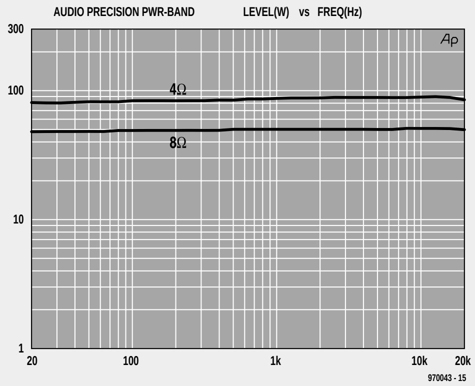

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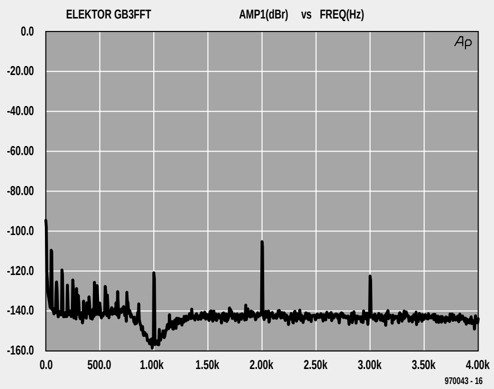

the battery-operated preamplifier.

In combination with current feedback this results in an amplifier that is appreciably faster than one with the traditional differential input stage.

The power amplifier has an open-loop bandwidth of some 40 kHz at the relative low amplification of ×2500. This enables the overall feedback, which many enthusiasts do not like, to be kept to a quite conservative level.

There is another side to everything and current feedback is no exception. A well-known but unavoidable drawback is the poor common-mode and supply-line suppression. In the present design, the effects of these have been countered to a large extent, however, by the use of two regulators in the power lines to all voltage amplifiers. This, in turn, has the drawback that the level of the supply voltage to these amplifiers is lower than that to the current amplifiers, although it should really be higher. This limits the maximum drive, but that is compensated by making the current amplifiers amplify ×2. Because of this, the output stage is not arranged as the usual emitter follower, but rather as a compound configuration.

Output transistors T_{17} and T_{18} are, as mentioned earlier, insulated gate bipolar types. These may be considered as transistors with a MOSFET input, which has the advantage of requiring a smaller current from the drivers, T_{15} and T_{16}. As this results in lower loading of the local feedback loop, it. means that the amplification of the drivers can be higher, which enhances the linearity of the current amplifiers.

INPUT STAGE

The input stage of the amplifier consists of emitter followers T_1 and T_2 and symmetrical voltage amplifiers T_5 and T_6 (**Figure 1**). Current feedback is obtained by coupling the output of the power amplifier to the emitters of T_5 and T_6.

The emitter followers provide impedance matching and arrange the base bias of T_5 and T_6. This arrangement relies on the fact that the drop across R_{10} and R_{11} is virtually the same as that across R_3 and R_4. The potential across the latter two resistors is held constant by current sources T_3 and T_4. Moreover, the reference voltages for these current sources are provided by D_1 and D_2, the current through which in turn is held constant by current sources T_7 and T_8.

To prevent the operating levels being upset by temperature effects in spite of these measures, T_5 and T_6 are thermally coupled to T_1 and T_2, while D_1 and D_2 are thermally coupled to current sources T_3 and T_4.

For best performance, it is advisable to select transistors T_1 and T_2 on the basis of equal base-emitter voltage and current amplification, but make sure that the (offset) potential across R_2 <50 mV. Since complete symmetry is virtually unattainable, there will always be some offset which is why the circuit based on IC_3 has been added. This stage compensates the input offset by placing an identical potential at feedback junction R_{10}-R_{11}. This will be reverted to later.

Combinations R_{14}-C_9 and R_{17}-C_{11} are compensating networks, while C_8 and C_{10} minimize the effects of the parasitic capacitances of T_5 and T_6.

VOLTAGE AMPLIFIERS AND OUTPUT STAGE

Voltage amplifiers T_5 and T_6 drive a push-pull amplifier based on T_9–T_{13}, which is arranged as a cascode stage for the following reasons. Firstly, such an arrangement limits the potential across, and the dissipation of, the actual voltage amplifier T_9-T_{10}. Secondly, a cascode configuration is ideal for obtaining high amplification coupled with a large bandwidth.

The d.c. operating point of the voltage amplifier is held stable by zener diodes D_3 and D_4, the current through which is in turn held constant by current source T_{11}.

The voltage amplifiers are loaded by the input impedance of T_{15}-T_{16}. Since this varies in accord with the output current of the power amplifier, it is linearized by resistor R_{28}. Capacitor C_{17} prevents any direct voltage from reaching the base of T_{16}.

The circuit based on T_{14} forms a temperature-independent transistor zener which has become a familiar aspect in many power amplifiers. This transistor is mounted on the heat sink adjacent to output transistors T_{17} and T_{18} and stabilizes the quiescent current through these devices. The level of this current is set with P_1.

Resistor R_{27} provides the additional negative temperature coefficient needed to compensate the time delay and thermal decay in the heat sink.

The output of the voltage amplifiers is applied to current amplifier T_{15}–T_{18}. Strictly speaking, this is a compound rather than a current amplifier, since, apart from current amplification, it also provides voltage amplification. The voltage gain (×2) is determined by resistors R_{29} and R_{30}; capacitor C_{18} serves to improve the response of the stage.

Since in a compound amplifier the collectors of the driver transistors form the output of the current amplifier, the gate-emitter potential does not influence the maximum drive from the voltage amplifier. This is an important benefit, because the gate-emitter potential may be several volts. The only limiting factor in the present amplifier, is, therefore, the knee voltage of T_{17} and T_{18}.

Emitter resistors R_{34} and R_{38} are low-inductance types to prevent any tendency to oscillation or other spurious effects.

Inductance L_1 wound on R_{39} enhances the performance of the power amplifier when the load is capacitive.

Zener diodes D_5 and D_6 protect the gates of T_{17} and T_{18} against overdrive.

In operation, the temperature of drivers T_{15} and T_{16} strongly affects the setting of the quiescent current. So as to make the power amplifier as stable with temperature variations as feasible, these transistors are, therefore, also mounted on the same heat sink as T_{14}, T_{17} and T_{18}. While it is true that the quiescent current will always vary to some (small) degree with temperature, the thermal coupling of the various transistors ensures that any drift is well within acceptable limits.

MISCELLANEOUS ASPECTS

So as to keep the power amplifier as compact as possible, it does not contain elaborate protection circuits.There is, however, a power-on delay to suppress annoying clicks and plops at on and off switching. Delayed on/off switching of the power lines is effected by relay Re_1. To avoid any noise on the signal paths via the supply lines, the relay has independent power lines. For this purpose, D_{10} and D_{11} rectify the secondary voltage of the mains transformer. The resulting direct voltage is buffered by C_{23} which ensures that the energizing voltage for the relay is 22–24 V.

After the power has been switched on, T_{19} begins to conduct slowly via R_{44}-R_{45}-C_{24}. It takes a few seconds before this transistor is fully on: only then will the relay be energized. When the power is switched off, C_{24} is rapidly discharged via R_{46} so that the relay is deenergized without any discernible delay.

To aid the offset setting, IC_3 compares the output of low-pass filter R_{40}-C_{21} with the direct voltage at the output of the power amplifier. If there is a difference, the operating point of T_5-T_6 is adjusted via R_{42} in such a manner that the average output of the power amplifier remains at earth potential. This arrangement ensures that the offset voltage at the output cannot exceed that of IC_3 (≤ 100 μV at 25 °C). It is a fact, however, that tiny variations caused by temperature changes cannot be eliminated entirely.

Zener diodes D_7 and D_8 reduce the supply voltage of ±23.2 V to the requisite level for IC_3.

The reasons for regulation of the supply lines to the input and voltage amplifiers has already been mentioned. The regulation is provided by IC_1 and IC_2. Type LM317 and LM337 respectively have been used for these regulators because they provide very good ripple suppression and tolerate a high peak input voltage. A further advantage of them is that their output potential can be adjusted very accurately by means of resistors R_{20}-R_{21} and R_{23}-R_{24} respectively. Capacitors C_{15} and C_{16} increase the ripple suppression to 70–80 dB.

CONSTRUCTION

The power amplifier is best built on the double-sided printed-circuit board shown in **Figure** 2. Making the board double-sided meant that it could be made very compact without the necessity of (too) long copper tracks at essential locations. The layout is unambiguous so that population of the board is straightforward.

Transistors T_{14}–T_{18} are located neatly in a row at one side of the board to facilitate their fixing on to the

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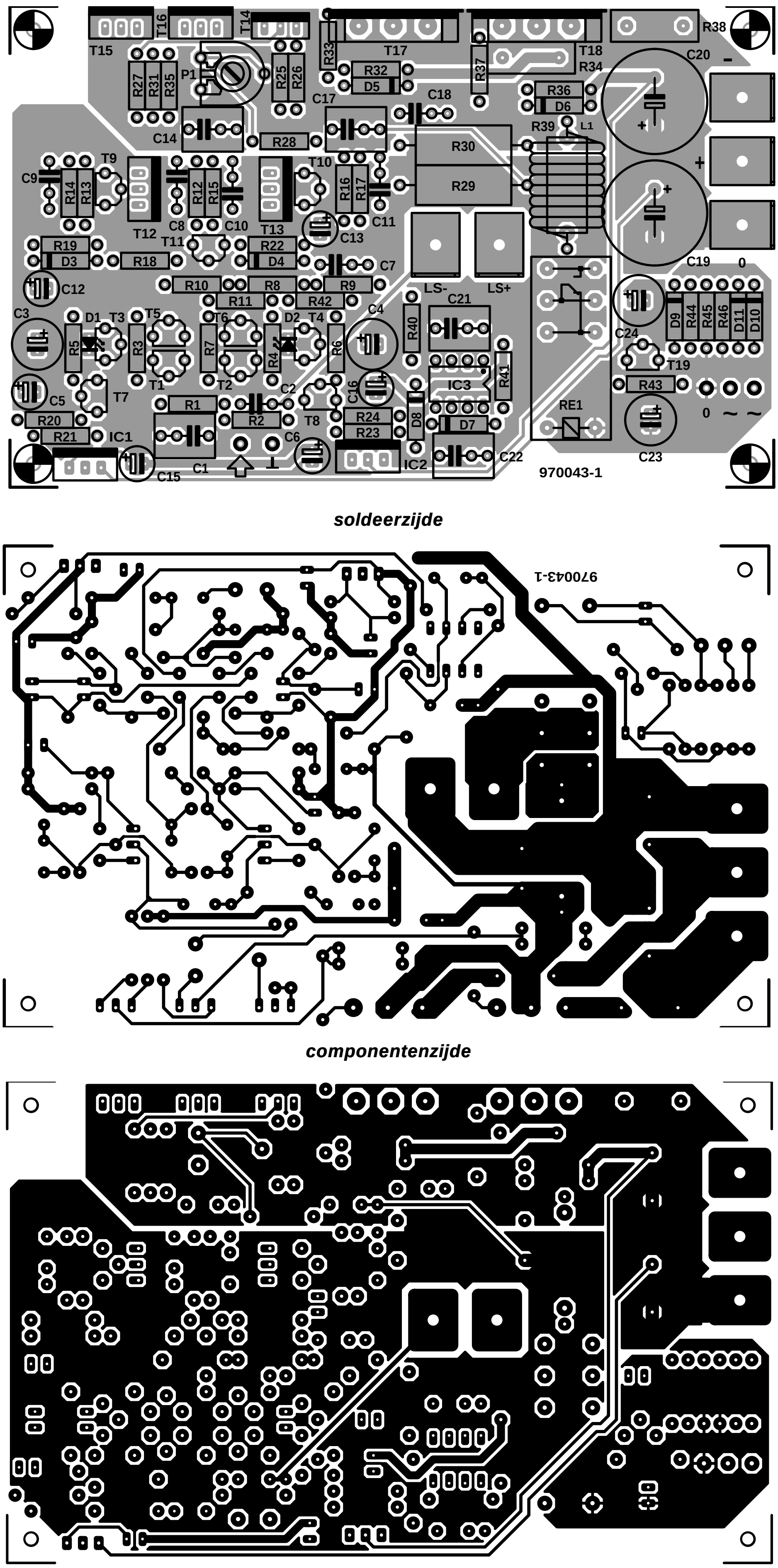
Figure 2. The double-sided printed-circuit board for the power amplifier. Large currents flow over only a few tracks.



Parts list

Resistors:
R_1, R_5, R_6 = 470 Ω
R_2 = 47 kΩ
R_3, R_4 = 47 Ω
R_7 = 36.5 Ω, 1%
R_8, R_9 = 340 Ω, 1%
R_{10}, R_{11} = 22.1 Ω, 1%
R_{12}, R_{15}, R_{20}, R_{23} = 270 Ω
R_{13}, R_{16} = 221 Ω, 1%
R_{14}, R_{17} = 150 Ω
R_{18}, R_{28} = 10 kΩ
R_{19}, R_{22} = 27.4 Ω, 1%
R_{21}, R_{24} = 4.7 kΩ
R_{25}, R_{46} = 1.8 kΩ
R_{26} = 1 kΩ
R_{27}, R_{31}, R_{33}, R_{35}, R_{37}, R_{43} = 22 Ω
R_{29}, R_{30} = 47 Ω, 5 W
R_{32}, R_{36} = 390 Ω
R_{34}, R_{38} = 0.22 Ω, 5 W, low inductance
R_{39} = 2.2 Ω, 5 W
R_{40}, R_{41} = 150 kΩ
R_{42} = 8.2 kΩ
R_{44}, R_{45} = 27 kΩ
P_1 = 1 kΩ preset

Capacitors:
C_1, C_{14}, C_{21}, C_{22} = 2.2 µF metallized polyester, pitch 5 or 7.5 mm
C_2, C_7, C_8, C_{10} = 1 nF
C_3, C_4 = 100 µF, 25 V
C_5, C_6, C_{15}, C_{16} = 10 µF, 16 V, radial
C_9, C_{11} = 10 nF
C_{17} = 1 µF, metallized polyester, pitch 5 or 7.5 mm
C_{18} = 4.7 nF
C_{19}, C_{20} = 1000 µF, 40 V, radial
C_{23} = 100 µF, 40 V, radial
C_{24} = 220 µF, 25 V, radial

Semiconductors:
D_1, D_2 = LED, 5 mm, flat
D_3, D_4 = zener diode, 4.7 V, 500 mW
D_5, D_6 = zener diode, 5.1 V, 1.5 W
D_7, D_8 = zener diode, 5.6 V, 500 mW
D_9–D_{11} = 1N4002
T_1, T_3, T_6, T_9, =BC560C
T_2, T_4, T_5, T_{10} = BC550C
T_7, T_8, T_{11} = BF245A
T_{12} = BF872
T_{13} = BF871
T_{14} = BD139
T_{15} = MJE15030 (Motorola)
T_{16} = MJE15031 (Motorola)
T_{17} = GT20D201 (Toshiba)
T_{18} = GT20D101 (Toshiba)
T_{19} = BC640

Integrated circuits:
IC_1 = LM317T
IC_2 = LM337T
IC_3 = OP77 (Analog Devices)

Miscellaneous:
L_1 = see text
Re_1 = relay, 24 V, 875 Ω with single make contact 16 A, 250 V, e.g. Siemens Type V23056-A0105-A101
5 off PCB mounting flat 3-way sockets
Heat sink 1.2 K W^{-1} or smaller – see text; e.g. Fischer SK85SA/75 mm from Dau (telephone 01243 553 031)
Insulating washer and insulating bushes for T_{14}–T_{18}
PCB Order No. 970043 – see Readers Services towards the end of this issue

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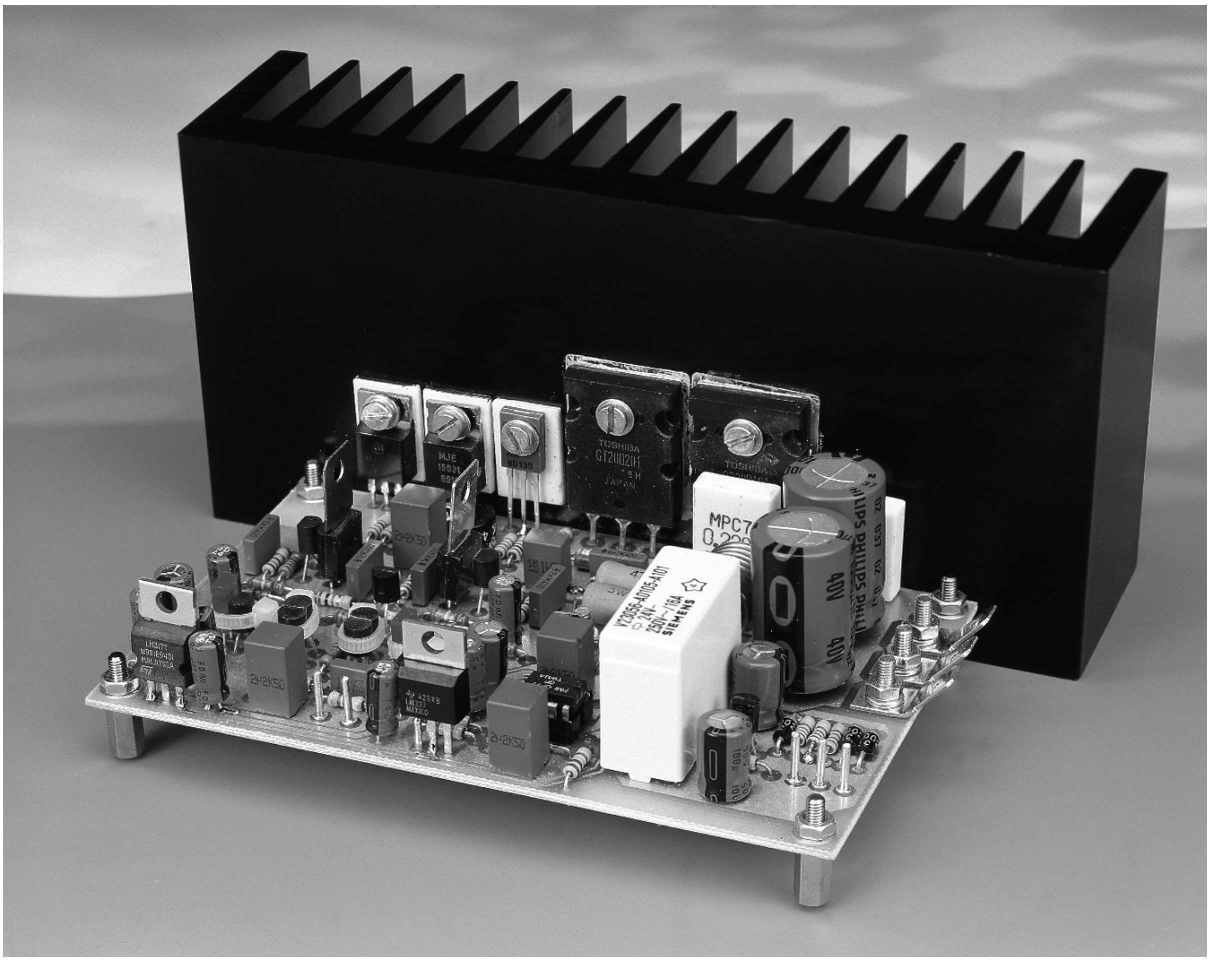

heat sink. They must, of course, be electrically isolated from each other with the aid of insulating washers and bushes for the fixing screws. In view of the required temperature stability, it is advisable to keep the terminals of T_{14} as long as possible to ensure that the device, when mounted, is at about the same height as the centre of T_{17}. This makes for good thermal coupling.

The other semiconductors and integrated circuits do not need extra cooling.

As mentioned earlier, good thermal coupling of a number of components (T_1-T_5; T_2-T_6; D_1-T_3; and D_2-T_4) in the input stages is also imperative for good performance. The easiest way of ensuring this is to secure the various pairs of components together with cable ties. In the case of D_1-T_3 and D_2-T_4 this is possible only if flat diodes are used. Tie the pairs together before fitting and soldering them on to the board. Mind the orientation of the diodes. It is advisable to mark the transistors with n-p-n or p-n-p, as the case may be, so that it is clear at a later stage which pair is which.

Inductor L_1 consists of eight turns of 1.5 mm dia. enamelled copper wire wound around a 9 mm drill bit. When it is wound, withdraw the drill bit, insert R_{39} into the coil and fit and solder the assembly on to the board at a height of a few millimetres above the surface.

All connections that carry large currents are output via PCB mounting flat sockets. The power connections are at the middle of one side of the board marked LS+ and LS–. The power lines to the relay are connected via ordinary

Figure 3. Photograph of the completed prototype board with heat sink. Two- and three-pin flat PCB mounting connectors are used for the loudspeaker and power line connections.

Figure 4. The power supply may be kept simple, but should have ample reserve power. Note that the diagram shows a design for a mono power amplifier.

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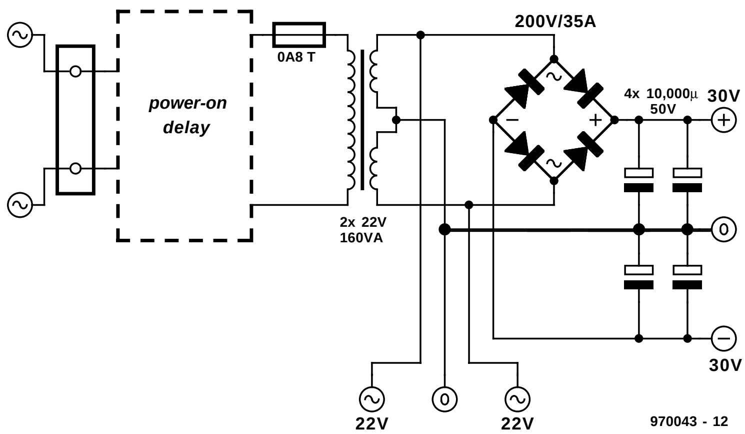


Suggested power supply (for mono amplifier):
Toroidal mains transformer, secondary 2×22 V, 160 VA
Bridge rectifier 200 V, 35 A
4 off electrolytic capacitor 10,000 µF, 50 V
Fuse holder and fuse 800 mA, slow

board terminals. Note that these lines consist of three wires: this means that the earth line must NOT linked to the amplifier earth.

The completed board with heat sink is shown in **Figure 3**. The heat sink must have a thermal resistance of 1.2 K W^{-1} or smaller.

POWER SUPPLY AND QUIESCENT CURRENT

When the board has been completed and thoroughly inspected for any faulty workmanship (including a one-by-one check of the components against the part list), a few operational tests need to be carried out. Naturally, a power supply is needed and a suitable arrangement for this is shown in **Figure 4**. This is a traditional setup of power-on delay, mains transformer, with a 2×22 V, 160 VA secondary, a 35 A bridge rectifier and four 10,000 μF electrolytic reservoir capacitors. This supply provides sufficient current even when 4 Ω loudspeakers are used. Note that the diagram is for a supply for a mono amplifier: two are needed for a stereo amplifier.

The diagram also shows how the power lines for the relay are obtained.

The power-on section is advisable but not obligatory. It serves to prevent large current peaks during switch-on.

As in all power supplies, good, firm, well-soldered connections are of prime importance. They ensure that there is no unnecessary resistance in the path of large currents.

Before the power is switched on, it is vital that preset P_1 on the power amplifier is turned completely anticlockwise. If this is not done, there is a risk that the quiescent current through the output transistors quickly reaches a very high level and the circuit is not designed to cope with this.

After the power has been switched on, check with a multimeter whether the potentials at the output of IC_1 and IC_2 are as indicated on the circuit diagram in Figure 1 (±23.2 V respectively). Then, measure the output voltage of the amplifier, which must be 0 V or very nearly so. If it is not, recheck the entire construction, particularly the input stages.

When up to this point all is well and the LEDs light, it may be assumed that the entire power amplifier is in good working order. It is then time to check all the potentials at the test points indicated in Figure 1.

Next, set the quiescent current with P_1. This is fairly high in this power amplifier: about 400 mA when a heat sink as specified is used; if a heat sink of ≤ 0.6 K W^{-1} is used, the current may even be as high as 500 mA. Connect a millivoltmeter set to its 200 or 250 mV range across R_{34} or R_{38} and turn P_1 slowly clockwise until the meter reads 88 mV (current = 400 mA) or 110 mV (current = 500 mA).

Leave the amplifier to warm up thoroughly and then readjust P_1 as required.

ASSEMBLY

The design of the printed-circuit board is intended for its use as a mono amplifier. From a quality point of view, this is the best way of building an amplifier. Use a compact enclosure, perhaps with integral heat sink, and assemble board and power supply in this to make a stand-alone mono amplifier. A stereo amplifier is obtained by building two of these mono units and simply stacking them.

Wiring up the assembly is straightforward. Use heavy-duty wire for the +ve, –ve and earth supply lines and link them to the appropriate terminals on the board. Use heavy-duty wire also for linking the loudspeaker terminals with the LS+ and LS– PCB mounting 4 mm sockets on the board.

Link the phono input socket at the rear of the enclosure to the two input terminals on the board via a short length of screened audio cable.

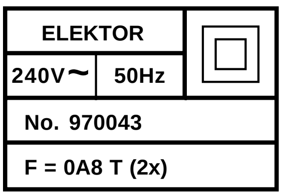
Next, link the two 22 V~ and 0 lines to the relay – remember that the 0 line must not be connected to the amplifier earth – see below.

The mains inlet should be of good quality, preferably with an integral fuse holder. The fuse rating is shown on the serial number label that should be affixed to the rear of the cabinet.

It is imperative that earth loops are avoided. This is best done by ensuring that there is link between the mains earth, the 0 supply line and the cabinet earth at only one point. Normally, and most conveniently this is at the signal input socket. If a non-insulated phono socket is used, this is automatically connected to the cabinet earth, so that no further action is needed.

It is, of course, possible to assemble two mono amplifiers using a single power supply in one cabinet, but this is not advisable. Even if it is desired to use only one cabinet, each of the mono amplifiers should have its own discrete power supply. The introductory photograph shows that this is how the prototype was constructed. The cabinet used for the prototype was provided with an integral heat sink and measured 445×75×305 mm (W×H×D).

[974003]



In passing ...

Most readers of this magazine know that it is published in identical form (at least as far as the articles are concerned) in Dutch, French and German, as well as in Greek, Polish, Portuguese, Spanish and Swedish in non-identical form. The Dutch, English, French and German issues are produced in the Netherlands, but published in the relevant country. These issues are the responsibility of an international team of writers/translators and technical editors.

On the face of it, there should be no problem in producing a supranational electronics magazine, since electronics is more international than any other branch of engineering. The electronics market is a world market: there is not one semiconductor or IC manufacturer who does not operate internationally. Also, norms and specifications tend to be internationally agreed.

Working in a multi-language team presents no real difficulties: most of the editorial staff have worked together for almost twenty years. Of course, technical translating is quite different from literary translating: for one, the technical translator has to know and understand the subject!

Most difficulties arise from the differences in standards laid down many years ago. For instance, an article on television will have to take into account that France (and eastern Germany) uses Secam, England PAL-I (but many other English-speaking countries NTSC), and Germany and the Netherlands PAL-B/G. The same applies to the different mains voltage outlets in the four countries. Astra transmissions are popular in Britain and in Germany, but not much in the Netherlands. Similarly, GSM mobile telephones for the E-band (1800 MHz) are in use in Britain and in Germany, but not in France and Holland.

Surprises still occur as well: chipboard or PVC tubing of the density specified in a German design are not available in the Netherlands. And Siemens products are not not easily available from retail outlets even in Germany. Strangest of all: try to find a bicycle rear light that is standard in all four countries!

[975043]

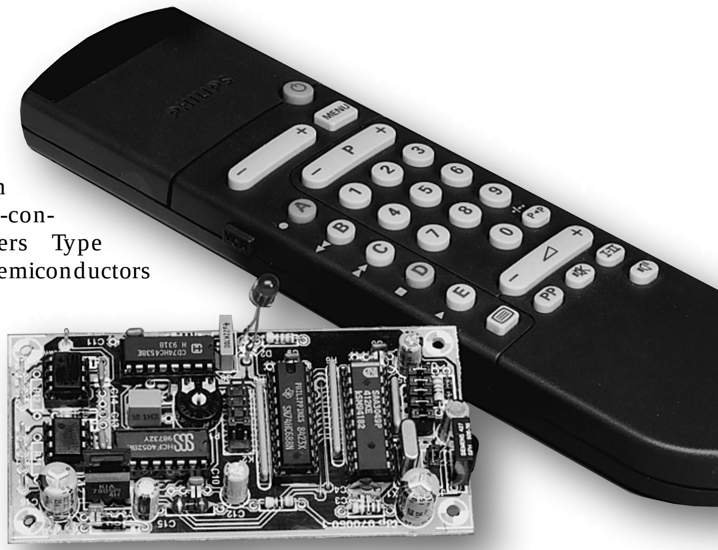
Contents

infra-red-controlled noiseless volume control

with Type DS1669 electronic potentiometer

The circuit described in this article is eminently suitable for those who appreciate comfort as well as sound quality. In the circuit two integrated electronic potentiometers are operated by an RC5 infra-red remote controller to provide a volume control that is not only free of crackles and other annoying noises, but is also free of wear and tear. It is intended to be built on a small printed-circuit board that can be conveniently built into almost any existing amplifier.

Design by T. Giesberts



ELECTRONIC POTENTIOMETER

The design is based on two electronically-controlled potentiometers Type DS1669 from Dallas Semiconductors that were described in the January 1997 issue (page 38) of this magazine.

Briefly, these potentiometers consist of a resistance track tapped at 64 positions separated in equal steps, a 64:1 multiplexer, control circuits, and an EEPROM. The devices are available in a DIP case (Type DS1669), or an SMD (SO8) case (Type DS1669S). Both types are available in one of three values: 10 kΩ, 50 kΩ, or 100 kΩ. The value is identified by adding the number 10, 50, or 100, as the case may be, to the type coding.

The 64 outputs of the resistance track are fed to the multiplexer, which determines which of the outputs is required; the relevant data is then stored in the EEPROM. This ensures that even when the supply to the device is switched off, the set value of resistance is retained.

The control circuits are driven by (1) an up/down switch connected between UC and earth (input DC is connected to the supply line), or (2)

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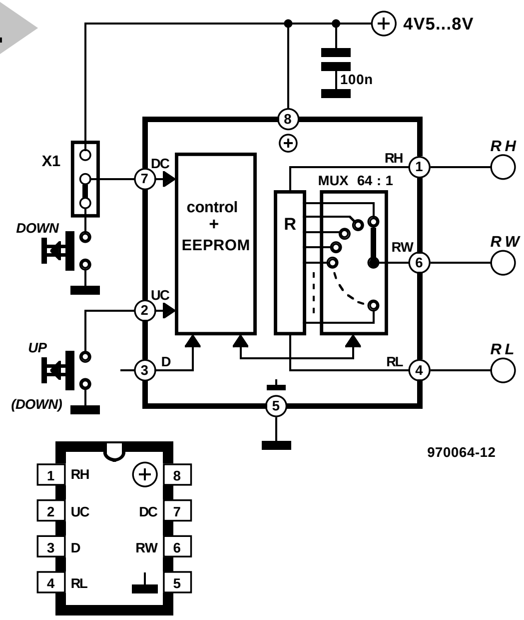

Figure 1. Block schematic of the Type DS1669 electronic potentiometer. The device is controlled by pulses provided by a microcontroller or generated manually.

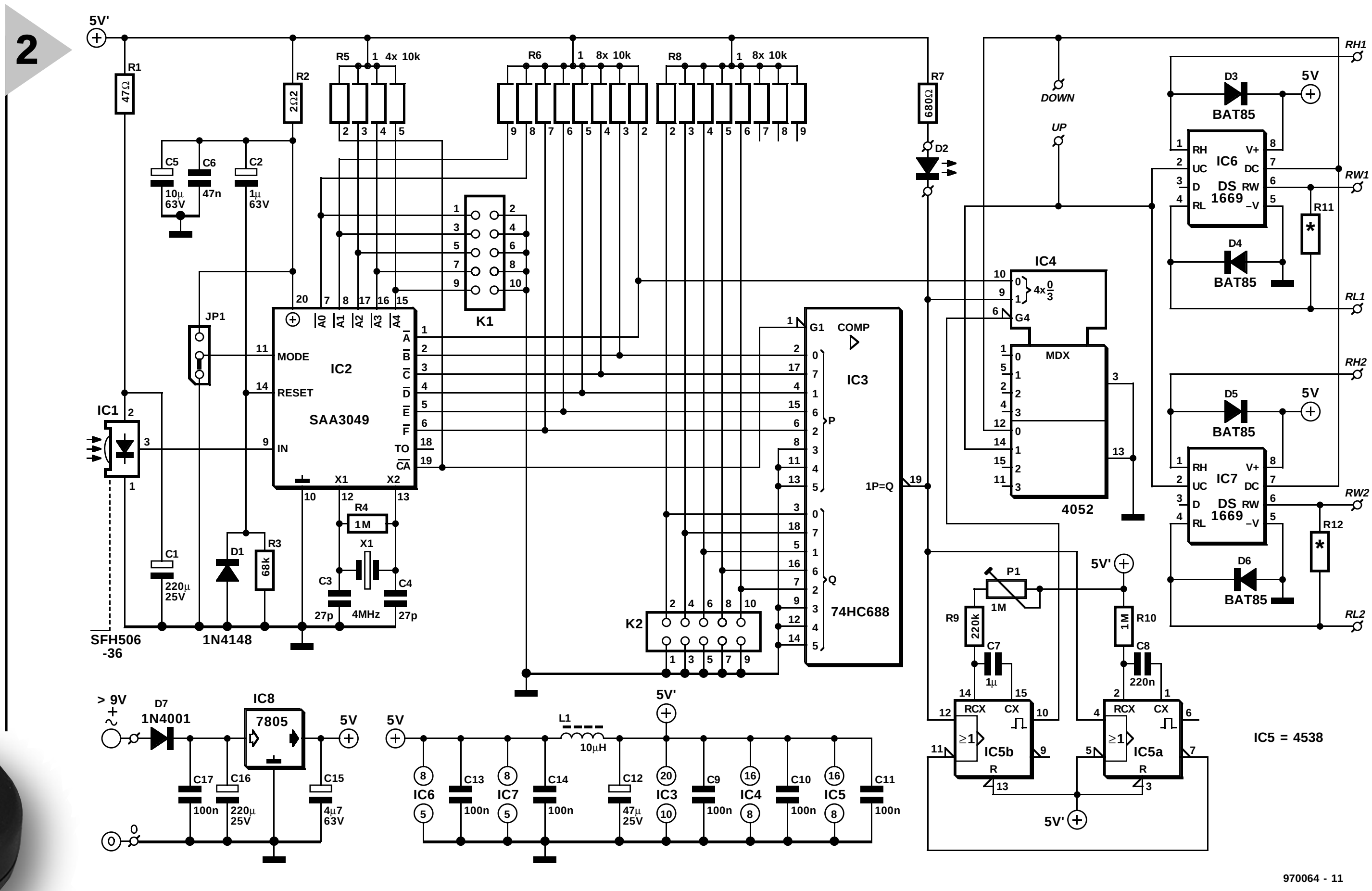

Figure 2. The circuit diagram of the volume control in which IC_6 and IC_7 are the actual potentiometers controlled by IC_4 and IC_5. The infra-red control signals are processed by IC_1–IC_3.

two down/up switches connected between DC and earth and UC and earth respectively, or (3) a microcontroller. The resistance track, *R*, is terminated into pins 4 (RL = *R* low) and 1 (RH = *R* high) (see **Figure 1**). In single-switch operation, when the switch is pressed, the imaginary wiper moves towards one end of the resistance track; when that is reached, it reverses direction and moves towards the other end. The same happens when the circuits are controlled by pulses from a microcontroller. In two-switch operation, when the 'down' switch is pressed, the imaginary wiper moves towards RL and when the 'up' switch is pressed, towards RH.

The D(igital) input intended for microcontrollers is internally debounced. For this purpose, the IC does not react immediately to a switch being pressed, but only after 1 ms. When a switch is pressed for less than 1 s, this is considered as one action. When the switch is held down longer than 1 s, the wiper is advanced one step every 100 ms; this is called the auto-repeat function. The wiper takes about 7 s to travel from one end of the track to the other.

The UC and DC inputs of the two potentiometers, IC_6 and IC_7 (see **Figure 2**) are switched by IC_4. In this, IC_{5a} and IC_{5b} perform a special function which will be reverted to later.

INFRA-RED CONTROL

The infra-red (IR) signal emitted by an RC5 controller contains two important data: the system address and the actual command. According to the RC5 code, the system address of a preamplifier is '16'. If this address is already occupied, or if the potentiometer is not fitted in a preamplifier but in another type of equipment, a different address may be used.

The address is set via inputs A_0–A_4. For address '16', inputs A_0–A_3 must be logic 0, which is arranged by short-circuiting pins 1 and 2, 3 and 4, 5 and 6, and 7 and 8, on K_1, and leaving 9 and 10 open. Address '0', reserved for television receivers, is set by short-circuiting all five sets of terminals. It is clear that the choice of key on the RC5 controller is determined by which of the pairs of terminals is short-circuited.

The signal from the RC5 controller is received by IC_1. This is a special device from Siemens which contains an IR photodiode and a complete receiver. The demodulated signal at its output, pin 3, is applied to decoder IC_2, which converts it into a digital signal. This signal is available as a logic level at outputs A–F. In the present circuit, only two commands are of interest:

address	F	E	D	C	B	A	command
16	0	1	0	0	0	0	louder
17	0	1	0	0	0	1	softer

Outputs B–F of IC_2 are linked to five of the P-inputs of digital comparator IC_3; the remaining three P-inputs are strapped to earth. The output, pin 19, of the comparator is low when the data word at the P-inputs is the same as that set at the Q-inputs via K_2. In that case, D_2 lights.

Output A of IC is purposely not linked to one of the comparator inputs, since the state of this bit (LSB) is different with commands 'louder' and 'softer' and might therefore upset the correct functioning of IC_3.

The receiver IC is decoupled by network R_1-C_1, and the decoder IC by network R_2-C_5-C_6.

Network R_3-C_2 provides a power-on reset for IC_2.

Jumper JP_1 enables decoder IC_2 to process RECS80 or RC5 codes. Normally, this will be the RC5 code, in which case the jumper must be between pin 11 and earth.

Crystal X_1, in conjunction with R_4, C_3 and C_4, ensures that the clock for IC_2 is correct.

Resistance arrays R_5, R_6, and R_8, function as pull-up resistors for the various inputs and outputs.

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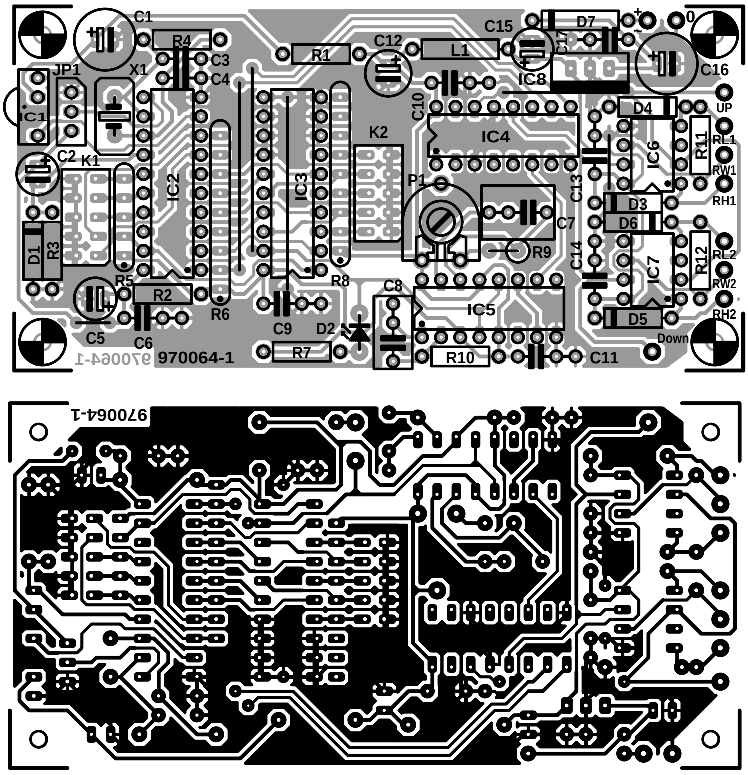


Figure 3. The printed-circuit board for the volume has been kept as small as possible to facilitate its being built into an existing equipment. Make sure that IC_1 is exposed to the outside world.

Parts list

Resistors:
R_1 = 47 Ω
R_2 = 2.2 Ω
R_3 = 68 kΩ
R_4, R_{10} = 1 MΩ
R_5 = array 4×10 kΩ
R_6 = array 8×10 kΩ
R_7 = 680 Ω
R_9 = 220 kΩ
R_{11}, R_{12} = optional, see text
P_1 = 1 MΩ preset potentiometer

Capacitors:
C_1, C_{16} = 220 µF, 25 V, radial
C_2 = 1 µF, 63 V, radial
C_3, C_4 = 27 pF
C_5 = 10 µF, 63 V, radial
C_6 = 0.047 µF, ceramic
C_7 = 1 µF, MKT (metallized polyester), pitch 5 mm or 7.5 mm
C_8 = 0.22 µF
C_9–C_{11}, C_{13}, C_{14}, C_{17} = 0.1 µF, ceramic
C_{12} = 47 µF, 25 V, radial
C_{15} = 4.7 µF, 63 V, radial

Inductors:
L_1 = 10 µH

Semiconductors:
D_1 = 1N4148
D_2 = LED, high efficiency
D_3–D_6 = BAT85
D_7 = 1N4001

Integrated circuits:
IC_1 = SFH506-36 (Siemens)
IC_2 = SAA3049P (Philips)
IC_3 = 74HC688
IC_4 = 4052
IC_5 = 4538
IC_6, IC_7 = DS1669 (Dallas Semiconductors)
IC_8 = 7805

Miscellaneous:
JP_1 = 3-way terminal strip with jumper
K_1, K_2 = dual 10-way terminal strip with 5 jumpers
X_1 = crystal, 4 MHz
PCB Order No. 970064 (see Readers Services toward the end of this issue)

SYNCHRONOUS CONTROL

It was stated earlier that the potentiometers are controlled by linking the UC and DC inputs to earth. This is effected by an analogue multiplexer, two of which are contained in IC_4. The use of a multiplexer makes it possible for two switches to be added if desired. For this purpose, pins ‘up’ and ‘down’ are provided on the printed-circuit board. The Type 4052 used for IC_4 has the added advantage of not needing additional logic circuits for the correct decoding of the rotary direction.

The address of the multiplexer is a combination of the decoded LSB (output A of IC_2) and the output of IC_3.

As mentioned earlier, when the UC or DC input of IC_6 or IC_7 is actuated longer than 1 s, their wiper is shifted automatically one step every 100 ms. In the present circuit, in which the ICs are controlled in parallel, the wipers may not always move in synchrony, and this is why the auto-repeat function is disabled. Instead, the ICs are driven by discrete pulse trains available at the CA output of IC_2. When the signal from the RC5 controller is sustained, a 15 ms wide pulse appears at the CA output every 120 ms. Both the width and the repetition frequency of these pulses are eminently suitable for repeated actuation of the potentiometers. More importantly, synchrony of operation is guaranteed.

DEAD TIME

The output of comparator IC_3 remains low as long as the CA output of IC_2 is active, which makes it difficult to ensure that the wipers of IC_6 and IC_7 do not move more than one defined step. Therefore, a dead time is provided by IC_5, which may be set between 0.22 s and 1.22 s with P_1. The arrangement is that the first wiper movement takes place immediately IC_3 outputs a pulse. Then follows the dead time, and then the pulse train.

The dead time is actuated by triggering monostable IC_{5b} at the leading edge of the first pulse output by IC_3. The Q-output of IC_{5b} is then high and prevents any further wiper movement by disabling the inhibit input of IC_4.

Since the Type 4538 used for the monostable is retriggerable, IC_{5a} has been added to prevent unwanted lengthening of the dead time. Both monostables are triggered simultaneously, so that IC_{5a} at once disables the trigger input of IC_{5b} by making pin 11 low. Since the mono time of IC_{5a} is longer than the repetition time of IC_3, any pulses repeated at the output of the comparator have no effect.

The length of the dead time is best determined empirically; normally, it will be sufficient to set P_1 to the centre of its travel.

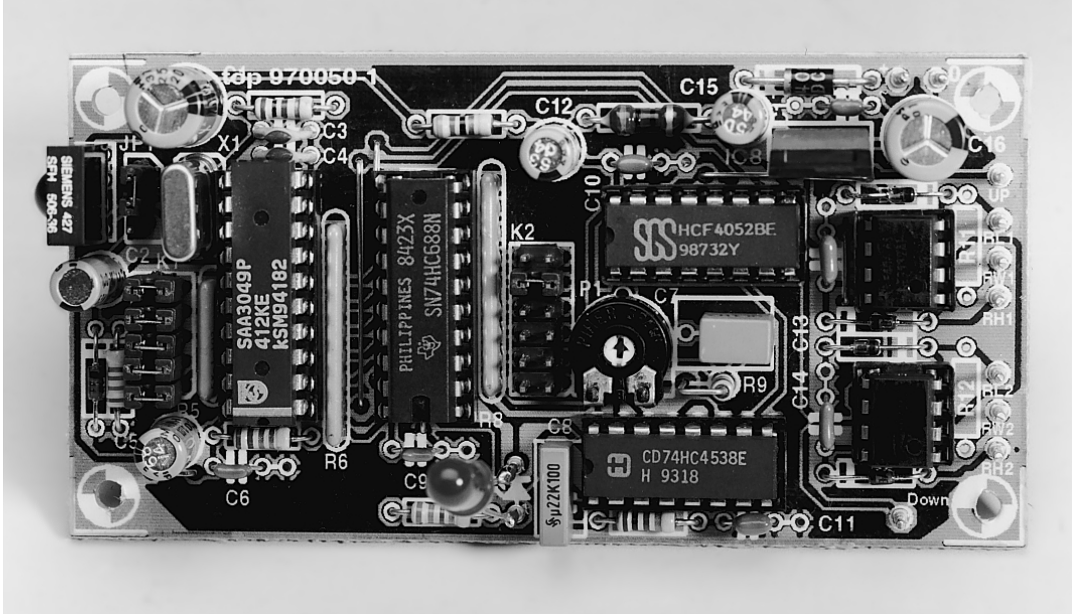
PRINTED-CIRCUIT BOARD

The circuit is best constructed on the printed-circuit board shown in **Figure 3**, which has been kept as small as possible since it is expected that most constructors will want to built it into an existing equipment to replace the volume control(s) in this. The connections to this control must be cut and linked to terminals RH, RW, and RL on the board.

FINALLY...

Although IC_6 and IC_7 are protected against high input voltages by diodes D_3–D_6, the applied input signal should not exceed 1.5 V r.m.s.

The linear characteristic of the potentiometers may be given a rather more logarithmic aspect by connecting



resistors R_{11} and R_{12} between the RW and RL terminals as shown in Figure 2. The value of these resistors should be $\frac{1}{4}$–$\frac{1}{16}$ of the value of the potentiometer. Be careful, however, not to damage the ICs, because the current flowing through the wiper must not exceed 1 mA. This means that the resistors should only be used with 100 kΩ potentiometers.

When the circuit is built into an existing equipment, it must, of course, be done in such a way that the IR receiver is exposed to the outside world. It may be necessary in some cases to extend the terminals of the IC with three short lengths of insulated circuit wire.

Since the circuit has its own rectifier (D_7) and regulator (IC_8), the input power supply may be any a.c. or d.c. source providing 9 V or more. This may be a mains adaptor, but it may also be possible to draw it from the equipment into which the volume control is to be fitted.

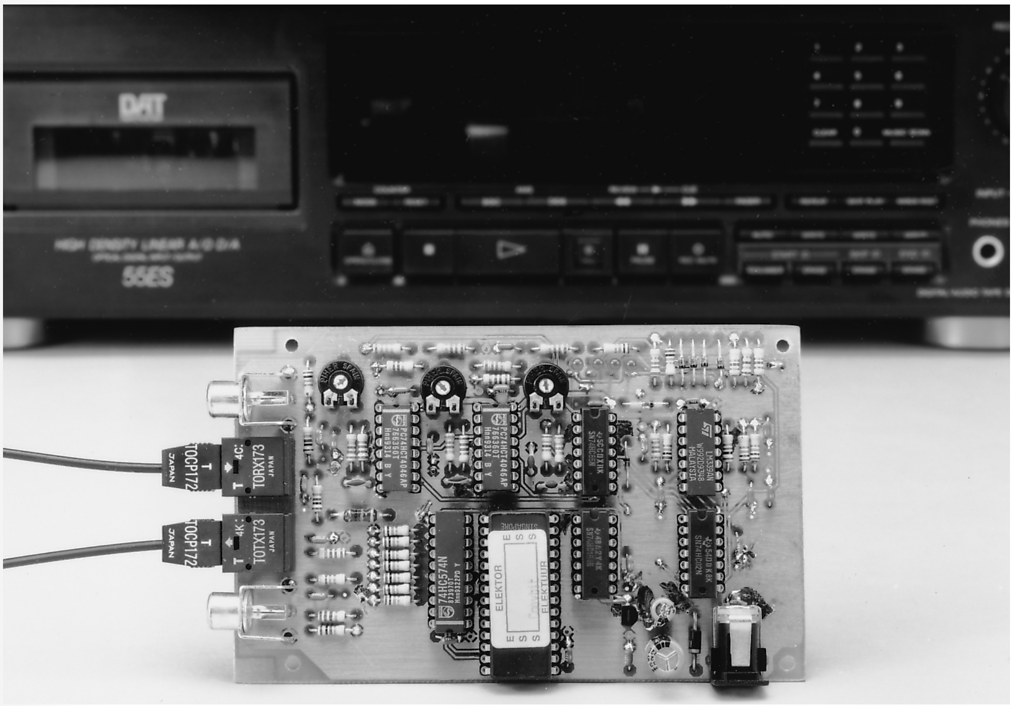
(970064)

Contents

simple copybit killer

enables limitless digital copying

The Serial Copy Management System (SCMS) prevents digital copying of audio material from the second generation onwards. Unfortunately, this protection also works on home recordings and so prevents home recordists from digitally copying their own musical work more than once without degradation by the SCMS. This problem has been highlighted before in this magazine, but it has not gone away – reason enough to describe a simple and inexpensive circuit for permanently removing the copy-prohibit-bit from the S/PDIF* audio signal.



Design by H. Hanft

GENERAL DESCRIPTION

The circuit described in this article is intended for the digital recording or copying and playback of non-commercial musical work. Such recording or copying is possible only when the copy-prohibit has been eliminated without any other effect on the audio signal.

The design of the circuit is such that there are no modifications required in the existing audio installation. The circuit is simply inserted in series with the digital (optical or coaxial) link between the relevant recording and playback equipment.

Briefly, the circuit:

- requires no modification to the digital audio equipment;
- is suitable for use with signals on optical as well as coaxial lines;
- uses readily available components;
- operates without programmable ICs such as PALs and EPLDs;
- is easily set up;
- has good clock regeneration through the use of PLLs (jitterkiller);
- gives clear indication (LEDs) of the sampling frequency (32 kHz, 44.1 kHz, or 48 kHz);
- automatically recognizes, and switches over to, the correct sampling frequency;
- draws a small current owing to the use of CMOS ICs.

CIRCUIT DESCRIPTION

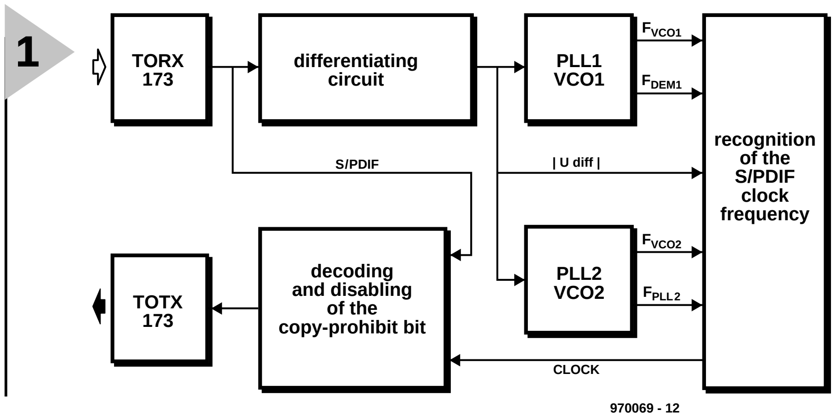
The block diagram of the copybit killer is shown in **Figure 1**. It shows that the copybit killer consists of:

- optical-to-electrical convertor for the S/PDIF signals;
- differentiating network;
- phase-locked loop (PLL) to regenerate the clock frequency;
- network for recognizing and processing the clock frequency;
- network for recognizing and dis-

** Sony/Philips Digital Interface Format – the consumer version of the AES/EBU standard. This standard was devised by the American Audio Engineering Society and the European Broadcasting Union to define the signal format, electrical characteristics, and connectors, to be used for digital interfaces between professional audio products.*

WARNING. *The information contained in this article is intended solely for the recording, processing, and copying, of private musical work. The Editor and Publishers disclaim all responsibility for its use that infringes any copyright vested in commercial compact disks and (digital) tape cassettes. Such infringement is entirely the responsibility of the perpetrator.*

Figure 1. Block diagram of the copybit killer. The core of the circuit, 'decoding and disabling of the copy-prohibit-bit', uses an EPROM.



abling the copy-prohibit-bit;
- electrical-to-optical convertor for the S/PDIF signals.

The circuit diagram proper is shown in **Figure 2**. The operation of the various networks outlined above is described in the following sections.

The **optical-to-electrical conversion of the S/PDIF signal** is carried out by IC_1, which is the well-known Type TORX173 integrated receiver. It converts the signal input from the fibre-glass cable into an electrical signal at TTL level. There is, of course, also a facility for inputting signals from a standard coaxial cable. This is possible via audio connector K_1 in parallel with the output of IC_1 via resistor R_2.

The **differentiating circuit**, consisting of XOR gates IC_{2a}–IC_{2c} and *RC* networks R_5-C_3 and R_6-C_4, serves to detect the rising or falling edges of the incoming S/PDIF signal. For each and every edge, a positive pulse of defined length is generated and used for synchronizing the following PLL.

The regeneration of the clock frequency contained in the S/PDIF signal is carried out by two discrete **phase-locked loops** (PLLs). The first one is for frequencies 6.144 MHz (sampling rate 48 kHz) and 5.6448 MHz (sampling frequency 44.1 kHz), and the second for frequency 4.096 MHz (sampling frequency 32 kHz)

So as to keep the circuit simple, both PLLs are Type 74HCT4046 ICs (IC_3 and IC_4). These circuits contain not only a phase comparator, but also a voltage-controlled oscillator (VCO). The PLLs circuits are virtually identical and differ only as far as the value of the resistor that sets the central frequency of the VCO is concerned (R_7 and R_{10}).

The network for **recognizing the clock frequency** serves to detect the sampling frequency of the S/PDIF signal and pass this on to the decoder. It consists of IC_{5a}–IC_{5d} and four NOR gates, IC_{6a}–IC_{6d}. The comparators recognize the incoming S/PDIF signal and ensure that the PLL locked in at that moment is included in the sigal processing. Also, when the sampling frequency is 48 kHz or 44.1 kHz, the VCO control

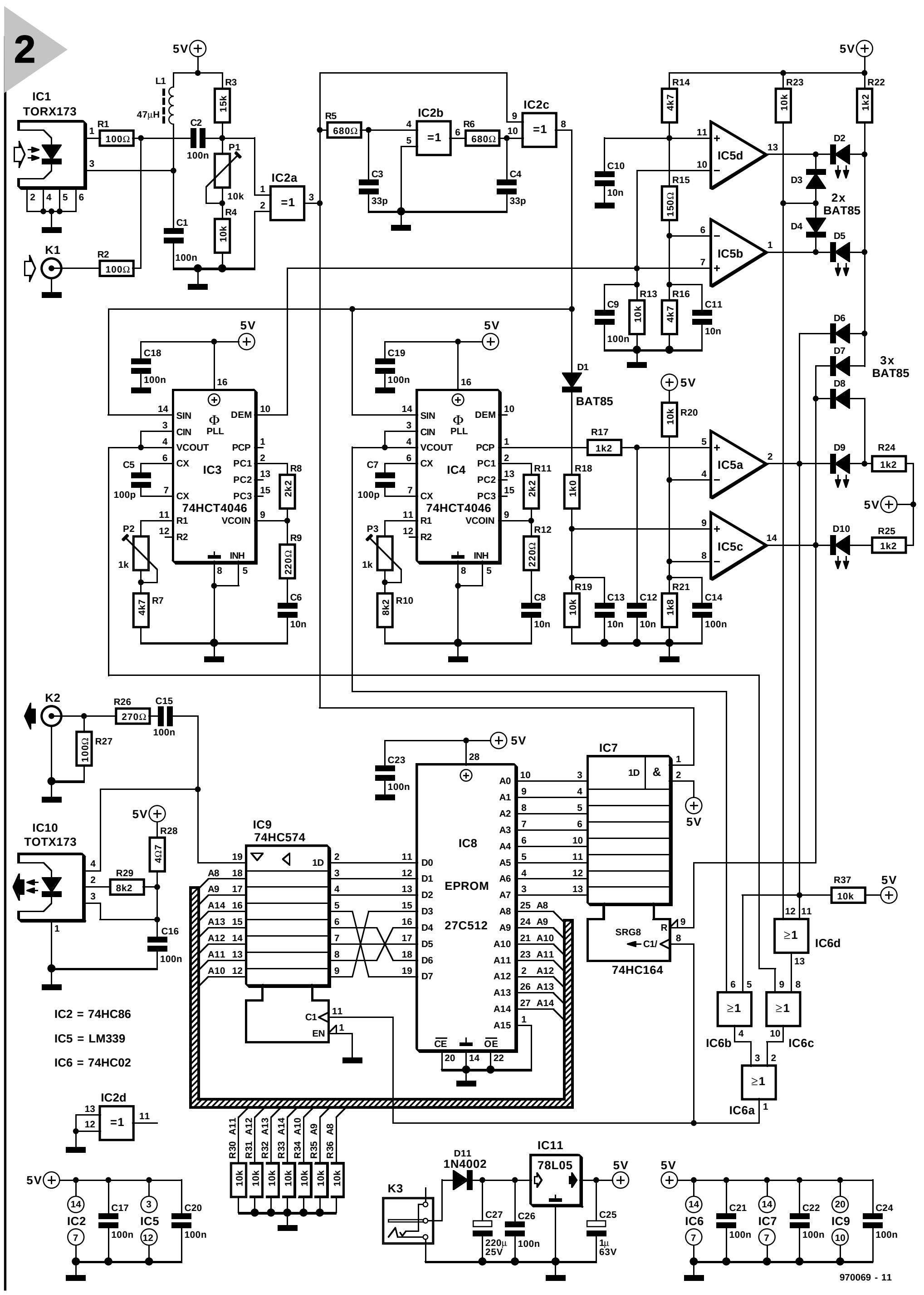


Figure 2. Circuit diagram of the copybit killer. All components are readily available and the circuit is easy to build and set up.

3

Parts list

Resistors:
R_1, R_{27} = 100 Ω
R_3 = 15 kΩ
R_4, R_{13}, R_{19}, R_{29}, R_{23}, R_{30}–R_{37} = 10 kΩ
R_5, R_6 = 680 Ω
R_7, R_{14}, R_{16} = 4.7 kΩ
R_8, R_{11} = 2.2 kΩ
R_9, R_{12} = 220 Ω
R_{10}, R_{29} = 8.2 kΩ
R_{15} = 150 Ω
R_{17}, R_{22}, R_{24}, R_{25} = 1.2 kΩ
R_{18} = 1.0 kΩ
R_{21} = 1.8 kΩ
R_{26} = 270 Ω
R_{28} = 4.7 Ω
P_1 = 10 kΩ preset potentiometer
P_2, P_3 = 1 kΩ preset potentiometer

Capacitors:
C_1, C_2, C_9, C_{14}–C_{24}, C_{26} = 0.1 µF, ceramic
C_3, C_4 = 33 pF
C_5, C_7 = 100 pF
C_6, C_8, C_{10}–C_{13} = 0.01 µF
C_{25} = 1 µF, 63 V, radial
C_{27} = 220 µF, 25 V, radial

Inductors:
L_1 = 47 µH

Semiconductors:
D_1, D_3, D_4, D_6–D_8 = BAT85
D_2, D_5, D_9, D_{10} = LED, high efficiency
D_{11} = 1N4002

Integrated circuits:
IC_1 = TORX173
IC_2 = 74HC86
IC_3, IC_4 = 74HCT4046
IC_5 = LM339
IC_6 = 74HC02
IC_7 = 74HC164
IC_8 = 27C512 (available ready programmed: Order no 976516* – see Readers Services towards the end of this issue)
IC_9 = 74HC574
IC_{10} = TOTX173
IC_{11} = 78L05

Miscellaneous:
K_1, K_2 = audio socket for PCB
K_3 = mains adaptor socket for PCB
PCB Order no. 970069* – see Readers Services towards the end of this issue

* These items may be purchased as a combination: Order no. 970069-C

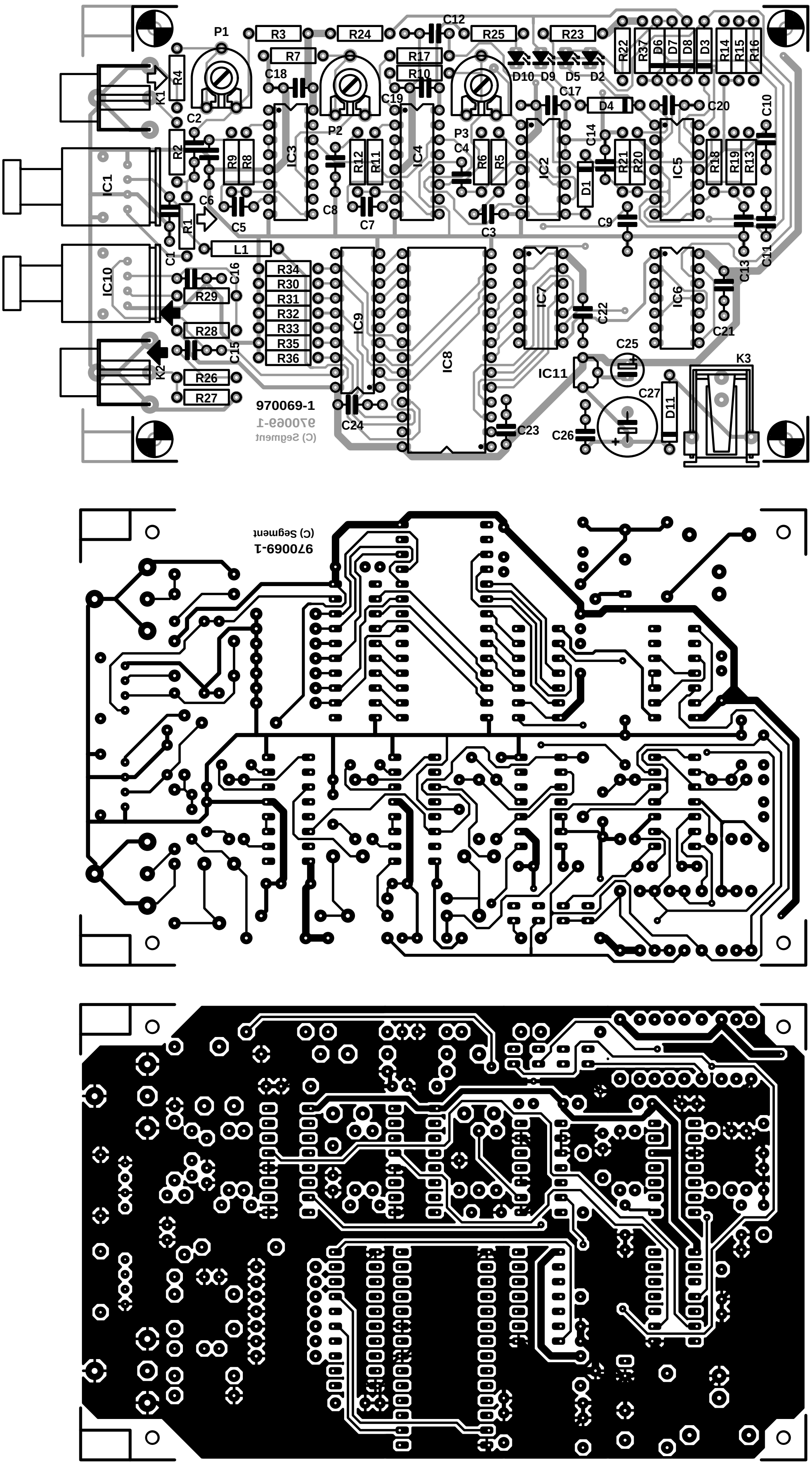
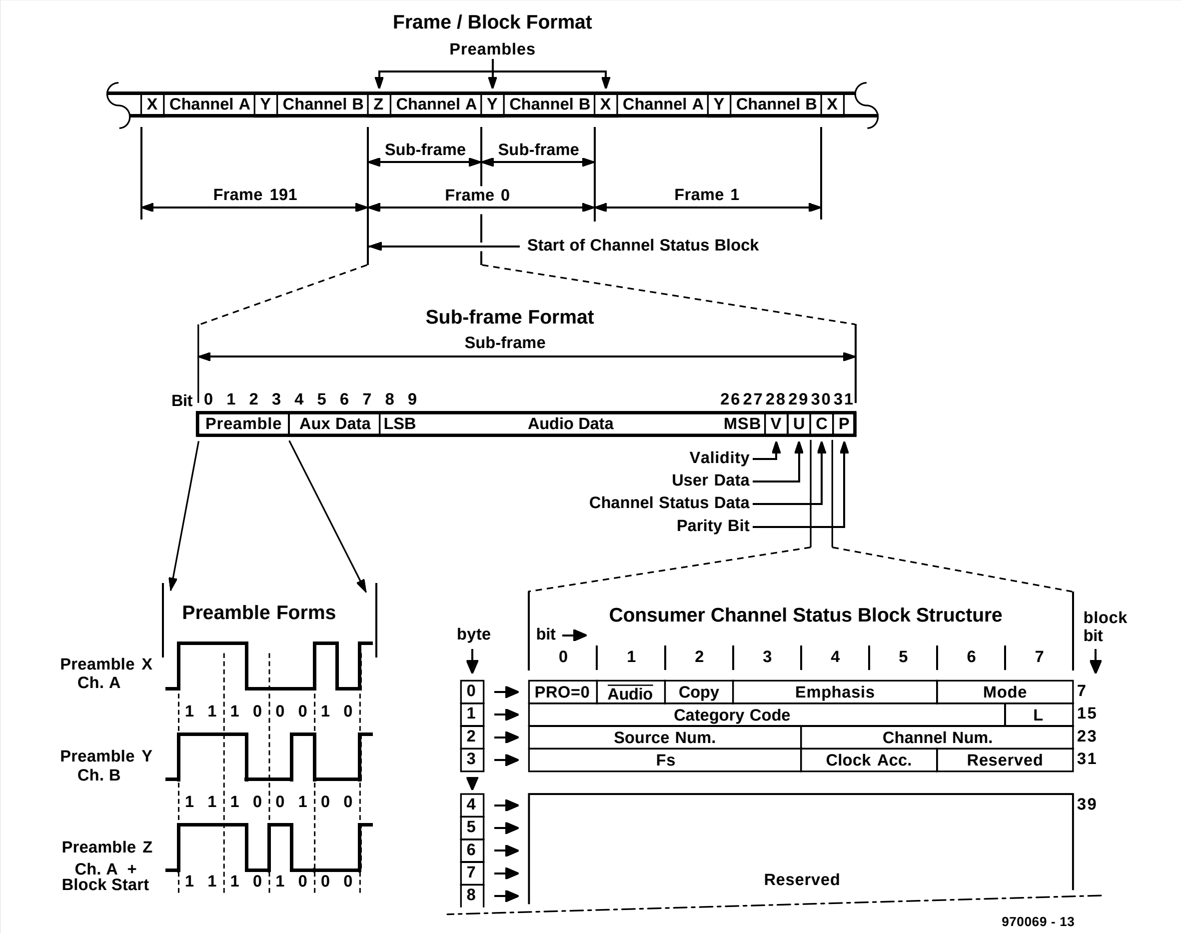


Figure 3. The printed-circuit board for the copybit killer is available ready-made.



Bit management

The digital part of the circuit, in conjunction with the associated programming code in the EPROM, provides a sequence control that recognizes and, if necessary, alters the status of the copy-prohibit-bit. The process attenuation of the S/PDIF signal is equal to one clock cycle. Shift register IC_7 continuously separates the last eight biphase-bit halves from the serial data stream and arranges that the halves are available at A_0–A_7.

The feedback of the data lines to the address lines via latch IC_9 divides the memory of the EPROM into 128 blocks of 256 bytes each. In this way it becomes possible within a block, depending on the status of address lines A_0–A_7, to select the same or another block which is then enabled at the rising edge of the next clock pulse. This arrangement provides processing control in up to 128 steps.

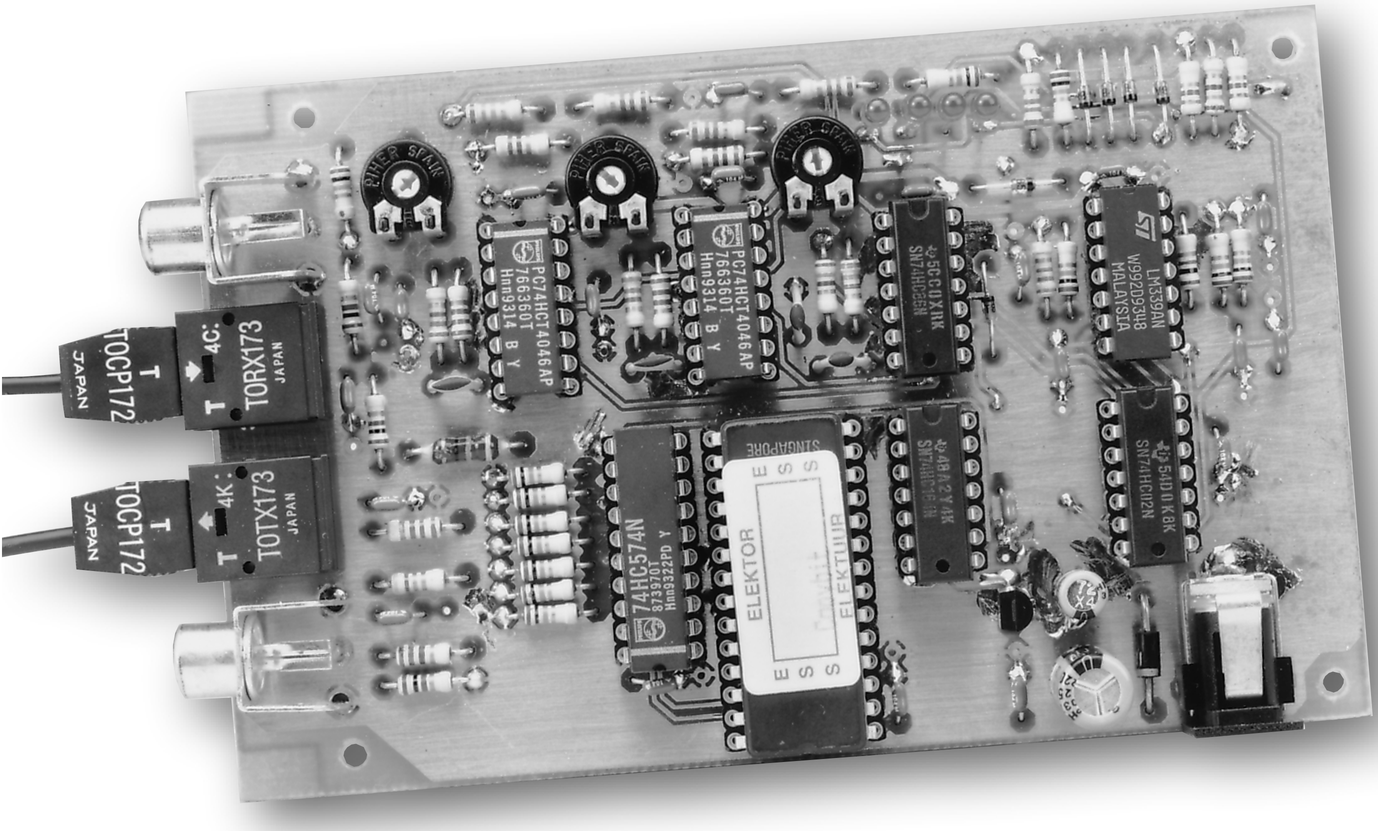
*From the format of the S/PDIF signal in the diagram above, it is seen that a transfer block consists of 192 frames, each of which is composed of two sub-frames. These sub-frames start with a preamble (**X, Y, Z**) and contain 32 bits. The preambles serve to mark the onset of a sub-frame; preamble **Z** also indicates the start of a new transfer block.*

Bit 2 of the Consumer Channel Status Block is of special importance for the present copybit killer, because unlimited digital copying can be carried out only when this bit is set. The Channel Status Bit is located at bit position 30 (biphase-bit-half positions 60 and 61) of a sub-frame and therefore occurs twice during each frame.

As far as the copy-prohibit-bit is concerned, these are the sub-frames of frame number 2.

In short, it is necessary that the status of bit number 30 in the two sub-frames of frame 2 be recognized and, if this bit has been erased, that it is set. If one of these bits is altered, the next parity bit (bit position 31) of the sub-frame must be inverted. This may give rise to eight different situations, which have to be taken into account in the programming code in the EPROM—see Table.

Bit Bi-Bit	User Data 29 58	 59	Channel Status Data 30 60	 61	Parity Bit 31 62	 63
Case 1 C=0, P=1	* ↓ *	* ↓ *	0 ↓ 0	0 ↓ 1	1 ↓ 0	0 ↓ 0
Case 2 C=0, P=0	* ↓ *	* ↓ *	0 ↓ 0	0 ↓ 1	1 ↓ 0	1 ↓ 1
Case 3 C=1, P=0	* ↓ *	* ↓ *	0 ↓ 0	1 ↓ 1	0 ↓ 0	0 ↓ 0
Case 4 C=1, P=1	* ↓ *	* ↓ *	0 ↓ 0	1 ↓ 1	0 ↓ 0	1 ↓ 1
Case 5 C=1, P=1	* ↓ *	* ↓ *	1 ↓ 1	0 ↓ 0	1 ↓ 1	0 ↓ 0
Case 6 C=1, P=0	* ↓ *	* ↓ *	1 ↓ 1	0 ↓ 0	1 ↓ 1	1 ↓ 1
Case 7 C=0, P=0	* ↓ *	* ↓ *	1 ↓ 1	1 ↓ 0	0 ↓ 1	0 ↓ 0
Case 8 C=0, P=1	* ↓ *	* ↓ *	1 ↓ 1	1 ↓ 0	0 ↓ 1	1 ↓ 0



voltage at the relevant PLL (IC_3) is measured by comparators IC_{5b} and IC_{5d}. At the same time, the appropriate sampling frequency is indicated by D_2 or D_5.

The core of the copybit killer is the circuit for **decoding and disabling the copy-prohibit-bit**. This digital circuit consists of edge-triggered 8-bit shift register IC_7, 32 kbyte EPROM IC_8, and edge-triggered 8-bit latch IC_9.

Address lines A_0–A_7 of the EPROM are controlled via the shift register, while latch IC_9 provides the feedback of databits D_1–D_7 to addresses A_8–A_{14}. This arrangement, in conjunction with the software in the EPROM, ensures that the copy-prohibit-bit is recognized and disabled. The modified S/PDIF signal is available at pin 19 of the latch.

It would have been possible to use a programmed controller for this part ofthe copybit killer, but the cicuit as is forms a less expensive, readily available alternative to an EPLD. Moreover, programming an EPROM with conventional means is straightforward, which is an advantage that must not be underestimated.

The data for the EPROM is provided by a small Pascal program that produces a binary file of 32768 bytes. Constructors need not concern themselves with this since the programmed EPROM is readily available through our Readers' Services.

The **electrical-to-optical conversion** of the (modified) S/PDIF signal is effected by IC_{10}, an integrated transmitter from the same stable as the integrated input receiver.

The TTL output signal from latch IC_9 is converted by IC_{10} into an equivalent optical signal that can be passed on via a standard fibre-glass cable. Just as at the input, IC_{10} is shunted by a coaxial audio connector, K_2.

CONSTRUCTION

The copybit killer is best built on the printed-circuit board in **Figure 3**. Populating the board is straightforward by consulting the circuit diagram and parts list as well as the board itself.

All components are standard items. As mentioned earlier, IC_8 is available ready-programmed through our Readers' Services.

The functions of the four LEDs are:

D_2 sampling frequency is 48 kHz;
D_5 sampling frequency is 44.1 kHz;
D_9 sampling frequency is 32 kHz;
D_{10} input signal is absent or poor.

When the board has been completed, compare it with the photograph of the prototype in **Figure 4**.

The copybit killer may be powered by a mains adaptor via K_3. The adaptor output must not exceed 9–10 V to avoid the dissipation limit of IC_{11} being exceeded. Since the output of some adaptors is already 9 V when it is set to 6 V, it is advisable to actually measure the output.

Figure 4. Illustration of the completed prototype board.

The circuit draws a current of about 80 mA.

SETTING UP

Th trigger level at the input of the copybit killer is set with P_1. This is best done with the aid of an oscilloscope by making the pulses at the output (pin 8) of differentiating network IC_8 coincide with one another. This setting, which gives the least jitter., may be checked with other signal sources and sampling frequencies so as to obtain a good average.

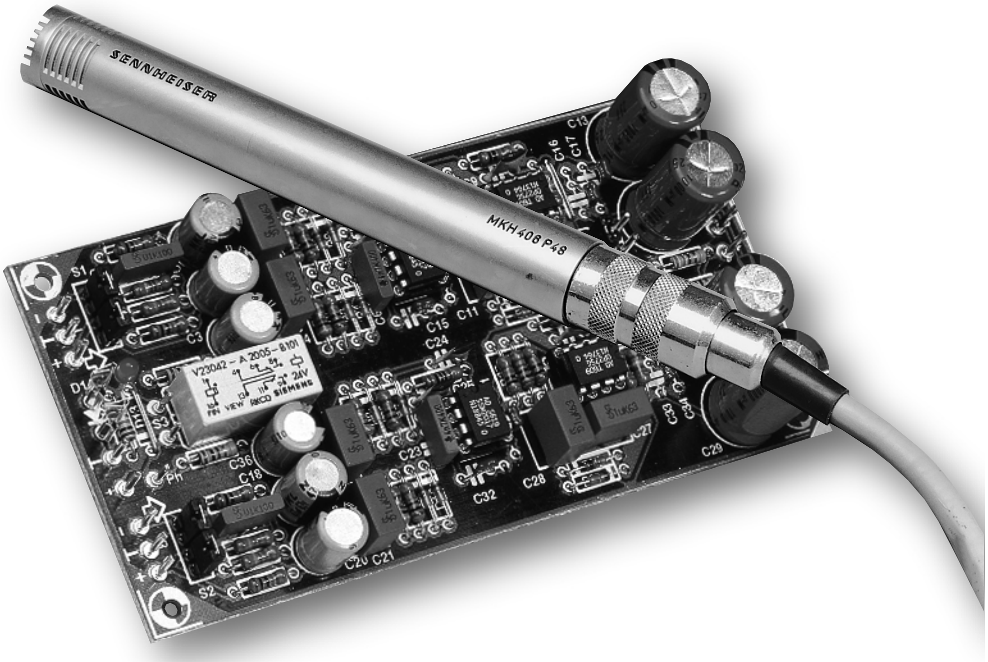
Setting the VCO (IC_3) with P_2 is a precise operation. It has to be ensured that the voltage variation of about 220 mV at the output (pin 10) of IC_3 is symmetrical with respect to the input voltage to IC_{5b} and IC_{5d} (a window of about 80 mV) when the sampling frequency is switched from 44.1 kHz to 48 kHz and vice versa. This measurement is best carried out with a digital voltmeter.

The setting of the second VCO with P_3 is not so critical. Make sure, however, that the lighting of the relevant LED accords with the sampling frequency in use.

[970069]

stereo microphone amplifier

with balanced inputs and phantom power supply



The magnification of very weak audio-frequency signals is and remains a precarious affair. That is why a microphone amplifier good enough for professional applications must have balanced inputs and indisputably good specifications. Foremost, of course, is the faithful transducing of sound. Although this is a subjective aspect, it is confidently expected that the design of the amplifier described here will satisfy even professional audio engineers and technicians. It is a stereo model with integral phantom power supply for electret microphones.

Design by J Brangé

INTRODUCTION

There are not all that many good methods to amplify the very small a.f. output signals of a microphone without affecting the quality of the input sound. One of the best is undoubtedly that in which a balanced instrument amplifier, discrete or integrated, is used.

Today, the accent in design is on integrated circuits, ICs, rather than on discrete transistors. One of the many a.f. ICs in the gamut of Analog Devices is the SSM2017, which is eminently suitable for use in a balanced microphone amplifier. This operational amplifier has a small noise factor and low total harmonic distortion, combined with a large bandwidth and high slew rate. Also, its amplification may be set within wide limits.

CIRCUIT DESCRIPTION

From the foregoing, it is clear that basing the design of the microphone amplifier on an SSM2017 is a sound decision. The circuit diagram in **Figure 1** may look extensive, but it should be borne in mind that it concerns a stereo circuit. This means that there are two of every device and component with the exception of those in the power supply. The following description will be limited to the left-hand channel.

The microphone is connected to the balanced input amplifier, IC_1, via terminals L+ and L–. The facility for short-circuiting R_1 and R_2 is provided for cases in which the circuit is to be used as a line amplifier. The switches are then open to provide an attenuation of 10 dB. When the circuit is used as microphone amplifier, the resistors are short-circuited. So, if use as a line amplifier is not envisaged, the switches and R_1, R_2 may be replaced by a wire bridge.

The supply voltage needed by electret microphones is provided via a phantom line. This means that the signal lines are used for carrying the supply voltage, which is, of course, not applied to the amplifier input. The volt-

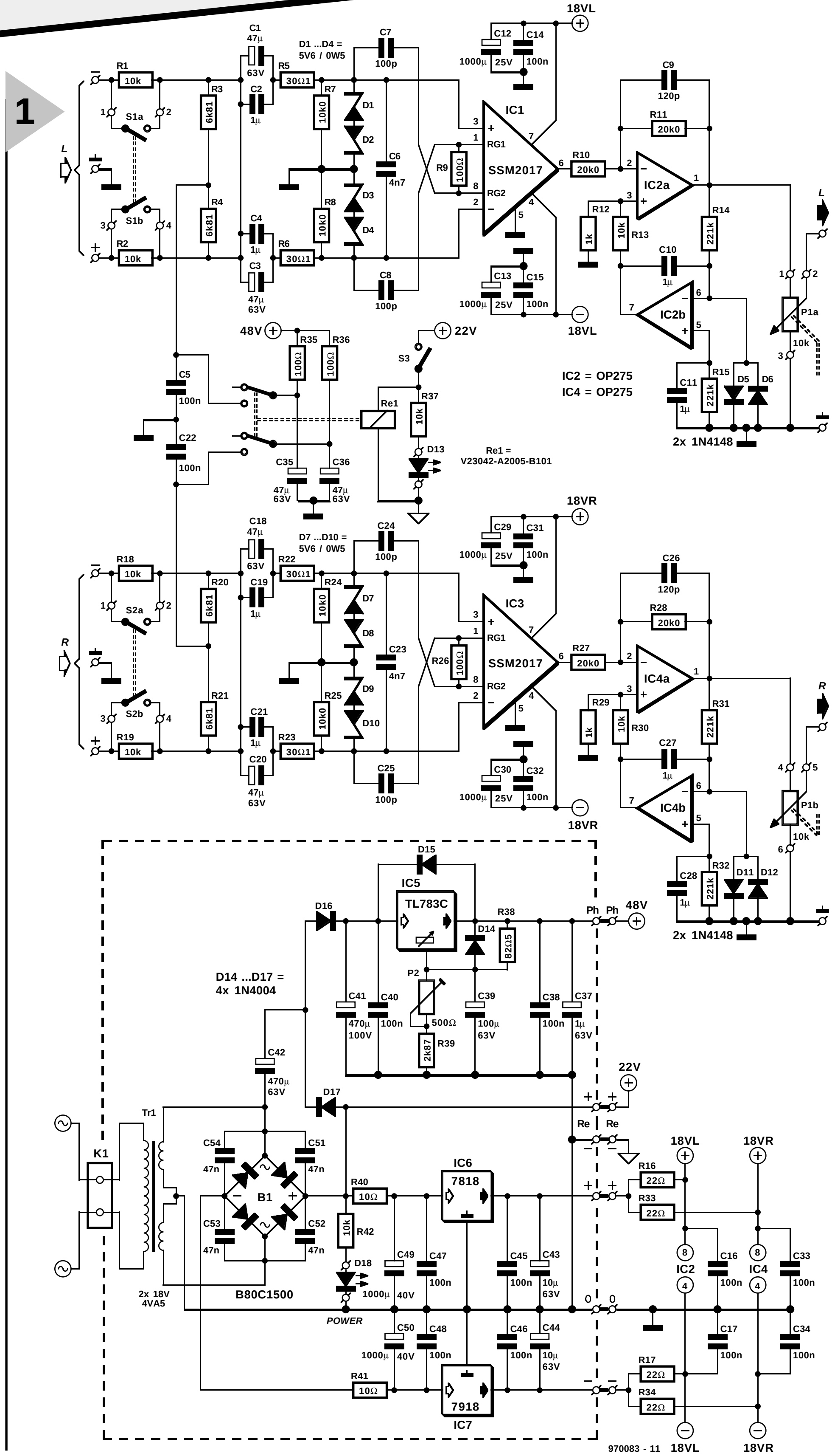
Features

(measured with an input signal of 2 mV into 50 Ω)

Signal-to-noise ratio		>74 dB (BW 22 kHz linear)
Total harmonic distortion (20 Hz–20 kHz)		<0.03% (BW 80 kHz)
Channel separation	(20 Hz)	120 dB
	(1 kHz)	95 dB
	(20 kHz)	83 dB
Bandwidth		> 60 kHz (source 50 Ω)
		> 37 kHz (source 600 Ω)

Figure 1. Circuit diagram of the stereo microphone amplifier, which is based on an SSM2017 operational amplifier from Analog Devices.

1



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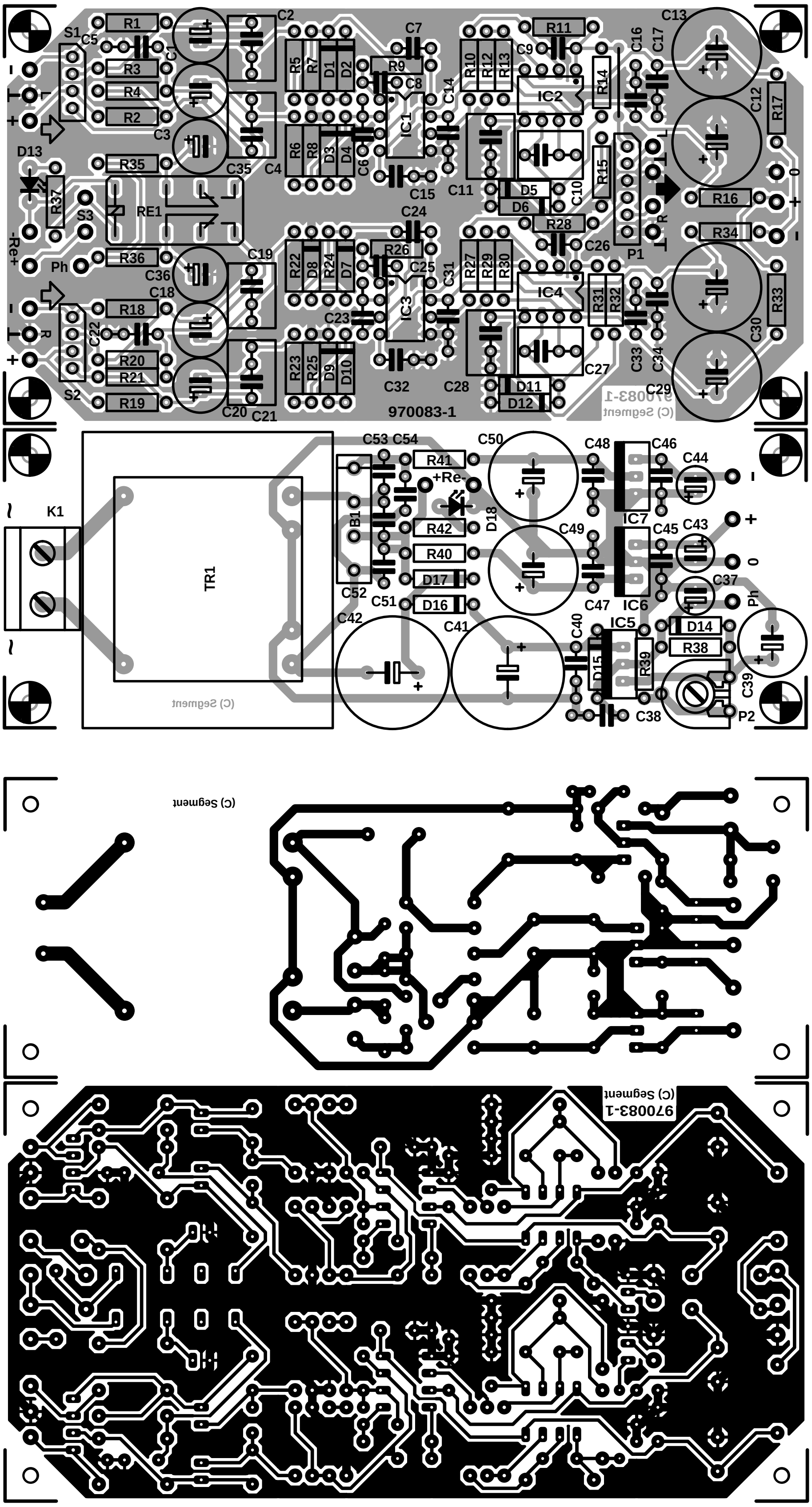


Figure 2. The printed-circuit board is in two parts to enable the amplifier and power sections to be kept isolated.

Parts list

Resistors:
R_1, R_2, R_7, R_8, R_{13}, R_{18}, R_{19}, R_{24}, R_{25}, R_{30} = 10.00 kΩ, 1%
R_3, R_4, R_{20}, R_{21} = 6.81 kΩ, 1%
R_5, R_6, R_{22}, R_{23} = 30.10 Ω, 1%
R_9, R_{26} = 100 Ω, 1%
R_{10}, R_{11}, R_{27}, R_{28} = 20.00 kΩ, 1%
R_{12}, R_{29} = 1.00 kΩ, 1%
R_{14}, R_{15}, R_{31}, R_{32} = 221 kΩ, 1%
R_{16}, R_{17}, R_{33}, R_{34} = 22 Ω
R_{35}, R_{36} = 100 Ω
R_{37}, R_{42} = 10 kΩ
R_{38} = 82.50 Ω, 1%
R_{39} = 2.87 kΩ, 1%
R_{40}, R_{41} = 10 Ω
P_1 = 10 kΩ, log, stereo potentiometer
P_2 = 500 Ω preset potentiometer

Capacitors:
C_1, C_3, C_{18}, C_{20}, C_{35}, C_{36} = 47 µF, 63 V, radial
C_2, C_4, C_{10}, C_{11}, C_{19}, C_{21}, C_{27}, C_{28} = 1 µF, metallized polyester (MKT), pitch 5 mm or 7.5 mm
C_5, C_{22} = 0.1 µF
C_6, C_{23} = 0.0047 µF
C_7, C_8, C_{24}, C_{25} = 0.001 µF
C_9, C_{26} = 0.12 µF
C_{12}, C_{13}, C_{29}, C_{30} = 1000 µF, 25 V, radial
C_{14}–C_{17}, C_{31}–C_{34}, C_{38}, C_{45}–C_{48} = 0.1 µF, ceramic
C_{37} = 1 µF, 63 V, radial
C_{39} = 100 µF, 63 V, radial
C_{40} = 0.1 µF, metallized polyester (MKT), 100 V
C_{41} = 470 µF, 100 V, radial
C_{42} = 470 µF, 63 V, radial
C_{43}, C_{44} = 10 µF, 63 V, radial
C_{49}, C_{50} = 1000 µF, 40 V, radial
C_{51}–C_{54} = 0.047 µF, ceramic

Semiconductors:
D_1–D_4, D_7–D_{10} = zener 5.6 V, 500 mW
D_5, D_6, D_{11}, D_{12} = 1N4148
D_{13} = LED, high efficiency, red
D_{14}–D_{17} = 1N4004
D_{18} = LED, high efficiency, green

Integrated circuits:
IC_1, IC_3 = SSM2017 (Analog Devices)
IC_2, IC_4 = OP275G (Analog Devices)
IC_5 = TL783C (Texas Instruments)
IC_6 = 7818
IC_7 = 7918

Miscellaneous:
K_1 = two-pin terminal block, pitch 7.5 mm
S_1, S_2 = double-pole on switch (or wire bridge – see text)
S_3 = single-pole on switch
B_1 = B80C1500 bridge rectifier
Re_1 = 22 V relay, 2 contacts
Tr_1 = mains transformer, 2×18 V secondaries, 4.5 VA
PCB Order no 970083-1 (see Readers services elsewhere in this issue)

3

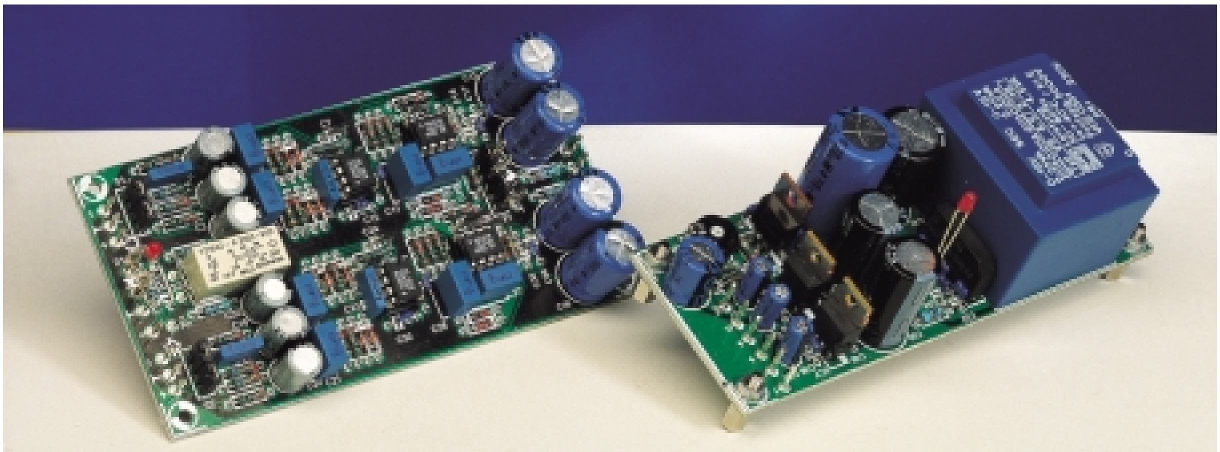


Figure 3. In the prototype, S_1, S_2 and P_1 have been omitted deliberately – see text.

age is 48 V and is linked to the input lines via relay Re_1, which is controlled by S_3. Network R_{35}-C_{35} provides additional smoothing of the supply line. The supply voltage is applied to the balanced microphone lines via potential divider R_3-R_4. RF decoupling is provided by C_5. The lighting of diode D_{13} indicates that the supply is on.

Capacitors C_1 and C_3 prevent any direct voltages from reaching the inputs of IC_1. They are bypassed for r.f. by C_2 and C_4.

Resistors R_5 and R_6, in conjunction with zener diodes D_1–D_4, suppress any peaks on the phantom supply lines that may be caused by the operation of S_3.

Resistors R_7 and R_8 form the input load of IC_1. As the pass-band of this IC is wide, it needs to be narrowed to reduce noise and distortion and also to largely suppress the effect of any r.f. radiation that may be present. The bandwidth is reduced by capacitors C_6–C_8, which must not be ceramic types.

One of the excellent features of the SSM2017 is that its voltage amplification is easily varied without affecting its input impedance and bandwidth. The amplification, α, is determined by the value of R_9 between pins 1 and 8. The relationship between these two quantities is given by

$$\alpha = (10^4/R_9) + 1.$$

With the value of 100 Ω specified for R_9 in the present circuit, this gives an amplification of ×100 (40 dB). For an amplification of ×31 (30 dB), the value of R_9 is 332 Ω, and for ×316 (50 dB), it is 31.6 Ω.

The output of IC_1 is applied to volume control P_{1A} via buffer amplifier IC_{2A}.This buffer is a special type of op amp, an OP275, which functions on the Butler amplifier principle. In this, a combination of bipolar and junction field-effect transistors is used to provide the low noise of the first and the speed and sound quality of the second. This arrangement results in a device with impressive specifications as regards noise contribution, distortion and slew rate.

The amplification of the buffer is set to unity with R_{10} and R_{11}.

The bandwidth of the buffer is limited by C_9.

To eliminate the fairly high offset (that is, imbalance) voltage produced by the SSM2017 (which may be as high as 200 mV, depending on the amplification), and since an electrolytic capacitor at the output was considered undesirable, integrator IC_{2B} is used. This op amp compares the potential at pin 1 of the buffer amplifier with that at its inverting input (pin 6). On the basis of this comparison, the integrator adjusts the input to the buffer in such a way that the overall offset voltage at the output is smaller than 1 mV at all times. Diodes D_5 and D_6 protect the integrator against high peak voltages.

POWER SUPPLY

The circuit needs three different supply voltages: a symmetrical one of ±18 V for IC_1–IC_4; a single one of 22 V for the relay; and a single one of 48 V for the microphone(s).

The ±18 V is provided by two voltage regulators, IC_6 and IC_7. The supply lines to the various ICs are bypassed for any noise and interference on the mains supply by resistors R_{16} and R_{17} and capacitors C_{12}–C_{17} (or R_{33} and R_{34}, and C_{29}–C_{34} in case of the right-hand channel).

The supply line for the relay is taken directly from the secondaries of the mains transformer since this need not be smoothed or regulated. Its level of 22 V is sufficient to drive the relay reliably.

Since the output of the mains transformer is not sufficient for directly generating the 48 V line for the microphone(s), a cascade network, D_{16}-D_{17}-C_{41}-C_{42} is used to boost the voltage across the bridge rectifier to about 70 V (open-circuit). This is reduced to +48 V by the combination of regulator IC_5 and preset P_2.

Diode D_{18} functions as on/off indicator.

CONSTRUCTION

It is obvious that in view of the very low a.f. input signal levels it is desirable to isolate the amplifier from the power supply, and this is why the printed-circuit board in **Figure 2** consists of two parts that should be separated before any construction work is commenced. In the final assembly, the two boards should be kept apart by at least 20 cm (8 in).

The construction itself is fairly straightforward, but a few points need to be borne in mind. Switches S_1 and S_2, if used, may be connected via 4-pin SIL headers.

In the same way, volume control P_1 may be connected via a 6-pin SIL header. Again, not everyone may deem this control necessary. When the amplification, and therefore the value of R_9 (and R_{26} in the right-hand channel), has been determined, the control is not really required, since the volume is set on the mixer panel or power amplifier. To enhance the signal quality it is then better to omit P_1 and, to protect the outputs, connect a 100 Ω resistor between pins 1 and 2 and 4 and 5 of the SIL connector.

It is advisable to house the power supply section in a well-insulated plastic (*Acrylonitrile-Butadiene-Styrene* – ABS) case.

The mains cable should be provided with a strain-relief.

The amplifier section is best housed in a metal case. The case should be connected to one of the earth connections on the amplifier board.

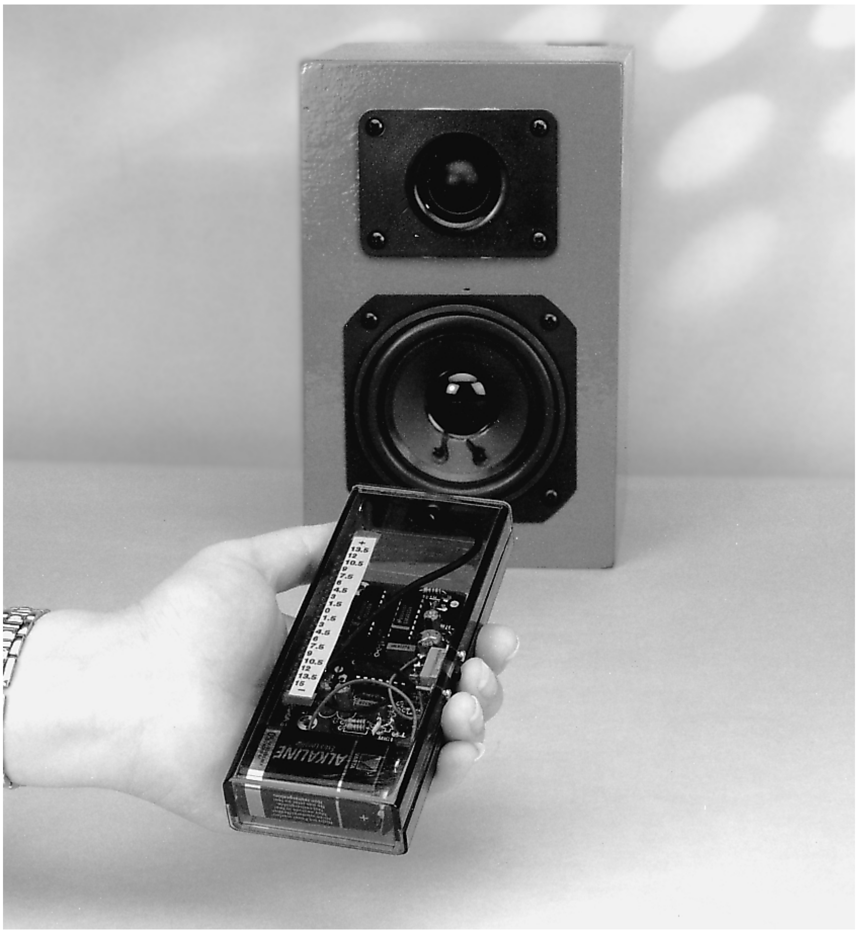
Finally, to keep any induced noise and interference to an absolute minimum, intertwine the three wires of the ±18 V line and earth, and the two wires of the +48 V line and earth.

[970083]

portable sound-pressure meter

sound analysis with LED indication

Few audio enthusiasts possess, or have access to, equipment required for accurately measuring the performance of a loudspeaker or the acoustics of a given hall or room. The unit described in this article is an instrument that does not give the performance of a professional meter but, in conjunction with a test CD, makes possible fairly accurate sound level measurements that enable the frequency response of an acoustic system to be ascertained. Moreover, its small size makes it a very handy unit to carry about.



Most home workshops have facilities for measuring current, voltage and resistance as well as other generated electrical signals with the aid of an oscilloscope and a function generator. However, equipment for measuring the frequency response of a loudspeaker, consisting of at least a sweep-oscillator system, a level recorder and a standard microphone, is in most cases not available.

Because of this lack, many amateurs would like a simple, inexpensive instrument with which the performance of a loudspeaker in a given hall or room can be assessed. Such an instrument, which, of course, cannot have the accuracy of a professional meter, is presented in this article. It consists of a simple-to-build sound pressure meter, equipped with a 20-LED display and has a resolution of 1.5 dB. In conjunction with a suitable test CD, it forms a very useful, compact and affordable miniature sound-level meter.

** In music and audio engineering, a third is a melodic and harmonic interval, taking three steps in a scale (major or minor) counting top and bottom notes. So, major third (C up to E), minor third (C up to Eb), and diminished third (C# up to Eb).*

Design by H. Baggen

1

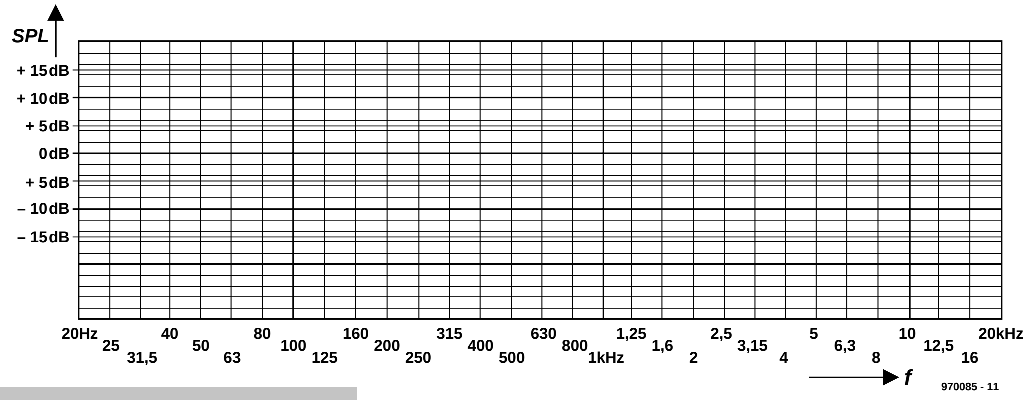


Figure 1. Model of a suitable graph paper to trace the measured frequency response of an acoustic system.

TEST SIGNALS

The circuit in **Figure 2** is, in essence, a fairly accurate sound level meter intended for carrying out relative, not absolute, measurements. Absolute sound levels are not of great import for determining the frequency response of a loudspeaker. What is of import is how the level of one sound compares with that of another, and that is a relative measurement.

In professional sound measurements, the sound source normally consists of an amplifier fed by a sweep-oscillator. Such a system is, however, not cheap and in the present circuit a much less expensive source, a test CD, is used. Most test CDs contain 30 discrete third-octave signals in the range 20 Hz to 20 kHz. When the output level produced by each of these signals is measured in identical conditions and plotted on graph paper as shown in **Figure 1**, a somewhat coarse, but nevertheless usable, frequency characteristic is obtained of the loudspeaker being tested.

It should be noted that the 30 signals available from the CD are recorded from a sweep-oscillator system generating third-octave warbled sine waves or noise (since, for measurement purposes, noise is much more to music). The frequency of a swept signal is not constant but swings between two values that are separated by 1/3 octave, that is, a third*. Third-octave noise is pink noise filtered to such a degree that only the frequencies at intervals of a third are retained.

Third-octave noise signals are used to eliminate the effects of the hall or room in which the loudspeaker is tested. Since many frequencies are generated simultaneously, the standard (test) microphone registers their mean level and this results in the averaging of the room (hall) resonances, which makes them less obtrusive.

CIRCUIT DESCRIPTION

Designing a circuit that picks up sounds and displays their relative level with reasonable accuracy is not very difficult – see Figure 2. This circuit consists of three sections: a microphone preamplifier (IC_{1a}), a full-wave rectifier

2

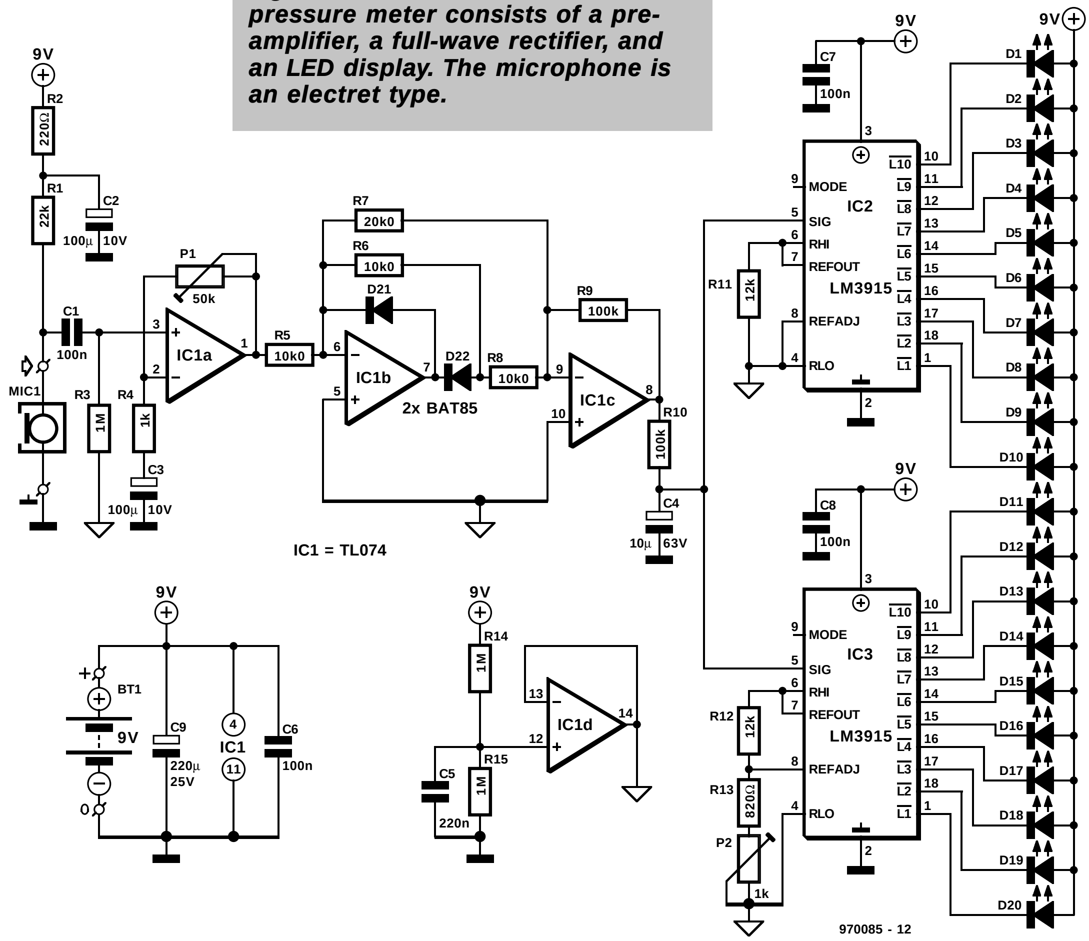
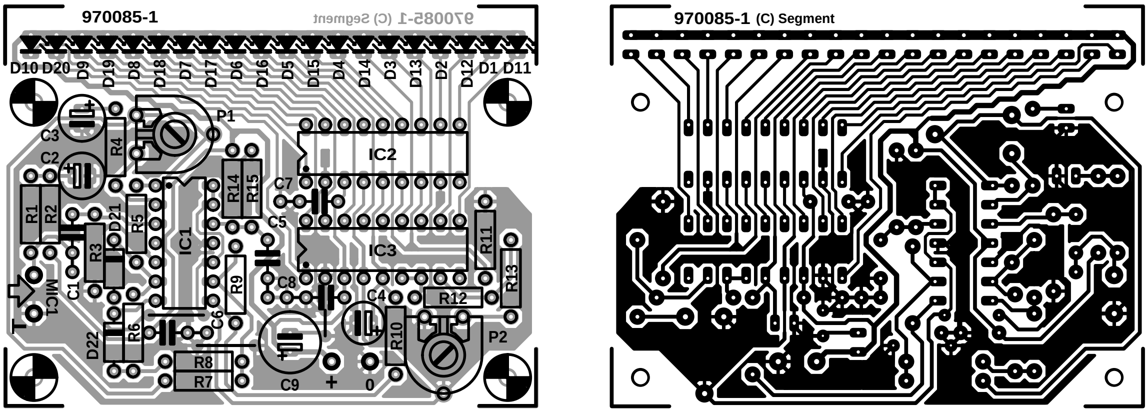


Figure 2.The circuit of the sound-pressure meter consists of a pre-amplifier, a full-wave rectifier, and an LED display. The microphone is an electret type.

3



Parts list

Resistors:
R_1 = 22 kΩ
R_2 = 220 Ω
R_3, R_{14}, R_{15} = 1 MΩ
R_4 = 1 kΩ
R_5, R_6, R_8 = 10.0 kΩ, 1% (but see text)
R_7 = 20.0 kΩ, 1% (but see text)
R_9, R_{10} = 100 kΩ
R_{11}, R_{12} = 12 kΩ
R_{13} = 820 Ω
P_1 = 47 kΩ (50 kΩ) preset
P_2 = 1 kΩ preset

Capacitors:
C_1, C_6, C_7, C_8 = 0.1 µF
C_2, C_3 = 100 µF, 10 V, radial
C_4 = 10 µF, 63 V, radial
C_5 = 0.22 µF
C_9 = 220 µF, 25 V, radial

Semiconductors:
D_1–D_{20} = LED, 3 mm, high efficiency
D_{21}, D_{22} = BAT85

Integrated circuits:
IC_1 = TL074CN
IC_2, IC_3 = LM3915N

Miscellaneous:
MIC_1 = electret microphone with rubber surround (e.g., MCE2000 from Monacor †)
BT_1 = 9 V battery with terminal clips
1 off single-pole on/off switch
Case: as desired - see text
PCB Order no 970085-1 (see Readers Services towards the end of this issue)

† Monacor
Inter-Mercador GmbH & Co, KG
Postfach 448 747
D-28286 Bremen
Germany
Telephone +49 421 78650
Fax +49 421 488415/488416

(IC_{1b} and IC_{1c}), and the LED display (IC_2, IC_3, and D_1–D_{20}). Op amp IC_{1d} is used for creating a virtual earth at half the supply voltage.

The microphone in this application must meet certain requirements, of course, even though it is used in a low-budget version of a sound level meter. It must, for instance, be fairly linear, otherwise the circuit cannot perform very well. It is therefore out of the question to use microphones of unknown origin with vague or uncertain properties. On the other hand, an expensive standard (test) microphone is a superfluous luxury. A good, linear electret microphone is a very good compromise between these extremes. The prototype instrument uses a type that is linear within ±2 dB over the

Figure 3. Printed-circuit board for the sound-pressure meter. Special care is needed when mounting the LED display.

4

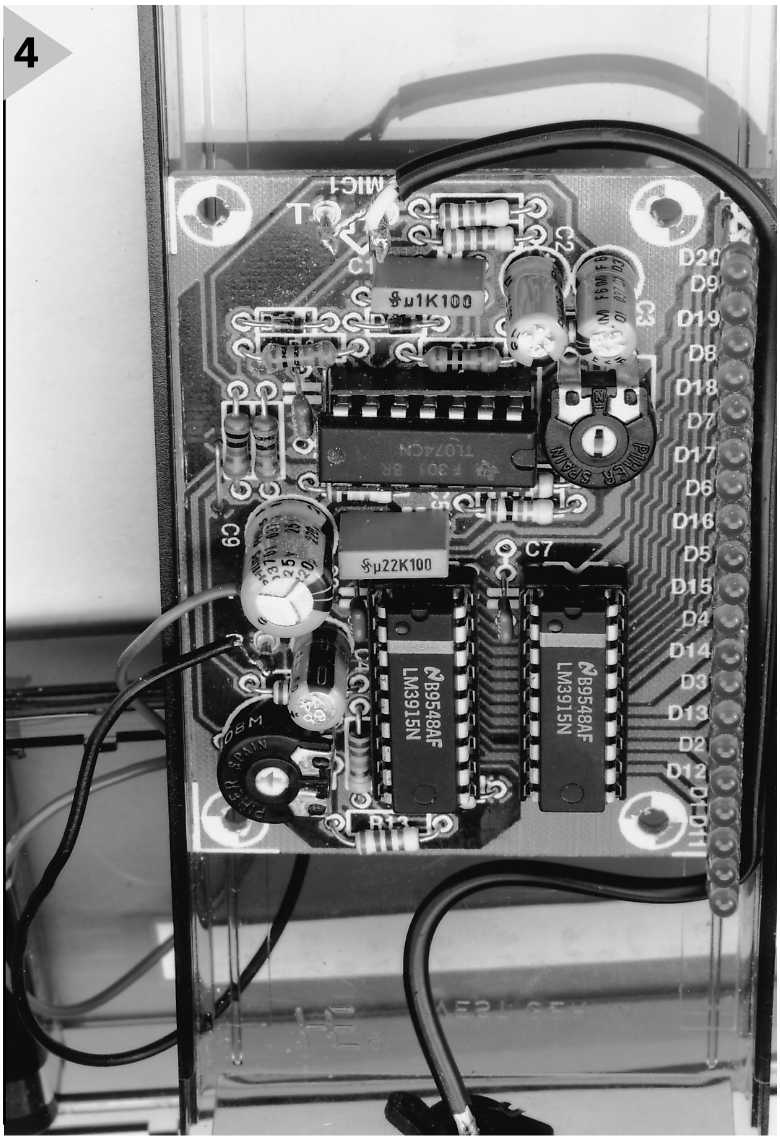


Figure 4. Inside view of the prototype sound-pressure meter in translucent case.

20 Hz to 20 kHz range.

The integral amplifier of microphone MIC_1 is held at about half the supply voltage with the aid of R_1. Network R_2-C_2 decouples the supply line.

The microphone signal is applied to preamplifier IC_{1a} via C_1. The cut-off frequencies of networks R_3-C_1 and R_4-C_3 are sufficiently low to result in a measurement error at 20 Hz of not more than 0.1 dB. The amplification, and thus the sensitivity of the microphone, is set with P_1.

The output of the preamplifier is applied to a conventional full-wave rectifier, IC_{1b} and IC_{1c}, So as to obtain a reasonable sensitivity without degrading the bandwidth of the instrument, IC_{1c} provides additional ×5 amplification. This results in an enhancement of the accuracy of the measurements at high frequencies.

To ensure a stable display, the rectified signal is differentiated by network R_{10}-C_4 (which constitutes a large time constant) before it is applied to the LED display.

The display is driven by the well-known Type LM3915 drive (IC_2). This IC contains a voltage reference source, a precise potential divider, and ten comparators, each of which can drive an LED directly. The level of the input voltage to the driver is displayed by the LED array in ten steps of 3 dB each.

Of course, 3 dB-steps are rather coarse and the resolution has, therefore, been enhanced by the addition of a second driver, IC_3, whose reference has been shifted by 1.5 dB. This is done by making the potential at the REFADJ(ust) pin (8) ×1.1885 higher than that at the corresponding pin of IC_2. This makes D_{11} the top of the decibel scale, followed by D_1, D_{12}, D_2, D_{13}, and so on. In other words, the LEDs driven by IC_2 and IC_3 are interlaced. This method has a slight drawback in that during measurements two LEDs light simultaneously: the correct test result lies somewhere between them. However, it was found that the operator quickly gets used to this.

POWER SUPPLY

Since the meter is intended for use as a portable instrument, the power supply must be battery-operated. The current drain is not greater than 19 mA, so that a 9-V battery will give about 100 hours service under normal conditions.

To ensure that op amps IC_{1a}–IC_{1c} remain within their common-mode range for as long as feasible, all three are powered by half the supply voltage. This is effected with the aid of a fourth op amp, IC_{1d}, and potential divider R_{14}-R_{15}, which is decoupled by C_5. The output of the op amp, pin14, constitutes a virtual earth at half the supply voltage above the real earth.

The reference pins of IC_2 and IC_3 are also connected to the virtual earth.

CONSTRUCTION

The instrument is best built on the printed-circuit board shown in **Figure 3**. The length of the board is determined by the dimensions of the display and the transparent case specified. It is, nevertheless, compact so that great care is required during soldering.

Potentiometer P_1 has intentionally been connected the wrong way around, that is, it has to be turned clockwise to reduce the amplification.

Note that although E96 type resistors are specified for R_5–R_8, E24 types may also be used if unavoidable. Their values should then be 11 kΩ instead of 10.0 kΩ or 22 kΩ instead of 20.0 kΩ as the case may be.

The prototype is housed in a translucent case, which has the advantage of not needing a cutout for the display. But, of course, any suitable or available case may be used, as long as the battery and finished board can be fitted comfortably inside.

Do not omit the rubber surround supplied with the microphone when fitting this on to the case. This surround damps spurious vibrations and makes the microphone less susceptible to reflections in or of the case.

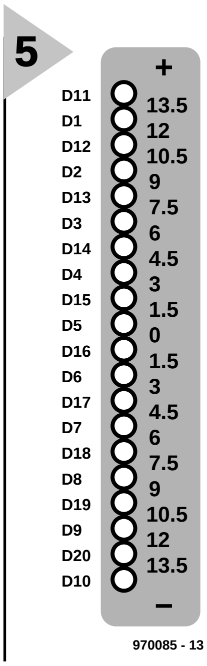
The display must, of course, be given a suitable scale of ±15 dB. The scale should have a 0 at its centre and this may be placed halfway between D_5 and D_{16} as shown in **Figure 5**. The other markings at 1.5 dB steps are placed accordingly.

SHIFTING THE REFERENCE OF IC_3

Preset P_1 is set according to the test CD used, and this will be discussed in the next section.

Preset P_2 sets the 1.5 dB shift of the reference voltage to IC_3, for which an accurate digital voltmeter is needed. Measure the potential, U_{REF1}, across pins 2 and 7 of IC_2 and then connect the voltmeter across pins 2 and 7 of IC_3. Adjust P_2 until a meter reading of 1.1885U_{REF1} is obtained.

Figure 5. Example of two usable scales for the sound-pressure meter. The total measuring range spans 30 dB.



USAGE

Usable test signals to obtain a frequency characteristic as described earlier may be obtained from the following test CDs: *The Test* (Stax, AXCD 92001); *Compact Test* (Pierre Verany, PV 784031); *Hi-fi Check* (Stereoplay); and *CD-2Check* (Monacor 30.0180).

When the frequency response is being measured, it is advisable to place the loudspeaker well away (≥ 1 metre) from reflecting walls or other objects. Set the volume of the audio system to a level where the test signals are well above the ambient noise level at all times. When a suitable level has been found, adjust P_1 on the sound-pressure meter to obtain a 0 reading on the LED display.

Bear in mind that at frequencies below about 200 Hz, effects of the room or hall are so strong that the measured levels say hardly anything about the performance of the loudspeaker being tested. This may be checked by holding the microphone right in front of the loudspeaker. It will be found that a number of peaks and troughs measured earlier (at the normal test distance of 1 metre) disappear.

An impression of the acoustic performance of the room or hall may be obtained by repeating the response measurement at a distance of 3–4 metres from the loudspeaker. At that distance, there is no question any longer of a flat response!

[970085]

the wall box

Brief parameters

Design	Two-way or three-way, vented
Woofer	WS13BF (Visaton)
Tweeter	DTW95NG (Visaton)
Wideband driver	FRWS5 (Visaton)
Frequency range	70 Hz – 20 kHz (–3 dB)
Crossover frequency	About 2 kHz
Music power	60 W
Nominal impedance	4 Ω
Volume	5 litres
Dimensions	300×418×138 mm (H×W×D)
Special aspects	Wideband unit provides sideways radiation

In these days of miniaturization, few people seem to feel a need for large loudspeakers in their living room. But, although they want small boxes, they still want good quality. The wall box presented in this article meets those requirements. It is a two-way system housed in a small, shallow enclosure that is inconspicuous in the living room.

Design by H. Baggen

There are many signs in the retail figures of the audio/hi-fi industry that the real hi-fi period of separates has peaked. These days, most customers specify a small or medium-sized system (mini towers are particularly popular right now) from which they expect exactly the same performance as from yesteryear's large installations. Loudspeakers are not excluded. Small, inconspicuous ones are wanted, but these must sound good and, more particularly, produce a good bass. Unfortunately, the unyielding laws of nature do not allow good bass reproduction from a small enclosure.

Nevertheless, if good drive units are used in a well-designed medium-sized box, good sound, small(ish) dimensions and affordability can go hand in hand, and this is what the present design is all about.

In the design, account was also taken of the fact that some people may use the present loudspeaker in a surround-sound system. Additional loudspeakers at right angles to the line connecting listener and main loudspeakers broaden the stereo reproduction.

The enclosure is intended primarily to be hung from a wall: it is tuned for that purpose.

DESIGN CONSIDERATIONS

As mentioned earlier, the design specification includes reasonable cost, good sound reproduction that fits in well with a modern (stereo) TV receiver or hi-fi system, and is not too conspicuous in the living room. The result is a medium-sized, shallow box intended for wall mounting.

The requirement of providing good bass response from a medium-sized,

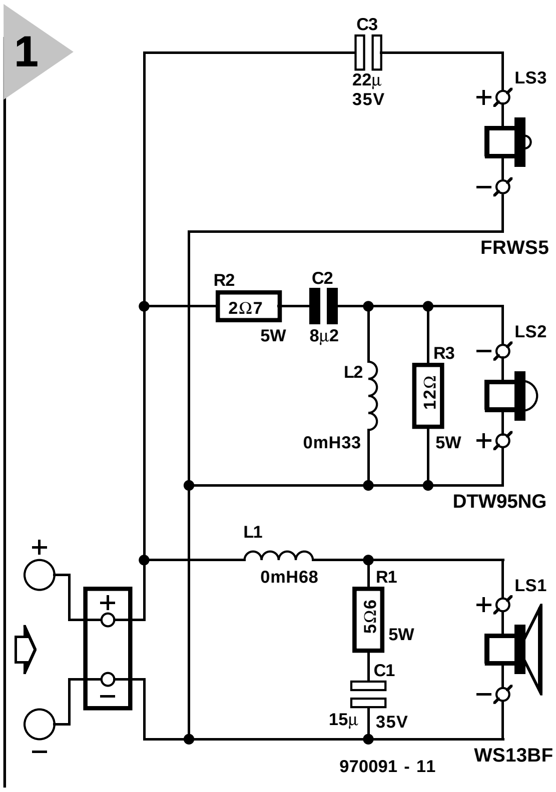


Figure 1. Circuit diagram of the crossover network for the wall box. The filter for the woofer and tweeter is a second-order one, while that for the wide-band unit is simply a capacitor.

not too expensive loudspeaker makes the choice of drive units highly critical. In the prototype, a 130 mm diameter bass/mid-range unit, a WS13BF, a 25 mm diameter tweeter, a DTW95NG, and a 50 mm diameter wideband unit, a FRWS5 are used. All three units come from the Visaton stable.

The Thiele/Small parameters of the WS13BF make it eminently suitable for use in a medium-sized box. It has a coated paper cone with a foam rubber surround and a 25 mm diameter voice coil.

The tweeter has a square mounting flange that goes well with the mid-range unit. It has an impregnated woven dome and heavily damped surround. The air gap is filled with magnetic ferrofluid, which provides good cooling and additional damping. Its frequency response is almost flat over the range 1.5–20 kHz.

The FRWS5 enhances the stereo effect. Strictly speaking, it may be omitted where this enhancement is not needed. However, in the prototype, it helped to produce a wide-angle stereo sound, which is particularly noticeable at the sides of the listening angle.

CROSSOVER NETWORK

The filter has also been kept as simple as possible without too many compromises. Its circuit diagram is shown in **Figure 1**. The network was designed with the aid of a computer simulation program that, based on the measured frequency and phase response of the drive units in the enclosure, calculates the overall response of the system including the crossover network. It also takes into account the positioning of the drive units and their radiation pattern.

The design of the network is a slightly modified second-order Butterworth filter with a few impedance corrections. This results in a second-order low-pass filter, L_1-C_1, for the woofer. Capacitor C_1 has another function as well: in conjunction with R_1 it provides the requisite impedance correction for the woofer, which is necessary for good filter performance.

The high-pass filter for the tweeter is formed by L_2-C_2, which, in conjunction with the response of the tweeter itself, results in a third-order rolloff. Resistors R_2 and R_3 lower the output of the tweeter by about 3 dB to match its output level to that of the woofer.

The cross-over frequency between woofer and tweeter is at about 2 kHz

The 'filter' for the wideband unit is simply a 22 µF bipolar electrolytic capacitor in parallel with the filters for the other two units. This cuts off the response of the wideband unit below about 800 Hz, which results in a satisfying additional spatial effect.

ENCLOSURE

The shallow box has a net volume of five litres and yet the 3 dB cut-off frequency is as low as about 68 Hz.

It will be seen from the introductory photograph and the construction plan in **Figure 3** that the woofer and tweeter are pointed slightly towards

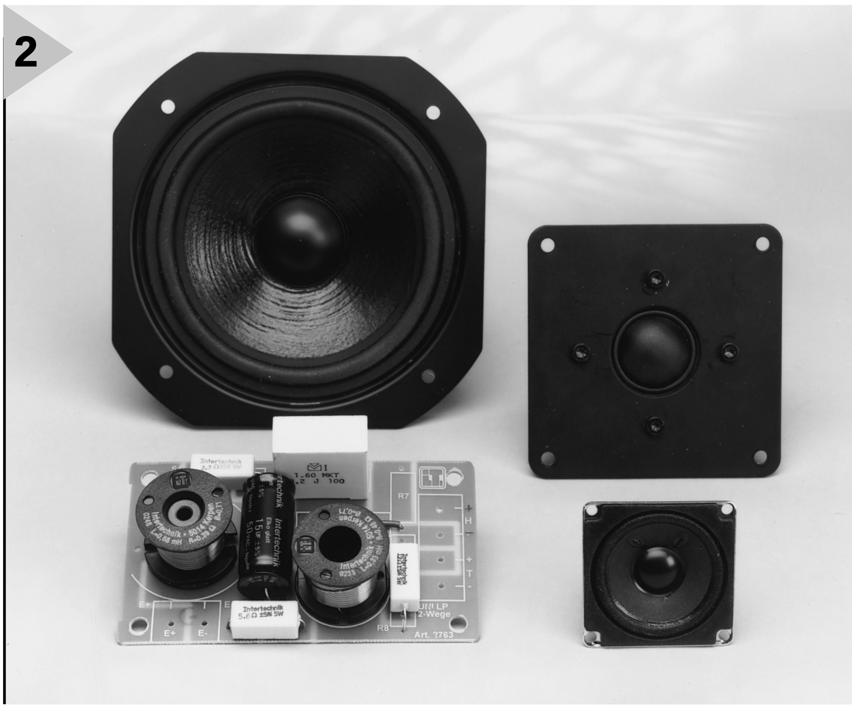


Figure 2. The drive units and crossover filter for the wall box (the electrolytic capacitor for the wideband unit is not shown).

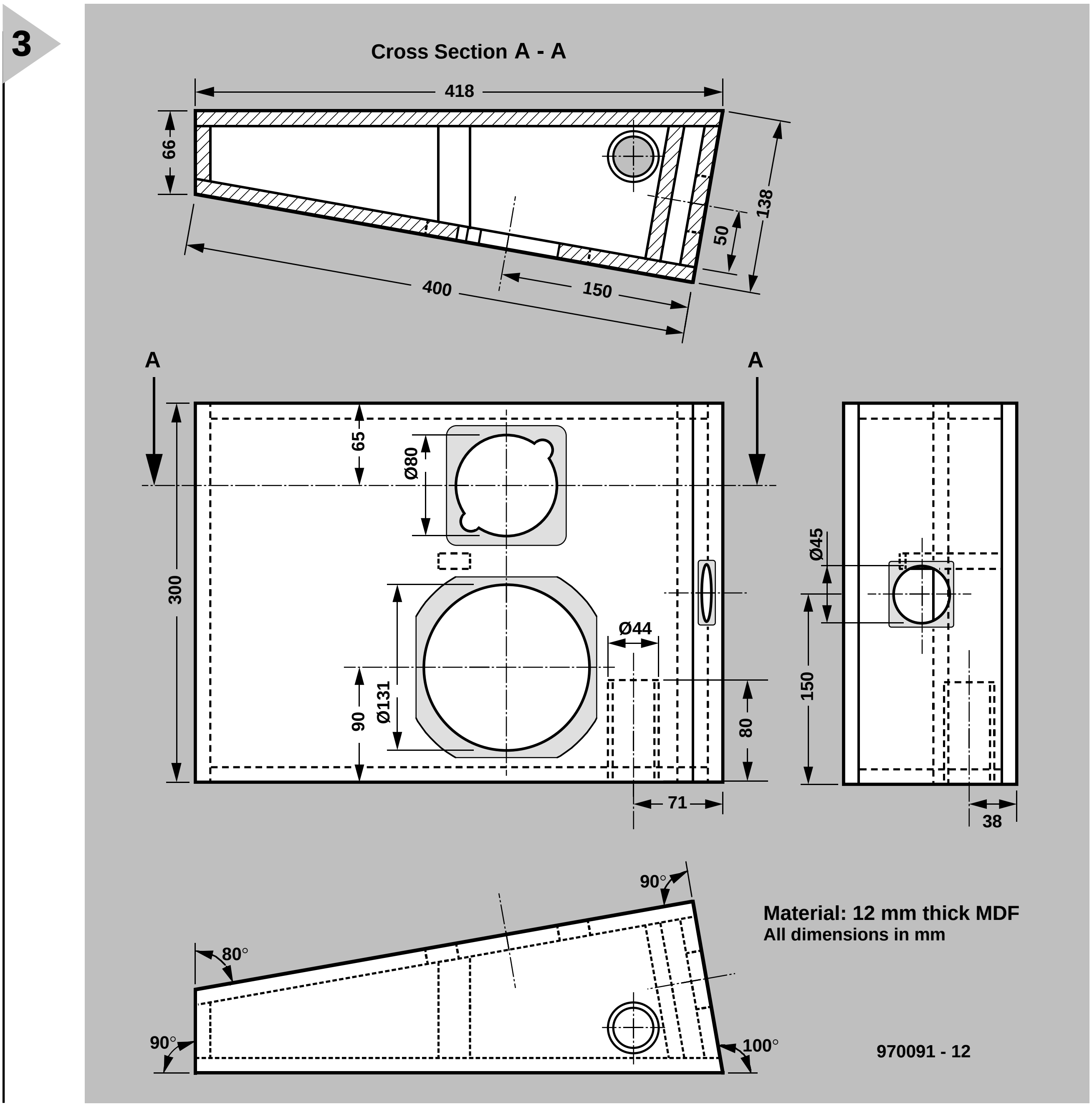


Figure 3. Construction plan for the box. Note that some of the panels are cut at 80° instead of the usual 90°.

the listener, whereas the wideband unit points slightly further forward. This makes it necessary for some of the constituent panels to be sawn at an angle of 80° instead of 90°. These angles make the enclosure look rather different from the usual rectangular box. Of course, if this angular construction is found too tedious, the box may be made rectangular with dimensions 300 × 160 × 160 mm.

Two holes are required in the front panel for the woofer and tweeter, and a single hole at the side for the wideband unit. The inside diameter of this latter hole should be enlarged with a suitable wood rasp or wood file to give the unit 'more air' at its rear.

A horizontal brace must be glued at the centre of the box between front and rear panel.

A partition at the side gives the wideband unit its own tiny chamber.

The bottom panel needs a suitable hole for the 44 mm diameter bass reflex port, which is 80 mm long. This port results in the the enclosure being tuned to about 50 Hz.

The crossover filter is best built on a piece of prototyping or similar board. Such boards can be bought from most specialist audio retailers. It is advisable to buy the components for the filters also from such a retailer, since electronics retailers normally do not stock them.

Fit the filter board in the box and feed the requisite connecting wires to the outside (those of the wideband unit through small holes in the partition). It is, of course, also possible to fit special (gold-plated) terminals to be underside of the box: this will look very pleasant.

Fit two small suspension plates at the top back of the box to enable it to be hung from the wall.

Fill the box with suitable wadding, wire up the drive units, and screw them into place. Mind the correct polarity, which is shown in the diagram (the tweeter should be in anti-phase with the woofer).

The box may be finished to personal taste and in accord with the decorations in the living room.

POSITIONING AND USAGE

The finished loudspeakers should be hung from a wall at ear-height (that is, at a height from the floor of about 1.6 metres). Their position should be

Parts list (per box)

Resistors:
R_1 = 5.6 Ω, 5 W
R_2 = 2.7 Ω, 5 W
R_3 = 12 Ω, 5 W

Capacitors:
C_1 = 15 µF, 35 V, bipolar electrolytic
C_2 = 8.2 µF, metallized polyester (MKT)
C_3 = 22 µF, 35 V, bipolar electrolytic

Inductors:
L_1 = 680 µH air-cored or suitably cored, $R_i \leq 0.5$ Ω
L_2 = 330 µH, air-cored, $R_i \leq 0.5$ Ω

Drive units:
LS_1 = WS13BF, 8 Ω (Visaton)
LS_2 = DTW95NG, 8 Ω (Visaton)
LS_3 = FRWS5, 8 Ω (Visaton)

Miscellaneous:
PVC pipe, 44 mm outer diameter, 80 mm long
Wadding, about 40×25 cm
2 off gold-plated loudspeaker terminals

Medium-density fibreboard (MDF) panels 12 mm thick:
1 off 418×300 mm
1 off 402×300 mm
1 off 420×300 mm
1 off 114×300 mm
1 off 107×300 mm
2 off 392×42×378×112 mm

4

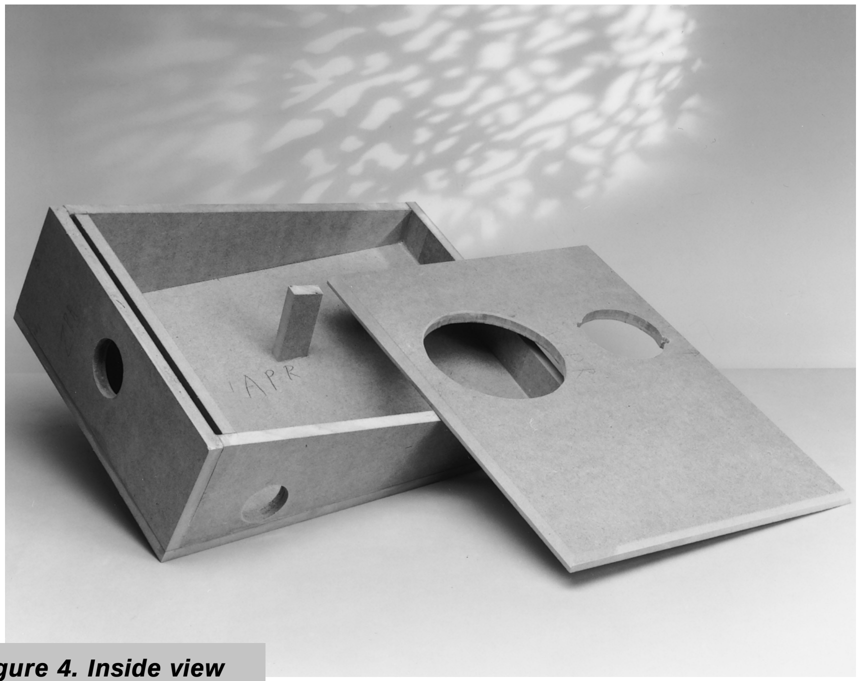
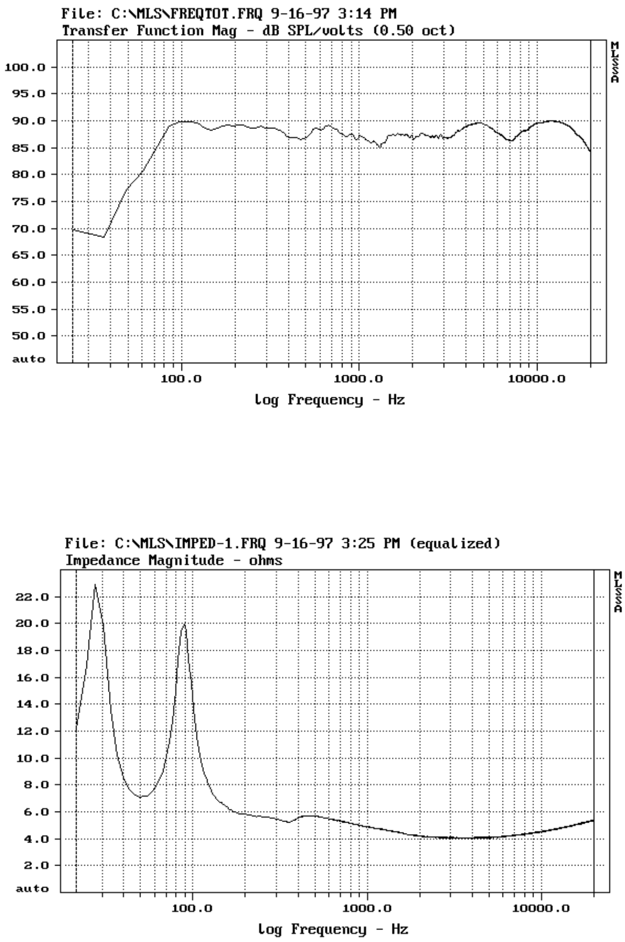


Figure 4. Inside view of the box before the top panel is put in place.

Figure 5. Frequency response curve and impedance curve of the wall box, incl. the wideband unit.

5



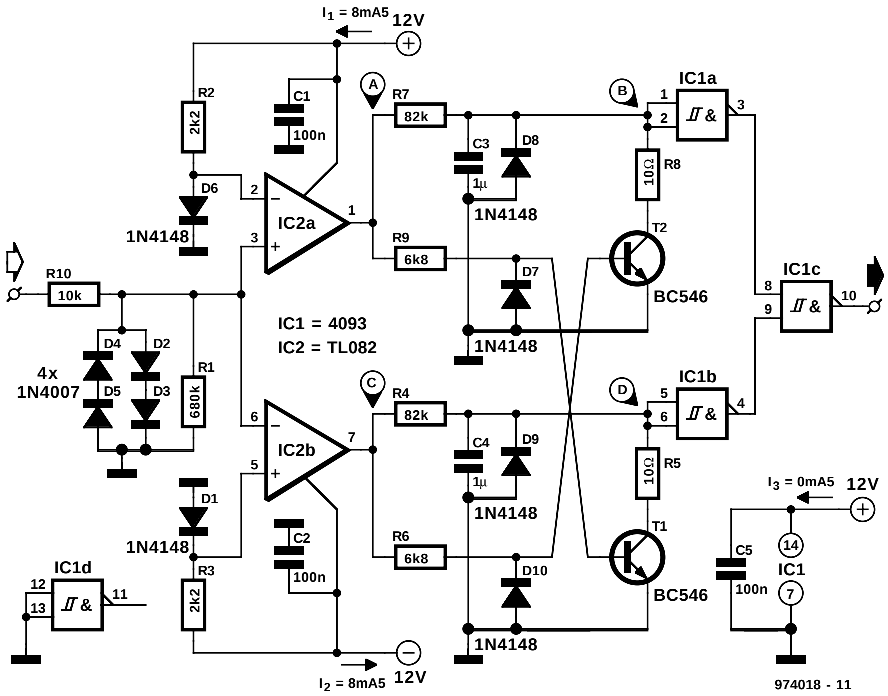
well away from corners so as not to lose the effect of the wideband unit. They should also not be placed too close (less than one metre) to a TV receiver, because the drivers are not shielded.

The frequency response and impedance characteristics are shown in **Figure 5**. Note that the minimum impedance is about 4 Ω at 10 kHz, which is a value that presents no risks to modern drive units.

A listening test on the prototype loudspeakers showed that, in spite of their modest dimensions, the sound quality and the bass reproduction are very good. Where top-class performance is required, they are best combined with a good-quality stereo system. The units are also very suitable for use as the front loudspeakers in a surround-sound installation.

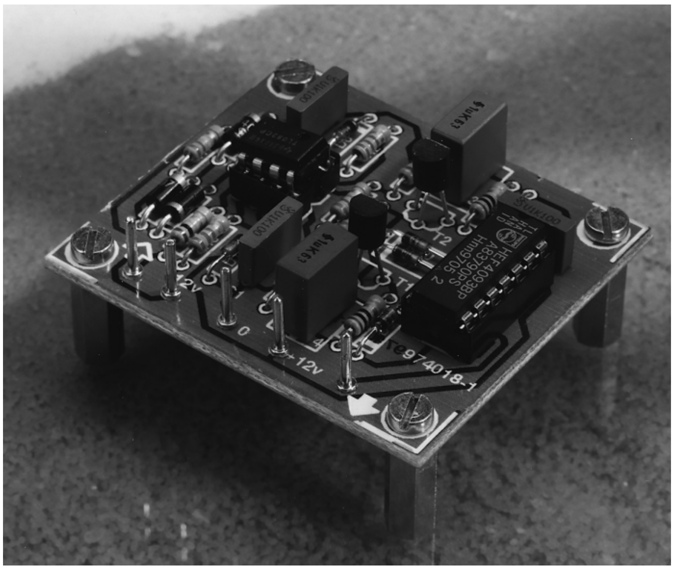
[970091]

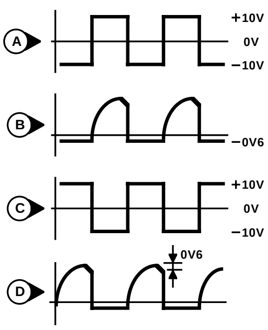
d.c. detector

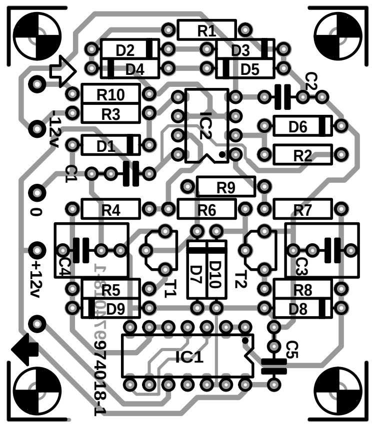


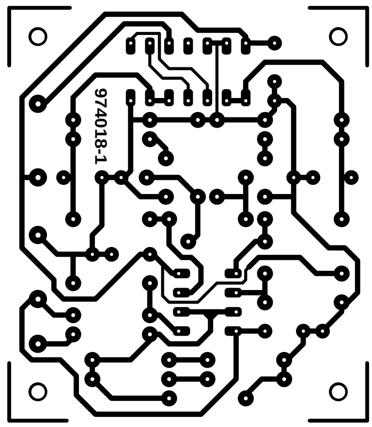
The detector is intended primarily to sense direct voltages at the output of power amplifiers. The signal so detected may be used to enable a protection circuit that, for instance, disconnects the loudspeakers from the amplifier. The circuit has the advantage of reacting at whatever level of direct voltage: always within 75 ms. It also reacts to signals >600 mV at very low frequencies below about 4 Hz, which are likely to damage the loudspeakers.

The circuit is configured symmetrically and may therefore be split into two. The upper part in the diagram processes positive input signals, and the lower part, negative signals.









The signal from the amplifier is applied to the sensor via R_{10}. Its level is limited by diodes D_2–D_5. The trip levels of comparators IC_{2a}–IC_{2b} are set to +600 mV and –600 mV by R_2-D_6 and R_3-D_1 respectively. This means that the output of IC_{2a} goes high when the input voltage is higher than +600 mV and that of IC_{2b} when the input voltage is lower than –600 mV.

It follows that the signals at the outputs of the comparators together form a square wave. This is used to charge C_3 and C_4 alternately to a potential that does not exceed the trip levels of the comparators. This situation changes, however, if, for instance because of a positive offset, the output of IC_{2a} remains high longer than usual. This causes C_3 to be charged to a higher potential, while at the same time T_1 is switched on via R_9 and C_4 is short-circuited. This causes T_2 to be blocked via R_6, so that the potential building up across C_3 cannot be removed via this transistor. This means that the trip level of IC_3 will be exceeded so that the output of the circuit changes from low to high.

The same kind of action occurs if because of a negative offset the output of IC_{2b} remains high longer than usual. It is then C_4, however, that is charged, while IC_{1b} functions as the trigger.

Diodes D_7 and D_{10} protect T_1 and T_2 by preventing their base voltage dropping below –700 mV.

Clearly, the response time of the sensor depends not only on the trigger level of IC_{1a} and IC_{1b}, but also on the time constants R_4-C_4 and R_7-C_3. The HEF4093 used in the prototype triggered at 7.5 V (V_{DD} = 15 V), which resulted in a response time of 57 ms. However, the spread of trigger voltages in the 4093 series is appreciable and it may, therefore, be necessary to lower the values of R_4 and R_7.

The detector is best built in the printed-circuit board shown, but this is not available ready made.

The symmetrical power supply may have an output between ±10 V and ±18 V. The prototype draws a current not exceeding 10 mA.

[Wolff – 974018]

Parts list

Resistors:
R_1 = 680 kΩ
R_2, R_3 = 2.2 kΩ
R_4, R_7 = 82 kΩ
R_5, R_8 = 10 Ω
R_6, R_9 = 6.8 kΩ
R_{10} = 10 kΩ

Capacitors:
C_1, C_2, C_5 = 0.001 µF
C_3, C_4 = 1 µF, MKT (metallized polyester)

Semiconductors:
D_1, D_6–D_{10} = 1N4148
D_2–D_5 = 1N4007
T_1, T_2 = BC546

Integrated circuits:
IC_1 = 4093
IC_2 = TL082CP

Miscellaneous:
5 off board pins

Contents

digital-audio-input selector

As the name indicates, the selector is intended to choose one of up to eight digital audio signal inputs, which it does with the aid of a multiplexer.

The multiplexer, IC_6 is controlled by preset up/down counter IC_2. The counter is set with DIP switch S_3 (note that the MSB switch is not used in this application).

The various inputs are selected with press-keys S_1 and S_2. Gates IC_{1d} and IC_{1e}, in conjunction with networks R_1-C_1 and R_3-C_2, provide effective debouncing of the keys.

Resistor R_5 and capacitor C_3 ensure that when the power is switched on, the counter is set.

If fewer than eight inputs are needed, the number can be reduced to four by resetting jumper J_1 so that pin 9 of IC_6 is linked to a fixed level. The non-used inputs of the multiplexer, pins 1, 2, 4, and 5, must be strapped to earth.

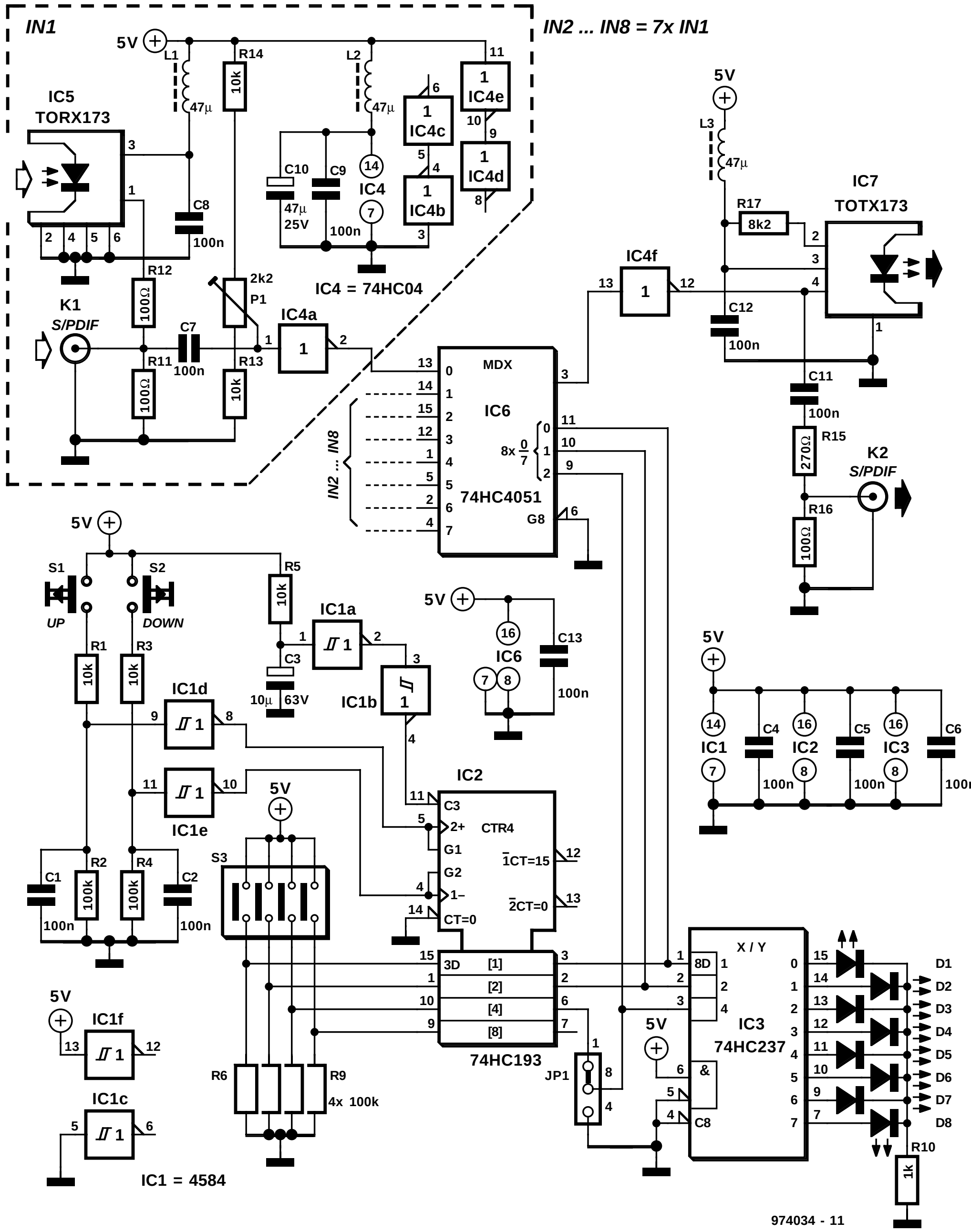
Which of the inputs is selected is indicated by one of four or eight LEDs that are controlled by 3-to-8 decoder IC_3 at the outputs pf IC_2. If four inputs are used, D_5–D_8 must be omitted.

Since the digital-audio- input circuits are identical, only one is shown (in dashed lines at the top left-hand side of the diagram). Each has an optical input (IC_5) and a coaxial input (K_1). It needs only one inverter (here IC_{4a}); the others (IC_{4b}–IC_{4e}) are strapped to earth.

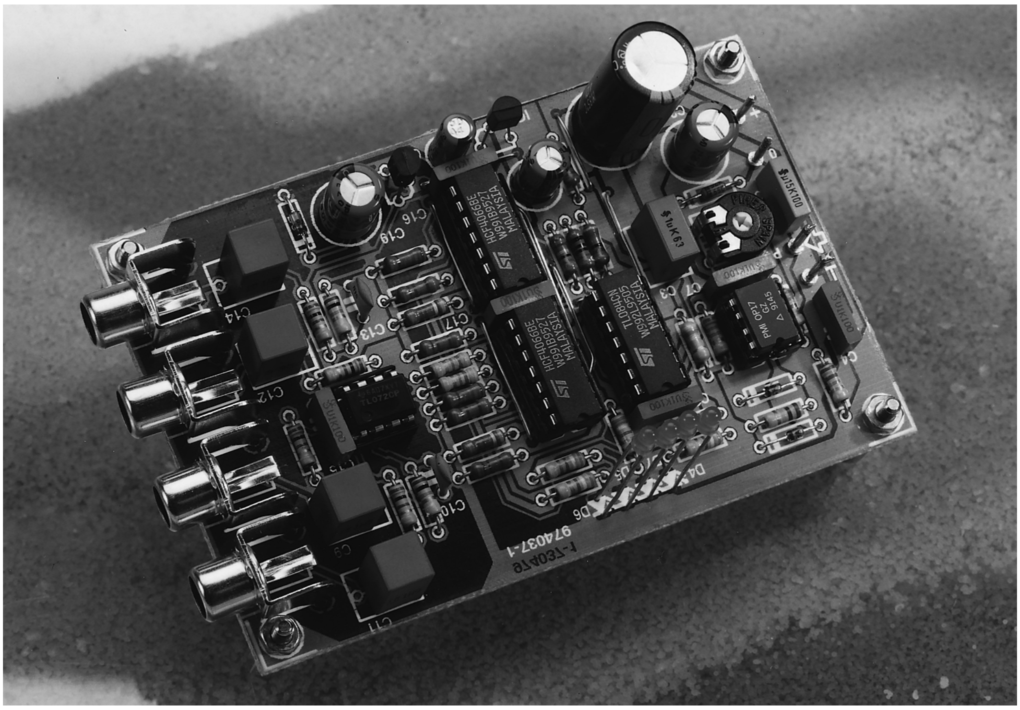
The output of the selector also has an optical output (IC_7) and a coaxial output (K_2).

The current drawn by the selector depends primarily on the number of optical modules (each of which draws 20–25 mA).

If standard LEDs instead of high-efficiency types are used, the value of R_{10} should be lowered to 220 Ω. The total current drain then rises by about 10 mA. [Giesberts – 97034]



auto volume control



The volume control is intended primarily for insertion between a car radio and its booster. It automatically adapts the volume to the amount of road and engine noise. This is done in four 5 dB steps based on the measured sound pressure in the interior of the car. This means that the volume can be increased by up to 20 dB with respect to the set volume level. This implies that care should be taken that the booster and loudspeakers do not become overloaded.

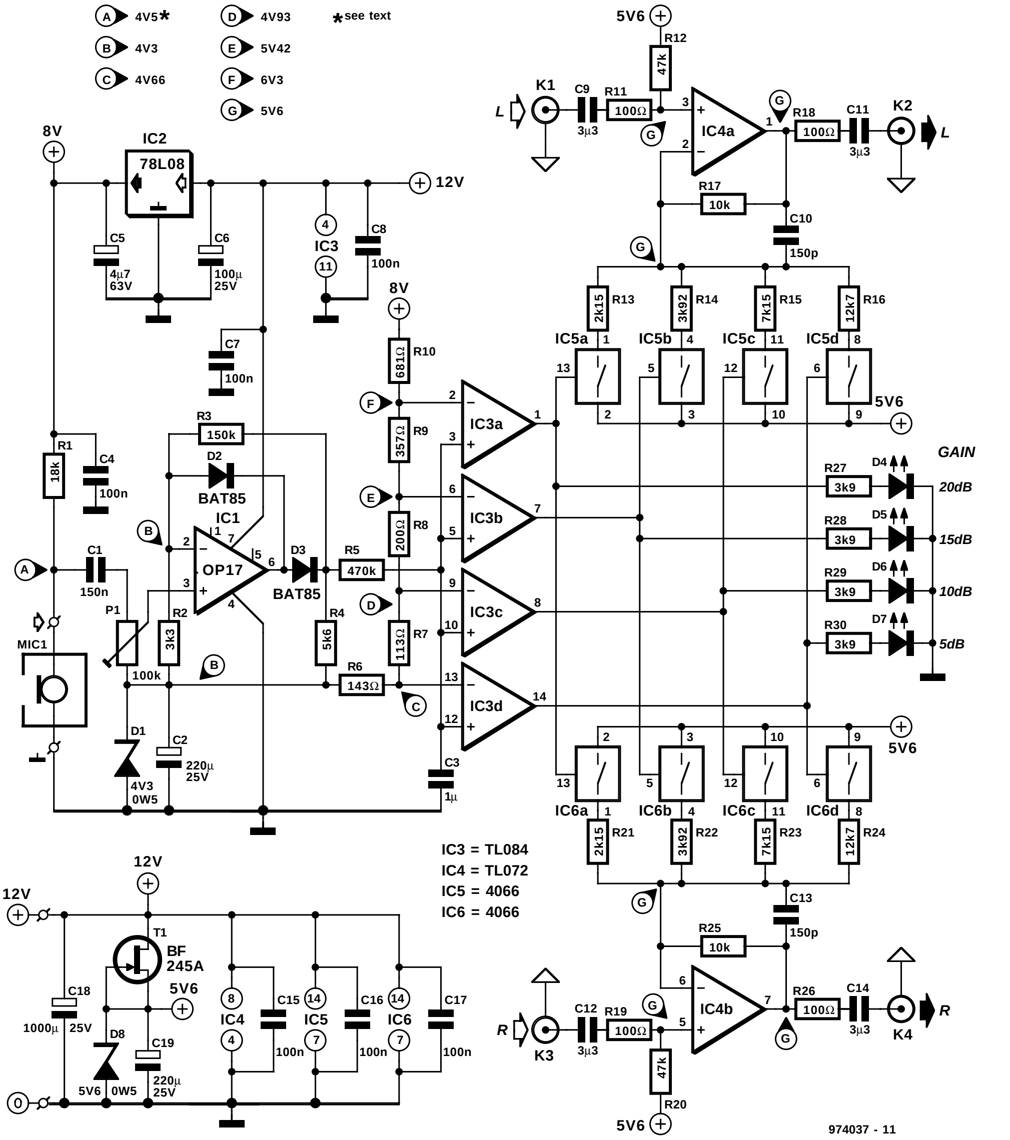
In the diagram in Figure 1, IC_{4a} and IC_{4b} operate as control amplifiers. The audio signal is input via K_1 and K_3 and applied to the booster via K_2 and K_4. The basis level is that registered with the electret microphone MIC_1.

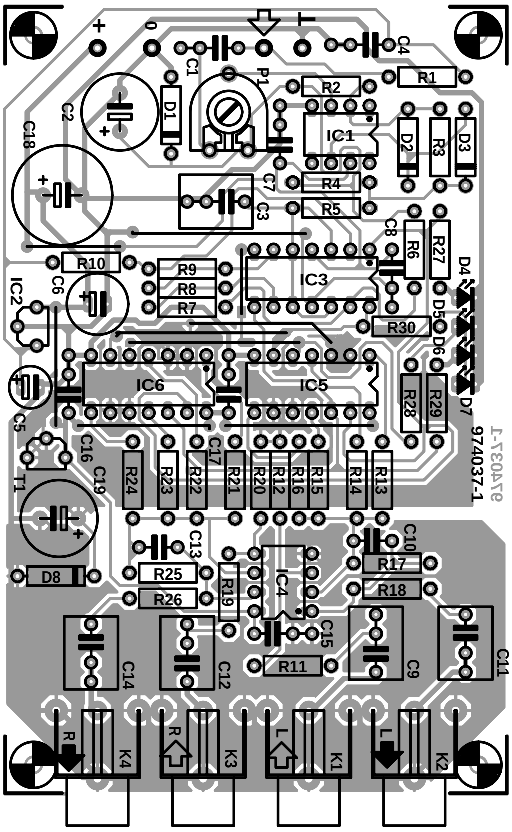
The microphone should not be too sensitive to avoid overdrive and acoustic coupling between it and the loudspeakers. Its d.c. setting is arranged with resistor R_1 while its sensitivity is set with P_1.

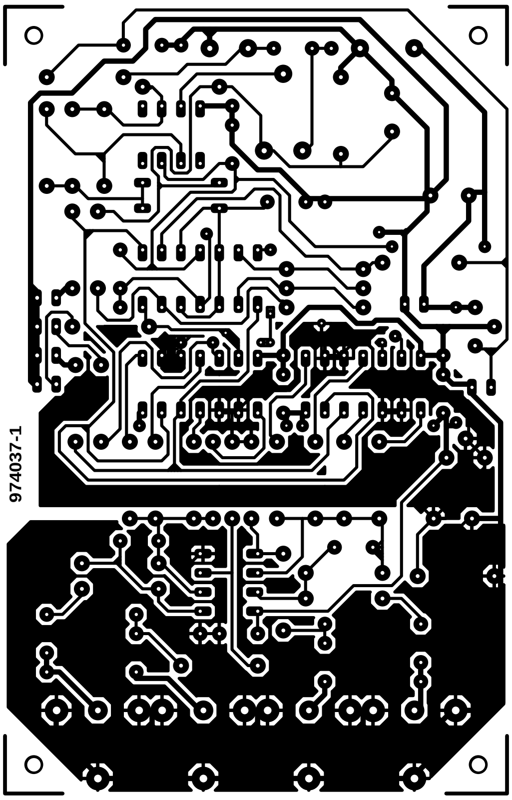
The output of the microphone is applied to fast op amp IC_1 via the wiper of P_1. The op amp, arranged as a rectifier/amplifier, provides an amplification of ×45. Its output is averaged by R_5-C_3 and then applied to comparators IC_{3a}–IC_{3d}.. These liken the amplified signal and averaged signal, U_{AA}, with the potentials at the junctions of divider R_6-R_{10}. Each of these potentials differs by 5 dB from the preceding or next one as the case may be.

The comparators control electronic switches IC_{5a}–IC_{5d} and IC_{6a}–IC_{6d}, which modify the degree of feedback of IC_{4a} and IC_{4b} on the basis of the control input. For instance, if none of the comparators in IC_3 has changed state, IC_{4a} operates as a voltage follower with unity gain. When U_{AA} exceeds the level at junction R_6-R_7, the gain of IC_{4a} is raised by 5 dB. When with increasing road and engine noise it exceeds the level at junction R_9-R_{10}, the switches are all closed so that R_{13}–R_{16} are in parallel, whereupon the gain of IC_{4a} is raised by 20 dB. The position of the automatic volume control is indicated by light-emitting diodes D_4–D_7.

The circuit is powered by the car battery. It is recommended that the battery voltage is well filtered.







The supply lines for the microphone and the voltage divider are held at 8 V by regulator IC_2. That for IC_4 is held at 5.6 V by T_1-D_8, irrespective of the battery voltage.

The circuit draws a current of 40 mA when the LEDs light.

The distortion of 0.0025% is well within the requirements for car hi-fi equipment.

The volume control is best built on the printed-circuit board in Figure 2, which is, however, not available ready made.

[974037]

Parts list

Resistors:
R_1 = 18 kΩ
R_2 = 3.3 kΩ
R_3 = 150 kΩ
R_4 = 5.6 kΩ
R_5 = 470 kΩ
R_6 = 143 Ω
R_7 = 113 Ω
R_8 = 200 Ω
R_9 = 357 Ω
R_{10} = 681 Ω
R_{11}, R_{18}, R_{19}, R_{26} = 100 Ω
R_{12}, R_{20} = 47 kΩ
R_{13}, R_{21} = 2.15 kΩ, 1%
R_{14}, R_{22} = 3.92 kΩ, 1%
R_{15}, R_{23} = 7.15 kΩ, 1%
R_{16}, R_{24} = 12.7 kΩ, 1%
R_{17}, R_{25} = 10 kΩ
R_{27}–R_{30} = 3.9 kΩ
P_1 = 100 kΩ preset

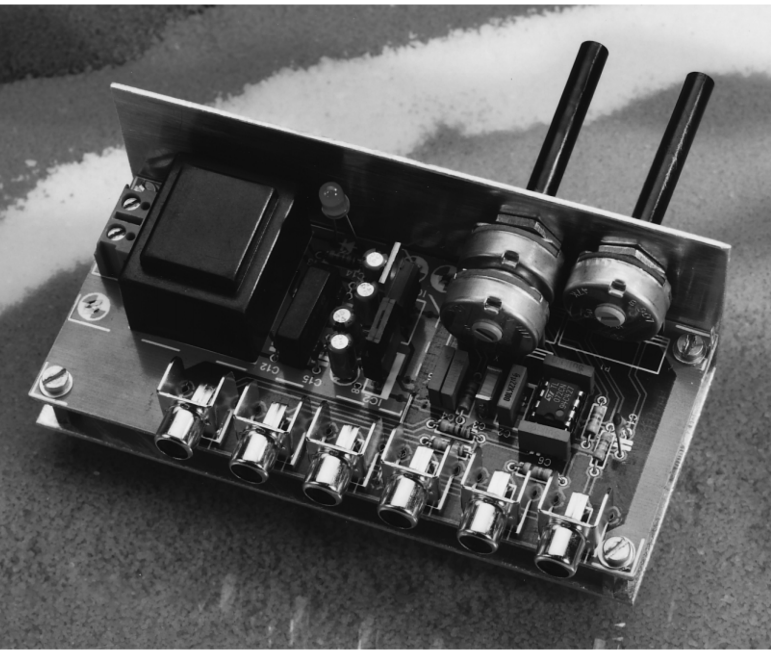
Capacitors:
C_1 = 150 nF
C_2, C_{19} = 220 μF, 25 V, radial
C_3 = 1 μF, MKT (metallized polyester), pitch 5 or 7.5 mm
C_4, C_7, C_8, C_{15}–C_{17} = 100 nF
C_5 = 4.7 μF, 63 V, radial
C_6 = 100 μF, 25 V, radial
C_9, C_{11}, C_{12}, C_{14} = 3.3 μF, MKT (metallized polyester), pitch 5 or 7.5 mm
C_{10}, C_{13} = 150 pF
C_{18} = 1000 μF, 25 V, radial

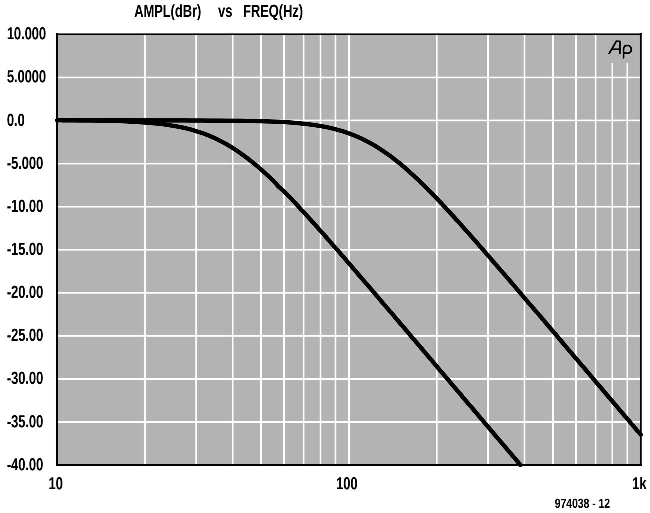
Semiconductors:
D_1 = zener, 4.3 V, 500 mW
D_2, D_3 = BAT85
D_4–D_7 = LED, high-efficiency
D_8 = zener, 5.6. V, 500 mW
T_1 = BF245A

Integrated circuits:
IC_1 = OP17
IC_2 = 78L08
IC_3 = TL084
IC_4 = TL072
IC_5, IC_6 = 4066

Miscellaneous:
K_1–K_4 = audio socket for board mounting
MIC_1 = electret microphone

bass extension for surround sound





The extension is intended primarily for surround-sound installations that need some boosting of the bass frequencies but where an additional subwoofer cannot be afforded. It is based on a disused mono a.f. amplifier and loudspeaker. If these provide reasonable bass performance, they can be converted into a fairly good subwoofer with the aid of an active low-pass filter—see Figure 1.

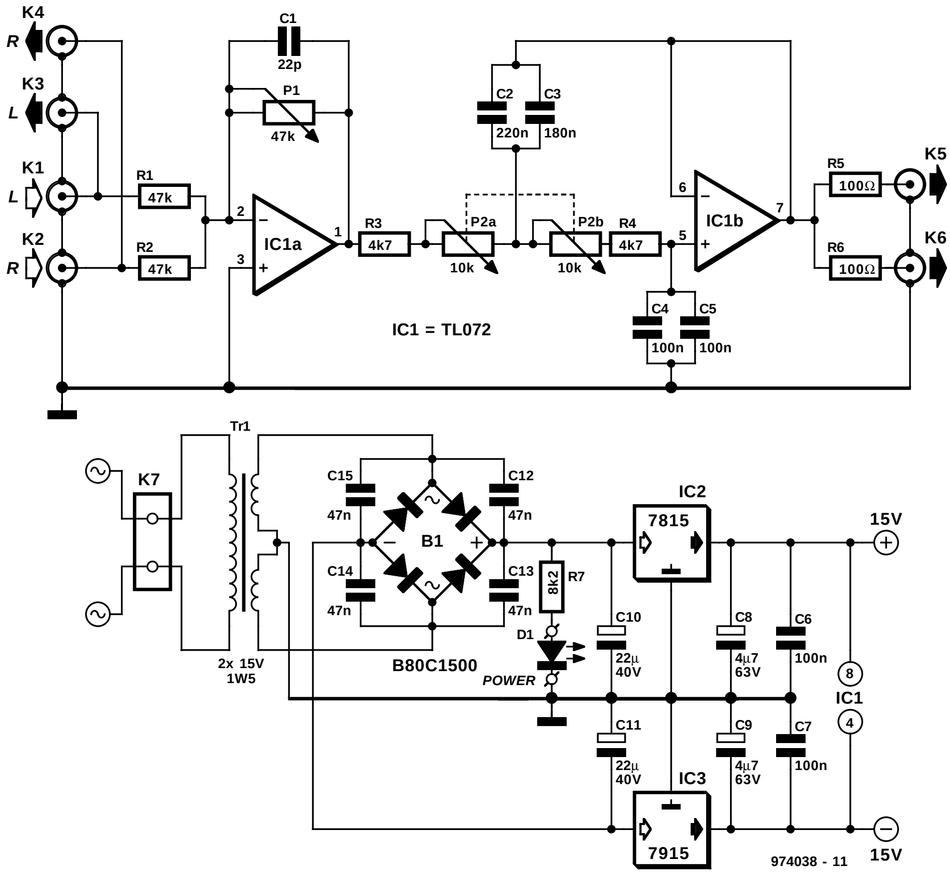
The input signals for the left-hand and right-hand channels are applied to audio sockets K_1 and K_2 respectively. They are output via audio sockets K_3 and K_4 to which the surround-sound decoder is connected.

The signals of the two channels are summed in IC_{1a}, which also functions as input amplifier. The amplification, and therefore the sensitivity of the 'subwoofer', can be adjusted with P_1.

The output of IC_{1a} is applied to a 2nd-order Butterworth low-pass filter. The cut-off frequency of this active filter can be set between 40 Hz and 120 Hz with stereo potentiometer P_2. The response characteristic of the filter at both these frequencies is shown in Figure 2. The actual cut-off point depends on individual taste.

The said mono amplifier is connected to audio output sockets K_5 and K_6.

The power supply for the circuit is simple and consists of a small mains transformer, Tr_1, a



Parts list

Resistors:
R_1, R_2 = 47 kΩ
R_3, R_4 = 4.7 kΩ
R_5, R_6 = 100 Ω
R_7 = 8.2 kΩ
P_1 = 47 kΩ logarithmic potentiometer
P_2 = 10 kΩ, linear stereo potentiometer

Capacitors:
C_1 = 22 pF
C_2 = 220 nF
C_3 = 180 nF
C_4–C_7 = 100 nF
C_8, C_9 = 4.7 µF, 63 V, radial
C_{10}, C_{11} = 22 µF, 40 V, radial
C_{12}–C_{15} = 47 nF ceramic

Semiconductors:
D_1 = LED, high efficiency

Integrated circuits:
IC_1 = TL072CP
IC_2 = 7815
IC_3 = 7915

Miscellaneous:
K_1–K_6, K_8–K_9 = audio socket for board mounting
K_7 = 2-way terminal block, pitch 7.5 mm
B_1 = B80C1500
Tr_1 = mains transformer, 2 × 15 V secondaries, 1.5 VA

bridge rectifier, B_1, antihunt capacitors C_{12}–C_{15}, a number of smoothing and decoupling capacitors, and two integrated voltage regulators, IC_2 and IC_3.

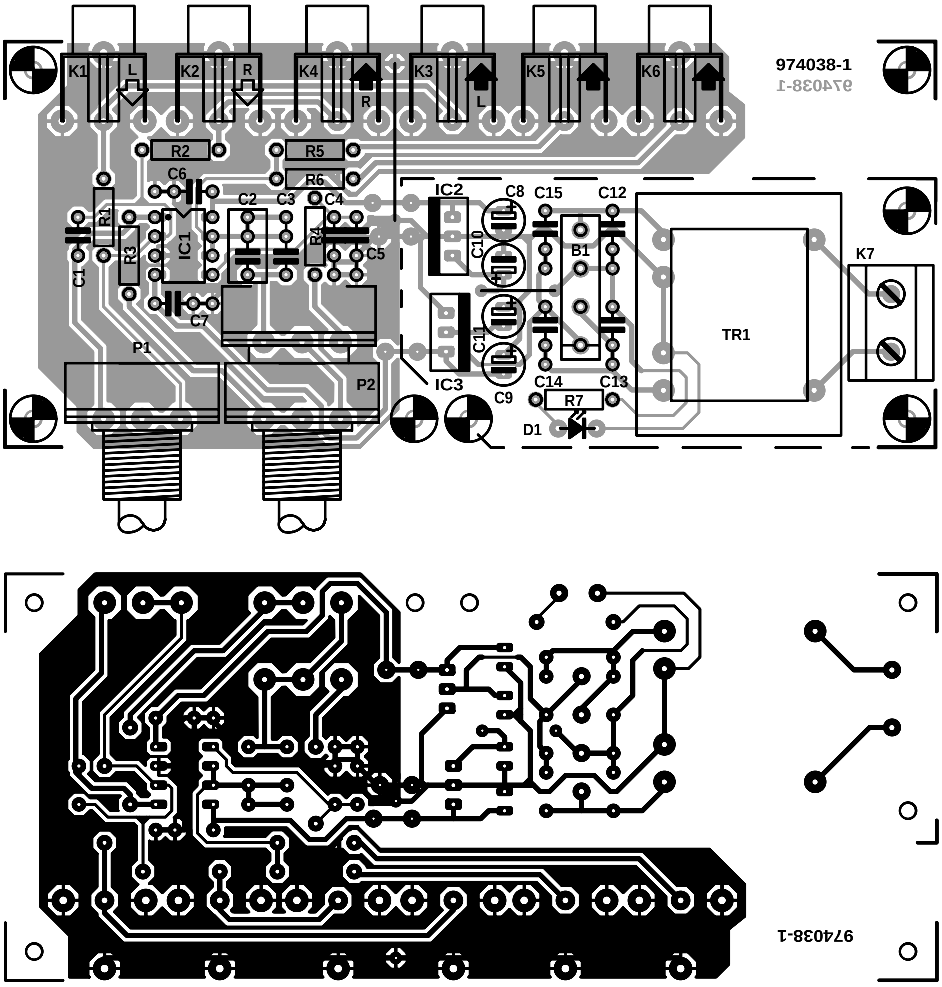
The filter circuit is best built on the printed-circuit board shown in Figure 3, which is, however, not available ready made.

The filter should be housed in a metal case. Moreover, P_1 and P_2 should preferably be types with a metal enclosure. Hum is prevented by earthing the case and the enclosures.

The harmonic distortion, with two input signals of 200 mV and a bandwidth of 22 kHz, is 0.0016% at 30 Hz.

Although not of prime importance at low frequencies, the polarity of the 'subwoofer' should be the reverse of that of the remainder of the system since the present circuit inverts the signals.

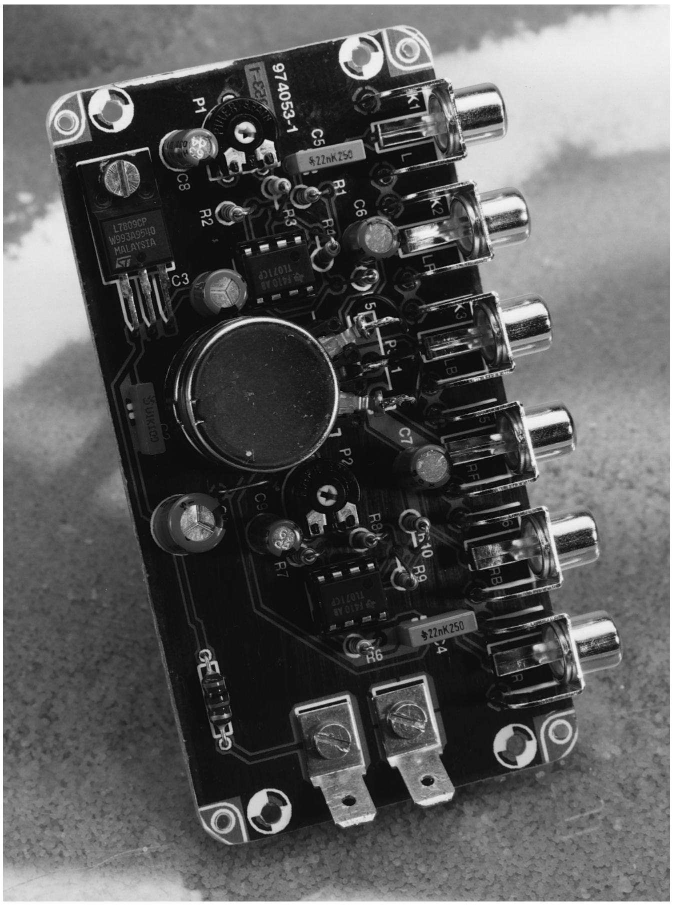
[974038]

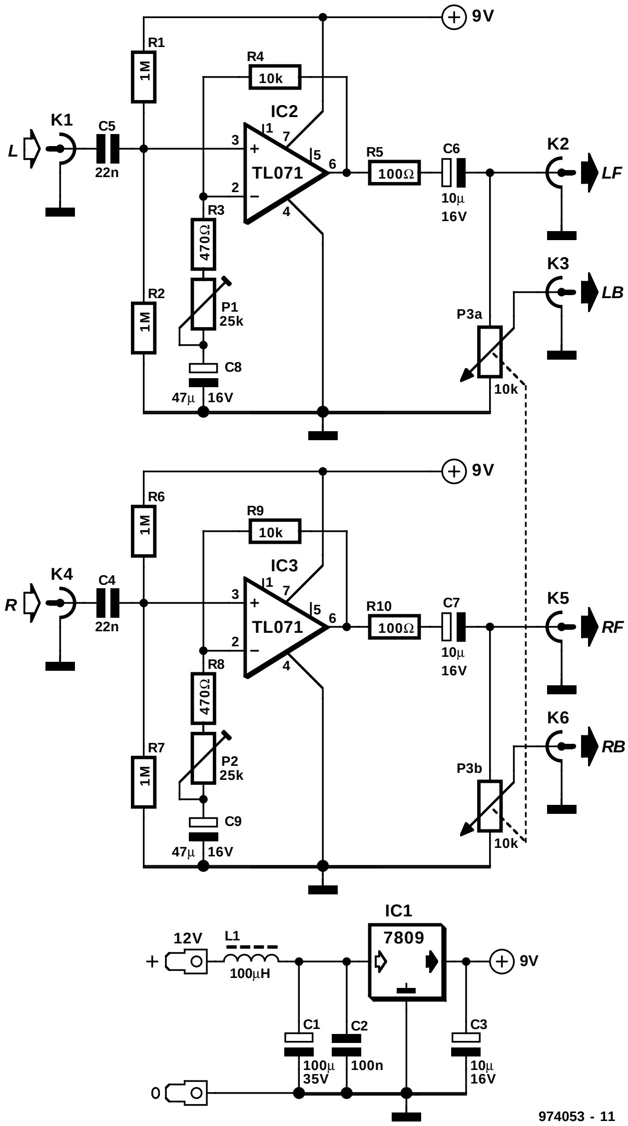


ELEKTOR	
240V~	50Hz
No. 974038	
P = 1VA5	

Contents

car booster adaptor





Judging by the cacaphony emanating from an increasing number of cars on the road, car radio boosters unfortunately remain popular with young people. Unfortunately, because deafness among these young people is becoming quite common.

From a technical point of view, the setup with a booster is often very in efficient, because these power monsters are normally connected simply to the loudspeaker terminals of the existing car radio installation via an attenuator. This puts the two output amplifiers in series, which is, as said, quite inefficient.

It is much better to take the signal from the wiper of the volume control in the car radio and

Components list

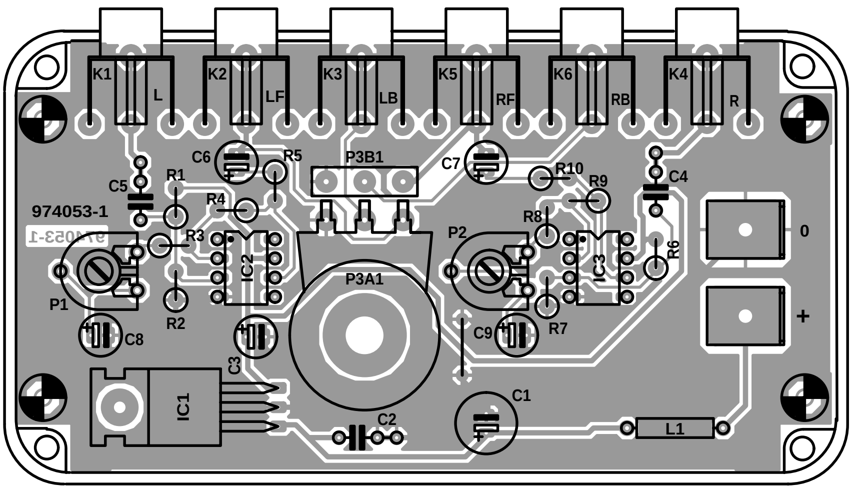
Resistors:
R_1, R_2, R_6, R_7 = 1 MΩ
R_3, R_8 = 470 Ω
R_4, R_9 = 10 kΩ
R_5, R_{10} = 100 Ω
P_1, P_2 = 25 kΩ preset
P_3 = 10 kΩ log stereo potmeter

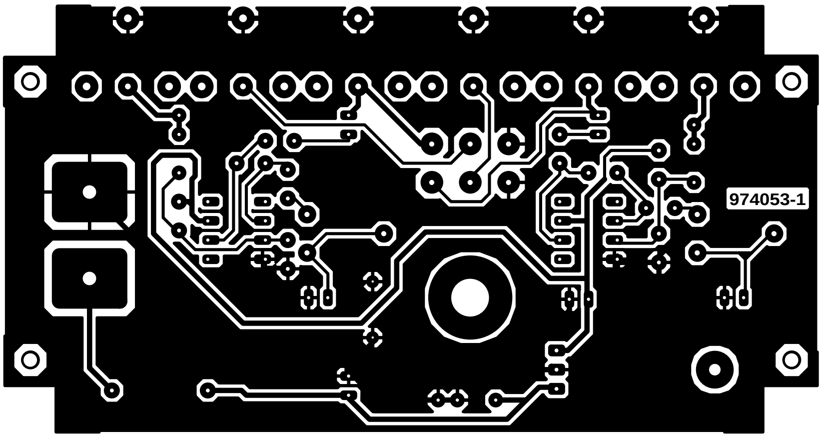
Capacitors:
C_1 = 100 µF, 35 V, radial
C_2 = 0.001 µF, high stability
C_3, C_6, C_7 = 10 µF, 16 V, radial
C_4, C_5 = 0.022 µF
C_8, C_9 = 47 µF, 16 V, radial

Inductors:
L_1 = 100 µH

Integrated circuits:
IC_1 = 7809
IC_2, IC_3 = TL071CP

Miscellaneous:
K_1–K_6 = audio socket for board mounting
2 off car-type connector for board mounting





use this as the input to the booster. This is normally not much of a job. The signal so obtained must, however, be buffered and sometimes also amplified.

The adaptor provides both these functions in a simple manner. The stereo signals are applied via K_1 and K_2 and buffered an amplified by an op amp in each channel. The amplification may be set between ×1.5 and ×22 with P_1 and P_2 respectively. These levels should be more than adequate for most situations. The peak output voltage is 2 V_{RMS}.

The output in each chan-nel is split into a front and a rear branch (left-hand front, LF, and left-hand back, LB, and RF and RB respectively). The volume of the rear speakers is set with P_3.

Regulator IC_1 provides a stable 9 V supply line for the op amps. The circuit draws a current of not more than 7 mA.

The adaptor is best built on the printed-circuit shown, which is, however, not available ready-made.

The input and output terminals are audio sockets for board mounting.

The battery voltage is applied to the circuit via two car-type connectors mounted on the board. When the adaptor is fitted in a small case, care must be taken that P_3 remains accessible.

[Bonekamp – 974053]

digital potentiometer

Xicor's digitally controlled E^{2}POT ICs provide ergonomic and long-lasting alternatives to mechanical potentiometers. The ICs in the X9CMME series have a 7-bit counter with reversible count direction and a decoder that enables one of the 100 analogue switches.

The outputs of the analogue switches serve as the wiper of a potentiometer, while the inputs are linked to a potential divider composed of 99 equal resistors. The counter state may be stored in a non-volatile EEPROM, so that it can serve as the output value at a subsequent start.

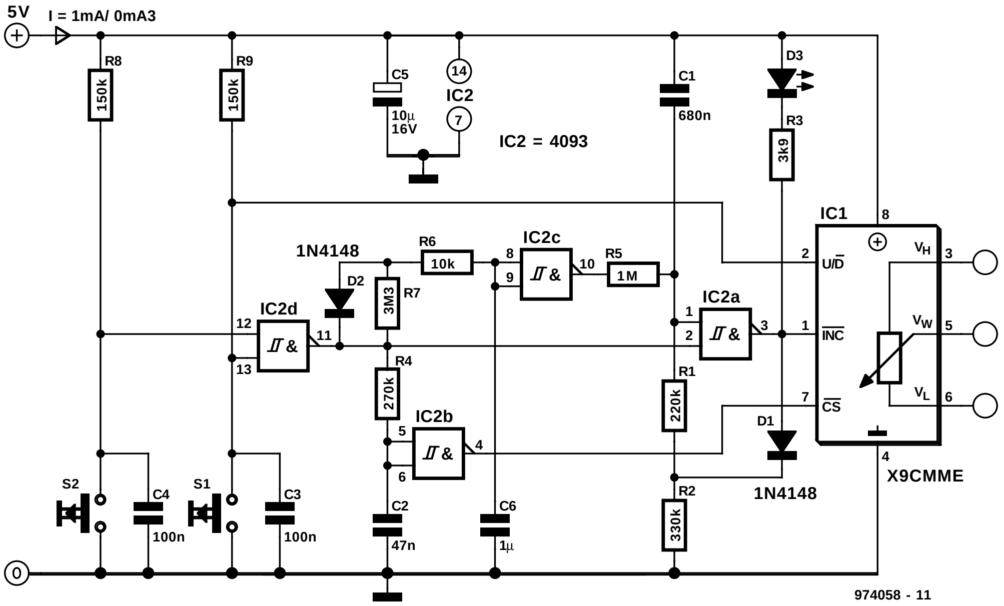
The X9CMM series is designed to operate from 5 V supply lines. The potential across the resistive divider must not exceed 10 V (only 4 V in case of the X9C102). The ON resistance of the analogue switch is about 40 Ω, so that the current through the wiper is limited to 1 mA.

E^{2}POT ICs have three inputs for the digital drive. The level at U/D determines whether a trailing edge at clock input INC lowers or raises the counter state. This action only takes place if chip select input CS is low. A leading edge at CS arranges for the counter state to be stored when INC is high. When CS is high, the IC is in the standby mode.

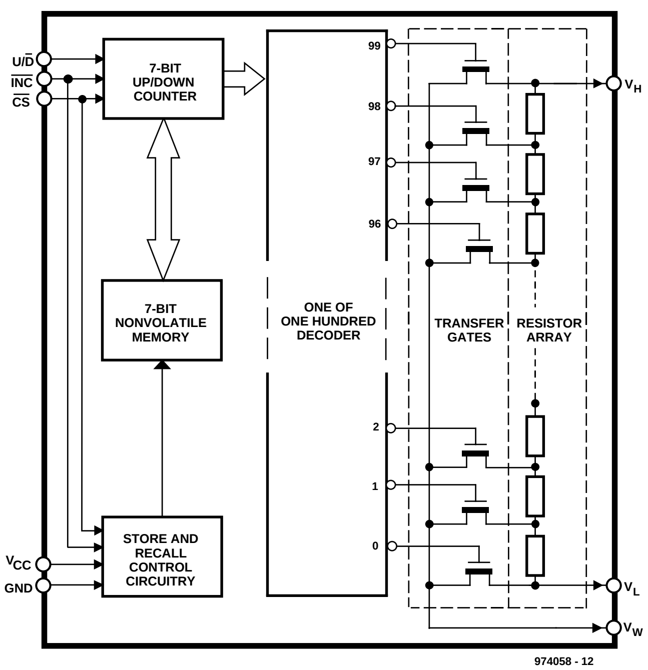
The circuit diagram shows a complete digital potentiometer based on a Type X9CMME. It is provided with two controls, S_1 and S_2, an optical indicator and a delayed frequency change-over of the clock generator.

When keys S_1 and S_2 are open, resistors R_8 and R_9 hold the inputs of IC_{2d}, a NAND, as well as the U/D input of IC_1 high. The low level at the output of IC_{2d} disables clock generator IC_{2a}. Frequency determining capacitor C_1 is discharged in the quiescent state.

When one of the keys is pressed (S_1 firmly, S_2 gently), the output of IC_{2d} changes state, so that the clock



generator and IC_1 (via IC_{2b}) are enabled. Capacitor C_1 is then charged via R_1 and R_2 until the input level of IC_{2a} goes low, whereupon the gate output linked to the clock input of IC_1 changes state (from low to high). When this happens, C_1 is discharged via R_1 and D_1 until the upper trigger level of IC_{2a} is attained. The gate then changes state again and the above action repeats itself.



The clock signal is optically monitored by D_3.

When the output of IC_{2c} is high, the gate draws a portion of the charging current from C_1, which results in the clock frequency at INC being relatively low.

At the same time that the generator is enabled, C_6 begins to be charged gradually via R_6 and R_7 until IC_{2c} changes states (from high to low). Circuit IC_2 then contributes to the charging current to C_1, whereupon the clock frequency increases: in the prototype, the frequency rose in four seconds from 1.3 Hz to 3.1 Hz.

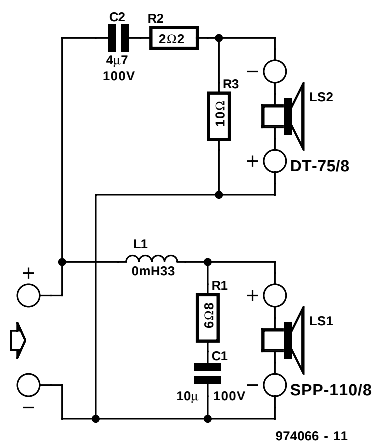
When the keys are released, the clock generator stops. At the same time, C_6 is discharged rapidly via R_6 and D_2, so that the frequency is low again when the keys are operated anew.

The switch-off delay owing to R_4-C_2 enables the actual counter state to be stored by the internal logic.

The circuit draws a current of 0.3–1.0 mA. [Kühne – 974058]

simple two-way loudspeaker

1



2

In this design of a simple loudspeaker enclosure an attempt has been made to achieve a reasonably good quality with a minimum of material. In spite of inexpensive drivers, this aim was met.

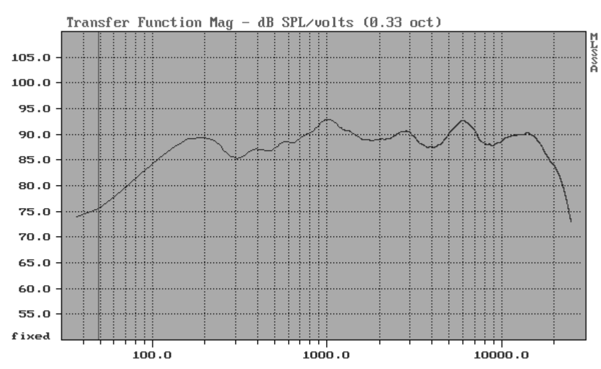
The cross-over filter has a 6-dB roll-off, which means only one component per driver: L_1 for the woofer and C_2 for the tweeter. There is also an impedance-correction network, R_1-C_1, for the woofer, which 'flattens' the rising impedance of this driver.

There is an attenuation network, R_2-R_3, to match the volume level of the tweeter to that of the woofer.

Note that owing to the position of the drivers, the polarization of the tweeter must be the opposite of that of the woofer.

The unit may be used as a rear speaker in a surround-sound system or with a multimedia computer. In the latter case, it must be placed well away from the monitor since the magnets of most inexpensive drivers are not screened.

3



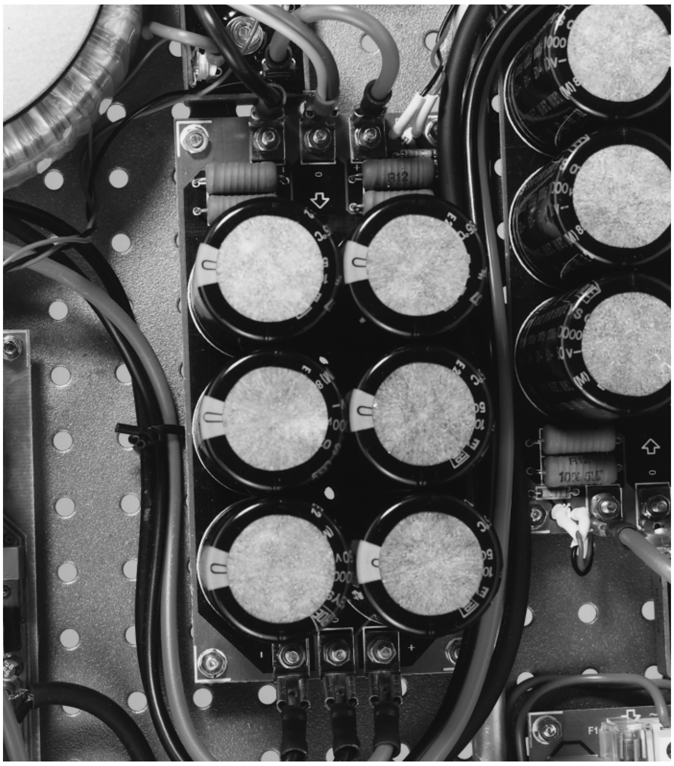
The bass-reflex enclosure (Figure 2) has a volume of 4.5 litres. The bass-reflex port is is a standard 40 mm dia. PVC pipe, 175 mm long (if its walls are 2 mm thick; if they are 3 mm thick, the length must be 150 mm). Te material used for the enclosure is 8 mm thick chip-board or similar.

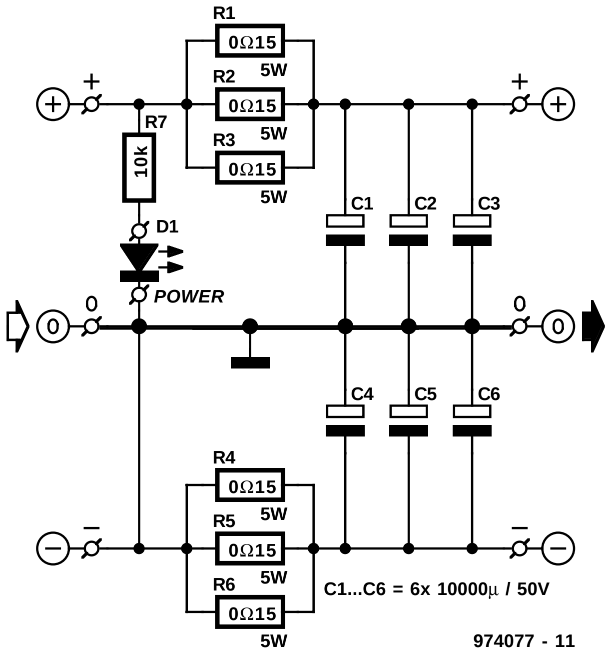
The nominal impedance of the system is 6 Ω. Maximum power input is 30 W. The cross-over frequency is 4 kHz. The frequency characteristic of the loudspeaker is shown in Figure 3.

If the coil is not obtainable ready-made, it may be wound on a non-metallic former, 28 mm dia and 28 mm long. The winding consists of seven layers of 1.5 mm dia. enamelled copper wire.

[Giesberts - 974066]

supply board for output amplifiers





Apart from their electronic configuration, all output amplifiers comprise the same elements: an amplifier board, a mains transformer, a bridge rectifier, and electrolytic smoothing capacitors. The board is normally screwed to the heat sink, while the transformer and bridge rectifier are fixed to the bottom of the enclosure. Often, there is no such defined location for the electrolytic capacitors. These are mounted on a piece of prototyping board, or to the bottom of the enclosure with suitable brackets, or …

Since this is a recurring difficulty, many constructors will be pleased with the board design shown here. Its layout is such that it is suitable for use with almost any type of output amplifier operating from a symmetrical power supply.

The board can accommodate six electrolytic capacitors with a value of up to 10,000 µF and a rating of 50 V. They are assumed to have a pitch of

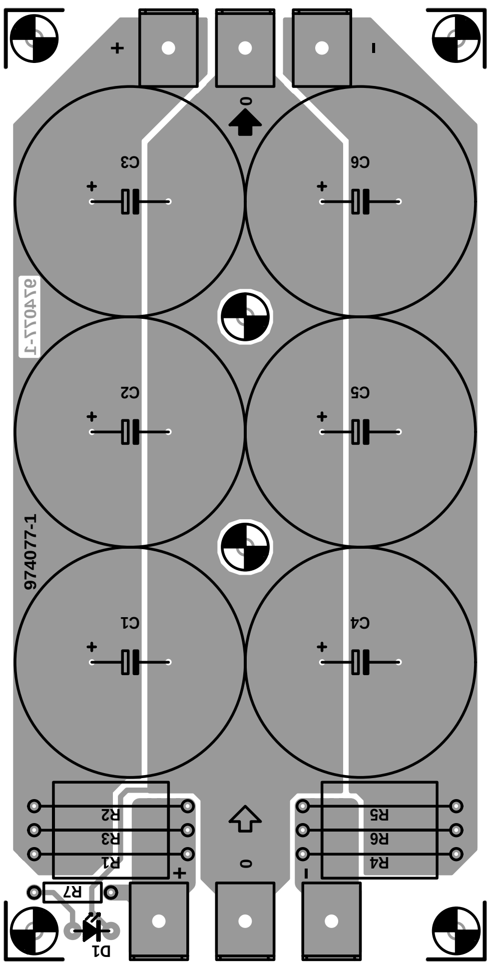
Parts list

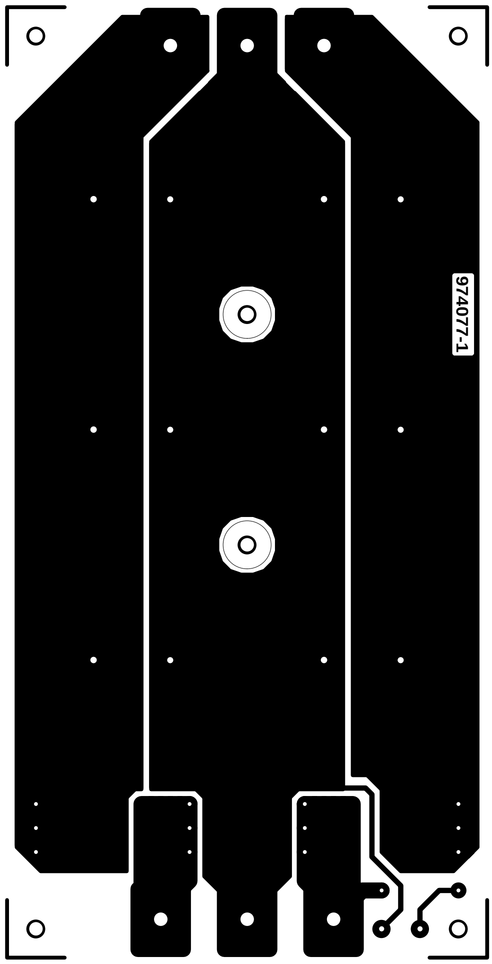
Resistors:
R_1–R_6 = 0.15 Ω, 5 W
R_7 = 10 kΩ

Capacitors:
C_1–C_6 = 10,000 µF, 50 V, pitch 10 mm, max. dia. 30 mm

Semiconductors:
D_1 = LED, high efficiency

Miscellaneous:
6 off single-pole PCB terminal block





10 mm and a maximum diameter of 30 mm.

The board also has space for 'soft switch-on' resistors with a value of 0.15 Ω and rated at 5 W. These resistors damp the peaks in charging current and also aid in smoothing spurious current peaks on the supply voltage.

Finally, the board has an on/off indicator in the shape of a high-efficiency LED and requisite series resistor.

Connections to the board are via single-pole PCB terminal blocks, which guarantee good contacts and can handle large currents.

[Giesberts – 974077]

mains on delay circuit

The delay is intended to switch on the mains to heavy loads gradually to ensure that the switch-on current remains within certain limits and to prevent the fuses from blowing. The elements that cause high currents at switch-on are, for instance, the electrolytic capacitors in the power supply of an output amplifier. Since these are not charged at switch-on, they constitute a virtual short-circuit on the supply lines. The current can, however, be kept within limits by inserting the present delay circuit between the mains outlet and the transformer primary. The amplifier is then powered in two stages: in the first instance, the current is limited by a number of heavy-duty series resistors; a second later these resistors are shunted (short-circuited) by a relay contact.

In the diagram, R_4–R_7 are the heavy-duty series resistors, each with a value of 10 Ω and rated at 5 W. They limit the switch-on current to about 5.5 A.

The relay is a type whose contact is rated at 2000 VA, which will be sufficient in most cases. Its supply is derived directly from the mains via potential divider R_3-C_1-B_1-relay coil. The resistor, R_3 limits the current at switch-on, after which C_1 limits the current in normal operation to about 20 mA. The delay time is determined by electrolytic capacitors C_2 and C_3 in parallel with the relay. The delay time may be altered by suitably changing the value of one or both of these capacitors.

For safety's sake, the board also has provision for a mains fuse, F_1. The rating of this depends, of course, on the current drawn by the load.

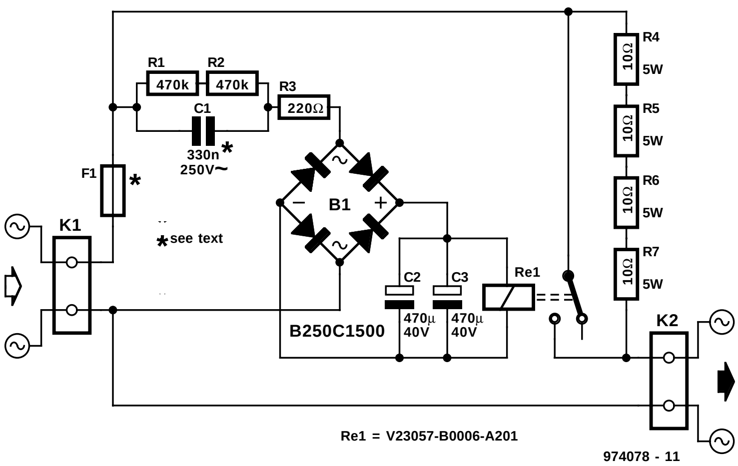
It should be noted that in the case of a double-mono stereo output amplifier (with separate power supplies) each of the mono amplifiers must be given a mains-on delay.

As mentioned earlier, the values of R_4–R_7 refer to a switch-on current of about 5.5 A. If the power rating of the load is lower than 200 VA, it is advisable to use resistors with a slightly higher value.

Note that C_1 is a metallized paper type, which is designed specially for mains voltage applications and meets stringent regulatory requirements.

Finally, at all times bear in mind that the circuit is connected to the mains, so do not touch anything inside the unit during operation and make sure that all wiring is safe and secure.

[Giesberts – 974078]

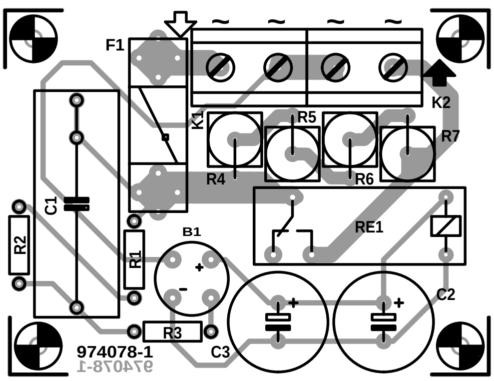


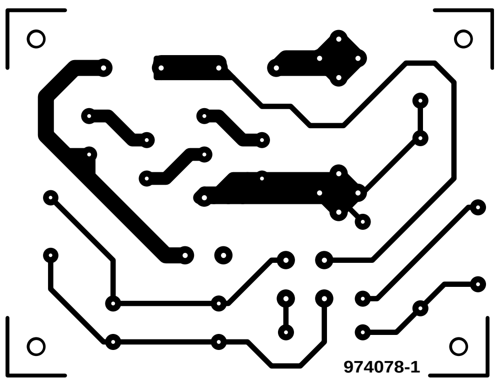
Parts list

Resistors:
R_1, R_2 = 470 kΩ
R_3 = 220 Ω
R_4–R_7 = 10 Ω, 5 W

Capacitors:
C_1 = 0.33 µF, 250 VAC, metallized paper
C_2, C_3 = 470 µF, 40 V

Miscellaneous:
K_1, K_2 = 2-way terminal block, pitch 7.5 mm
B_1 = B250C1500, round
Re_1 = contact rating 250 V, 8 A, coil 24 V, 1200 Ω
F_1 = see text





Contents

AES/EBU-to-S/PDIF converter

The converter is intended primarily for use with the sample rate converter published in the October 1996 issue of this magazine.

The conversion of a symmetrical signal to an asymmetrical one requires no more than a small transformer. Amplification is not required since the AES/EBU signal is strong enough to generate the S/PDIF signal (500 mV_{pp} into 75 Ω). However, the quality of the conversion depends entirely on that of the DIY transformer.

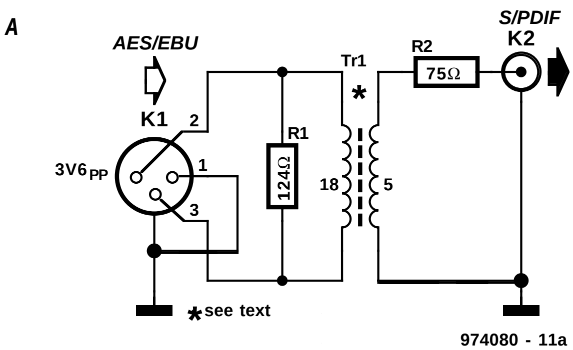
The simplicity of the circuit means that the turns ratio depends on the level of the symmetrical input voltage. This is the reason that the diagram shows two versions. Version A is suitable for inputs of 3.6 V_{pp} and Version B for inputs of 5 V_{pp}.

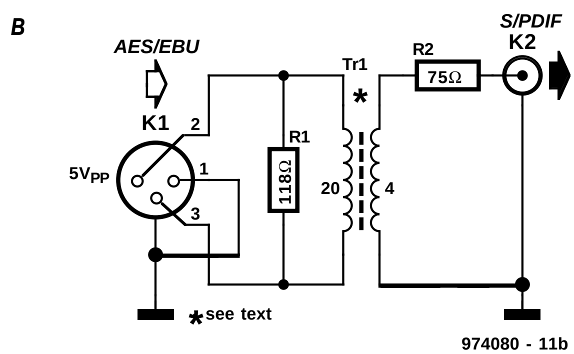
The transformer is wound on a Type G2-3/FT12 core. The primary as well as the secondary are wound from enamelled copper wire of 0.5 mm dia. The transformer for Version A needs a primary winding of 18 turns and a secondary of 5 turns. That for Version B requires a primary of 20 turns and a secondary of 4 turns.

The secondary impedance is transformed to the primary winding. Assuming that the system has a correctly terminated output of 75 Ω, the primary winding needs to be shunted by a resistor, R_1, of 124 Ω (Version A) or 118 Ω (Version B) to give an input impedance of 110 Ω. This arrangement ensures a correct input impedance over a wide range of input frequencies. Only at 60 kHz (Version A) or 50 kHz (Version B) does the impedance drop by about 20%.

The bandwidth of the converter is ≥ 20 MHz.

[Giesberts – 974080]





S/PDIF-to-AES/EBU converter

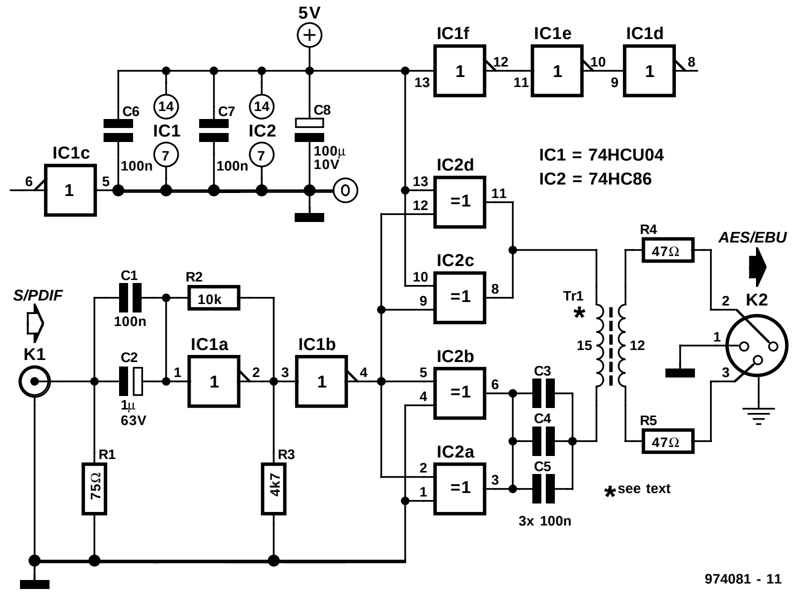
This magazine assumes that most readers use consumer appliances. However, in the case of the sample rate converter (October 1996), many readers have asked for a conversion from the asymmetrical S/PDIF format to the symmetrical (professional) AES/ EBU format and such a converter is presented here.

The timing and levels comply with the AES3-1992 Standard. This means that: (a) the output voltage must be 2–7 V_{pp} (transmitter load 100 Ω); (b) the rise and decay times must be 5–30 ns; (c) the output impedance must be 110 Ω ±20% (within the bandwidth of 0.1–6 MHz). These requirements are met in the design in the diagram (30 ns; 3.6 V_{pp}; 115 Ω respectively).

The circuit at the input, based on IC_1, converts the S/PDIF signal to HC levels. Op amp IC_{1a} is an analogue amplifier, while IC_{1b} raises the signal to the level of the supply lines. Resistor R_3 pulls IC_{1a} slightly from its centre of operation, so that the output buffer attains a logic level even in the absence of an input signal.

The buffer to drive the output transformer is formed by a symmetrical circuit based on IC_{2a}–IC_{2d}. This arrangement ensures that the rise and decay times are equal and that the output voltage is large enough. The use of XOR gates ensures that the transfer times for inverting and non-inverting of the output of IC_{1b} are equal. Since the primary transformer voltage is 9.5 V, the secondary voltage could be decreased slightly. This is beneficial for the linearity of the impedance and the bandwidth of the converter.

The transformer is wound on a Type G2-3/FT12 core: the primary on one side and the secondary on the other. Both windings consist of enamelled copper wire of 0.5 mm dia. The core can accommodate a tin-plate screen for maximum common-mode suppression. Regulations



require this to be ≥ –30 dB w.r.t. the nominal output level; in the present circuit it is –48 dB (with screen).

The output impedance, ignoring R_4 and R_5, is about 22 Ω. If a figure of exactly 110 Ω is wanted, R_4 and R_5 should have a value of 44.2 Ω.

Capacitors C_3–C_5 prevent any direct current flowing through the transformer in the absence of a signal as this would short-circuit IC_2. The use of three capacitors in parallel ensures that the impedance and loss resistance of them (ceramic, high-stability types) are low.

The AES/EBU signal is output via XLR connectors (to IEC268-12). Note that versions with male pins and female shells are used. Pin 1 is for the screen or the signal earth; pins 2 and 3 are for the signals – the phase is not important.

The circuit requires a 5 V power supply from which a current of about 26 mA is drawn.

If the converter is used with the sample rate converter published in the October 1996 issue or the 20-bit analogue-to-digital converter in the December 1996 issue, do not forget to use the CS8402A in the professional mode.

[Giesberts – 974081]

AF input selection

Contents

Input selection in the battery operated preamplifier published in the January 1997 issue of this magazine is by rotary switch. This does not give the best possible performance as far as crosstalk and channel separation are concerned. This article proposes a better means, which also enables the number of input to be extended to 12. In this design, each input source is linked to the circuit via a bistable relay (described elsewhere in this issue).

Use is made of a single-pole, 12 position rotary switch, S_1, with which and 12 pull-down resistors the input source is selected. Since only one resistor is linked to the positive supply line at any one time, the current drawn by the circuit is only 15 µA, which, in the case of a battery operated preamplifier, is an important advantage.

The 12 outputs of S_1 are linked to parity checker IC_3. The output of this device is high only if an odd number of inputs is high. When S_1 is turned, all inputs go low briefly and so the output of S_3 also becomes low for an instant. The output of IC_3 then triggers monostable multivibrator (MMV) IC_{4a}. Since this is retriggerable, its output will be a single pulse even with contact bounce of S_1. As long as trigger pulses arrive during the period the output is active, the output pulse is stretched. To make absolutely sure, the time can be set with P_1 between 0.1 s and 1 s.

The outputs of S_1 are also linked to D-type bistables IC_1 and IC_2, which ensure a stable change-over of the output levels. The bistables have the advantage that they can be reset. This facility is made use of by resetting all relays before a change of input, so ensuring that only one input is linked to the circuit at any one time. This arrangement provides a dead time between the releasing of one relay and the tripping of another. This dead time corresponds to the sum of the mono times of IC_{4a} and IC_{4b}. MMV IC_{4b} serves to clock the inputs of all the D-type bistable. Since this requires a pulse of only 10 µs, the dead time is determined primarily by IC_{4a}.

The Q output of IC_{4a} is used to reset the D-type bistables, but it also provides the reset pulse for all relays together. Accordingly, the new data from S_1 is accepted by the D-type bistables 10 µs after the reset pulse. To enable the position of S_1 to be assumed during the rise time of the supply lines, the bistables need an additional pulse and this is provided by R_5-C_3.

About 4 seconds after the supply has been switched on, the 13th input of the parity checker changes state, which results in the output of IC_3 changing from low to high and the triggering of IC_{4a}. This means that all relays are reset after switch-on, immediately followed by the enabling of the relevant input. This entire process must be completed before the output of the preamplifier becomes active.

The circuit requires a power supply of about 15 V. The diagram shows how this may be derived from the ±7.2 V supply of the battery operated preamplifier.

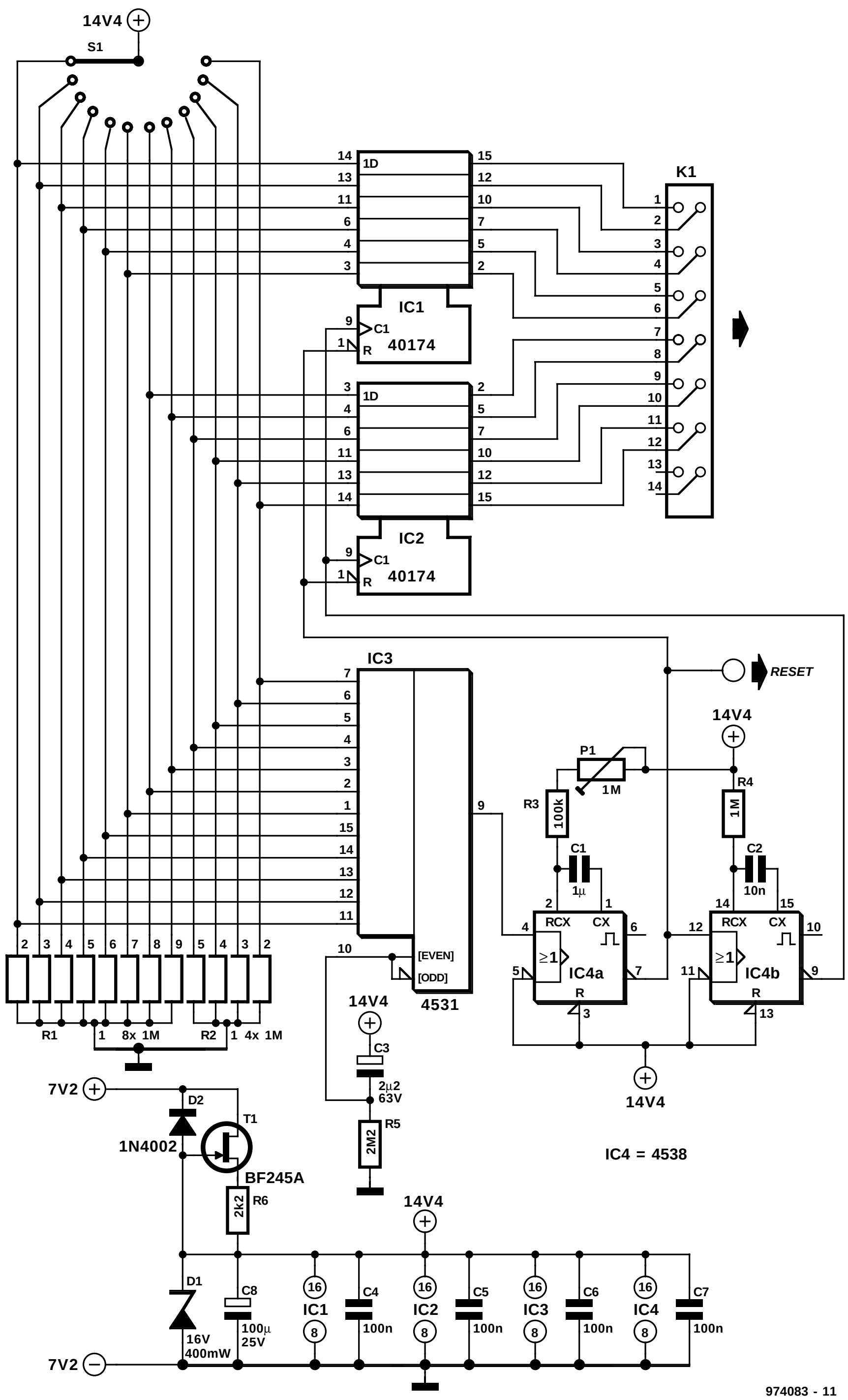
The ICs are protected against overvoltage by zener diode D_1. To ensure that the current through this diode is held within limits when the battery voltage is high, current source T_1 is provided in series with D_1.

When the supply voltage is lower than 15 V, the drop across R_6 and T_1 may be ignored, but when it is higher, the current is limited to about 400 µA.

The value of capacitor C_8 is purposely large since this component provides the energy required for changing over the inputs. This becomes clearer on reading the 'AF input module' elsewhere in this issue.

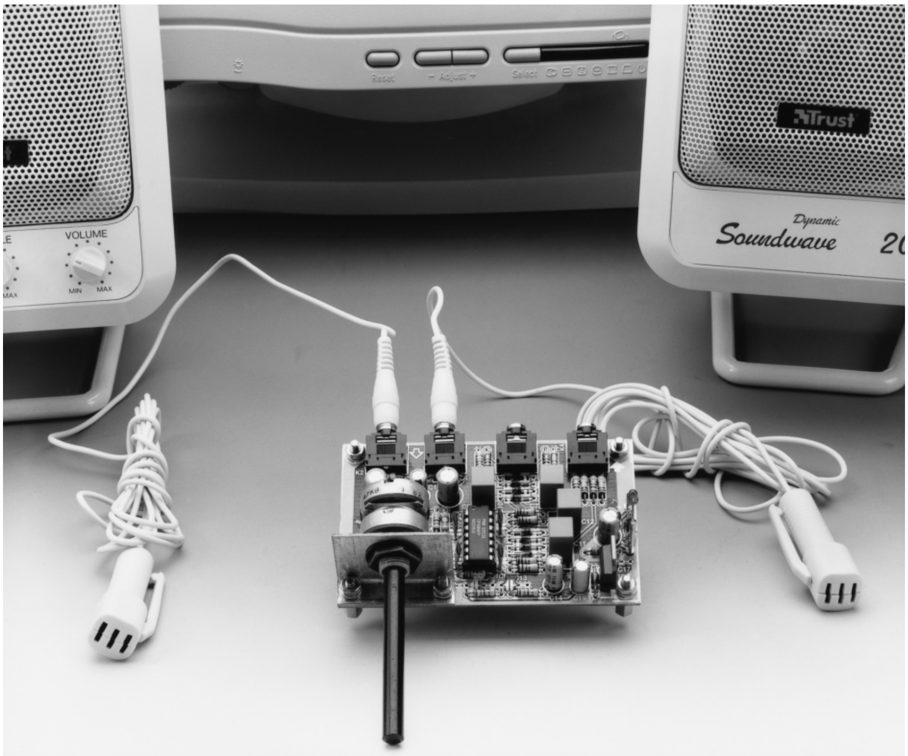
Finally, diode D_2 prevents C_8 being discharged via T_1.

[Giesberts – 974083]



stereo microphone input adaptor for PC

useful extension of sound card



There are occasions when it is desirable or even essential for a stereo microphone to be connected to a personal computer. Unfortunately, the microphone input of most sound cards used in PCs is monophonic. This article describes a simple adaptor to convert the mono(phonic) input into a stereo(phonic) input. It may also be used to provide a cassette deck with a stereo(phonic) microphone input.

Design by T.Giesberts

Nowadays, there is not the sharp dividing line of yesteryear between consumer audio, TV, video, and computer, equipment. In fact, today it is sometimes difficult to decide where one ends and the other begins. The audio installation may be used to reproduce the sound of a film on the video recorder; the CD-ROM drive of a PC may be used to play an audio CD; and the PC may be used for processing audio and video signals.

The PC can serve not only for the reproduction of the simple sounds that support certain software, but also for processing complex music signals. In that case, the audio signals are first quantized via the sound card and subsequently processed as desired. In fact, hard-disk recording is no longer a novelty.

Unfortunately, the microphone

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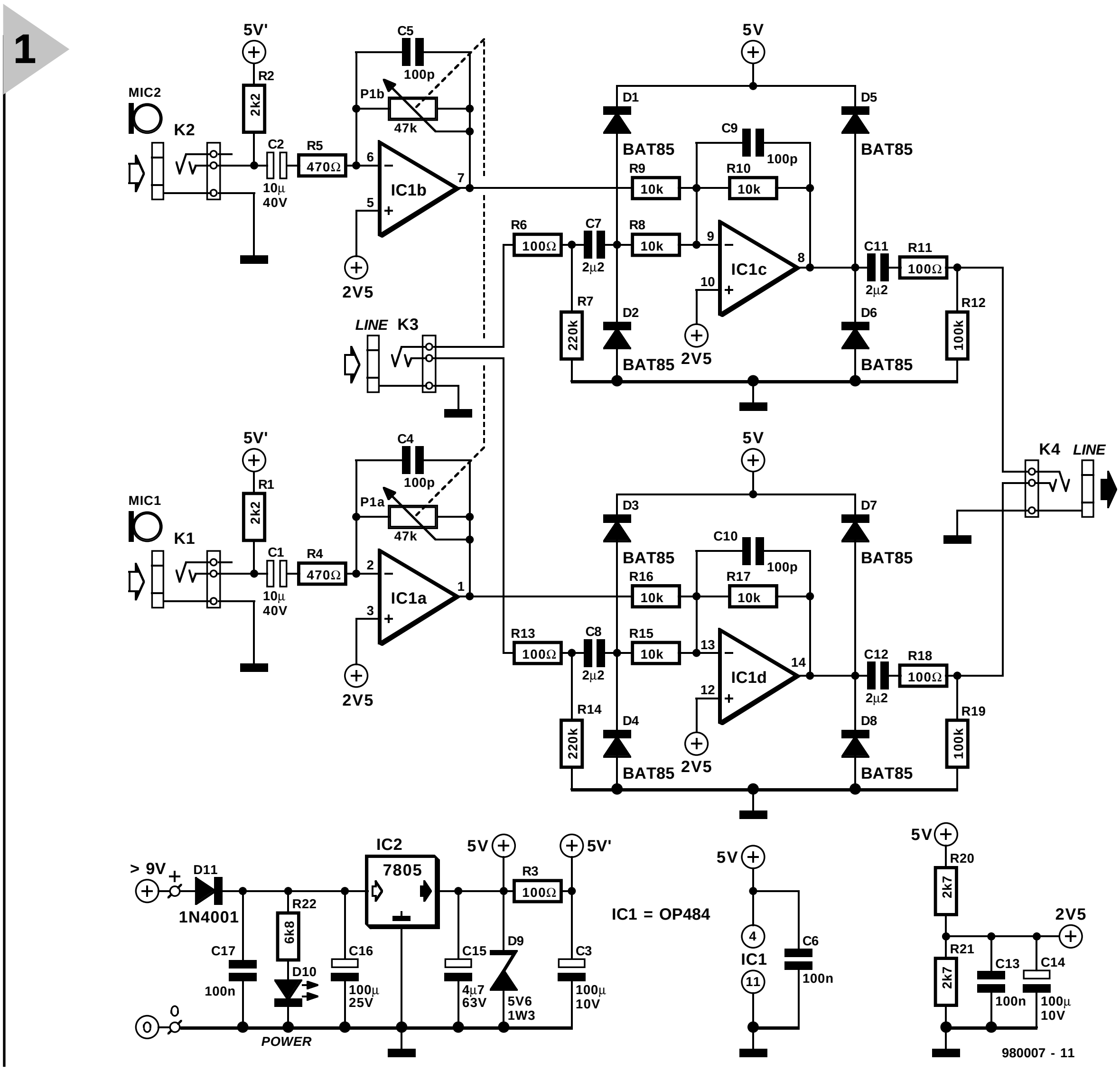

Figure 1. Circuit diagram of the stereo microphone adaptor for PCs.

input of most sound cards used in consumer PCs is monophonic. Luckily, however, many sound cards have a stereophonic line input. which can be converted to a stereo microphone input.

CIRCUIT DESCRIPTION

The conversion of a line input to a microphone input normally entails first of all raising the microphone output signal (a few millivolts to a few hundred millivolts) to line level (standard: 775 mV across 600 Ω designated 0 dBm, but in consumer equipment the r.m.s. level may range from 100 mV to 5 V). It is, however, convenient to retain the original line input function, and this is so in the present circuit.

In the circuit diagram in **Figure 1,** jack socket K_4 serves to connect the adaptor to the PC, socket K_3 is the 'new' line input, and sockets K_1 and K_2 form the stereophonic microphone input. The microphone output signals are amplified by operational amplifiers IC_{1a} and IC_{1b}, while IC_{1c} and IC_{1d} serve as adders and output stages.

Op amps IC_{1a} and IC_{1b} are straightforward inverting amplifiers whose amplification is determined by the ratio P_{1a}:R_1 (P_{1b}:R_2). In the prototype, this is ×23, which is sufficient for the electret microphones used.

Resistors R_1 and R_2 also provide the supply voltage for the FET impedance adaptor in these microphones. (FET = field-effect transistor).

Potentiometer P_1 serves to set the sensitivity of the microphone input or the level of the amplified microphone signal.

The configuration of the adders/output stages is similar to that of the preamplifiers, but their amplification is unity and the output impedance is rather higher.

Resistors R_6-R_7 and R_{13}-R_{14} ensure stable operation with unusual line signals.

Since the supply voltage is only 5 V, the line inputs and outputs are protected against overvoltage by diodes D_1–D_8. Zener diode D_9 makes certain that the supply voltage cannot exceed 5.6 V in any circumstances.

The supply voltage is obtained from a standard 9 V mains adaptor, which need not be regulated nor rated for high currents (the circuit draws only about 10 mA). Regulator IC_2 holds the output voltage steady at 5 V. This low voltage ensures that the sound card cannot be overdriven or damaged by overvoltage.

The amplifier stages are powered by half the supply voltage via potential divider R_{20}-R_{21}, which is decoupled by capacitors C_{13} and C_{14}.

The supply lines to the microphones are decoupled by network R_3-C_3.

Diode D_{11} protects the lines against polarity reversal, while D_{10} is the on/off indicator.

CONSTRUCTION

The adaptor is best built on the printed-circuit board shown in **Fig-**

Parts list

Resistors:
R_1, R_2 = 2.2 kΩ
R_3, R_6, R_{11}, R_{13}, R_{18} = 100 Ω
R_4, R_5 = 470 Ω
R_7, R_{14} = 220 kΩ
R_8, R_9, R_{10}, R_{15}, R_{16}, R_{17} = 10 kΩ
R_{12}, R_{19} = 100 kΩ
R_{20}, R_{21} = 2.7 kΩ
R_{22} = 6.8 kΩ
P_1 = 47 kΩ stereo, logarithmic potentiometer for board mounting

Capacitors:
C_1, C_2 = 10 µF, 40 V, bipolar, radial
C_3, C_{14} = 100 µF. 10 V, radial
C_4, C_5, C_9, C_{10} = 100 pF
C_6, C_{13}, C_{17} = 0.1 µF
C_7, C_8, C_{11}, C_{12} = 2.2 µF, metallized polyester (MKT)
C_{15} = 4.7 µF, 63 V, radial
C_{16} = 100 µF, 25 V, radial

Semiconductors:
D_1–D_8 = BAT85
D_9 = zener diode, 5.6 V, 1.3 W
D_{10} = LED, high-efficiency
D_{11} = 1N4001

Integrated circuits:
IC_1 = OP484FP (Analog Devices) (see text)
IC_2 = 7805

Miscellaneous:
K_1–K_4 = stereo jack socket, 3.5 mm, for board mounting
PCB Order no. 980007-1 (see Readers Services section towards the end of this issue)

ure 2. All jack sockets are at one side of the board and the volume/sensitivity control at the opposite side.

When the (straightforward) construction has been completed and the correct operation of the adaptor has been verified, the adaptor should be fitted in a suitable enclosure. This is preferably a small metal case to which the earth of the circuit is strapped via one central point (near one of the jack sockets).

OPTIONAL MODIFICATIONS

The amplification of the circuit specified earlier is sufficient for the (electret) microphones used with the prototype. If desired (or required), it may be raised by lowering the value of R_9 and R_{16}, but not below 2 kΩ.

The operational amplifier used in the IC_1 position is a Type OP484 from Analog Devices. This device combines a rail-to-rail input and output with a very low noise factor and a range of supply voltages that extends to well below that of most other types. Nevertheless, other types of op amp, such as the TLC272, may also be used.

[980007]

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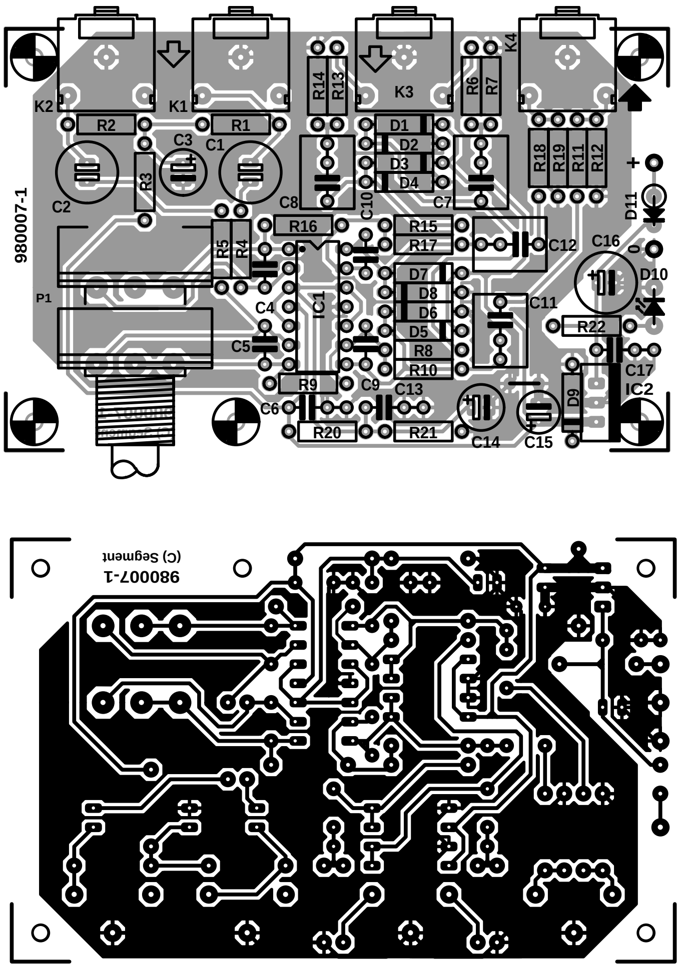

Figure 2. The various jack sockets and potentiometer can be fitted directly to the printed-circuit board.

Figure 3. Photograph of the completed prototype stereo microphone adaptor board.

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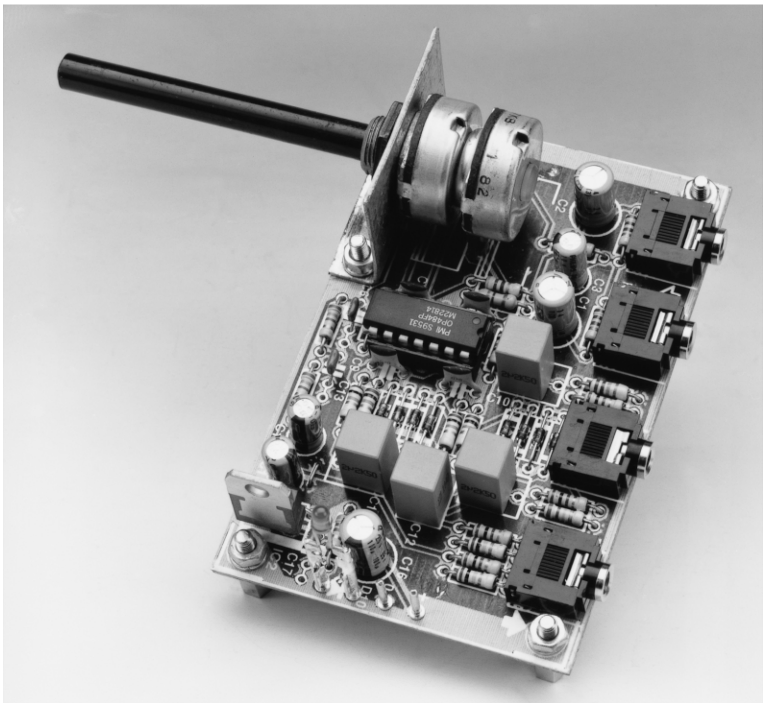

Contents

AVC for PCs

limits differences in sound level

An annoying phenomenon (not restricted to PCs) is that each and every programme that produces sound does so at a different level. This means almost constant adjustment of the volume control to ensure audibility of one programme and protection of the ear drums with another one. The control circuit described in this article is designed to obviate this nuisance: it constantly monitors the signal even at the output of the soundcard and adjusts it when required. Use of the circuit is not restricted to PCs; it may also be used as a dynamic limiter in existing audio equipment.

Design by T. Giesberts

Brief specification

Power output	*1.2 W*
Maximum input	*1 V*
Compression	*10:1*
Supply line	*12 V, 6 VA*
Output load	*8 Ω (LSP); 10 kΩ (line)*
Input sensitivity	*280 mV (gain line in to out = 0 dB; distortion at output = 1%)*
	120 mV (gain line in to out = max; distortion at output = 1%)
Line in to LSP out (input voltage = 200 mV)	
THD+N	*0.25% (2×0.5 W)*
Signal-to-noise	*70 dB for 0.5 W output at maximum gain*
Channel separation	*>45 dB*
Line in to line out (input voltage = 200 mV; no loudspeaker connected)	
THD+N	*0.047%*
Signal-to-noise	*80 dB*
Channel separation	*>73 dB*

In general, the signal levels in current audio equipment are equalized and standardized (although there are still some exceptions). Software manufacturers do not seem to know or care about this. Anyone who has ever opened two different sound programs will know of the quite different levels various effects often have. This is obviously an annoying situation and one which makes the constant adjusting of the volume control a necessity.

The present circuit offers a solution to this problem. It consists of a dynamic compressor with a control range of 10:1 which ensures that very loud and very soft sound passages are attenuated or amplified respectively. This results in a much narrower dynamic range of audio signals which makes adjusting the volume control a much less frequent necessity. It proves that something that appears difficult in software can be easily achieved by a small electronic circuit.

DESIGN

The block diagram of the automatic volume control (AVC) circuit is shown in **Figure 1.** The stereo audio signal at the output of a sound card used in a multimedia PC is applied to the line input. The active part of the circuit consists of two integrated amplifiers that contain a variable preamplifier and a compact output amplifier.

The signal from the output amplifier is freed from any direct voltage and then applied to a discrete rectifier. After the rectified signal has been processed, it is used to control the amplification factor of the preamplifiers. The control circuit is based on a number of discrete operational amplifiers. The design is an OR-type, so that the sound channel (lefthand or righthand) with the highest peak signal level determines the amplification factor of

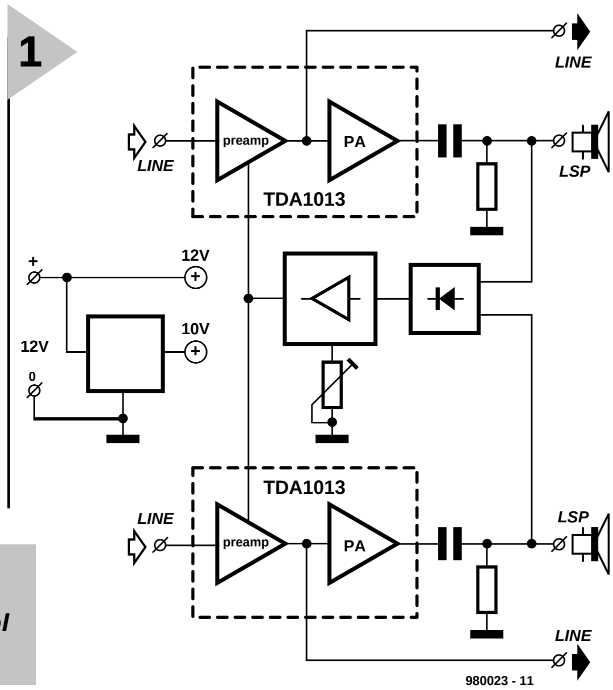

Figure 1. Block diagram of the automatic volume control circuit for PCs.

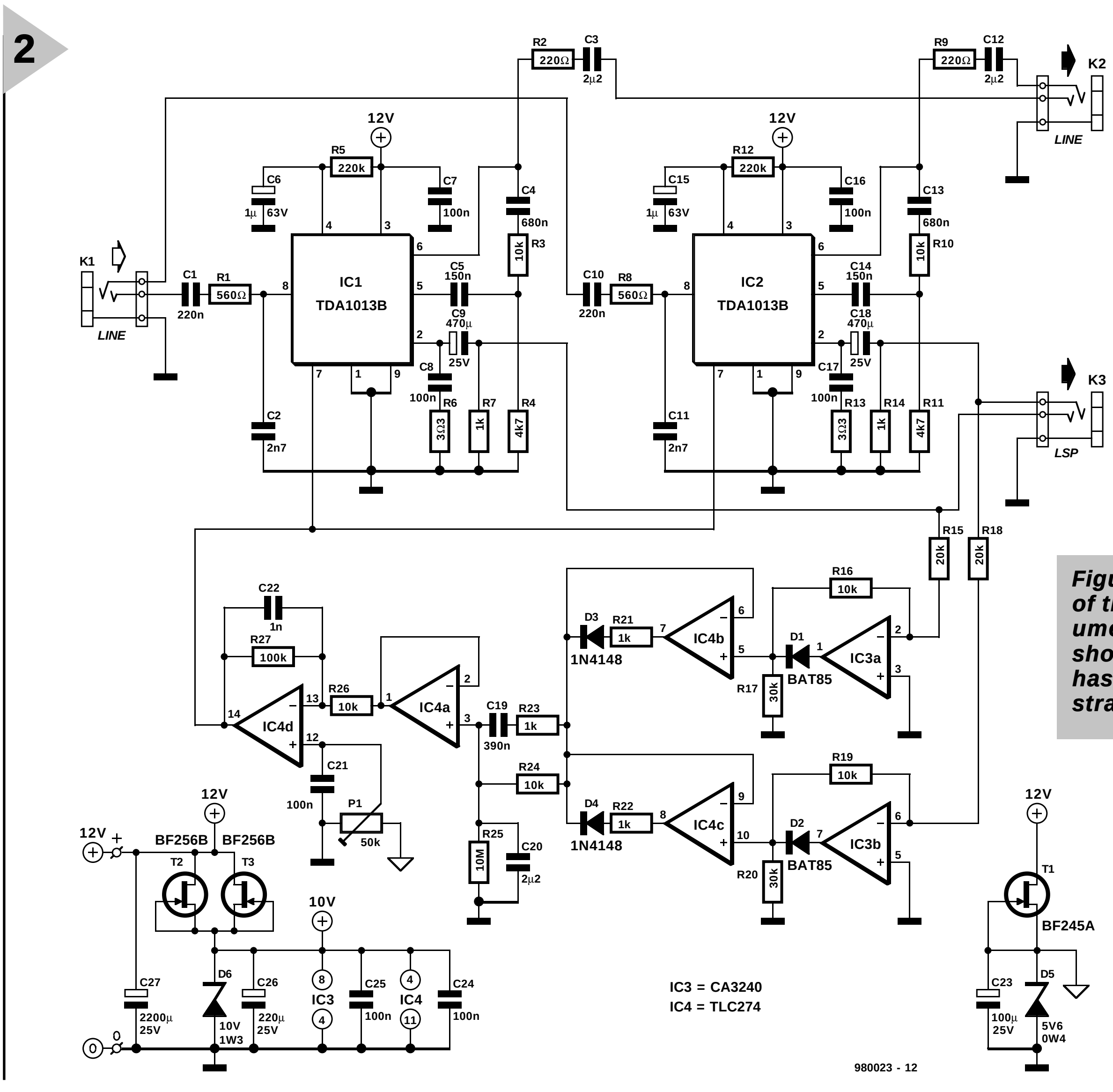

Figure 2. The diagram of the automatic volume control for PCs shows that the circuit has been kept straightforward.

Parts list

Resistors:
R_1, R_8 = 560 Ω
R_2, R_9 = 220 Ω
R_3, R_{10}, R_{16}, R_{19}, R_{24}, R_{26} = 10 kΩ
R_4, R_{11} = 4.7 kΩ
R_5, R_{12} = 220 kΩ
R_6, R_{13} = 3.3 Ω
R_7, R_{14}, R_{21}, R_{22}, R_{23} = 1 kΩ
R_{15}, R_{18} = 20 kΩ
R_{17}, R_{20} = 30 kΩ
R_{25} = 10 MΩ
R_{27} = 100 kΩ
P_1 = 50 kΩ (47 kΩ) preset

Capacitors:
C_1, C_{10} = 0.22 µF
C_2, C_{11} = 0.0027 µF
C_3, C_{12}, C_{20} = 2.2 µF metallized polyester (MKT), pitch 5 or 7.5 mm
C_4, C_{13} = 0.68 µF
C_5, C_{14} = 0.15 µF
C_6, C_{15} = 1 µF, 63 V, radial
C_7, C_8, C_{16}, C_{17}, C_{21}, C_{24}, C_{25} = 0.1 µF
C_9, C_{18} = 470 µF, 25 V, radial
C_{19} = 0.39 µF
C_{22} = 0.001 µF
C_{23} = 100 µF, 25 V, radial
C_{26} = 220 µF, 25 V, radial
C_{27} = 2200 µF, 25 V, radial

Semiconductors:
D_1, D_2 = BAT85
D_3, D_4 = 1N4148
D_5 = zener diode 5.6 V, 400 mW
D_6 = zener diode 10 V, 1.3 W
T_1 = BF245A
T_2, T_3 = BF256B

Integrated circuits:
IC_1, IC_2 = TDA1013B
IC_3 = CA3240E
IC_4 = TLC274CN

Miscellaneous:
K_1–K_3 = 3.5 mm stereo audio socket for board mounting
PCB Order no. 980023-1 (see Readers Services towards the end of this issue)

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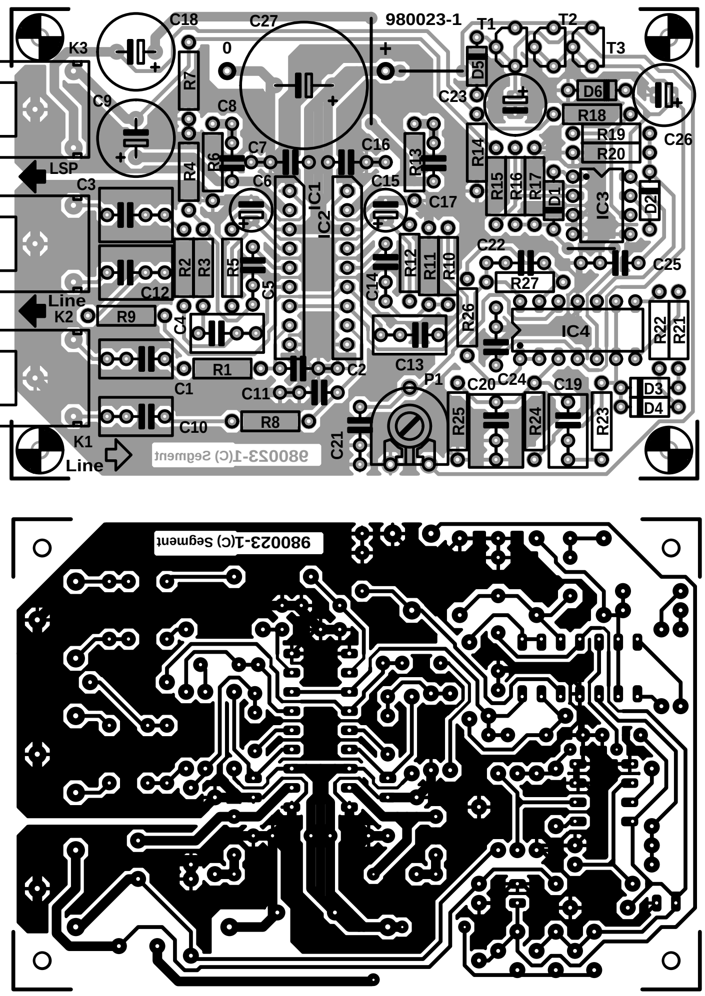

Figure 3. The printed-circuit board for the automatic volume control.

the stereo preamplifier.

The values of various components in the control circuit are chosen to ensure a fast attack time and a long release time. This ensures that short-duration signal peaks are effectively suppressed, whereupon the circuit recovers (relatively) slowly from the damping action.

Power for the circuit is derived from a standard 12 V mains adaptor.

CIRCUIT DESCRIPTION

In the circuit diagram in **Figure 2**, the preamplifier-output amplifier combination is contained in IC_1 and IC_2. This type of IC is a compact 4 W audio amplifier with integral voltage-controlled volume control. The range of the logarithmic volume control is 80–90 dB with control voltages between 6.5 V (+5 dB) and 2.0 V (–80 dB).

The control voltage is applied to pin 7 of the IC. The line output is at pin 6, which is linked via a capacitor to pin 5, the input of the output amplifier..

The circuit has three stereo terminals: line in, line out, and power out.

The power output is 2.5 W for a loudspeaker impedance of 8 Ω and a supply line of 18 V, which is sufficient for most applications.

The analogue input signal at the line input, K_1, is applied to pin 8 of IC_1,raised in the preamplifier and output via pin 6. The transfer between pins 8 and 6 depends on the control voltage at pin 7.

The line signal is attenuated and its level made suitable for inputting to the output amplifier by networks R_3-R_4 and R_{10}-R_{11}. Assuming a supply line of 12 V, the output amplifier is driven fully ($P_{o(max)}$ = about 1.2 W into 8 Ω) by an input signal of 90 mV.

RC networks are provided at the inputs (R_1-C_1-C_2 and R_8-C_{10}-C_{11}) and the line outputs (R_2-C_3 and R_9-C_{12}).

The output amplifier outputs are provided with large electrolytic capacitors, C_9 and C_{18}.

The supply lines are decoupled by C_7 and C_{16}.

Filters R_6-C_8 and R_{13}-C_{17} ensure that the amplifiers remain stable at high frequencies.

RECTIFICATION AND REGULATION

The audio signal to be rectified is taken from the loudspeaker terminals and applied to IC_{3a} and IC_{3b}. The following description is based on IC_{3a}.

Negative signals are inverted by the op amp and amplified by a factor that depends on the ratio R_{15}:R_{16}. In the present circuit, this is –2, that is, attenuation. With positive signals, the op amp is overdriven and its output negative. Diode D_1 is then cut off and half the input voltage is available at its cathode [$R_{17}/(R_{15}+R_{16}+R_{17})$]. This means that the op amp behaves as a full-wave rectifier/amplifier, whose amplification is the same (0.5) for both

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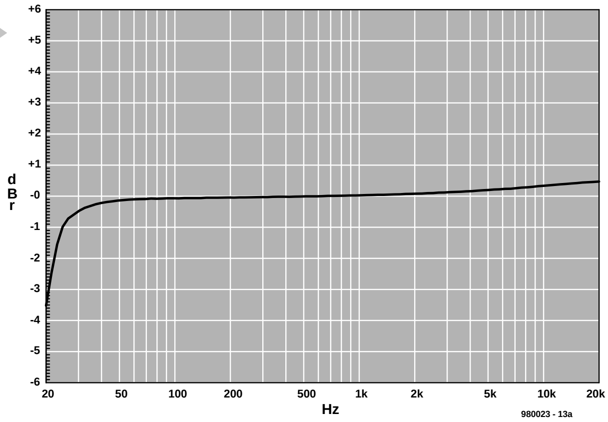
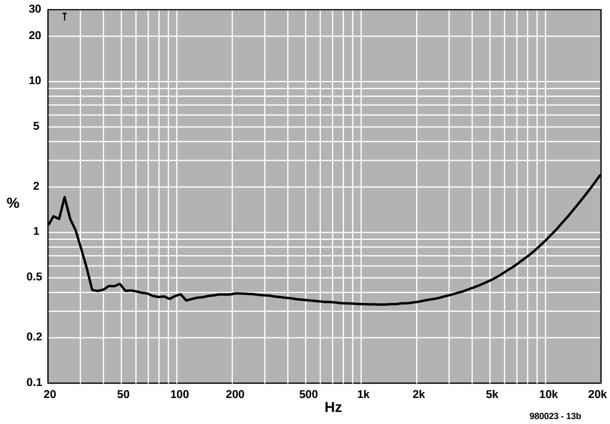




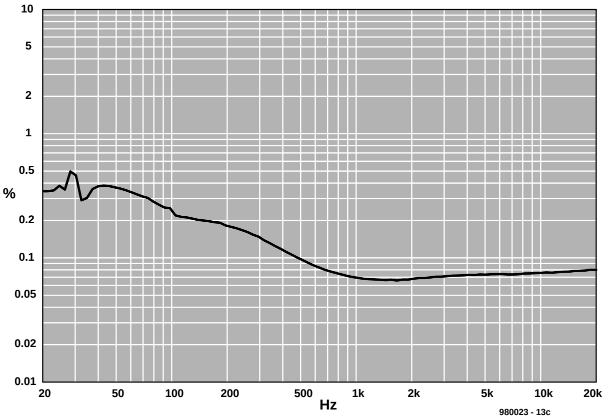


Figure 4. Some graphs to show a few of the electrical properties of the AVC. (a) is the frequency response curve measured between the line input and loudspeaker input terminals. (b) is the distortion characteristic measured at the loudspeaker output terminals (5 W into 8 Ω). (c) is the distortion characteristic measured at the line output terminals terminated into 10 kΩ.

halves of the input signal.

Operational amplifiers IC_{4a} and IC_{4b} are half-wave rectifiers whose outputs are interlinked by diodes D_3 and D_4. Because of these diodes, the output with the highest potential determines the extent to which capacitor C_{20} is charged via resistor R_4. Network R_{23}-C_{19} has been added to ensure that fast signal fluctuations are passed on very rapidly.

Capacitor C_{20} is discharged slowly via resistor R_{25}, so that the control circuit returns to its default setting when no or a smaller input has been applied for some time. The potential across C_{20} is buffered by IC_{4a}, while IC_{4d} ensures that the (fixed) default level is added to the signal. The resulting control signal is applied to the control input (pin 7) of IC_1 and IC_2.

With component values as specified, the compression is 10:1; in other words, a 20 dB change at the input results in a 2 dB change at the output.

The setting of P_1 depends on the signal level at the input of the circuit. Since this level varies largely from one sound card to another, the design provides a wide control range.

SUPPLY LINES

As mentioned earlier, the circuit is powered by a standard 12 V mains adaptor, which is applied directly to the output amplifier. All other circuit elements are supplied with a regulated 10 V potential. This voltage is produced with the aid of current source T_2-T_3 and zener diode D_6.

The reference voltage of 5.6 V is produced with the aid of current source T_1 and zener diode D_5.

CONSTRUCTION

The circuit is best built on the printed-circuit board shown in **Figure 3** (see Readers Services towards the end of this issue). Start the construction with placing audio sockets K_1–K_3, the three wire bridges, and all solder pins, and follow these with first the passive components, and then the active ones. Mind the polarity of the electrolytic capacitors, diodes, transistors, and ICs.

After it has been fitted, set the preset to minimum volume (anticlockwise).

Solder the output leads from the standard mains adaptor from which power is derived to the relevant pins on the board. If the board is to be housed in an enclosure, a plug-and-socket arrangement should be used for linking the output from the adaptor to the board.

Check that the output voltage of the adaptor does not rise above 18 V with small loads.

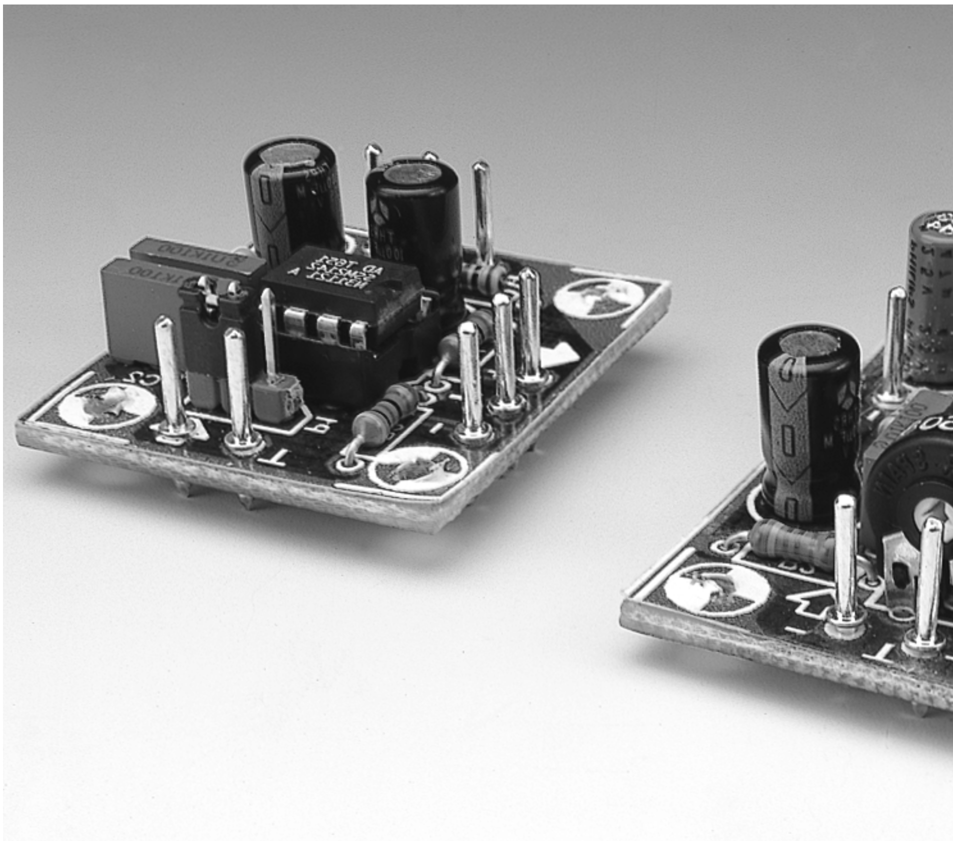
When all is connected, the circuit can be tested. Passive loudspeakers may be linked directly to the LSP output terminals, but active ones should be connected to the line output terminals.

Finally, connect a sound source, for instance, the line output of a sound card or the output of a Walkman™ to the input of the circuit and adjust P_1 for the desired volume. From then on, any fluctuations in the signal input level will be minimized automatically.

[980023]

Balanced/unbalanced converters for audio signals

for high-end applications



In the Quest For Quality, a quasi-metaphysical activity zestfully practised by countless audiophiles, the use of 'balanced' (or symmetrical) signals is believed to contribute considerably to the end result. In this article we take pleasure in presenting balanced/unbalanced converters for audio signals. The designs, we are convinced, offer a solution to many problems you may stumble on when it comes to conveying small audio signals over large distances.

Design by J. F. Brangé

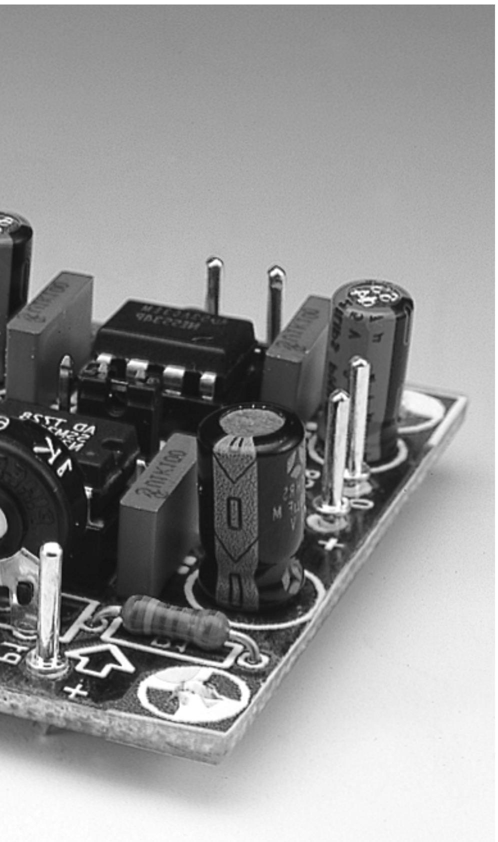
First things first. Let's recall that an unbalanced (asymmetrical) signal is defined as existing with respect to the ground line in a circuit. Consequently, conveying an unbalanced audio signal from one preamplifier stage to another by way of a cable may pose various problems including parasitics and radiation which degrade the quality of the audio signal. The use of shielded cable is an insufficient remedy, particularly when the signal source supplies low signal levels (say, a couple of millivolts).

The panacea in these cases is to make the signal balanced, that is, floating with respect to ground. Doing so allows a weak source signal to be conveyed over long distances (if necessary) without it being 'modified' underway. Sure, the problem of parasitics remains. However, the balanced signal being converted to unbalanced again at the input of the 'receiver', noise is effectively cancelled out by the differential effect. The floating signal on the two wires arrives at the inputs of the differential circuit. Whereas the wanted audio signal arrives with opposite phases on the two wires, any noise picked up by the symmetrical cable will have the same phase on the two wires. Consequently, this noise is effectively eliminated by the subtracting operation of the differential circuit.

There are, without doubt, many circuits which enable an audio signal to be converted from balanced to unbalanced and the other way around. The operational amplifier (opamp) lends itself quite well to this kind of operation. Provided you use quality audio opamps, ample results will be obtained. However, a couple of precautions should be taken to prevent degrading the performance that may be achieved in theory. One of these conditions is the use of 'hand picked' resistors with a tolerance of 0.1 per cent or better.

Unbalanced-to-balanced conversion

There is a modern solution to this problem. Analog Devices has developed integrated circuits which are totally geared to this application. These ICs boast internal resistors which are laser-trimmed to a precision of 0.0001%! Not surprisingly, these ICs achieve 'professional-grade' performance as far as noise rejection, para-



sitics suppression and distortion are concerned. These ICs now being relatively well distributed in Europe, we have few hesitations about presenting you practical circuits for a stereo application. Obviously, the two stereo channels being identical, it will be sufficient to describe only one of these.

The circuit diagram of the unbalanced-to-balanced converter is shown in **Figure 1**. The SSM2142 opamp from Analog Devices is a buffer/amplifier with an internal differential output driver. Its main function is to convert an unbalanced input signal into a high-level balanced signal. Based on an electronically balanced cross-coupled chip topology, the SSM2142 comes close to achieving the performance of balancing circuits that make use of a transformer for line driving. As a matter of course, the IC has the advantage of a much smaller footprint than that of a transformer, while offering compara-

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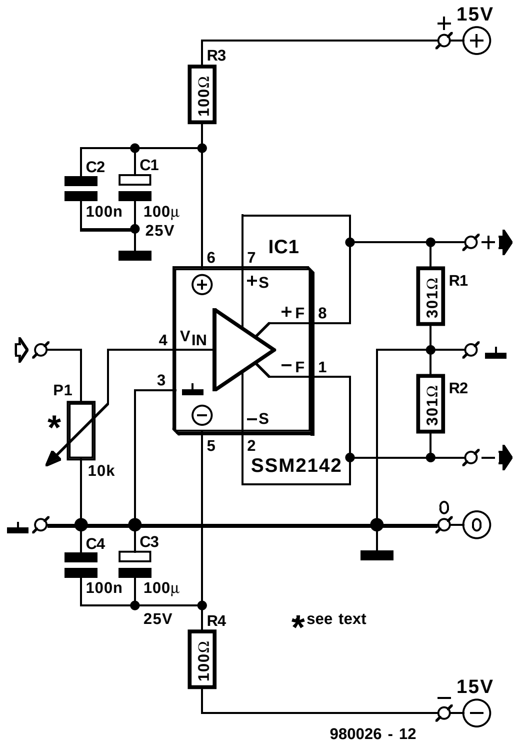

ble common-mode rejection. Those of you who are interested in the chief technical specs of the SSM2142 will no doubt find this month's *Datasheet* pages of particular interest.

The input signal is applied to the chip via a 10-kΩ preset, P1, whose function is to adjust the output signal level while also matching the IC input impedance recommended by the manufacturer. The preset may, of course, be replaced by either a potentiometer with the same value, or a 3-way pin-header on to which a jumper is

Figure 1. Circuit diagram of the unbalanced-to-balanced converter based on the SSM2142 from Analogue Devices.

Figure 2. The balanced-to-unbalanced converter is designed around the SSM2141 and an NE5534 buffer opamp.

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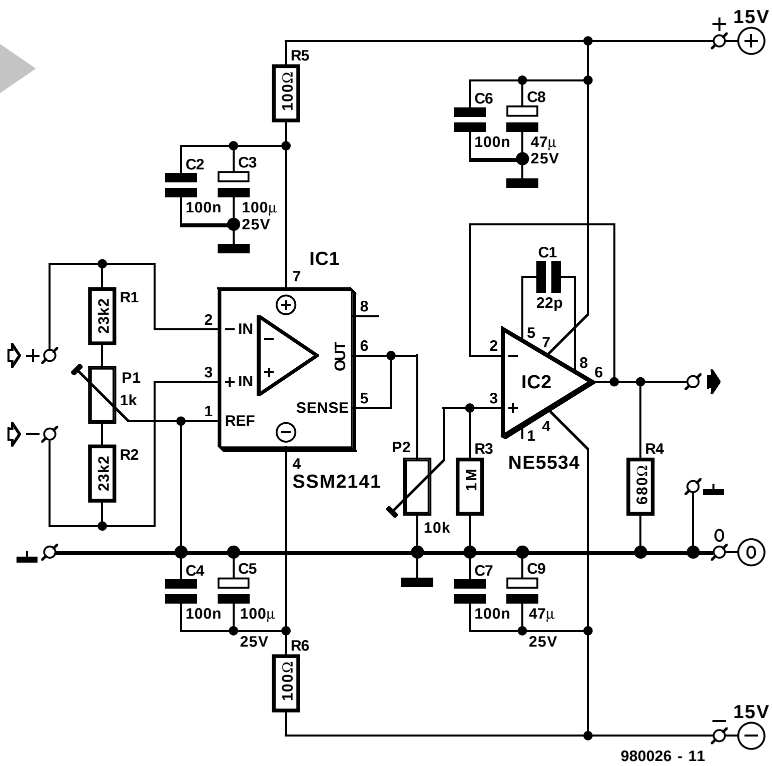

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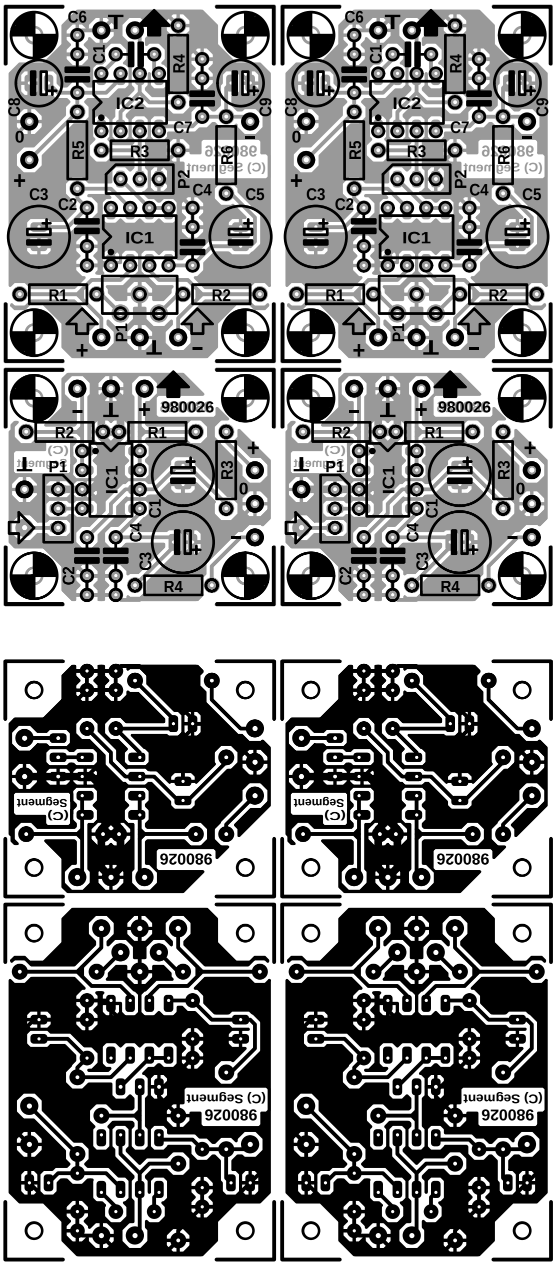

Figure 3. Copper track layout and component mounting plan of the printed circuit board designed for two balanced-to-unbalanced converters and two unbalanced-to-balanced converters (board available ready-made through the Readers Services).

installed which takes the signal from the 'input' pin to the centre pin. This is the solution we adopted. The output is also simple: pin 8 of the SSM2142 supplies in-phase (+) output signal, while pin 1 supplies the inverted (–) signal. Since both outputs are loaded with a 301-Ω resistor to ground, an output impedance of about 600 Ω is created.

The SSM2142 is protected against parasitic signals arriving by way of the supply lines. This is achieved by connecting elementary RC filter networks comprising of R3-C1-C2 and R4-C3-C4 to the respective supply pins of the SSM2142. The output of the circuit supplies an audio signal which should be worthy of the very best home-brew audio projects.

A final word or two about the power supply: although a ±12 V symmetrical supply will be fine in many cases, we recommend using ±15 V as shown in the circuit diagram because that value results in an improved dynamic range. The absolute maximum supply voltage is ±18 V.

BALANCED-TO-UNBALANCED CONVERTER

For this purpose you need the SSM2141. This IC, a direct relative of the SSM2142, is an integrated differential-amplifier designed to receive balanced 'line' signal levels in audio

Figure 4. Finished prototype of each of the converters.

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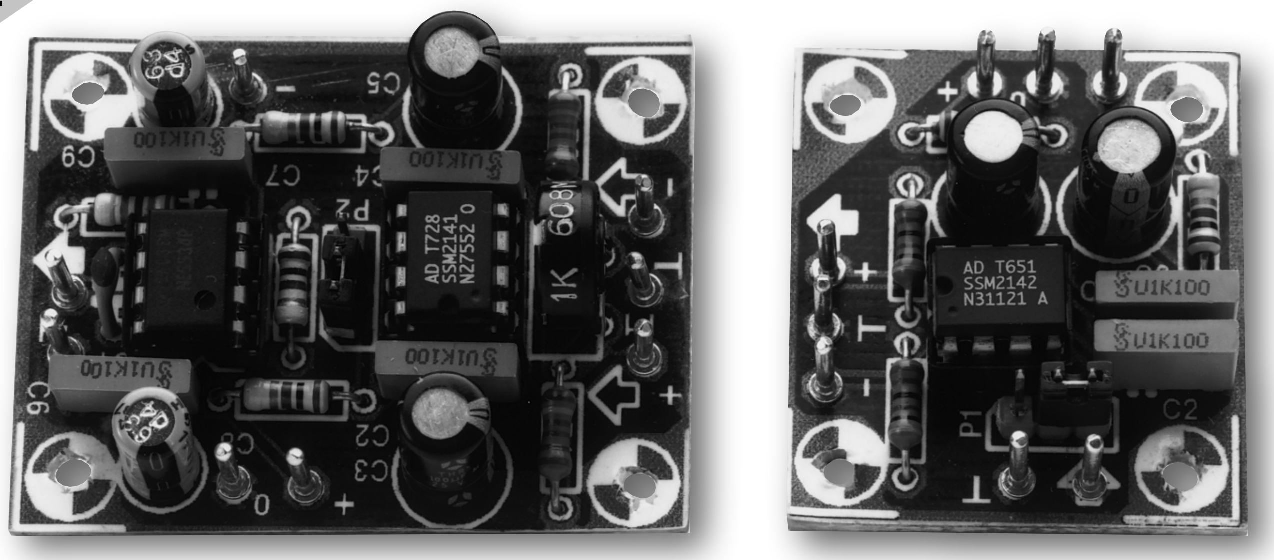

COMPONENTS LIST

Printed circuit board: order code 980026-1.

Balanced-to-unbalanced converter

Resistors:
R1,R2 = 23kΩ22 1%
R3 = 1MΩ
R4 = 680Ω
R5,R6 = 100Ω
P1 = 1kΩ preset vertical
P2 = 10kΩ logarithmic potentiometer (may be replaced by jumper)

Capacitors:
C1 = 22pF
C2,C4,C6,C7 = 100nF
C3,C5 = 100µF 25V radial
C8,C9 = 47µF 25V radial

Semiconductors:
IC1 = SSM2141 (Analog Devices)
IC2 = NE5534 (Philips Semiconductors)

Unbalanced-to-balanced converter

Resistors:
R1,R2 = 301Ω 1%
R3,R4 = 100Ω
P1 = 10kΩ logarithmic pot (or jumper)

Capacitors:
C1,C3 = 100µF 25V radial
C2,C4 = 100nF

Semiconductor:
IC1 = SSM2142 (Analog Devices)

circuits requiring high noise immunity and common-mode noise rejection. This IC achieves a typical CMR (common-mode rejection) spec of 100 dB. By comparison, an opamp with four regular resistors around it will be hard pressed to achieve a CMR rating of anything over 40 dB or so, which is by no means enough for high-end audio designs. Let's cast a look at **Figure 2** which shows the schematic of this subcircuit. The resistor networks between the SIG+ (pin 3) and SIG– (pin 2) inputs of the SSM2141 fix the input impedance at about 47 kΩ. Preset P1 (1 kΩ) allows the CMR value to be fine-tuned (see also further on). This component is optional, however, and may be omitted. As indicated by the component overlay of the balanced-to-unbalanced converter, it may be replaced by wire links. This was also done on our prototype. Note, however, that the source impedance has to be perfectly controlled, as the slightest imbalance of the source resistance will reduce the achievable CMR value. For example, a difference of just 5 Ω is punished with a CMR increase of no less than 20 dB.

The output signal of the SSM2141 is applied to an NE5534 voltage follower by way of a 10-kΩ preset. The (low-impedance) output of the NE5534 should be able to drive almost any preamplifier input. The remarks on the supply filtering of the SSM2142 also apply to the SSM2141.

If used, the CMR fine-tuning preset has to be adjusted with the aid of a differential input signal. What better way to generate such a signal than use the SSM2142? Apply a 50-Hz, 100-mV signal to the input of the 2142. Connect its output signal to the 2141. Next, tweak P1 for the smallest possible signal at the output. This setting corresponds to the best possible CMR. Those of you who do not have an oscilloscope (or access to one) may replace the 1-kΩ preset by two wire links, as mentioned earlier. If you can get hold of two 23.2-kΩ resistors with a tolerance of 0.1%, no adjustment should be necessary.

As already mentioned, the circuits are powered by a ±15-V symmetrical supply. Current consumption being very modest indeed, you can make do with a mains adaptor with stabilized ±15 V outputs.

Construction

As you can see from the artwork in **Figure 3**, a printed circuit board was designed for the two converters. The artwork comprises the copper track layout and the component mounting plan (overlay). The PCB design for each converter is duplicated so you need just this one board for a stereo application.

As a matter of course, it is best to start by separating the four small boards. The two smaller boards are used to build the unbalanced-to-balanced converters, while the boards with two IC sockets on them are intended for the balanced-to-unbalanced converters. Neither of these circuits should present undue difficulty when populating the boards. The only points to be made here are to observe the polarity of the electrolytic capacitors and the orientation of the integrated circuits when they are inserted into their sockets. Also be sure not to mix up the two 8-pin integrated circuits on the balanced-to-unbalanced converter board.

The large unetched copper areas on the boards acts as ground planes which help to make the circuits immune to noise and other stray signals.

Having finished the construction of the converter boards you may mount them in small boxes, and wire them up to the external parts. The input and output connectors may be mini-DIN types of which only three pins are used. The unbalanced-to-balanced converter is best located close to the signal source. Its complement, the balanced-to-unbalanced converter, will typically be installed near the 'receiver'. Note the connections: the outputs of the unbalanced-to-balanced converter are the mirror-image of the inputs of the balanced-to-unbalanced converter.

The circuits should function spot-on. The photograph in **Figure 4** should allow you to compare your own efforts at building the circuits with those of our engineering laboratory. In particular, you should be able to detect missing components immediately in the (unlikely) case of a problem.

A final note aimed at those with a keen interest in figures: Below are the very encouraging results of exhaustive measurements on a pair of these converters built up in our design lab:

- unbalanced-to-balanced converter: THD (total harmonic distortion) between 0.0008% and 0.0015% from 20 Hz to 20 kHz.
- balanced-to-unbalanced converter: THD between 0.0008% and 0.0011% from 20 Hz to 20 kHz; CMR between –140 dB and –70 dB from 20 Hz to 20 kHz.

All measurements were made with an Audio Precision test system.

(980026-1)

clipping indicator

for compact disc

A little while ago a reader wrote to say that he had found overdrive on some of his compact discs. This sort of news comes of course like a thunderbolt since it is assumed by most people that compact discs are examples of the quality of today's digital technology. The first reaction to such an allegation is one of outright disbelief or at least scepticism. Moreover, it has been alleged by other readers that several producers have admitted (sic!) to overdriving, that is clipping, of CDs at the request of the relevant artists. Be that as it may, it was reason enough to design an indicator to bring overdrive to light and help the consumer in his/her quest not to buy flawed CDs*.

*It should be noted that overdrive on a CD is not a legal reason for asking your money back.

Designed by T. Giesberts

Most people will not believe that there are CDs that are overdriven by the producer: they generally assume that manufacturers know what they are doing and supply discs that are technically correct within the confines of modern digital technology. If the experiences of some of our readers are accepted, this may not always be true.

One reader wrote to say that he had noticed that some CDs in his collection sounded 'less than perfect' and others even 'downright poor'. Since he thought that his ears were playing him tricks, he decided to check the level with a VU (visual unit). To his surprise he found that the level varied around 0 dB. A surprise, indeed, for the level on a CD should reach 0 dB only during very brief peaks in the signal. The average signal strength should be not less than 6 dB and preferably 10–12 dB below 0 dB.

In view of these findings, our reader decided to take his investigation a little further and connected an oscilloscope to the output of his CD player. This showed that on certain CDs the signal was clipped; on one or two, the clipping led to 'audible distortion'. **Fig-**

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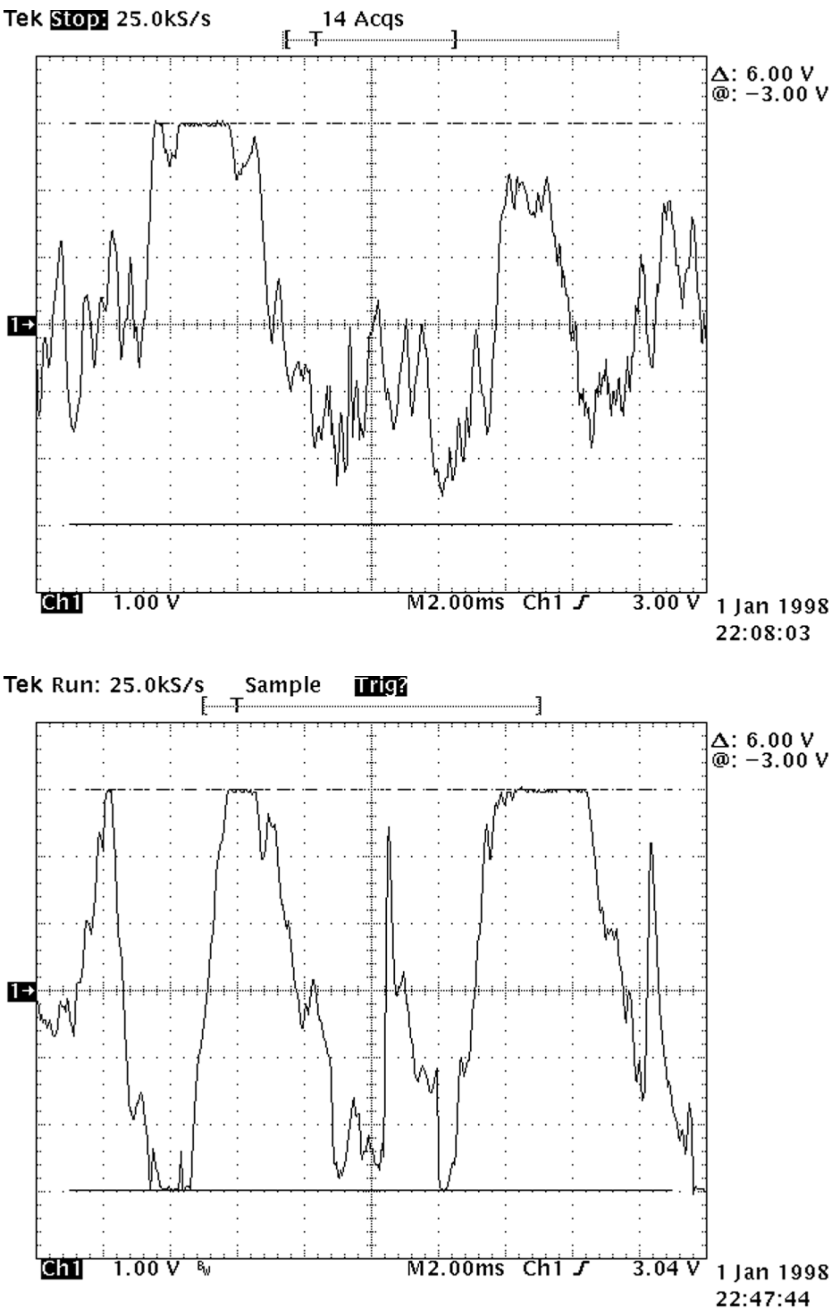
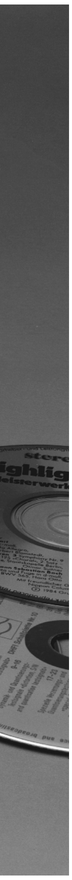


Figure 1. Two clear cases of overdrive allegedly measured on modern compact discs (Sony Music 099748 393227/1996 and 099748 698421/1997).



ure 1 shows a few examples (not necessarily the worst!).

Apart from leading to distortion, clipping also results in another phenomenon. Since the average signal strength is too high, the dynamic range of the music is reduced, so that the reproduced sound is much too flat, which can easily lead to ‘listening fatigue’.

COINCIDENCE?

Could these findings be coincidence? It is hard to say, but evidence from other readers and our own measurements seem to indicate that there are CDs on which the signal has been purposely overdriven. [The allegation by another reader that two producers admitted to him that they sometimes used overdrive on CDs at the request of the relevant artists seems far-fetched – because it would be contrary to their commercial acumen – but not impossible. Also, the reason given by these producers that they ‘dared not go against the wishes of these artists’ seems highly suspect. No artist is bigger than a bona fide recording studio. Editor].

DETECTION

What can the consumer do to avoid buying a flawed disc? After all, a CD cannot be repaired or enhanced. Overdrive used in the recording studio cannot even be eradicated during manufacture of the disc.

The only thing a consumer can do is not to buy the suspect CD. But how is he/she to detect that a certain CD suffers from overdrive? Listening to it at the retailer’s premises normally does not indicate anything awry, but once it is played on a good-quality installation at home a deficiency may come to light.

The solution appears to be a small portable indicator that can be taken to the retailer, assuming that it is possible to connect it to the retailer’s playback equipment.

What should the indicator react to? It is clear that on a good-quality CD the 0 dB level will be reached only during short high-signal peaks. If the 0 dB level is sustained for more than a fraction of a second, there may be reason to be suspicious. Consequently, the indicator is designed so that an LED lights when two or more samples

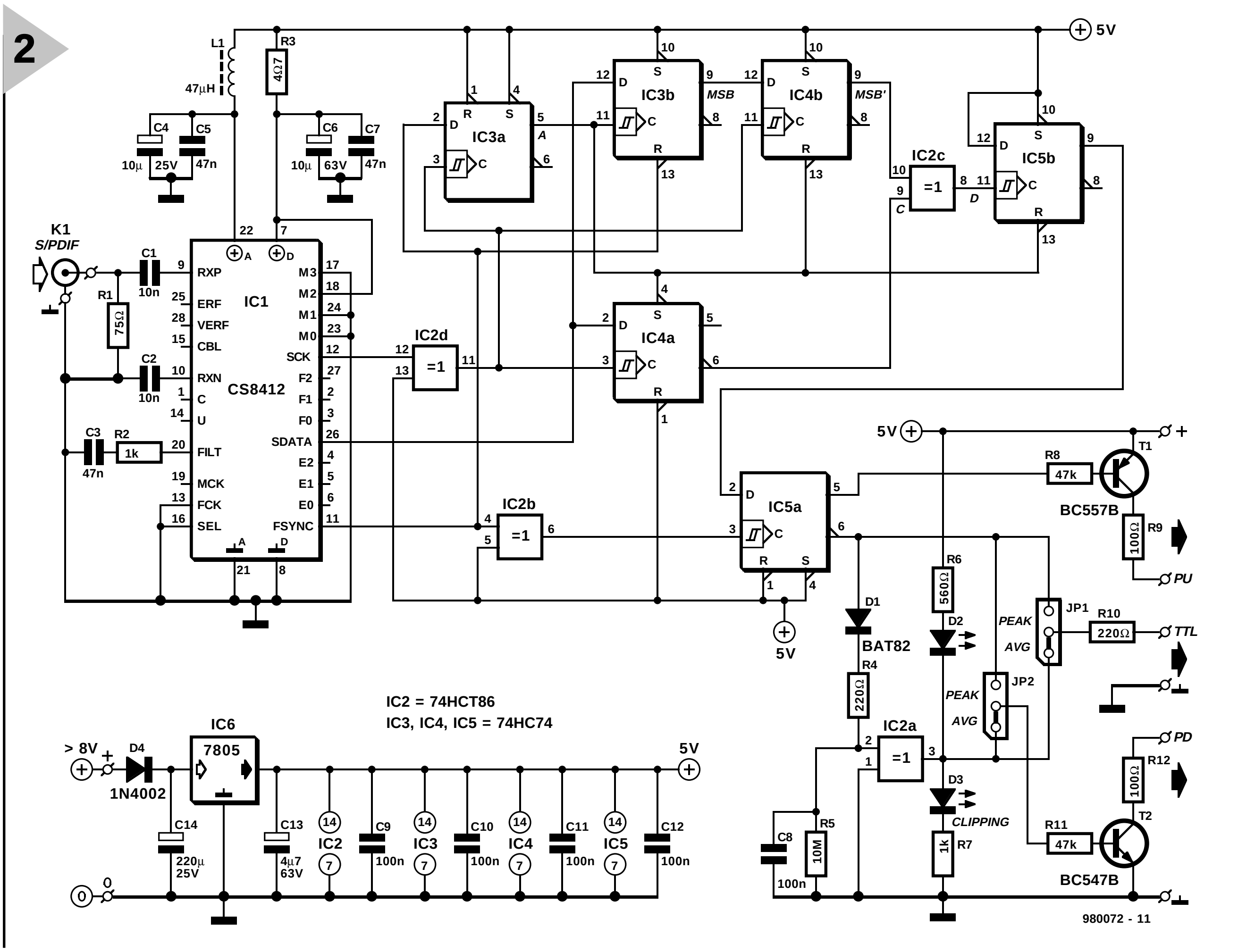


Figure 2. The circuit of the indicator consists of an integrated digital audio interface receiver and a number of gates and bistables forming the signalling section.

of the signal reach the peak value. The probability that some clipping then occurs is great. When the LED lights only once or twice per track, it must be assumed that this is caused by a couple of strong signals. If it lights more often, or it remains on for longer than a second, there is something not quite right.

To make a possible error indication as clear as possible, two LEDs are used: the green one lights as long as all is well, and the red one when there is something amiss. For those who have not the patience to keep an eye on the LEDs during the entire time the CD is played, there is an optional facility for connecting a counter module. This shows how many times during the playback clipping may have occurred.

In the design of the indicator it was assumed that the studio recording was transferred 1:1 to the manufactured compact disc.

CIRCUIT DESCRIPTION

The circuit diagram of the indicator is shown in **Figure 2**. Audio socket K_1 is for linking the indicator to the digital output of the CD player.

The relevant data are retrieved from the S/PDIF† signal with the aid of IC_1, an integrated interface receiver Type CS8412 (see Data Sheets elsewhere in this issue). This circuit can handle virtually all current sampling frequencies. The serial audio data (SDATA) are read with the aid of a bit-clock and a word-clock (SCK and FSYNC respectively). The output of the IC is set to a special format (normal mode FMT 4: $M_0=M_1=0$; $M_2=1$ $M_3=0$) in which a clock pulse follows each audio sample, irrespective of left or right.

The audio data are coded in 2s complement. To check whether a peak value has been reached, the MSB (most significant bit) must be inspected and compared with the remaining bits. In the case of digital minimum and maximum values, all remaining bits must be the opposite of the MSB. The minimum and maximum values are checked with an XOR function.

Gates & bistables

The comparing and indexing of the bits is carried out by a number of gates and D-bistables. The timing diagram of the most important signals is shown in **Figure 3**.

At the start of a new sample, a clock signal is generated exclusively for the MSB in D-bistable IC_{3a}-IC_{3b}. The FSYNC signal is clocked into IC_{3a} by the inverted bit-clock (IC_{2d}). The output of IC_{3a} (signal A) forms the clock for the MSB.

The clock input of IC_{3b} goes high in the middle of the MSB, after which the bit is held for the duration of the audio sample (pin 9 of IC_{3b}).

To ensure that the MSB is applied to comparator IC_{2c} simultaneously with the next bit, it is clocked again in D-bistable IC_{4b}.

The remaining bits are clocked by IC_{4a}. Since signal A is applied to the S-input of this bistable, the inverted output remains low until the first of the remaining bits is clocked (signal C).

Use of the inverted output ensures that all bits there have the same level as the MSB if and when a minimum or maximum value is reached. If the level is not the same, IC_2 goes high at a certain moment, which causes IC_{5b} to be clocked. This means that the output of IC_{5b} is high when there is neither a minimum nor a maximum value.

So as to enable the actual state of each sample to be determined, most bistables, including IC_{5b}, are reset by signal A. Before this happens, the status of IC_{5b} for each sample is clocked to IC_{5a} by the FSYNC signal inverted by IC_{2b}. Since IC_{5a} is not reset by signal A, the data transferred to it are retained. Pin 6 of this bistable therefore remains low as long as there is no sample that contains a minimum or maximum value. If a peak value does occur, pin 6 briefly goes high.

Converting the change in level at the output of IC_{5a} into a usable optical indication is not too difficult. The design basis was to make a (red) LED (D_3) light for about one second if two or more consecutive samples reach peak value. Time constant R_4-C_8 averages a number of samples, while the discharge time of C_8, determined by R_5, results in an afterglow of about one second. The potential across C_8 is buffered by IC_{2a}. The indication becomes much clearer by the addition of a second (green) LED (D_2). The light-up behaviour of this diode is the opposite of that of D_3, so that when clipping occurs there is a distinctive change of colour.

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Figure 3. Timing diagram of the most important signals in the indicator.

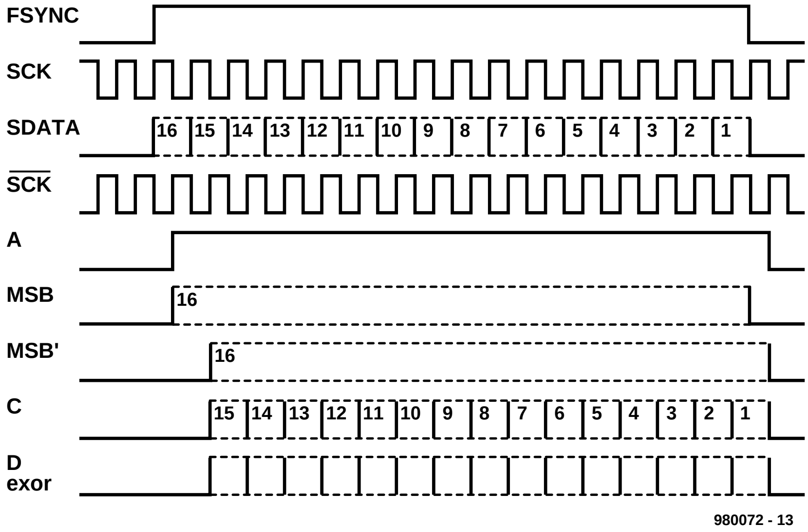

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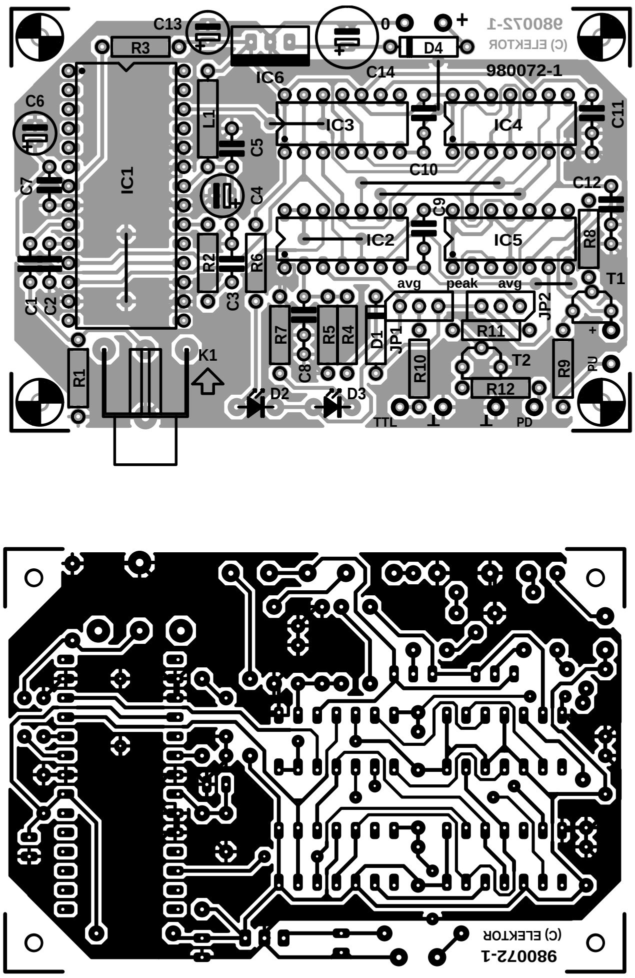
Figure 4. The printed-circuit board for the indicator makes construction child's play.



Parts list

Resistors:
R_1 = 75 Ω
R2, R_7 = 1 kΩ
R_3 = 4.7 Ω
R_4, R_{10} = 220 Ω
R_5 = 10 MΩ
R_6 = 560 Ω
R_8, R_{11} = 47 kΩ
R_9, R_{12} = 100 Ω

Capacitors:
C_1, C_2 = 0.01 µF, ceramic
C_3 = 0.047 µF
C_4, C_6 = 10 µF, 63 V, radial
C_5, C_7 = 0.047 µF, ceramic
C_8–C_{12} = 0.1 µF
C_{13} = 4.7 µF, 63 V, radial
C_{14} = 220 µF, 25 V, radial

Semiconductors:
D_1 = BAT82
D_2 = LED, green, high efficiency
D_3 = LED, red, high efficiency
D_4 = 1N4002
T_1 = BC557B
T_2 = BC547B

Integrated circuits:
IC_1 = CS8412 (Crystal Semiconductor)
IC_2 = 74HCT86
IC_3, IC_4, IC_5 = 74HC74
IC_6 = 7805

Miscellaneous:
L_1 = choke 47 µH
JP_1, JP_2 = 3-way, 2.54 mm pin strip and pin jumper (Maplin)
K_1 = audio connector (male) for board mounting
Enclosure 120 × 65 × 41 mm (L× W× H), e.g., Bopla 430 (available from Phoenix 01296 398355)
PCB Order no. 980072 (see Readers Services towards the end of this issue)

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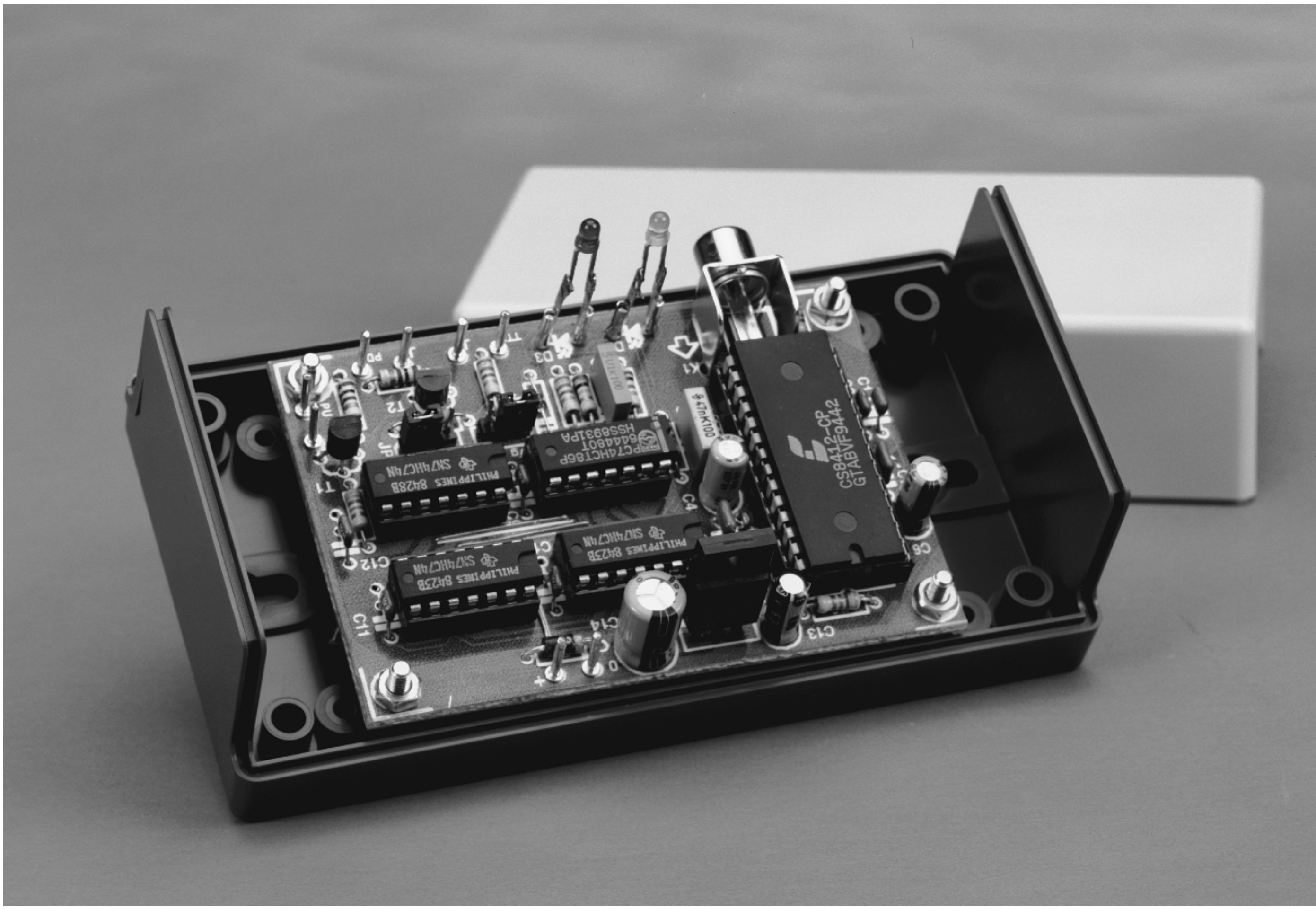


Figure 5. Photograph of the completed prototype indicator board.

COUNTER OPTION

As mentioned earlier, there is provision for linking a counter module to the indicator to show the number of times that clipping has occurred over a given period. There are three outputs: TTL, pull-down (PD), and pull-up (PU), so that almost any current type of module can be used.

Owing to the averaging by R_4-C_8, the output remains active even when brief interruptions occur. If, however, the output of IC_5 is used, count pulses are obtained for all discrete samples or strings of them. Both facilities may be used thanks to JP_1 and JP_2. This arrangement gives a choice at the TTL or PD output of either an averaged count of the number of times clipping has occurred or a count giving the peak value.

In practice, peak counting may be a little too severe, since normally nothing much happens when the peak signals just reach the 0 dB level. The averaged count is a more realistic measure of the number of clipping occurrences. A drawback of the averaged count is that the toggling of IC_{2a} may cause high-frequency pulses that may adversely affect fast counter modules. However, most modern modules are immune to these pulses and in any case the risk can be removed by connecting a 1 µF capacitor across the counter input.

CONSTRUCTION

The indicator is best built on the PCB shown in **Figure 4**. Populating the board with reference to the components list and the circuit diagram should not present any undue difficulties. Sockets should be used for the ICs. Mind the the polarity of the electrolytic capacitors and diodes. Note that D_1 must be of the type specified in view of the permissible leakage current.

Since the circuit provides for a 5 V regulator, IC_6, a mains adaptor with an output of not less than 8 V may be used as power source. The circuit draws a current of about 25 mA. This low current also facilitates the use of a 9 V battery if portable use is desired. A dry battery will give some 10 hours operation.

For portable use it is, of course, essential that the circuit is housed in a small, neat enclosure such as that specified.

[980072]

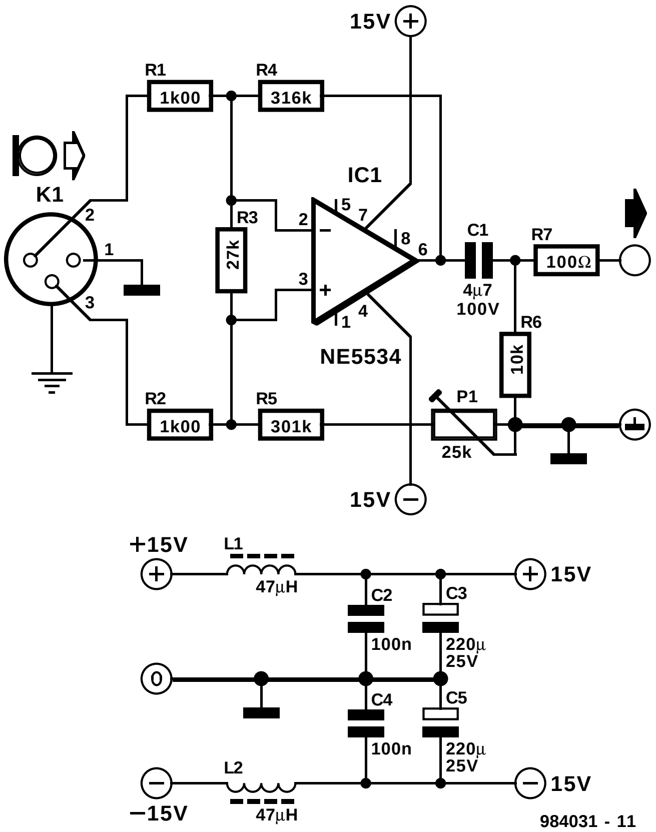
027 balanced microphone preamplifier

Design: T. Giesberts

The preamplifier is intended for use with dynamic (moving coil – MC) microphones with an impedance up to 200 Ω and balanced terminals. It is a fairly simple design, which may also be considered as a single stage instrument amplifier based on a Type NE5534 op amp.

To achieve maximum common-mode rejection (CMR) with a balanced signal, the division ratios of the dividers (R_1-R_4 and R_2-R_5 respectively) at the inputs of the op amp must be identical. Since this may be difficult to achieve in practice, a preset potentiometer, P_1, is connected in series with R_5. The preset enables the common-mode rejection to be set optimally.

Capacitor C_1 prevents any direct voltage at the input, while resistor R_7 ensures stability of the amplifier with capacitive loads. Resistor R_3 prevents the amplifier going into oscillation when the input is open circuit. If the microphone cable is of reasonable length, R_3 is not necessary, since the parasitic capacitance of the cable ensures stability of the amplifier. It should be noted, however, that R3 improves the CMR from >70 dB to >80 dB.



Performance of the preamplifier is very good. The THD + N (total harmonic distortion plus noise) is smaller than 0.1% with an input signal of 1 mV and a source impedance of 50 Ω.

Under the same conditions, the signal-to-noise ratio is –62.5 dBA.

With component values as specified, the gain of the amplifier is 50 dB (×316).

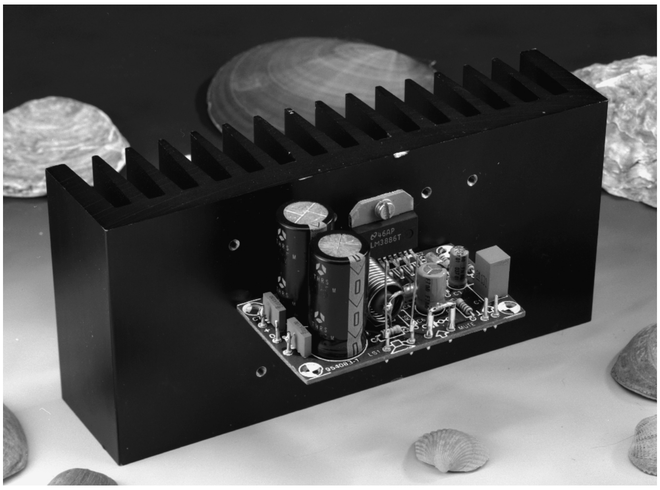
After careful adjustment of P_1 at 1 kHz, the CMR, without R_3, is 120 dB.

The supply voltage is ±15 V. The amplifier draws a current at that voltage of about 5.5 mA. Note the decoupling of the supply lines with L_1, L_2, C_2–C_5.

[984031]

003

100-watt single-IC amplifier



Specifications *(8 Ω/1 kHz unless otherwise noted)*

Input sensitivity:	*1 V_{rms} (63 W into 8 Ω)*
Output power, 8 Ω:	*63 W (THD <1%)*
Output power, 4 Ω:	*108 W (THD <1%)*
Damping factor (8 Ω)	*>450 at 1 kHz*
	>170 at 20 kHz
Slew rate:	*>10 V/µs (rise time = 5 µs)*
Power bandwidth:	*8 Hz to 90 kHz*
Signal/noise ratio:	*94 dBA (1 W into 8 Ω)*

Design: T. Giesberts

According to National Semiconductor, the LM3886 is a high-performance 150W Audio Power Amplifier with Mute. The performance of the LM3886, say NS, utilising its Self Peak Instantaneous Temperature (°Ke) (SPIKe) protection circuitry, puts in a class above discrete and hybrid amplifiers by providing an inherently, dynamically protected Safe Operating Area (SOA). The LM3886T comes in an 11 (staggered-) lead non-isolated TO220 package.
We put the LM3886T through its paces, using two earlier publications (Ref. 1, 2) and an existing printed circuit board as a basis. For test purposes, the prototype of the amplifier was powered by a stabilised ±35-V supply. A maximum undistorted output power of about 63 watts into 8 ohms was obtained at a drive level of 1 Vrms. Dropping the load impedance to 4 ohms pushed the output power to no less then 108 watts. In practice, these power levels can be taken to mean 'music power', but do remember that the amplifier will not normally be powered from a regulated supply!
Great attention should be paid to the cooling of the amplifier IC. The cooling capacity offered by a heatsink as specified in the parts list is really only sufficient for load impedances of 6 ohms or more. Even if a heatsink with a thermal resistance lower than 1 K/W is employed, the amplifier IC will cause a 'hot spot' on the heatsink surface where the actual thermal resistance is much higher locally than the specification! With this in mind, it is recommended to drop the supply voltage to about ±30 V if the amplifier is used to drive a 4-ohm load. Also, bear in mind that heatsink isolating materials like mica and even ceramics tend to raise the thermal resistance by 0.2 K/W to 0.4 K/W. The metal tab at the back of the IC is at the negative supply potential.
Boucherot network C6-R6 is not normally required in this application, and should be omitted unless the amplifier is found to be unstable as a result of an application which is widely different from the one shown here.
Populating the amplifier board itself will be a piece of cake, and most of the time required to build the amplifier will go into drilling, cutting, mounting and isolating the heatsink. The printed circuit board shown here is available ready-made through the Publishers' Readers Services. Note that the radial electrolytic capacitors are rated at 40 volts, so you have to make sure that the supply voltage can never exceed that level. The performance of the prototype amplifier built and tested in our design lab is expressed by the Specifications box.

(984062-1)

COMPONENTS LIST

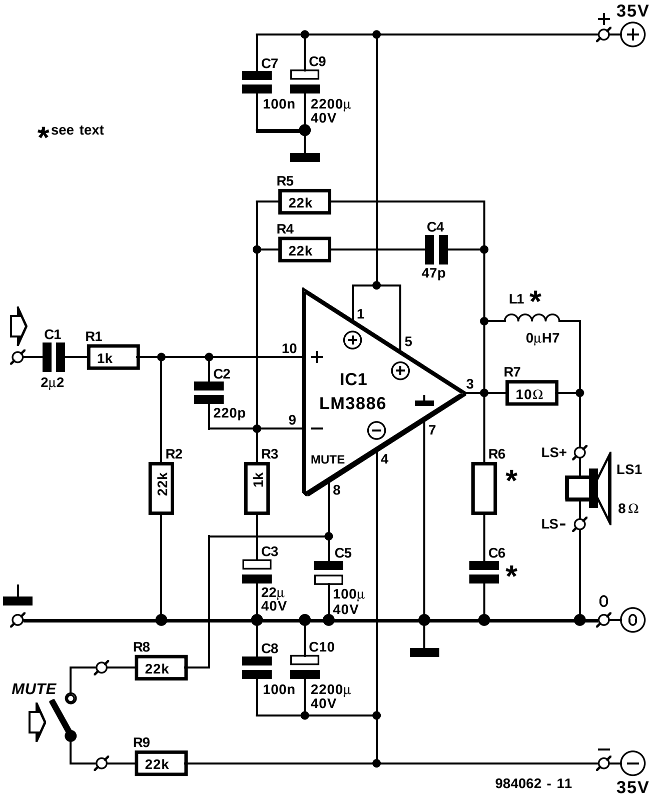
Resistors:
R1,R3 = 1kΩ
R2,R4,R5,R8,R9 = 22kΩ
R6 = not fitted, see text
R7 = 10Ω, 5W

Capacitors:
C1 = 2µF2, MKT (Siemens), pitch 5mm or 7.5mm
C2 = 220pF, 160V, axial, polystyrene (Siemens)
C3 = 22µF, 40V, radial
C4 = 47pF, 160V, axial, polystyrene (Siemens)
C5 = 100µF, 40V, radial
C6 = not fitted, see text
C7,C8 = 100nF
C9,C10 = 2200µF, 40V, radial, max. diameter 16mm

Inductor:
L1 = 0.7µH, 13 turns of 1.2-mm diameter (#18 SWG) enamelled copper wire, 10mm internal diameter, wound around R7.

Semiconductor:
IC1 = LM3886T (National Semiconductor)

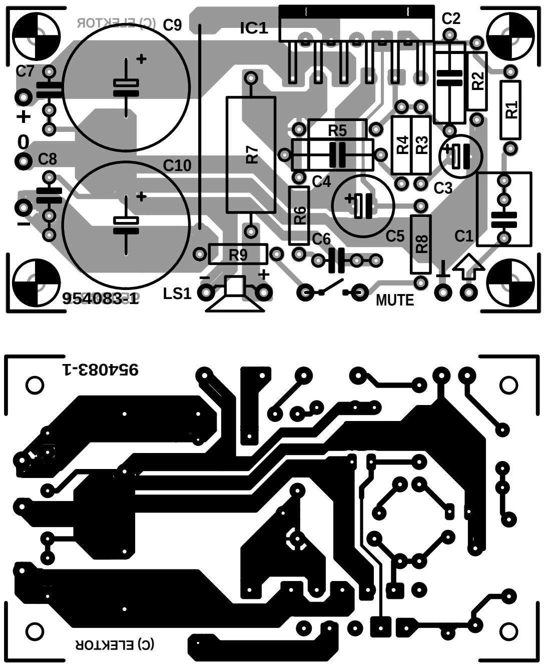
Miscellaneous:
Heatsink for IC1: specification R_{th} < 1K/W
Printed circuit board, order code 954083-1.



References:

1. LM3886 150W Audio Power Amplifier with Mute (Application Note), *Elektor Electronics* May 1995.

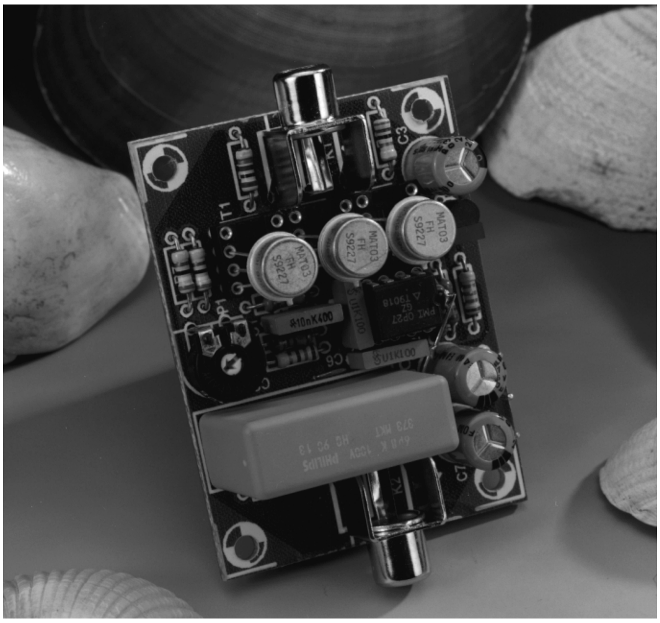
2. Single-chip 50 W AF Amplifier, *Elektor Electronics* December 1995.



Contents

009

ultra low-noise MC preamplifier



COMPONENTS LIST

Resistors:
R1,R12 = 100Ω
R2 = 15kΩ
R3 = 82Ω
R4,R5 = 1kΩ50
R6 = 150Ω
R7,R8 = 39Ω
R9 = 5Ω62
R10 = 82Ω5
R11 = 511Ω
R13 = 100kΩ
P1 = 50Ω preset H

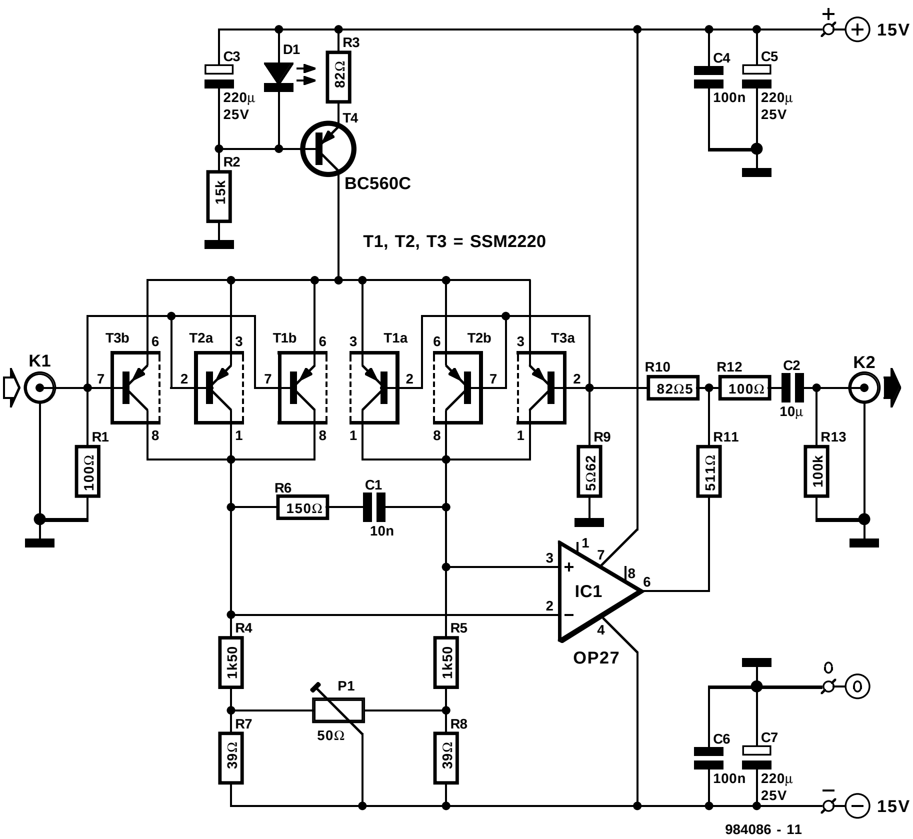
Capacitors:
C1 = 10nF
C2 = 10μF MKT (Siemens) raster 22.5mm or 27.5mm
C3,C5,C7 = 220μF 25V radial
C4,C6 = 100nF

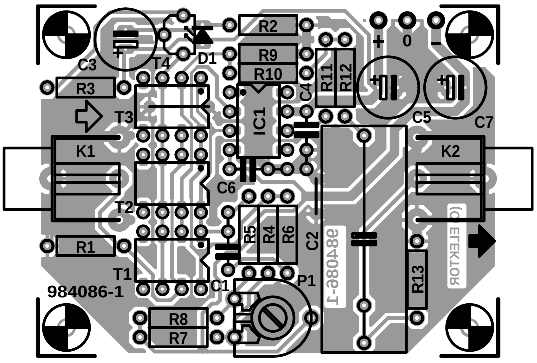
Semiconductors:
D1 = red LED, flat
T1,T2,T3 = SSM2220 or MAT03 (Analog Devices)
T4 = BC560C
IC1 = OP27GP (Analog Devices)

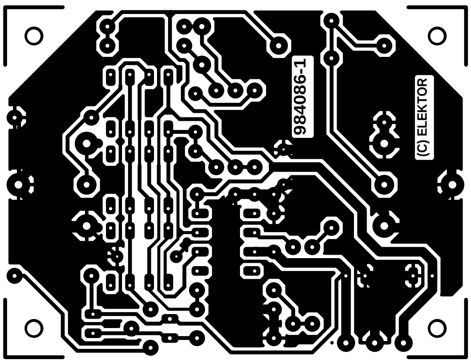
Miscellaneous:
K1,K2 = phono (line) socket, PCB mount, gold-plated, e.g. T-709G from Monacor/Monarch (available from C-I Electronics or Stippler Electronics)

This preamplifier was designed for low-impedance signal sources like MC (moving-coil) pick-up cartridges used in high-end record players (yes, they still exist). The actual input impedance of the preamplifier is 100 Ω. To keep the input noise as low as possible, three dual transistors type SSM2220 or MAT03 transistors are connected in parallel to form a discrete difference amplifier. By connecting this amplifier ahead of an opamp (OP27), the input noise of the opamp becomes immaterial. The base connections of the discrete amplifier then function as the inputs of a super-opamp with a very low input noise level. An advantage of the p-n-p transistors used here over their n-p-n counterparts is their much lower low-frequency noise level. On the down side, a fairly large bias current of about 5.5 μA is created at the input. This is the result of the 2-mA setting for each transistor in combination with the relatively low gain of the p-n-p devices.

Preset P1 and resistors R7/R8 enable you to iron out any tolerances on R4 and R5 in the difference amplifier output. Transistor T4 and LED D1 ensure a stable current setting for the difference amplifier. D1 should be a flat, red, LED which is fitted face-to-face against T4 for thermal coupling. Because the input noise level amounts to 0.4 nV/√Hz (theoretical value for a 10-Ω resistor), it is essential that the feedback adds as little as possible to the overall noise figure. Consequently, the impedance of the feedback circuit must be much lower than 10 Ω. Furthermore, the OP27 demands a certain minimum load impedance, so that the feedback impedance may not be less than 600 Ω. To ensure that a low value can be used for R9, a compromise had to be found between maximum gain (here, approx. 24 dB or 15.7 times) on the one hand, and the value of R9. By fitting an additional resistor, R11, ahead of the actual feedback, the opamp is not excessively loaded, while R9 adds 'only' 0.3 nV/√Hz to the input noise level, which, based on measurement data, amounts to 0.52 nV/√Hz. If more gain is needed, a noise figure of about 0.4 nV/√Hz may be achieved at a lower value of R9. The obvious disadvantage of adding R11 is a higher internal gain, causing a smaller bandwidth and a lower drive margin. Fortunately, these factors are of little consequence in the case of moving-coil elements.
There are two ways to adjust P1. The first is to adjust the output







voltage to nil (measure at IC1 pin 6). The second option is to measure the input offset, for example, 0.55 mV across 100 Ω. Assuming that the offset caused by T1, T2 and T3 is negligible, then the output voltage should be 15.68 x 0.55 mV for perfect symmetry, in other words, junction R10-R11-R12 should be at 8.62 mV with respect to ground.

Those of you who like to experiment may want to try the effects of reducing the number of input transistors from three to just one. You may want to do this, for example, to reduce the input bias current. Resistor R3 then has to be changed into 249 Ω. Do remember, however, that the input noise level then rises by 2.5 dB!

The output has a large, solid 10 μF MKT (metal theraphtelate, ask your local Siemens distributor) capacitor to prevent a large offset voltage being applied to the input of an MD amplifier.

The preamplifier is powered by a symmetrical, regulated 15-V supply, and draws about 16 mA on each rail. Finally, here are a few key figures measured on our prototypes:

Configuration: 3 x SSM2220/MAT03		
	signal: 0.5 mV/25 Ω	input short-circuited
S/N (BW = 22 kHz)	71.2 dB	74 dB
	74 dBA	76.2 dBA
Configuration: 1 x MAT03 (R3 = 249 Ω)		
S/N (BW = 22 kHz)	69.5 dB	71 dB
	72.3 dBA	73.7 dBA

The preamplifier is best built on the printed circuit board whose artwork is shown here. Construction is uncritical, but do not forget the wire links under transistor T3 and next to capacitor C2. The PCB is unfortunately not available ready-made from the Publishers.

(984086-1, Gb)

020 1-watt BTL audio amplifier

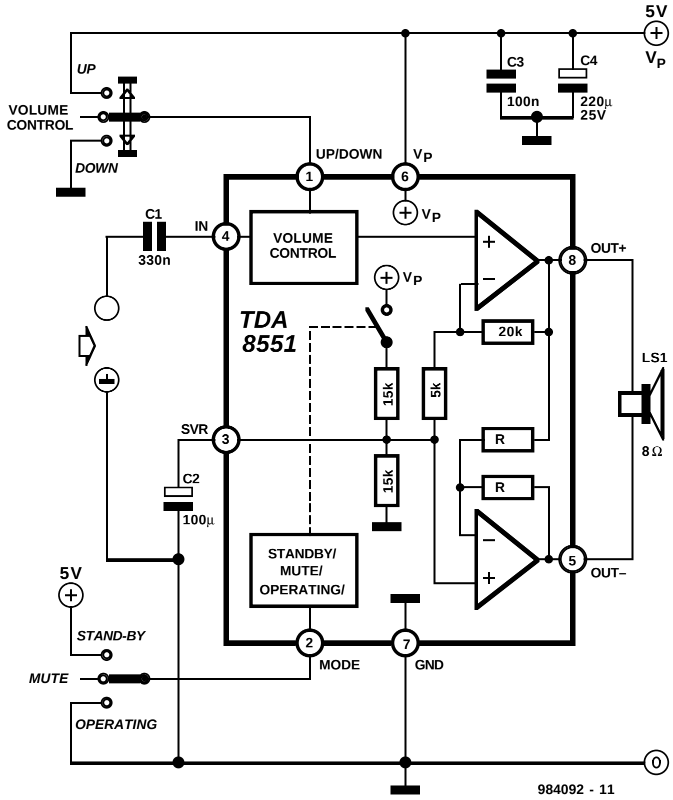
Source: Philips Semiconductors Preliminary Specification

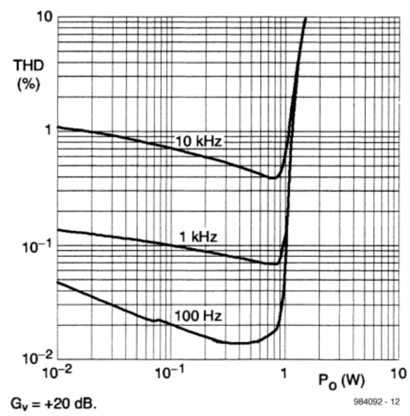
The TDA8581(T) from Philips Semiconductors is a 1-watt Bridge Tied Load (BTL) audio power amplifier capable of delivering 1 watt output power into an 8-Ω load at THD (total harmonic distortion) of 10% and using a 5-V power supply. The schematic shown here combines the functional diagram of the TDA8551 with its typical application circuit. The gain of the amplifier can be set by the digital volume control input. At the highest volume setting, the gain is 20 dB. Using the MODE pin the device can be switched to one of three modes: standby (MODE level between V_P and V_P–0.5 V), muted (MODE level between 1 V and V_P–1.4 V) or normal (MODE level less than 0.5 V). The TDA8551 is protected by an internal thermal shutdown protection mechanism.

The total voltage loss for both MOS transistors in the complementary output stage is less than 1 V. Using a 5-V supply and an 8-Ω loudspeaker, an output power of 1 watt can be delivered.

The volume control has an attenuation range of between 0 dB and 80 dB in 64 steps set by the 3-state level at the UP/DOWN pin: *floating*: volume remains unchanged; *negative pulses*: decrease volume; *positive pulses*: increase volume Each pulse at he Up/DOWN pin causes a change in gain of 80/64 = 1.25 dB (typical value). When the supply voltage is first connected, the attenuator is set to 40 dB (low volume), so the gain of the total amplifier is then –20 dB. Some positive pulses have to be applied to the UP/DOWN pin to achieve listening volume. The graph shows the THD as a function of output power. The maximum quiescent current consumption of the amplifier is specified at 10 mA, to which should be added the current resulting from the output offset voltage divided by the load impedance.

(984092-1, Gb)





Contents

004 playback amplifier for cassette deck

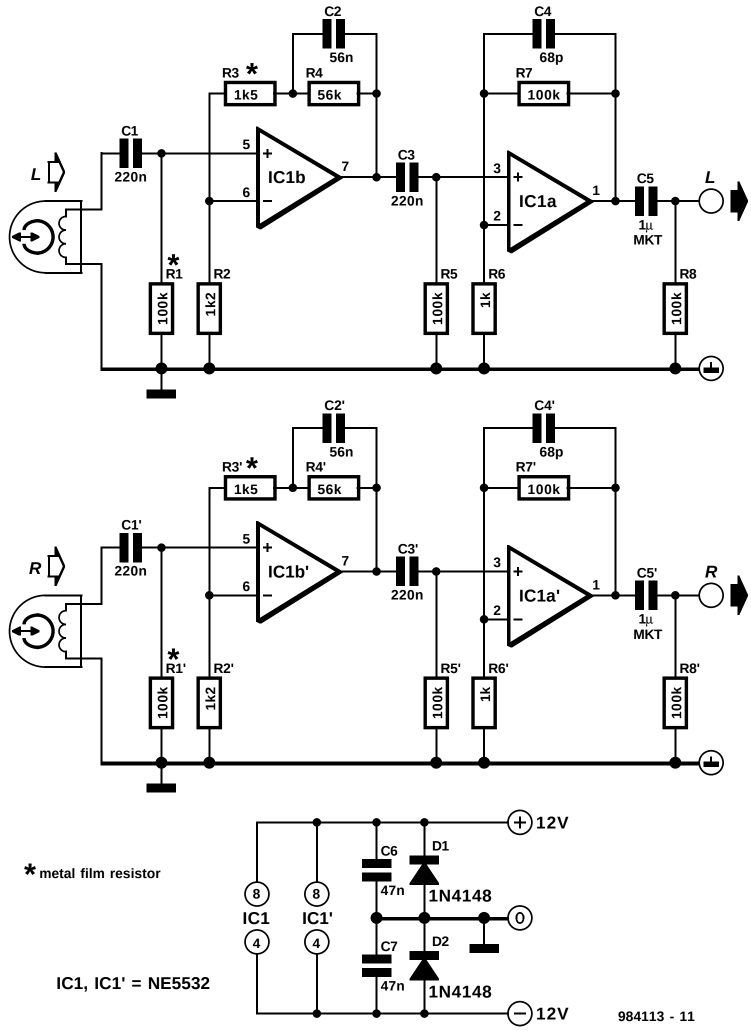
Design: T. Giesberts

For some time now, there have been a number of tape cassette decks available at low prices from mail order businesses and electronics retailers. Such decks do not contain any electronics, of course. It is not easy to build a recording amplifier and the fairly complex magnetic biasing circuits, but a playback amplifier is not too difficult as the present one shows.

The stereo circuits in the diagram, in conjunction with a suitable deck, form a good-quality cassette player. The distortion and frequency range (up to 23 kHz) are up to good standards. Moreover, the circuit can be built on a small board for incorporation with the deck in a suitable enclosure.

Both terminals of coupling capacitor C_1 are at ground potential when the amplifier is switched on. Because of the symmetrical ±12 V supply lines, the capacitor will not be charged. If a single supply is used, the initial surge when the capacitor is being charged causes a loud click in the loudspeaker and, worse, magnetizes the tape.

The playback head provides an audio signal at a level of 200–500 mV. The two amplifiers raise this to line level, not linearly, but in accordance with the RIAA equalization characteristic for tape recorders. Broadly speaking, this characteristic divides the frequency range into three bands:



- Up to 50 Hz, corresponding to a time constant of 3.18 ms, the signal is highly and linearly amplified.
- Between 50 Hz and 1.326 kHz, corresponding to a time constant of 120 µs, for normal tape, or 2.274 kHz, corresponding to a time constant of 70 µs, for chromium dioxide tape, the signal is amplified at a steadily decreasing rate.
- Above 1.326 kHz or 2.274 kHz, as the case may be, the signal is slightly and linearly amplified.

This characteristic is determined entirely by A_1 (A_1'). To make the amplifier suitable for use with chromium dioxide tape, add a double-pole switch (for stereo) to connect a 2.2 kΩ resistor in parallel with R_3 (R_3').

The output of A_1 (A_1') is applied to a passive high-pass rumble filter, C_3-R_5 (C_3'-R_5') with a very low cut-off frequency of 7 Hz. The components of this filter have exactly the same value as the input filter, C_1-R_1 (C_1'-R_1').

The second stage, A_2 (A_2') amplifies the signal ×100, that is, to line level (1 V r.m.s.).

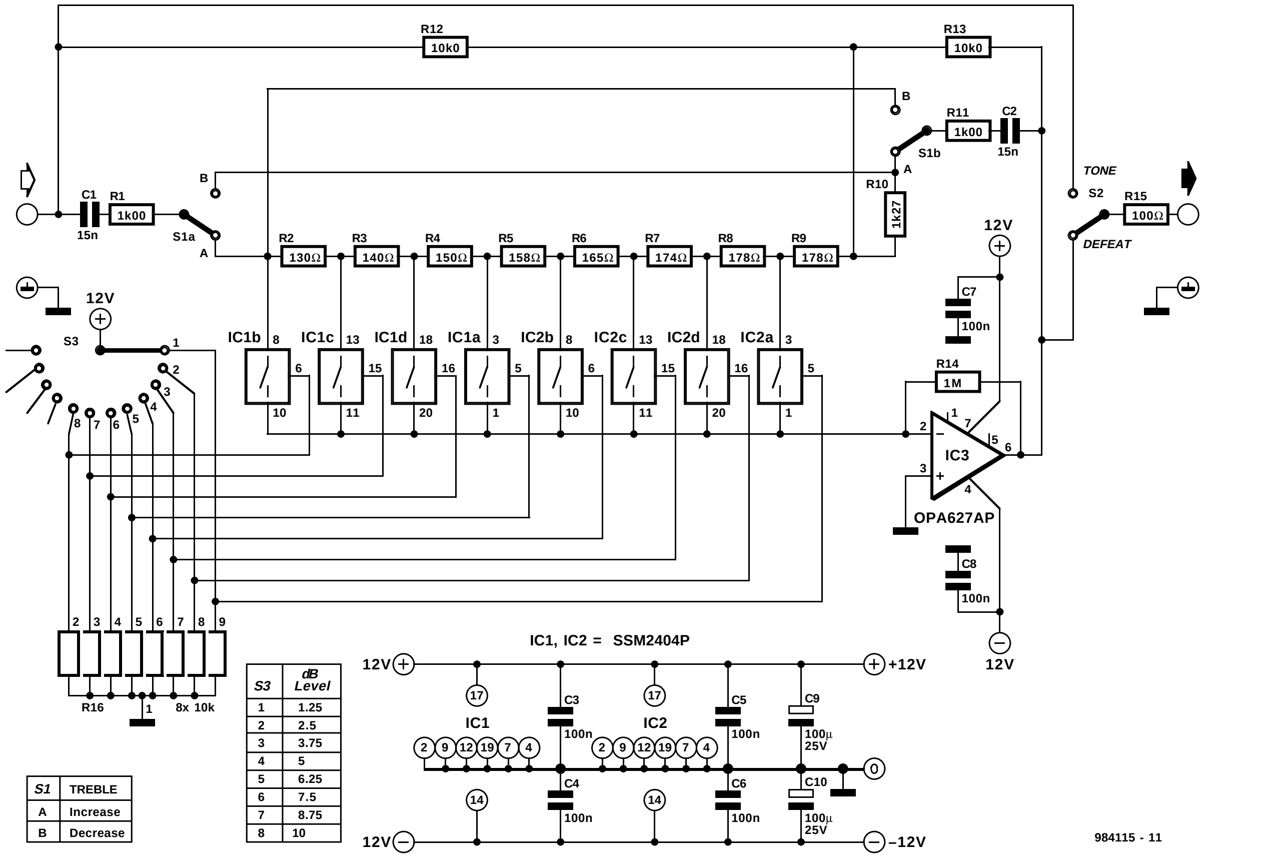
Capacitor C_4 limits the upper frequency range to avoid r.f. interference and any tendency of the amplifier to oscillate.

The amplifier needs a symmetrical ±12 V power supply that can provide a current of up to 0.5 A. The greater part of this current is drawn by the motor of the deck; the electronic circuits draw only 15 mA.

[984113]

001

treble tone control



S3	dB Level
1	1.25
2	2.5
3	3.75
4	5
5	6.25
6	7.5
7	8.75
8	10

S1	TREBLE
A	Increase
B	Decrease

Design: T. Giesberts

The treble control works in a similar manner as the bass control elsewhere in this issue, but contains several modifications, of course. One of these is the series network C_1-C_2–R_1–R_{11}.

The d.c. operating point of IC_3 is set with resistors R_{12} and R_{13}. To ensure that these resistors do not (adversely) affect the control characteristics, they are coupled to the junction of R_9 and R_{10}. In this way they only affect the low-frequency noise and the load of the op amp. Their value of 10 kΩ is a reasonable compromise.

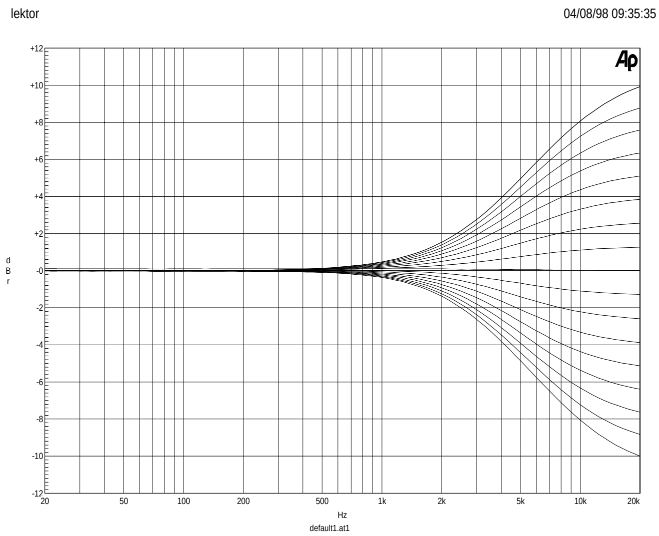
The functions of switches S_1–S_3 are identical to those of their counterparts in the bass tone control; their influence is seen clearly in the characteristics.

Good symmetry between the left-hand and right-hand channels is obtained by the use of 1% versions of R_1–R_{13} and C_1, C_2.

The value of resistors R_2–R_{10} is purposely different from that of their counterparts in the bass tone control. In the present circuit, the control range starts above 20 kHz. To make sure that a control range of 10 dB is available at 20 kHz, the nominal amplification is ×3.5 (11 dB).

The control circuit draws a current of about ±10 mA.

[984115]



039

up/down drive for tone control

Design: T. Giesberts

The up/down drive is intended primarily for use with the tone controls described elsewhere in this issue.

The tone controls use electronic switches that are operated by a multi-position selector. The present circuit is intended as a replacement for this selector and has facilities for operating the tone controls via an up and a down key. A third key enables the user to switch over rapidly to a preprogrammed position of the relevant tone control.

The electronic switches are driven by a BCD-to-decimal decoder Type 4028 (IC_3), which in turn is controlled by a 4-bit preset up/down counter (IC_2). The counter uses the three lowest bits only. The MSB of decoder IC_3 is permanently low. Only the eight lowest outputs of the decoder are used and these are linked via K_1 to the control inputs of IC_1 and IC_2 in the tone controls.

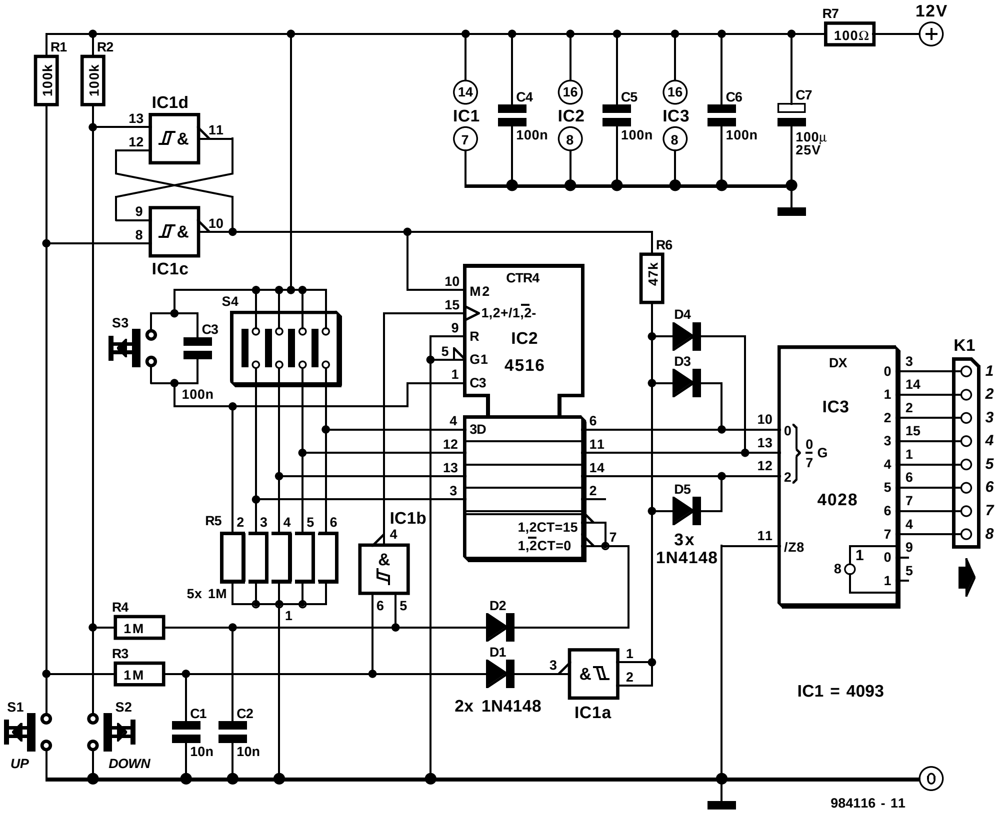
The circuit is operated with S_1 and S_2. Switch S_3 is the earlier mentioned preset key. The data for the preset inputs are set with DIP switch S_4. Capacitor C_3 ensures that when the supply voltage is switched on, the preset data are automatically adopted by the counter.

Each of switches S_1 and S_2 drives an S/R bistable (US: flip-flop), which determines the level at the U/D input of counter IC_2.

Networks R_3-C_1 and R_4-C_2, in conjunction with Schmitt trigger IC_{1b} provide a thorough debouncing and at the same delay the clock pulse slightly. This delay guarantees that the clock pulse (output of IC_{1b}) arrives after the state of the counter has been defined.

To prevent the counter jumping from minimum to maximum or vice versa, the clock pulse is disabled in the outermost positions. In the minimum state, this is achieved simply by use of the carry-out terminal (pin 7) of the counter. In the maximum state, an auxiliary network, consisting of R_6, D_3, D_4, D_5/IC_{1a}, and D_1, was found necessary.

Diode D_2 ensures that pin 5 of IC_1 remains low when the minim state is reached until S_1 is pressed. The same is achieved by diode D_1 in regards of pin 6 of the IC when the maximum state is reached. Resistor R_6 serves to reset the clock disabling during down counting; when the down key is pressed, the output of IC_{1a} goes high again.



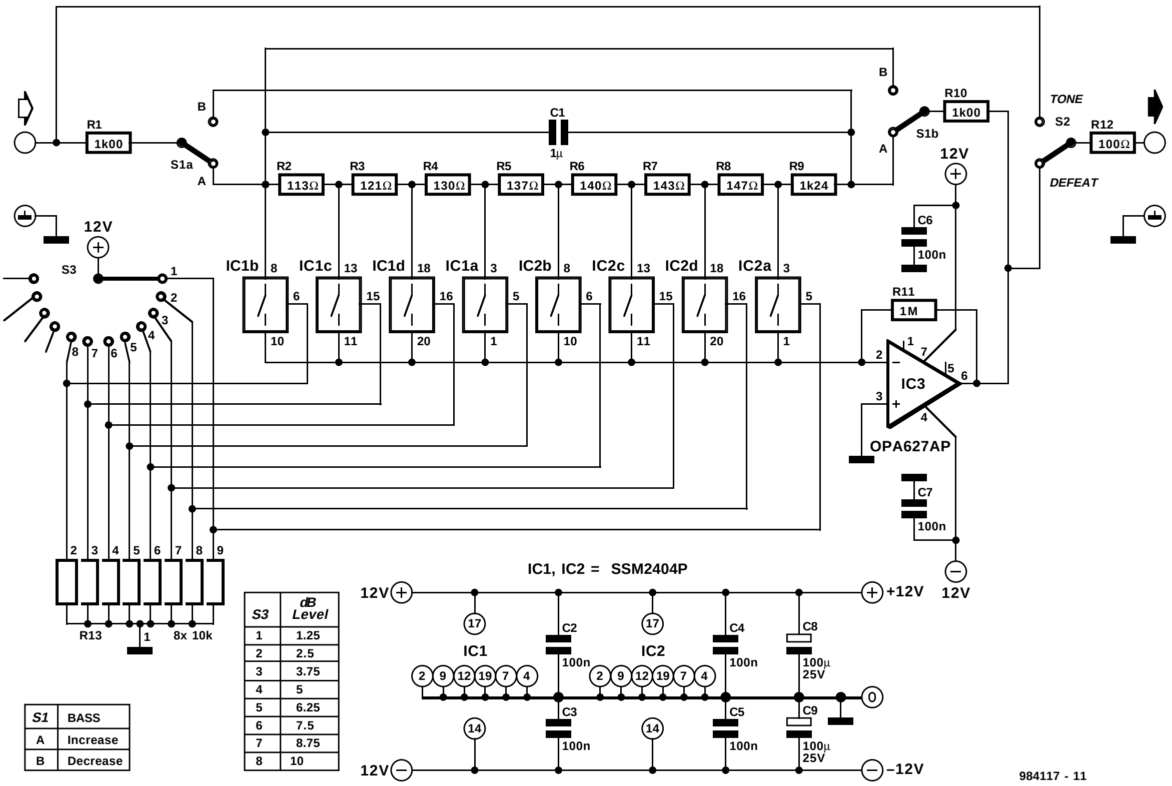
If an indication is desired of the actual state of the up/down drive, eight high-efficiency LEDs may be added at the output of IC_3 (anodes to the output, cathodes via a common 10 kΩ resistor to ground).

An indication whether amplification or attenuation occurs may be given by an additional LED at the output of IC_{1c} or IC_{1d}.

During quiescent operation, the circuit draws a current of 20 µA, which rises to about 140 µA when S_1 or S_2 is pressed. Network R_7-C_7 provides effective decoupling of the digital circuit from the analogue supply. [984116]

008

accurate bass tone control



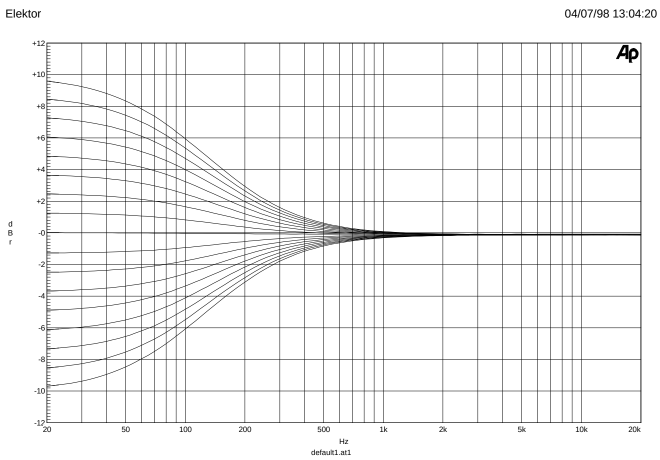
S3	dB Level
1	1.25
2	2.5
3	3.75
4	5
5	6.25
6	7.5
7	8.75
8	10

S1	BASS
A	Increase
B	Decrease

Design: T. Giesberts

A difficult problem in the design of conventional stereo tone controls is obtaining synchronous travel of the potentiometers. Even a slight error in synchrony can cause phase and amplitude differences between the two channels. Moreover, linear potentiometers are often used in such controls, and these give rise to unequal performance by human hearing. Special potentiometers that counter these difficulties are normally hard to obtain in retail shops.

A good alternative is a control based on a rotary switch and a discrete potential divider. The problem with this that for good tone control more than six steps are needed, and switches for this are also not readily available. Fortunately, electronic circuits can remove these difficulties.



The analogue selectors used may be driven by mechanical switches, standard logic circuit or a microcontroller. The selectors used in the present circuit are Type SSM2404 versions from Analogue Devices, which switch noiselessly. Each IC contains four selectors, so that a total of eight are used. The step size is 1.25 dB at 20 Hz with a maximum of 10 dB.

The circuit can be mirrored with S_1, which means that a selection may be made of amplification or attenuation of bass frequencies. The user can choose between attenuation only and extending the range by dividing R_9. The control can be bridged by switch S_2.

To prevent the output impedance of the circuit having too much effect on the operation of the circuit, the output impedance must be ≤ 10 Ω. Resistor R_{12} protects the circuit against too small a load.

At maximum bass amplification at U_{in} = 1 V r.m.s., the THD + N < 0.001% for a frequency range of 20 Hz to 20 kHz and and a bandwidth of 80 kHz.

The circuit draws a current of about 10 mA.

[984117]

024 ten-band equalizer

Design: P. Staugaard

The equalizer presented in this article is suitable for use with hi-fi installations, public-address systems. mixers and electronic musical instruments.

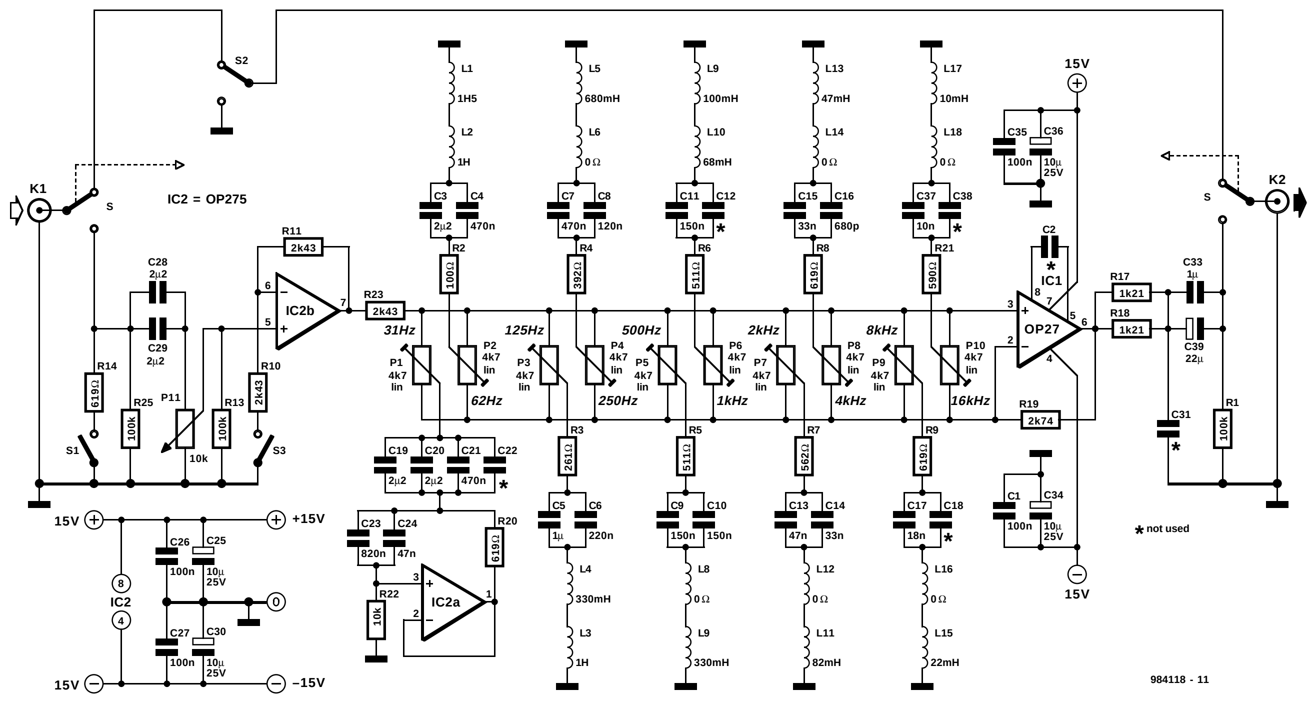
The relay contacts at the inputs and outputs, in conjunction with S_2, enable the desired channel to be selected. The input may be linked directly to the output, if wanted. The input impedance and amplification of the equalizer are set with S_1 and S_3. The audio frequency spectrum of 31 Hz to 16 kHz is divided into ten bands.

Ten bands require ten filters, of which nine are passive and one active. The passive filters are identical in design and differ only in the value of the relevant inductors and capacitors. The requisite characteristics of the filters are achieved by series and parallel networks. The filter for the lowest frequency band is an active one to avoid a very large value of inductance. It is based in a traditional manner on op amp A_1.

The inductors used in the passive filters are readily available small chokes. The filter based on L_1 and L_2 operates at about the lowest frequency (62 Hz) that can be achieved with standard, passive components.

The *Q*(uality) factor of the filters can, in principle, be raised slightly by increasing the value of R_{19} and R_{23}, as well as that of P_1–P_{10}, but that would be at the expense of the noise level of op amp IC_1.

With component values as specified, the control range is about ±11 dB, which in most case will be fine. A much larger range is not attainable without major redesign.



The input level can be adjusted with P_1, which may be necessary for adjusting the balance between the channels or when a loudness control is used in the output amplifiers.

Several types of op amp can be used:in the prototype, IC_1 is an LT1007, and IC_2, an OP275. Other suitable types for IC_1 are OP27 or NE5534; and for IC_2, AD712, LM833 and NE5532. If an NE5534 is used for IC_1, C_2 is needed; in all other cases, not.

The circuit needs to be powered by a regulated, symmetrical 15 V supply. It draws a current of not more than about 10 mA.

[984118]

Titan 2000

High-power hi-fi and public-address amplifier

It could be argued that most of the output amplifiers published in this magazine lack power. Although this is a debatable point, it was felt that a true heavyweight output amplifier would make a welcome change for many constructors. The Titan 2000 can produce 300 watts into 8 Ω, 500 watts into 4 Ω, and 800 watts into 2 Ω. For those who believe that music power is a reputable quantity, the amplifier can deliver 2000 watts of this magical power into 4 Ω.

Brief parameters

Sine-wave power output	*300 W into 8 Ω; 500 W into 4 Ω; 800 W into 2 Ω*
*Music power**	*2000 W into 4 Ω*
Harmonic distortion	*<0.005%*
Slew limiting	*85 V μs^{-1}*
Open-loop bandwidth	*55 kHz*
Power bandwidth	*1.5 Hz – 220 kHz*

**See text about the validity of this meaningless quantity.*

Design by T. Giesberts

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INTRODUCTION

Amplifier output has been a cause of argument for as long as there have been audio power amplifiers. For domestic use, a power rating of 2×50 W is more than sufficient. With the volume control at maximum and the use of correctly matched good-quality loudspeakers, this will provide a sound pressure level (SPL) equivalent to that of a grand piano being played forte in the same room.

However, not all amplifiers are intended for domestic use: many are destined for discos, small music halls and other large rooms. But even here, what power is really required? Since doubling the amplifier output increases the SPL by a barely audible 3 dB, it was felt that 300 watts sine wave power into 8 Ω would appeal to many.

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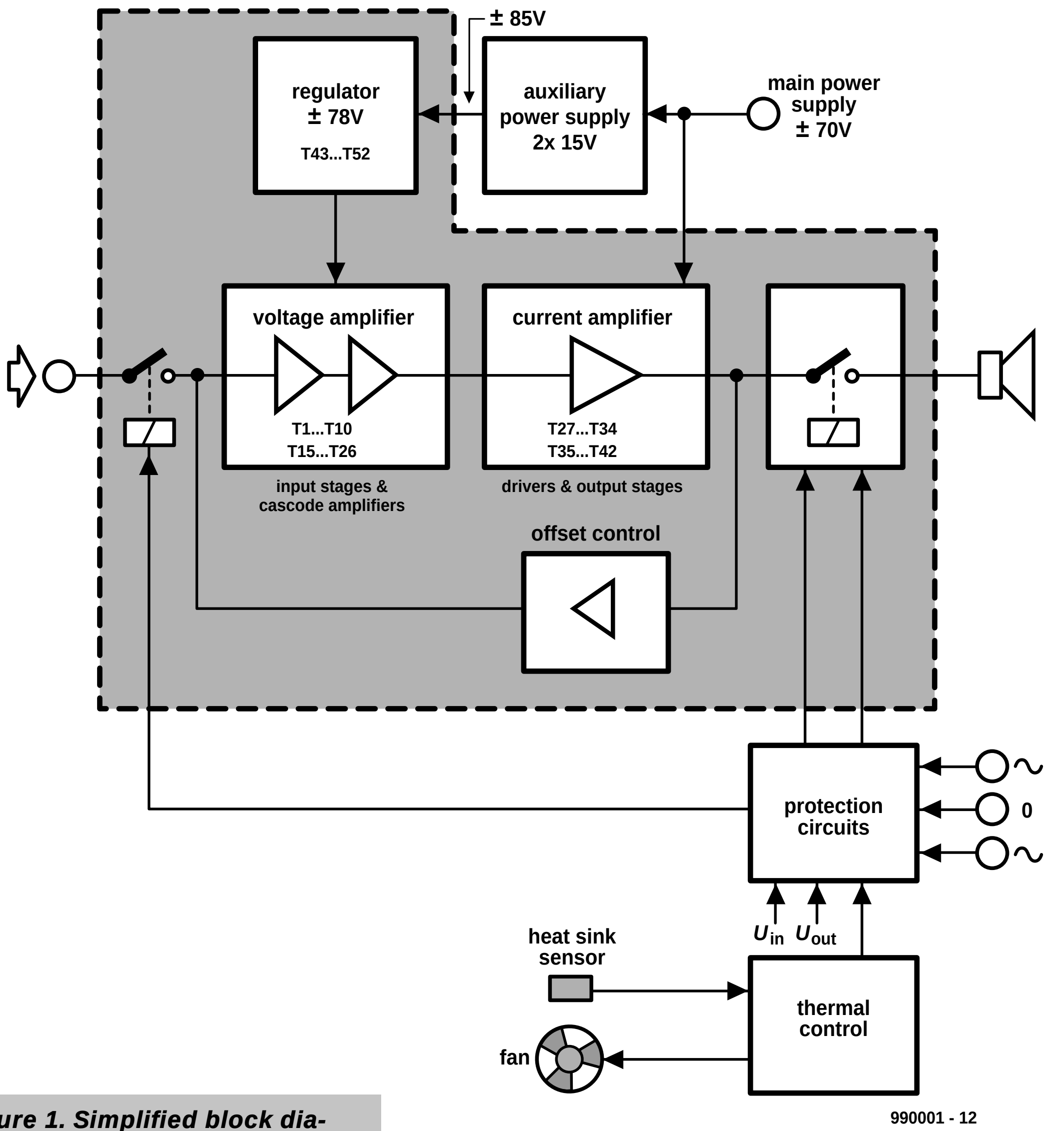

Figure 1. Simplified block diagram of the Titan 2000. The auxiliary power supply, protection networks and thermal control are discrete circuits built on discrete PCBs.

'PROGRAMMABLE' POWER OUTPUT

The amplifier has been designed in such a manner that its output is 'programmable' as it were. With a sine wave input, it delivers an average power of 300 W into an 8 Ω load, which should meet the requirements of all but the power drunk. Compared with the output of 50 W from a domestic audio amplifier, this gives an increase in SPL of 7.5 dB. If even higher outputs are needed, the load impedance may be lowered to 4 Ω, which will give an increase in SPL of 10 dB compared with a 50 W output.

Although music power is a deprecatory term, since it does not really give the true power rating of an amplifier, readers may note that the Titan 2000 can deliver 2 kW of this magical power into 4 Ω. (*True power is average power, that is, the product of the r.m.s. voltage across the loudspeaker and the r.m.s. current flowing into the speaker. The term music power is generally meaningless, because to some manufacturers it means the product of the peak voltage and peak current; to others it means merely double the true power; and to yet others, even more disreputable, it means quadrupling the true power*).

However, power is not the only criterion of an amplifier. Low distortion, good slew limiting, and an extended power bandwidth, as possessed by the Titan 2000, are also hallmarks of a good amplifier.

Power bandwidth denotes the frequency range over which the power falls to not less than half its maximum value. This is much more telling than the frequency response, which is usually measured at a much lower output level.

Slew limiting is the maximum input voltage change that can occur in one

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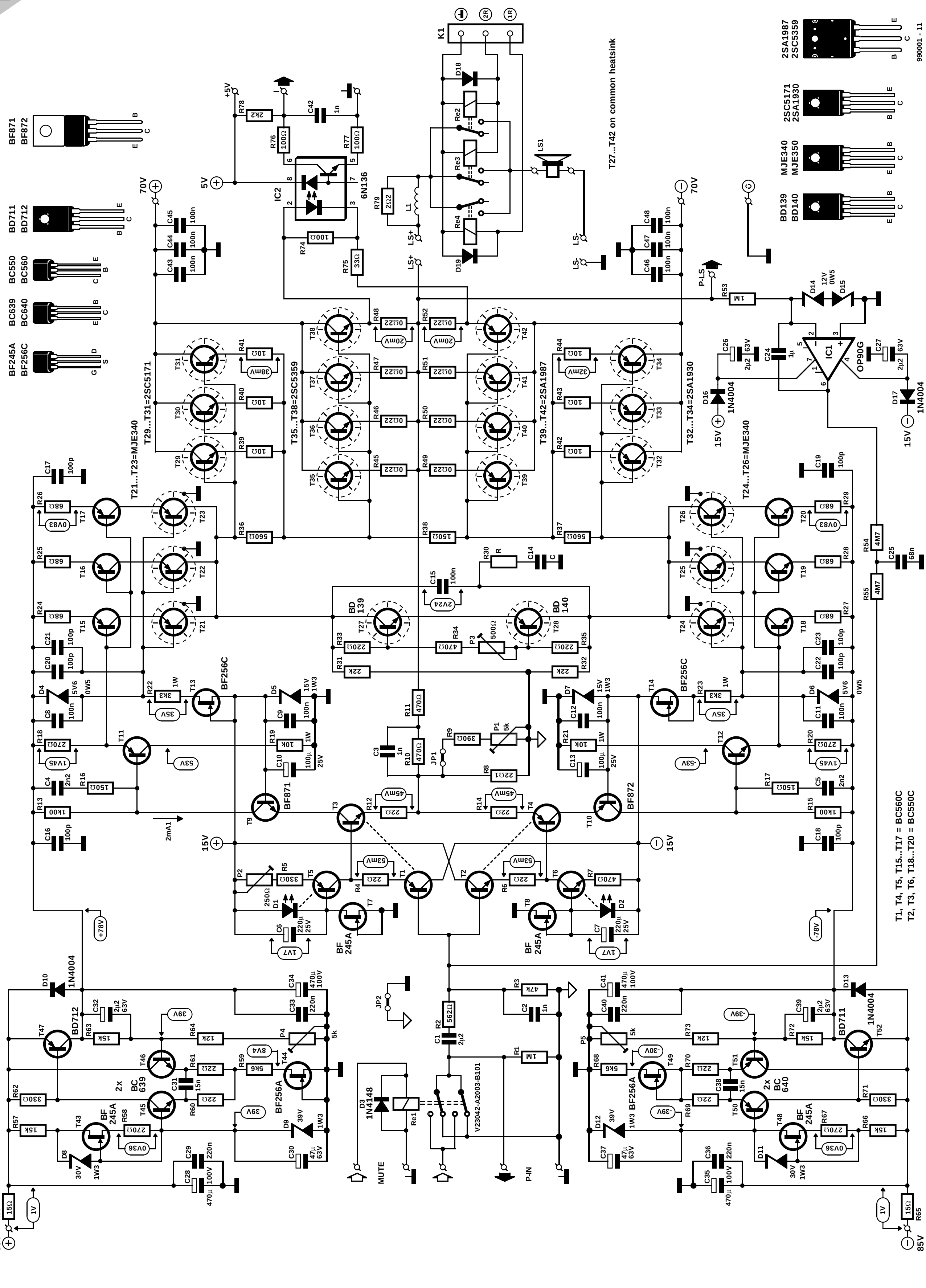
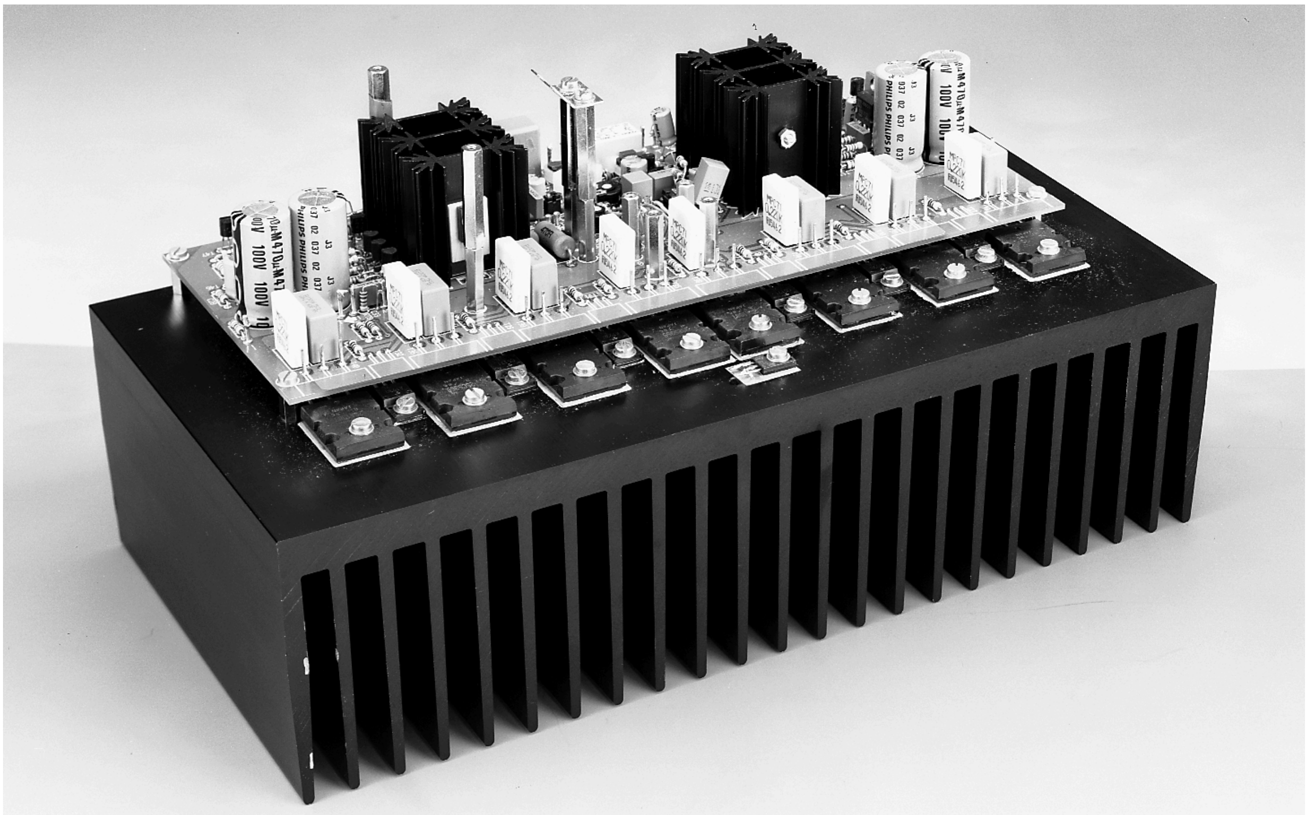


Figure 2. Although the circuit diagram gives the impression of a highly complex design, the amplifier is, in essence, fairly straightforward.



microsecond, and to which the amplifier can respond.

DESIGN CONSIDERATIONS

The Titan 2000 is based on the 'compact power amplifier' published in the May 1997 issue of this magazine. That was a typical domestic amplifier with a power output of 50 W into 8 Ω or 85 W into 4 Ω. The special property of this fully balanced design was the use of current feedback instead of voltage feedback, which resulted in a fast-responding amplifier with a large open-loop bandwidth. The amplifier performed well both as regards instrument test and measurements and listening tests. However, to serve as a basis for the Titan 2000, its output current and drive voltage range had to be increased substantially.

For a start, the supply voltage has to be more than doubled, which means that transistors with a higher power rating have to be used in the power supply. The higher supply voltage also results in larger potential drops across a number of components, and this means that dissipation problems may arise.

The large output current required for the Titan 2000 makes a complete redesign of the current amplifier used in the 'compact power amplifier' unavoidable, since that uses insulated-gate bipolar transistors (IGBTs). Although these are excellent devices, the large spread of their gate-emitter voltage makes their use in parallel networks next to impossible. To obtain the requisite output power, the use of parallel networks of symmetrical pairs of transistors is inevitable.

In view of the foregoing, bipolar transistors are used in the current amplifier of the Titan 2000. However, these cannot be driven as readily as IGBTs, which means that current drive instead of voltage drive is used. This entails a substantial upgrading of the driver stages and the preceding cascode amplifiers (which also consist of a couple of parallel-connected transistors). The good news is that the power transistors in the Titan 2000 are considerably less expensive than IGBTs: an important factor when eight of these devices are used.

Finally, the protection circuits have been enhanced in view of the higher voltages and currents. The circuits protecting against direct voltages and short-circuits are supplemented by networks protecting against overload and (too) high temperatures. The latter is coupled to a proportional fan control.

In short, a large part of the Titan 2000 is a virtually new design rather than a modified one.

BRIEF DESCRIPTION

The block diagram of the Titan 2000 is shown in **Figure 1.** The voltage amplifier consists of input stages T_1–T_{10}, and cascode amplifiers/pre-drivers T_{15}–T_{26}. The current amplifier is formed by driver transistors T_{27}–T_{34}, and output transistors T_{35}–T_{42}.

The offset control stage prevents any direct voltage appearing at the output of the amplifier.

The loudspeaker is linked to the amplifier by three heavy-duty relays.

The current amplifier operates from a ±70 V supply, which is provided by two 50 V mains transformers. To enable the voltage amplifier to drive the current amplifier to its full extent, it needs a slightly higher supply voltage to compensate for the inevitable losses caused by inevitable voltage drops. This is accomplished by superimposing a ±15 V potential from an external auxiliary supply on to the main ±70 V supply and dropping the resulting voltage to ±78 V with the aid of regulator T_{43}–T_{52}.

The combined protection circuits constantly compare the input and output voltage of the amplifier: any deviation from the nominal values leads to the output relays disconnecting the loudspeaker and the input relay decoupling the input signal.

The thermal protection circuit monitors the temperature of the heat sink and, if necessary, switches on a fan. If, with the fan operating, the temperature approaches the maximum permissible limit, the output relays are deenergized and disconnect the loudspeaker.

CIRCUIT DESCRIPTION

The circuit diagram of the Titan 2000 is shown in **Figure 2**. In spite of the large number of components, the basic cir-

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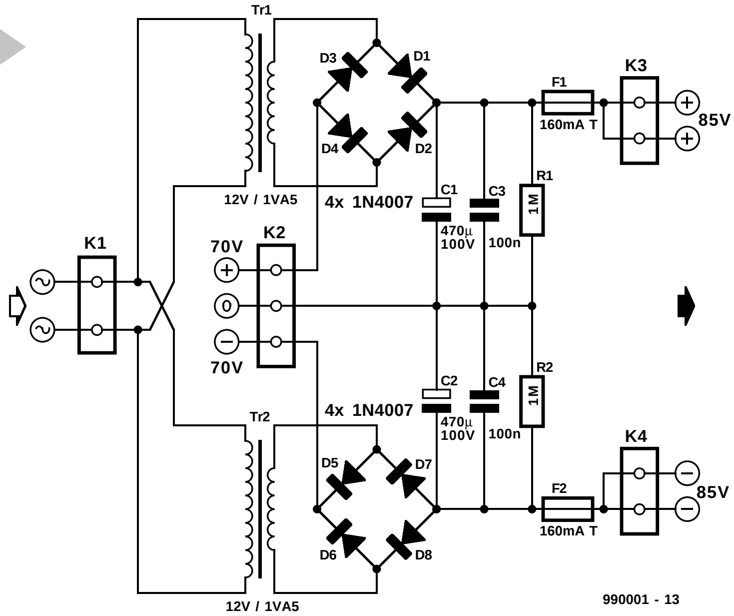


Figure 3. Circuit diagram of the requisite auxiliary power supply.

cuit is straightforward.

As already noted in the previous paragraph, transistors T_1–T_{10} form the input amplifier, T_{11} and T_{12} are buffers, T_{13} and T_{14} are current sources, T_{15}–T_{26} form the cascode amplifier/pre-driver stage, T_{27}–T_{34} are the driver transistors in the current amplifier, T_{35}–T_{42} are the output transistors, and T_{43}–T_{52} form a sophisticated supply voltage regulator.

Input amplifier

Strictly speaking, the input amplifier is formed by transistors T_3–T_4. Cascode stages T_9–T_{10} serve merely to enable the input section handling the high voltages. These voltages are limited by zener diodes D_5 and D_7, which are part of the potential divider that also sets the operating points of T_{21}–T_{26}. In view of the requisite stability, the current through the zener diodes is held constant by current sources T_{13} and T_{14}. Resistors R_{22} and R_{23} limit the potential across, and thus the dissipation in, these field-effect transistors.

Otherwise, the input section is virtually identical to that of the ‘compact power amplifier’. The drop across the emitter resistors of buffers T_1 and T_2 determines the drop across the emitter resistors of T_3 and T_4, and consequently the setting of the operating point of the overall input section. To eliminate the influence of temperature variations, T_1 is thermally coupled to T_3 and T_2 to T_4.

Since the operating point of buffers T_1 and T_2 is critical, current sources T_5 and T_6 have been added. The reference for these current sources is provided by light-emitting diodes (LEDs) D_1 and D_2. The current through these diodes is determined by current sources T_7 and T_8. In view of the requisite stability, diode D_1 is thermally coupled to T_5 and D_2 to T_6.

Any imbalance of the input stages is compensated by making the current through T_5 equal to that through T_6 with potentiometer P_2.

Cascode amplifiers/pre-drivers

The large output current of the Titan 2000 necessitates a proportionally large pre-drive voltage, which is provided by three parallel-connected cascode amplifiers, T_{15}–T_{26}. The current through these amplifiers is arranged at 10–15 mA, but the current feedback used may cause this level to be appreciably higher. This is the reason that the transistors used in the T_{21}–T_{26} positions are types that can handle currents of up to 50 mA when their collector-emitter voltage is 150 V.

The input section is linked to the cascode amplifiers by buffers T_{11} and T_{12}, which results in a lowering of the input impedance. The arrangement also enables an increase in the values of R_{13} and R_{15}, which results in a 3 dB increase in amplification of the input section.

The function of resistors R_{19} and R_{21} is threefold: they limit the dissipation of the buffers; they obviate the need of an additional voltage to set the operating point of the buffers; they limit the maximum current through the buffers, and thus the cascode amplifiers, to a safe value.

The open-loop amplification of the Titan 2000 is determined solely by those of the input section and cascode amplifiers. The amplification of the input section depends on the ratios R_{13}:($R_{12}+R_8$) and R_{15}:($R_{14}+R_8$) and, with values as specified is $\times 10$ (i.e., a gain of 20 dB).

The amplification of the cascode amplifiers is determined largely by the ratio of parallel-connected resistors R_{31} and R_{32} and the parallel network of R_{24}–R_{26}. With values as specified, the amplification is about $\times 850$ (remember, this is a push-pull design), so that the overall amplification of input section plus cascode amplifiers is $\times 8500$ (a gain of close to 80 dB).

Current amplifier

Since one of the design requirements is that the amplifier is to work with loads down to 1.5 Ω, the output stages consist of four parallel-connected pairs of transistors, T_{35}–T_{38} and T_{39}–T_{42}. These transistors have a highly linear transfer characteristic and provide a direct-current amplification that remains virtually constant for currents up to 7 A.

Like the output transistors, the driver stages need to remain within their safe operating area (SOA), which necessitates a threefold parallel network. The transistors used in the driver stages are fast types ($f_T = 200$ MHz).

Setting the bias voltage for the requisite quiescent current is accomplished by balanced transistors T_{27} and T_{28}. These transistors are mounted on the same heat sink as the output transistors and driver transistors to ensure good thermal coupling and current control. Of course, the current rises during full drive conditions, but drops again to its nominal level when the amplifier cools off. The quiescent current is set to 200 mA with potentiometer P_3.

Owing to the large output current, the connection between amplifier output and loudspeaker is not arranged via a single relay, but via three. Two of these, Re_3–Re_4, are controlled in synchrony by the protection circuits. When they are deenergized, their disabling action is delayed slightly to give the contacts of the third relay, Re_2, time to open, which is of importance in a fault situation.

Input relay Re_1 is switched off in synchrony with Re_2 to ensure that there is no input signal by the time Re_3 and Re_4 are deenergized.

Optoisolator IC_2 serves as sensor for the current protection circuits. The light-emitting diode in it monitors the voltage across R_{48}–R_{52} via potential divider R_{74}–R_{75}, so that the positive as well as the negative output currents are guarded. The use of an optoisolator prevents earth loops and obviates compensation of the ±70 V common-mode voltage. The +5 V supply for the optoisolator is derived from the protection circuits.

Feedback

The feedback loop runs from the output of the power stages to the junction of T_3 and T_4 via resistors R_{10} and R_{11}. This is current feedback because the current through T_3 and T_4 depends on the potential across R_8, which is determined largely by the current through R_{10} and R_{11}. The overall voltage amplification of the output amplifier is determined by the ratio R_8:($R_{10}+R_{11}$).

Compensation

Capacitors C_3–C_5 and resistors R_{16}, R_{17} form part of the compensation network required for stable operation.

Low-pass filter R_2–C_2 at the input is essential to prevent fast, that is, high-frequency, signals causing distortion. This filter is also indispensable for stability's sake.

Coupling capacitor C_1 is needed because the available offset compensation network merely redresses the bias current of the input buffers and is not intended to block any direct voltages at the input.

Relay Re_1 at the input enables the input signal to be 'switched off'. It forms part of the overall protection and in particular safeguards the input section against overdrive. The overall protection circuit will be discussed in detail next month.

Network R_9-P_1 is intended specifically for adjusting the common-mode suppression when two amplifiers are used in a bridge arrangement. It is needed for only one of these amplifiers, and may be interconnected or disabled by jumper JP_1 as needed.

Offset compensation is provided by integrator IC_1, which ensures that if there is any direct voltage at the output of the amplifier, the operating point of T_1-T_2 is is shifted as needed to keep the output at earth potential. The operational amplifier (op amp) used draws only a tiny current (20 µA) and has a very small input offset (450 µV).

Supply voltage for IC_1 is taken from the ±15 V line for the input section via diodes D_{16} and D_{17}. This arrangement ensures that the supply to the IC is retained for a short while after the main supply is switched off so that any interference is smoothed out.

Diodes D_{14} and D_{15} safeguard the input of IC_1 against (too) high input voltages in fault conditions.

The values of resistors R_{54} and R_{55} arrange the level of the compensating current at not more than 1 µA, which is sufficient to nullify the difference between the base currents of T_1 and T_2.

Regulation

Although current feedback has many advantages, it also has a serious drawback: poor supply voltage suppression. This makes it essential for the supply voltage for the voltage amplifier to be regulated. In view of the requisite high symmetrical potential and the fact that the unregulated voltage that serves as input voltage can vary substantially under the influence of the amplifier load, two discrete low-drop regulators, T_{43}–T_{47} and T_{48}–T_{52} are used.

As mentioned before, owing to inevitable losses through potential drops, the supply voltage for the input section and cascode amplifiers needs to be higher than the main ±70 V line. Furthermore, the input voltage to the regulators must be higher than the wanted output voltage to ensure effective regulation.

Fortunately, the current drawn by the voltage amplifier is fairly low (about 70 mA) so that the input voltage to the regulators can be increased with a simple auxiliary supply as shown in **Figure 3**. This consists of two small mains transformers,two bridge rectifiers, D_1–D_4 and D_5–D_8, and the necessary reservoir and buffer capacitors.

The ±15 V output is linked in series with the ±70 V line to give an unregulated voltage of ±85 V.

The 39 V reference is provided by zener diode D_9. This means that the regulator needs to amplify the reference voltage ×2 to obtain the requisite output voltage.

The zener diode is powered by current source T_{43}, to ensure a stable reference, which is additionally buffered by C_{30}.

Differential amplifier T_{45}-T_{46}, whose operating point is set by current source T_{44}, compares the output voltage with the reference via potential divider R_{63}-R_{64}-P_4. This shows that the output voltage level can be set with P_4.

Transistor T_{47} is the output stage of the regulator. The output voltage remains stable down to 0.2 V below the input voltage.

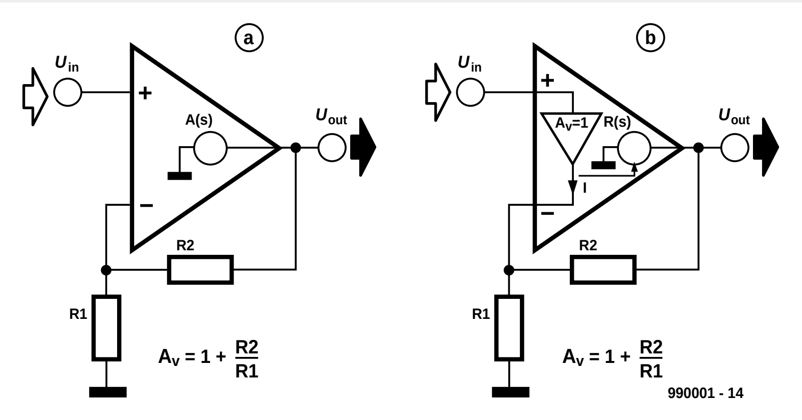
Current-feedback

In an amplifier using voltage feedback (Figure a), the differential voltage at its inputs is multiplied by the open-loop amplification. The feedback loop forces the output voltage to a level that, divided by network R_1-R_2, is equal to the input voltage.

Whereas an amplifier with voltage feedback has high-impedance inputs, an amplifier with current feedback (Figure b) has an high-impedance and a low-impedance input. Its input stage consists of a buffer with unitary gain between the inverting and non-inverting inputs. Essentially, the inverting input is the low-impedance input. The buffer is followed by an impedance matching stage that converts the output current of the buffer into a directly proportional output voltage.

The current feedback loop operates as follows. When the potential at the non-inverting input rises, the inverting input will also rise, resulting in the buffer current flowing through resistor R_1. This current, magnified by the impedance matching stage, will cause the output voltage of the amplifier to rise until the output current flowing through resistor R_2 is equal to the buffer current through R_1. The correct quiescent output voltage can be sustained by a very small buffer current. The closed-loop amplification of the circuit is determined by the ratio $(1+R_2)$:R_1.

A interesting property of an amplifier with current feedback is that the closed-loop bandwidth is all but independent of the closed-loop amplification, whereas that of an amplifier with voltage feedback becomes smaller in inverse proportion to the closed-loop amplification – a relation known as the gain-bandwidth product.



Resistor R_{57} and diode D_8 protect T_{43} against high voltage during switch-on, while D_{10} prevents current flowing through the regulator in the wrong direction.

Capacitors C_{31} and C_{32} enhance the rate of operation of the regulator.

Network R_{56}-C_{28}-C_{29} provides additional smoothing and r.f. decoupling of the ±85 V lines.

NEXT MONTH

Next month's second and concluding instalment of this article will describe details of the protection circuits, the fan control, and the construction of the amplifier. The instalment will also include detailed specifications and performance characteristics.

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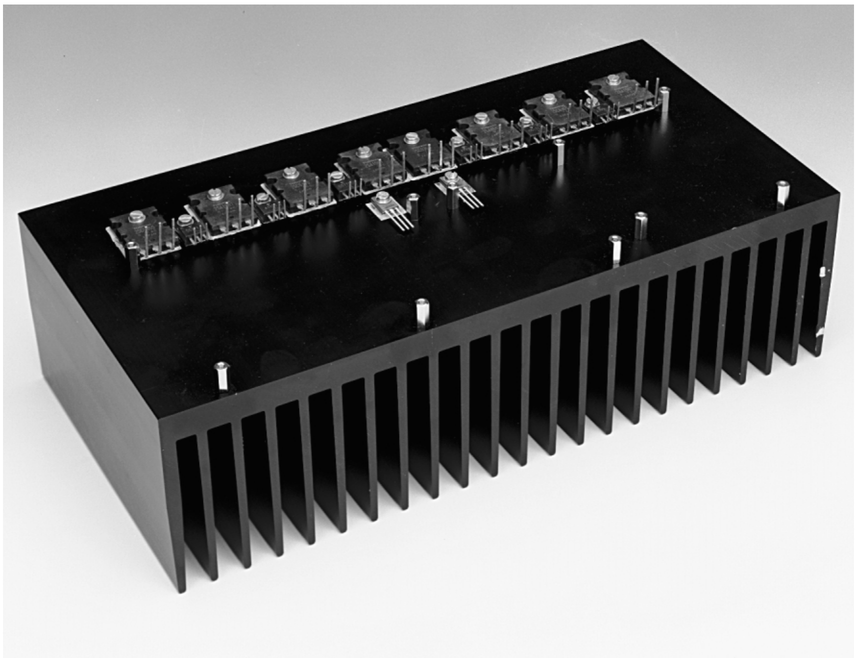

Titan 2000

Part 2: protection network

This second of four parts deals primarily with the protection network incorporated in the amplifier. This indispensable network safeguards the amplifier and the loudspeakers connected to it against all kinds of error that may arise. The network is an independent entity with its own power supply.

Correction. In last month's first part of this article, it was stated erroneously that the article consists of two parts, whereas in fact it will be described in four parts.

Design by T. Giesberts

INTRODUCTION

As mentioned briefly in Part 1, extensive and thorough protection is a must in an amplifier of this nature. It may well be asked why this is so: is there such a likelihood of mishaps arising? Or is the amplifier so vulnerable? On the contrary: extended tests on the prototype have shown that the Titan 2000 is a very stable and reliable piece of equipment. In fact, unusual means had to be used to actuate the protection circuits during these tests, since not any standard test prompted the amplifier into an error situation.

The extensive protection is necessary because by far the largest number of mishaps occur owing to actions by the user, not because of any shortcomings in the amplifier. For example, the most robust and reliable amplifier can not always cope with extremely high overdrive or overload conditions.

SIX FUNCTIONS

The integrated protection network consists of six sub-circuits:

- power-on delay
- transformer voltage sensor
- temperature sensor
- current sensor
- direct-current sensor
- overdrive sensor

The power-on delay ensures that the relays in the amplifier are energized 50–100 milliseconds after the supply has been switched on to prevent switch-on clicks.

The transformer voltage sensor reacts to the cessation of the secondary voltage of the mains transformers to prevent switch-off clicks and crackles.

The temperature sensor responds to excessive heat sink temperatures, but it should be noted that this works only in

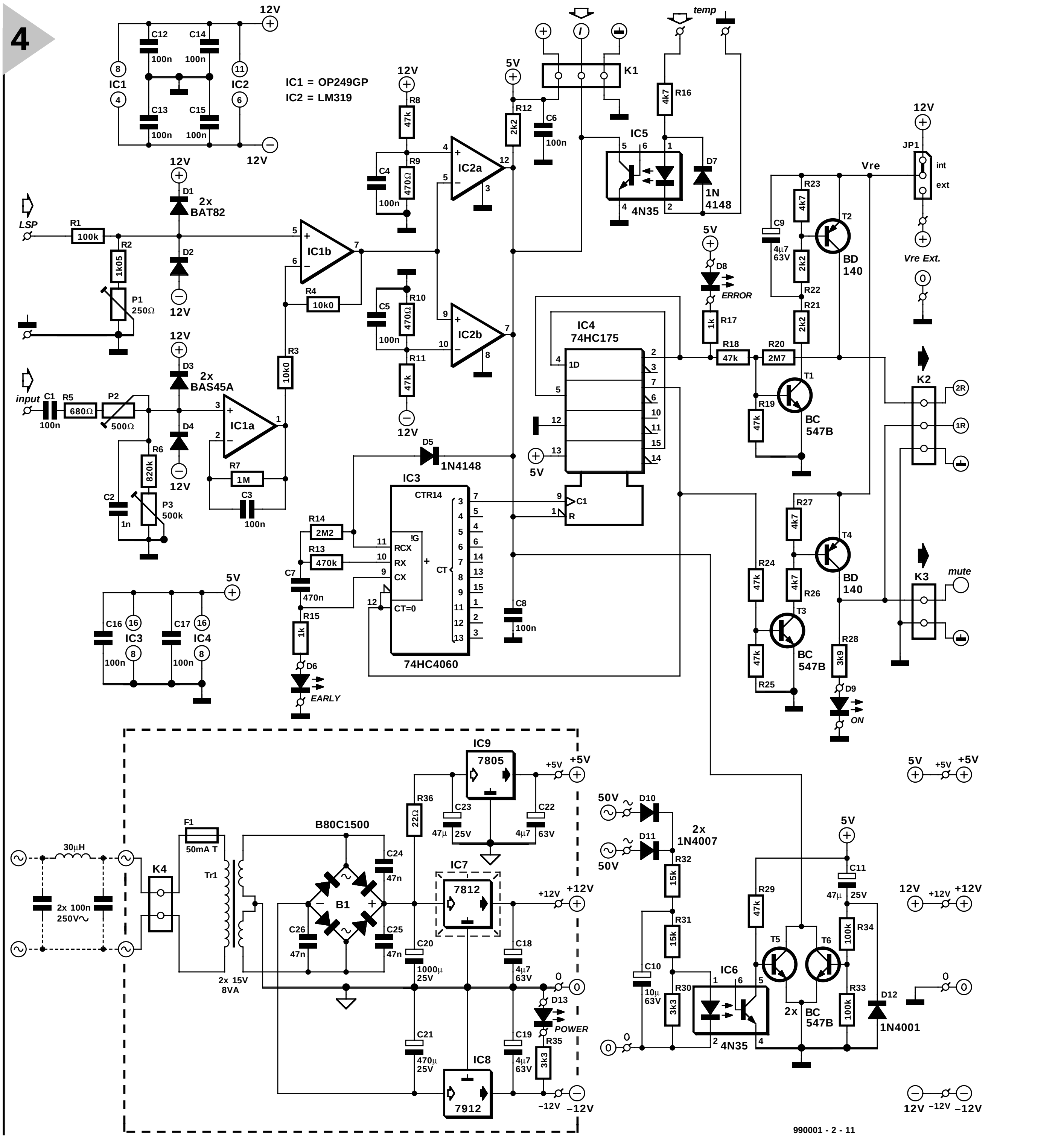

Figure 4. The protection network consists of six sensor circuits each of which causes the input and output relays of the amplifier to be deenergized when a fault occurs.

conjunction with the fan drive, which is reverted to later in this article.

The current sensor monitors the output current, while the direct-current and overdrive sensors form a combined circuit that monitors differences between the input and output signals, and reacts to excessive direct-current levels or distortion. This circuit is the most important and 'intelligent', but also the most complex of the six.

All sensors, when actuated, react in the same way: they cause the output relays and the mute relay at the input of the amplifier to be deenergized immediately. This action causes the input signal and the output load to be disconnected from the amplifier. After the fault causing the sensor action has been removed or remedied, the relevant protection circuit is disabled, whereupon the amplifier relays are reenergized after a short delay.

When the protection network is actuated, a red LED lights to indicate an error. When the fault has been removed or remedied, the red LED remains on, but a yellow LED flashes to indicate that the amplifier will be reenabled shortly. The red LED then goes out, shortly followed by the yellow, whereupon a green LED lights to indicate that all is well.

COMMON SECTION AND POWER-ON DELAY

The circuit of the integrated protection network, including the +5 V and ±12 V power supplies, is shown in **Figure 4**.

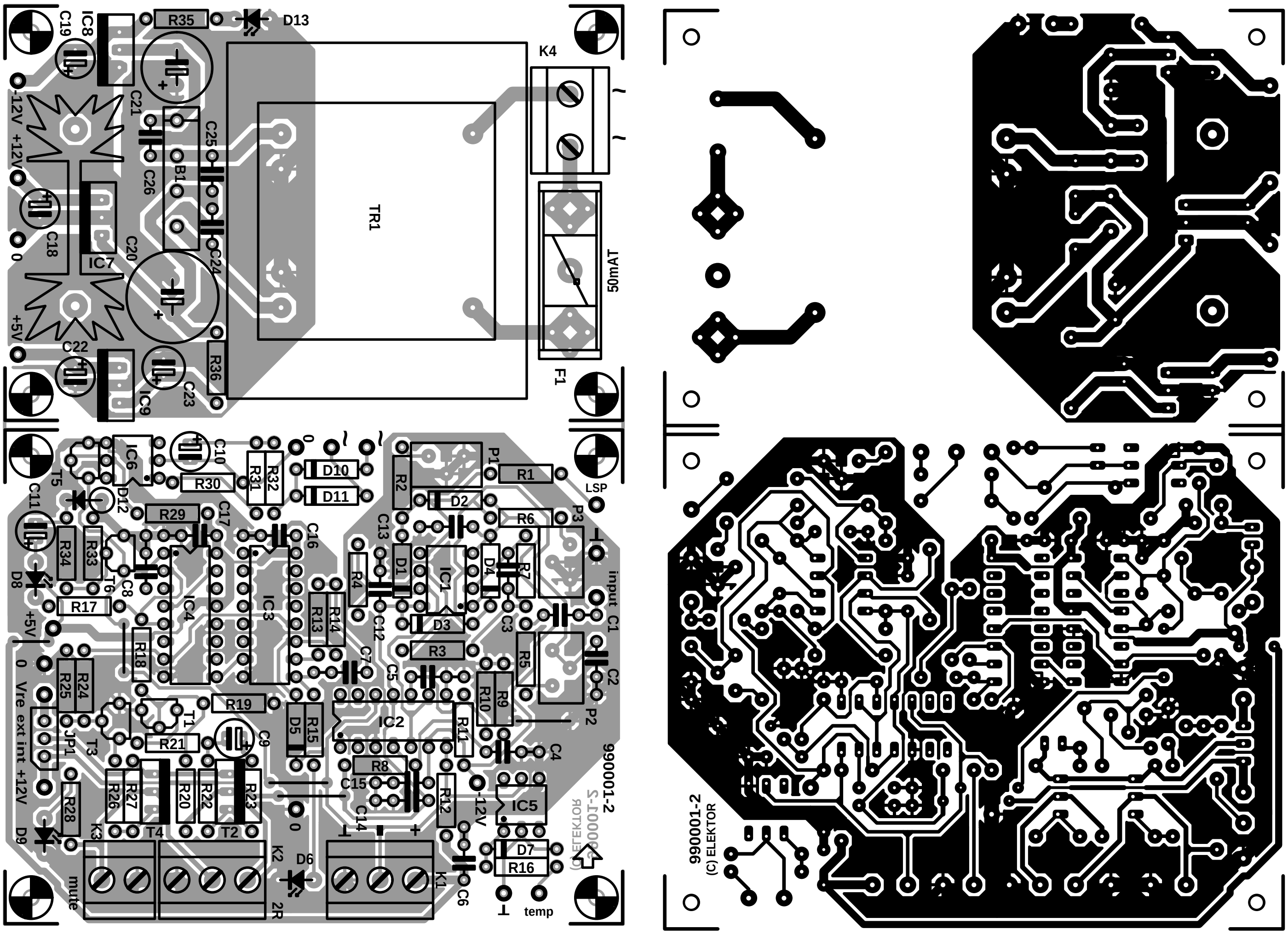

Figure 5. The printed-circuit board of the overall protection network.

Parts lists

Protection network

Resistors:
R_1, R_{33}, R_{34} = 100 kΩ
R_2 = 1.05 kΩ
R_3, R_4 = 10.0 kΩ
R_5 = 680 Ω
R_6 = 820 kΩ
R_7 = 1 MΩ
R_8, R_{11}, R_{18}, R_{19}, R_{24}, R_{25}, R_{29} = 47 kΩ
R_9, R_{10} = 470 Ω
R_{12}, R_{21}, R_{22} = 2.2 kΩ
R_{13} = 470 kΩ
R_{14} = 2.2 MΩ
R_{15}, R_{17} = 1 kΩ
R_{16}, R_{23}, R_{26}, R_{27} = 4.7 kΩ
R_{20} = 2.7 MΩ
R_{28} = 3.9 kΩ
R_{30}, R_{35} = 3.3 kΩ
R_{31}, R_{32} = 15 kΩ
R_{36} = 22 Ω
P_1 = 250 Ω, multiturn preset (upright)
P_2 = 500 Ω, multitun preset (upright)
P_3 = 500 kΩ, multiturn preset (upright)

Capacitors:
C_1, C_3 = 0.1 µF
C_2 = 0.001 µF
C_4, C_5, C_6, C_8, C_{12}–C_{17} =0.1 µF, ceramic
C_7 = 0.47 µF
C_9, C_{18}, C_{19}, C_{22} = 4.7 µF, 63 V, radial
C_{10} = 10 µF, 63 V, radial
C_{11}, C_{23} = 47 µF, 25 V, radial
C_{20} = 1000 µF, 25 V, radial
C_{21} = 470 µF, 25 V, radial
C_{24}–C_{26} = 0.047 µF, ceramic

Semiconductors:
D_1, D_2 = BAT82
D_3, D_4 = BAS45A
D_5, D_7 = 1N4148
D_6, D_8, D_9, D_{13} = 3 mm high-efficiency LED (yellow, red, green, green respectively)
D_{10}, D_{11} = 1N4007
D_{12} = 1N4001
T_1, T_3, T_5, T_6 = BC547B
T_2, T_4 = BD140

Integrated circuits:
IC_1 = OP249GP (Analog Devices)
IC_2 = LM319N
IC_3 = 74HC4060
IC_4 = 74HC175
IC_5, IC_6 = 4N35
IC_7 = 7812
IC_8 = 7912
IC_9 = 7805

Miscellaneous:
JP_1 = 2.54 mm pin strip and pin jumper
K_1, K_2 = 3-way terminal block, pitch 5 mm
K_3 = 2-way terminal block, pitch 5 mm
K_4 = 2-way terminal block, pitch 7.5 mm
B_1 = bridge rectifier, rectangular, Type B80C1500
F_1 = fuse, 50 mAT and fuse holder
Tr_1 = mains transformer, 15 VA, with 2×15 V secondary
Heat sink (for IC_7) = e.g. Fischer SK104, 50 mm
Mains interference filter

The network is linked to the input and output of the amplifier via terminals ‘input’ and ‘LSP’ respectively (to terminals ‘P-IN’ and ‘P-LS’ on the amplifier board).

Terminals ‘50 V≈’ are connected to the secondary windings of the mains transformers.

The three output relays and the mute relay in the amplifier are linked to the protection network via K_2, and K_3 respectively.

The current sensor is connected to the output of optoisolator IC_2 in the amplifier (‘I->’ on the amplifier board) via K_1.

The terminals marked ‘temp’ are intended to be linked to the output of the fan control circuit.

As mentioned earlier, the action of each sensor results in the deenergizing of the output and mute relays in the amplifiers. This implies that the outputs of the the various sensor circuits are interlinked. This is effected by combining the open-collector outputs of these circuits into a wired OR gate with R_{12} functioning as the common pull-up resistance. The combined output signal serves to reset a number of

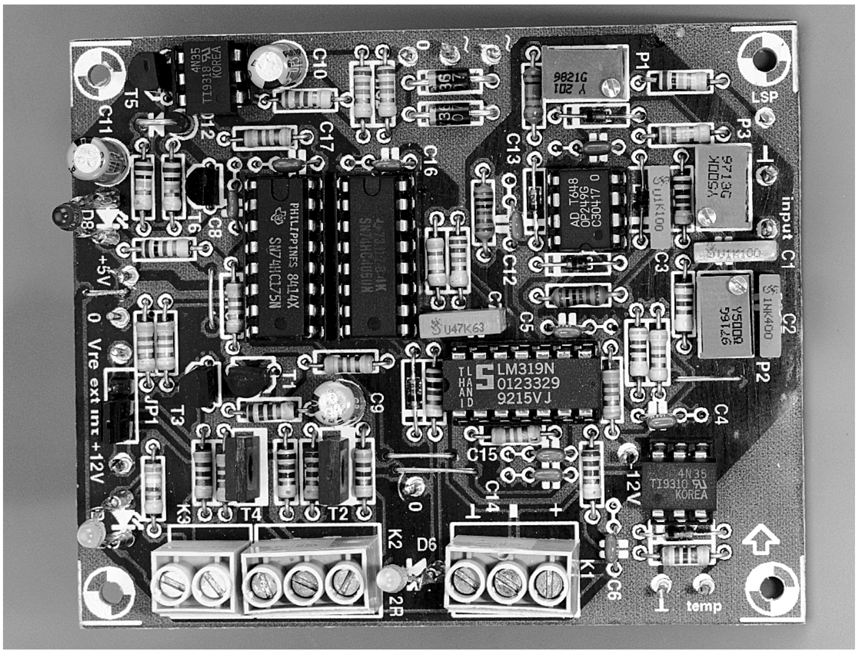

Figure 6. Completed prototype of the protection network.

D-type bistables (flip-flops), contained in IC_4, which are interconnected to form a shift register. Note that D-type bistables are essential since these can be set and reset in a defined manner.

The outputs of IC_4 are used to drive two level converters, T_1-T_2 and T_3-T_4 respectively, which bridge the difference between the 5 V level of the logic ICs and the 12 V supply for the relays. Jumper JP_1 enables a different, external supply voltage (V_{RE}) to be used if 12 V relays are not employed.

Transistors T_1 and T_2 drive Re_1 and Re_2, which are the first to be energized (synchronously). On switch-off, capacitor C_9 ensures that T_2 remains on for some milliseconds longer during which period Re_3 and Re_4 are deenergized (see Part 1).

The power-on delay, which also operates after a fault situation, is more complex than usual. To start with, after the supply voltage us switched on, input CLR of IC_4 is held low (active) for a few seconds by the circuit around T_6. When, after this period, CLR is made high by R_{12} –which happens only when there is no error situation (any longer)–the internal oscillator of IC_3 is enabled via D_5. This results after a few seconds in a clock pulse appearing at the CLK input of IC_4 , whereupon Q_4 goes high. The period between the oscillator being enabled and the appearance of the first clock pulse is not defined since, owing to the presence of T_6, a power-on reset is purposely not provided. To ensure a minimum delay in the energizing of Re_1 and Re_2 in spite of this, a high level is clocked into Q_4 after IC_3 has been enabled. The precise moment at which this happens varies, therefore, only when the supply voltage is switched on for the first time.

A period of IC_3/Q_3 later, Q_1 of IC_4 goes high, whereupon Re_1 and Re_2 are energized. After another period, Q_2 of IC_4 becomes high, whereupon Re_3 and Re_4 are energized. At the same time, IC_3 is disabled since its reset is interlinked with Q_2 of IC_4.

The red LED, D_8, in parallel with Q_1 of IC_4 lights when the relays in the amplifier are not energized, either because the amplifier is (not yet) switched on, or owing to an error.

The yellow LED, D_6, is linked to the output of the oscillator in IC_3, causing it to flash until IC_4 is clocked.

The green LED, D_9, is connected in parallel with Re_3 and Re_4, so that it lights only when the amplifier is fully switched on.

TRANSFORMER VOLTAGE SENSOR

The 50 V≈ secondary voltages of the mains transformers in the amplifier are rectified by diodes D_{10} and D_{11}, and smoothed by R_{30}-R_{31}-R_{32}-C_{10}. The values of these components ensure that the LED in optoisolator IC_6 lights sufficiently to hold the associated photo transistor on. This transistor pulls the base of T_5 to ground, causing T_5 to cut off. When the secondary voltages fail, T_5 is switched on immediately via R_{29}, whereupon the D-type bistables in IC_4 are reset.

Use is made of an optoisolator purposely to avoid any risk of earth loops between the supply return and the ground of the protection network, which is linked to the input ground of the amplifier.

TEMPERATURE SENSOR

The temperature sensor works in a manner similar to that of the transformer voltage sensor. The optoisolator in this circuit is IC_5, which, in contrast to IC_6, is normally cut off and comes on only when the heat sink becomes excessively hot.

The sensor reacts to the fan control circuit switching the fan speed to maximum (because the heat sink is getting too hot). A comparator in the fan control circuit then toggles, whereupon IC_5 is actuated via the 'temp' input and resets the D-type bistables in IC_4. This situation changes only after the heat sink has cooled down to an acceptable temperature (although the fans may still be rotating).

CURRENT SENSOR

To nullify high common-mode voltages and to prevent any risk of earth loops, the current sensor also uses an optoisolator, IC_2 (**Figure 5**). However, this is not located on the protection board, but directly at the output of the amplifier.

The values of the relevant components cause the sensor to be actuated when the output current is about 40 A. This may appear a very large current, but this is due entirely to the specified requirement that the amplifier must be capable of delivering 60 V into a load of 1.5 Ω without the protection circuit being actuated. The current level may be lowered to some extent by increasing the value of R_{74} in the amplifier.

Output resistor R_{78} is in parallel with R_{12} by linking terminals 'I', '+5 V' and ground on the amplifier board to K_1 on the protection board via three lengths of insulated, stranded circuit wire twisted together. This arrange-

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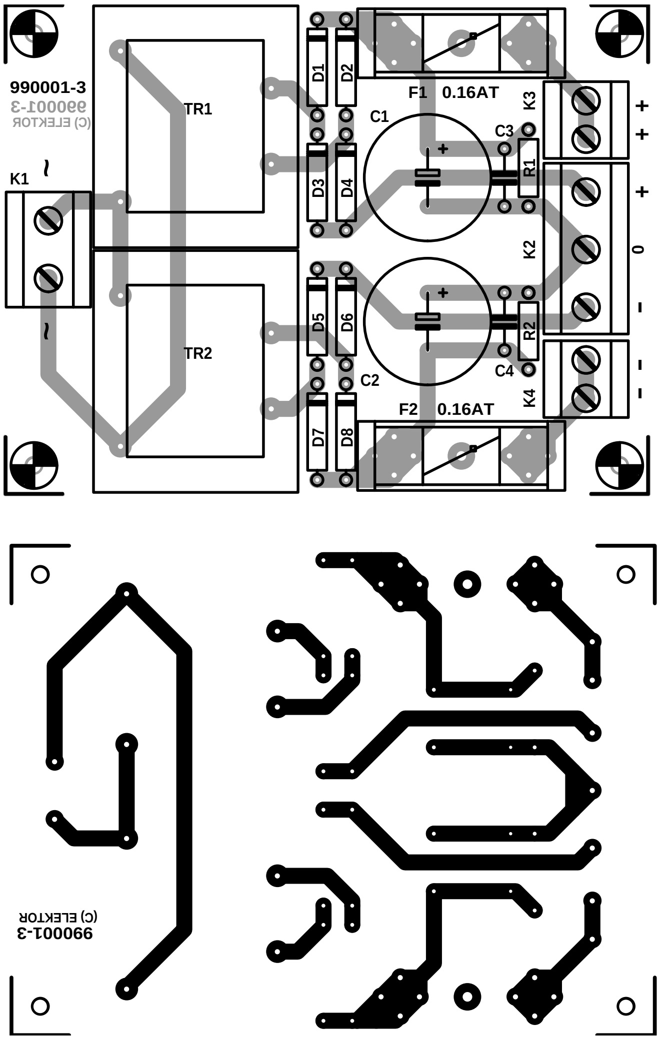

Figure 7. Printed-circuit board for the auxiliary power supply described in Part 1.

Parts lists

Auxiliary power supply

Resistors:
R_1, R_2 = 1 MΩ

Capacitors:
C_1, C_2 = 470 µF, 100 V, radial
C_3, C_4 = 0.1 µF, 100 V, pitch 7.5 mm

Semiconductors:
D_1–D_8 = 1N4007

Miscelleneous:
K_1 = 2-way terminal block, pitch 7.5 mm
K_2 = 3-way terminal block, pitch 7.5 mm
K_3, K_4 = 2-way terminal block, pitch 5 mm
Tr_1, Tr_2 = mains transformer, 1.5 VA, with 12 V secondart
F_1, F_2 = fuse, 160 mAT, and fuse holder

ment ensures a low impedance to any interference and a high reaction speed.

DIRECT-CURRENT AND OVERDRIVE SENSOR

The d.c. and overdrive sensor constantly compares the input and output signals of the amplifier and reacts when the difference between the two is too great. The comparison is effected with the aid of operational amplifier IC_1 which has a very low bias current and a very low offset. It is, of course, essential that during the comparison of the two signals by differential amplifier IC_{1b} the differences in phase and transit times do not lead to error detection. At the same time, the voltage amplification (×43) of the amplifier must be taken into account.

The amplification is compensated by potential divider R_1-R_2-P_1 at input LSP. The potentiometer is a multiturn type to ensure accurate adjustment.

The phase difference is compensated by the circuit based on IC_{1a}. The transit at high and low cut-off points is simulated by first-order networks that can also be adjusted very accurately with multiturn potentiometers P_2 and P_3.

The inputs of IC_{1a} and IC_{1b} are protected by diodes. Since any leakage current of these diodes, combined with the high input impedance (≈ 1 MΩ) of IC_{1a}, might lead to an appreciable offset, and therefore to an unwanted error detection, the diodes, D_3 and D_4, are special types with a leakage current of only 1 nA.

The output of differential amplifier IC_{1b} is monitored by a window comparator formed by IC_{2a} and IC_{2b}. The value of the components used in potential dividers R_8-R_9 and R_{10}-R_{11} ensures that the protection circuit is actuated when the direct voltage reaches a level of ±5 V or the distortion becomes 2.5 per cent. Such distortion will normally be the result of overdrive, but the circuit reacts equally well to oscillations or other spurious signals that cause too large a difference to be detected.

CONSTRUCTION AND SETTING UP

The integrated protection network is best built on the printed-circuit board shown in **Figure 5**. Populating this board should not present any undue difficulties, but it should be noted that diodes D_6, D_8, D_9 and D_{13}, are not located on the board, but are linked to it via flexible, stranded circuit wire. They are fitted to the front of the enclosure.

Jumper JP_1 will normally be in position 'intern' unless relays with a coil voltage other than 12 V are used.

A prototype of the completed protection board is shown in **Figure 6**.

All input and output terminals of the board are clearly marked with the same symbols as shown in Figure 4. Most interconnections can be made in thin, stranded hook-up wire to DEF61-12, but the input and output links ('input' and 'LSP') must be screened audio cable.

Although the power supply for the protection network can be fitted on the same board, the relevant section may be cut off and fitted elsewhere. Of course, the supply lines must then be linked to the relevant terminals on the protection board via insulated, stranded hook-up wire.

The power supply is straightforward. From the secondary output of the specified mains transformer, Tr_1, a symmetrical ±12 V supply is obtained with the aid of regulators IC_7 and IC_8. From the same secondary, a +5 V supply for the digital circuits is obtained with the aid of regulator IC_9. Since the relays are fed by the +12 V line, regulator IC_7 must be fitted on a heat sink.

To ensure that the protection network is not actuated by interference on the mains supply, it is advisable to precede the power supply by a suitable noise filter. This may be made from a 30 µH choke and two 0.1 µF, 300 V≈ capacitors as shown in dashed lines in Figure 4.

The network is set up by maximizing the common-mode suppression with the aid of an oscilloscope or a multimeter with sufficient bandwidth. Measurements need to be made at 1 kHz, 20 kHz, and 20 Hz. The open-circuit amplifier is driven as far as possible by a suitable sine-wave generator or CD player with a test CD.

With a signal of 1 kHz, set P_1 for minimum sign al at the output of IC_{1b}, follow this with a signal of 20 kHz and adjusting P_2, and finally, with a signal of 20 Hz, by adjusting P_3. Since the settings influence one another to some extent, the potentiometers should be set a couple of times, perhaps also at some different audio frequencies.

POWER SUPPLY

The auxiliary power supply described in Part 1 is best constructed on the printed-circuit board shown in **Figure 7**. The mains voltage is linked to K_1, the ±70 V to K_2 and the +85 V and –85 V lines to K_3 and K_4 respectively. Since all currents are low level, the wiring may be made in thin, insulated, stranded hook-up wire. A completed prototype board is shown in **Figure 8**.

The main supply for the amplifier is a straightforward, unregulated type, providing an output of ±70 V. Its circuit diagram is shown in **Figure 9**.

Since the specified requirements call for a 2 Ω load, the supply must be rated at 1000 VA, which necessitates two toroidal transformers. To prevent unforeseen equalizing currents, the dual secondaries are not linked in parallel, but are individually connected to a bridge rectifier. The outputs of the rectifiers can be connected in parallel without any problem. The rectifiers need to be mounted on a suitable heat sink such as a Type SK01.

It should be clear that the wiring of the power supply must allow for the large output currents of the amplifier. In the prototype, the electrolytic capacitors are linked by 3 mm thick strips of aluminium. The remainder of the wiring should be in insulated, high-current wire to BS6231 with a conductor size of 50/0.25 mm (2.5 mm^2) or better. The use of car-type connectors is recommended.

Note that the power supply as described is intended for use with a mono(phonic) amplifier that can deliver 800 W into 2 Ω and should remain stable with loads of 1.5 Ω. If you are certain that you will always use 4 Ω or 8 Ω loads, the power supply requirements may be relaxed to some extent. A reasonable relaxation is the use of 2×50 V/300 VA transformers and 10,000 µF/100 V smoothing capacitors. The rating of the primary fuses may then be reduced to 1.5 AT.

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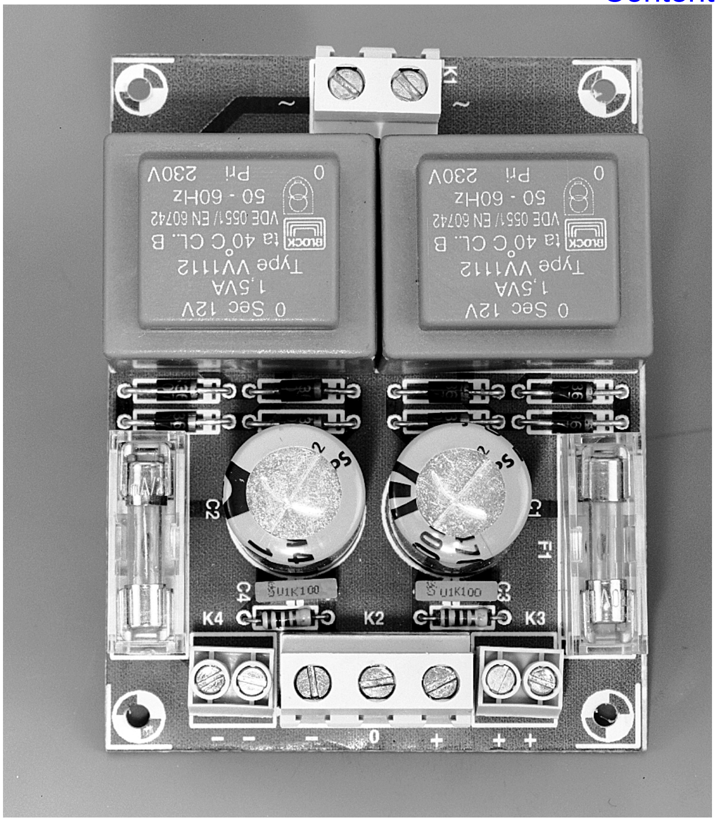

Figure 8. The auxiliary power supply is small enough to fit in most enclosures.

MAINS-ON DELAY

The use of a mains-on delay is recommended when heavy loads are to be switched on, as in the case of the present amplifier. Such a delay circuit switches on the mains to the load gradually to ensure that the switch-on current remains within certain limits and to prevent the mains fuses from blowing.

The most recently published (in this magazine) mains-on delay is found in the July/August 1997 issue (p. 74), whose circuit diagram is reproduced in **Figure 10**. Its printed-circuit board is readily connected with the primary windings of the two mains transformers. The board is not available ready-made, however, and its diagram is, therefore, reproduced in **Figure 11**.

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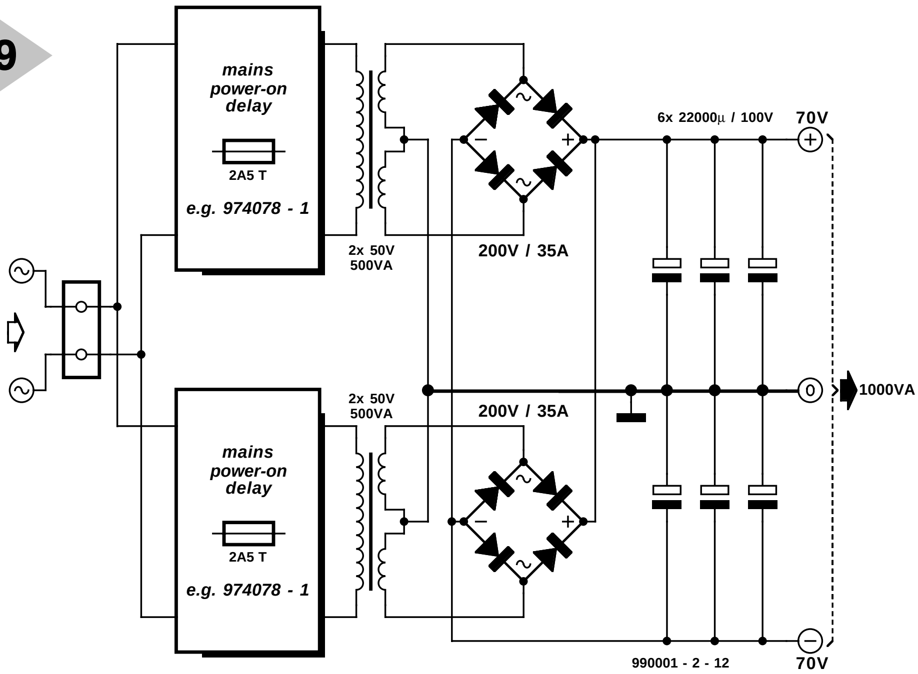

Figure 9. The main power supply for the amplifier is a heavy-duty entity in which the six capacitors are particularly impressive.

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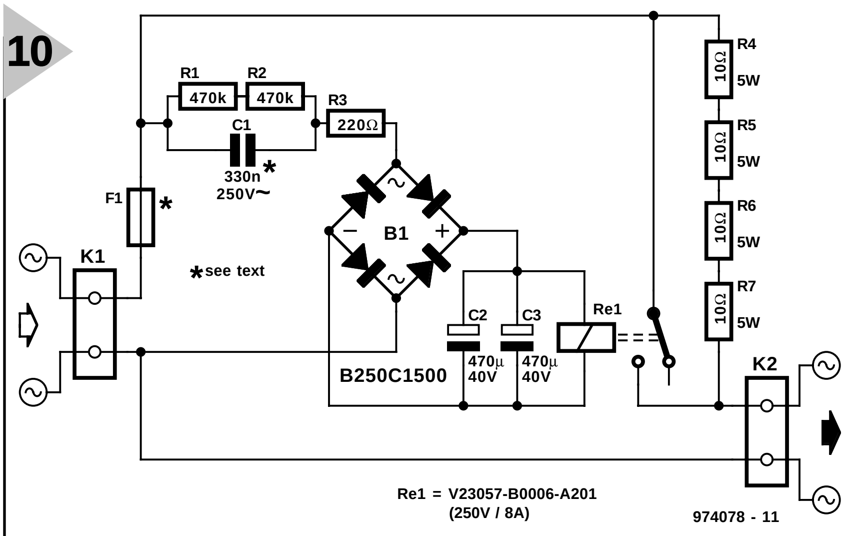

Figure 10. The mains-on delay ensures that the switch-on current remains within certain limit. Two of these delays are required for each Titan 2000.

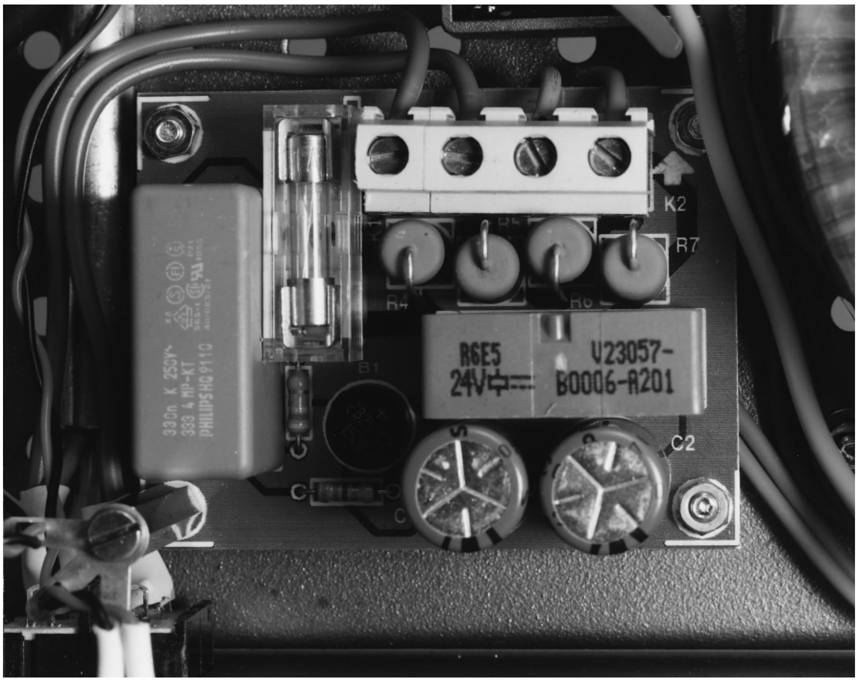

Figure 11. Printed-circuit board for the mains-on delay circuit, which is not available ready made.

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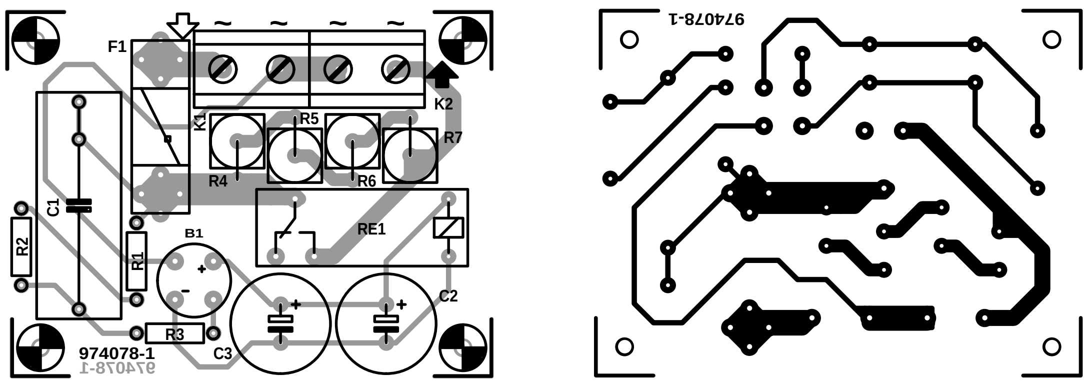


The delay arranges for the load, that is, the Titan 2000, to be switched on in two stages. In the first of these, the switch-on current is limited by series network R_4–R_7. After the delay determined by capacitors C_2 and C_3, the series network is shorted by a relay contact, whereupon the full current flows between K_1 and K_2.

Relay Re_1 can switch up to 2000 VA. Its supply voltage is obtained from the mains with the aid of rectifier B_1, capacitor C_1 and resistor R_3.

Since the amplifier power supply uses two mains transformers, two mains-on delay circuits are needed.

Fuse F_1 functions as a primary mains fuse for the amplifier.

Capacitor C_1 is a metallized paper type intended especially for use with mains voltage applications.

Bear in mind that the circuit is linked directly to the mains supply and thus carries lethal voltages.

Next month's third instalment of this article deals with the construction of the amplifier, a few other practical matters, and some measurements.

[990001-2]

Parts lists

Mains-on delay circuit

Resistors:
R_1, R_2 = 470 kΩ
R_3 = 220 Ω
R_4–R_7 = 10 Ω, 5 W

Capacitors:
C_1 = 0.33 µF, 300 V a.c.
C_2, C_3 = 470 µF, 40 V

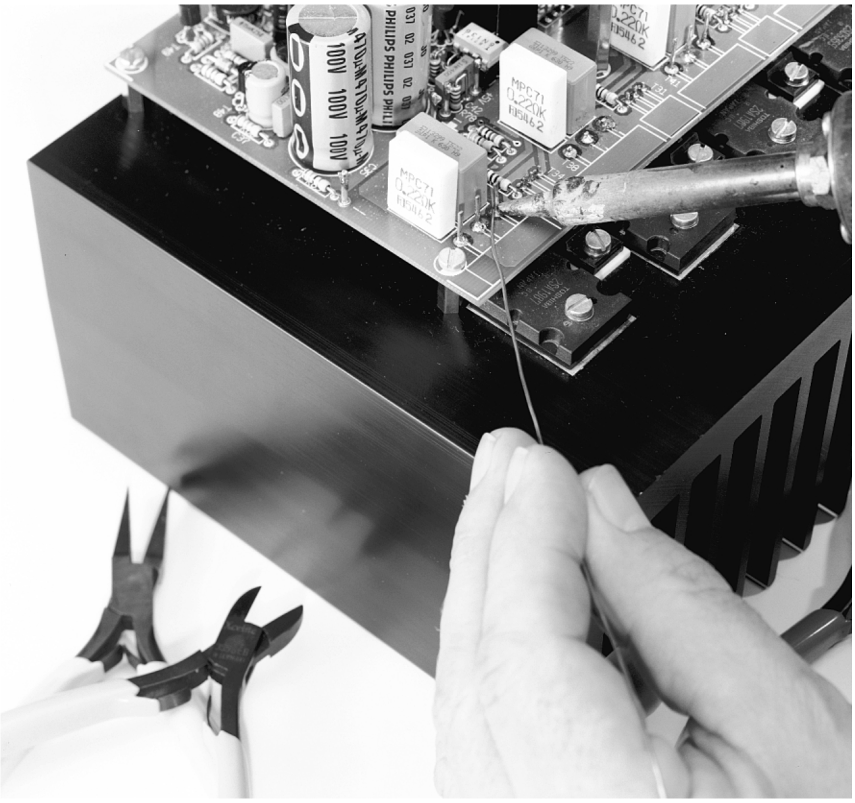
Miscellaneous:
K_1, K_2 = 2-way terminal block, pitch 7.5 mm
B_1 = bridge rectifier, round, Type B250C1500
Re_1 = relay, coil 12 V, 1200Ω; contact rating 250 V, 8 A
F_1 = see text

Contents

Titan 2000

Part 3: construction and setting up

This third of four parts deals primarily with the construction of the amplifier and ends with a brief resume of its performance and specifications. Let the constructor beware, however: the Titan 2000 is not an easy project and certainly not recommended for beginners in electronic construction.



Design by T. Giesberts

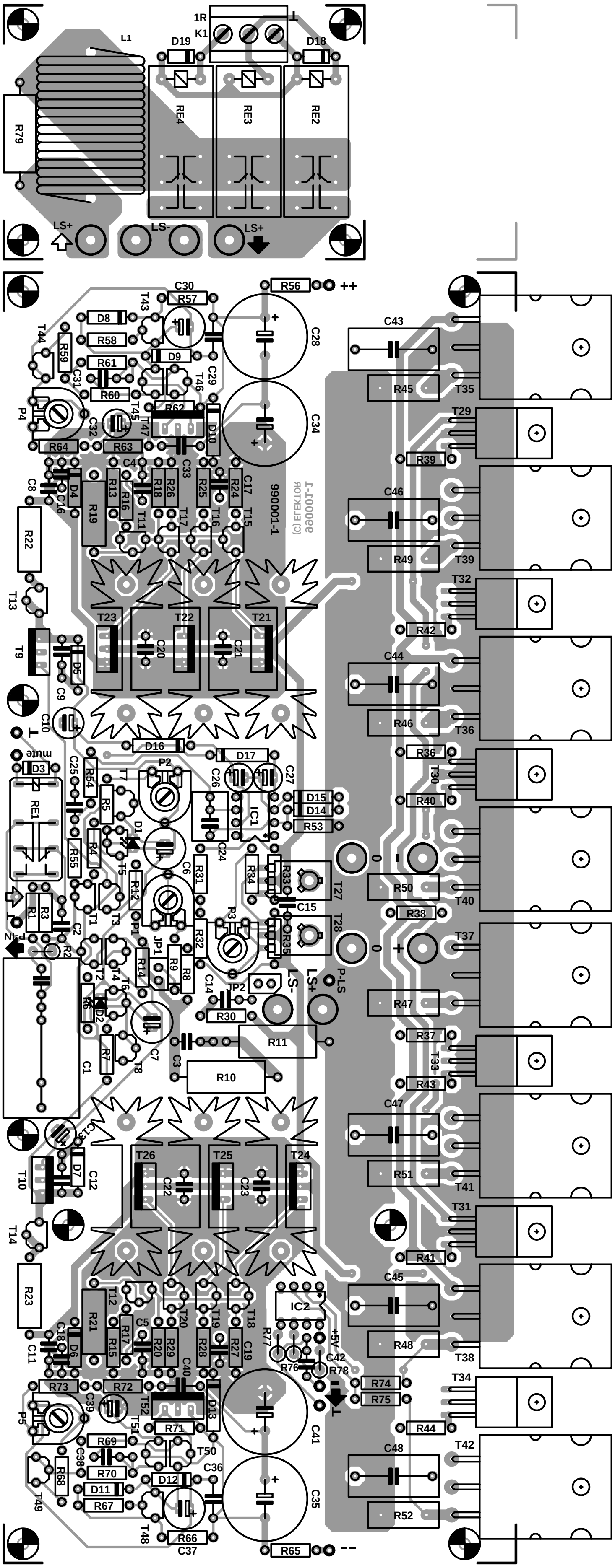
INTRODUCTION

It is clear from the first two parts of this article that the Titan 2000 is a complex unit that needs to be constructed and wired up with with great care to ensure the specified performance. For that reason, the construction notes will be more detailed than is usual with projects in this magazine. It is assumed that the protection network and auxiliary power supply have already been built and tested.

MOTHER BOARD

It must be borne in mind that in the case of a fast power amplifier like the Titan 2000, with a gain/bandwidth product of about 0.5 GHz, the board must be an integral part of the circuit. The mother board is therefore designed together with the remainder of the circuit. The length of the tracks, the area of the copper pads, the positions of the decoupling capacitors, and other factors, are vital for the proper and stable operation of the unit. Constructors who make their own boards are therefore advised to adhere strictly to the published layout.

Owing to the power requirements, the various stages are parallel configurations. When these are mounted on the heat sinks, a fairly large parasitic capacitances to earth ensue. This is because for reasons of stability all seven heat sinks must be strapped to earth. It



Parts lists

It is regretted that, owing to circumstances beyond our control, component codings in the various sections have been duplicated. Consequently, the mother board, protection network board, and auxiliary power supply board contain many components with the same identification (R_1-R_{36}, C_1-C_{26}, D_1-D_{12}, T_1-T_6, IC_1-IC_2, JP_1, K_1).

Amplifier

Resistors:

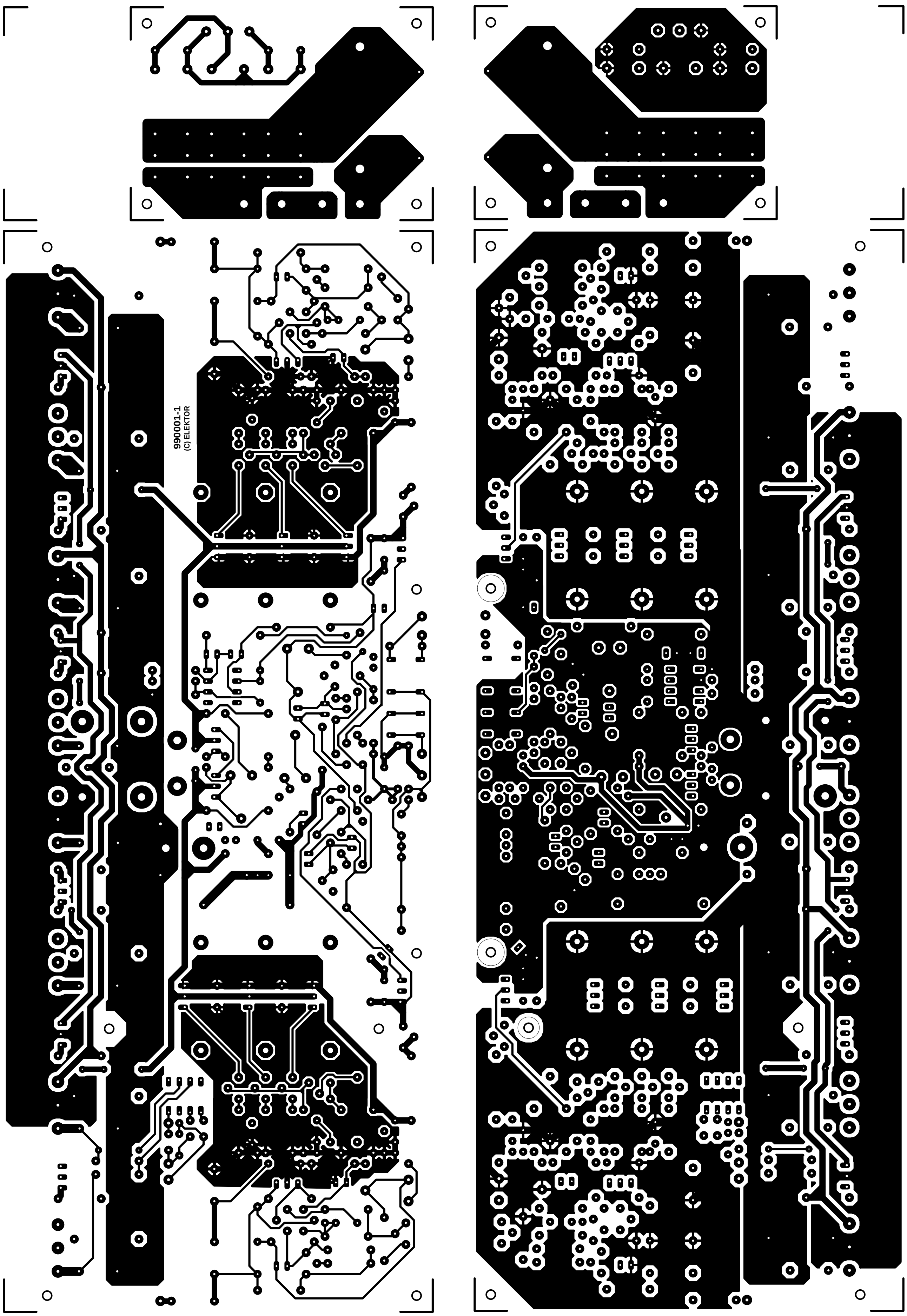
R_1, R_{53} = 1 MΩ

R_2 = 562 Ω
R_3 = 47 kΩ
R_4, R_6, R_{12}, R_{14}, R_{60}, R_{61}, R_{69}, R_{70} = 22 Ω
R_5, R_{62}, R_{71} = 330 Ω
R_7, R_{34} = 470 Ω
R_8 = 22.1 Ω
R_9 = 390 Ω
R_{10}, R_{11} = 470 Ω, 5 W
R_{13}, R_{15} = 1.00 kΩ
R_{16}, R_{17}, R_{38} = 150 Ω
R_{18}, R_{20}, R_{58}, R_{67} = 270 Ω
R_{19}, R_{21} = 10 kΩ, 1 W
R_{22}, R_{23} = 3.3 kΩ, 1 W
R_{24}–R_{29} = 68 Ω
R_{30} = see text
R_{31}, R_{32} = 22 kΩ
R_{33}, R_{35} = 220 Ω
R_{36}, R_{37} = 560 Ω
R_{39}–R_{44} = 10 Ω
R_{45}–R_{52} = 0.22 Ω, inductance-free
R_{54}, R_{55} = 4.7 MΩ
R_{56}, R_{65} = 15 Ω
R_{57}, R_{63}, R_{66}, R_{72} = 15 kΩ
R_{59}, R_{68} = 5.6 kΩ
R_{64}, R_{73} = 12 kΩ
R_{74}, R_{76}, R_{77} = 100 Ω
R_{75} = 33 Ω
R_{78} = 2.2 kΩ
R_{79} = 2.2 Ω, 5 W
P_1, P_4, P_5 = 4.7 kΩ (5 kΩ) preset
P_2 = 250 Ω, preset
P_3 = 500 Ω, preset

Capacitors:

C_1 = 2.2 µF, metallized polyester (MKP)
C_2, C_3, C_{42} = 0.001 µF
C_4, C_5 = 0.0022 µF
C_6, C_7 = 220 µF, 25 V, radial
C_8, C_9, C_{11}, C_{12}, C_{15} = 0.1 µF
C_{10}, C_{13} = 100 µF, 25 V, radial
C_{14} = see text
C_{16}–C_{23} = 100 pF, 100 V
C_{24} = 1 µF, metallized polypropylene (MKT)
C_{25} = 0.68 µF
C_{26}, C_{27}, C_{32}, C_{39} = 2.2 µF, 63 V, radial

Figure 12. The double-sided printed-circuit board is intended to be combined with the heat sink into a single entity. Before that can be done, however, the section for the output relay and the inductor must be cut off the main section.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(C) ELEKTOR

C_{28}, C_{34}, C_{35}, C_{41} = 470 µF, 100 V, radial
C_{29}, C_{33}, C_{36}, C_{40} = 0.22 µF, 100 V
C_{30}, C_{37} = 47 µF, 63 V, radial
C_{31}, C_{38} = 0.015 µF
C_{43}–C_{48} = 0.1 µF, 630 V

Inductors:
L_1 = see text

Semiconductors:
D_1, D_2 = LED, red, flat
D_3, D_{18}, D_{19} = 1N4148
D_4, D_6 = zener, 5.6 V, 500 mW
D_5, D_7 = zener, 15 V, 1.3 W
D_8, D_{11} = zener, 30 V, 1.3 W
D_9, D_{12} = zener, 39 V, 1.3 W
D_{10}, D_{13}, D_{16}, D_{17} = 1N4004
D_{14}, D_{15} = zener, 12 V, 500 mW
T_1, T_4, T_5, T_{15}–T_{17} = BC560C
T_2, T_3, T_6, T_{18}–T_{20} = BC550C
T_7, T_8, T_{43}, T_{48} = BF245A
T_9 = BF871
T_{10} = BF872
T_{11}, T_{50}, T_{51} = BC640
T_{12}, T_{45}, T_{46} = BC639
T_{13}, T_{14} = BF256C
T_{21}–T_{23} = MJE350
T_{24}–T_{26} = MJE340
T_{27} = BD139
T_{28} = BD140
T_{29}–T_{31} = 2SC5171 (Toshiba)
T_{32}–T_{34} = 2SA1930 (Toshiba)
T_{35}–T_{38} = 2SC5359 (Toshiba)
T_{39}–T_{42} = 2SA1987 (Toshiba)
T_{44}, T_{49} = BF256A
T_{47} = BD712
T_{52} = BD711

Integrated circuits:
IC_1 = OP90G
IC_2 = 6N136

Miscellaneous:
JP_1, JP_2 = 2.54 mm, 2-way pinstrip and pin jumper
K_1 = 3-way terminal block, pitch 5 mm
Re_1 = relay, 12 V, 600 Ω
Re_2–Re_4 = relay, 12 V, 16 A, 270 Ω
Heat sink for T_{21}–T_{26} = 38.1 mm, 11 K W^{-1} (Fischer Type SK104-STC; TO220)
Heat sink for drivers/output transistors, 150 mm, 0.25 K W^{-1}, Fischer Type SK157
Ceramic isolation washers for T_{21}–T_{34}: Fischer Type AOS220
Mica isolating washers for T_{35}–T_{42}
PCB Order no 990001-1 (see Readers Services towards end of this magazine)

is, of course, of paramount importance that these capacitances are as small as feasible. For this reason, it is vital that in the thermal coupling of T_{21}–T_{34} 1.5 mm thick ceramic—not mica—isolating washers are used. Mica washers may, however, be used with the output transistors since parasitic capacitances there are of no significance.

The component and track layouts of the mother board are shown in **Figure 12**. It will be seen that the board consists of two sections: the mother board proper and the output-relay board. The latter must be cut off before any other work is done. Later, when it is built up, it is mounted on the mother board with the aid of four 50 mm long metal spacers in such a way that the LS– and LS+ terminals on the two boards are above each other. The spacers also provide the electrical link between the boards.

The completed relay board is shown in **Figure 13**. Inductor L_1 is made from a doubled-up length of 1.5 mm enamelled copper wire wound in two layers of eight turns each around a 16 mm former (such as a piece of PVC pipe). After the coil has been wound, the PVC pipe is removed and the four windings connected in parallel. See **Figure 14**.

Ignoring the drivers and output transistors for the moment, the construction of the mother board is traditional. As always, great care must be taken during the soldering and placing of components. Do not forget the thermal coupling of T_1-T_3, T_2-T_4, D_1-T_5, D_2-T_6, T_{45}-T_{46}, and T_{50}-T_{51}, as already pointed out in Part 1. Also, T_{21}–T_{23} and T_{24}–T_{26} must be mounted on a heat sink, and isolated from it by means of a ceramic washer. When this is done, fit the composite heat sinks on the board, and link them to earth.

The input signal and the ±85 V supply lines are linked to the board via standard solder pins.

For connecting the ±70 V supply lines and the relay board, 3 mm screw holes are provided. Metal spacers are to be fixed to these and cable connectors to the top of the spacers.

MAIN HEAT SINK

When the mother board has been completed, and carefully checked, as far as described, it and the drivers and output transistors, T_{27}–T_{42}, must be mounted on the main heat sink. This is a 150 mm high Type SK157 from Fischer with a thermal resistance of 0.25 K W^{-1}. This is admittedly a very tedious job. It is vital that all requisite fixing holes are drilled accurately in the heat sink and preferably tapped with 3 mm thread. The template delivered with the ready-made board is almost indispensable for this work.

When the holes have been drilled (and, possibly, tapped) transistors T_{27} and T_{28} should be fitted first (this is important because they become inaccessible after the board has been fitted). They must be located as close as possible to the output transistors and not in the position indicated on the board. Again, the template makes all this clear. Their terminals must then be extended with the aid of short lengths of equipment wire, which are later fed through the relevant holes on the board and soldered to the board via, for instance, a three-way pin header.

The terminals of the drivers and output transistors must be bent at right angles: those of the former at the point where they become thinner and those of the latter about 5 mm from the body of the device. When this is done, screw all transistors loosely to the heat sink, not forgetting the isolating washers. If it is intended to use fan cooling, the requisite temperature sensor—that is, a Type BD140 transistor— should also be attached to the heat sink at this stage. The template does not show a location for the sensor, but it seems sensible to fit it at the centre close to T_{37} or T_{40}.

The next step is to fit all ten spacers to the heat sink: these should all be 10 mm long. In the prototype, spacers with a 3 mm screwthread at one end were used. Two of the spacers merely provide additional support for the relay board and another two form the electrical link between the negative supply line and the heat sink.

When all this work is done, the board should look more or less like that in **Figure 15**. Note that because of tests later on, there are, as yet, no ceramic isolating washers fitted on the prototype.

The next, and most tedious, step is to combine the board and heat sink. It is, of course, vital that all spacers are exactly opposite the relevant fixing holes and—even more tedious—that the terminals of all transistors are inserted into the correct mounting holes. Bear in mind that the metal

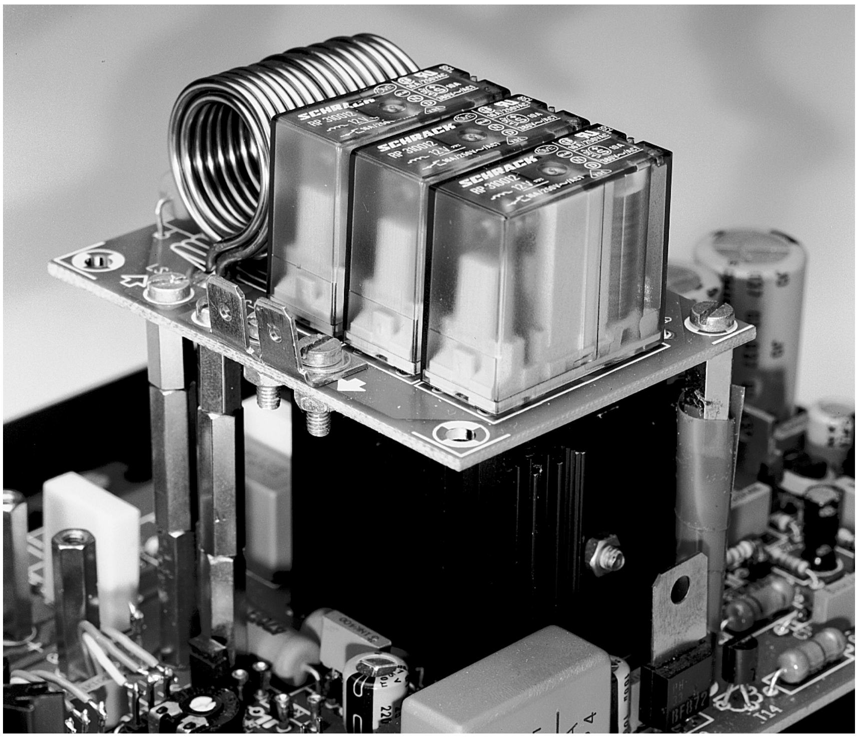


Figure 13. Illustrating how the relay board is mounted on the mother board with the aid of spacers.

spacers for linking –, +, LS+, and LS–, are already on the board. As the terminals of the output transistors are slightly longer than those of the drivers, it may be possible to do this work in two stages: output transistors first and drivers second. It may prove necessary to turn one or more of the transistors slightly, which is the reason that the fixing screws have not yet been tightened. When all terminals are correctly inserted, these screws must, of course, be tightened firmly.

The final step is to fix the relay board on the spacers that form the link for the LS– and LS+ terminals.

SETTING UP

Before the amplifier module can be taken into use, presets P_2–P_5 must be set as required. Preset P_1 is intended only for possibly adjusting the balance in case of a bridge configuration.

Start by turning P_3 (the quiescent-current control) fully anticlockwise and P_2, P_4, and P_5, to their centre position. Check the outputs of the power supply and auxiliary power supply and, if these are correct, link the +70 V line to pins '+' and '0', the –70 V line to '–' and '0', the +85 V line to '+ +' and the -85 V line to '--'. For absolute safety, link the ±70 V lines temporarily via a 10 Ω, 5 W resistor.

Next, set P_4 and P_5 for voltages of +78 V and –78 V respectively at the cases of transistors T_{47} and T_{52} respec-

Figure 14. Air-cored inductor L_1 is formed by laying two windings each of eight turns of doubled-up each on top of one another. The former is a length of 16 mm diameter PVC pipe as used by plumbers. The resulting four windings are simply connected in parallel.

tively (the cases of these transistors are linked to the output of the relevant regulator). It is important that the negative and positive voltages are numerically identical.

Since the parameters of the n-p-n and p-n-p transistors in the input stage are never exactly identical, there may be a slight imbalance. This may be corrected by adjusting the output of current source T_5 with the aid of preset P_2 to give a potential of exactly 0 V at the output (pin 6) of IC_1 (when 'cold').

Finally, insert an ammeter (set to 500 mA or 1 A range) in the +70 V or –70 V line, and adjust P_3 carefully for a quiescent current of 200 mA (cold condition—that is, immediately after switch-on). With a large drive signal, the quiescent current may increase to some 600 mA, but at nominal temperatures, its level will stabilize at 200–400 mA. Note that these fluctuations have no noticeable effect on the performance of the amplifier.

CHECK AND TEST

When the amplifier has been switched on for about half an hour, the voltages shown in Figure 2 (Part 1) may be verified. Note that voltage levels depending on the setting of current sources habitually show a substantial spread: 30 per cent is quite common. All measurements should be carried out with a good digital voltmeter or multimeter with a high-impedance input.

Other than the test voltages in the circuit diagram, there are some others that may be checked. For instance, the proper functioning of the output transistors may be ascertained by measuring the voltage across R_{45}–R_{52}. Hold one test probe against the loudspeaker terminal and with the other measure the potential at the emitters of all output transistors. The average value should be about 20 mV, but deviations of up to 50 per cent occur.

The voltage amplifier operation may be checked by measuring its current drain: if this is within specification, the voltage across R_{56} and R_{65} must be within 0.8–1.1 V (after the amplifier has been on for at least half an hour).

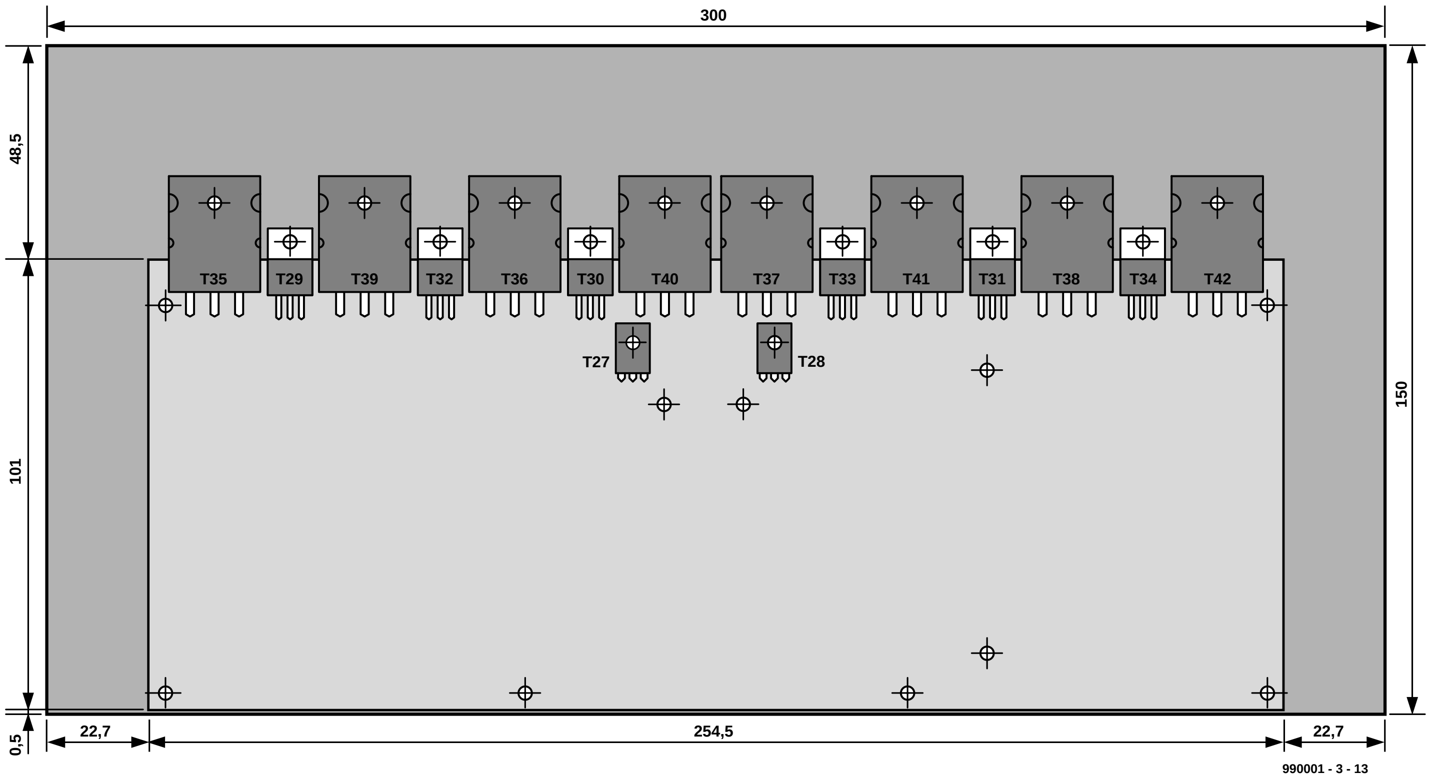
Finally, the potential drops across the emitter resistors of differential amplifiers T_{45}-T_{46} and T_{50}-T_{51} must not differ by more than a factor 2. Too large a factor is detrimental to the stable operation of the amplifiers. A too large difference may be corrected by changing the value of R_{62} or R_{71}, as the case may be. If this is unsuccessful, the relevant transistor pair will have to be replaced.

When all is well, the resistors in series with the ±70 V lines should be removed. Note that a rectified voltage of 70 V, let alone one of 140 V, is lethal. It is therefore absolutely essential to switch off the power supply and verify that the residual voltages have dropped to a safe value before doing any work on the amplifier.

Next month's instalment will deal with the wiring up of the amplifier and its performance, including specifications.

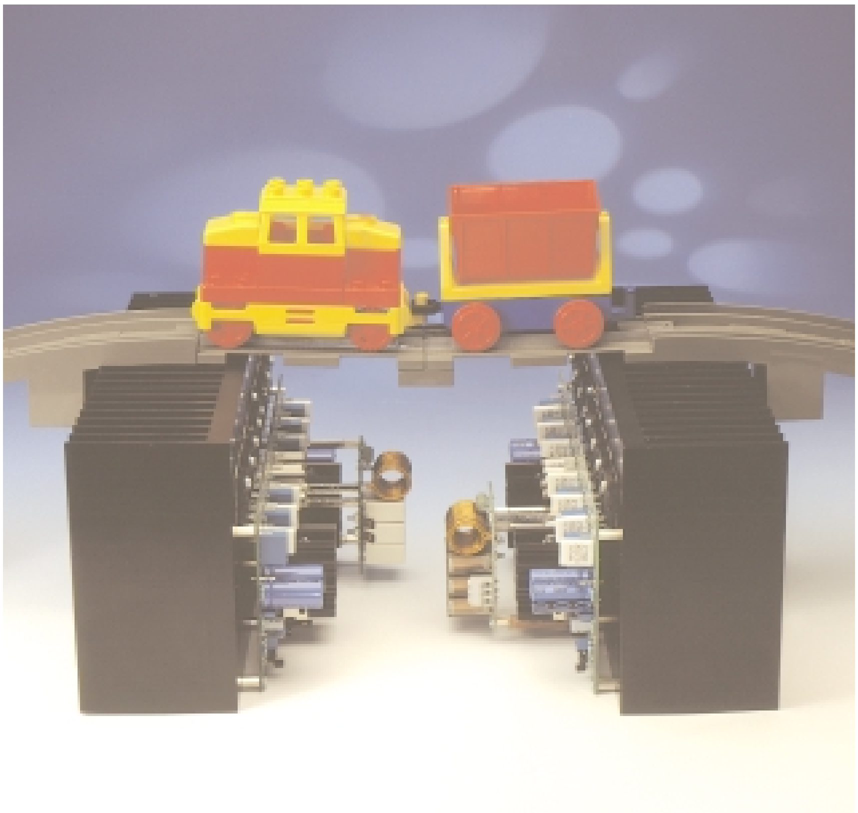
[990001-3]

Figure 15. The PCB is delivered with a template to ensure that the transistors are fitted at the correct location on the heat sink.



Titan 2000

Part 5: half-bridging two single amplifiers



In the introduction to Part 1 it was stated that the Titan 2000 could deliver up to 2000 watts of 'music power', a term for which there is no standard definition but which is still used in emerging markets. Moreover, without elaboration, this statement is rather misleading, since the reader will by now have realized that the single amplifier cannot possibly provide this power. That can be attained only when two single Titan amplifiers are linked in a half-bridge circuit. The true power, that is, the product of the r.m.s. voltage across the loudspeaker and the r.m.s current flowing into the loudspeaker, is then 1.6 kilowatts into a 4-ohm loudspeaker.

Design by T. Giesberts

BRIDGING: PROS AND CONS

Bridging, a technique that became fashionable in the 1950s, is a way of connecting two single output amplifiers (valve, transistor, BJT, MOSFET, push-pull, complementary) so that they together control the passage of an alternating current through the loudspeaker. This article describes what is strictly a half-bridge configuration, a term not often used in audio electronics. When audio engineers speak of bridge mode, they mean the full-bridge mode in which four amplifiers are used.

In early transistor audio power amplifiers, bridging was a means of achieving what in the 1960s were called public-address power levels as high as,

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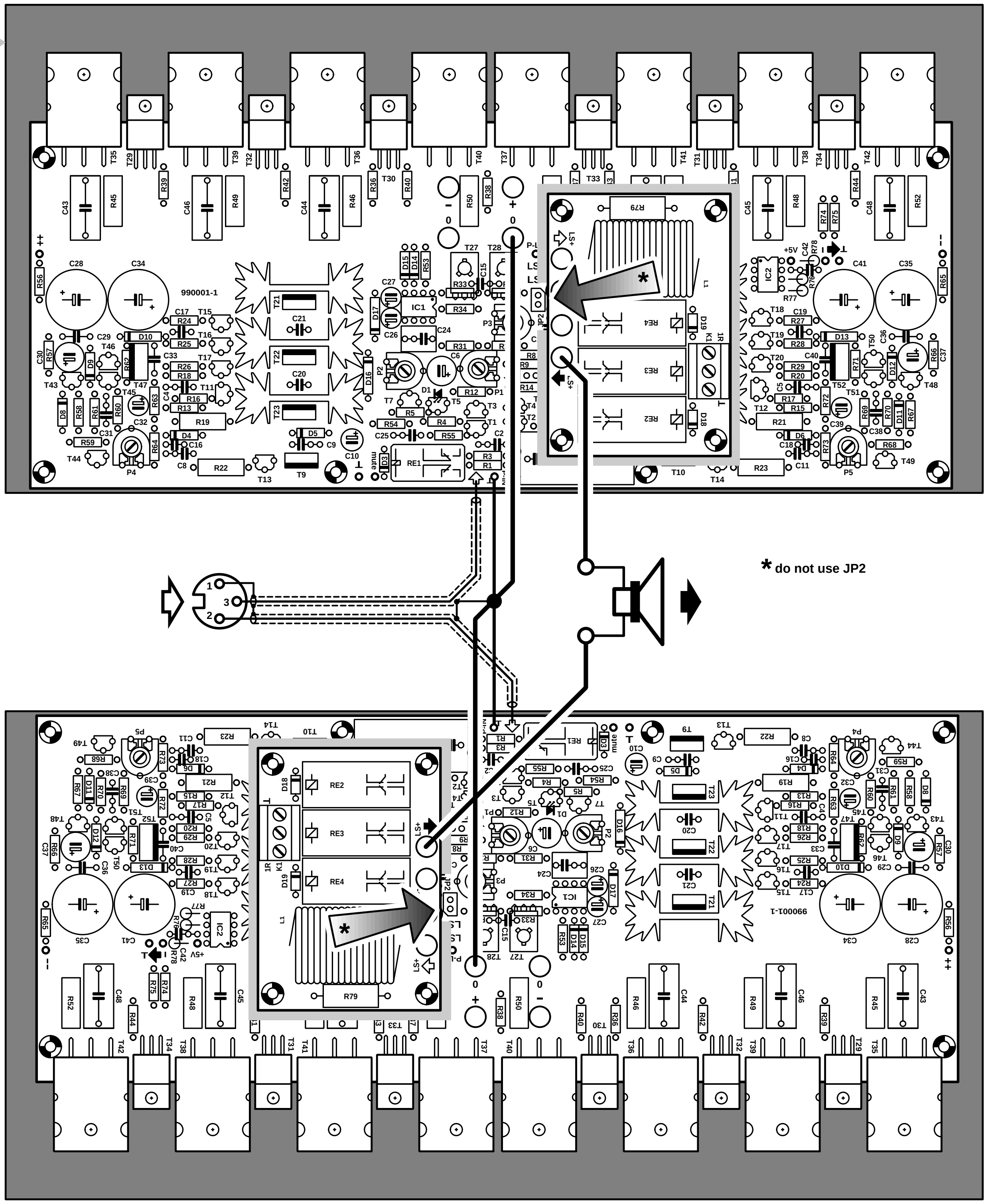


Figure 17. The interlinking required to form a half-bridge amplifier from two single Titan 2000 units. Note that the resulting balanced input may be reconverted to an unbalanced one with the Brangé design (Balanced/unbalanced converters for audio signals) published in the March 1998 issue of this magazine. The PCB for that design (Order no. 980026) is still stocked.

say 50–80 W into 8 Ω. Such power levels were then way beyond of what the voltage ratings of output transistors would permit.

Bridging is considered by many to be a good thing, since it automatically provides a balanced input (drive). However, opponents will quickly point out that it halves output damping, doubles the circuitry and virtually cancels even-order harmonics created in the amplifier.

Opponents also claim that bridging amplifiers is tedious and requires too much space. It is, however, not simple either to design a single amplifier with the same power output and the requisite power supply. A single 2 kW amplifier requires a symmetrical supply voltage of ±130 V, that is, a total of 260 V. The power supply for this would be quite a design. And where would a designer find the drivers and output transistors for this? Advocates point out that bridging amplifiers have the advantage of requiring a relatively low supply voltage for fairly high output powers.

Bridging just about doubles the rated output power of the single amplifier. Again, opponents point out that loudness does not only depend on

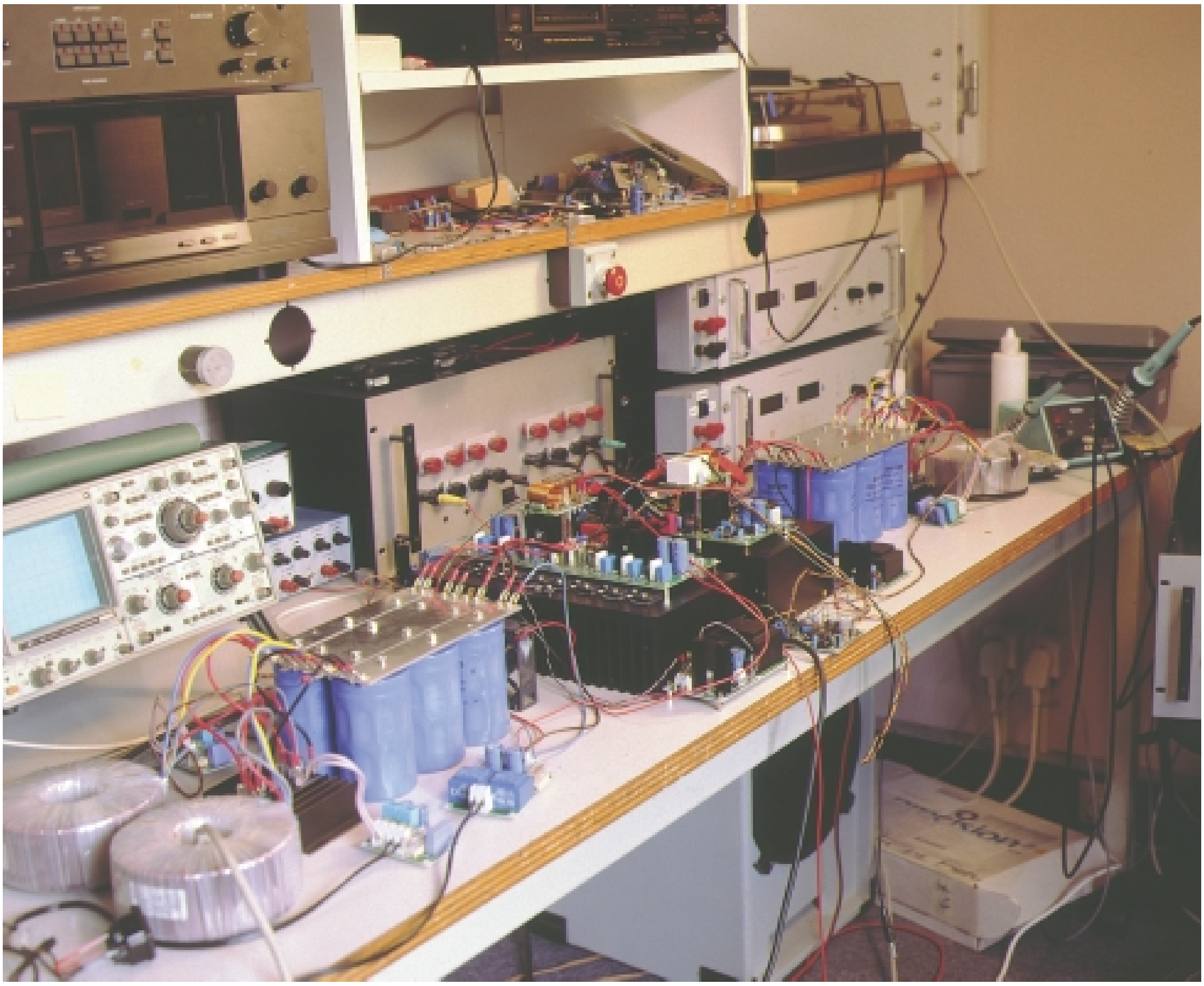


Figure 18. Test setup for the prototype half-bridge amplifier (centre). Note the large power supplies at the left and right of the amplifier.

the amplifier, but also on the loudspeaker. Bear in mind, they say, that just changing a loudspeaker with a sensitivity of, say, 90 dB_{SPL} per watt per metre to one with a sensitivity of 93 dB_{SPL} per watt per metre is equal to doubling the amplifier power rating.

Clearly, bridging two amplifiers is a mixture of good and bad audio engineering and sonics.

INTERCONNECTING

It is, of course, necessary that two completed single Titan 2000 amplifiers are available, each with its own power supply. It should then be possible to simply interlink the earths of the two units, use the inputs as a common balanced input, and connect the loudspeaker between terminals LS+ on the two amplifier. However, a few matters must be seen to first.

Owing to the requisite stability, it is imperative that the two amplifiers are juxtaposed with the space between them not exceeding 5 cm (2 in). They should, of course, be housed in a common enclosure.

The interwiring is shown in **Fig-ure 17**. Make sure that the power supplies are switched off and that the smoothing capacitors have been discharged before any work is carried out.

Start by interlinking the negative supply lines (terminals 0) with insulated 40/02 mm wire. Remove the insulation at the centre of the length of wire since this will become the central earthing point for the new (balanced) input. Link the ⊥ terminals on both boards to the new central earth with 24/02 mm insulated wire.

Connect the loudspeaker terminals to the LS+ terminals on the two boards with 40/02 mm insulated wire.

Link pins 2 and 3 of the XLR connector to the input terminals on the boards with two-core screened cable. Solder the screening braid to pin 1 of the XLR connector and to the new central earthing point.

Finally, on both boards remove jumper JP_2 from the relevant pin strip.

FINALLY

When all interconnections between the boards as outlined have been made, the single amplifiers form a half-bridge amplifier. If all work has been carried out as described, there should be no problems.

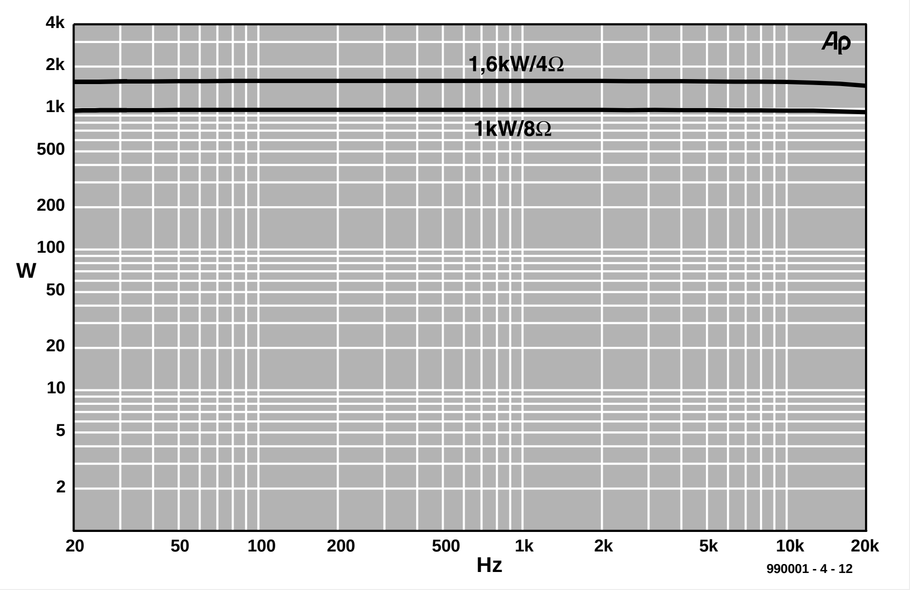
In the design stages, network R_9-P_1, inserted into the circuit with pin jumper JP_1 (see Part 1), was considered necessary for common-mode suppression. However, during the testing of the prototype, the network was found to be superfluous. It may be retained if the half-bridge amplifier is to be used with a second half-bridge amplifier for stereo purposes, when it may be used to equalize the amplifications of the two half-bridge amplifiers.

[990001]

Parameters

With a supply voltage of ±70 V (quiescent ±72 V) and a quiescent current of 200–400 mA

Input sensitivity	2.1 V r.m.s.
Input impedance	87 kΩ
True power output for 0.1% THD	950 W into 8 Ω; 1.5 kW into 4 Ω
True power output for 1% THD	1 kW into 8 Ω; 1.6 kW into 4Ω
Power bandwidth	1.5 Hz – 220 kHz
Slew limiting	170 V μs^{-1}
Signal+noise-to-noise ratio (at 1 W into 8 Ω)	97 dB (A-weighted
	93 dB (B=22 kHz)
Total harmonic distortion (B=80 kHz)	
at 1 kHz	0.0033% (1 W into 8 Ω)
	0.002% (700 W into 8 Ω)
	0.0047% (1 W into 4 Ω)
	0.006% (700 W into 4 Ω)
at 20 kHz	0.015% (700 W into 8 Ω)
	0.038% (1200 W into 4 Ω)
Intermodulation distortion	
(50 Hz:7 kHz = 4:1)	0.0025% (1 W into 8 Ω)
	0.0095% (500 W into 8 Ω)
	0.004% (1 W into 4 Ω)
	0.017% (500 W into 4 Ω)
Dynamic intermodulation distortion	
(square wave of 3.15 kHz and	0.0038% (1 W into 8 Ω)
sine wave of 15 kHz)	0.0043% (700 W into 8 Ω)
	0.005% (1 W into 4 Ω)
	0.0076% (1200 W into 4 Ω)
Damping (with 8 Ω load)	≥ 350 (at 1 kHz)
	≥ 150 (at 20 kHz)
Open-loop amplification	×8600
Open-loop bandwidth	53 kHz
Open-loop output impedance	3.2 Ω

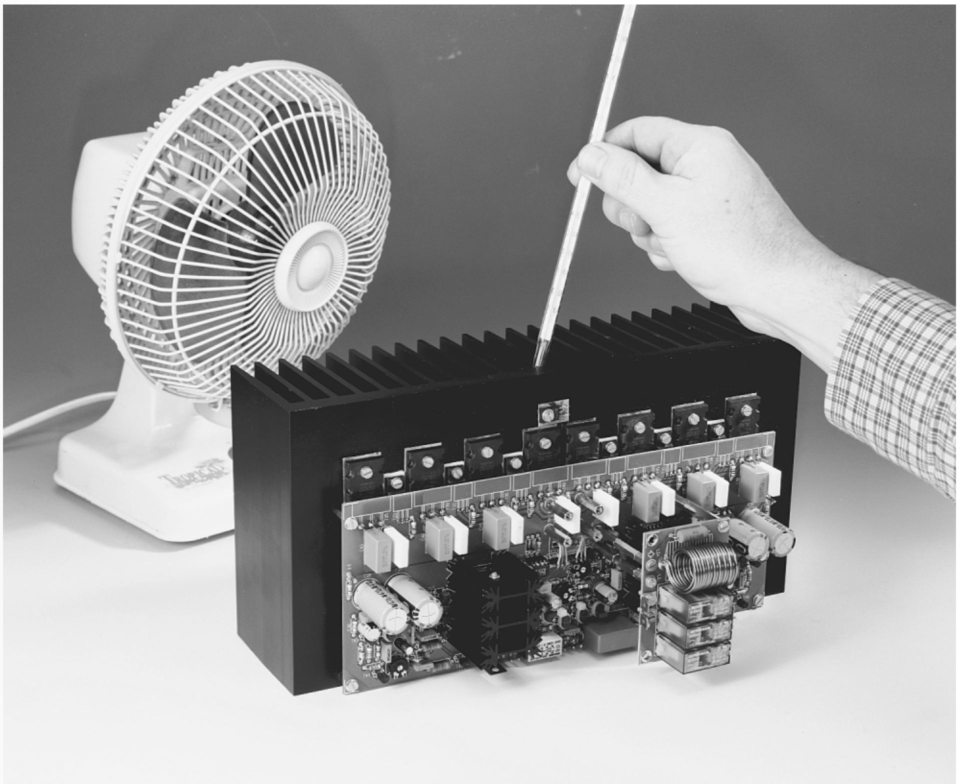


A comparison of these parameters with the specifications given in Part 4 ((May 1999 issue) show that they are generally in line. In fact, the intermodulation distortion figures are slightly better. Because of this, no new curves are given here other than power output (1 kW into 8 Ω and 1.6 kW into 4 Ω) vs frequency characteristics for 1 per cent total harmonic distortion.

During listening tests, it was not possible to judge the half-bridge amplifier at full volume, simply because there were no loudspeakers available that can handle this power output. However, up to 200 W true power output, the half-bridge amplifier sounds exactly the same as the single amplifier. Instrument test figures show no reason to think that the performance at higher output powers will be degraded.

fan control

for high-power output amplifiers



Most domestic amplifiers are cooled solely by convection, but it may happen that a standard heat exchanger does not provide adequate cooling for a certain output amplifier. In such a case other means must be used, normally a fan or fans. The circuit described in this article provides a temperature-dependent, proportional control for such a cooling fan. It enables the turn-on threshold and the control characteristic to be set as needed within a given range. It also provides over-temperature indication which may be used to actuate the protection circuit of the amplifier.

Design by T. Giesberts

INTRODUCTION

Although the heat sink of the Titan 2000 is of reasonable dimensions, the amplifier, like other output amplifiers, has its limits as regards dissipation. With loads down to 4 Ω the heat sink is perfectly capable of dissipating the generated heat, but with lower loads and maximum drive levels, forced cooling must be used.

The design of the present circuit ensures that when a predetermined heat sink temperature is exceeded, two small fans are switched on which rotate at a speed that is directly proportional to the heat sink temperature. When maximum fan speed is reached, the over-temperature indication is enabled and the protection circuit of the ampli-

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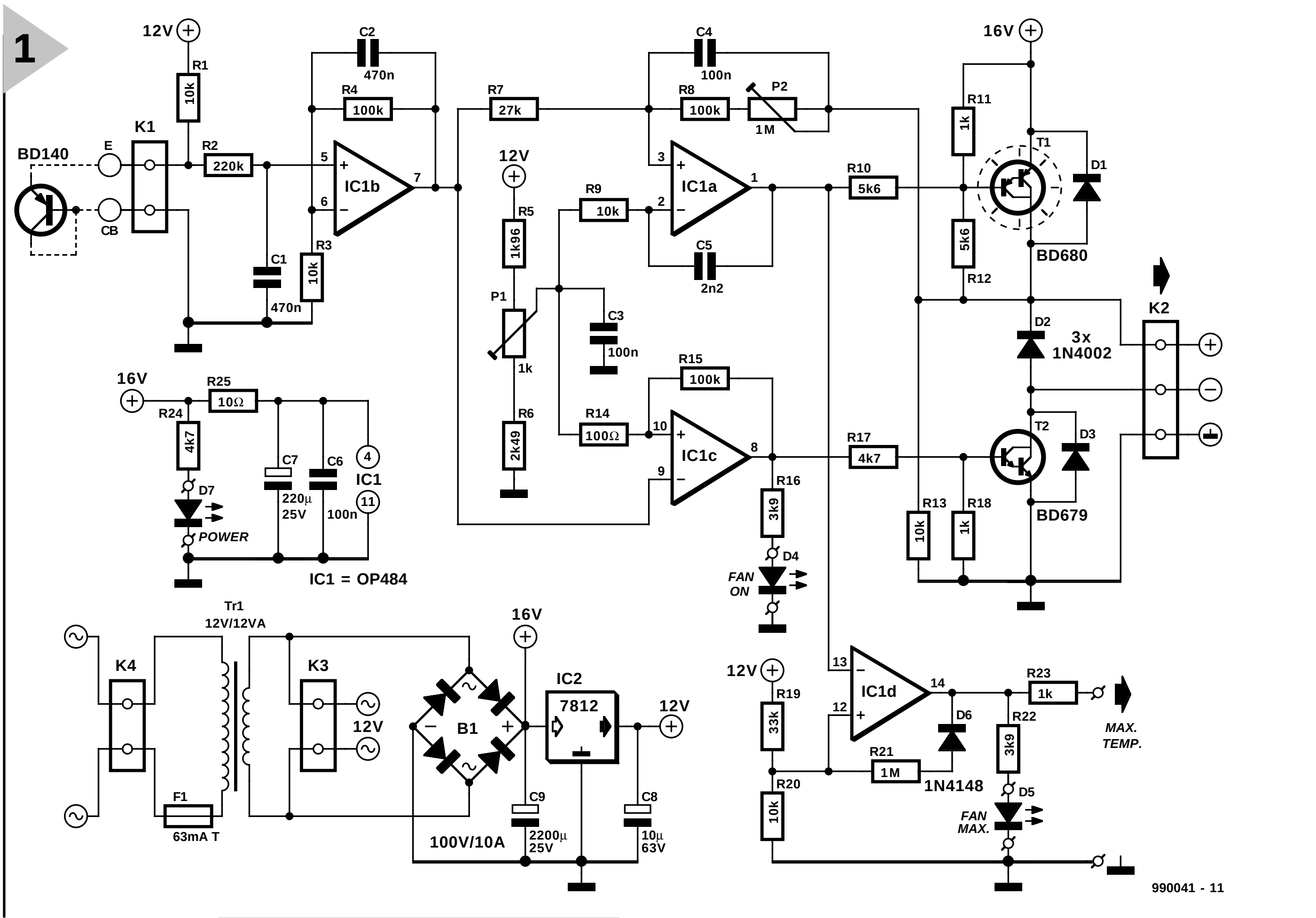

Figure 1. Two amplifiers, a pair of comparators and a couple of light-emitting diodes (LEDs) provide the desired control characteristic and indications.

fier may be actuated via a separate output.

There are two indicator LEDs: one for the turn-on threshold (when the fans are turned on), and the other for maximum temperature. The fans in the prototype are small 12 V models. The rotational speed of these can be varied readily with the supply voltage: they start reliably from voltages as low as 5 V. It is of great importance that the fans are very quiet types to avoid intrusive fan noise in the listening room during low programme levels.

The temperature sensor is a standard Type BD140 transistor mounted on the heat sink. Its forward voltage (like that of any silicon transistor or diode) decreases by exactly 2 mV °C^{-1}, which makes it eminently suitable for use as a sensor.

DESIGN

The circuit diagram of the control is shown in **Figure 1**.The over-temperature signal appears on pins 'max temp', the fans are linked to connector K_2, and the sensor is connected to K_1.

The sensor, which is arranged as a diode, is biased via resistor R_1. The temperature-dependent potential across the p-n junction is filtered by network R_2-C_1 and then amplified ×11 by operational amplifier IC_{1b}. To make sure that there is adequate control voltage for the fans, amplifier IC_{1a} provides additional gain. This op amp is arranged a an inverting amplifier to ensure that the control voltage is directly proportional to the heat sink temperature. Transistor T_1 provides the level of current needed by the fans.

Preset P_2 in the feedback loop sets the gain of IC_{1a} and thus the degree to which the control voltage follows the temperature-dependent sensor output.

The fans are switched on via IC_{1c}, which compares the output signal of IC_{1b} with the reference potential derived from network R_5-P_1-R_6. At a given threshold, set with P_1, IC_{1c} changes state, whereupon the –ve pin of K_2 is linked to earth, resulting in the fans being switched on. Diode D_4 lights to indicate this happening. The threshold set with P_1 is also used as a reference for IC_{1a} to ensure that the initial control voltage is equal to the switching threshold.

Circuit IC_{1d} provides the over-temperature indication signal. This comparator guards the output voltage of amplifier IC_{1a} and changes state when the fans are working at full speed. Diode D_5 lights and a high level is passed to the 'max temp' output via

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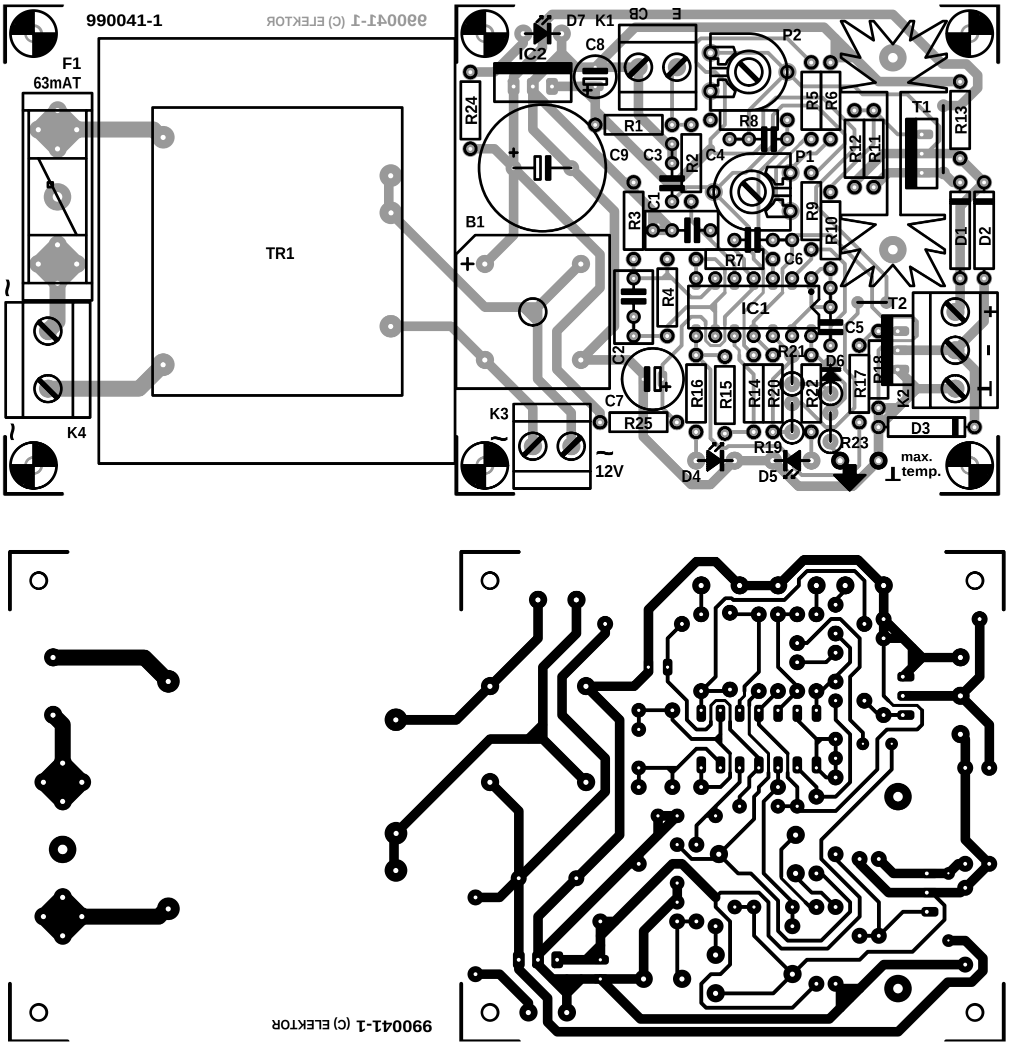
990041-1
990041-1 (C) ELEKTOR
F1
63mAT
TR1
K4
K3
12V
B1
IC1
IC2
T1
T2
P1
P2
K1
K2
D1
D2
D3
D4
D5
D6
D7
C1
C2
C3
C4
C5
C6
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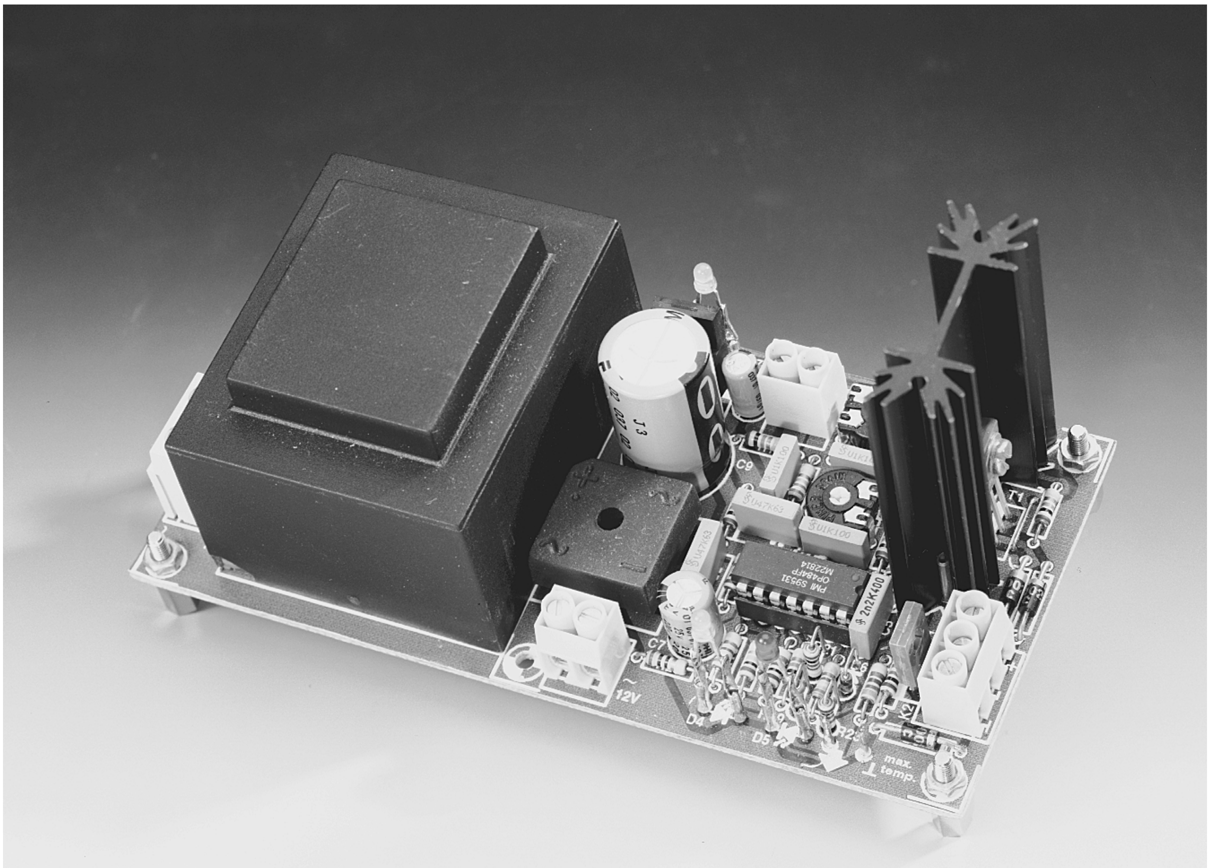

Parts list

Resistors:
R_1, R_3, R_9, R_{13}, R_{20} = 10 kΩ
R_2 = 220 kΩ
R_4, R_8, R_{15} = 100 kΩ
R_5 = 1.96 kΩ
R_6 = 2.49 kΩ
R_7 = 27 kΩ
R_{10}, R_{12} = 5.6 kΩ
R_{11}, R_{18}, R_{23} = 1 kΩ
R_{14} = 100 Ω
R_{16}, R_{22} = 3.9 kΩ
R_{17}, R_{24} = 4.7 kΩ
R_{19} = 33 kΩ
R_{21} = 1 MΩ
R_{25} = 10 Ω
P_1 = 1 kΩ preset
P_2 = 1 MΩ preset

Capacitors:
C_1, C_2 = 0.47 µF
C_3, C_4, C_6 = 0.1 µF
C_5 = 0.0022 µF
C_7 = 220 µF, 25 V, radial
C_8 = 10 ≤ F, 63 C, radial
C_9 = 2200 µF, 25 V, radial

Semiconductors:
D_1–D_3 = 1N4002
D_4, D_5, D_7 = high-efficiency LED, yellow red, green respectively
D_6 = 1N4148
T_1 = BD680
T_2 = BD679

Integrated circuits:
IC_1 = OP484FP (Analog Devices)
IC_2 = 7812

Miscellaneous:
K_1, K_3 = 2-way terminal strip for board mounting, pitch 5 mm
K_2 = 3-way terminal strip for board mounting, pitch 5 mm
K_4 = 2-way terminal strip for board mounting, pitch 7.5 mm
B_1 = rectifier, 100 V, 10 A, horizontal, 19×19 mm
Tr_1 = mains transformer, 12 V, 12 VA (see text for external transformer or mains adaptor)
F_1 = slow fuse 63 mA, complete with holder for board mounting
Heat sink (T_1), e.g., Fischer SK104/50
Two off 12 V fans (see text)
PCB Order no. 990041 (see Readers Services towards the end of this issue)

Figure 2. The design of the board allows the transformer section to be cut off.

resistor R_{23}. This high level may be communicated to the protection board in the Titan 2000 for use as a switching signal. In other amplifiers, it may be used via a buffer to, say, energize an output relay.

The supply for the circuit has been kept as simple as possible. The fans are driven by the rectified secondary voltage (16 V) of the transformer. The 16 V line is additionally decoupled by network R_{25}-C_6-C_7 for use as a supply for IC_1. The +12 V supply needed for the various references and setup networks is obtained by applying the 16 V voltage to IC_2. Diode D_7 functions as on/off indicator.

Figure 3. Photograph of the completed prototype control.

SETTING UP

The turn-on threshold is set with P_1 and the range of rotational speed of the fans with P_2. These settings are empirical and to individual requirements, for which an accurate thermometer is required.

Two extreme settings are possible. The first is with the turn-on threshold fairly high (60–65 °C) and maximum gain. The maximum rotational speed will then occur at a temperature only 6–7 °C higher than the turn-on threshold. The second is to set the turn-on threshold fairly low, say, 50 °C and the amplification low. This provides the largest possible control range, but the fans cannot reach maximum speed.

It is clear that the most practical setting is somewhere between these extremes. It is advisable to use a maximum heat sink temperature of 70 °C, in which case D_5 should light.

CONSTRUCTION

The control is best built on the printed-circuit board shown in **Figure 2**.

Since it may happen that the specified transformer cannot be obtained, the board may also be used with an external transformer or mains adaptor. The secondary of the transformer or the output of the mains adaptor is linked to the board via a two-way terminal strip for board mounting (K_3). If required, the transformer section of the board may then be cut off.

Since the dissipation of T_1 may rise to 6 W, the component must be mounted on a heat sink. If a Fischer Type SK104 is used, the supporting pins of this may be soldered to the board. An isolating washer is not strictly needed, but, since the collector of the transistor is linked to the case, care should be taken to ensure that the heat exchanger does not touch other components.

Figure 3 shows the completed prototype.

The sensor is connected to K_1 via two twisted lengths of thin, stranded, insulated circuit wire. Also, the output “max temp’ is linked to input ‘temp’ on the protection board via two twisted lengths of similar wire.

The light-emitting diodes (LEDs) are, of course, not placed on the board, but mounted on the front panel of the amplifier. They, too, are linked to the relevant positions on the board by twisted lengths of thin, insulated, stranded circuit wire.

The fans are connected in parallel to K_2.

When the control is used with the Titan 2000, all connections are shown on the wiring diagram of that amplifier (Part 4, Figure 16 elsewhere in this issue).

When the control is used with a public-address amplifier (PA), the fans may be connected to ‘+’ and ‘⊥’ on K_2 instead of to ‘+’ and ‘–’. If this is done, the fans are on continuously, but still at a speed that depends on the heat sink temperature.

The peak output current of the circuit is 1 A, which is more than adequate since most fans need only 200–250 mA.

The peak voltage available at K_2 is equal to the rectified transformer voltage less the knee voltages of T_1 and T_2. In practice, this is about 13 V.

[990041]

DIY: from vinyl to compact disk

with a PC and sound card

Nowadays, with the availability of personal computers and compact-disk (CD) writers, there is nothing in the way of transferring one's collection of vinyl records on to compact disks. All that may be needed in addition to the equipment already mentioned is a suitable preamplifier, such as the one presented in this article.

Brief parameters

Input sensitivity	
(moving-coil)	*2 mV*
(dynamic)	*about 0.2 mV*
Nominal output signal	*200 mV*
Signal-to-noise ratio	
(moving-coil)	*78 dBa (750 Ω in)*
	88 dBA (input short-circuited)
(dynamic)	*70 dBA (25 Ω in)*
	71 dBa (input short-circuited)

Design by T. Giesberts

INTRODUCTION

The DIY making of compact disks is rapidly becoming a commonplace. One of the applications that is particularly attractive to many people is the digitizing of their valuable collection of vinyl records. There are, of course, other advantages than creating space (CDs take much less storage space than vinyl records): a compact disk has a longer life than a vinyl record (although it is not, as some people believe, infinite), and it becomes possible to select and shuffle sections of the recording if and as desired.

When a personal computer is available that incorporates a CD recorder (many modern ones are) and a good-quality sound card, the copying of vinyl records is straightforward. All that is then required is a means of linking the pickup output to the sound card. When the record player is placed next to the computer, the line outputs of the amplifier may be used. When this is not possible, there are a few difficulties. The output voltage of a dynamic pickup is about 3 mV and that of a moving-coil type around 0.3 mV. Clearly, these potentials are insufficient to drive the line input of the sound card. Moreover, the frequency response of the signal must be corrected.

RIAA CORRECTION

A vinyl record is cut tangentially, that is, the cutter traverses the disk in a straight line from disk edge to centre. The cutter response is called constant velocity, which means that its velocity is the same for all frequencies. Therefore, the amplitude increases as the frequency drops (at a rate of 6 dB/octave). It would thus be 16 times greater at 30 Hz than at 15 kHz.

Large low-frequency stylus excursions during playback are avoided by cutting the bass and boosting the treble frequencies to improve the signal to noise ratio. These contours roll off at either side of a short flat region centred on 1 kHz to form the RIAA (Recording Industry Association of America) characteristic. The playback amplifier or preamplifier has a frequency response that is a mirror image of the RIAA characteristic (see Figure 1).

DESIGN

The design of the preamplifier allows the output of dynamic as well as moving-coil pickups to be connected to its input.

Although the preamplifier is intended primarily for use as a converter between record player and personal computer, it is equally suitable for use with a hi-fi amplifier that has no integral phone input.

The block schematic of the preamplifier is shown in **Figure 2**. Each of the two stereo channels is linked to an input amplifier for moving coil pickups, which can be taken out of circuit by a wire bridge, followed by a standard amplifier for dynamic elements. The RIAA frequency-correction circuit is incorporated in this latter amplifier.

Note that for cases where the record player is linked to the computer for long periods, the line input is retained for other applications. To avoid the cumbersome changing of plug-and-socket connections, there is a change-over relay at the output, which ensures that when the moving-coil preamplifier is not used, the line input is connected to the relevant terminal(s) on the computer.

The power supply provides the ±15 V lines for the operational amplifiers, as well as the single +20 V line for the relay. It is preceded by a filter to eliminate any mains hum and interference.

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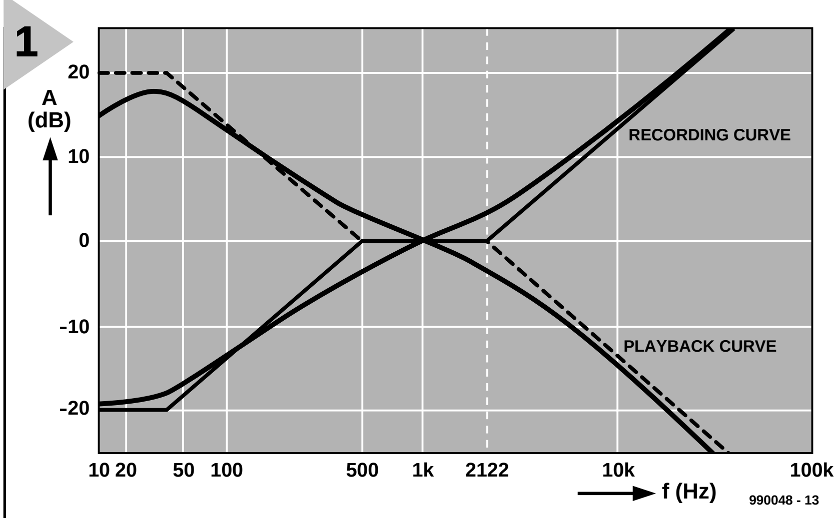

Figure 1. RIAA recording and playback characteristics.

Figure 2. Block schematic of the preamplifier.

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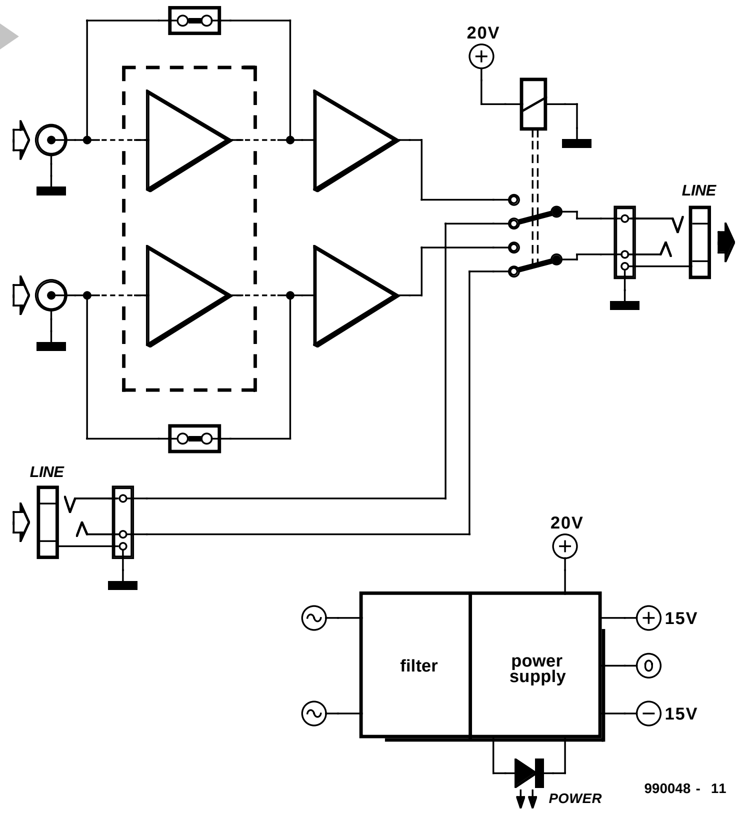

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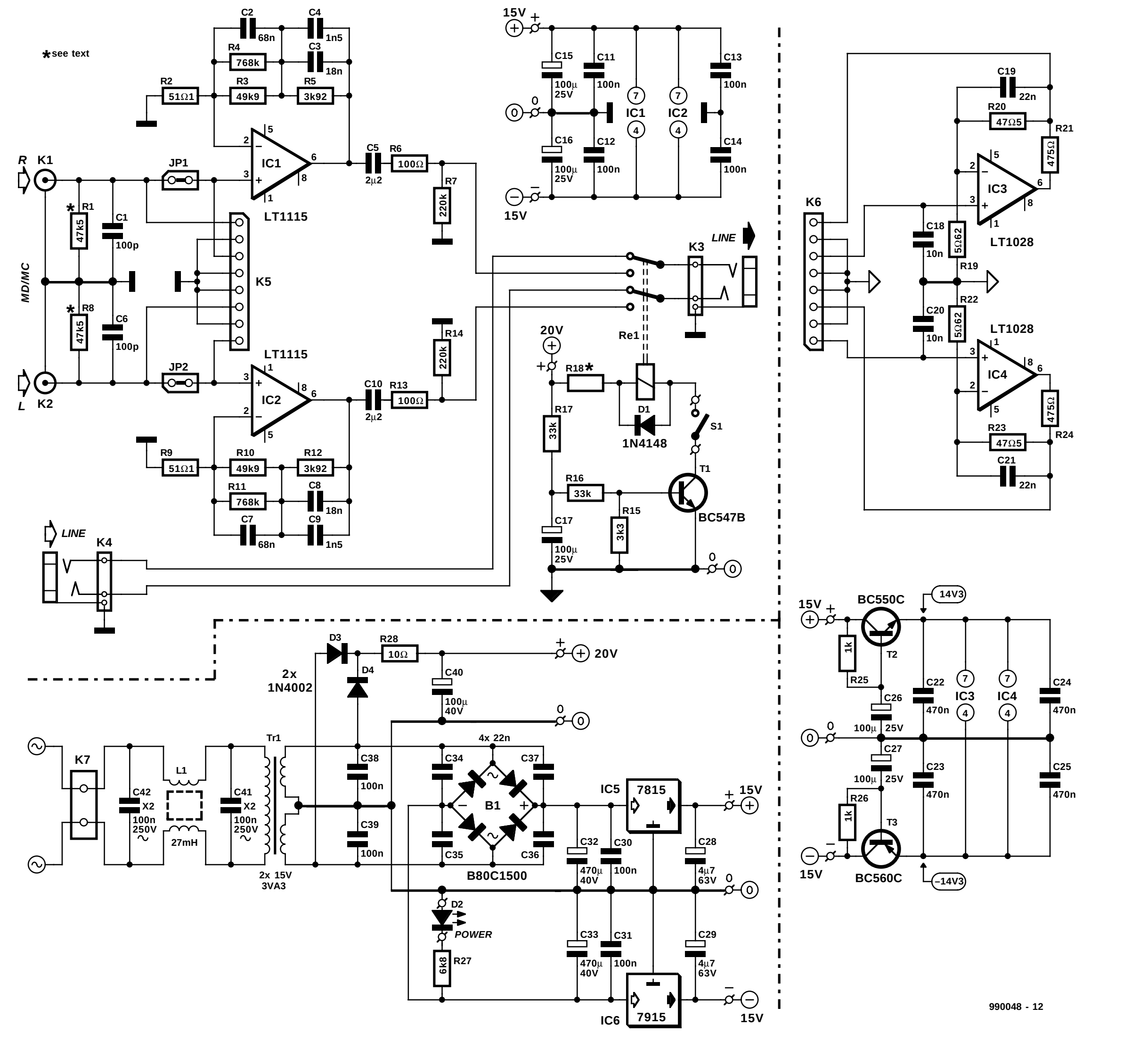

Figure 3. The circuit diagram may conveniently be split into three sections.

CIRCUIT DESCRIPTION

In **Figure 3**, the preamplifier for moving-coil elements is formed by IC_3 and IC_4, while the main amplifier is based on IC_1 and IC_2.

When the output of a dynamic pickup is linked to terminals K_1 and K_2, the input impedance has the standard value of 47 kΩ – determined almost exclusively by R_1 and R_8. Capacitors C_1 and C_6 determine the frequency response between 10 kHz and 20 kHz, which means that their value depends to some extent on the type of dynamic element used.

Operational amplifiers IC_1 and IC_2 are typified by a very low noise figure, a reasonably low bias current, and low input offset. When the output is 200 mV and the input is short-circuited, the amplifiers have a signal-to-noise ratio of 88 dB. In practical use, the noise of the amplifier is produced primarily by the pickup element. Note that the resistance and inductance of an average dynamic element are about 750 Ω and 450 mH respectively.

The gain of IC_1 and IC_2 is 40 dB at 1 kHz. The RIAA correction network is included in the negative-feedback loop between pins 2 and 6. Capacitors C_5 and C_{10} decouple any offset, while resistors R_6 and R_{13} protect the operational amplifiers against capacitive loads. Resistors R_7 and R_{14} ensure that C_5 and C_{10} are charged in the absence of a load, which helps to prevent switch-on phenomena.

When the power is switched on, relay Re_1 is energized, whereupon the output of the amplifiers is linked to terminal K_3. When the supply is switched off, the relay is disabled, whereupon the additional line input at terminal K_4 is linked to K_3.

To avoid switch-on clicks and plops, the relay is energized with some delay provided by capacitor C_{17} via transistor T_1. Resistor R_{15} ensures that the relay is deenergized rapidly to guarantee that the supply to the amplifiers is switched off instantly.

Switch S_1 serves to enable manual switching of the relay between amplifiers and line input terminal; K_4 without the need of switching off the supply.

When a dynamic pickup element is used, jumper terminals JP_1 and JP_2 are closed. The sections based on IC_3 and IC_4 are then not used and need not be built.

When a moving-coil pickup element is to be used, JP_1 and JP_2 must remain open and resistors R_1 and R_8 must be replaced by 100 Ω types. Amplifiers IC_3 and IC_4 are included in the signal path via terminals K_5 and K_6. These amplifiers provide an amplification of about ×10.

To ensure a low noise figure, the values of R_{19} and R_{22} are very low. To prevent this forming too large a load for the op amps, an additional resistor is used in the negative-feedback loop (R_{21} and R_{24} respectively). The resulting narrowing of the bandwidth is negated to a large extent by the use of very fast operational amplifiers.

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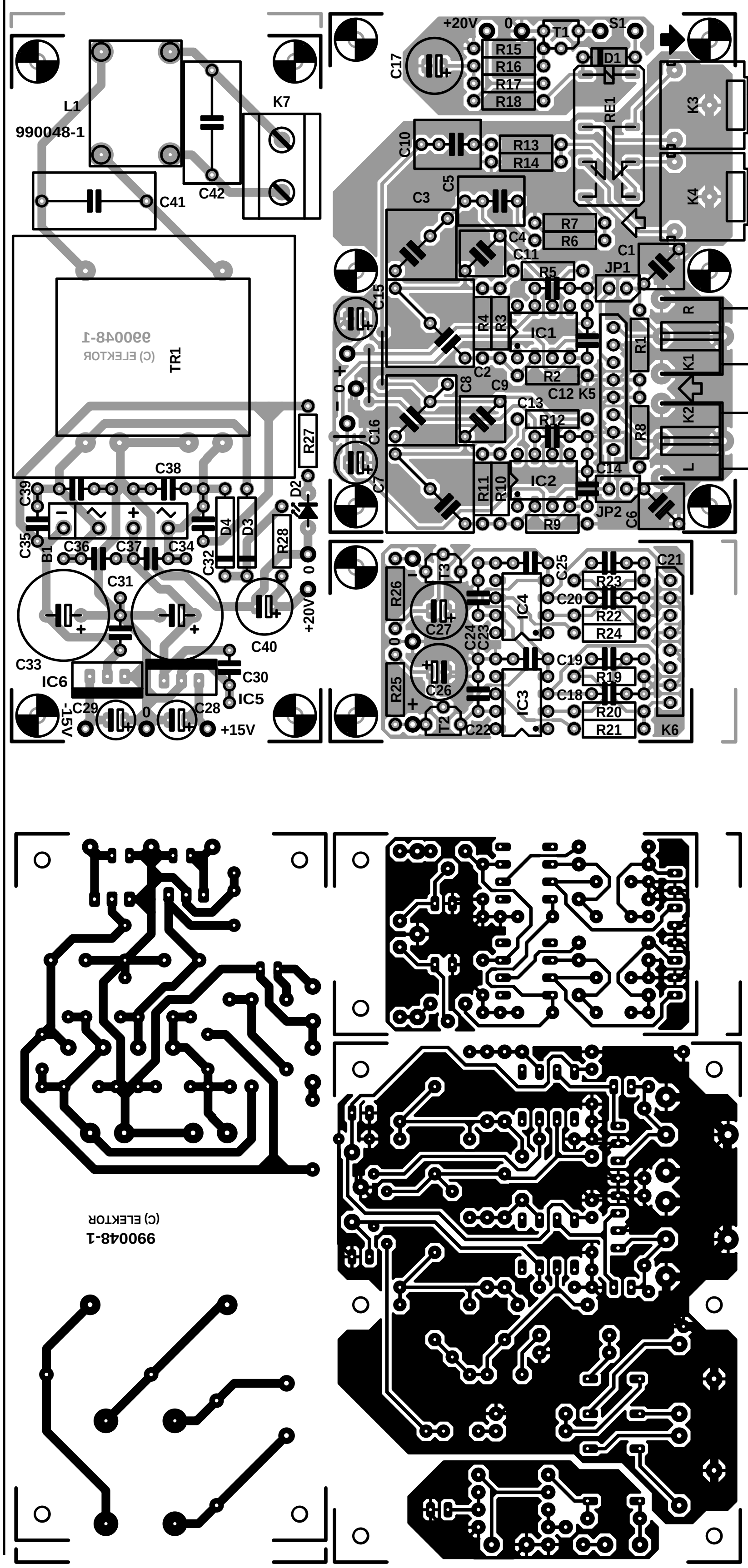


Figure 4. The printed-circuit board for the preamplifier is intended to be cut into two or three sub-boards.

Parts list

Resistors:
R_1, R_8 = 47.5 kΩ or 100 Ω (see text)
R_2, R_9 = 51.1 Ω
R_3, R_{10} = 49.9 kΩ
R_4, R_{11} = 768 kΩ
R_5, R_{12} = 3.92 kΩ
R_6, R_{13} = 100 Ω
R_7, R_{14} = 220 kΩ
R_{15} = 3.3 kΩ
R_{16}, R_{17} = 33 kΩ
R_{18} = see text
R_{19}, R_{22} = 5.62 Ω
R_{20}, R_{23} = 47.5 Ω
R_{21}, R_{24} = 475 Ω
R_{25}, R_{26} = 1 kΩ
R_{27} = 6.8 kΩ
R_{28} = 10 Ω

Capacitors:
C_1, C_6 = 100 pF, 63 V, 1%
C_2, C_7 = 0.068 µF, 63 V, 1%
C_3, C_8 = 0.018 µF, 63 V, 1%
C_4, C_9 = 0.0015 µF, 63 V, 1%
C_5, C_{10} = 2.2 µF, metallized polyester, pitch 5 mm or 7.5 mm
C_{11}–C_{14}, C_{30}, C_{31}, C_{36}, C_{38} = 0.1 µF
C_{15}–C_{17}, C_{26}, C_{27} = 100 µF, 25 V, radial
C_{18}, C_{20} = 0.01 µF
C_{19}, C_{21} = 0.022 µF
C_{22}–C_{25} = 0.47 µF
C_{28}, C_{29} = 4.7 µF, 63 V, radial
C_{32}, C_{35} = 470 µF, 40 V, radial
C_{34}–C_{37} = 0.022 µF, ceramic
C_{40} = 100 µF, 40 V, radial
C_{41}, C_{42} = 0.1 µF, 250 VAC, Class X_2

Inductors:
L_1 = 2×27 mH, 400 mA, 250 VAC

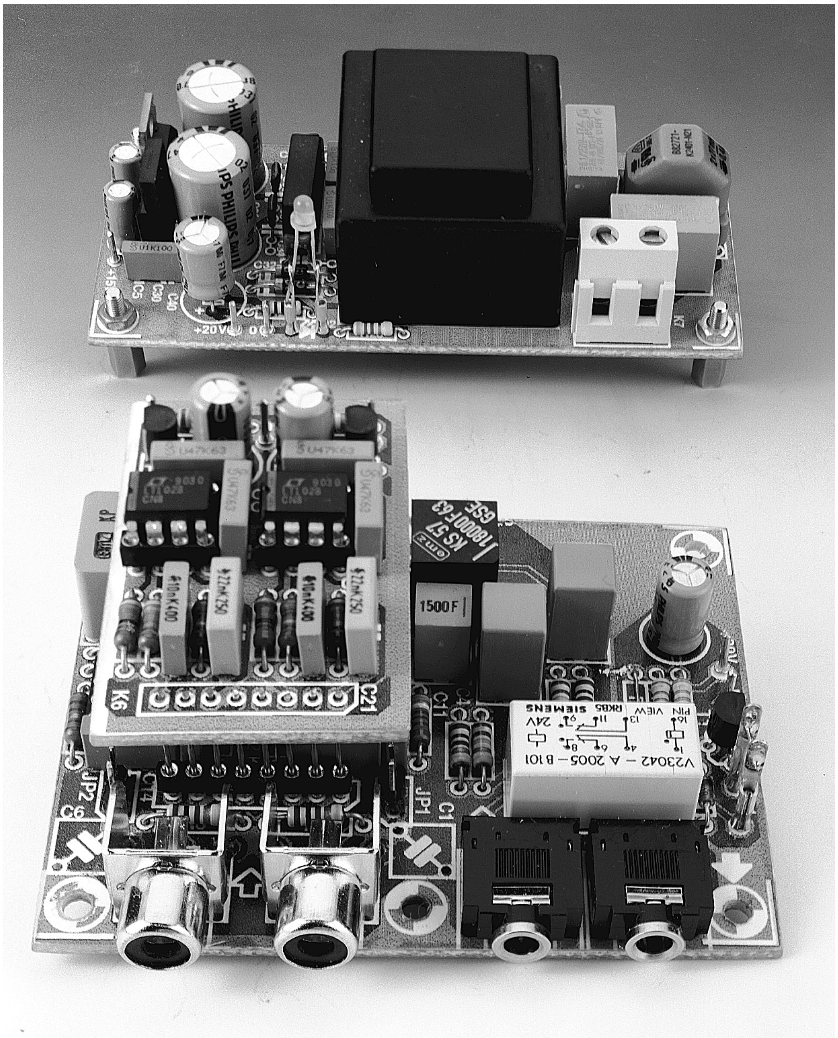
Semiconductors:
D_1 = 1N4148
D_2 = LED, green, high efficiency
D_3, D_4 = 1N4002
B_1 = B80C1500 (straight)
T_1 = BC547B
T_2 = BC550C
T_3 = BC560C

Integrated circuits:
IC_1, IC_2 = LT1115CN6 (Linear Technology)
IC_3, IC_4 = LT1028CN8 (Linear Technology)
IC_5 = 7815
IC_6 = 7915

Miscellaneous:
JP_1, JP_2 = 2-way pin header and jumper
K_1, K_2 = audio socket for board mounting
K_3, K_4 = 3.5 mm PCB mounting audio socket
K_5, K_6 = 8-way SIL (see text)
K_7 = 2-way terminal block for PCB mounting, pitch 7.5 mm
S_1 = single-pole, single-throw switch
Re_1 = 24 V relay, 2.2 kΩ
Tr_1 = mains transformer, 2×15 V secondary, 3.3 VA

5

Figure 5. The completed prototype preamplifier for use with dynamic and/or moving-coil pickup elements.



Capacitors C_{18} and C_{20} suppress any r.f. radiation. Since the impedance of moving-coil pickup elements is very low, the values of C_1 and C_6 are too low, which results in too wide a bandwidth. This is, therefore, narrowed by capacitors C_{19} and C_{21}.

Any interference on the supply lines to IC_3 and IC_4 is additionally decoupled by gyrators T_2 and T_3.

Regulators IC_5 and IC_6 provide stabilized ±15 V lines from a traditional power supply. The 20 V supply for the relay is separately rectified and smoothed. Resistor R_{28} provides some filtering of the line.

Note that, because of the small signal voltages, the supply contains rather more r.f. decoupling than usual. Since the mains voltage in the vicinity of a personal computer often is not too 'clean', mains filter L_1-C_{42} is provided at the primary of mains transformer Tr_1.

Diode D_2 is the obligatory on/off indicator.

CONSTRUCTION

The preamplifier is best built on the printed-circuit board shown in **Figure 4**. The board consists of three sections, which may be cut apart. This is highly advisable as far as the power supply is concerned, since, in view of stray fields around the mains transformers, this is best kept as far away as possible from the amplifier section(s).

Construction should present no problems provided it is done with constant and careful reference to the circuit diagram and the parts list. There are, nevertheless, a few points that need special mention.

The output of the pickup element is linked to the preamplifier via audio terminals K1 and K_2. For best, long-life performance, use gold-plated types.

The line input and output terminals, K_3 and K_4, are standard 3 mm audio sockets.

Note that a 24 V relay has been used, since this draws a smaller current than a 12 V type, which means that the (adverse) effect on the preamplifier of the ripple superimposed on this current is smaller.

The relay needs an energizing voltage of not less than 18 V, so that the 20 V provided in the present design is more than adequate. If a relay other than that specified is used, it may be possible to lower the current drawn by it by altering the value of R_{18}. Note that this resistor is not needed when the specified relay is used.

Table 1.
To lower the gain to 30 dB, alter the values of the following components as indicated.

R_2, R_9	=	162 Ω
R_3, R_{10}	=	49.9 kΩ
R_4, R_{11}	=	845 kΩ
R_5, R_{12}	=	3.83 kΩ
C_3, C_8	=	0.02 µF
C_4, C_9	=	0.0012 µF

Table 1 shows the values of which components need to be altered if the line input of the sound card in the computer needs a lower level.

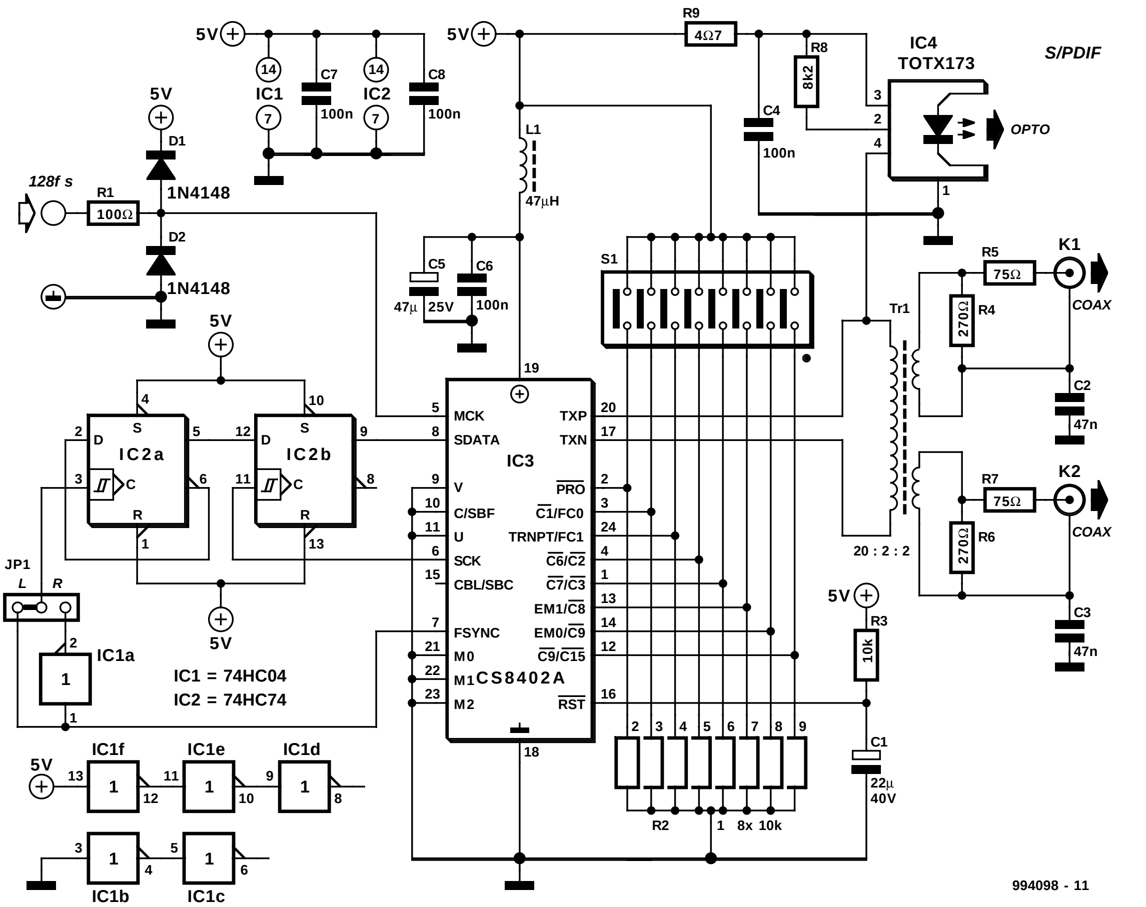
The board for the moving-coil type pickup element is linked to the main amplifier board via an 8-way single-in-line (SIL) connector, K_5, which is in essence a half IC socket. This, as well as the corresponding connector K_6, may also consist of an 8-way terminal strip. The two connectors or strips are linked by eight 15 mm lengths of 0.8 mm dia. insulated circuit wire. See the photograph in **Figure 5**.

To avoid any interference between the signal lines and the supply lines of the moving-coil board, the latter do not enter via K_6, but via three additional solder pins at the back of the board.

As mentioned earlier, the interior of a personal computer, and the space immediately surrounding it, are not exactly free of interference. It is, therefore, highly advisable to house the preamplifier in a well-screened metal enclosure.

[990048]

S/PDIF test generator



T. Giesberts

The generator is intended primarily for checking S/PDIF (Sony/Philips Digital Interface Format) receivers and any associated digital-to-analogue converters (DAC) and/or output filters. The external clock – standard TTL level – enables 128 sample frequencies to be generated. The clock may also be used for generating standard frequencies with the remaining inverters serving as crystal oscillators (provided a 74HCU04 is used).

The sender is a Type CS8402A digital audio interface transmitter from Crystal. In this short article it is not possible to list all settings that may be obtained with switch S_1: the reader is referred to the data sheet of the IC or to the 'sampling rate converter' published in the October 1996 issue of this magazine. The connections to the switch are exactly as described in that article.

There is an optical (IC_4) as well as a coaxial output (K_1, K_2). Toroidal transformer Tr_1 provides electrical isolation of the coaxial sockets and also serves to prevent earth loops. Capacitors C_2 and C_3 provide the earth connections for the sockets.

The transformer is wound on a TN13/7.5/5-3E25 core with a transformation ratio of 20:2:2 since TXP and TXN (on IC_3) are differential outputs. The primary voltage is 10 V_{pp} to give a signal across the 75 Ω coaxial outputs of 0.5 V_{pp}. After a reset, both outputs are low and are not short-circuited by Tr_1. A coarse audio signal is added to prevent, for instance, muting of the outputs.

Jumper JP_1 enables either the left-hand or right-hand signal to contain a rectangular signal at peak value and half the sampling frequency. This enables, for instance, the channel separation and the combination of digital and analogue signals to be checked. In most DACs, filter action commences at half the sampling frequency. At that instance, there is hardly any attenuation by the analogue filters, so that the level of the sinusoidal signal more or less coincides with that of a 0 dB signal. At this frequency, it is also clearly discernible whether de-emphasis correction is present ($S_{1–4}$ off: de-emphasis on) and, if so, whether this provides the requisite attenuation of 10 dB.

The CS8402A is used in mode 0 (low level at inputs M_0–M_2). This mode is really intended for interfacing with analogue-to-digital converters (ADC), but is used here since it enables the FSYNC of the L/R clock and the bit clock, SCK, to be derived internally from the MCK clock, and to be arranged as outputs. The data for half the sampling frequency are obtained by halving the L/R clock in IC_{2a}. Since the data must be the 2s complement, they are shifted by one clock period in IC_{2b}, so that, depending on the phase of the L/R clock, that is, inverter IC_{1a}, either the left-hand or the right-hand channel contains a peak-level signal. The other channel then toggles one LSB at identical frequency.

It should be noted that some DACs, particularly 1st generation 1-bit types, fail to operate correctly with 0 dB signals, which may cause difficulties with overdriven CDs (see 'Clipping and the CD' in the Readers' Letters column in the April 1999 issue of this magazine). This may be checked with the present generator. If the audio signal is not wanted, the SDATA input should be linked to earth and IC_1 and IC_2 omitted.

Resistor R_1 and diodes D_1, D_2 protect the MCK input against excessive or unbalanced clock signals.

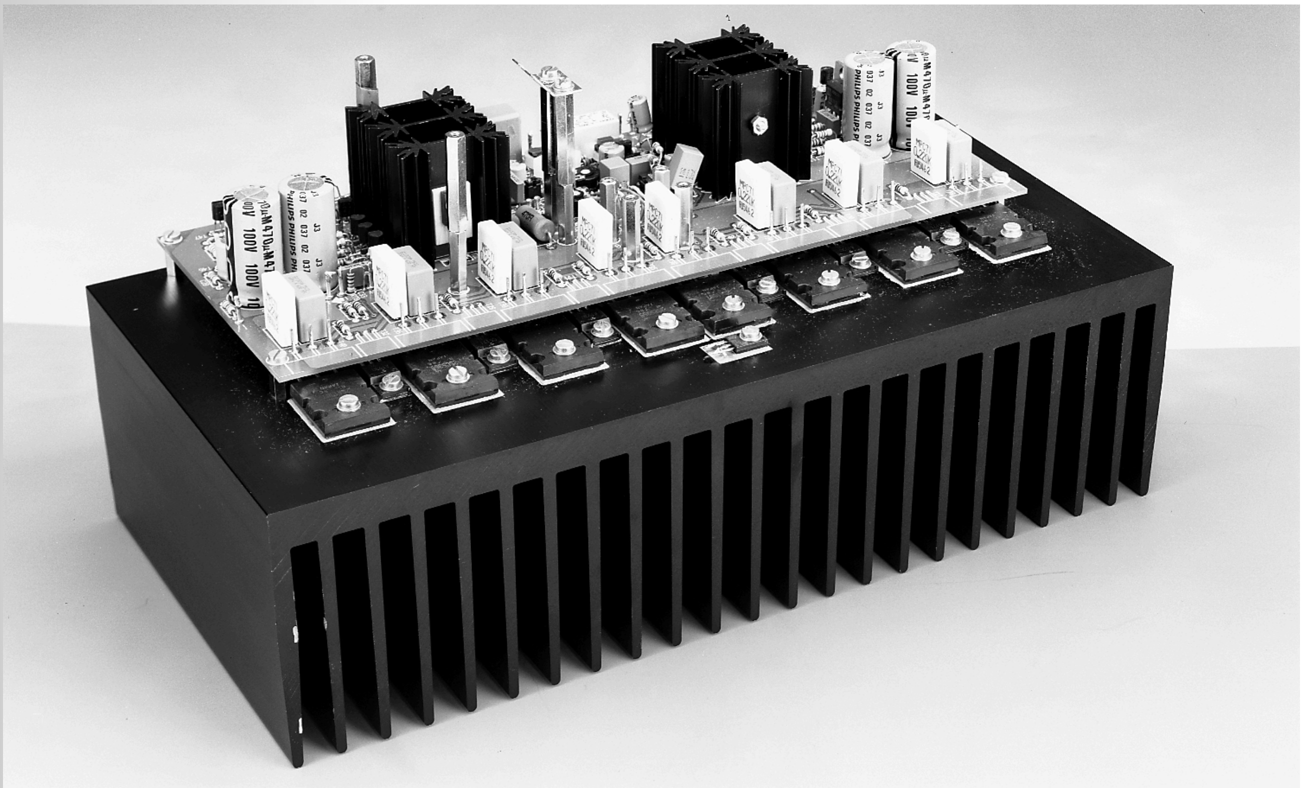
The generator draws a current of about 30 mA.

[994098]

Contents

Titan 2000

Part 4: wiring and performance



This fourth of five parts deals primarily with the wiring up of the amplifier and ends with a brief resume of its performance and specifications. The fifth and final part of the article in a forthcoming issue will deal with the temperature control, bridge configuration and some other practical hints.

Design by T. Giesberts

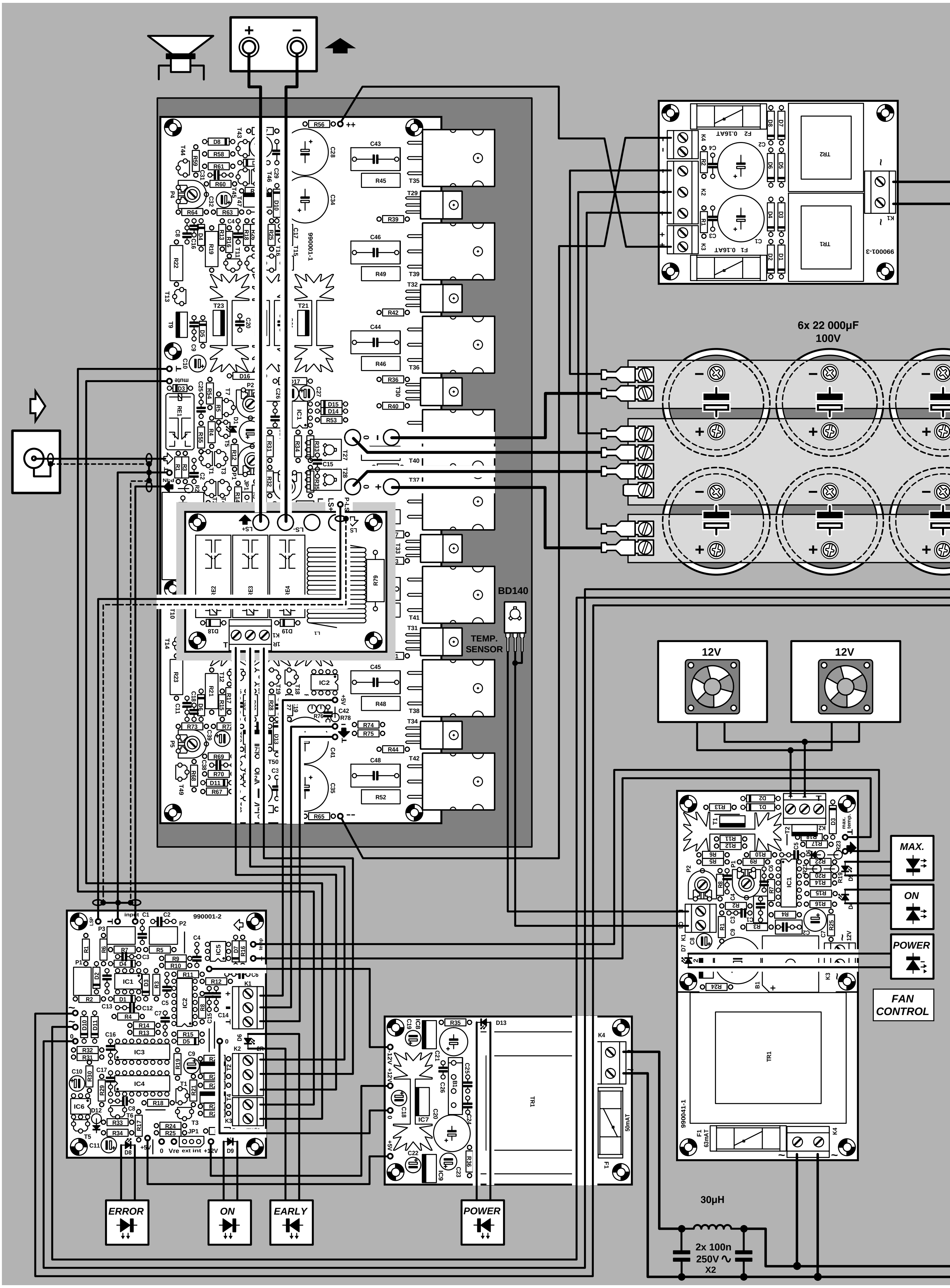
WIRING UP

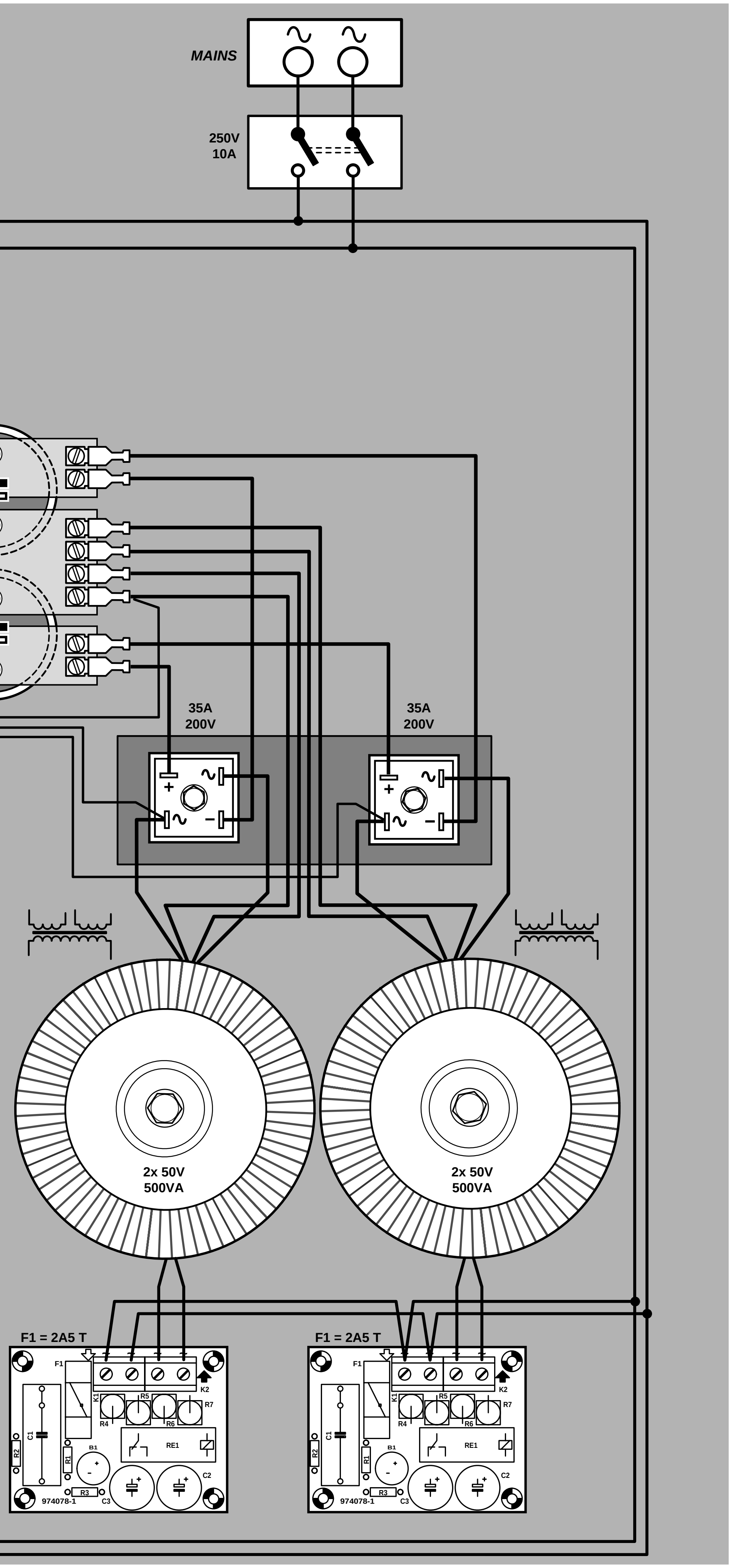
How the various board, power supplies, controls and terminals are combined into an effective and interference-free unit is shown in **Figure 16**.

As already mentioned in Part 2, all wiring carrying the main supply voltage (±70 V) must be insulated, high-current wire to BS6321 with a conductor size of 50/0.25 (2.5 mm²). This wire should also be used to link the output terminals of the power transistors and the loudspeaker terminals. Any wiring between smoothing capacitors and the board should not exceed 15 cm and be preferably much shorter. This kind of wire is best terminated into car-type connectors.

Other wiring may be made in light-duty, stranded, insulated hook-up wire. It is advisable (and may prove to be very helpful in case of problems) to use wire with different colour insulation for dissimilar functions.

The connections between the input socket and board must, of course, be in screened audio cable. To avoid earth loops, the socket should be isolated from a metal enclosure. Bear in mind that the supply earth and the enclosure are linked by metal spacers between the two '0' terminals and the heat sink. It is, therefore, essential that the heat sink is firmly strapped to the metal enclosure.


6x 22 000µF
100V
BD140
TEMP.
SENSOR
12V
12V
MAX.
ON
POWER
FAN
CONTROL
ERROR
ON
EARLY
POWER
30µH
2x 100n
250V
X2
990001-1
990001-2
990001-3
990041-1



The on/off indicator, the functional indicators, and the mains on/off switch should, of course, be fitted on the front panel of the enclosure. The mains on/off switch must be a 10 A or 15 A type.

If the output power of the amplifier is limited to no more than 500 W, in which case the enclosure does not need fan cooling, the heat sink may be mounted at the outside of the enclosure or even form the sidewall or back of a home-made enclosure.

For greater output powers, cooling fans with relevant apertures at the front and back of the enclosure are a must. The heat sink must then be located in the enclosure in such a position that it is directly between the two fans, ensuring a continuous supply of cooling air.

PERFORMANCE

The specification and associated comments in the box cannot, of course, give a full impression of the performance of the amplifier. It is a well-known fact that amplifiers with an almost identical specification, and using identical loudspeakers, can sound quite different.

Particularly at low frequencies, the amplifier maintains good control over the loudspeaker, which results in a clean fast (i.e., taut over the whole audio range) sound, totally lacking in reverberation. High and medium frequencies were also reproduced with excellent definition and without any trace of tizziness.

The overall impression is that the amplifier has plenty of reserve and is not strained in any circumstances.

In next month's final instalment, the temperature control and possible bridge configuration will be discussed.

[990001-3]

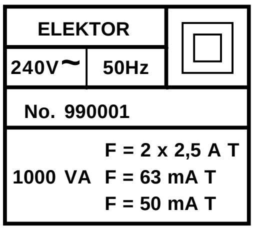


Figure 16. The wiring diagram clearly illustrates how the various parts of the amplifier are combined into a single unit.

Contents

Technical specifications

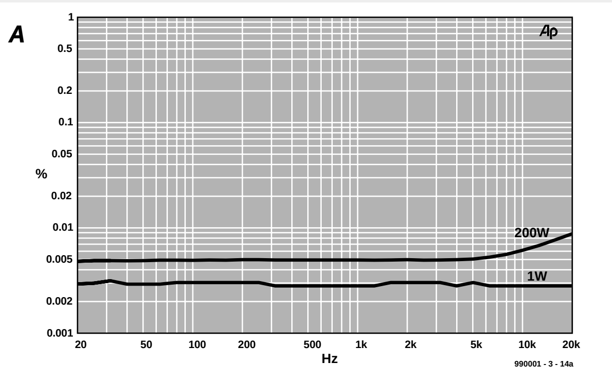
(Supply voltage = ±70 V; quiescent current = 200–400 mA)

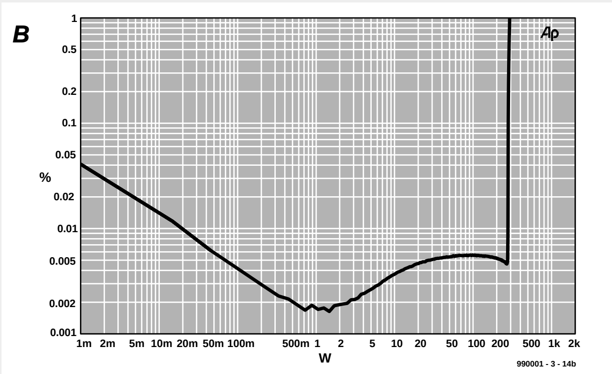
Input sensitivity	1.1 V r.m.s.
Input impedance	47.5 kΩ
Sine-wave power output (0.1% THD)	280 W into 8 Ω; 500 W into 4 Ω; 800 W into 2 Ω
Music power* (1% THD)	300 W into 8 Ω; 550 W into 4 Ω; 1000 W into 2 Ω
Slew limiting	85 V μs^{-1}
Open-loop bandwidth	53 kHz
Open-loop amplification	×8600
Power bandwidth	1.5 Hz – 220 kHz
Signal-to-noise ratio (1 W into 8 Ω)	101 dB (A-weighted); 97 dB (B = 22 kHz)
Damping factor (at 8 Ω)	>700 (1 kHz); >300 (20 kHz)
Output impedance	1.6 Ω

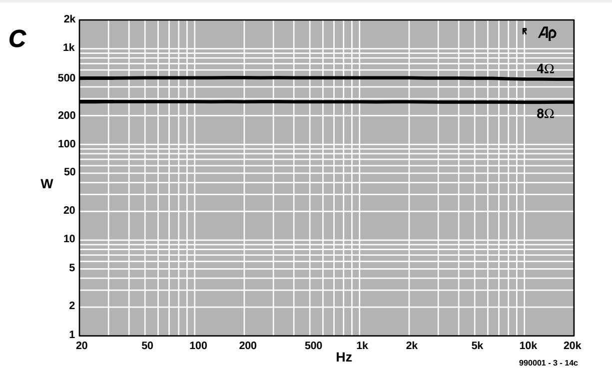
Harmonic distortion (THD) (B = 80 kHz)	8 Ω	4 Ω	2 Ω
at 1 kHz	0.003% (1 W)	0.0046% (1 W)	0.01% (1 W)
	0.005% (200 W)	0.0084% (400 W)	0.02% (700 W)
at 20 kHz	0.009% (200 W)	0.018% (400 W)	0.07% (700 W)
Intermodulation distortion (IM)			
(50 Hz:7 kHz = 4:1)	0.004% (1 W)	0.01% (1 W)	0.034% (1 W
	0.016% (150 W)	0.025% (300W)	0.07% (500 W)
Dynamic IM			
(square wave 3.15 kHz with sine wave 15 kHz)	0.003% (1 W)	0.0036% (1W)	0.0055% (1 W)
	0.003% (200 W)	0.005% (400 W)	0.0085% (700 W)

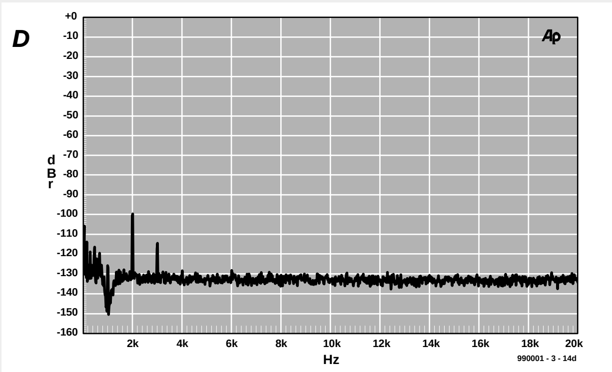
*See Part 1 about the validity of this meaningless quantity.

The specified figures were measured after the amplifier had been switched on for two hours. The figure show that the Titan 2000 compares favourably with most amplifiers. The slew limiting is a measure of the speed of the amplifier, which is exceptionally good in the Titan 2000.
Figure A shows the total harmonic distortion plus noise (THD+N) for an output of 1 W into 8 Ω (lower curve) and for 200 W into 8 Ω. The latter figure corresponds with 70% of the peak sine wave power and the curve shows that the distortion increases clearly only above 10 kHz.
Figure B shows the THD+N at 1 kHz as a function of the drive with an output impedance of 8 Ω. The curve is purposely drawn for a bandwidth of 22 kHz so that the noise above 20 kHz does not degrade the performance of the amplifier. From about 2 W, the distortion increases slightly with increasing drive, which is normal in most amplifiers.
Figure C shows the peak output of the amplifier at a constant distortion of 0.1% and a load of 4 Ω (upper curve) and 8 Ω. The bandwidth was 80 kHz.
Figure D shows a Fourier analysis of a reproduced 1 kHz signal at a level of 1 W into 8 Ω. It will be seen that the 2nd harmonics are down just about 100 dB, while the 3rd harmonics are down to –114 dB. Higher harmonics lie below the noise floor of –130 dB.









The Audio Collection

Contents

Amplifiers

The Audio Collection

30 W AF amplifier for cars
50 W a.f. amplifier
100-watt single-IC amplifier

Digital audio

AES/EBU-to-S/PDIF converter
clipping indicator
copybit eliminator revisited
copybit inverter
digital-audio-input selector
opto-to-coaxial audio converter
sampling rate converter
simple copybit killer
S/PDIF test generator
S/PDIF-to-AES/EBU converter
splitter for S/PDIF coax/optical output

Loudspeakers

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loudspeaker protector
simple two-way loudspeaker
surround speakers
surround-sound subwoofer - 1
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surround-sound subwoofer - 3
the wall box

Test and measurement

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digital VU meter - 2

The Audio Collection

passive VU meter
portable sound-pressure meter
sound pressure meter

Miscellaneous

accurate bass tone control
active potentiometer
active power buffer
AF input selection
AF level matching
auto volume control
AVC for PCs
balanced microphone preamplifier
balanced/unbalanced converters for audio signals
bass extension for surround sound
car booster adaptor
d.c. detector
delay line
digital potentiometer
DIY: from vinyl to compact disk
fan control
front/rear car radio fader
infra-red-controlled noiseless volume control
Insulated Gate Bipolar Transistors (IGBTs)
mains on delay circuit
parametric equalizer
pick-up input becomes line input
playback amplifier for cassette deck
silent volume control
stereo microphone amplifier

The Audio Collection

stereo microphone input adaptor for PC
surround sound indicator
surround sound processor
symmetrical supply in cars
ten-band equalizer
treble tone control
ultra low-noise MC preamplifier
µP-controlled mixer board - 1
µP-controlled mixer board - 2
up/down drive for tone control
20-bit A-D converter