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Elektor Electronics

Droitwich time standard

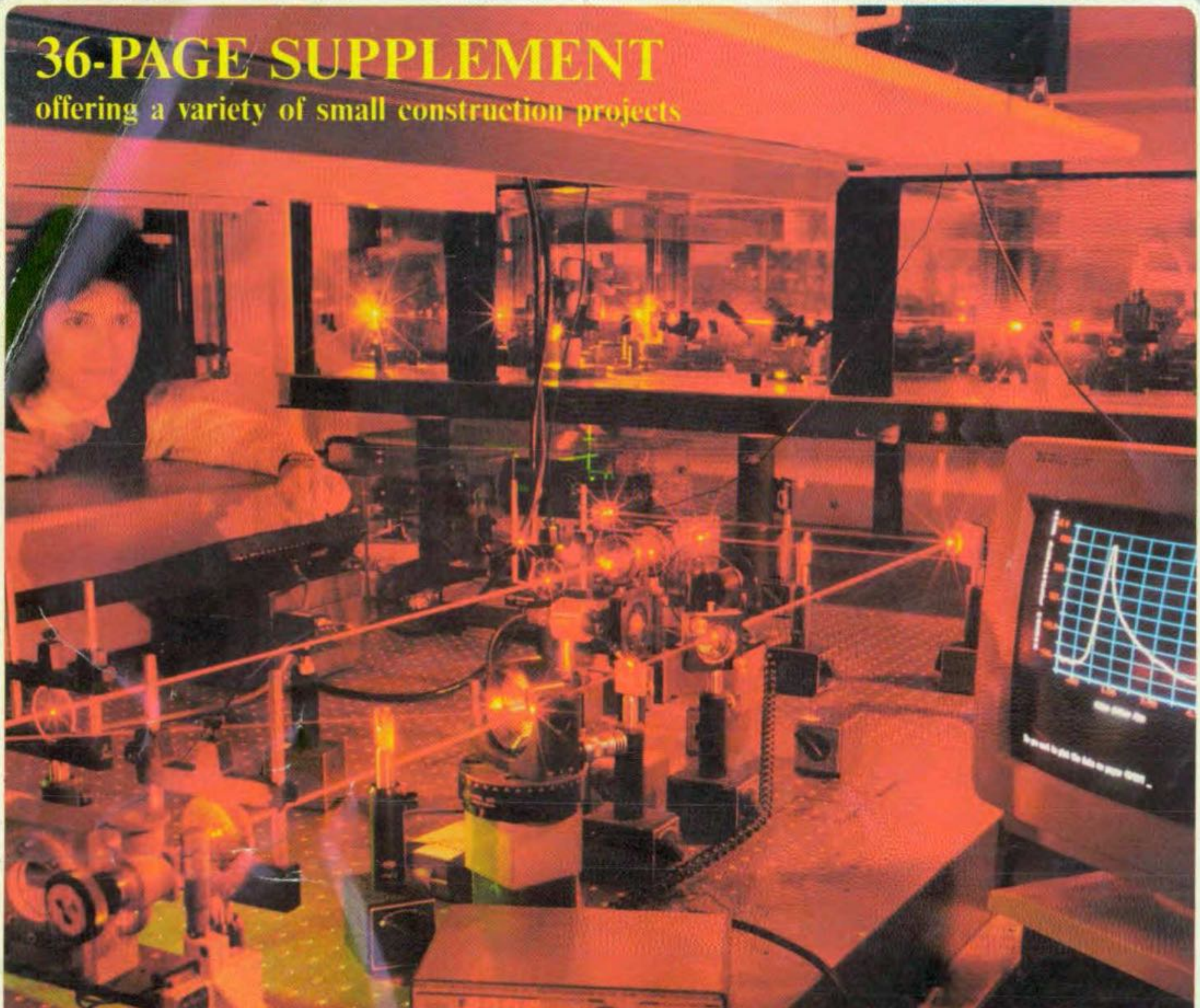
Chopper opamps

Milliohm meter

Metal transmission lines

PC videotext decoder

Line pulse fundamentals



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Elektor Electronics

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Front cover

Semiconductor devices, built from organic material rather than silicon, are being tested on this femtosecond laser system at Cambridge University's Cavendish Laboratory, a new facility for studying the behaviour of materials on very short time scales. Although the devices are quite large in area, they are composed of polymer no more than 200 Å, or about 100 molecules, thick. The Femtosecond Laser Group is a world leader in molecular electronics, in particular in making semiconductor devices from organic materials.

A sample is excited by a laser beam, while a second laser beam is used to measure the change in colour of the sample. Such non-linear optical processes could be exploited in optical computers which might be up to a million times faster than current supercomputers. Future work will include studies of biological material, in particular the genetic material DNA in the form of a virus, and the visual pigment rhodopsin.

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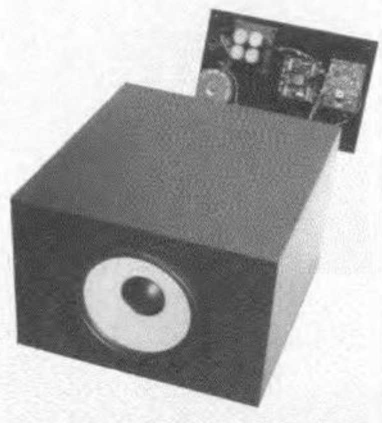
Cavendish Laboratory
Madingley Road
CAMBRIDGE CB3 0HE
England**

We thank all our readers wherever they are for their continued support and wish them all a Prosperous and Peaceful New Year!

WE APOLOGIZE
for the unfortunate omission from the left-hand column on page 31 of our November 1990 issue of the following 'autogram'

45 DOTS



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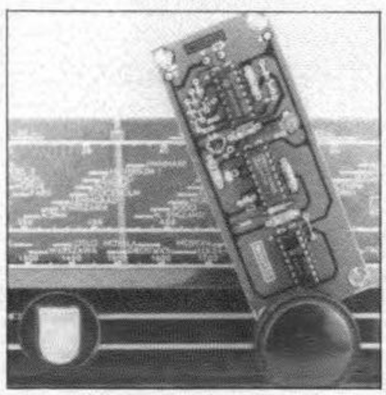

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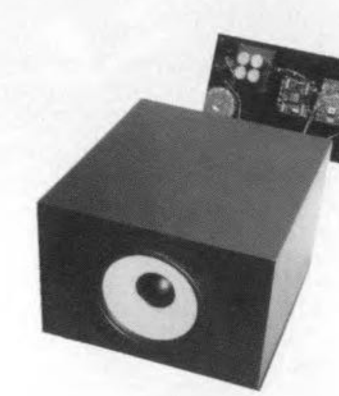
ACTIVE MINI SUBWOOFER - PART 2

by T. Giffard

THIS second part of the article describes an output amplifier designed for the subwoofer; the fitting of the electronics in the enclosure; and how the subwoofer can be connected to an existing audio system.



Output amplifier

Although in principle any output amplifier that can deliver about 50 watts into 8 Ω may be used with the subwoofer, we felt that many readers would want a complete system and so we designed an output amplifier especially for them.

The amplifier is a hybrid circuit consisting of a control section based on an opamp, and a power section that uses discrete transistors. Its circuit diagram is shown in Fig. 8.

The opamp, a Type OP16 from PMI, is a precision type with JFET inputs and a slew rate of 25 V/μs. It has its own power supply of ±15 V, which is derived from the 30-V main supply via R15/D4 and R16/D5.

The input signal is taken to the non-inverting input of the opamp via C1. The input impedance is determined almost entirely by R1 (since the opamp has JFET inputs).

The bandwidth of the OP16 is restricted to some extent by a 2.2 nF capacitor between the output and inverting input, and a 100 Ω resistor between the inverting input and ground. This arrangement may be compared to the compensation capacitor between the outputs of the first differential amplifier in a conventional output stage.

The output of the opamp drives the power section via a current source based on T1. This source ensures a stable setting of the quiescent current through the output transistors. The voltage reference in the source is provided by a high-efficiency LED (D1).

The power section consists of a complementary compound configuration, T3–T6. Normally, a kind of super emitter follower is used in the output to ensure adequate current amplification. In the present design, current amplification alone (a typical characteristic of an emitter follower) is not sufficient, because the signal excursion at the output of the opamp is limited to about ±12 V. Some additional amplification is therefore needed. A compound circuit provides current as well as voltage amplification.

The voltage amplification in the present circuit is determined by the amplification factor of the output transistors and the potential divider, R9-R10, between the output transistors and the drivers. To make sure that the opamp does not provide too high an output voltage, which would limit the output current, the amplification of the compound output circuit has been made ×4 (12 dB).

Notable in this output stage configuration is the location of the emitter resistors of the output transistors, which are connected to the power rails.

Setting of the quiescent current level is accomplished with variable 'zener diode' T2-P1-R4. Transistor T2 is clamped to the heat sink between the output transistors to ensure good thermal coupling. Capacitors C7 and C13 provide a.c. decoupling of the 'zener'.

The feedback loop of the overall amplifier consists of resistors R2 and R3, which set the overall amplificatiopn to ×23 (27 dB).

The circuit around T7 and Re1 provides a delay of a few seconds between power on and connection between the loudspeaker and the

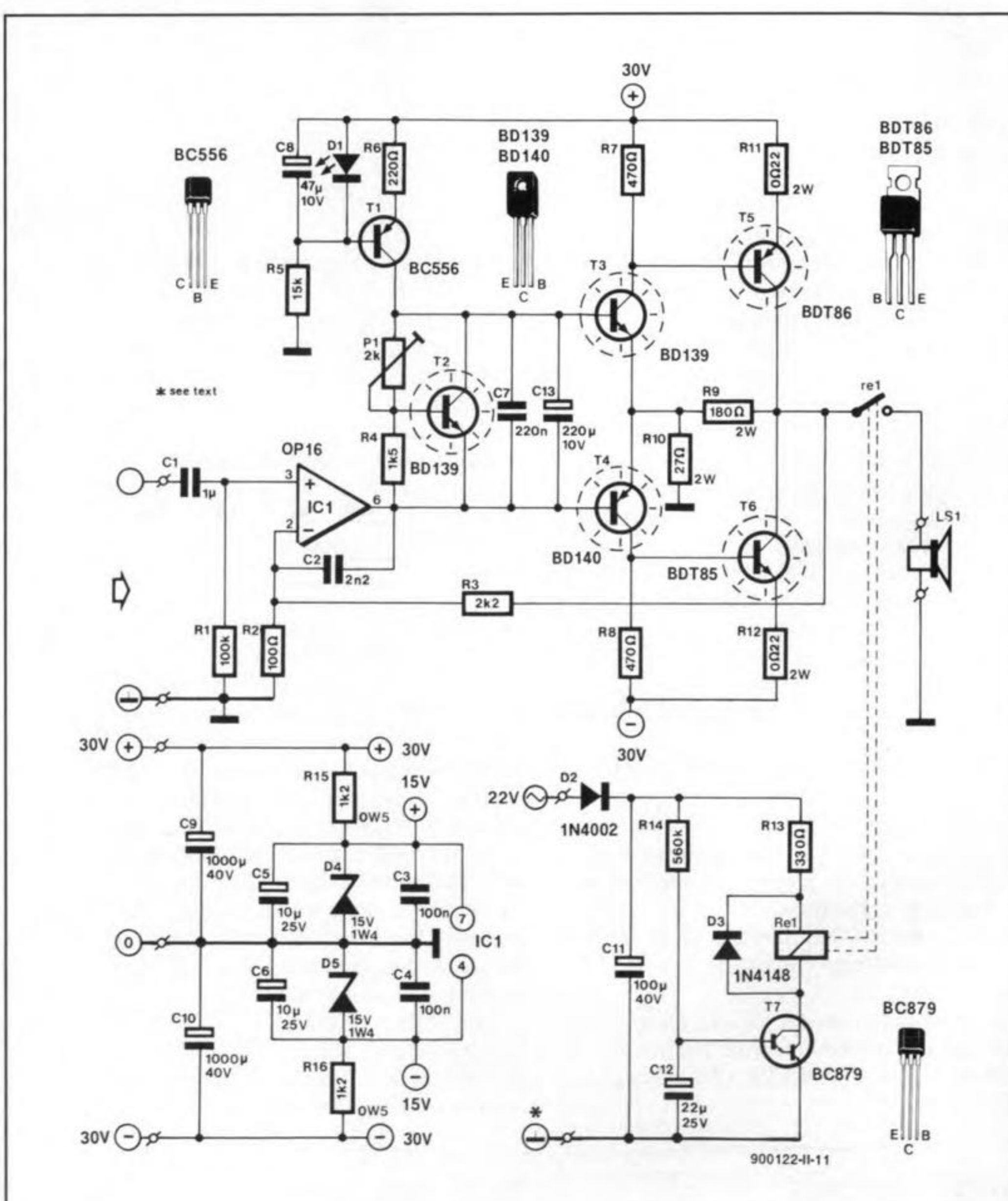


Fig. 8. Circuit diagram of the output amplifier specially designed for use with the subwoofer.

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output stage being made. It derives power from the main power supply via D2: this ensures that the relay is deenergized as soon as the power is switched off.

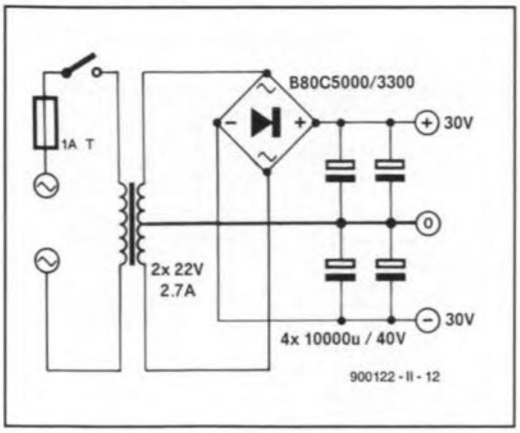

Fig. 9. Power supply for the output amplifier.

The circuit of the power supply is straightforward—see Fig. 9. Apart from the four 10 000 μF capacitors shown here, two more 1000 μF capacitors on the board provide additional decoupling of the power lines.

Construction

The amplifier is best built on the PCB shown in Fig. 10. Apart from the mounting of transistors T2–T6, the construction should not present any problems.

Transistors T2–T6 may be fitted in various ways, depending on the mechanical construction. If use is made of an aluminium L-section, they can be fitted above the board and fastened to the L-section, which in turn is screwed to the heat sink.

It is, however, also feasible to screw the amplifier and filter boards on to an aluminium sheet of suitable size, which then serves as the heat sink. In that case, fit T2–T6 to the sheet first, bend their terminal wires upwards a couple of millimetres above their body and pass these through the relevant holes in the PCB. Make sure that sufficient space is left between the board and sheet to allow solder connections to be made. Also, bear in mind that the transistors must be insulated from the sheet.

For clarity's sake, the latter construction, on a 3 mm thick aluminium sheet, is shown in Fig. 11. The dimensions of the sheet allow it to be fitted in the space in the back of the subwoofer enclosure. For that purpose, glue four triangular wooden supports in the corners of that space to which the built-up sheet is screwed later on.

Fit the boards to the sheet with the aid of 10 mm spacers.

COMPONENTS LIST

Resistors:
R1 = 100 kΩ
R2 = 100 Ω
R3 = 2k2
R4 = 1k5
R5 = 15 kΩ
R6 = 220 Ω
R7, R8 = 470 Ω
R9 = 180 Ω; 2.5 W
R10 = 27 Ω; 2.5 W
R11, R12 = 0.22 Ω; 5 W
R13 = 330 Ω; 1 W
R14 = 560 Ω
R15, R16 = 1k2; 0.5 W
P1 = 2 kΩ; multi-turn preset; top adjust

Capacitors:
C1 = 1 μF
C2 = 2n2
C3, C4 = 100 nF
C5, C6 = 10 μF; 25 V
C7 = 220 nF
C8 = 47 μF; 10 V
C9, C10 = 1000 μF; 40 V
C11 = 100 μF; 40 V
C12 = 22 μF; 25 V
C13 = 220 μF; 10 V; radial

Semiconductors:
D1 = 3 mm LED; red; high efficiency
D2 = 1N4002
D3 = 1N4148
D4, D5 = zener diode 15 V; 1.4 W
T1 = BC556
T2, T3 = BD139
T4 = BD140
T5 = BDT86 or BD912
T6 = BDT 85 or BD911
T7 = BC879
IC1 = OP16

Miscellaneous:
Re1 = relay; 24 V; 1 change-over
Mains transformer,
secondary 2×22 V, 2.7 A
4 electrolytic capacitors 10 000 μF; 40 V
Bridge rectifier B80C5000/3300
PCB Type 900122-2

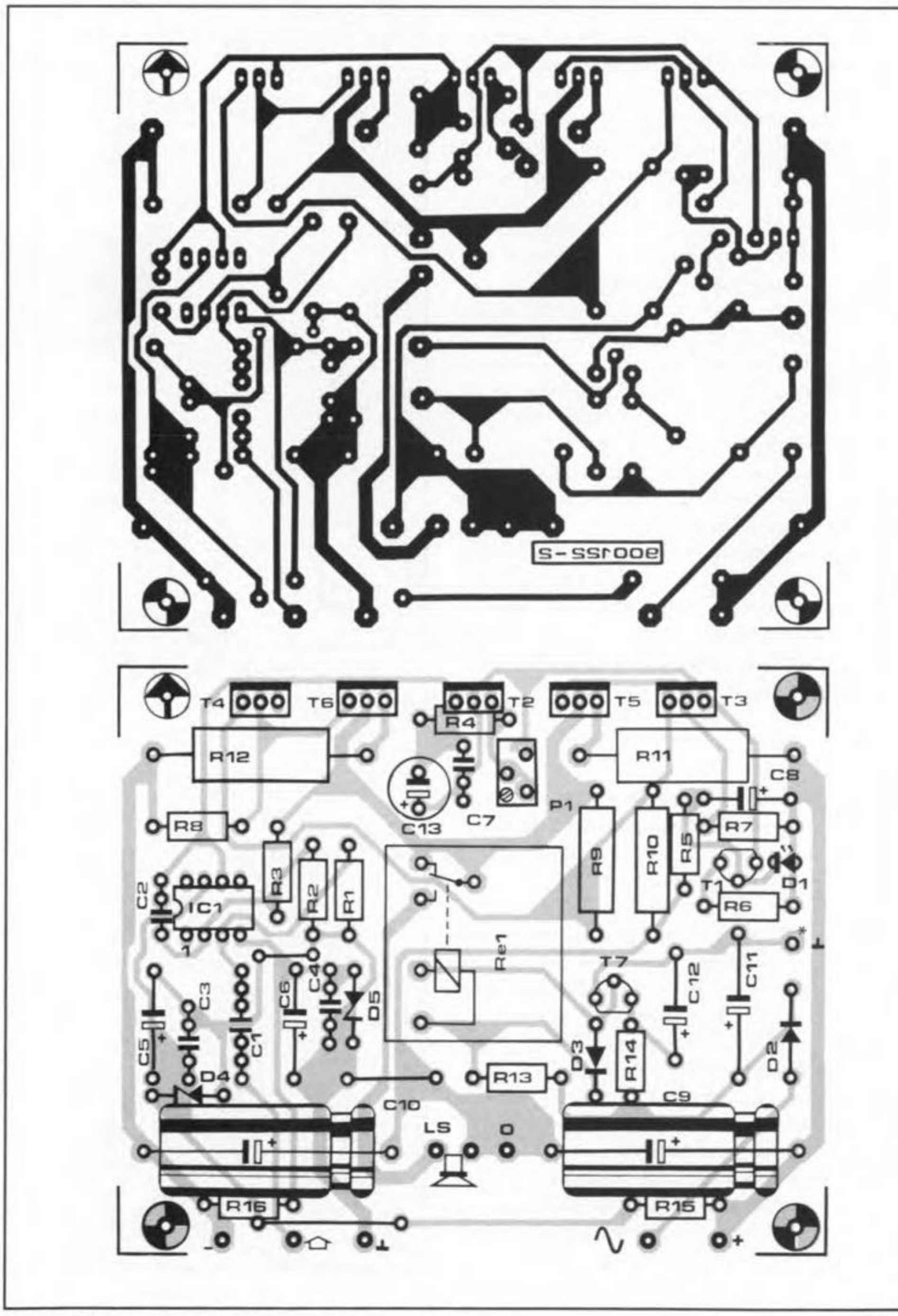

Fig. 10. Printed circuit board for the output amplifier.

The power supply is fitted as far away from the boards as possible to avoid any possibility of hum.

Note the separate earth connection for the delay circuit (indicated on the PCB by an earth symbol and asterisk) to the central earthing point. **Do not make a direct connection between the two earthing points on the amplifier board.**

Do not yet connect the loudspeaker to the amplifier.

When everything is ready, first set P1 for minimum resistance and then switch on the mains. Next, adjust P1 for a quiescent current through the output amplifier of 100 mA: this is measured with a millivoltmeter across R11 or R12 where the reading should be 22 mV.

Finally, switch off the mains, connect the loudspeaker to the amplifier and close the loudspeaker box.

Connecting the subwoofer

There are two ways in which to connect the subwoofer to an existing audio system. If the system has discrete pre- and output-amplifiers, or an external connection between these units when integrated, the best way is to feed the output of the pre-amplifier to the subwoofer via a screened audio cable. If that is not possible, connect the (second pair of) loudspeaker terminals of the system to the banana sockets on the subwoofer.

When the connections between the audio system and its loudspeaker boxes are long, it is possible to extend them from these boxes to the subwoofers, since the latter should in any case be near the loudspeakers for optimum performance.

The low cut-off point of the existing system and the subwoofers may be matched in several ways. When separate pre- and output-amplifiers are used, a simple first-order high-pass filter may be provided by adapting the input capacitor of the power amplifier. If the input impedance, Z, of the power amplifier is known, the value of the capacitor for a cut-off frequency, f, is given by:

$$C = 1/2\pi f Z \qquad [F]$$

Another way is adapting the cross-over network in the loudspeaker boxes. This is not so simple, however, because in the low frequency range the resonance peak of the subwoofer will have an effect, so that the filter cannot be terminated into a pure resistance.

A third possibility is to leave everything as it is. Particularly with small loudspeaker boxes where the low cut-off frequency is in any case fairly high—normally 75–100 Ω— it is perfectly all right to just connect the subwoofers into the system.

A fourth solution would be to precede the present output stage by a cross-over network of a type of which we have published several during the past few years. This is a rather exaggerated solution, but it is there if you want.

The location of the subwoofers is not very important, but they should preferably be not too far from the loudspeakers. Critical listeners may like them between the loudspeakers. The sound level may be set with the potentiometer at the back of the sub woofers.

Finally, the input signals may be inverted with the aid of the phase switch if needed. Some experimentation here may well prove to be interesting. ■

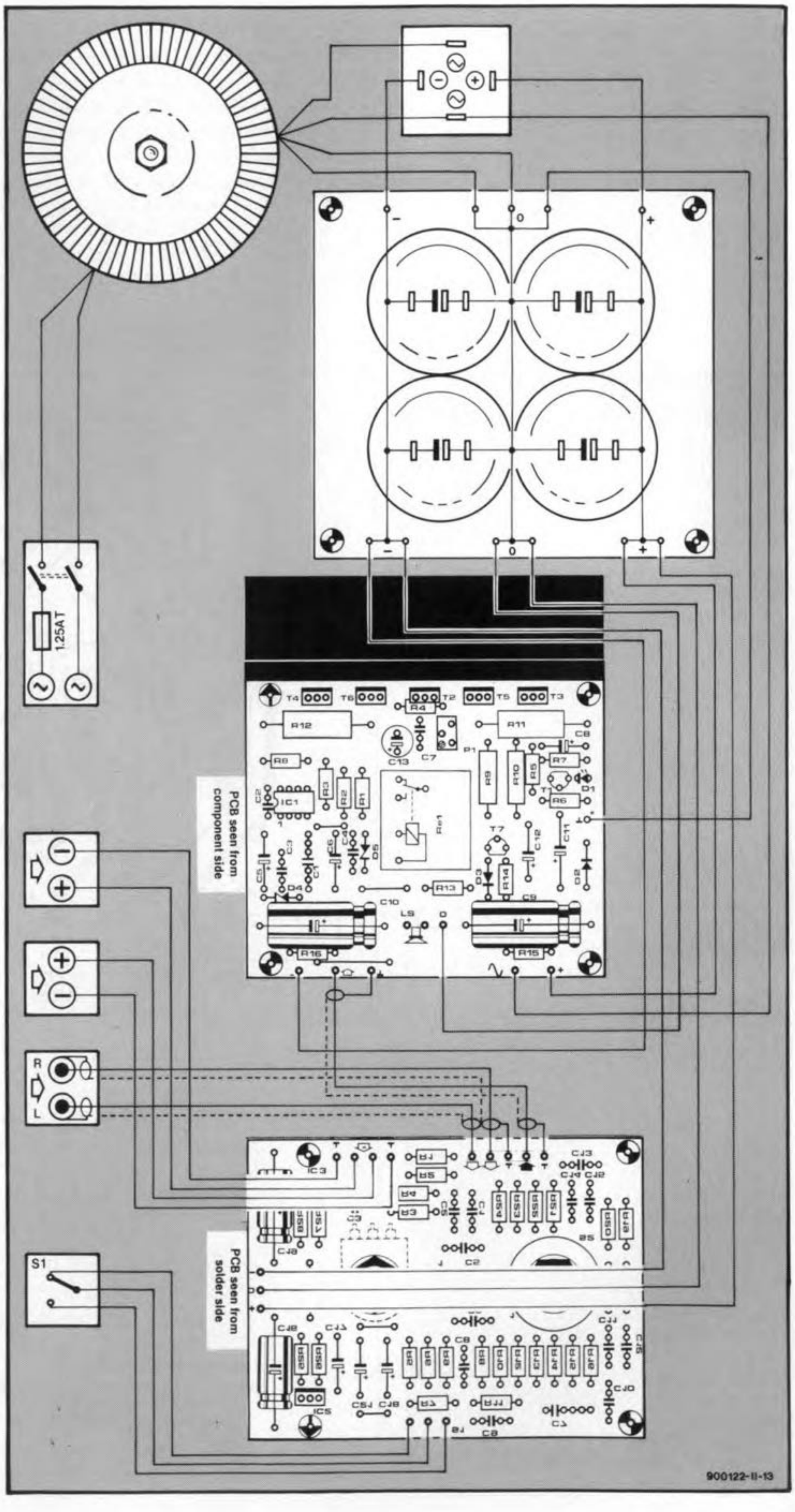

Fig. 11. Wiring diagram of the output amplifier complete with its power supply.

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R4 = 1k5
R5 = 15 kΩ
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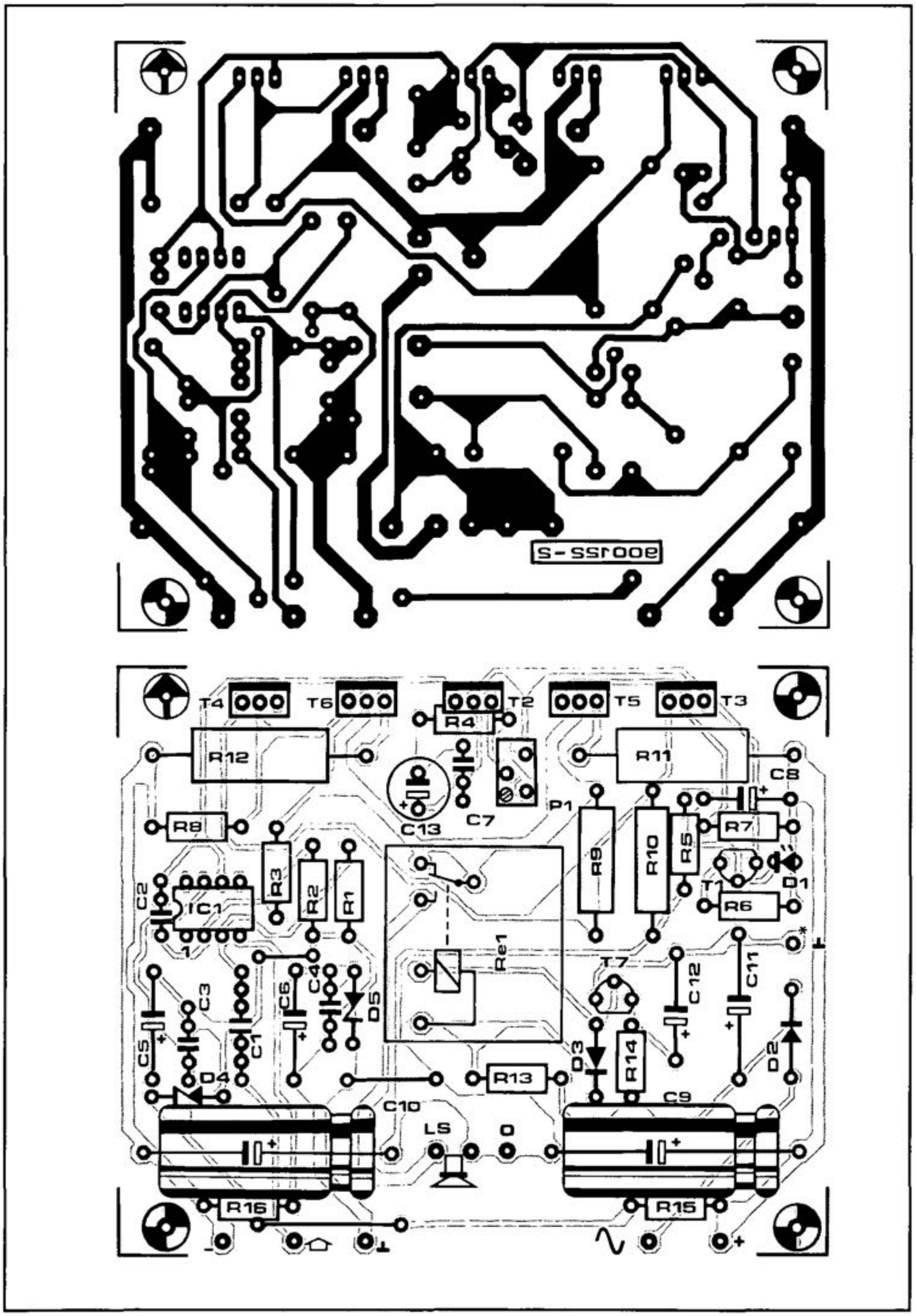

Fig. 10. Printed circuit board for the output amplifier.

LINE PULSE FUNDAMENTALS:

SOME PROBLEMS UNKNOTTED

By Bryan Hart

Introduction

Throw a stone the size of a golf-ball into a can of water the size of a tea-cup and see what happens: virtually all points on the surface of the water are disturbed simultaneously. This is a rough mechanical analogy to the case of a 'lumped' electrical circuit, e.g., a simple resistive potentiometer, comprising two resistors, subjected to a transient input.

Throw the same stone into the middle of a village pond and observe a different effect: all points on the surface of the pond are not affected simultaneously; they are disturbed only as ripples move outward from the point where the stone falls. This is a crude mechanical analogy to the case of a 'distributed' electrical circuit, notably a transmission line, with a transient input.

The difference between the two cases arises through the finite time taken for disturbances to be transmitted. A variation of the pond analogy, which has long been used in the study of wave transmission, is the 'canal' analogy. In this, we consider what happens when a straight plank is dropped into a canal in a direction perpendicular to its length. Straight ripples, parallel to the length of the plank, move outward from the place where it falls. This analogy is more appropriate in the discussion that follows, because propagation is characterized by movement principally in one dimension.

Transmission lines are very important in digital electronics because of their use in the distribution of fast logic signals, but their operation is sometimes a puzzle to budding engineers (some with a predominantly mechanical engineering background), who have been taught the basic principles of lumped-circuit electronics but who have not studied established Electromagnetic Theory (or been convinced by it, even if they had!)

This introductory article sets out to clarify the understanding of some fundamental aspects of the pulse operation of transmission lines, particularly the popular twisted pair line (t.p.l.). The aim is to concentrate on the basic circuit theory aspects and practically observable waveforms that support the theoretical background.

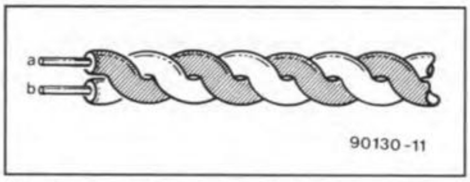

Fig. 1. A twisted pair line (t.p.l.).

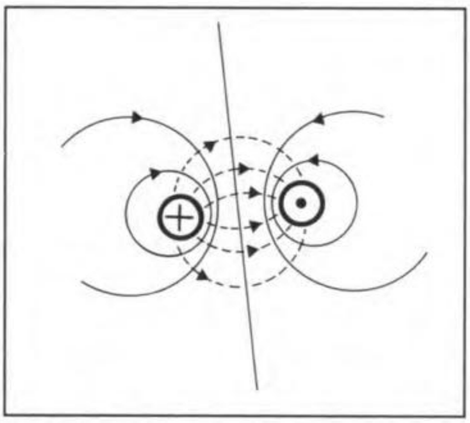

Fig. 2. Field patterns at a point on a t.p.l. under d.c. conditions. Solid lines = magnetic field; dashed lines = electric field.

Line modelling

A section of t.p.l. is shown diagrammatically in Fig. 1. In reel form, this can be purchased commercially (e.g. from RS Components), but for line lengths of a few metres, a t.p.l. may be made up by twisting together, uniformly, two pieces of PVC insulated wire (26 gauge, say) with a pitch of about 5 cm.

If we imagine the t.p.l. as laid along an x–axis perpendicular to the plane of this page, the field patterns that exist when equal-magnitude direct currents flow into the page at 'a' and 'b' respectively are shown in Fig. 2, where the solid lines indicate the nature of the magnetic field and the dashed lines the configuration of the electric field. These field patterns correspond also to those of the basic propagation mode for line transients discussed throughout this article.

The magnetic flux linking the wires is proportional to the current. The flux per unit current is represented by a series-inductance L per unit length. L is a parameter dependent on conductor geometry and can be estimated by analytical principles well known in field theory but a knowledge of L, by itself, is rarely required by t.p.l. users and, if needed, is best inferred from other readily measurable parameters. The electric field and flux associated with the conductors and the line charge on them are proportional to the p.d. between them, so the t.p.l. has also a per-unit-length capacitance, C. As with L, this can be estimated theoretically, if required, but is readily determined practically.

Series losses may be represented by a per-unit-length resistance R and shunt losses resulting from leakage, through wire insulation, by a per-unit-length conductance, G. The t.p.l., although distributed in nature, can nevertheless be considered as made up from as large a number as we wish of tiny lumped sections, each of length ∂x, connected in series. The idea of using a large number of small discrete lumps to simulate a continuous variable is not unfamiliar in electronics. Thus, a digital time base for an oscilloscope based on a counter and D-A converter produces a horizontal pattern of dots on the

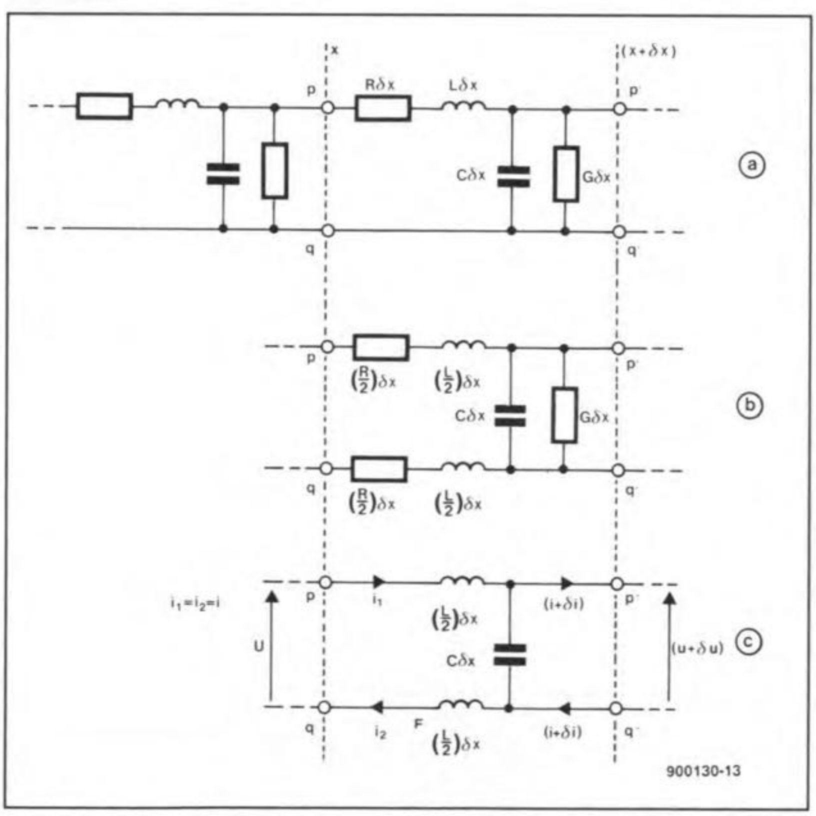

Fig. 3. (a) a t.p.l. made up from lumped 'L-shaped' sections; (b) equivalent form for (a); (c) reduced form for (b) for lossless line [$R = G = 0$].

screen. However, for a 10-bit converter, the number of dots exceeds 1000 and on a 10-cm screen the trace appears continuous.

The specific configuration of series and shunt components adopted to model an elemental section of line is a matter of sensible choice. All choices must, by definition, be equivalent in electrical characterization. We could use a 'T-section', but the 'L-section' shown in Fig. 3(a) is analytically more convenient.

Figure 3(a) is often used for coaxial cable lines in which the outer conductor is 'earthed' but this can be misleading, particularly for a t.p.l., because it may give the false impression that one of the conductors behaves in a different way, electrically, from the other. The alternative model shown in Fig. 3(b) shows R and L as equally shared between the two conductors of the t.p.l. and in that respect is conceptually more attractive.

For a t.p.l. a few metres long, series and shunt losses can usually be neglected and the section reduces to the 'ideal' or 'lossless' form ($R = G = 0$); it is tempting to say that this is 'fortunate' for were it not so, the t.p.l. would be of very restricted use.

For this case, the relevant equations lend themselves simply to pictorial interpretation and the essential features of line operation are not obscured by second-order effects.

Line equations

Consider the section shown in Fig. 3(c). The currents, i_1, i_2, shown flowing in the upper and lower inductance elements must be equal in magnitude to i, say.

The reason for this is as follows. If we imagine the line to the right of the points p and q to be contained within the 'black box', the Law of Conservation of charge requires that $\int(i_1 - i_2)dt = 0$ This is true only, irrespective of the timescale t, if $i_1 = i_2 = i$.

Applying Kirchhoff's Voltage Law for loop voltage drops,

$$u = (u_L/2) + (u + \partial u) + (u_L/2),$$

where u_L, the inductive voltage drop, is given by

$$u_L = (L\partial x)(\partial i/\partial t).$$

Substituting for u_L and rearranging:

$$-(\partial u/\partial x) = L(\partial i/\partial t) \qquad [1]$$

In passing, it may seem contrary to write the p.d. at $(x + \partial x)$ as $(u + \partial u)$, with a plus sign for the increment, when physical considerations tell us that it must be less than u. However, this is in the tradition of differential calculus. The physics of the problem gives the negative sign in [1].

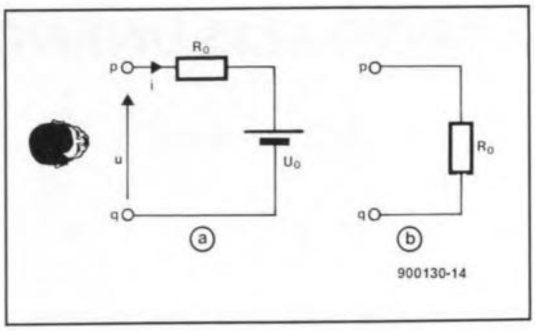

Fig. 4. View of line looking right between p and q at $t = 0$; (a) intitally charged line; (b) initially uncharged line.

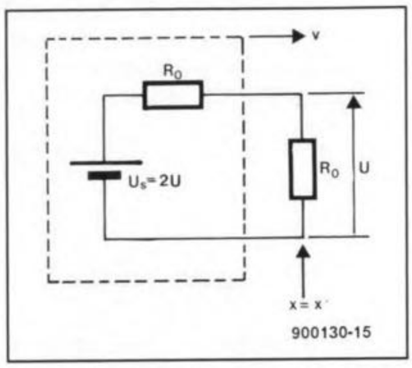

Fig. 5. Sliding source description of step progress.

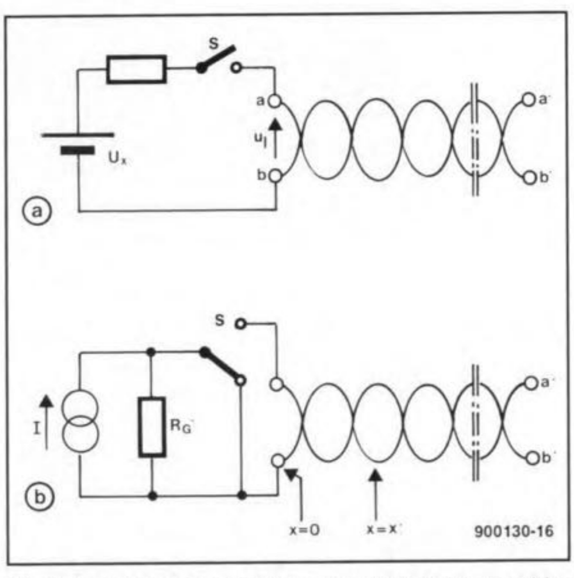

Fig. 6. Applying a step input to a line: (a) voltage-step drive (Sw closes at $t = 0$); (b) current step drive.

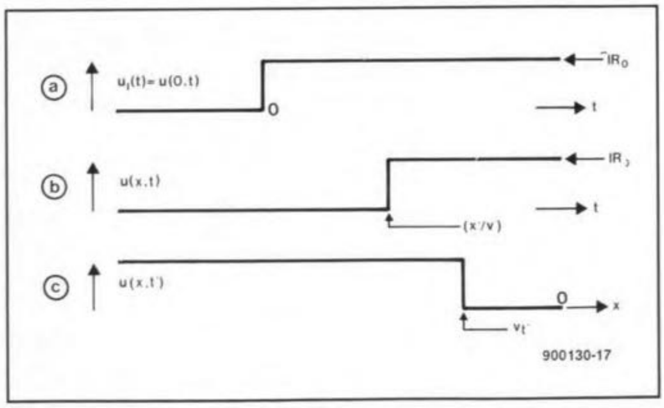

Fig. 7. Voltage sketches for Fig. 6(b).

Kirchhoff's Current Law for Fig. 3(c) gives

$$i = (i + \partial i) + (C\partial x)\{\partial(u + \partial u)/\partial t\}.$$

For $\partial u << u$, a condition always achievable if ∂x is small enough, this reduces to

$$-(\partial i/\partial x) \approx C(\partial u/\partial t) \qquad [2]$$

In the limit case $\partial x \to 0$, $\partial t \to 0$, the approximation sign becomes an equality symbol. We have not proceeded to this limit yet, because the aim is to avoid the distraction of partial differential relationships that arise when a function is dependent on two or more variables. Indeed, combining [1] and [2] to eliminate ∂i or ∂u leads to the (partial differential) 'wave equation' for an ideal line, but such a procedure requires us to solve the equation or, at least, quote solutions for it.

An alternative approach is to show that a voltage step at the input to the line travels along it with constant amplitude and uniform velocity. To do this, we must first establish a relationship between u and i and then derive an expression for step velocity that is independent of x.

u / *i* relationship: characteristic resistance, R_o

Dividing each side of [1] by the corresponding side of [2] gives:

$$(\partial u/\partial i) = (L/C)(\partial i/\partial u),$$

or

$$(\partial u/\partial i)^2 = (L/C).$$

Taking the square root and proceeding to the limit,

$$(du/di) = \sqrt{(L/C)} = R_o, \text{ say.} \qquad [3]$$

We are entitled to express [3] in total differential form because it is valid irrespective of t. Equation [3] gives the limit case for small changes. The ratio is given the symbol R_o, because $\sqrt{(L/C)}$ has the dimensions of resistance. R_o is known as the 'characteristic resistance'. It is characteristic of the line alone and not dependent on the nature of u or i, and is the incremental resistance looking to the right (or left) between terminals p and q, or p' and q'.

The expression 'characteristic impedance' is often used but is unnecessary for a lossless line. It conjures up thoughts of the frequency variable w (or jw) and we are operating here strictly in the time domain.

R_0 is unlike a normal resistor in that it dissipates no power: it is a parameter, dependent on line geometry, that fixes a relationship between the instantaneous changes in i and u either of which can be regarded as a stimulus while the other is regarded as a response. In particular, a step change in u produces a step change in i and vice versa.

Integrating [3], the instantaneous value of u is:

$$u = iR_0 + U_0 \qquad [4]$$

Equation [4] is illustrated in Fig. 4, in which U_0 is any initial line voltage, shown here arbitrarily as positive, and for changes in u and i is accessible only via the series resistor R_0. This accessibility to a line voltage source only via a series resistor R_0 is true also looking to the left at a point on the line, because the line has no built-in directional properties for pulse propagation.

For an initially uncharged line, treated from now on, $U_0 = 0$ and the circuit looking to the right between p and q reduces to the simpler form of Fig. 4(b).

A step voltage of magnitude U appearing at one moment between p and q appears at a later time between p' and q', charging up the line as it progresses with velocity v. There is no loss in amplitude as there are assumed to be no line losses.

Propagation velocity *v*

The propagation velocity, v, is found as follows. Multiplying each side of [1] by the corresponding side of [2]:

$$\{(\partial u)(\partial i) / (\partial x)^2\} = = LC\{(\partial u)(\partial i) / (\partial t)^2\}.$$

Thus, in the limit,

$$v = (dx / dt) = 1 / LC \qquad [5]$$

Since v is independent of x, the velocity is constant along the line. The time, t_u, is

$$t_u = 1/v = \sqrt{(LC)} \qquad [6]$$

Let us check [5] another way. We assume that v is constant and apply the principle of charge conservation. If a step wavefront U travels from $x = 0$ to $x = x'$ in a time $t' = (x'/v)$, the charge supplied to theline by the source is $i(x'/v) = =(U/ / R_0)(x' / v)$. This must equal the charge accumulated by the line capacitance from $x = 0$ to $x = x'$ and this is $Cx'U$. Thus,

$$(Ux' / R_0v) = Cx'U$$

and

$$v = 1/CR_0. \qquad [7]$$

Substituting for R_0 (from [3]) in [7] gives the same value for v as in [5].

Note that R_0 and t_u are two basic parameters required to be known of a line. From these can be found C and L, if needed, with the aid of equations [3] and [6].

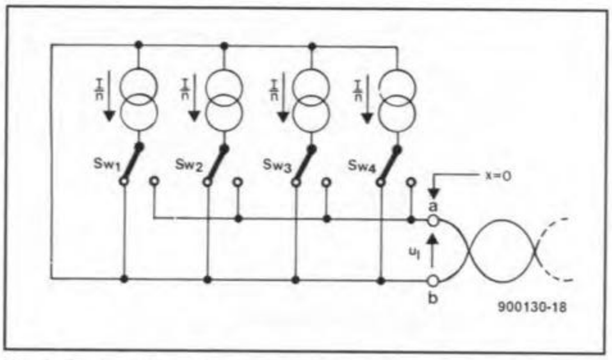

Fig. 8. T.p.l. with multiple step current drive.

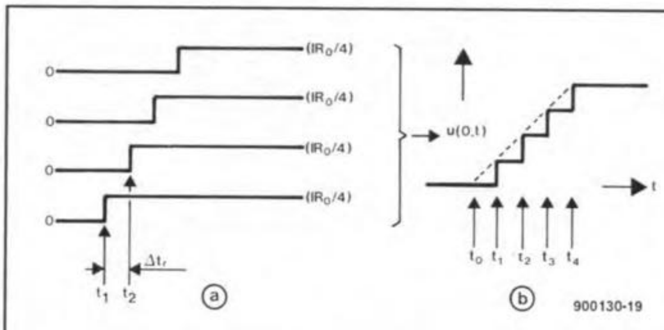

Fig. 9. (a) input voltage contributions for Fig. 8 for $n = 4$; (b) resultant staircase voltage input.

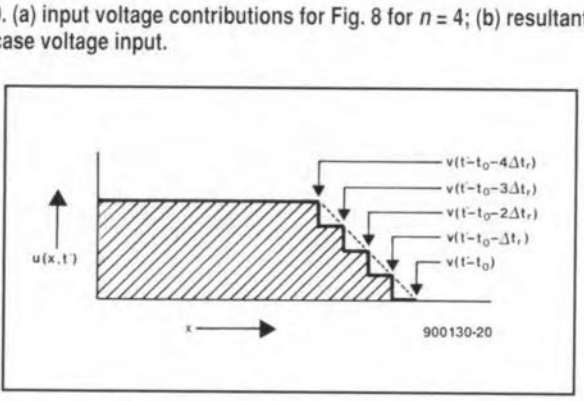

Fig. 10. $v(x;t')$ derived from Fig. 9(b).

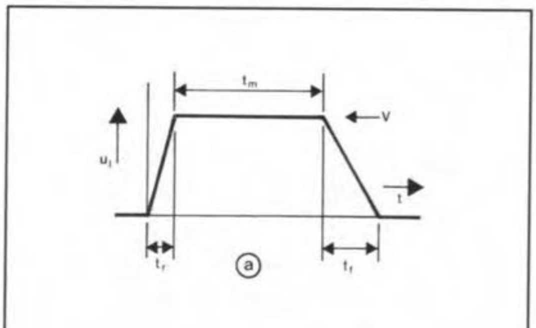

Fig. 11. (a) digital input signal to t.p.l.; (b) $u(x, t)$ derived from (a).

Models for step waveform progress

Progress of a step waveform is so basic that it merits further study. We consider a mechanical analogy and an electric circuit model.

In a mechanical analogy, we may consider a stationary hopper containing sand over a conveyor belt that is moving to the right. At a chosen moment, the exit pipe from the hopper is opened suddenly. The result is a constantly lenghtening, uniform-thickness, trace of sand on the belt. Sand here is, of course, analogous to electric charge.

A model for step progress attractive to the engineer more at home with lumped circuit theory involves the concept of a 'sliding source'. Consider the progress of a step voltage wavefront of magnitude U along an initially uncharged line in the direction of increasing x. Looking to the right at any point x', the remainder of the line appears as a resistor R_0 as shown in Fig. 5. Looking backwards, towards $x = 0$, the line appears as a voltage source U_s accessible via a source resistor R_0 (inside dashed rectangle), both of which appear to slide along the line with velocity v. To produce a step of magnitude U at $x = x'$, it is obviously necessary that $U_s = 2U$.

This sliding source approach is helpful in calculating what happens at the end of a line of finite length l.

Line voltage *u* (*x*,*t*): step and pulse drive

To investigate, experimentally, a t.p.l. subjected to a step input, the line can be 'voltage-driven' or 'current-driven' as shown in Fig. 6. In both cases, the condition of switch S is assumed to change at $t = 0$.

Simple experimental predictions of terminal voltage behaviour based on Fig. 6(a) require a knowledge of R_0 and the certainty of its constancy over the range of the output voltage swing. It is not possible to guarantee constancy using standard saturated transistor logic circuits (e.g. TTL) to voltage-drive the line.

In the current-drive scheme of Fig. 6(b), the output resistance, R_g, is generally much greater than R_0 and can be ignored by comparison with it. This is the case, in practice, with a switched long-tail pair driver stage. $u(x, t)$ denotes line voltage as a function of vartiables x and t. Of special interest are: $u(0, t)$, the variation with t at $x = 0$, i.e., the input waveform, $u_I(t)$; $u(x', t)$ the waveform at an arbitrary point $x = x'$; $u(x, t')$, a plot of line voltage as a function of x at a specific time $t = t'$.

In Fig. 7(a), $u_I(t)$ is a step of magnitude IR_0 because the line appears initially, at its input terminals, as a pure resistance R_0.

$u(x', t)$ in Fig. 7(b) is $u(0,t)$ delayed by a time interval (x'/v): the line is uncharged at x' till the step reaches that point. If the switching action occurs at $t = t_0$, the line is charged up to the point $v(t'-t_0)$.

Unlike $u_I(t)$ and $u(x',t)$, which can be monitored, $u(x,t')$ is not a waveform. However, if we choose an appropriate scale on the paper as in Fig. 7(c), we can make the graph appear complementary to that of Fig. 7(b). This means that the sum of ordinates of the two graphs, at a given point on the horizontal axis, gives a constant value. This scale changing 'trick' is useful in deriving $u(x,t')$ from $u_I(x',t)$ for the general case of a line signal that is not a step, as we will show now.

In Fig. 8, n current sources each of strength I/n are connected to a t.p.l. via switces S1–Sn.

S1 changes state at $t = (t_0 + \Delta t_r)$, where $\Delta t_r = \{(t_n - t_0) / n\}$, and pumps a current I/n into the line. This is followed at successive time intervals Δt_r by S2–Sn, respectively, causing additional current steps I/n to be applied in sequence to the line input.

Equation [4] specifies a linear relationship between u and i, so the Principle of Superposition is applicable and we can add algebraically the effects of each input taken separately to obtain the overall response.

The resulting waveform for $u_I(t)$ is a voltage 'staircase', which, for $n = 4$, is shown in Fig. 9(b). The dashed line joining the edges of the treads intersects the t-axis at $t = t_0$. Suppose now that instead of $n = 4$ we let $n \rightarrow \infty$. The staircase edge then assumes the profile of the dashed line in Fig. 9(b) and Fig. 10. We have thus deduced $u(x, t')$ for a ramp input. The general case for an input of arbitrary shape is worked out similarly by considering steps of unequal magnitude and—if necessary—opposite polarity when the switches in Fig. 8 change state.

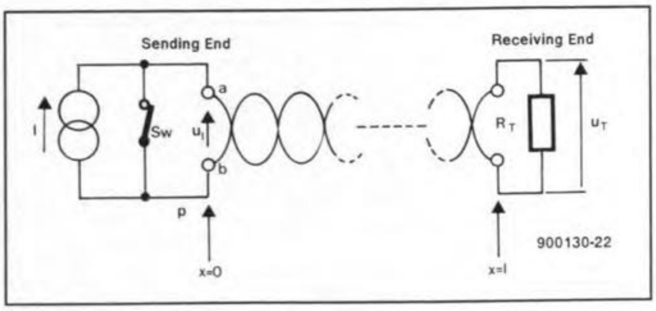

Fig. 12. Current-driven terminated line.

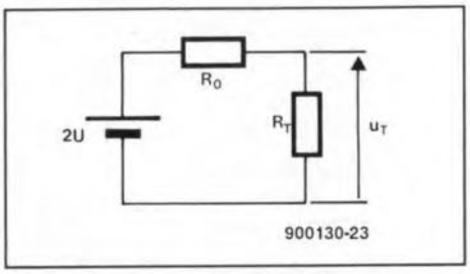

Fig. 13. Calculation of terminal voltage at $t = t_d$.

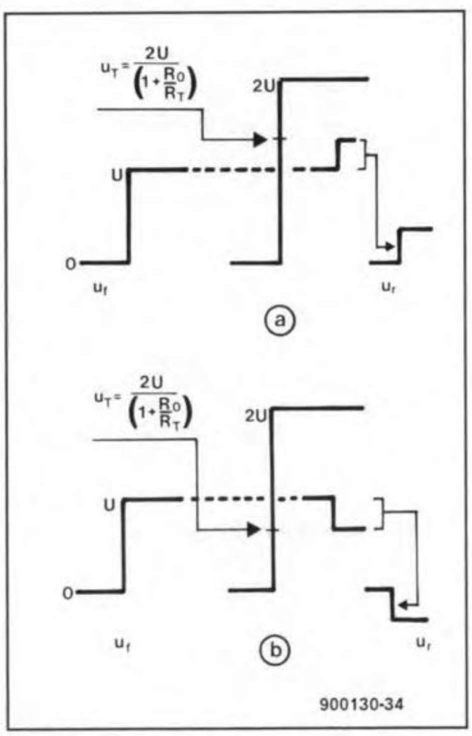

Fig. 14. Generation of v_r for $R_T \neq R_0$: (a) $R_t > R_0$; (b) $R_T < R_0$.

Thus, the digital input signal of Fig. 11(a), with transition times t_r and t_f purposely chosen unequal, produces $u(x, t')$ in Fig. 11(b). Figure 11(c) may be regarded as a scaled mirror image of Fig. 11(a) displaced along the horizontal axis. An alternative graphical method for obtaining $u(x, t')$ from $u(0, t)$ is given in the reference at the end of this article.

Reflections

It is convenient to imagine a semi-infinite line, stretching from $x = 0$ to $x = \infty$, in an initial discussion of lines because it simplifies the presentation. However, once the progress of a step wavefront is understood we can consider what happens at the end of a line of finite length l when a pulse edge or a pulse of arbitrary shape reaches it.

Consider the scheme shown in Fig. 12, where R_T is a terminating resistor. A voltage wavefront u_f of amplitude $U = IR_0$, which we call the forward wavefront, starts down the line at $t = 0$ when S opens. It reaches the end of the line in the one-way delay time $t_d = l / v = lt_u$.

The terminal voltage u_T at $t = t_d$ is calculated from the sliding source equivalent circuit of Fig. 13: $u_T(t) = 2UR_T / (R_T + R_0)$. Now, $u_T = u_f$ (the terminal voltage step is equal to the amplitude of the forward voltage wavefront on the line) if $R_T = R_0$.

This is the case of a line 'matched' or 'correctly terminated' at the receiving end. Then R_T dissipates energy at the same rate as it is supplied to the line from the source. No energy is reflected, that is, sent back to the source. As far as any effect on the sending end is concerned, the line may just as well be considered semi-infinite despite its actual finite length. There is an analogy here in radar. If the energy in a radar beam is completely absorbed by a target, there is no reflection, that is, the target is 'invisible'. As far as the radar receiving equipment is concerned, the target may be regarded as located at a point an infinite distance away. Suppose, however, that $R_T \neq R_0$. Then, $u_T \neq u_f$, all the energy associated with u_f cannot be absorbed by R_T, and a reflected wavefront u_r is pro-

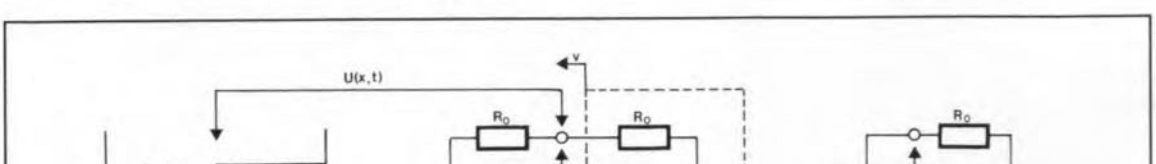

Fig. 15. (a) line voltage for $2t_d \geq t > t_d$; (b) sliding circuit equivalent circuit form for (a); (c) circuit for calculating $v_i(2t_d)$.

duced. The amplitude and polarity of u_r must be such that the Principle of Superposition is applicable at the termination. Thus,

$$u_f + u_r = u_T,$$

or

$$u_r = u_T - u_f. \quad [9]$$

Substituting for u_T from [8] and $u_f = U$, gives

$$u_r = [\{2UR_T/(R_T+R_o)\} - U] = \rho_{VT}U \quad [10]$$

where ρ_{VT} is the voltage reflection coefficient at the termination and is defined by

$$\rho_{VT} \equiv (R_T - R_o)/(R_T + R_o) \quad [11]$$

Figure 14 shows a geometrical construction giving u_r for the cases:

(a) $R_T > R_o$, and hence $\rho_{VT} > 0$, and

(b) $R_T < R_o$, and hence $\rho_{VT} < 0$.

For either condition, the reflected voltage wavefront travels back to the source.

A plot of line voltage for $2t_d \geq t > t_d$ is shown in Fig. 15(a) for $\rho_{VT} = 0$. This results from adding u_r to the existing line voltage giving a total line voltage $(1 + \rho_{VT})U$ at the position of the wavefront.

The total line voltage is also obtained from the sliding source equivalent circuit which, in this case, comprises a generator $2\rho_{VT}U$ in series with an output resistance R_o as shown in Fig. 15(b):

$$u(2t_d) = IR_o + 2\rho_{VT}U = U(1 + 2\rho_{VT}). \quad [12]$$

Since there is already a line voltage U and $u_r = \rho_{VT}U$, this means a further forward, reflected wavefront of amplitude $\rho_{VT}U$. This also follows from [11] since the voltage reflection coefficient is unity for an ideal current source.

The current-driven line of Fig. 12 with $R_T \neq R_o$ is of restricted use. Two cases of reflection of practical interest for a current-driven line with an intentional mismatch at the receiving end are considered next.

With reference to Fig. 16, in which a shunt matching resistor is incorporated at the sending end, the two cases correspond to $R_T = 0$ and $R_T = \infty$.

Consider first the case $R_T = 0$. Writing U for $I'R_o/2$, it follows that $u_1(0+) = U$. From [11], $\rho_{VT} = -1$. The equivalent circuit for calculating $u_1(2t_d)$, and $u(x)$ for $t > 2t_d$ is shown in Fig. 17.

In Fig. 18, $u_1(t)$ is a pulse of amplitude U and duration $2t_d$. The line input current, i_1, and the energy supplied by the source, W_s, are shown in Fig. 18(b) and Fig. 18(c) respectively.

An argument based on the Principle of Conservation of Energy leads to an algebraic expression for t_u. Thus,

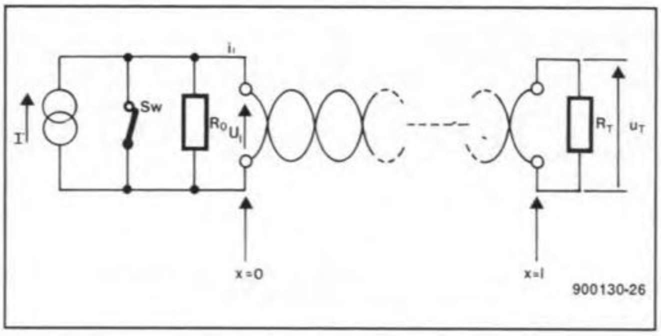

Fig. 16. Current-driven line, matched at the sending end.

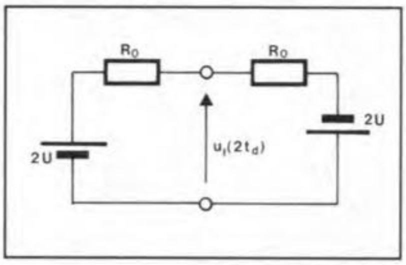

Fig. 17. Equivalent circuit for calculating $v_1(2t_d)$ for $R_T = 0$.

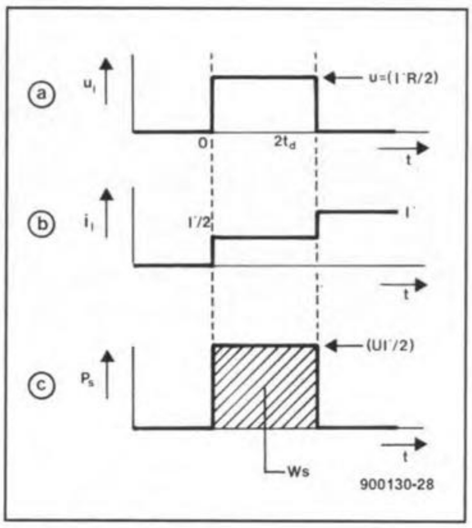

Fig. 18. For $R_T = 0$, the energy, W_s, supplied to the line by the source (c) is obtained by multiplying graphs (a) and (b).

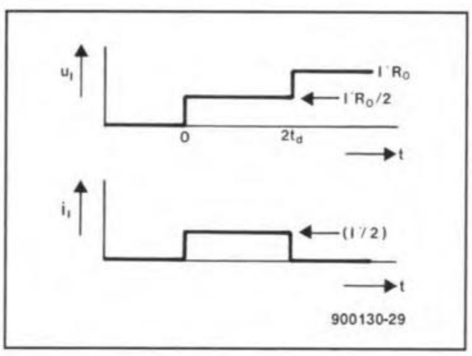

Fig. 19. $v_1(t)$, $i_1(t)$ for the case $R_T = \infty$.

$$W_s = (I'U/2)2t_d = I'U = (I')^2R_o/2 \quad [13]$$

At $t = 2t_d$, the line stores no energy in its electric field since at that time $u_1 = 0$. All the energy, W_m, is stored in the magnetic field:

$$W_m = Ll(I')^2/2 \quad [14]$$

Equating W_m and W_s yields:

$$t_u = (t_d/l) = (L/R_o) \quad [15]$$

However, $R_o = \sqrt{(L/C)}$, so that

$$t_u = \sqrt{(LC)} \quad [16]$$

as previously shown in [6].

With reference to Fig. 16, the case $R_T = \infty$ gives the waveforms for $u_1(t)$ and $i_1(t)$ in Fig. 19.

Conclusion

This article has dealt in detail with some aspects of line pulse operation that are either ignored or skimpily covered in the literature.

Reference:

Digital Signal Transmission: Line Circuit Technology by B.L. Hart, Van Nostrand Reinhold (UK), 1988 (Chapter 3). ■

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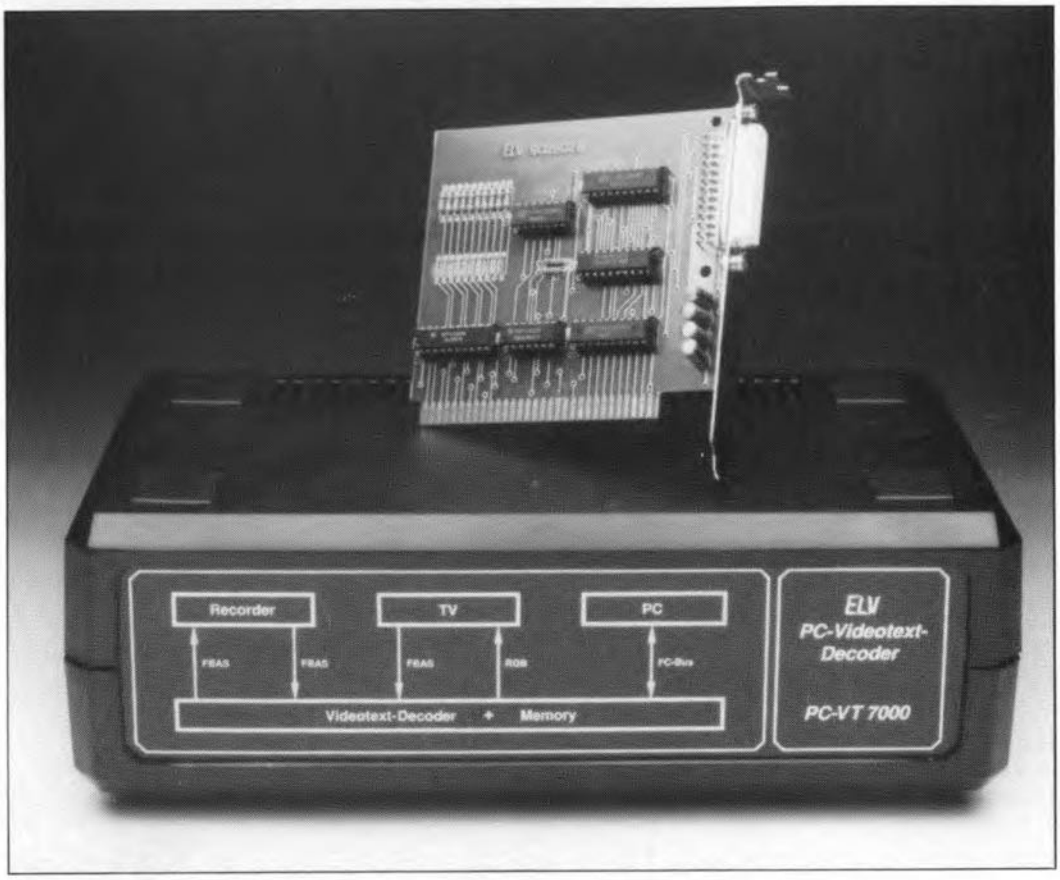
PC-CONTROLLED VIDEOTEXT DECODER PC-VT7000

PART 1: INTRODUCTION AND DESCRIPTION OF THE DECODER



This Videotext decoder, designed and marketed as a kit by ELV, allows the decoding and storage of Videotext (or Teletext) pages on an IBM PC or compatible. Among the special features of the PC-VT7000 are fast access to subpages, the possibility of using a video recorder for separate processing of subtitles (particularly useful for the deaf and hard of hearing), and the use of a SCART-compatible TV set for displaying the decoded pages.

Videotext, Teletext, CEEFAX and Oracle are but a few names given by broadcasters to a special information service transmitted during the blanking period of TV signals. The information is brought to the viewer via pages of text and graphics, which can be called up by entering the appropriate number on the remote control of the TV set. Among the subjects in the Videotext service are news items, sports, weather information and TV programme overviews. In most cases, the pages are updated by the broadcaster's editorial staff for the Videotext service.

In the PAL TV system, 625 TV lines are transmitted as two interlaced fields of 312.5 lines each. About 50 of these lines fall inside the vertical retrace (or blanking) period, which is not normally visible on the TV screen. These 50 lines are used to convey test signals and digital information (see also Ref. 1).

Teletext is usually conveyed via lines 11 to 14, and 20 and 21, in the first field, and 324 to 327, and 333 and 334, in the second field. At a field frequency of 50 Hz, the maximum text line rate is about 300 per second, corresponding to about 12 pages per second.

To be able to receive Videotext pages, you need a special decoder. Most modern TV sets, and even some of the latest video recorders, have such a decoder as a built-in unit. Where a decoder is not part of the TV, it may often be purchased and installed as an upgrade.

After entering the requested Videotext page number on the remote control, the decoder starts to search for it. The search process is indicated by the three-digit page counter in the upper left-hand corner of the TV screen. When the page is found, the search process stops, and the relevant information is shown on the screen. Unfortunately, finding a particular page may take quite some time — depending on the reception conditions and the number of pages in the service, wait times of up to 10 s are not uncommon. Particularly when frequent use is made of Videotext pages, the long wait time before they are available is a real disadvantage of an otherwise extremely useful information service.

The PC-VT7000 has a number of advantages over a conventional Videotext decoder built into a TV set. To use the unit, you require either a video recorder with a CVBS (chrominance-video-blanking-synchronization, also called composite video) output, or a TV set (with or without a Teletext decoder) with a SCART socket. The CVBS signal taken from this socket is fed to the PC-VT7000. After decoding and processing, the Videotext pages may be displayed either on the TV set (which takes in the video signal via the SCART socket), or on the monitor of the PC. A video recorder may be connected to the second SCART socket on the PC-VT7000 to enable Videotext pages as well as TV pictures with subtitles to be recorded. The latter option is of particular interest to the deaf and the hard of hearing.

A further special feature of the decoder is its ability to produce hard copy of Videotext pages on a printer. By using this option you are in a position to print out, say, the day's programme overview, or the current weather situation (which consists of charts and tables). The final advantage of the PC-VT7000 over a conventional Teletext decoder is that it enables you to have immediate access to subpages. Most conventional Teletext decoders allow you to enter the main page only. To view the subpages that belong with this main page, you have to sit and wait for the decoder to show them one after the other. Normally, a subpage is shown 10 seconds or so before the next appears. There is, however, no way to skip subpages to get at the one you do want to read. Many Videotext users find this irritating and a waste of time. The PC-VT7000 has a special page memory that solves this problem by offering you immediate access to any subpage.

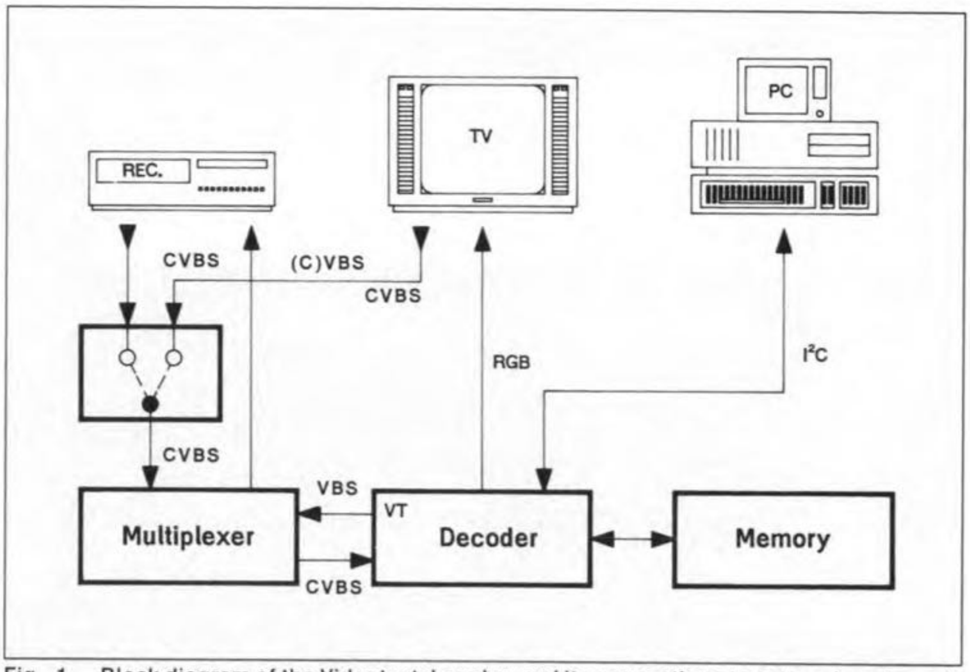


Fig. 1. Block diagram of the Videotext decoder, and its connections to external equipment.

Connecting the decoder

The PC-VT7000 consists of two units: (1) an insertion card for PCs that provides an IC bus interface, and (2) the decoder proper.

The inputs and outputs of the decoder are found on the rear panel of the ELV 7000-series enclosure. These inputs and outputs are used to connect the PC insertion card and the video equipment.

The minimum equipment to run the system is a TV set with a SCART socket, which must be connected to the PC-VT7000 via a SCART cable. One of the pins on the SCART socket of the TV supplies the composite video signal, which is used by the decoder to extract the Videotext information. This information is processed and turned into a video signal that is fed back into the TV set, again via the SCART connection, which thus functions as a bidirectional link.

The second SCART socket on the rear panel of the PC-VT7000 allows a video recorder to be connected. The special use of the VCR for recording TV programmes with a subtitling service has already been mentioned.

The toggle switch on the rear panel is used to select either the TV set or the video recorder as the source of the CVBS input signal for the decoder. When this switch is set to TV, the TV set must be switched on — otherwise, the tuner can not supply a CVBS signal to the decoder. For the same reason, when the switch is set to the other position, the VCR must be on or in stand-by mode, i.e., its tuner must supply a video signal that contains Videotext information. To enable the PC-VT7000 to store Videotext pages on the PC, or programmes with subtitling on the VCR, the input source switch must be set to TV.

There is one special equipment configuration in which a TV set is not required: when a VCR is used as the CVBS signal source, and a computer screen only to display the Videotext pages.

The IC insertion card is powered by the PC. The cable between this interface card and the decoder also carries the required supply voltage, so that a separate power supply is not required to use the system.

The hardware: an overview

The block diagram in Fig. 1 shows the way in which the previously discussed units are interconnected. The heart of the circuit is formed by the Videotext decoder, which communicates with the other units via a two-way multiplexer that forms part of the main decoder. The CVBS signal that contains the Videotext information of the relevant broadcaster is supplied by the tuner in the TV set. As already mentioned, the decoder may also accept the CVBS signal of a video recorder, provided this is not used to play a tape. Unfortunately, owing to their limited bandwidth and recording method, very few videorecorders are capable of reproducing a usable Videotext signal from tape. However, the VCR is perfect for recording and reproducing decoded Videotext pages and programmes with superimposed subtitles.

The CVBS signal applied to the decoder is analysed to extract and store the TV lines that contain Videotext information (see also Ref. 1). Next, the information is either sent to the PC via the IC bus (see Ref. 2), or fed to the display controller which uses it to build a complete picture that can be displayed on the TV screen.

The system also allows decoded Videotext pages as well as subtitles with the current programme to be recorded (on the VCR) or stored (on the PC). It should be noted that the stored Videotext pages and the subtitles are displayed in black and white, while the VCR recordings are, of course, in colour.

Control program

The functions of the PC-VT7000 are controlled from a PC running a special program, loaded from floppy disk or hard disk. This program is called up by typing **VT** followed by a carriage return. The program automatically prompts the Videotext decoder to search and display page 100 on the PC monitor or the TV screen. The command to do so is issued by the PC insertion card and sent to the decoder via the I²C bus. Page 100 is provided as a default value: by simple programming, the software can be changed to load any other page on starting the system.

Although the control program supplied with the PC-VT7000 is largely self-explanatory, a help function giving details of all essential actions may be called up at any time by pressing function key **F1**.

The three-digit number of the requested page is entered on the PC keyboard. A page selection window appears on the PC screen when the first digit is typed. Since the page number invariably consists of three digits, the **CR** key need not be pressed when the number is complete. The requested page is displayed as soon as it is found in the Videotext datastream. If the page is not active, or can not be found, a message appears after a short while.

As already mentioned, the PC-VT7000 offers a fast way of calling up subpages. After

Fig. 2. Example of a Teletext screen (BBC TV Europe programme via satellite).

loading the main page, simply press the ↑ or the ↓ key to leaf through the subpages. The action on part of the Videotext decoder is virtually immediate. The other two arrow keys, ← and →, are used to leaf through the main pages. Since the main pages are not stored sequentially, this may take more time than with subpages.

The current Videotext page may be sent to a printer by pressing **F2**.

Function key **F3** allows the currently displayed page to be converted into a data file. After pressing **F3** you are prompted to enter comment which helps you identify the page when it is retrieved later. The program automatically assigns the page number to the file as an identifier.

Videotext data files may be retrieved by pressing **F4**. You are prompted to enter the number of the requested page, which is subsequently loaded from disk and displayed on the PC screen. Note that this function is available even when neither the IC insertion card, the main decoder, nor the TV set are connected.

The list of Videotext datafiles stored on the computer, along with the associated comment, may be called up by pressing the **F5** key.

Finally, the control program may be terminated by pressing the **ESC** key.

The Videotext decoder

The circuit diagram of the decoder is given in Fig. 4. As already mentioned, this circuit is fitted into a series-7000 type enclosure. The CVBS signal is applied to the decoder either via pin 20 of SCART socket BU$_1$ (TV set), or via pin 20 of BU$_2$ (video recorder). The CVBS signal supplied by the TV set is terminated with resistor R$_1$, and applied to pins 4 and 11 of analogue multiplexer IC$_5$. When the CVBS signal from the VCR is used, this is allied to pins 2 and 15 of the same IC. The required terminating resistance is then formed by the TV set, whose CVBS input is connected to pin 19 of socket BU$_1$. In case a TV set is not connected, switch S$_1$ must be set to the upper position to allow R$_2$ to function as a terminating resistance.

The SCART sockets, BU$_1$ and BU$_2$, are wired in a manner that allows the video recorder and the TV set to be used for recording programmes and playing back tapes just as if the PC-VT7000 were not connected, and without having to change any cable or connection.

The CVBS signal with the Videotext information in its vertical blanking period is applied to electronic switch (multiplexer) IC$_5$. Depending on the source selection (Rec/TV) set with S$_2$, either the CVBS signal from the TV set, or that from the tuner inside the VCR, is routed to the parallel-connected IC outputs, pins 3 and 13. This is achieved by S$_2$ determining the logic level at address selection input A of the 4052.

SAA5231 VIP2

The composite video signal arrives at the input of the video processor, IC$_1$ (SAA5231) via coupling capacitor C$_6$. The SAA5231 VIP (Video Interface Processor) extracts the Videotext information from the data carried in the previously mentioned TV lines in the vertical blanking interval. The block diagram of the SAA5231, which is manufactured by Philips Components, is given in Fig. 3. Its tasks include:

- separating and regenerating the Videotext information;
- generating a clock signal that is synchronous to the current picture;
- supplying data to the display controller that follows it;
- extracting the synchronization components from the CVBS signal;
- supplying the synchronization components to the display controller that follows it;
- switching to internally generated synchronization when the external synchronization fails;
- supplying the synchronization components at positive and negative polarity;
- locking the internal 13.5-MHz quartz-controlled oscillator to the applied CVBS signal;
- adjusting itself to the level of the applied CVBS signal.

The CVBS signal applied to pin 27 of IC$_1$ is fed to an internal adaptive data separation stage with a slicing level of 50% of the CVBS signal amplitude. The slicing level is set to 50% to achieve the highest possible noise immunity. The 8-bit data supplied by the VIP2 consists of 7 databits and 1 parity bit.

As shown in Fig. 3, the CVBS signal is also fed to the input of an adaptive sync separation circuit. The slicing level of this circuit is adjusted automatically as a function of the input amplitude. This is done to compensate low-frequency level variations.

The VIP2 supplies the Videotext data and the associated clock pulses at output pins 15 and 14 respectively, for use by the display controller that follows it.

The output frequency of the 6-MHz VCO (voltage-controlled oscillator) on board the VIP2 is controlled via a phase detector, with the aid of a line-frequency clock signal at pin 28. This is done to ensure that the generated Videotext characters are synchronized to the current picture as required for the subtitling service. The synchronization pulses obtained from the input video signal are also applied to the phase detector. The 6-MHz clock, which is phase-locked to the sync pulses, is coupled out via pin 17 and applied to the relevant input, pin 9, of the display controller, a SAA5243.

Pin 28 of IC$_1$ accepts the composite synchronization signal generated in the SAA5243. When the synchronization signal at the CVBS (TV programme) input of the VIP2 fails, this chip automatically switches to the replacement sync signal furnished by

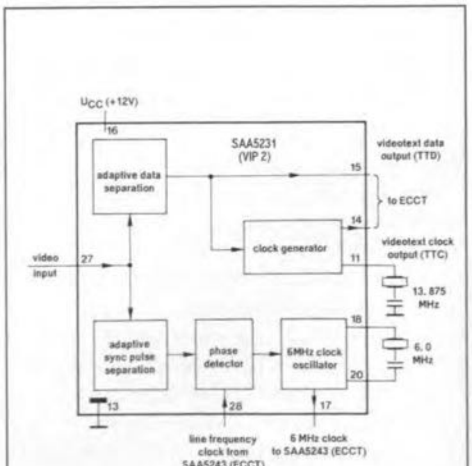


Fig. 3. Block diagram of the SAA5231 Video Interface Processsor (VIP2).

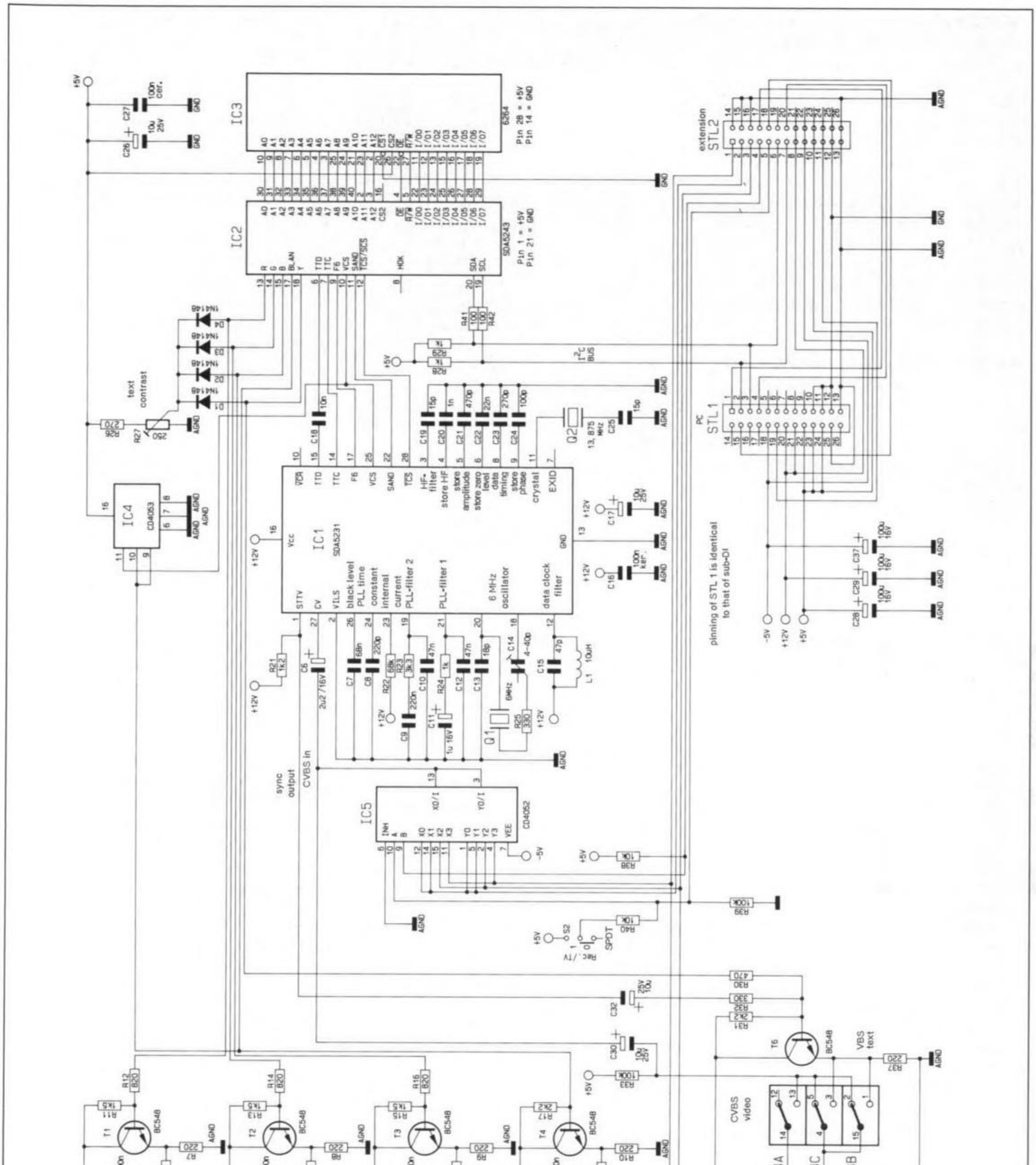


Fig. 4. Circuit diagram of the VideoText decoder. Note that this ciruit is controlled via an I²C interface connected to PCB header STL1.

the ECCT. The ECCT also generates the sandcastle pulse, which is fed to the VIP2 for use in the Videotext data slicer.

Capacitor C15 and inductor L1 form the external components required to make the 6.938 MHz data clock filter operate. Similarly, quartz crystal Q2 and capacitor C25 enable the 13.875 MHz oscillator to operate.

Pin 1 of the VIP2 supplies the composite-sync signal for the TV set. Resistor R12 sets the polarity of this signal to positive by pulling it to the +12-V supply rail. The sync-locked 6-MHz oscillator operates with external components Q1, C13, C14 and R25. Trimmer C14 allows the synchronization to be adjusted.

SAA5234 (ECCT) and page memory

The full identification of the SAA5243, another Philips Components IC, is Enhanced Computer-Controlled Teletext Chip, which is mercifully abbreviated to ECCT. Together with a RAM Type 6264 and the VIP2, the ECCT forms the heart of the present decoder. It should be noted that the VIP2 and the ECCT are also available under the respective type numbers SDA5231 and SDA5243 from 'second source' Siemens.

As shown in the block diagram in Fig. 5, the ECCT contains a character generator, a data acquisition circuit, an I^{2}C interface, a clock driver and a memory interface. These standard functions are boosted by the following extras:

- an integrated character generator with 160 alphanumeric and 2×64 graphical characters, each built in a 12 (H) by 10 (V) matrix;
- user-controlled double-height characters for the upper or lower half of the Videotext page;
- insertion of all characters and colours via commands on the IC control bus;
- the current character position may be identified with a cursor;
- status information above or below the main text (line 25);
- automatic switching of the character set to one of six languages by special control bits in the page header;
- simultaneous searching process for up to four pages;
- data capture in all lines of the frame (full-channel mode), offering fast page access.

The clock driver in the ECCT communicates with the VIP2 video processor via pins 9 to 12. After checking their validity, the ECCT accepts the data and clock signals received from the VIP2. These data are written to the external page memory RAM, a 6264, via the memory interface. The data acquisition is organized such that four Videotext pages can be searched for, and stored in RAM, simultaneously. Thus, these four pages are updated at the same time. The page memory is accessed with the aid of signals $\overline{\text{OE}}$ (output enable) and R/$\overline{\text{W}}$ (read/write). Data is carried via pins 22 to 29, and addresses via pins 2, 3 and 30 to 40.

The ECCT supplies the picture information via its three colour output pins, R, G and

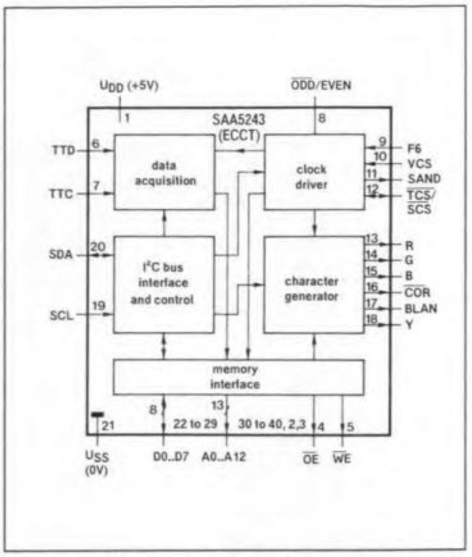


Fig. 5. Block diagram of the SAA5243 Enhanced Computer-Controlled Teletext chip (ECCT).

B. The character generator has 256 characters, a selection of which is available in each of the six national character sets that can be called up by an appropriate software command.

The blanking signal for use with the RGB components is available at pin 17 of the ECCT. This signal is used during mixed picture operation as, for instance, Videotext subtitling. The system has two modes of operation, which are selected by software:

- character insertion (superimpose)
- background suppression

The Y-signal (luminance or brightness) is provided independently of the selected colour at pin 2 of the ECCT, and is thus only valid for the Teletext characters. A flash function is not provided as standard.

Output circuits

The R, G, B, blanking and Y outputs are of the open-drain type, and require external pull-up resistors. Resistors R11-R12 are fitted at the R output, R13-R14 at the G output, and R15-R16 at the B output. The ratios of these resistors determine the signal level at the base of the associated RGB transistor driver stage.

Preset R27 enables the output level of the RGB drivers and that of the Y output to be set to a value that produces optimum contrast of the Videotext characters in relation to the TV picture.

The four outputs are decoupled by diodes D1 to D4. The drivers for the RGB and blanking signals are built around four transistors in common-collector circuits, T1 to T4. Resistors R3 to R6 determine the output impedance and ensure optimum signal matching to the loads formed by the TV inputs. The signals are fed out of the circuit via the SCART socket for the TV set.

VCR output

As already mentioned, the PC-VT7000 offers the user the possibility of recording Videotext subtitles on a VCR. This works as follows. The Y signal at pin 18 of the ECCT is fed to the base of emitter follower T6 via R30. The composite-sync signal is added to the video via R32. Capacitor C32 provides the necessary d.c. decoupling. The combination of R30 and R31 forms the pull-up resistor at the open-drain Y output, pin 18, of the ECCT.

The buffered VBS (monochrome) signal at the emitter of T6 is fed to the inputs, pins 1 and 3, of electronic switches IC4B and IC4C. Each second input of these switches, pins 2 and 5, has on it the CVBS signal (the original TV picture). This means that the system can switch between these signals. To make sure that the CVBS signal is at the right level, it is fed, via C30, to a clamping circuit composed of IC4A, R33, R35, R36 and C31. Since the positive sync pulses supplied by the VIP2 control the electronic switches, the CVBS input signal is clamped at a potential fixed by R35-R36. This ensures the correct d.c. levels at the second inputs, pins 2 and 5, of the electronic switches.

The control of electronic switches IC4B and IC4C is determined by the blanking signal. The relevant output, pin 15, allows one of three signal configuration to be selected:

- the original composite video signal;
- the Videotext image;
- a mixture of these (superimpose).

When the third configuration is used, the output supplies a signal composed of the CVBS TV signal and the VBS Videotext signal. This mixed signal is fed to a buffer, T5, via coupling capacitor C5. The buffered signal is taken from the emitter of T5, and fed to the video recorder input via pin 19 of the relevant SCART socket.

Interface to I^{2}C card

The connection marked STL1 links the Videotext decoder to the PC. This connection carries the supply voltages for the decoder board, and the data.

All functions of the Videotext decoder are controlled via the IC bus interface, pins 19 and 20, of the ECCT. The relevant control signals are conveyed via the IC interface card in the PC. As already mentioned, this card forms part of the project.

Finally, connector ST12 carries a number of control and data signals that may be used for future extensions.

Next month's second and final instalment of this article will deal with the operation of the I^{2}C card, and the construction.

MILLIOHMMETER

As you are probably aware, measuring small resistance values is difficult, if not impossible, with conventional digital and analogue multimeters. While only a few of these instruments have a 1-Ω range with limited practical use, the meter presented here allows very small resistances in the range from 10 mΩ to 5 Ω to be measured reliably.

A. Rigby

That most multimeters have a lowest resistance range of 100 Ω or 1 kΩ is not surprising. The measurement of small resistances poses a number of special problems that do not occur in the kΩ ranges. Take, for instance, the measurement system, which in many cases has to be changed just for the sake of the lowest range. There is, however, a more serious problem in the range up to 10 Ω: the contact resistance of the test lead plugs and the sockets on the instrument, and, of course, the resistance of the test leads themselves. A connection formed by a banana plug and a mating socket, both in new condition, represents a typical resistance smaller than 1 mΩ. This resistance rises to several milliohms as the contact surfaces start to oxidize. Although a few mΩ may not seem much to start worrying about, such values are significant since the instrument discussed here has a resolution of 2 mΩ. The resistance of the test leads is also a factor of some importance. A test lead with a length of 1 m and a cross-sectional area of 1 mm^2 has a typical resistance of 17 mΩ. For a similar lead with a cross-sectional area of 2.5 mm^2, this value becomes 7 mΩ. Relating these values to 1 Ω, the error factors are 1.7% and 0.7% respectively. In other words, our measurement starts to become unreliable when these parasitic resistances are not taken into account. Fortunately, there exists a measurement principle that eliminates the effects of these unwanted resistances. This principle is called four-point resistance measurement.

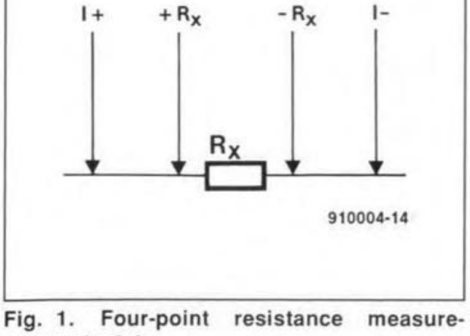


Fig. 1. Four-point resistance measurement principle.

Two terminals, four wires?

Using four wires to connect a resistor with only two terminals to a meter system may seem strange at first. However, since these wires may be divided into two pairs with the same functions, this method allows us to eliminate the effects of parasitic resistances. The principle is illustrated in Fig. 1. The unknown resistor, *R*, is connected with four wires. The outer two cause a current flow through *R*. The present meter sends a constant current through *R* via terminals I+ and I–. The advantage of using a constant-current source is that it is not affected by the parasitic resistance. Hence, we know exactly how much current flows through *R*. To determine the value of *R*, all we have to do is

MAIN FEATURES

- **Ranges:** 100 mΩ, 200 mΩ, 500 mΩ, 1 Ω, 2 Ω, 5 Ω
- **Resolution:** 2% of f.s.d. value
- **Principle:** 4-point measurement with pulsed constant current
- **Measurement current:** I_p = 1 A; I_{rms} = 10 mA; pulse length approx. 1 ms; repeat rate approx. 10 Hz.
- **error detection:** too low test current
- **current consumption:** max. 70 mA

measure the voltage across it as a result of the constant current. This voltage is fed to the instrument via wires +Rx and –Rx. These wires are connected as close as possible to the resistor body, or to the terminals to which a resistor is to be connected later. In this way, only the voltage drop across the resistor is measured, without the additional voltage across all kinds of parasitic resistances. The system also eliminates the resistance of the test leads, and the contact resistance at the plugs and sockets.

Since the current flow into the voltage meter is negligible with respect to the constant current sent through the resistor under test, it may be concluded at this point that the four-point resistance measurement offers a reliable method of determining the value of small resistors at an accuracy that is not normally achievable with a multimeter.

1 A, and no heat?

Good as the four-point measurement system may be as a basis for the design of a milliohmmeter, there are more aspects to such an instrument that need to be given thought. Among these factors is the heat dissipated by the resistor. To make sure that a low-value resistor produces a voltage drop that is readily measured, it must pass a relatively high current. We can not make the current as high as we wish, however, since the maximum permissible dissipation of the resistor must be taken into account. A 1-Ω resistor with a power rating of 0.25 W, for instance, will not survive the constant current of 1 A supplied by the instrument. The solution to this problem is found in the use of a pulsed constant-current source (see the block diagram in Fig. 2). The resistor under test is fed with an effective current of only 10 mA since the 1-A current source is pulsed at a duty factor of 0.01 (1 ms on, 100 ms off). Even a 0.25-watt resistor will not mind such a low effective current. Unfortunately, the use of a pulsed test current has one disadvantage in that resistors with a relatively high reactive component (stray inductance or capacitance) can not be measured reliably.

The test current through the resistor is pulse-shaped because the constant-current source is switched on and off by a pulse generator. The same generator controls a sample-and-hold circuit that stores the measured voltage during the 'off' period of the current. This means that the output of the sample-and-hold supplies a constant voltage whose value is in direct proportion to the measured resistance. Depending on the selected range, this voltage is amplified or attenuated before it is fed to a moving-coil meter provided with an ohm scale.

The circuit helps you avoid measurement errors by signalling over-range conditions. This is achieved by monitoring the output current of the current source. When a too large resistor is connected, or when the current wires, I+ and I–, are broken, the current source will no longer be able to supply 1 A, so that the voltage measured across the resistor is no longer a direct measure for the resistance value. However, the meter will still indicate 'something' because the measurement circuit and the resistor supply are separate circuits. The fault condition is simple to recognize because the current source then pulls terminal I– to ground. A detector circuit that measures the voltage between the I– terminal and ground is all that is required to signal over-range conditions. When these occur, the detector causes the ERROR LED to light.

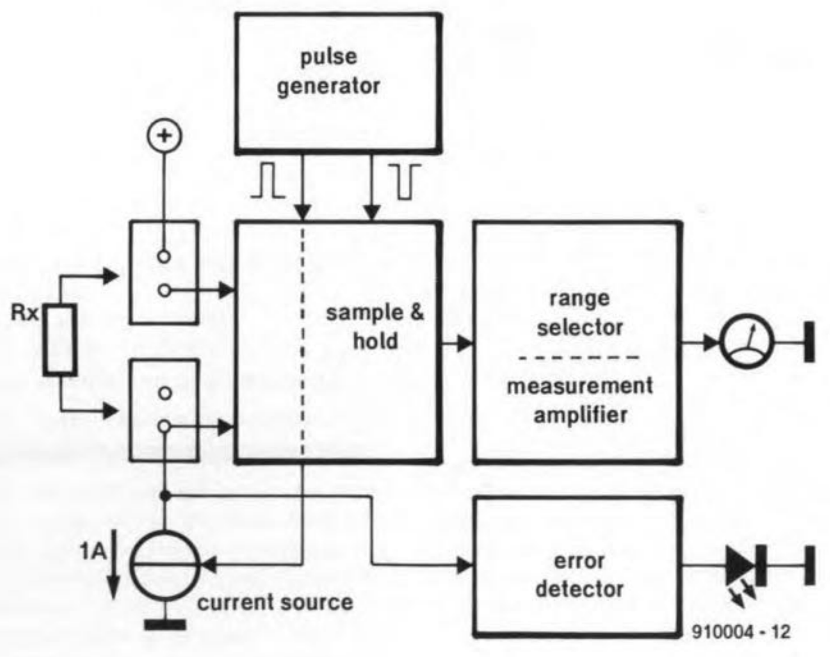


Fig. 2. Block diagram of the milliohmmeter. The resistor to be measured, R_x, is connected into a four-point network that supplies constant current pulses, and feeds the voltage developed across R_x to a sample-and-hold meter circuit.

Circuit description

Having explained the principle of operation of the milliohmmeter, we can start to look at the way the circuit is realized in practice. Figure 3 shows the circuit diagram of the instrument. The pulse generator is built around opamp IC2a. Resistors R1, R2 and R3 cause the opamp to function as a Schmitt-trigger inverter, while components R4, R5, D1 and C1 provide the function of an oscillating pulse generator. The operation of the generator is as follows: when the output of IC2a is high, capacitor C2 is charged via diode D1 and resistor R4, until the voltage across it reaches the upper switching threshold of the Schmitt-trigger. This takes about 1 ms. Next, the output of IC2a goes low, so that C2 is discharged to the lower switching threshold. This takes about 100 ms. The output of the opamp goes high again, and the cycle is repeated. Transistor T1 inverts the output signal of the pulse generator.

The current source in the instrument is built around opamp IC4. This provides a drive signal to transistor T2 that results in a voltage across emitter resistor R25 equal to the voltage at the +input of the opamp. When this voltage is constant, the emitter current is constant too. Since there is a fixed relation between the emitter current and the collector current of T2, it follows that the collector current is also constant. The magnitude of the collector current (which is the test current through the unknown resistor) depends on the value of R25 and the voltage at the +input of IC4. That voltage is supplied by preset P4, and is stabilized by a precision zener diode, D2. The zener diode is powered by the pulse generator. As a result, the voltage set by P4 at the +input of IC4 will vary between nought and the set peak value. Hence, the test current will also vary between nought and the set peak value of 1 A.

The current sent through the resistor under test can not be drawn direct from voltage regulator IC5 because the peak value (1 A) is about equal to the maximum current the 7810 is capable of supplying. However, since the peak current has a relatively short 'on' time, the necessary energy may be obtained from a large electrolytic capacitor, in this case, C5. It will be clear that the voltage across this capacitor is far from constant. This is of little consequence, however, since these variations are compensated by the current source. Resistor R30 between C5 and the voltage regulator keeps the charge current within limits. The relatively long 'off' time of the current pulses ensures sufficient time for the capacitor to be charged via this resistor.

The test current sent through the unknown resistor via terminal I– gives rise to a voltage which is fed to the sample-and-hold circuit via the Rx terminals. The sample-and-hold stores the measured voltage during the

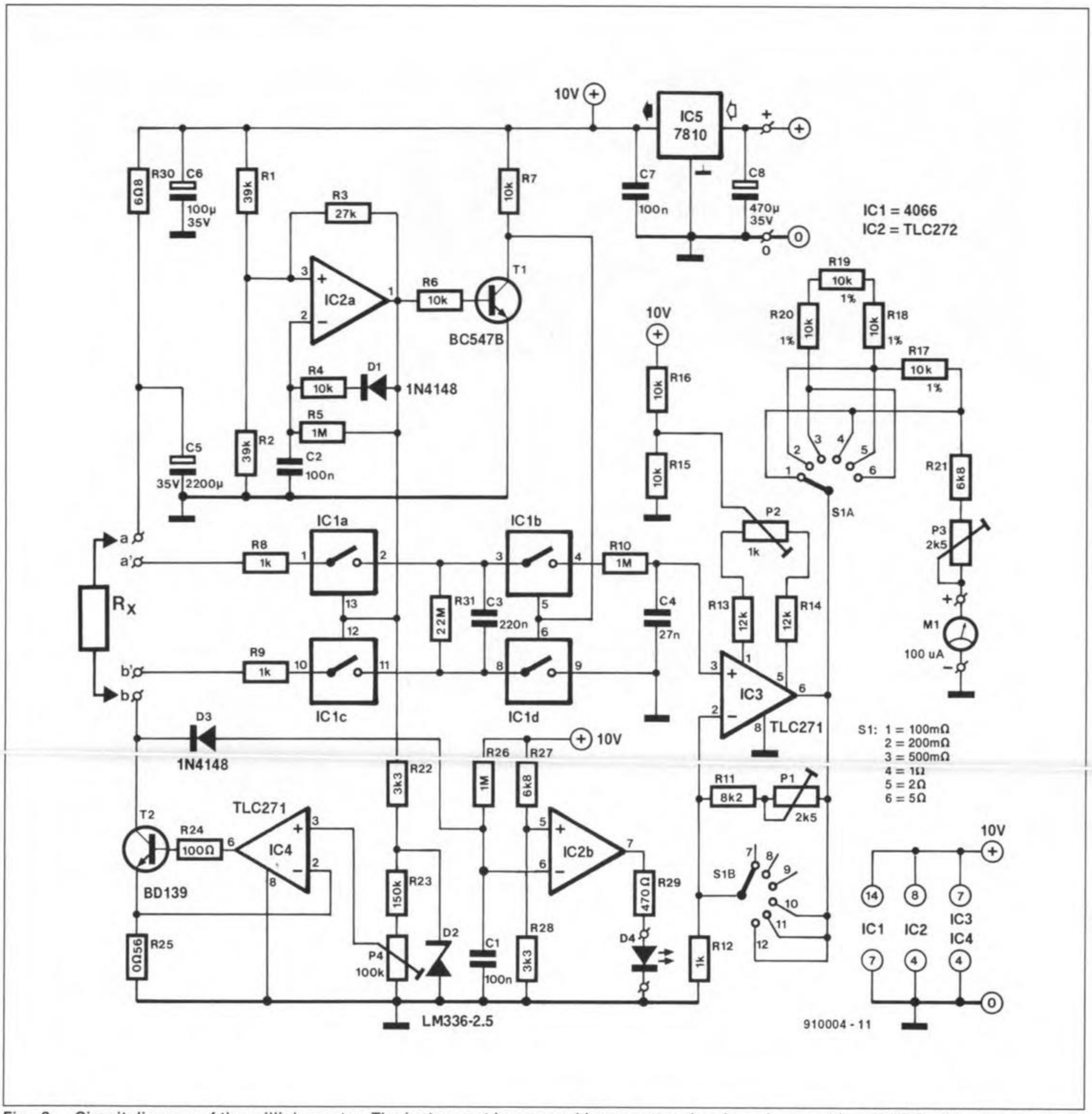


Fig. 3. Circuit diagram of the milliohmmeter. The instrument is powered by an external mains adapter with a 15 VDC, 100 mA output.

'off' time of the test current. In addition, it converts this voltage from floating into one that can be measured with respect to ground. Four CMOS bilateral switches are used to achieve this. When the current source is on, switches IC1a and IC1c are closed, while IC1b and IC1d are open. Capacitor C3 is connected in parallel with Rx via resistors R8 and R9, and will be charged until the voltage across it equals that across Rx. The resistors and C3 form a low-pass filter to suppress interference. The moment the current source is switched off, switches IC1a and IC1c are opened, while IC1b and IC1d are closed. This results in C3 being connected to ground via IC1d. The switching can be done without the risk of a short-circuit occurring, because the connection with the floating voltage across Rx is broken. Next, the voltage across C3 is fed to C4. This capacitor ensures that the measurement amplifier, IC3, is provided with an input voltage during the time C3 is connected to Rx.

Switch S1b selects between an amplification of one, and an amplification of 10, for opamp IC3. These amplification factors are used for the ranges 1 Ω, 2 Ω and 5 Ω (×1), and 100 mΩ, 200 mΩ and 500 mΩ (×10). The offset of IC3 is compensated by adjusting P2. The attenuator circuit that follows IC3 consists of a number of switchable potential dividers that drive moving-coil meter M1. The use of 1% resistors in the attenuator obviates any adjustments. The attenuator is followed by the moving-coil meter with its series resistors R21-P3.

The over-range detector is formed by comparator IC2b. Resistors R27 and R28 define the switching threshold of this comparator at about 3.3 V. The comparator compares this reference level to the voltage across capacitor C1, which is charged via R26 and can only be discharged when the current source is off. Then, the minimum voltage across C1 is about 0.6 V higher than the collector voltage of T2. When this voltage drops below 2.7 V as a result of a too high resistance between the I+ and the I− terminals, the voltage across C1 drops below the switching threshold of the comparator. Consequently, this toggles, so that LED D4 lights. Calculating the resistance value at which this happens, we find a value of about 7 Ω between the I terminals.

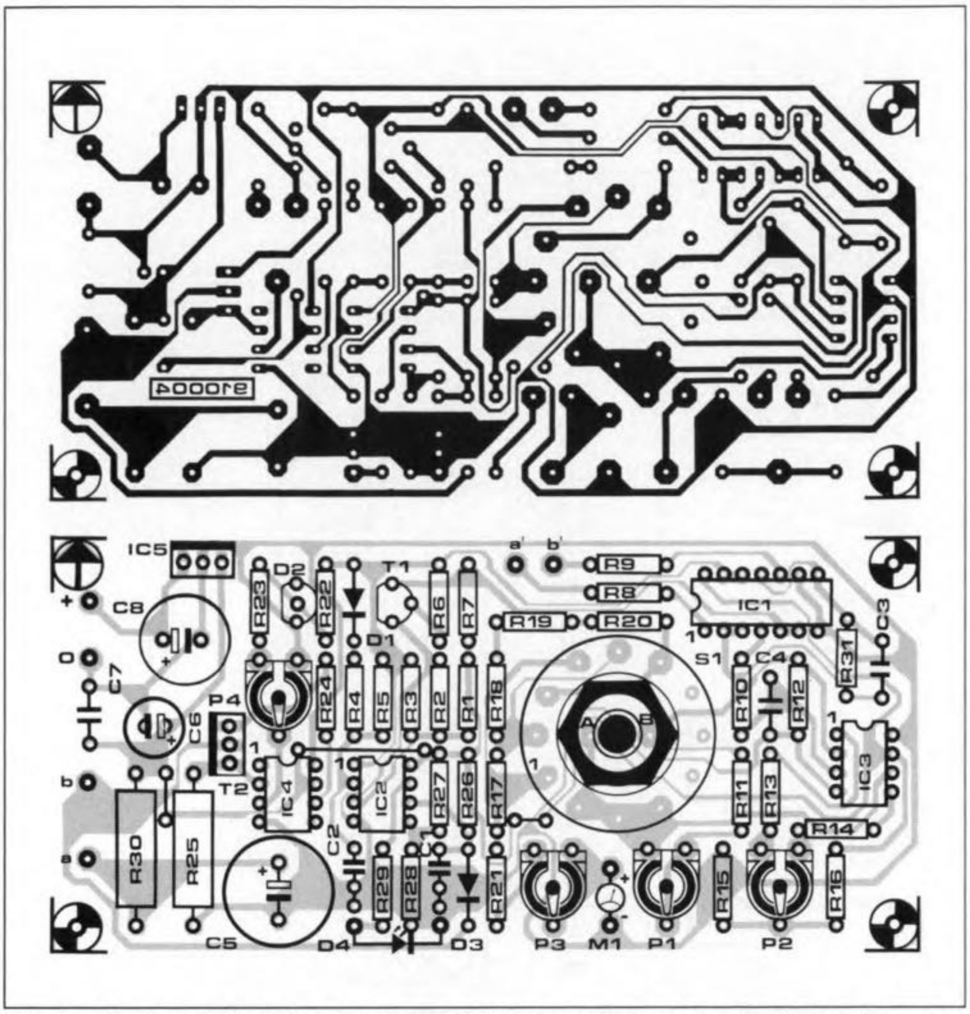


Fig. 4. Single-sided printed-circuit board for the milliohmmeter. Note that the range switches are fitted direct onto the PCB.

Construction

When the PCB shown in Fig. 4 is to be fitted into the enclosure mentioned in the parts list, the corner near IC1 will have to be cut off. Next, fit the parts on to the PCB, starting with the three wire links. Zener diode D2 comes in two different enclosures: a metal type and a plastic type. If you have a metal version, pay attention to the correct polarization (see Fig. 6). The plastic version presents no problems since its orientation is printed on the component overlay.

As with previous test instruments in this series (see the list at the end of this article), the milliohmmeter is powered by a mains adapter. In this case, an adapter with a rating of 15 VDC at about 100 mA is recommended.

The prototype of the milliohmmeter is shown in Fig. 7. The completed PCB is fitted vertically at a suitable distance behind the front panel. Use short pieces of solid wire to connect the banana sockets to the relevant points on the PCB. The range selection switch is a type for PCB-mounting that obviates any wiring. The front panel is not fitted as yet.

COMPONENTS LIST

Resistors:		
2	39kΩ	R1;R2
1	27kΩ	R3
5	10kΩ	R4;R6;R7; R15;R16
3	1MΩ	R5;R10;R26
3	1kΩ	R8;R9;R12
1	8kΩ2	R11
2	12kΩ	R13;R14
4	10kΩ 1%	R17 – R20
2	6kΩ8	R21;R27
2	3kΩ3	R22;R28
1	150kΩ	R23
1	100Ω	R24
1	0Ω56	R25
1	470Ω	R29
1	6Ω8	R30
1	22MΩ	R31
2	2kΩ5 preset H	P1;P3
1	1kΩ preset H	P2
1	100kΩ preset H	P4
Capacitors:		
3	100nF	C1;C2;C7
1	220nF	C3
1	27nF	C4
1	2200µF 35V radial	C5
1	100µF 35V radial	C6
1	470µF 35 V radial	C8
Semiconductors:		
2	1N4148	D1;D3
1	LM336-2.5V	D2
1	LED	D4
1	BC547B	T1
1	BD139	T2
1	4066	IC1
1	TLC272	IC2
2	TLC271	IC3;IC4
1	7810	IC5
Miscellaneous:		
1	100µA moving-coil meter	M1
1	2-pole 6-way rotary switch for PCB mounting	S1
1	metal enclosure, e.g., Telet LC850 (supplier: C-I Electronics). Approx. dimensions: 80×200×180 mm	
1	printed-circuit board 910004	
1	front-panel foil 910004-F	

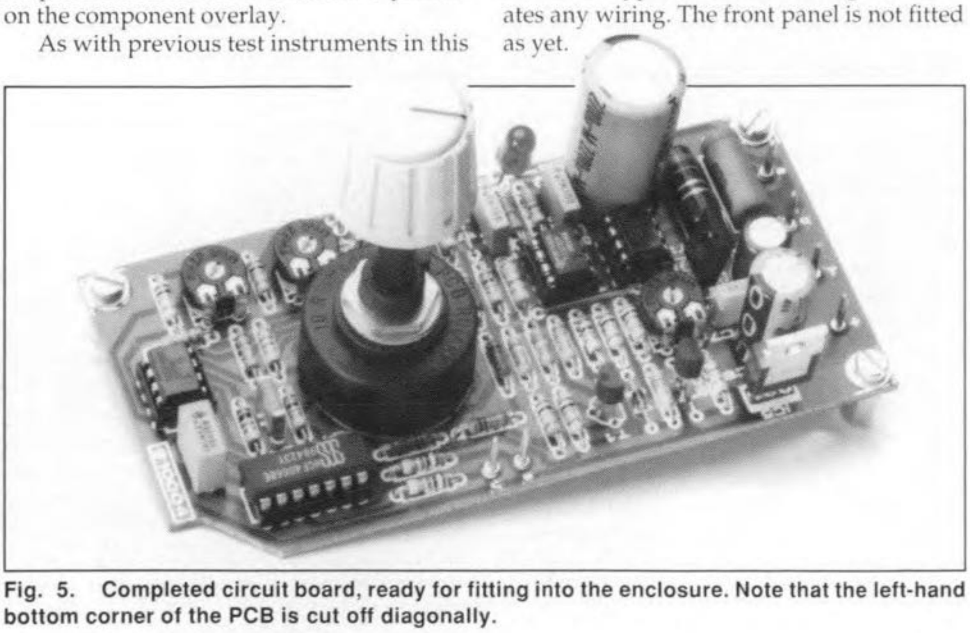


Fig. 5. Completed circuit board, ready for fitting into the enclosure. Note that the left-hand bottom corner of the PCB is cut off diagonally.

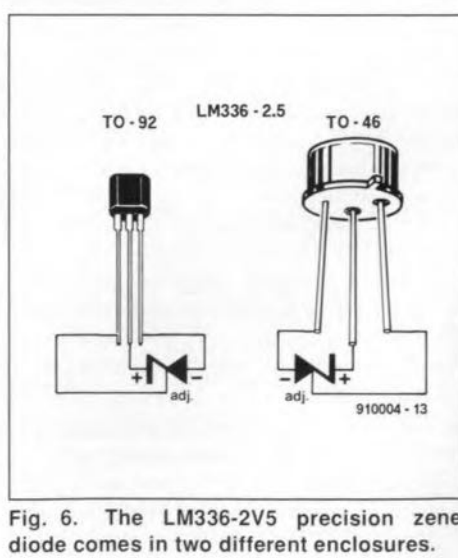


Fig. 6. The LM336-2V5 precision zener diode comes in two different enclosures.

Fig. 7. **Internal view of the instrument.**

Adjustment

To adjust the instrument you require two 1% resistors: one of 1 Ω and one of 0.5 Ω (preferred value) or smaller. Where these resistors are not available, two pieces of 0.5-Ω/m resistance wire may also be used with good results. The 1-Ω resistor then has a length of 2 m, and the 0.5-Ω resistor a length of 1 m. In the first case, an error of 1 cm corresponds to a resistance error of 0.5% — in the second case, to a resistance error of 1%. Resistance wire with a different specification may also be used, although the required values of 1 Ω and 0.5 Ω will be a little more difficult to calculate.

The indicated length of the resistance wire applies to where it is connected to the +Rx and −Rx terminals. This means that the wires must be made slightly longer than 2 m or 1 m to allow the ends to be connected to terminals I+ and I−. Having prepared the calibration resistors, put them aside for the moment.

First, null the moving-coil meter mechanically by adjusting the screw on the front. Switch on the instrument, and turn the range switch to select the 100-mΩ range. Connect the +Rx and −Rx terminals, and adjust P_2 for maximum meter deflection. Next, re-adjust P_2 until the meter just indicates zero. Do not turn P_2 any further, since this may cause an unwanted, negative, off-set. Remove the connection between the test terminals. The meter may start to deflect slowly. This is no cause for alarm, however, since it indicates that C_4 is charged by the input off-set current. This effect disappears as soon as a resistor is connected to the Rx terminals.

Next, P_4 must be adjusted. If you do not have access to an oscilloscope, set the preset to the centre of its travel (this does not affect the accuracy of the instrument). If you do have an oscilloscope, connect the 1-Ω resistor between the I terminals of the instrument. Do not connect the resistor to the Rx terminals as yet. Connect the oscilloscope as close as possible to the resistor body, or, when you use resistance wire, at the distance you have previously calculated to produce a resistance of 1 Ω. Adjust P_4 until the peak value of the measured voltage is 1 V. This sets a peak current of 1 A. Remove the scope connections, and connect the 1-Ω resistor to the Rx terminals. Switch to the 1-Ω range, and adjust P_3 for full-scale deflection of the meter.

Finally, connect the 0.5-Ω resistor, and switch the instrument to the 0.5-Ω range. Adjust P_1 until the meter indicates 0.5 Ω.

This concludes the adjustment of the milliohmmeter. At this point, you may fit the front panel, and apply the ready-made two-colour self-adhesive foil that gives the instrument a professional look. ■

Other test instruments in this series are:

- RF inductance meter. *Elektor Electronics* October 1989.
- LF/HF signal tracer. *Elektor Electronics* December 1989.
- Simple AC millivoltmeter. *Elektor Electronics* January 1990.
- Q meter. *Elektor Electronics* April 1990.
- Budget sweep/function generator. *Elektor Electronics* May 1990.
- High-current h_{FE} tester. *Elektor Electronics* September 1990.
- 400-W laboratory power supply. *Elektor Electronics* October 1990 and November 1990.

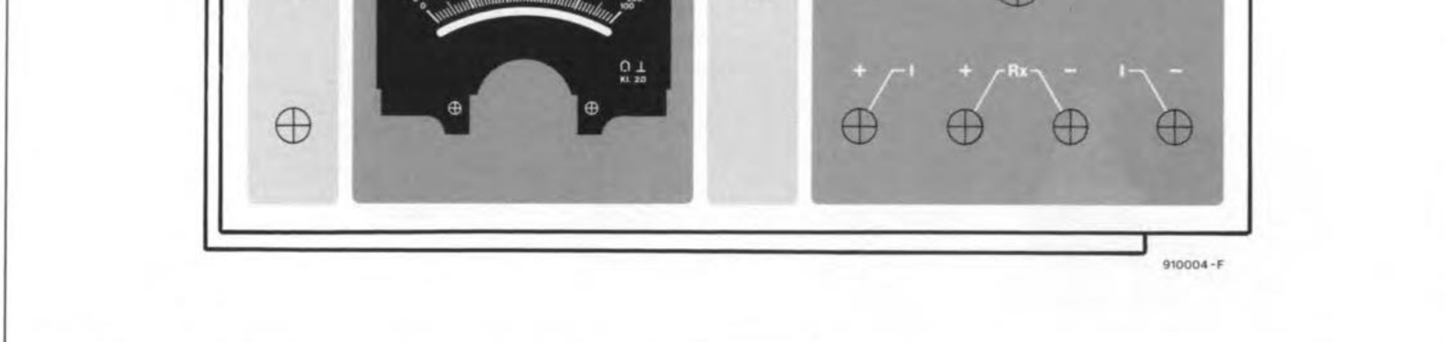


Fig. 8. Front-panel designed for the milliohmmeter. For technical reasons, the meter scale is reproduced in black here, although it is really white. The scale can be cut out of the self-adhesive foil, to replace the one that comes with the moving-coil meter.

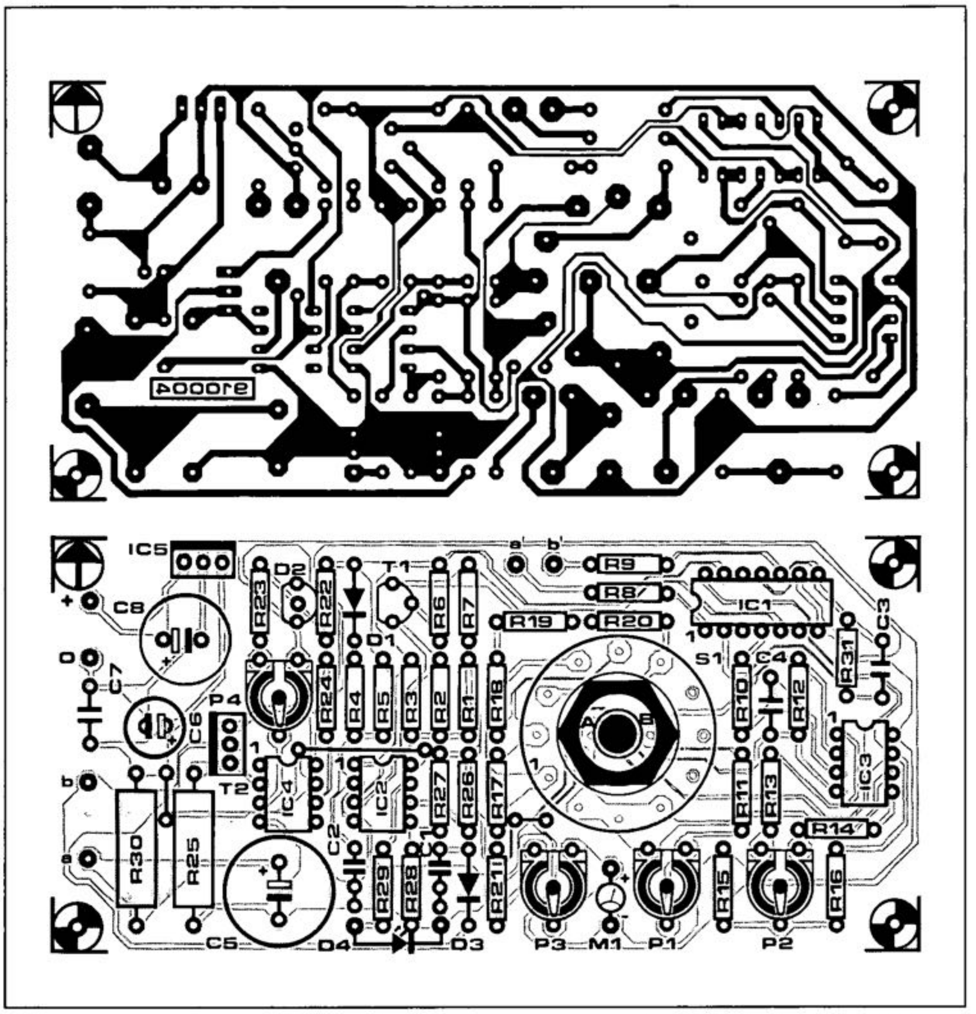


Fig. 4. Single-sided printed-circuit board for the milliohmmeter. Note that the range switches are fitted direct onto the PCB.

Construction

When the PCB shown in Fig. 4 is to be fitted into the enclosure mentioned in the parts list, the corner near IC1 will have to be cut off. Next, fit the parts on to the PCB, starting with the three wire links. Zener diode D2 comes in two different enclosures: a metal type and a plastic type. If you have a metal version, pay attention to the correct polarization (see Fig. 6). The plastic version presents no problems since its orientation is printed on the component overlay.

As with previous test instruments in this series (see the list at the end of this article), the milliohmmeter is powered by a mains adapter. In this case, an adapter with a rating of 15 VDC at about 100 mA is recommended.

The prototype of the milliohmmeter is shown in Fig. 7. The completed PCB is fitted vertically at a suitable distance behind the front panel. Use short pieces of solid wire to connect the banana sockets to the relevant points on the PCB. The range selection switch is a type for PCB-mounting that obviates any wiring. The front panel is not fitted as yet.

COMPONENTS LIST

Resistors:

2	39kΩ	R1;R2
1	27kΩ	R3
5	10kΩ	R4;R6;R7; R15;R16
3	1MΩ	R5;R10;R26
3	1kΩ	R8;R9;R12
1	8kΩ2	R11
2	12kΩ	R13;R14
4	10kΩ 1%	R17 – R20
2	6kΩ8	R21;R27
2	3kΩ3	R22;R28
1	150kΩ	R23
1	100Ω	R24
1	0Ω56	R25
1	470Ω	R29
1	6Ω8	R30
1	22MΩ	R31
2	2kΩ5 preset H	P1;P3
1	1kΩ preset H	P2
1	100kΩ preset H	P4

Capacitors:

3	100nF	C1;C2;C7
1	220nF	C3
1	27nF	C4
1	2200μF 35V radial	C5
1	100μF 35V radial	C6
1	470μF 35 V radial	C8

Semiconductors:

2	1N4148	D1;D3
1	LM336-2.5V	D2
1	LED	D4
1	BC547B	T1
1	BD139	T2
1	4066	IC1
1	TLC272	IC2
2	TLC271	IC3;IC4
1	7810	IC5

Miscellaneous:

1	100μA moving-coil meter	M1
1	2-pole 6-way rotary switch for PCB mounting	S1
1	metal enclosure, e.g., Telet LC850 (supplier: C-I Electronics). Approx. dimensions: 80×200×180 mm	
1	printed-circuit board 910004	
1	front-panel foil 910004-F	

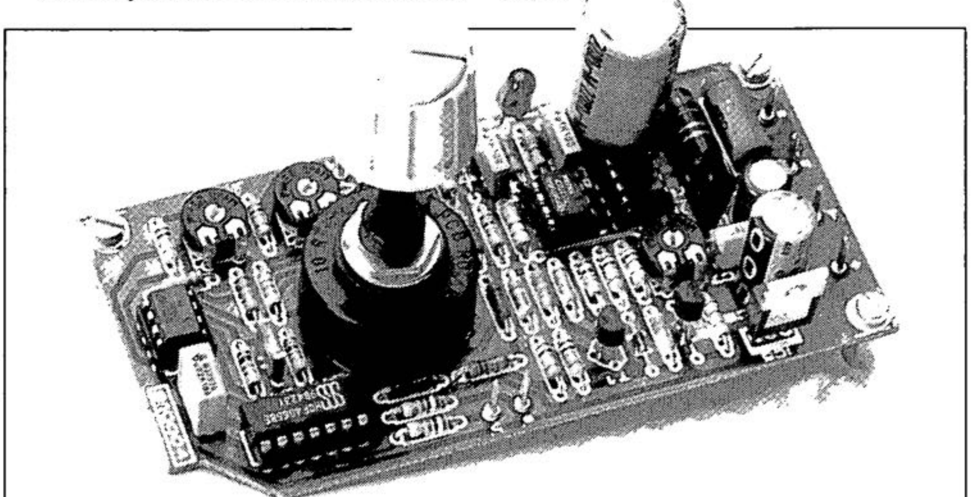


Fig. 5. Completed circuit board, ready for fitting into the enclosure. Note that the left-hand bottom corner of the PCB is cut off diagonally.

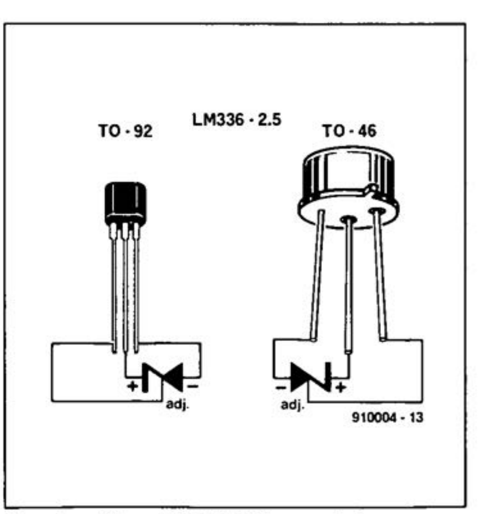


Fig. 6. The LM336-2V5 precision zener diode comes in two different enclosures.

MEASUREMENT TECHNIQUES (2)

by F.P. Zantis

After the brief discussion on measuring, errors and tolerances in Part 1, we now turn our attention to practical measurements, more particularly the measurement of voltages.

Measurement of direct voltages

Even measuring a direct voltage is not the straightforward job it is often assumed to be. This may be because of the level of the voltage: very low voltages lie under the noise level and their measurement requires special equipment and techniques, whereas very high voltages require the use of an external prescaler, such as a capacitive voltage divider. But even well away from these extremes there exists the danger that the result will be distorted by the internal resistance of the measuring instrument.

An ideal voltmeter has an infinite internal resistance, but in practice that is, of course, unattainable. Perhaps that is just as well, because a very large resistance produces a high noise voltage and this will affect the measurement. In practice, the internal resistance of the instrument should be appreciably higher than the resistance across which the voltage is being measured, but it should not approach infinity.

Depending on the nature of the measurement, there are two types of voltmeter on the market: those whose internal resistance depends on the selected meter range and those whose internal resistance is constant (normally between 1 MΩ and 10 MΩ).

The first kind includes most low-priced multimeters, a typical example of which is shown in Fig. 8. One of the quality criteria of these instruments is their characteristic resistance, which is expressed in ohms per volt. The internal resistance, R_{in}, is calculated by dividing the characteristic resistance (Ω) by the full-scale deflection (V) of the relevant meter range.

Fig. 8. Typical analogue multimeter.

The inverse value of the characteristic resistance gives the current that flows through the network at full-scale deflection.

For instance, if the characteristic resistance of the instrument is 100 kΩ/V, its internal resistance over the 1-V range is 100 kΩ, over the 3-V range, 300 kΩ, and over the 5-V range, 500 kΩ.

The input amplifier stage of a typical multimeter shown in Fig. 9 is a grounded-emitter circuit with current feedback. Were the voltage measured across R_2, there would be a problem because, if, correctly, the 3-V range is selected, the internal resistance of the instrument is 300 kΩ. Unfortunately, this resistance is in parallel with R_2, so that the ratio of divider R_1–R_2 changes according to the selected range and this will, of course, give rise to incorrect measurements. In such a case, it may, therefore, be better to select a higher range. True, the error will then be larger, but so will the internal resistance and this makes the effect on the divide ratio smaller. None the less, even then precise measurements are not possible.

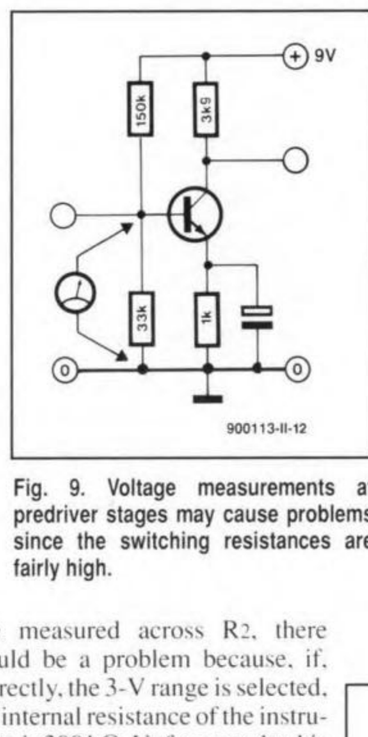


Fig. 9. Voltage measurements at predriver stages may cause problems since the switching resistances are fairly high.

From the above, it is clear that measurements in high-resistance circuits, such as opamp inputs and base and gate inputs of transistors and FETs respectively, require instruments with a high internal resistance.

It is interesting to calculate how much greater the measurement error is when a 20 kΩ/V instrument is used instead of, say, a 50 kΩ/V one. Once you know the problem, it is quite possible to use a low-priced multimeter for most voltage measurements and guestimate the error. However, in the long run, that is not a satisfactory solution to the problem. Fortunately, manufacturers are aware of this and modern instruments have a much higher internal resistance than their predecessors. This is achieved on the one hand by far more sensitive meters and on the other hand by the use of, for instance, impedance converters (simulating the valve voltmeters of yesteryear).

Instruments with input amplifiers

Instruments with input amplifiers generally have a high input resistance, at least 1 MΩ, and this value is constant, i.e., independent of the metering range. Such instruments are much better suited for use in high-resistance circuits. None the less, even they have their limitations. For instance, in circuits using FETs or electronic valves, an internal resistance of even 1 MΩ can cause errors.

To understand the function of an instrument with input amplifier, consider the circuit in Fig. 10. This shows the layout of an electronic voltmeter, which may actually be

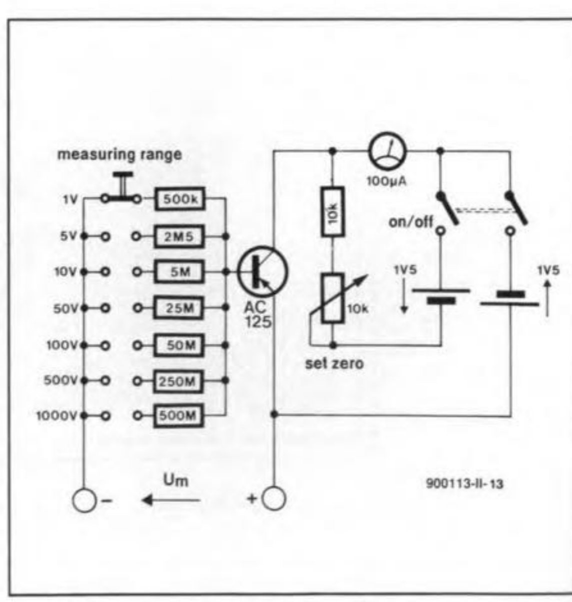


Fig. 10. Circuit of a simple meter amplifier.

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constructed by any electronics enthusiast. Because of the transistor amplifier, the branches of the input divider have a very high resistance. For instance, that for the 1-V range is 500 kΩ, which is equal to the characteristic resistance of the instrument.

A disadvantage of this type of circuit is the temperature dependence of the quiescent collector current. To counter this current, an equal current of opposite polarity, derived from an auxiliary battery, is passed through the instrument. This current is limited by the series network consisting of a 10 kΩ resistance and a potentiometer. Prior to each measurement, the potentiometer must be set to ensure zero reading of the meter.

Figure 11 shows the circuit of a commercial impedance converter for multimeters, which may be used in virtually any kind of multimeter. In use, the instrument must be set to the most sensitive current range. Calibration is effected with R6. Zero setting is accomplished with R8.

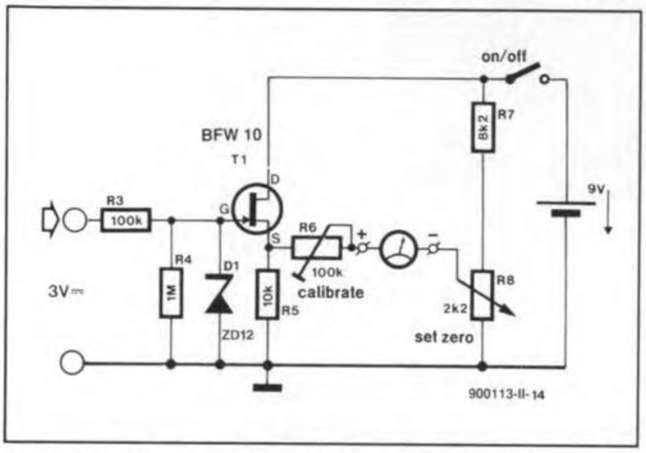


Fig. 11. Typical commercial impedance converter for multimeters.

Multimeters for industrial use have rather more complex circuits than that in Fig. 11, but they are not necessarily any more exact. Here, as almost everywhere, you get what you pay for: if you want good accuracy, you have to pay a good price.

Figure 12 shows a popular analogue multimeter with integral input amplifier. Its input resistance is 10 MΩ and is independent of the selected metering range.

Digital multimeters generally also have a high input resistance (up to 10 MΩ), but the measurand* must additionally be translated by an analogue/digital converter. Their accuracy is, therefore, dependent on the accuracy of the converter.

As mentioned before, the last digit of the read-out of a digital multimeter is error-prone and should, therefore, not be taken into account where great precision is required.

Measurement of alternating voltages

What has been said about the internal resistance of measuring instruments for direct voltage is equally applicable to those for measuring alternating voltages. There are some additional difficulties as well. For instance, in most instruments, the internal resistance for alternating voltages is lower than that for direct voltages, and is typically 20 kΩ/V= and 5 kΩ/V~. This is because for alternating voltages to be measured by moving coil meters, it is necessary for them to be rectified. Now, every rectifier diode has a fairly significant threshold voltage and for that reason the full-scale deflection on the lowest metering range cannot be very small.

To reduce the effect of the threshold voltage, it is normal to use diodes in only one section of the bridge rectifier and resistors in the other section—see Fig. 13. This arrengements leads to an additional current through the resistors and this lowers the internal resistance and also the sensitivity. When the instrument has an input amplifier these aspects are of no consequence, because the impedance converter at the input isolates the measurand* from the meter section.

There are two other problems in measuring alternating voltages: (1) the r.m.s. value is shown only if the measurand* is sinusoidal and (2) the instrument does not function properly at fairly high frequencies.

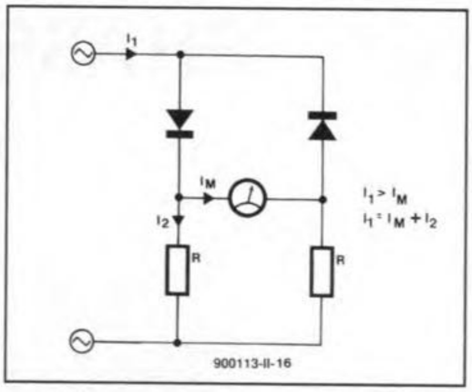


Fig. 13. Many multimeters do not use a full rectifier bridge in order to reduce the effect of the threshold voltage of the diodes.

The first problem is not so bad when a moving coil meter is used, since in that case the arithmetic mean value of the rectified voltage will be calculated and indicated. In digital instruments, the reading is not reliable if the measurand* is not sinusoidal.

The second problem is again not too serious in analogue instruments, since the frequency range of them is generally considerably higher than that of digital instruments. Many digital instruments have an upper frequency range as low as 400 Hz so that even measuring audio signals with these becomes problematic. To add to the problems, there is no indication of the frequency range on many low-priced digital multimeters: that information is normally hidden in the small print in the specification contained in the operating manual. Figure 14 shows the, dramatically different, frequency ranges of two digital multimeters. One may be used up to 100 kHz, whereas the other becomes unreliable above 1 kHz.

(to be continued)

* measurand = electrical quantity to be, or being, measured.

Fig. 12. Analogue electronic multimeter.

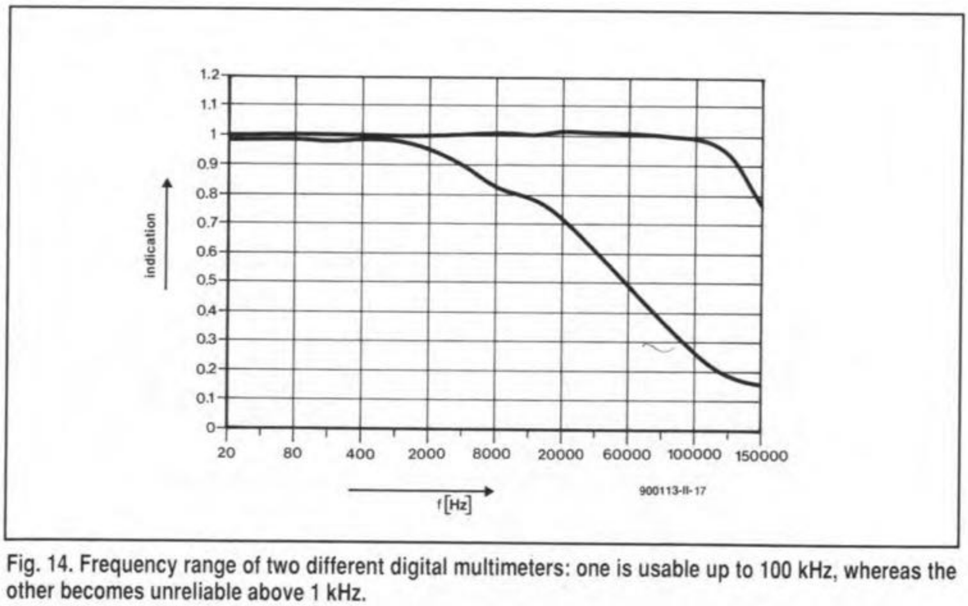


Fig. 14. Frequency range of two different digital multimeters: one is usable up to 100 kHz, whereas the other becomes unreliable above 1 kHz.

PHASE CHECK FOR AUDIO SYSTEMS

While setting up and connecting audio equipment it is important to have all the units — microphone, loudspeakers and everything in between — 'in phase', that is, interconnected with the right polarity. The low-cost instrument described here is particularly handy for checking out the phase of almost any audio system, whether installed in a living room, in a car, in a studio, or on a stage.

K. Orlowski

REVERSED phase connections in an audio equipment system give strange and unpredictable effects such as the unwanted attenuation or boosting of a particular frequency range, jet-plane effects, whistling noises, or amplifier output power which does not seem to produce any usable sound level. To avoid these problems, use the simple instrument described here. Based on a transmitter and a receiver with a simple good/fault indication, the instrument will check out the system from the input (microphone or line input) right through to the output (loudspeaker or line output).

The transmitter supplies positive or negative needle pulses, which are fed either electrically to an equipment input, via the line-cinch output socket, or acoustically to a microphone, via the built-in loudspeaker. Accordingly, the receiver has an electrical (line) input and an acoustic (microphone) input.

The drawings in Fig. 1 illustrate two ways of using the transmitter and the receiver for phase tests on audio equipment. Figure 1a shows the set-up used to check the polarity of a microphone, and Fig. 1b that used to ensure a loudspeaker is connected the right way around. The LEDs on the receiver provide a quick indication whether or not the received pulses have the same polar-

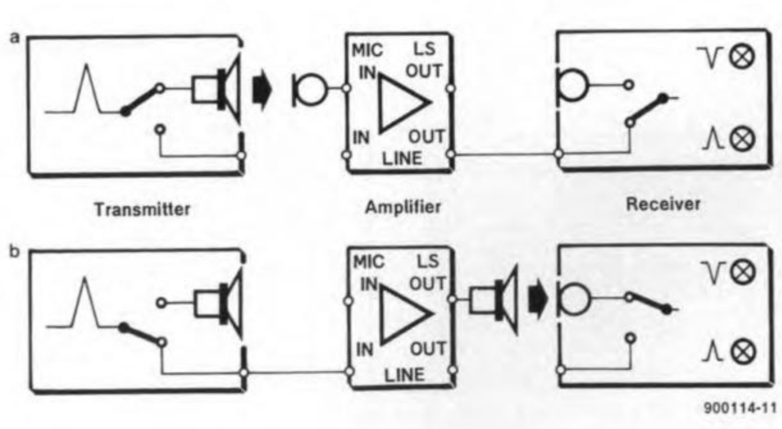


Fig. 1. Application examples of the phase-check system.

ity as the transmitted pulses. If the receiver indicates the opposite polarity of the transmitter, the chances are pretty high that there is a reversed signal connection somewhere in the system.

The pulse transmitter

The needle pulses are generated by oscillator IC1A (see Fig. 2), which is built from a NAND gate with two Schmitt-trigger inputs. After applying the supply voltage, these inputs take on complementary logic levels, i.e., one is high, the other is low. Consequently, the output of the gate is logic high. Capacitor C2 is charged via resistor R1, until the voltage on it reaches the high threshold voltage of about 5.5 V. Next, the output of the Schmitt-trigger toggles to 0, so that C2 is discharged via D1 and R2, until the low threshold voltage of about 3 V is reached. The NAND gate toggles, and the charging of C2 starts again.

The above process is cyclical and results in a self-oscillating circuit. Since R2 is much smaller than R1, the discharge time of C2 is much shorter than the charge time. As a result, the on-off (mark-space) ratio of the output signal is about 2 ms/1 s, or 0.002. Mind you, 'off' means 'logic high' here since we are dealing with a NAND gate.

The oscillator output signal is fed to two sub-circuits. One is a small loudspeaker driver based on emitter follower T2. The loudspeaker connections can be swapped by switch contacts S1c and S1d. When an oscilloscope is connected to the loudspeaker, it indicates negative-going needle pulses with the switch set to the centre position, and positive-going pulses with the switch set to the upper position. Likewise, in the other signal branch, the polarity is changed by switching transistor T1 from a common-emitter circuit (S1b at centre position) to a common-collector circuit (S1b at centre position). Coupling capacitor C3 takes the test signal to an attenuator that supplies output levels of 1 Vpp (0 dBV), –20 dBV and –40 dBV.

The receiver

The circuit diagram of the receiver (Fig. 3) shows that two almost identical detectors are used. The test signal is supplied to the two voltage amplifiers T1-T2 and T4-T5 either by the electret microphone, or by the signal source connected to K1. In the latter case, the signal is taken through a high-pass filter, R1-C3, before it arrives at a voltage limiter, D1-D2. The input source, microphone or line, is selected with switch S1. The voltage amplifiers are complementary circuits: T1-T2 amplifies the negative pulses, T4-T5 the positive pulses.

The two monostables in IC1 have different networks at their trigger inputs to enable them to respond to negative pulse edges (IC1A) or positive pulse edges (IC1B). To prevent the trailing edge of a pulse triggering the wrong monostable, IC1A and IC1B disable one another when one of them is actuated. The monostables thus allow the circuit to determine whether a pulse starts with a

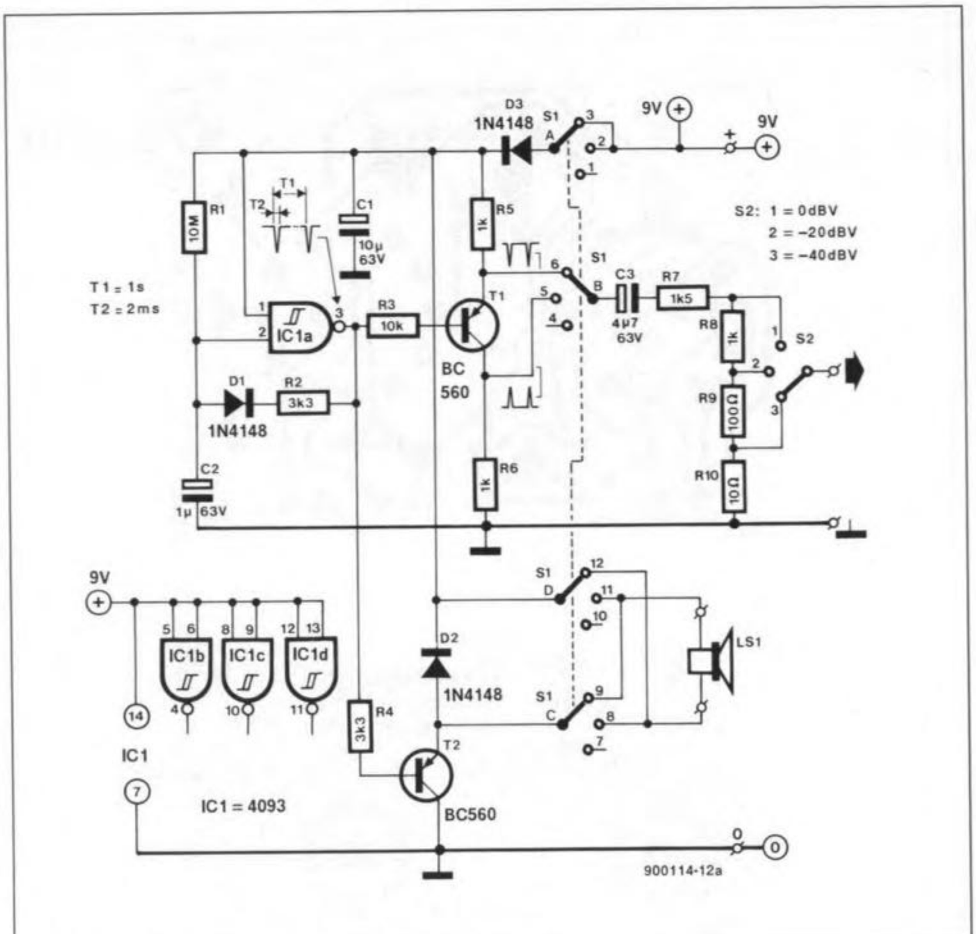


Fig. 2. Circuit diagram of the pulse transmitter.

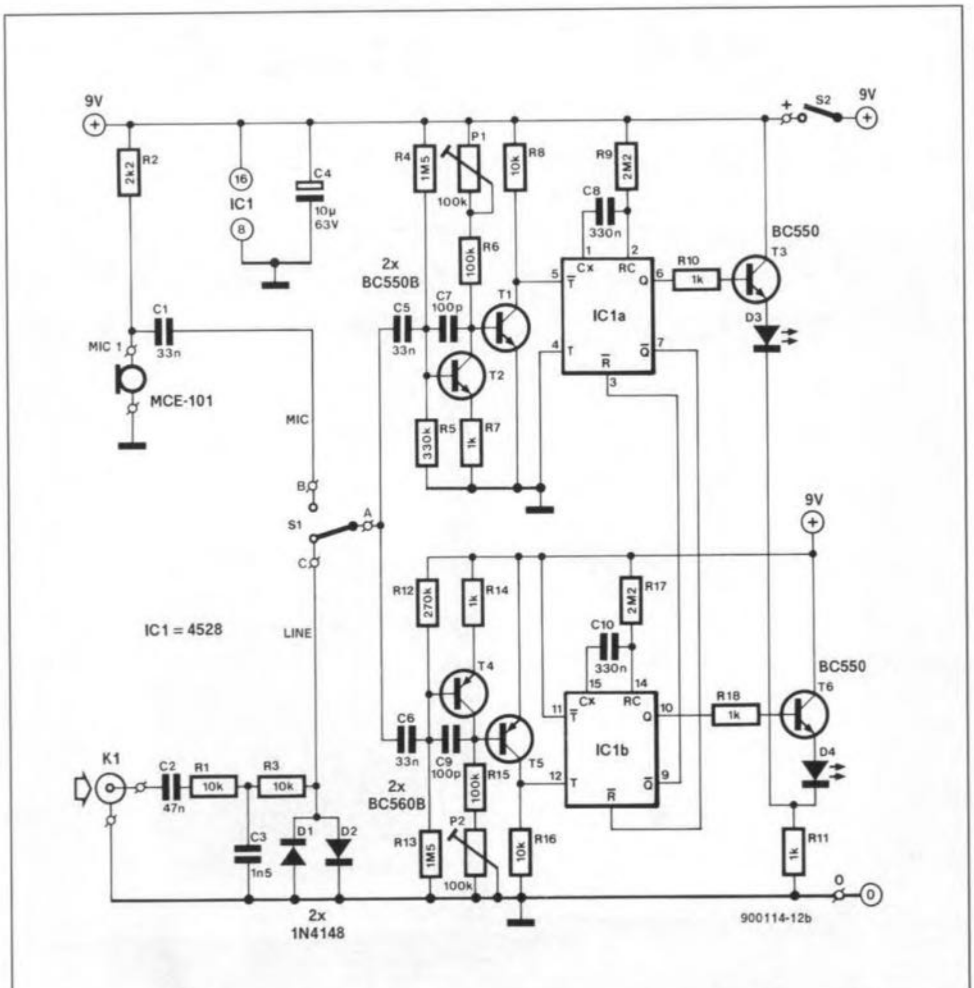


Fig. 3. Circuit diagram of the pulse receiver. The polarity of the measured signal is indicated by two LEDs, D3 and D4.

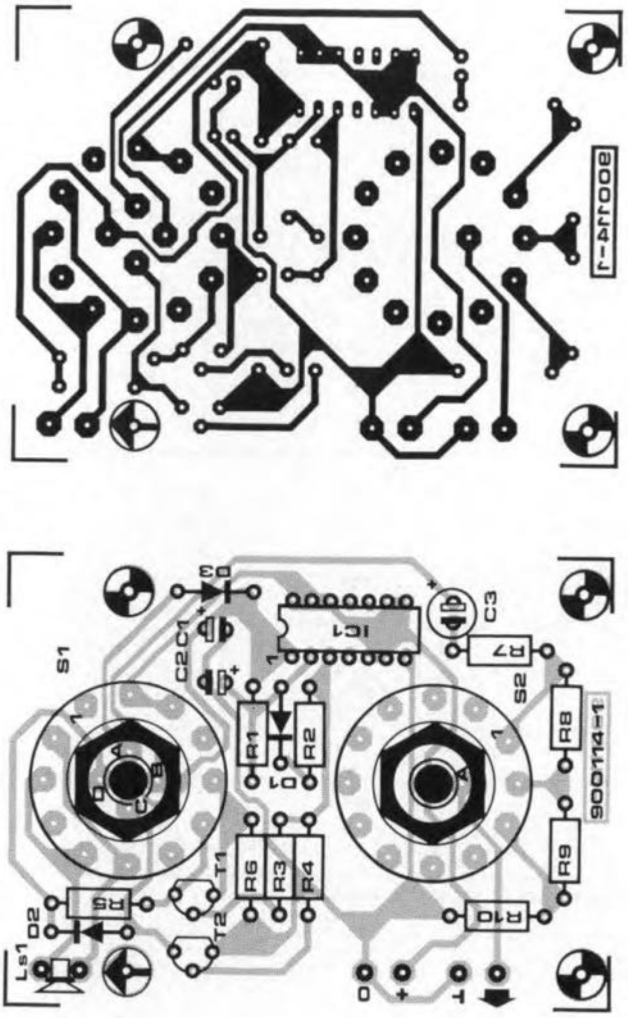


Fig. 4a. Single-sided printed circuit board for the pulse transmitter.

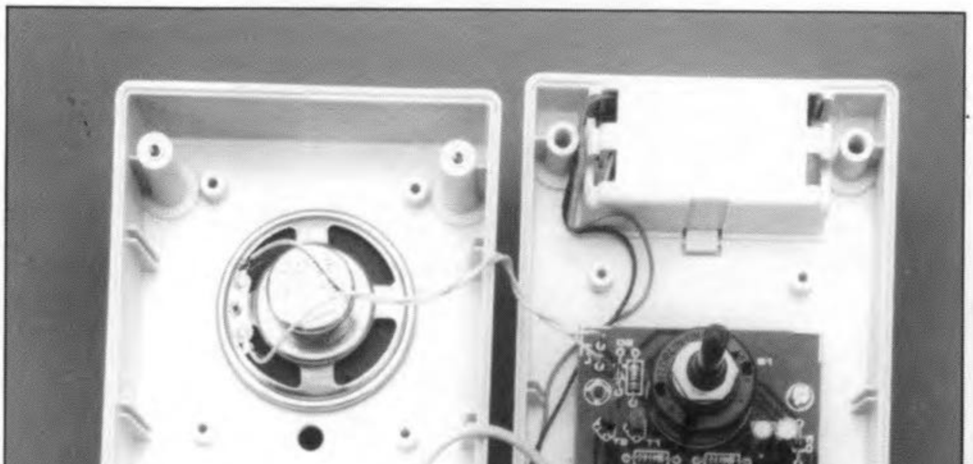


Fig. 5. A look inside the completed pulse transmitter.

COMPONENTS LIST

TRANSMITTER:

Resistors:

1	10MΩ	R1
2	3kΩ3	R2;R4
1	10kΩ	R3
3	1kΩ	R5;R6;R8
1	1kΩ5	R7
1	100Ω	R9
1	10Ω	R10

Capacitors:

1	10μF 63V radial	C1
1	1μF 63V radial	C2
1	4μF7 63V radial	C3

Semiconductors:

3	1N4148	D1;D2;D3
2	BC560	T1;T2
1	4093	IC1

Miscellaneous:

1	3-way 4-pole rotary switch for PCB mounting	S1
1	12-way 1-pole rotary switch for PCB mounting	S2
1	8-Ω loudspeaker, dia. 50 mm	LS1
1	ABS enclosure, e.g., OKW A9409126	
1	clip for 9-V battery	
1	phono socket	
1	printed-circuit board 900114-1	

RECEIVER:

Resistors:

4	10kΩ	R1;R3;R8;R16
1	2kΩ2	R2
2	1MΩ5	R4;R13
1	330kΩ	R5
2	100kΩ	R6;R15
5	1kΩ	R7;R10;R11;R14; R18
2	2MΩ2	R9;R17
1	270kΩ	R12
2	100kΩ preset H	P1;P2

Capacitors:

3	33nF	C1;C5;C6
1	47nF	C2
1	1nF5	C3
1	10μF 63V radial	C4
2	100pF	C7;C9
2	330nF	C8;C10

Semiconductors:

2	1N4148	D1;D2
1	red LED	D3
1	green LED	D4
4	BC550B	T1;T2;T3;T6
2	BC560B	T4;T5
1	4528	IC1

Miscellaneous:

1	electret microphone	Mic1
1	phono socket	K1
1	miniature SPDT switch	S1
1	miniature SPST switch	S2
1	clip for 9V PP3 battery	
1	ABS enclosure, e.g., OKW A9409126	
1	printed-circuit board 900114-2	

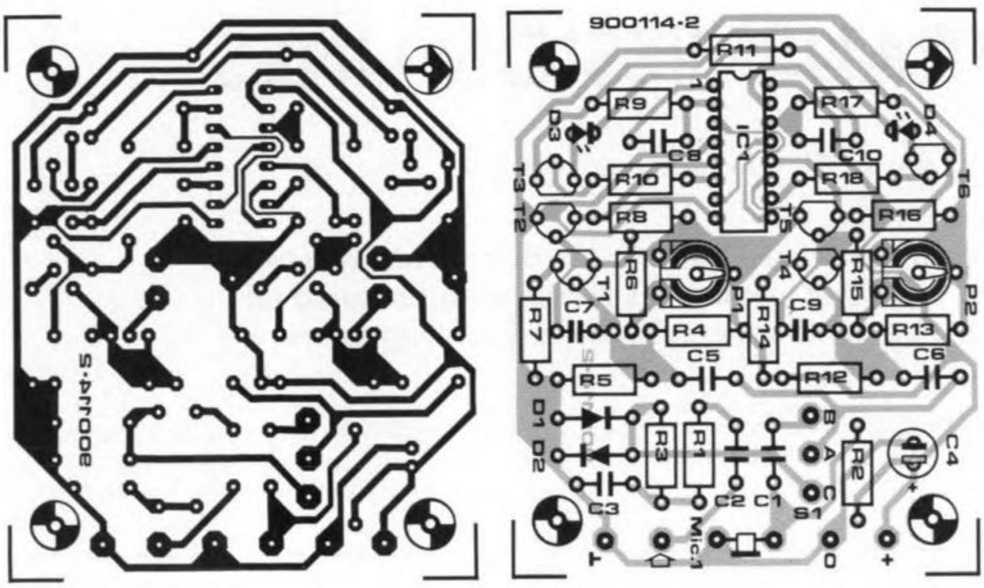


Fig. 4b. Single-sided printed-circuit board for the pulse receiver.

positive (rising) or a negative (falling) edge. The two LEDs, D_3 and D_4, indicate the respective polarities. The monostable times are set at about 0.5 s with R_9-C_8 and R_{17}-C_{10}. This causes the active LED to flicker.

Building and testing

The receiver and the transmitter are best built on the printed-circuit boards shown in Fig. 4. Be sure to fit all polarized components (electrolytic capacitors, ICs, transistors and diodes) the right way around. Also make sure that the two rotary switches on the transmitter PCB are fitted as shown on the overlay (note the '1' mark, and the letters that indicate the poles). On completion of the two units, apply the self-adhesive foils shown in Fig. 6 to the enclosure front panels.

Interconnect the transmitter and the receiver via their line sockets, and check that the LED indication on the receiver is in accordance with the polarity set on the transmitter. When the LEDs remain off, IC_1 in the receiver may not have sufficient gain. In that case, adjust P_1 and P_2 until the receiver does trigger correctly. ■

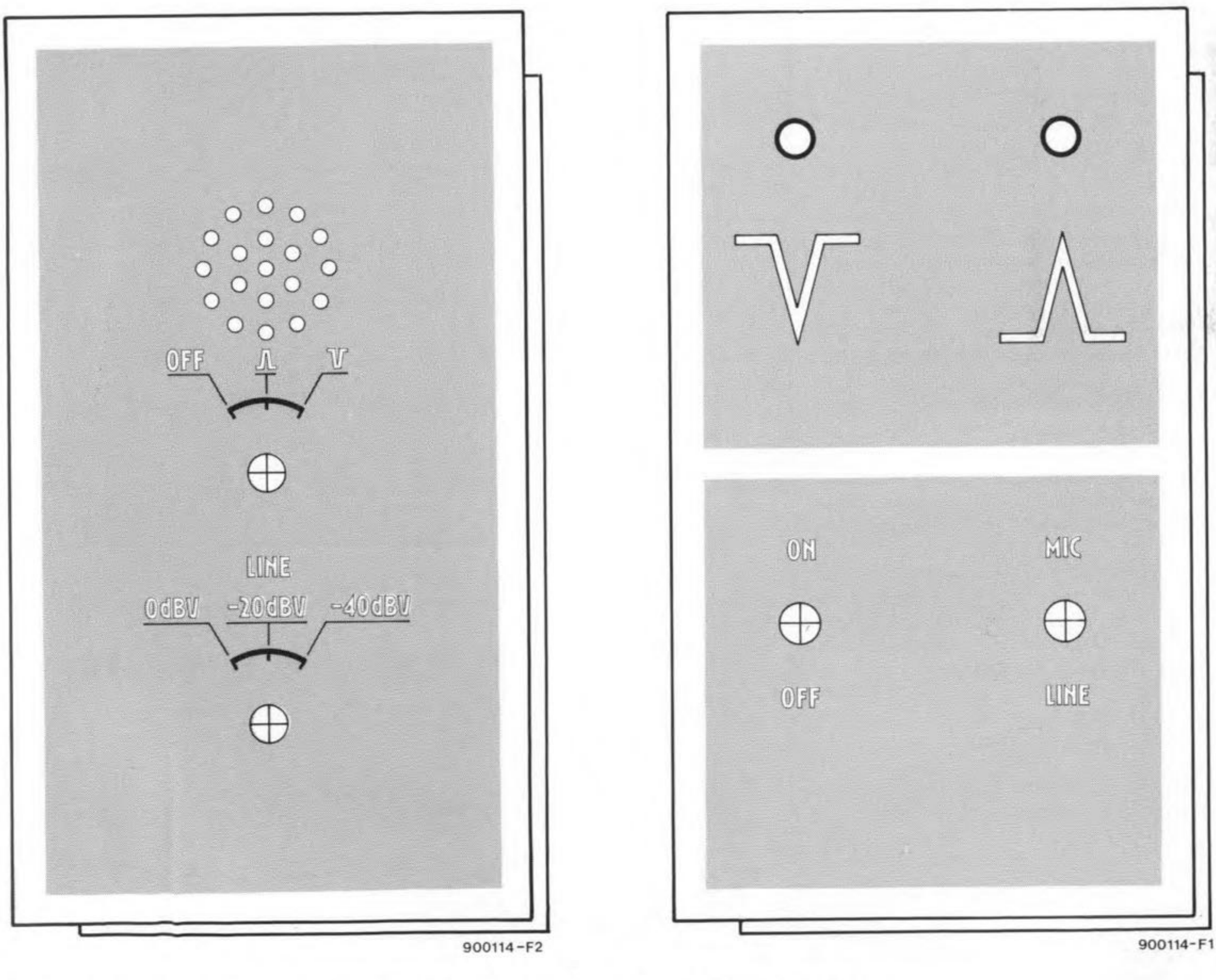
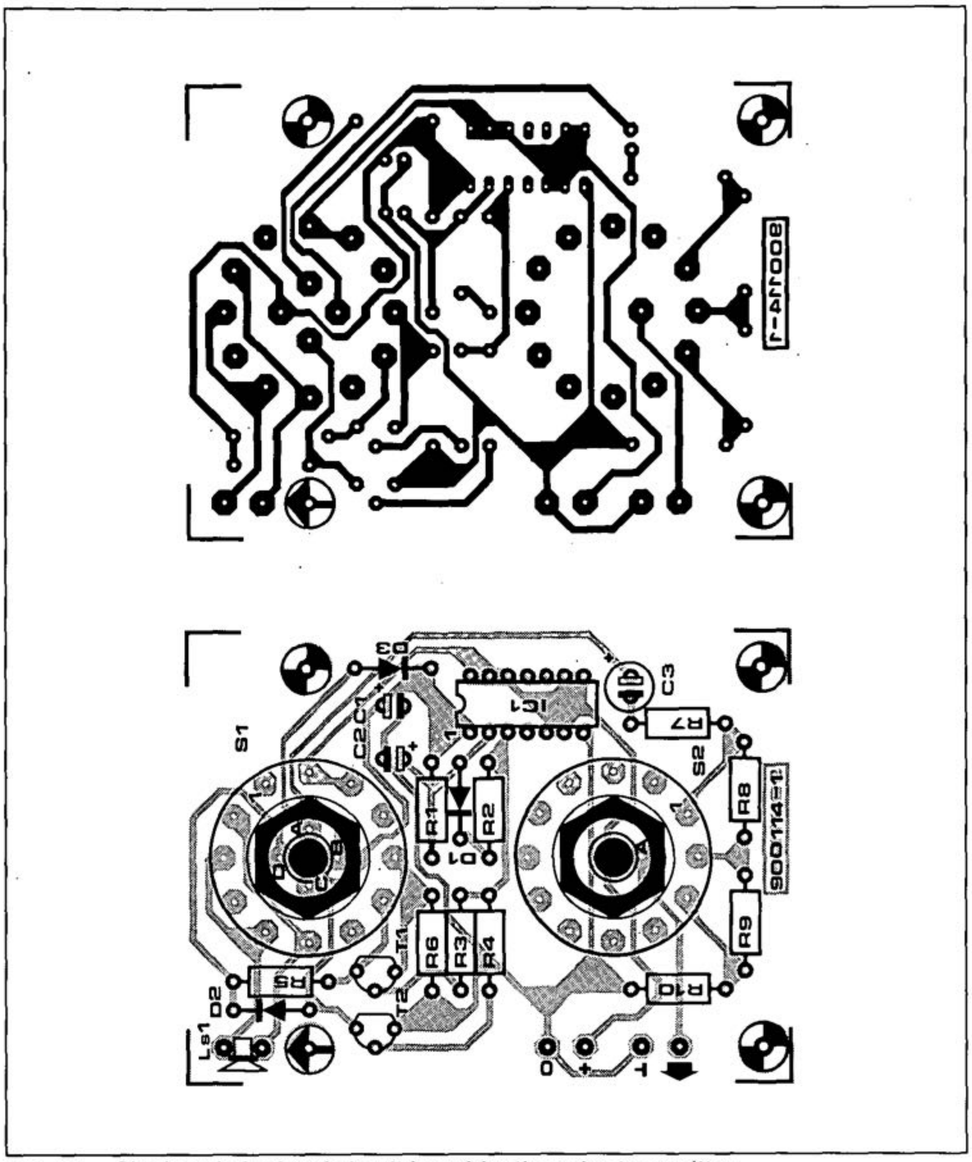


Fig. 6. Design of the self-adhesive front panel foil for the transmitter (left) and the receiver (right).



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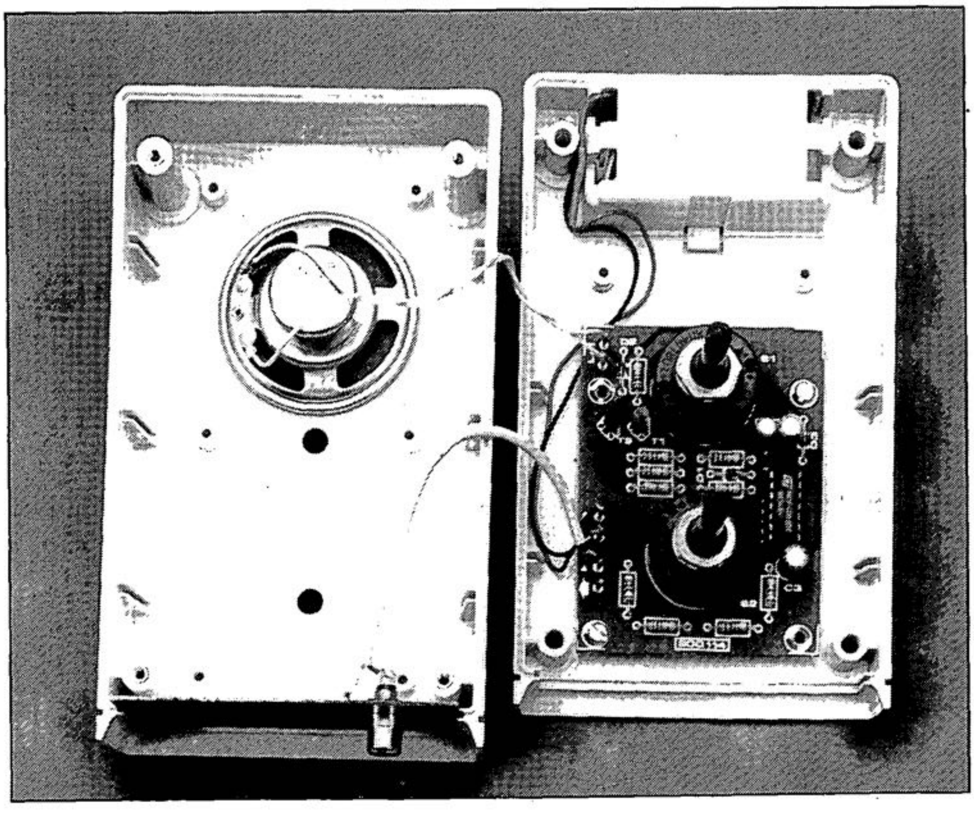

Fig. 5. A look inside the completed pulse transmitter.

COMPONENTS LIST

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Resistors:

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3	1kΩ	R5;R6;R8
1	1kΩ5	R7
1	100Ω	R9
1	10Ω	R10

Capacitors:

1	10µF 63V radial	C1
1	1µF 63V radial	C2
1	4µF7 63V radial	C3

Semiconductors:

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Miscellaneous:

1	3-way 4-pole rotary switch for PCB mounting	S1
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1	8-Ω loudspeaker, dia. 50 mm	LS1
1	ABS enclosure, e.g., OKW A9409126	
1	clip for 9-V battery	
1	phono socket	
1	printed-circuit board 900114-1	

RECEIVER:

Resistors:

4	10kΩ	R1;R3;R8;R16
1	2kΩ2	R2
2	1MΩ5	R4;R13
1	330kΩ	R5
2	100kΩ	R6;R15
5	1kΩ	R7;R10;R11;R14; R18
2	2MΩ2	R9;R17
1	270kΩ	R12
2	100kΩ preset H	P1;P2

Capacitors:

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Miscellaneous:

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1	miniature SPDT switch	S1
1	miniature SPST switch	S2
1	clip for 9V PP3 battery	
1	ABS enclosure, e.g., OKW A9409126	
1	printed-circuit board 900114-2	

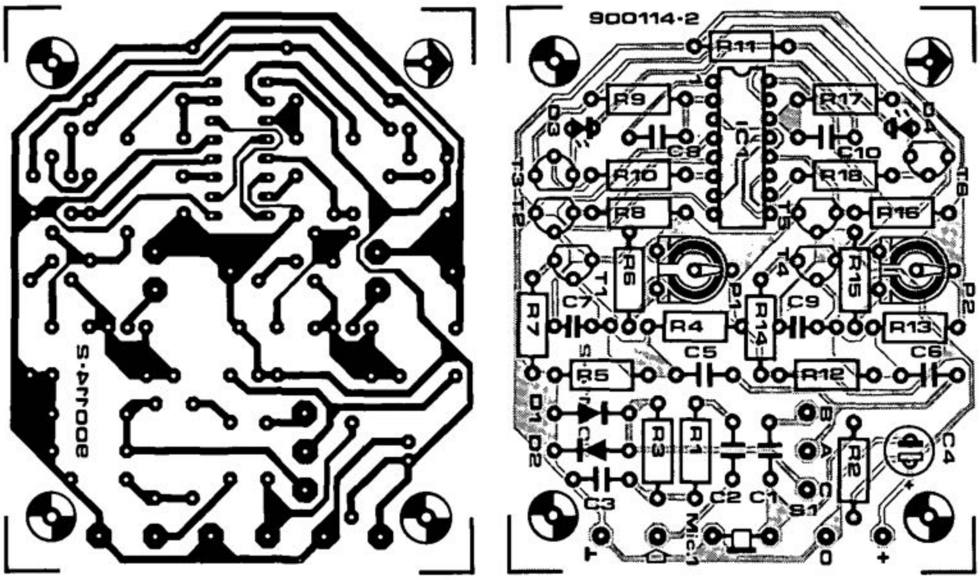

Fig. 4b. Single-sided printed-circuit board for the pulse receiver.

positive (rising) or a negative (falling) edge. The two LEDs, D3 and D4, indicate the respective polarities. The monostable times are set at about 0.5 s with R9-C8 and R17-C10. This causes the active LED to flicker.

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Interconnect the transmitter and the receiver via their line sockets, and check that the LED indication on the receiver is in accordance with the polarity set on the transmitter. When the LEDs remain off, IC1 in the receiver may not have sufficient gain. In that case, adjust P1 and P2 until the receiver does trigger correctly. ■

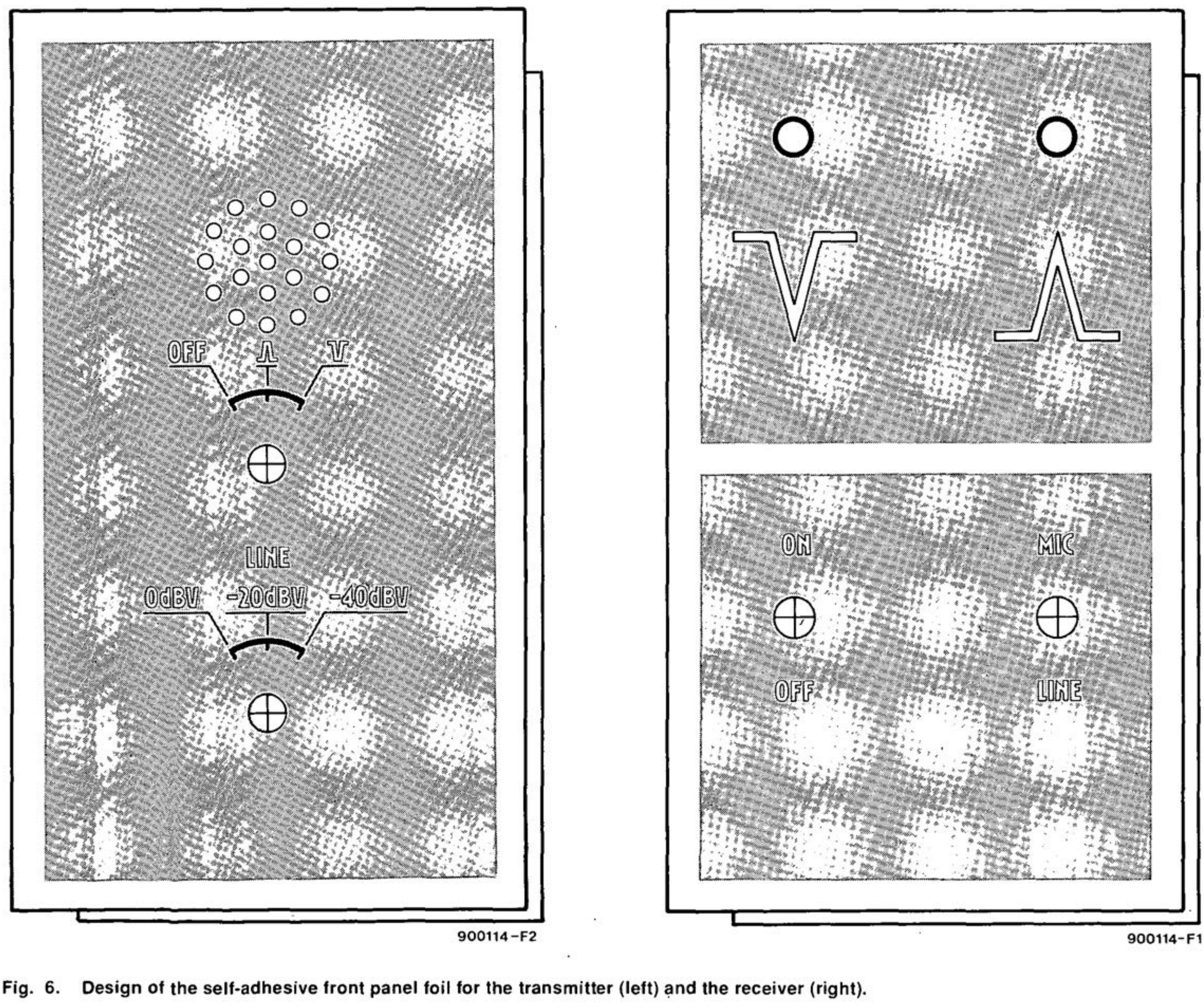

Fig. 6. Design of the self-adhesive front panel foil for the transmitter (left) and the receiver (right).

INTRODUCTION TO METAL TRANSMISSION LINES

by Roy C. Whitehead, C.Eng., MIEE

Transmission lines may be used both for the direct transmission of information and as circuit elements, sometimes substituted for such components as transformers, capacitors and inductors.

THE two main types of metal line are the balanced and the coaxial types as shown in Fig. 1. The familiar pair of wires mounted on porcelain insulators, supported on wooden poles, and the twisted or parallel pairs embedded in solid insulation are shown at (a) and (b) respectively. Where several such pairs are run together, it is customary to employ a physical transposition process, so that mutual interference between pairs, encountered along one length of line, is partially balanced out by reversed interference along another length. Such lines are normally operated in the 'balanced' condition, neither conductor being earthed, although sometimes the centre point of an associated transformer or amplifier may be earthed.

A coaxial line, with its central conductor insulated from its outer conductor is shown at (c). These lines are operated in the 'unbalanced' condition, that is, the outer conductor is earthed. The outer conductor does not always provide a very efficient screen at low frequencies, so in some circumstances signals are confined to the spectrum above 50 kHz.

A very important characteristic of any transmission line is its 'characteristic impedance' Z_0, which is the ratio V/I for a line of infinite length as shown in Fig. 2. But, of course, there is no such thing as a line of infinite length. However, if a line of finite length be connected with a variable resistor R_d to its remote or distal end, there will be one specific value of R_d that produces a constant ratio V/I for all frequencies and for all lengths of that particular type of line. It is around that particular value of R_d, that is, Z_0, that complete telecommunication systems are built, just as railway systems are built upon the 'gauge', or spacing, of the rails (which is 4 ft. 8½ in. = 1435 mm in Britain and many other countries).

The two ends of a line are sometimes referred to as the 'proxal' or sending end and the 'distal' or receiving end. Subscripts p and d respectively will be used accordingly.

The characteristic impedance Z_0 of a line is governed by the ratio D/d shown in Fig. 1 and the value of the permittivity, k, of the dielectric.

Simplified equivalents to balanced and unbalanced lines are shown in Fig. 3. For most practical purposes, the value of Z_0 may be taken as $Z_0 = \sqrt{(L/C)}$, where L is measured with the distal end short-circuited and C with it open-circuited. Details are given in the Appendix.

In Britain, open-wire lines and twisted pairs, singly or in multi-pair cables usually have $Z_0 = 600\ \Omega$. Coaxial cables on the other

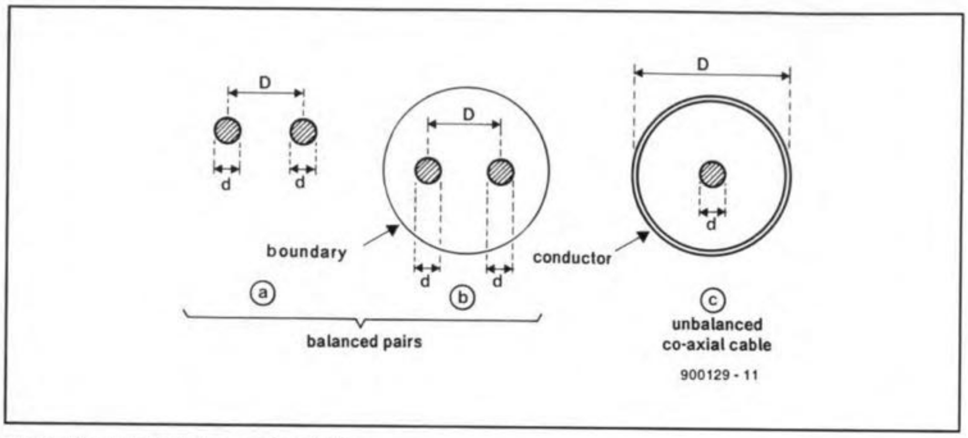

Fig. 1. The two main types of metal line.

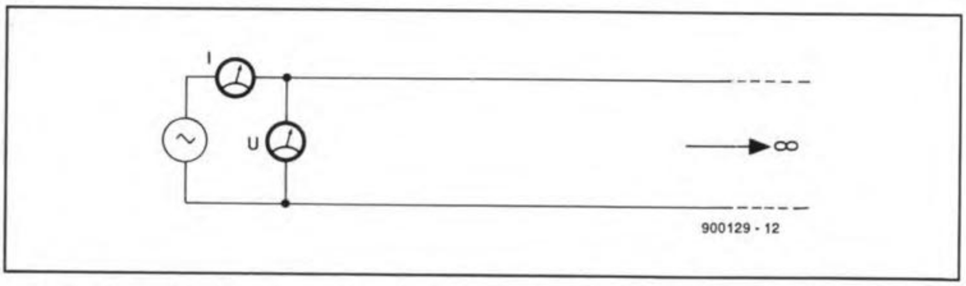

Fig. 2. Testing an imaginary line to determine its characteristic impedance Z_0.

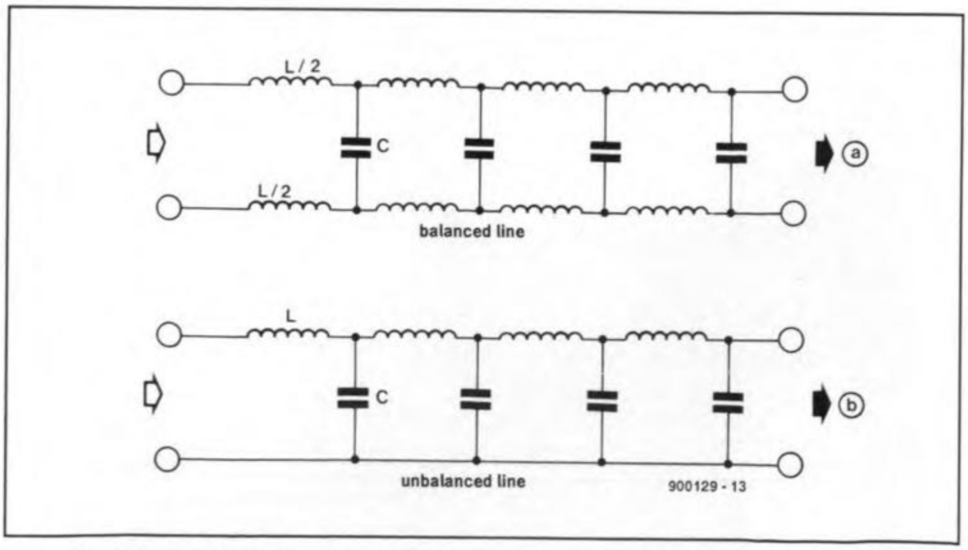

Fig. 3. (a) a balanced line; (b) an unbalanced line.

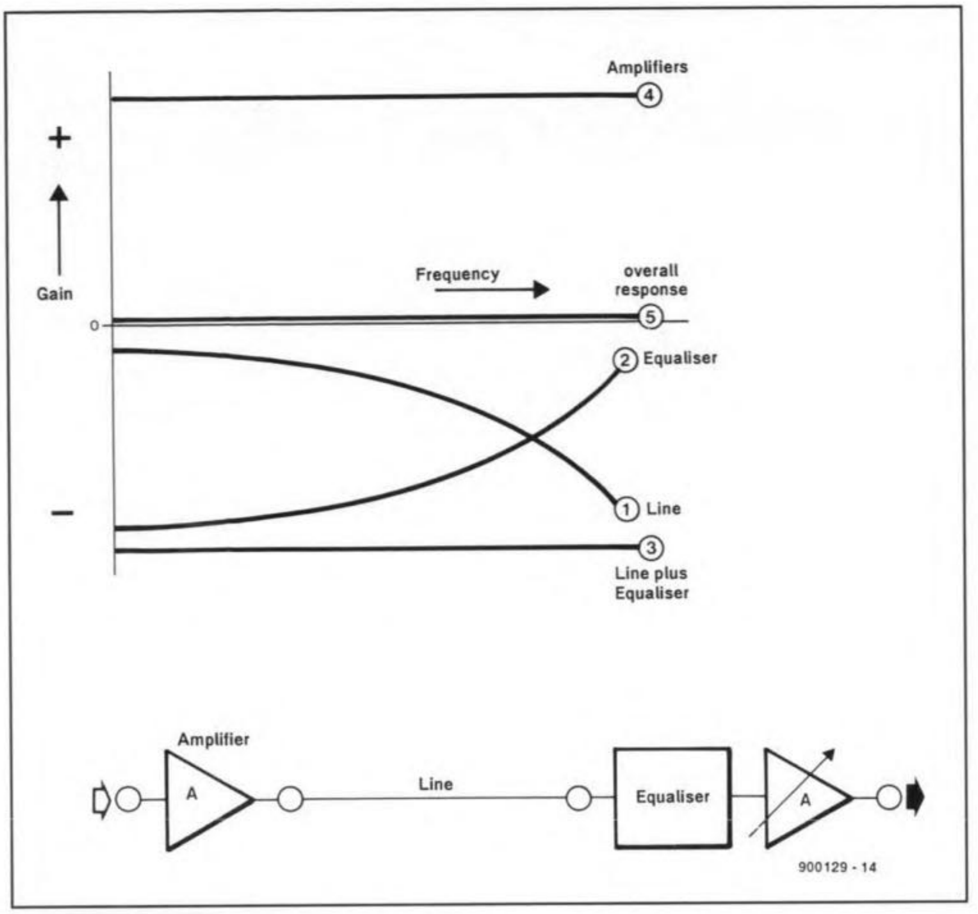

Fig. 4. A simple equalized telecommunication link.

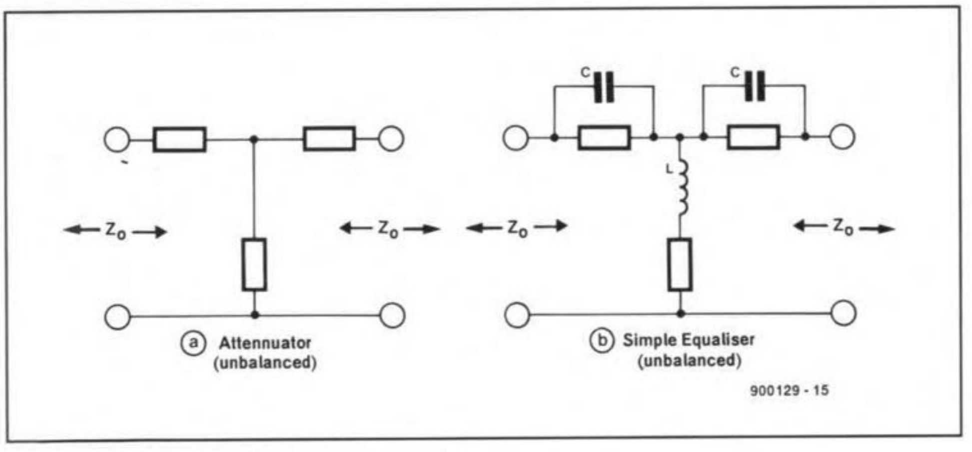

Fig. 5. Attenuators for different frequency bands.

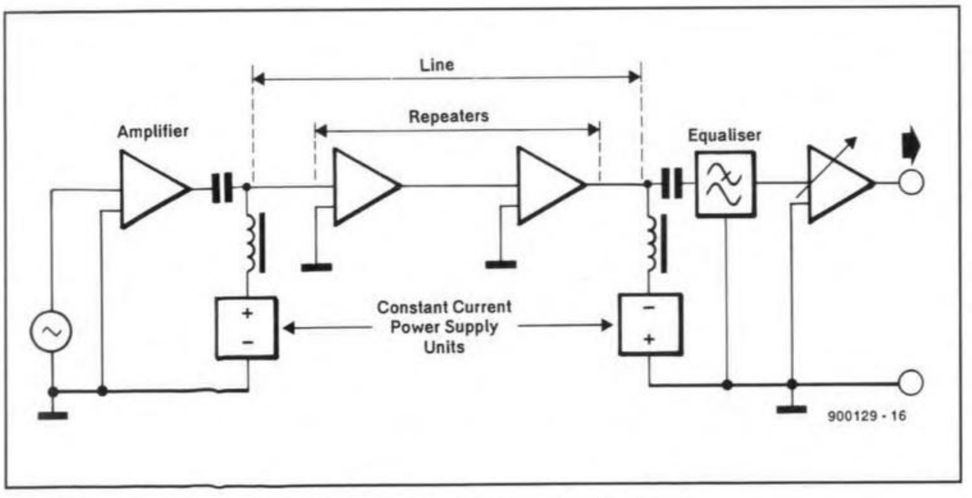

Fig. 6. A repeatered line with power fed to repeaters along the signal line.

hand usually have $Z_o = 75\ \Omega$ or, at very high frequencies, 50 Ω.

The velocity of propagation in an ideal line that has vacuum insulation and no supports, would equal the velocity, c, of an electromagnetic wave in free space, that is, $c = 3\times10^8$ metres/second. For a line with minimum supports and with air as insulation, the velocity is only slightly less. For a line with solid or gaseous insulation that has a permittivity k, the velocity v is $v = c/\sqrt{k}$. The velocity v/c of a line is known as the 'velocity ratio'. This is usually quoted by the manufacturers; if k varies between 1.2 and 2.8, the value of v/c lies between 0.9 and 0.6.

The type of insulation for concentric cables that is most commonly experienced in laboratories or small installations is polytetrafluoroethylene, normally called PTFE. This has the great merit of being flexible.

For high-power transmitters, where very high voltages are incurred, the insulation may be air, nitrogen under pressure, or helium. No, or very few, intermediate physical spacers may be incurred when transmission lines are installed vertically up masts.

The relationship between attenuation per unit length and frequency is given by the empirical equation: attenuation = $a\sqrt{f} + bf$, where $b \ll a$. Up to about 16 MHz, the second term may be ignored, but above that frequency the attenuation rises faster.

The increase in attenuation at high frequencies has two causes. The first is that the losses in the insulation rise with frequency. The second is the well-known 'skin effect' that takes place in conductors that operate at high frequencies. The higher the frequency, the less deep is the penetration of current into the conductor surface. For this reason, conductors that must carry high levels of current at very high frequencies usually take the form of tubes that have conductivities which are equal to those of solid conductors of similar diameter).

Complete transmission links, say between cities, are usually engineered to produce what is known as 'zero equivalent', that is, the combination of attenuation and amplification equals zero decibels. This is to enable communication to be established readily, either directly between two points, or indirectly via other points without change of amplitude of the received signal. To achieve this result, a complete link includes terminal amplifiers to counteract attenuation and 'equalizers' to counteract the variations of attenuation over the frequency band. An example of the various parts of a link is shown in Fig. 4, starting with LINE (1), EQUALIZER (2), and so on. The design of an equalizer starts with the design of an attenuator that has an attenuation which is slightly greater than the variation of attenuation of the line over the operating frequency range. A simple attenuator (for an unbalanced line) is shown in Fig. 5 (a). Reactive elements are then added to reduce attenuation at the higher frequencies as shown in Fig. 5 (b). This produces finally an attenuation/frequency characteristic of line plus equalizer that is approximately flat. The equalizer is located at the receiving end of the line so that it will attenuate not only the lower frequency components of the signal, but also random noise and cross-talk that has been picked up along the line. Finally, a variable-gain amplifier is added to achieve the zero equivalent condition. Along a line there is a limit to the attenuation that can be tolerated between the two terminal amplifiers, otherwise the signal-to-noise ratio of the received signal would be unacceptable. Along a lengthy line, this effect is combated by the introduction, at various stages, of amplifiers that are referred to as 'repeaters'. The power that is required to operate these repeaters is sometimes fed along the signal line as shown in Fig. 6.

The introduction of terminal amplifiers and repeaters implies that such lines can be operated only unidirectionally, so that to enable a conversation to take place, two lines are required. In order that the high-level signal at one end of one line shall not interfere with the low-level signal of an adjacent line that is operating in the reverse direction, two groups of lines are formed physically with screening between them. Each group consists entirely of lines that operate in a given direction as is shown in Fig. 7.

When transmission lines are used for communication, it is usual to operate them between resistive terminations that are equal to the characteristic impedances Z_0 of the lines. This produces an attenuation/frequency curve that is smooth as was shown in Fig. 4, curve number 1, enabling a simple equalizer to be designed as was shown in Fig. 5.

It is customary for telecommunication authorities to specify the maximum amplitude of the signals that may be fed into the lines, which is to avoid overloading the amplifiers and also to minimize cross-talk between the various users. Thus, if all users feed into their lines signals of approximately the same magnitude, the overall signal-to-cross-talk ratios will be maximized.

A line that is terminated with a resistance $R_d = Z_0$ will (ignoring attenuation) have a distribution of voltage and current along its length as shown in Fig. 8 (a).

When electrical energy starts to travel down a line, it does so at a rate that is determined by the details of the generator, the velocity ratio of the line and the value of Z_0. This is said to constitute a 'travelling wave'. If the termination has a value of $R_d = Z_0$ and energy reaches the termination, a stable condition is established and electrical energy is converted into thermal energy at the same rate at which it was admitted to the line. If, however, R_d does not equal Z_0, the termination can no longer disspate energy at that rate, so information is communicated back to the source by a 'reflective wave' to reduce the rate at which energy is admitted. The final result is a combination of the two waves.

An example of how current and voltage are distributed as travelling waves along a line where $R_d = Z_0$, and there are no reflective waves, is shown in Fig. 8 (a). But if the termination is an open circuit, and consequently no current can flow in it, the travelling and reflected current waves are in opposition, resulting in no current at the termination. But the voltage travelling and reflected waves are in phase, resulting in a doubling of voltage there as shown in Fig. 8 (b).

The reverse condition applies to a short-circuited line as shown in Fig. 8 (c).

A compromise condition, where R_d is finite but does not equal Z_0, is shown in Fig. 8 (d). Because the magnitude of U is rising at the approach to the load, it follows that R_d is greater than Z_0.

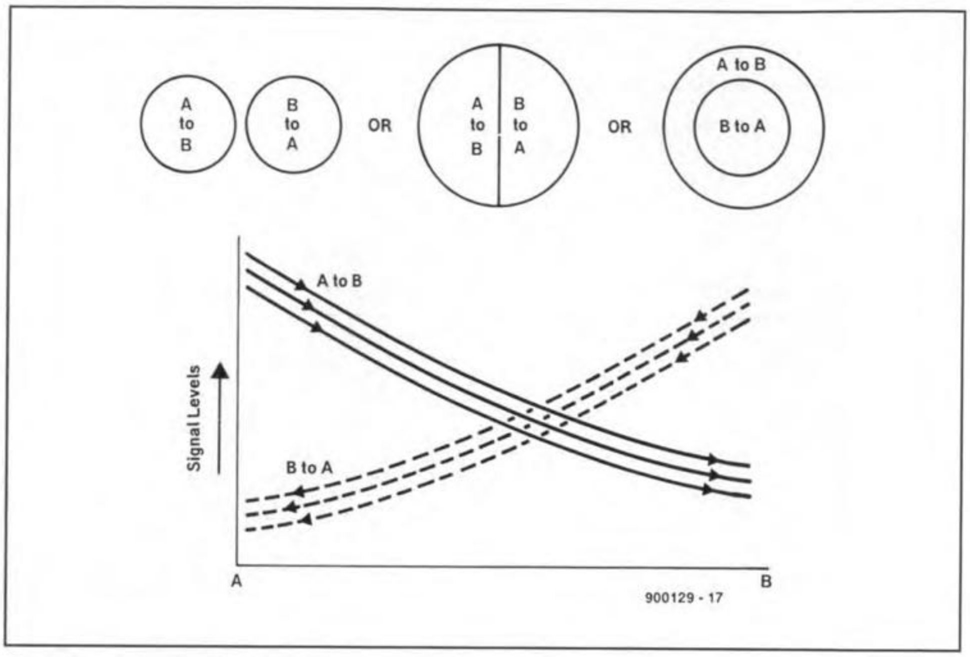

Fig. 7. Signal levels in isolated groups arranged to minimize cross-talk.

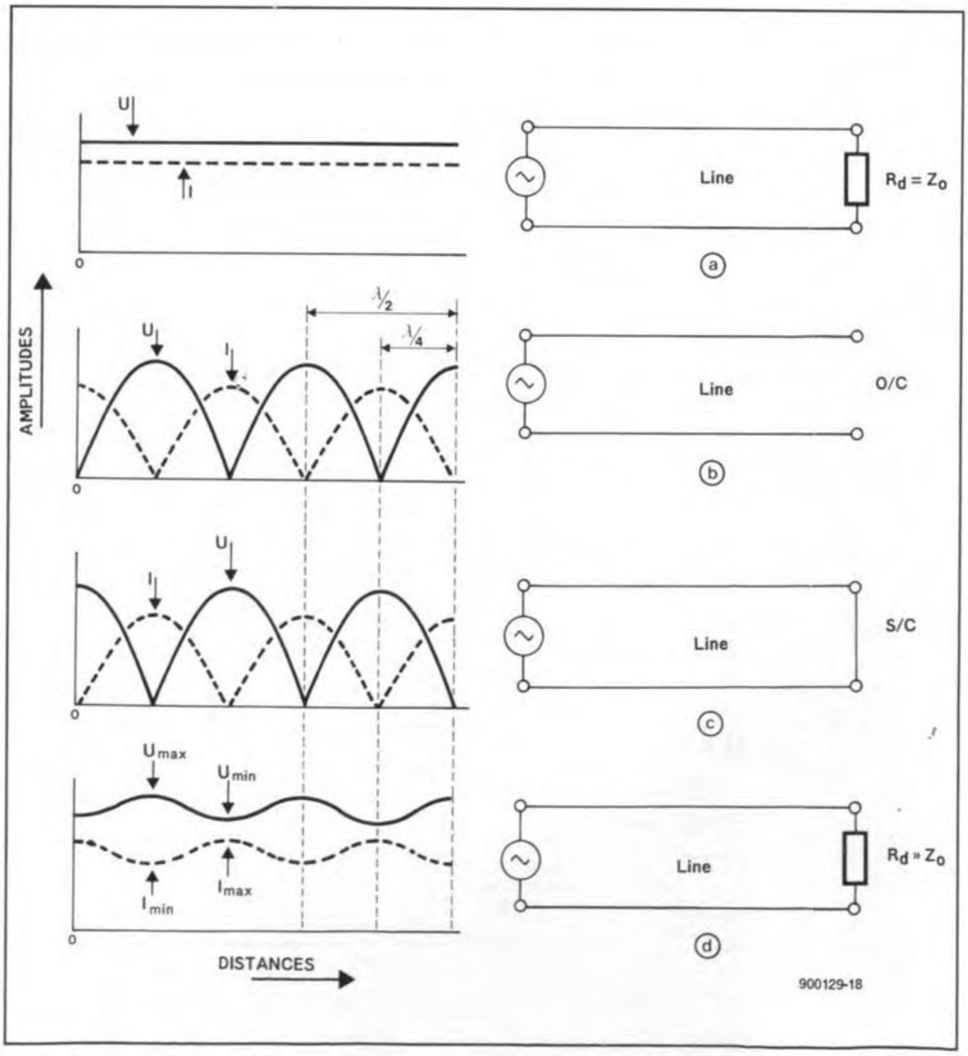

Fig. 8. Distributions of voltage and current along lines having various terminations.

Consider again Fig. 8 (d) and note that in comparison with Fig. 8 (a) the values of U and I vary cyclically along the line. Because the dieelectric and conductor losses are proportional to the squares of U and I respectively, the increased losses around the peaks of the waveforms are not compensated completely by the decreased losses around the troughs. The ratios U_{max}/U_{min} and I_{max}/I_{min} are known as the 'standing-wave ratios'. Considering particularly cases of high-power transmitters feeding aerials, the excess amplitudes may cause breakdowns. However, the small losses of radiated power are usually considered to be significant only in the case of very-high-power transmitters (see References). Standing-wave ratios may be measured using a commercial standing-wave ratio measuring instrument, but a simple test may be carried out with the aid of a neon tube mounted at one end of a stick of insulating material and running this along the line to explore the peaks and troughs.

Having considered the line as a device for the transmission of information, let us now consider it as a circuit element equivalent to, for instance, a capacitor, inductor or transformer.

Consider Fig. 8 (b) and (c). At any point along the axes where U is finite and I equals zero, the impedance looking towards the termination must equal infinity. Where U equals zero and I is finite, the impedance must equal zero. Where U is rising, energy is being converted from kinetic into potential form: the nature of the impedance must therefore be capacitive. But where I is rising, energy is being converted from potential into kinetic form and the nature of the impedance must be inductive. Of special interest are the distributions of voltage and current along quarter- and half-wavelength lines and the extraction of information from Fig. 8 (b) and (c); the distributions for open- and short-circuit conditions are given in Fig. 9.

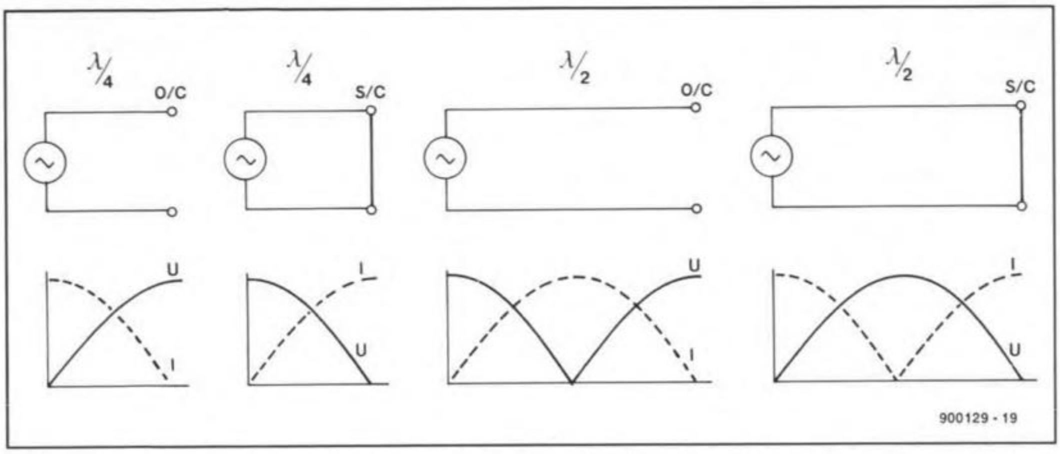

Fig. 9. Distributions of voltage and current along quarter- and half-wavelength lines with open and short-circuit terminations.

Extending considerations to a wider variety of effective lengths, some examples with their equivalents are shown in Fig. 10. Where a choice exists, it is preferable to use a line with a short-circuit rather than an open-circuit termination, since such a line is easier to fix mechanically. Figure 10 (9), which represents a line with a sliding short-circuiting bar, can be adjusted to present a wide variety of such equivalents.

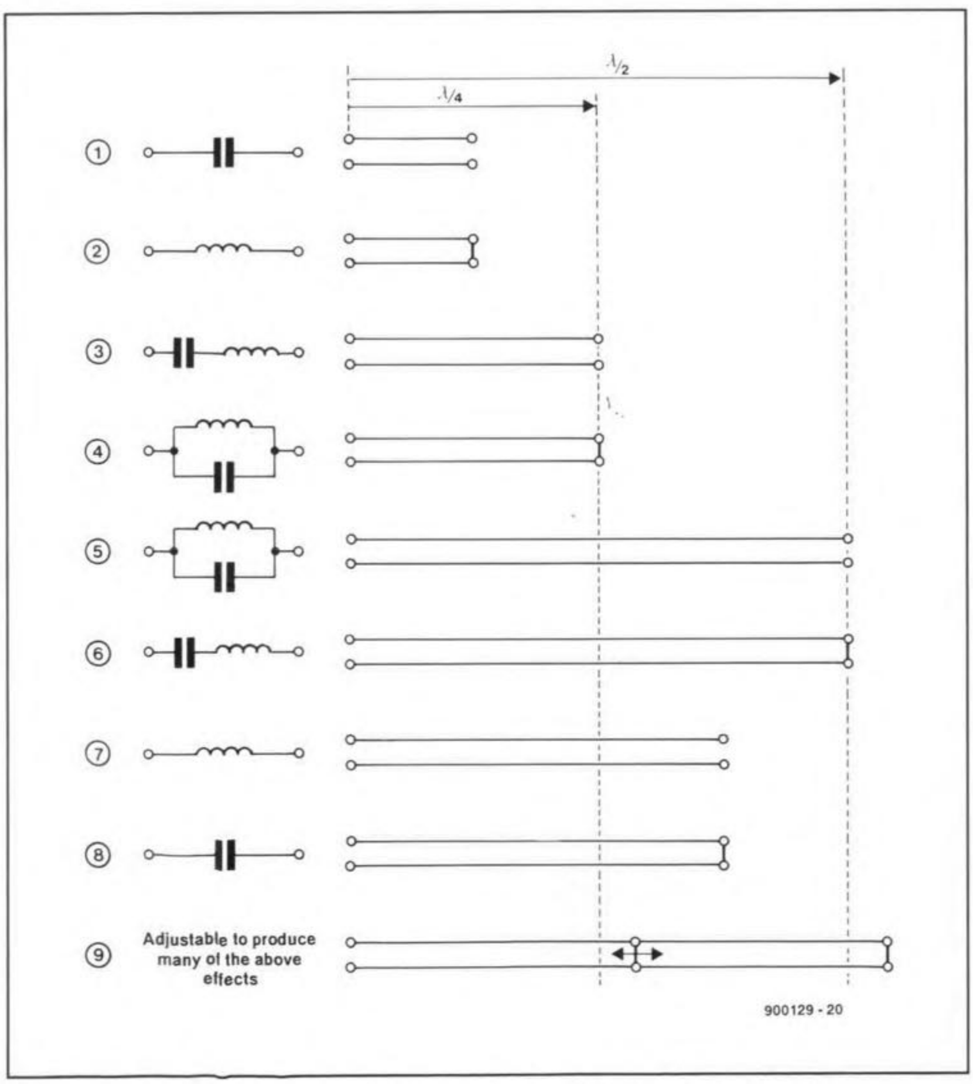

Fig. 10. Lengths of transmission lines being used to produce different reactive equivalents.

Although the lines shown are balanced, similar results may be obtained with the use of unbalanced lines. In the case of Fig. 10 (9) using unbalanced lines, however, similar results require the provision of 'trombones', that is, devices that are similar to their musical equivalents in that they provide paths of variable lengths. It must be emphasized, however, that these effects hold good only over very narrow bands of frequency.

A particular use of Fig. 10 (9) is in transmitting stations that house many high-power transmitters which operate on different frequencies. Power that is radiated from one aerial might be picked up by another aerial and this might affect the operation of the second transmitter. This may be avoided by connecting a line of the type shown in Fig. 10 (9) across the output terminals of each transmitter, adjusting the bridge to produce a quarter-wave condition for that transmitter and earthing the short-circuiting bridge. This technique also provides protection against lightning strikes.

The effects of open- and short-circuit terminations have already been dealth with. Now it is necessary to consider the results of employing various other types of termination.

Since the magnitudes of the voltage and current at both ends of a $\lambda/2$ line are similar, it follows that if an impedance Z_d be connected at one end of such a line, a similar impedance will appear at the other end, that is, the line acts as a 1:1 transformer. This will be so irrespective of the relationship between the values of the load and the characteristic impedance of the lines as is shown in

Fig. 11 (a).

The quarter-wave line has been seen to have an inverting effect. Therefore, if the terminating impedance Z_d has a magntitude that equals nZ_o, the impedance that appears at the other end will have a magnitude Z_o/n as is shown in Fig. 11 (b). But not only is the magnitude of the impedance inverted but so also is the sign, e.g., capacitive reactance is transformed into inductive reactance.

Two practical examples of λ/4 lines used as transformers are shown in Fig. 12. If the line and aerial impedances be Z_o = 600 Ω and Z_{ae} = 75 Ω, the matching impedance will be $Z_m = \sqrt{(600\times75)}$ = 212 Ω.

One of the equations in the Appendix gives for a balanced line:

$$D/d = {}^1/_2 \text{ antilog } (Z_o\sqrt{k}/276) =$$

$$= {}^1/_2 \text{ antilog } (212/276) =$$

$$= {}^1/_2 \text{ antilog } 0.768 = 2.93$$

That is, the spacing/diameter ratio will be 2.93 to provide the transformation. The line conductors may then be sqeezed closer together to just under three times their diameter as shown in Fig. 12 (a). This, however, might bring them dangerously close together. The alternative would be to maintain the same spacing for the transformer as for the line and to construct the transformer from tubes that have diameters of one third of the 600 Ω line spacing as shown in Fig. 12 (b).

It is interesting to consider whether the presence of standing waves could present a problem on power supply transmission lines operating at 50 Hz. The worst possible case would be where the line represented the λ/4 condition:

$$\lambda/4 = V/4f \quad \text{[metres]};$$

and allowing for a velocity ratio of 0.8:

$$\lambda/4 = 3\times10^8\times0.8/(4\times50) \quad \text{[metres]} =$$

$$= 1200 \text{ km},$$

which is the distance from, say, London to Madrid, Venice or Oslo. Therefore, even with 60 Hz mains supplies, there is little or no trouble likely to be experienced even with very long lines.

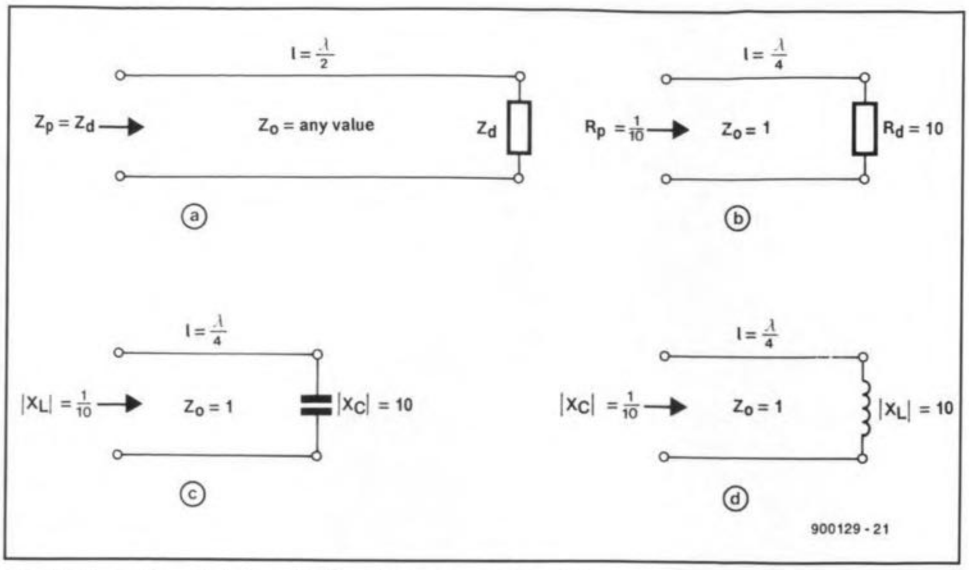

Fig. 11. Some characteristics of half- and quarter-wavelength lines.

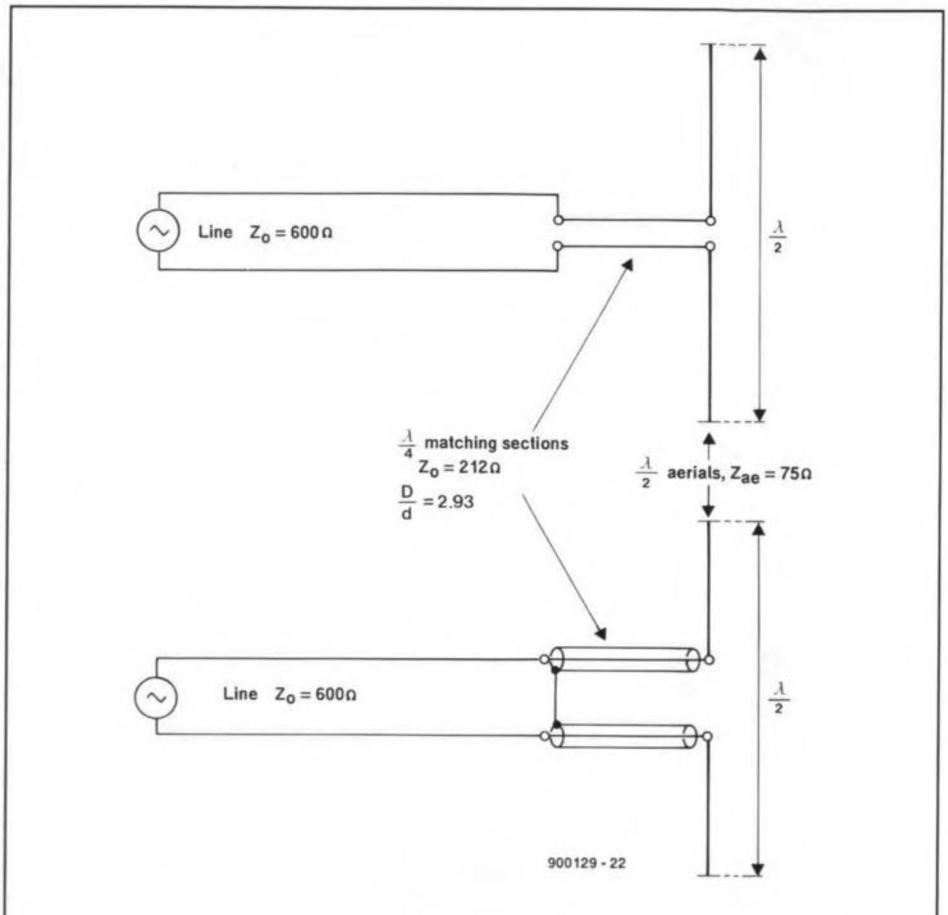

Fig. 12. Matching a 600 Ω line to a 75 Ω aerial. Alternative transformers.

Appendix

Consider the line formed by a drum of cable or two identical lines that are looped at the distal end, represented by inductors and capacitors as shown in Fig. 13. The two variable resistors are ganged and always have equal values of resistance. The generator voltage is kept fixed and its frequency is varied. *R* is then varied to produce a constant value of *U*. Then, $Z_o = R$.

The kinetic energy stored in the inductors will be ${}^1/_2 LI^2 = {}^1/_2 LV^2/R^2$, which is dependent on the value of *R*.

Equal energy will be stored in capacitances and inductances when the value of *R*

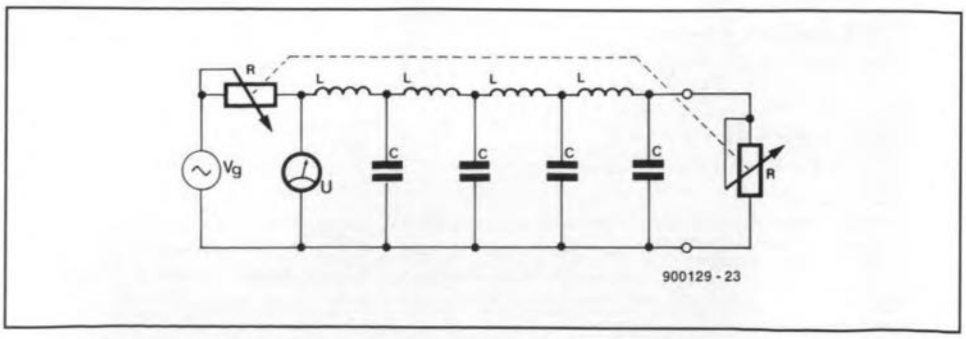

Fig. 13. Using a generator of slowly varying frequency but constant voltage, *R* is adjusted to produce a constant value of voltage *U*. Then, $R = Z_o$.

is such that

$$\tfrac{1}{2}CV^2 = \tfrac{1}{2}LV^2/R^2$$

$$\therefore C = L/R^2;$$

and $R = \sqrt{(L/C)}$.

This value of R is termed the 'characteristic impedance' and is given the symbol Z_0.

A fuller equation, taking into account the series resistance R of the inductors and the shunt leakance G (in siemens) of the insulation, is

$$Z_0 = \sqrt{\frac{R + j\omega L}{G + j\omega C}} \quad (\Omega)$$

where $j = \sqrt{-1}$ and $\omega = 2\pi f$.

However, as $R \ll \omega L$ except at very low frequencies, and $G \ll \omega C$ except at very high frequencies, the simpler equation is normally accepted as adequate for practical purposes.

It is also possible to determine the value of Z_0 from the physical construction of the line by the use of one of the following two equations.

For a twin-wire line:

$$Z_0 = \frac{138 \times 2}{\sqrt{k}} \times \frac{\log 2D}{d} \quad (\Omega)$$

or

$$\frac{D}{d} = \frac{1}{2}\,\text{antilog}\,\frac{Z_0\sqrt{k}}{276}$$

where D is the distance between the centres of the conductors, d is the wire diameter, and k is the permittivity of the insulation (= unity for air).

For a concentric line:

$$Z_0 = 138/\sqrt{k} \times \log (D/d) \quad (\Omega)$$

or

$$\frac{D}{d} = \text{antilog}\left(\frac{Z_0\sqrt{k}}{138}\right)$$

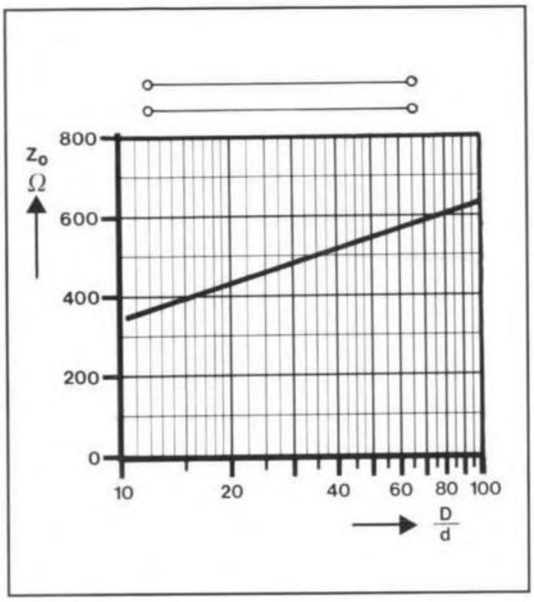

Fig. 14. Balanced lines with air dielectrics: divide Z_0 by $\sqrt{k}$ for other dielectrics.

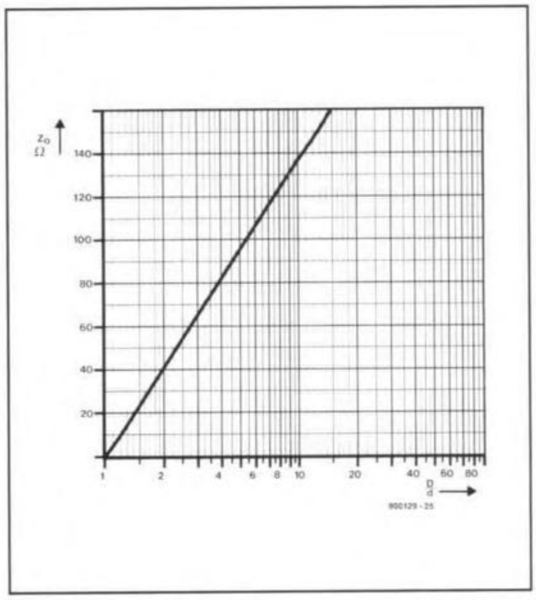

Fig. 15. Coaxial lines with air dielectrics: divide Z_0 by $\sqrt{k}$ for other dieelectrics.

References:

Electronics Communication Systems by George Kennedy, Mc-Graw-Hill Book Company, Gregg Division.

Amateur Radio Techniques by P. Hawker, RSGB.

HF Antennas by L.A. Moxon, RSGB.

1987 ARRL Handbook.

VHF UHF Manual, RSGB.

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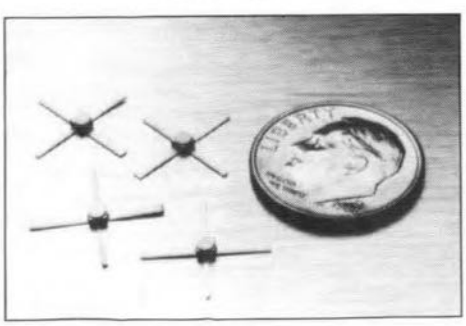

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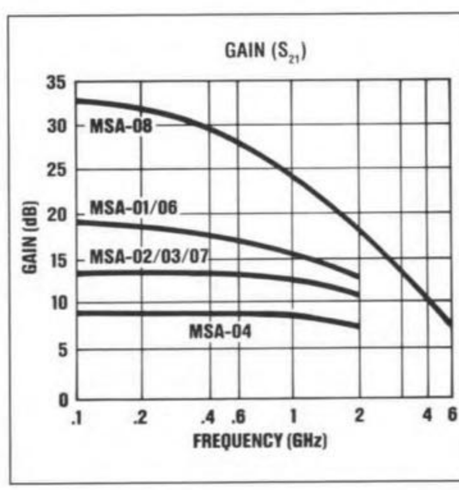

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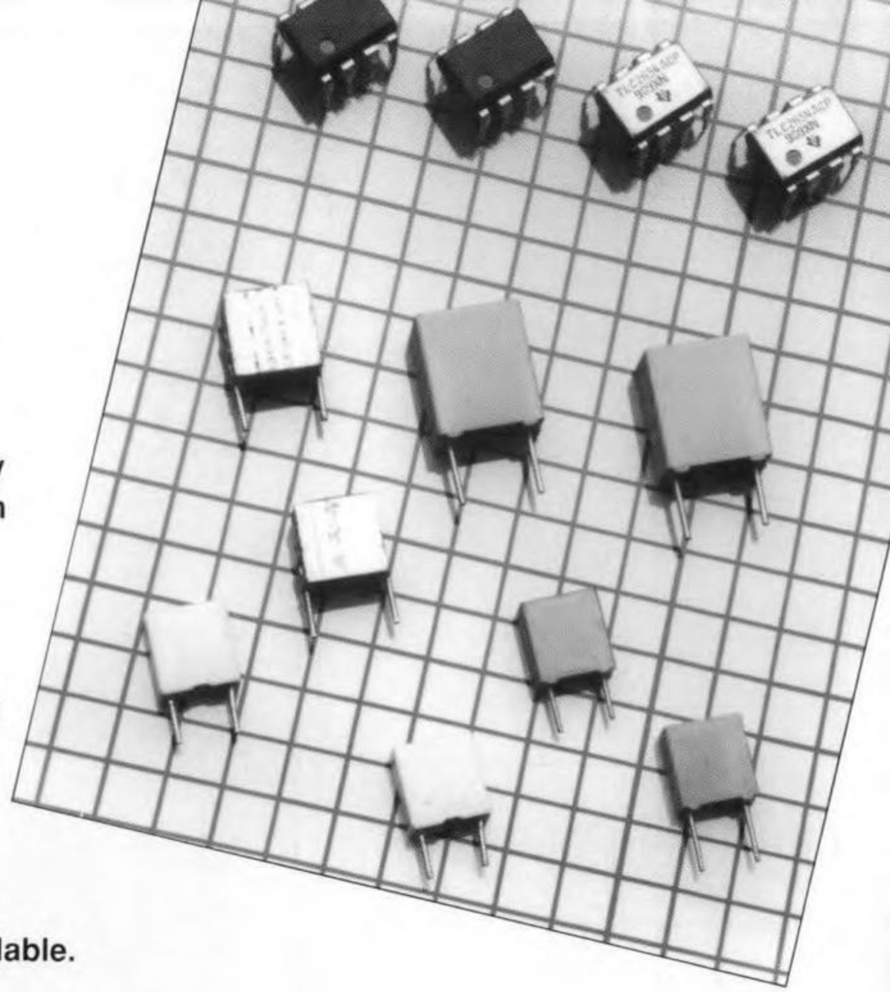
CHOPPER-STABILIZED OPERATIONAL AMPLIFIERS



Chopper-stabilized opamps are in many cases the only feasible alternative when we want to amplify very small direct voltages. In this article we will explore why chopper opamps have such excellent d.c. characteristics. A theoretical background to the operation of interesting new devices is given, followed by a discussion of some inherent problems (and, of course, proposed solutions). The article is closed off with an overview of the most popular chopper opamps currently available.

by J. Ruffell, with contributions from B. Marshall (Texas Instruments) and G.J. van Os (Acal Auriema)

FOR a long time to come, instrumentation amplifiers will be required to operate at the highest possible accuracy. This expectation is based on the trend towards ever higher resolution of DACs (digital-to-analogue converters) and ADCs (analogue-to-digital converters). It will be clear that high resolution in a measurement is not achieved just by the use of converters with a high resolution. After all, it makes little sense to perform a measurement at an accuracy of 18 bits when the analogue amplifier used has a maximum resolution of, say, 16 bits. In practice, the accuracy of the hardware for analogue signal conditioning must be doubled for every additional bit to be measured.

Analogue signals are preferably conditioned and/or amplified by a.c.-coupled circuits, mainly because these can be built by relatively simple means and at low cost. There are, however, many applications where the wanted signal is applied in the form of a direct voltage or a direct current. Devices used in such applications include thermocouples, photodiodes and, on a larger scale, the digital multimeter, which is an example of a data acquisition system. Since these devices and circuits can only be d.c. coupled, the designer is faced with off-set voltages and drift of the linear amplifier he intends to use. The origins of input off-set voltages and their stability is discussed in an earlier article on new opamps, see Ref. 1.

Although conventional operational amplifiers such as the OP07 and the OP77 are good choices for d.c. signal conditioning, there are devices whose extremely low drift and off-set voltage make them far better suited to the application. The type of operational amplifier we have in mind is generally referred to as a chopper opamp, or, more accurately, a chopper-stabilized opamp.

Chopping: the classic approach

During the valve era, the terms chopper amplifier and indirect d.c. amplifier were familiar to almost anybody in the field of electronics. At that time, chopping was taken very literally. A kind of electronic guillotine was used to convert the low-frequency alternating voltage (or the direct voltage) to be amplified, into a signal with a higher frequency. Next, this 'high-frequency' signal was raised in an a.c. coupled amplifier, and subsequently restored to its original frequency by a synchronous detector. In practice, the chopping element used to be a relay or, a little later, a bipolar transistor or a FET.

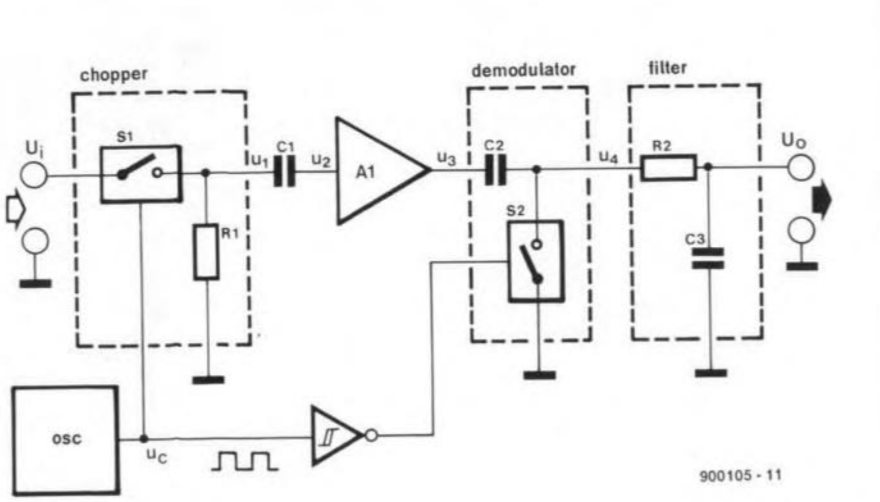
Figures 1a and 1b show the basic schematic of a classic chopper amplifier and the associated waveforms. The input voltage, U_i, is converted to a pulsating waveform, u_1, by switch S_1. The d.c. component is removed before u_2 is amplified by a.c. coupled amplifier A_1. It will be clear that the original waveform (with a higher amplitude) must be recovered from u_3. The recovering, or demodulation, of u_3 is effected by switch S_2. This electronically operated switch connects the right-hand side of capacitor C_2 to ground on every second half-cycle of the oscillator signal. The waveform of u_4 indicates that the switching results in a shift of the direct voltage level. Finally, an integrating filter recovers the amplified voltage, U_o, from u_4.

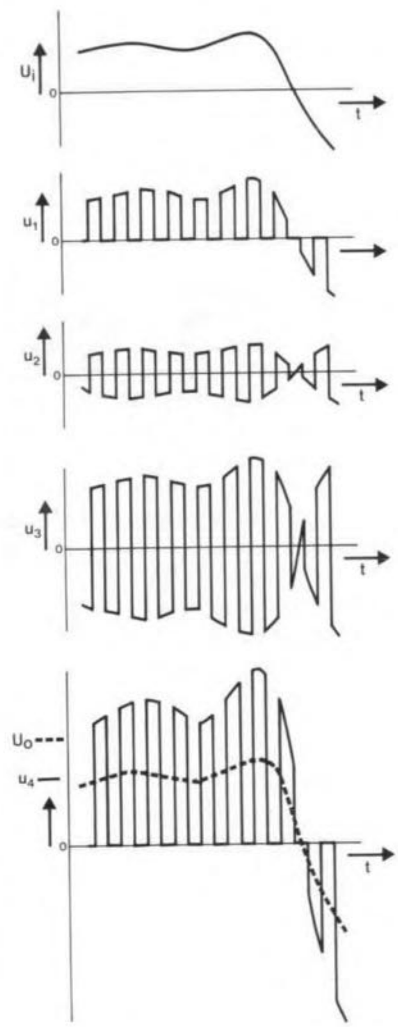
Although this type of amplifier allows good drift specifications to be achieved, it suffers from a number of inherent shortcomings. The chopper, for instance, often introduces glitches at the output. Also, the amplifier lacks a differential output, while its bandwidth is limited to a few hundred hertz.

Integrated

Modern chopper opamps no longer work as described above. These days, the signal to be amplified is no longer chopped to pieces and then rebuilt. Instead, use is made of a control loop which compensates the input off-set voltage of a normal differential amplifier. As a result, these new circuits look quite similar to the standard opamps you have grown accustomed to in many circuits in this magazine.

Chopper opamps, like standard opamps, have a differential input circuit. Because of this likeness, and because their principle of operation is based on the old chopper model, the new devices are generally called chop-





per-stabilized operational amplifiers, or chopper opamps. A typical application circuit of a chopper opamp is shown in Fig. 2.

Fig. 1. Schematic diagram of a classic chopper amplifier (1a), and the waveforms pertaining to this type of circuit (1b).

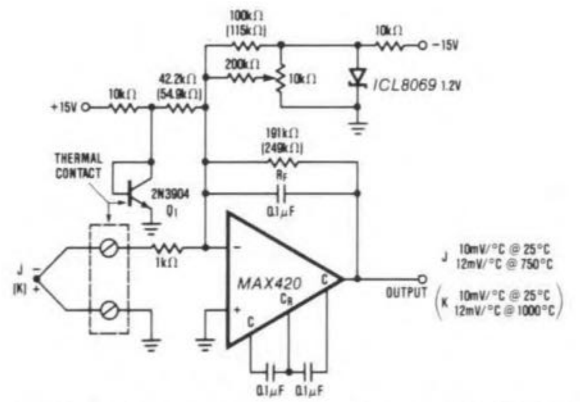
Automatic off-set compensation

The off-set compensation control applied with chopper opamps is in many ways similar to a technique used to compensate the input off-set voltage, U_{OS}, of a standard opamp. This technique entails off-set compensation by fitting a voltage source that supplies $-U_{OS}$ in series with the non-inverting input of the opamp (see Figs. 3 and 4). Automatic input off-set voltage compensation thus requires a circuit capable of measuring U_{OS}, and supplying an accurate 'negative copy', $-U_{OS}$, at the non-inverting input.

You may start wondering at this point how U_{OS} can be measured when the opamp is already part of an existing circuit. Assuming that a simple electronic circuit is used, it can be shown that the input off-set voltage is best measured between the input terminals of the opamp in question. Figure 5 shows how this is done in an inverting amplifier set up around the ideal opamp model. Equation 1 describes the voltage between the non-inverting and the inverting input of the opamp. True, the equation looks fairly complex. However, assuming for the moment that U_i does not contain an alternating voltage component, you will easily discover that the expression in equation 1 is virtually equal to $-U_{OS}$. This is because the open-loop gain, A_{OL}, is high (say, 100,000), so that E (see equation 2) approaches 1. The upshot is that equation 1 can be simplified to give equation 3. The output voltage is approximated as described by equation 4.

The schematic in Fig. 6 shows a circuit designed on the basis of the above discussion. An auxiliary amplifier is used to measure and compensate the input off-set voltage of the main opamp. Equation 5, which describes the output voltage, indicates that the effect of the input off-set voltage is reduced by a factor of $1-E$. Assuming an open-loop gain of 100,000, and $R_1 = R_2$, the reduction amounts to no less than 50,000 times. Compared to the off-set error of about $2U_{OS}$ in the output signal of the circuit in Fig. 5, a specification of the order of $1/25{,}000U_{OS}$ is quite impressive for the circuit in Fig. 6. Thus, equation 6 may be applied with confidence for d.c. applications.

It should be noted that the off-set of the opamp can only be compensated successfully if the auxiliary amplifier is sufficiently compensated. This is why we have shown the auxiliary amplifier as an ideal device, i.e. an opamp without input off-set. It will be clear that such a device does not exist. And yet, the circuit can be extended in a way that does allow automatic off-set compensation to be achieved. Basically, the auxiliary am-



Note 1: Q_1 and connection terminals must be at the same temperature.
Note 2: Values in parentheses are for type K thermocouple.
Note 3: Connections to inverting input of op-amp should be kept as short as possible to reduce noise pickup.
Note 4: All circuit power is ±15V.

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Fig. 2. Typical application of a chopper-stabilized opamp in a thermocouple amplifier with cold-junction compensation (illustration courtesy Maxim).

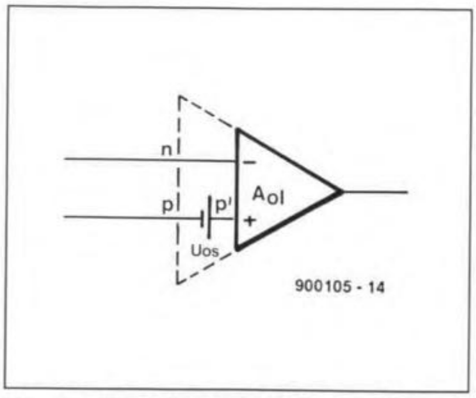


Fig. 3. **Operational amplifier model with input off-set voltage U_{os}.**

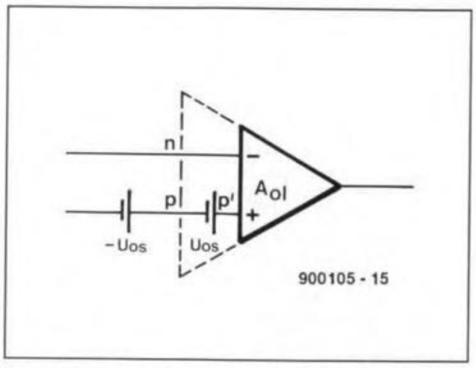


Fig. 4. **The input off-set voltage may be compensated by placing a voltage source $-U_{os}$ in series with the non-inverting input.**

$$(u_p - u_n) = -\frac{(1-E)\cdot R_2}{R_1+R_2}\cdot u_i + E\cdot U_{os} \qquad \text{Eq. [1]}$$

$$E = \frac{1}{1+\dfrac{R_1+R_2}{A_{ol}\cdot R_1}} \qquad \text{Eq. [2]}$$

$$(u_p - u_n) = -U_{os} \qquad [A_{ol}\to\infty] \qquad \text{Eq. [3]}$$

$$u_o = E\cdot\left\{\left(1+\frac{R_2}{R_1}\right)\cdot U_{os} - \frac{R_2}{R_1}\cdot u_i\right\} \qquad \text{Eq. [4]}$$

$$u_o = E\cdot\left\{\left(1+\frac{R_2}{R_1}\right)\cdot(1-E)\cdot U_{os} - \frac{R_2}{R_1}\cdot(2-E)\cdot u_i\right\} \qquad \text{Eq. [5]}$$

$$u_o \approx -\frac{R_2}{R_1}\cdot u_i \qquad [A_{ol}\to\infty] \qquad \text{Eq. [6]}$$

plifier must measure and compensate its own input off-set voltage before handling the off-set of the main opamp. The necessary extensions are shown schematically in Fig. 7.

Off-set compensation thus consists of two successive phases. During the first phase, the electronic switch, S_1, is set to position A. This causes the inputs of the auxiliary opamp to be short-circuited, so that the output voltage of this amplifier is virtually equal to its own input off-set voltage, U_{os1}. Just before S_1 switches to position B, a sample-and-hold circuit, S&H-1, connects U_{os1} in series with the inverting input of the auxiliary amplifier. This results in compensation of the off-set error of this amplifier at the start of the second phase. During the second phase, S_1 connects the positive input of the auxiliary amplifier to the positive input of the main opamp. This, in fact, creates the circuit in Fig. 6. The sample-and-hold circuit still compensates the off-set of the auxiliary amplifier, whose output is at a potential of practically $-U_{os2}$. To retain this voltage, a second sample-and-hold, S&H-2, is introduced. As shown in Fig. 7, this causes $-U_{os2}$ to be connected in series with the non-inverting input of the main opamp. At least in theory, the result is as may be expected: the input off-set voltage is automatically compensated.

The off-set compensation of the two amplifiers may be optimized by repeating the two phases. Depending on the repeat rate, input off-set drift as a result of temperature changes or supply voltage fluctuations may be eliminated, preventing these factors from

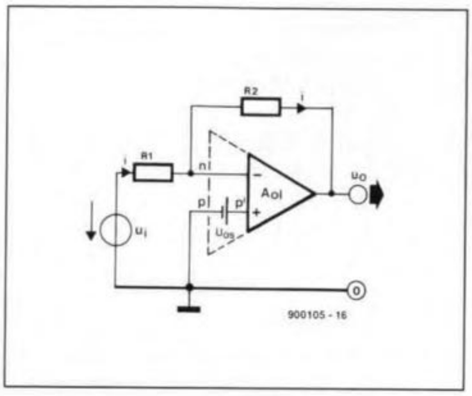


Fig. 5. **This basic circuit allows us to prove, by calculation, that the voltage difference between the inverting and the non-inverting input of the opamp is practically equal to $-U_{os}$ if u_i does not contain an alternating voltage component.**

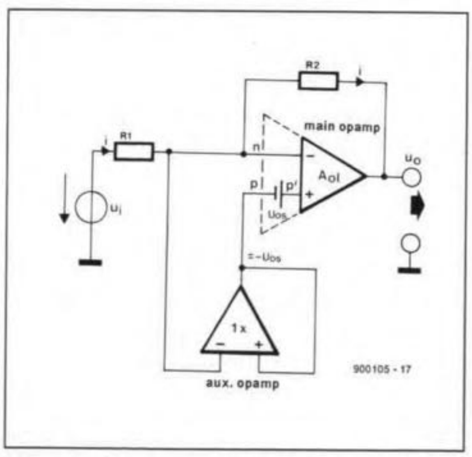


Fig. 6. **First design of a control circuit for automatic compensation of the off-set voltage.**

affecting the stability of the instrumentation amplifier.

Main amps and null amps

The above information will, no doubt, enable you to take a well-prepared look at the block schematic diagram of a chopper-stabilized opamp. The functional diagram used by most manufacturers is shown in Fig. 8. The term main amp refers to the main operational amplifier, while the term null amp is meant to identify the auxiliary amplifier. The switches and the oscillator should not surprise you by now. The two sample-and-hold circuits are not so easily discovered, because they appear in the form of two capacitors, C_A and C_B. The only new blocks are a clamping circuit and a circuit to suppress intermodulation. These two sub-circuits are of vital importance to a good chopper opamp, and their function will therefore be reverted to a little further on in this article.

During the first phase, also called the clock phase, the null amp compensates itself. Switch S_1 is closed, and short-circuits the amplifier inputs. The output voltage is stored in external capacitor C_A via switch S_{1a}. Since there is no input signal, the voltage on C_A is equal to the input off-set voltage of the null amp. Furthermore, the capacitor voltage is fed back to an additional inverting input, so that the off-set error of the null amp is eliminated. During the second period of the clock signal, switch S_2 is closed, and S_1 is open. The null amp then measures the input off-set voltage of the main opamp, and stores it in capacitor C_B. At the same time, the measured voltage is applied to the non-inverting input of the main amp, so that the input off-set voltage is compensated. Thus, the system compensates U_{os} of both amplifiers at the rate of the clock- or chopper-frequency, f_C.

It will be noted that the chopping operation is effected only by the main opamp. The glitches mentioned at the close of the section on the classic chopping amplifier are virtually absent with chopper opamps because the amplified signal is always passed via the continuously operating main opamp.

Recovery time

The decision to use chopper opamps in a practical circuit instead of standard opamps may lead to some surprising problems. First, chopper opamps typically require a much longer time to recover from an overdrive condition, which may occur, for instance, when the output circuit is driven into saturation. Saturation occurs readily and is perfectly normal in, for instance, a comparator circuit.

After an overdrive condition, the main amp no longer works as a linear amplifier. As a result, the voltage difference between the inverting and the non-inverting input is large relative to U_{os}. The auxiliary opamp responds to this condition by charging the two capacitors, C_A and C_B, to the maximum level, i.e., the supply voltage. Inevitably, the main opamp requires some time to remove these capacitor charges when the overdrive condition is passed. In the datasheets, the discharge time is referred to as the overload recovery time. For a conventional opamp, this time is about 10 µs. A chopper opamp, however, may need up to 4 s to recover.

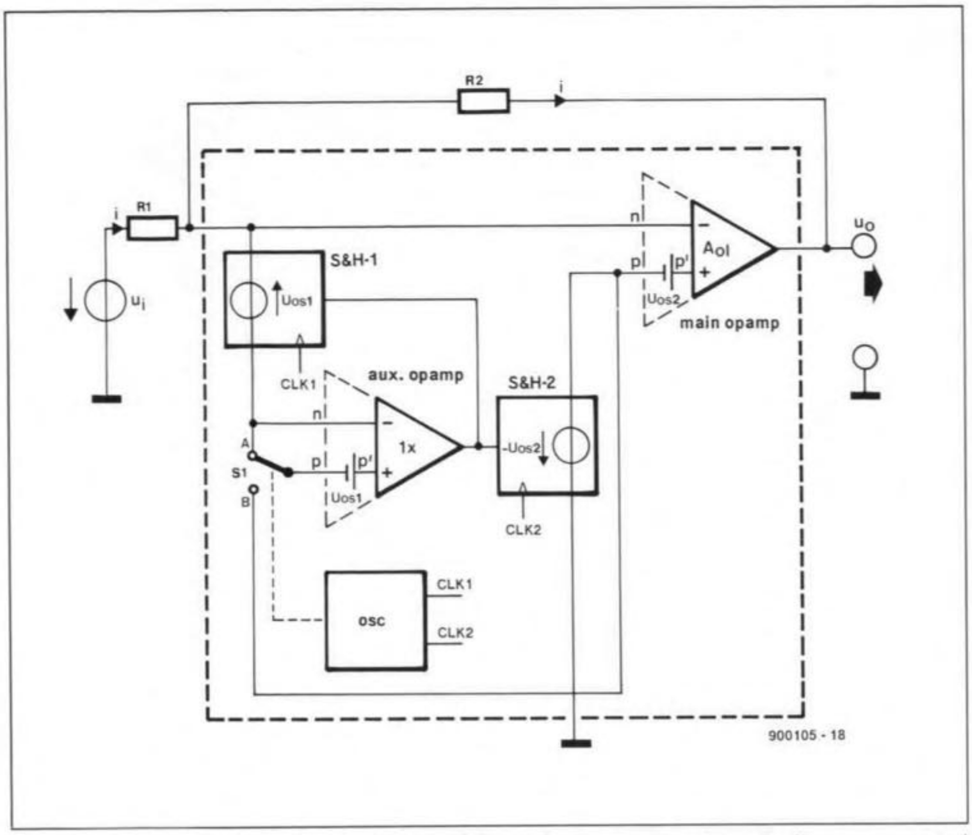


Fig. 7. In this circuit, the input off-set of the main opamp is automatically compensated during two phases of the clock signal.

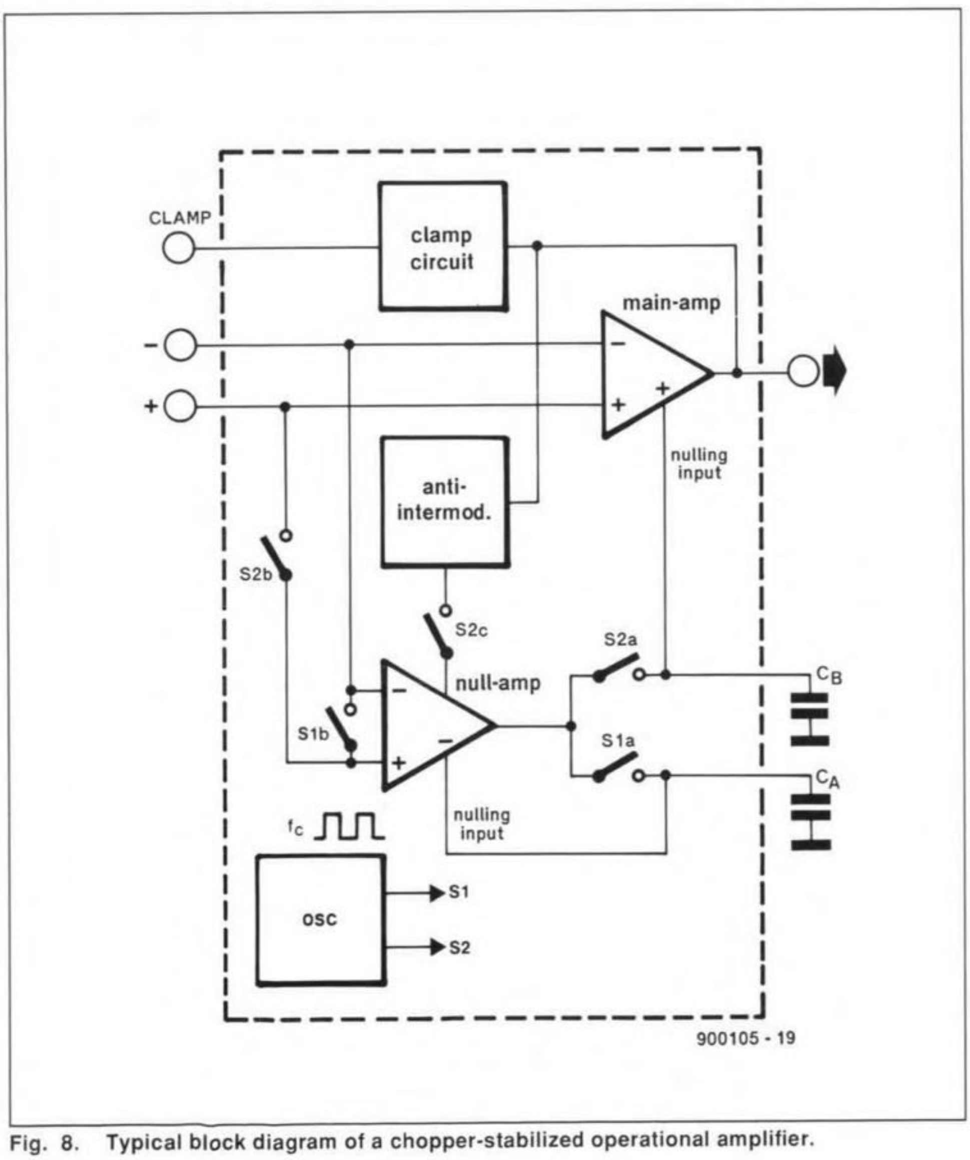


Fig. 8. Typical block diagram of a chopper-stabilized operational amplifier.

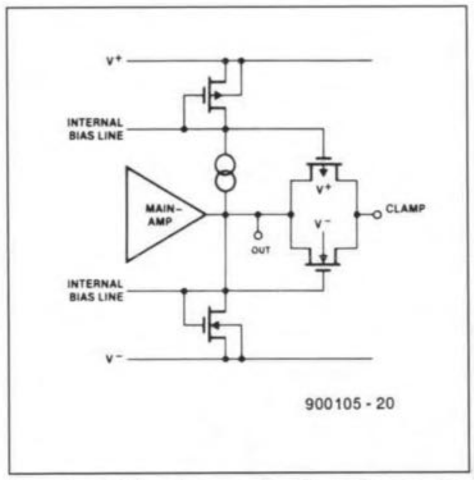


Fig. 9. This clamp circuit reduces the overload recovery time of the ICL7650.

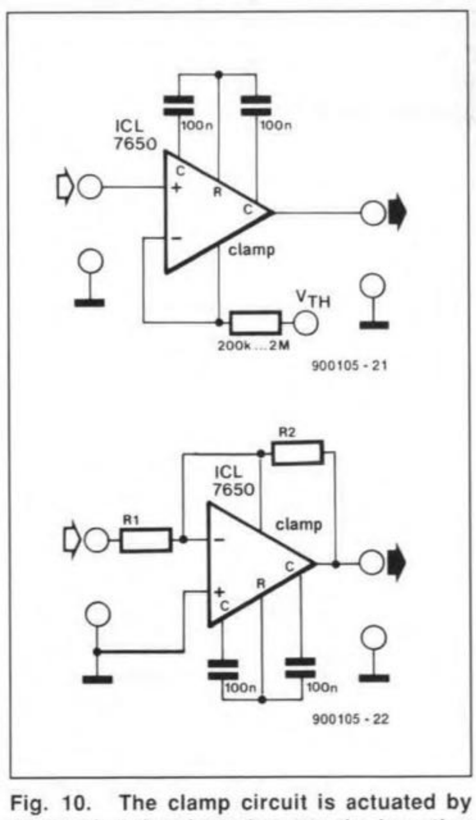


Fig. 10. The clamp circuit is actuated by connecting the clamp input to the inverting input of the opamp. Figure 10a shows a comparator with very low off-set, and Fig. 10b an inverting direct voltage amplifier.

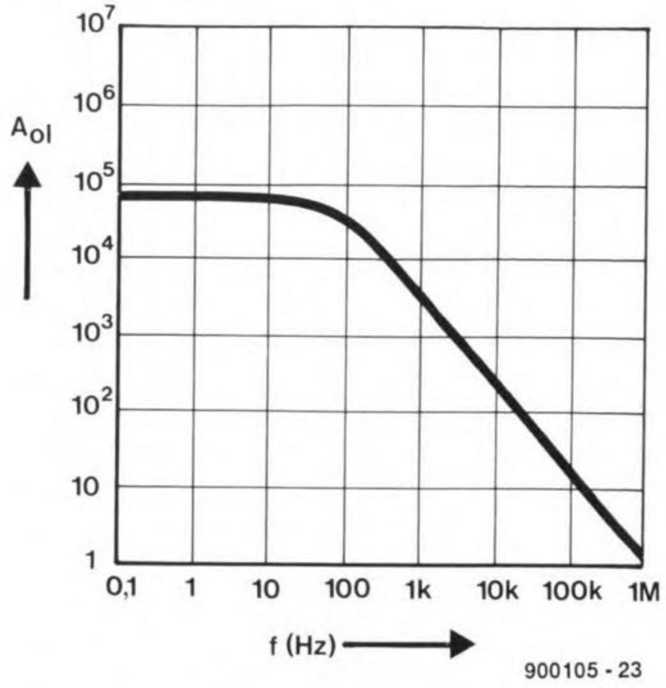


Fig. 11. Open-loop gain, A_{ol}, as a function of frequency.

The clamp circuit provided in the latest chopper opamps serves to reduce the recovery time. The ICL7650, manufactured by Maxim and Teledyne, for instance, has a recovery time of only 300 ms. The clamp circuit used in this chip is shown in Fig. 9. The circuit is actuated by connecting the clamp terminal to the inverting input of the amplifier. Figure 10 shows two circuits that make use of this option.

The clamp circuit is really quite simple, and consists of a mere switch that closes automatically when the output voltage is too close to the supply voltage. When that happens, the switch shunts the externally connected feedback resistor, so that the amplification is reduced. The clamp thus effectively prevents the amplifier being driven into saturation. The very latest chopper opamps have an additional circuit that limits the voltage across the sample-and-hold capacitors. The result is an even shorter recovery time—Texas Instruments' TLC2652, for instance, has a recovery time of only 40 ms.

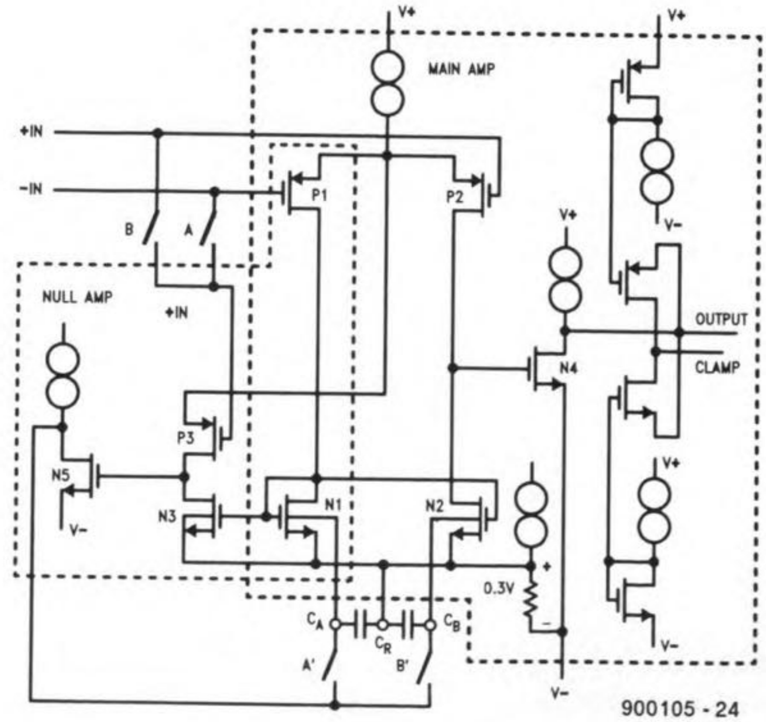


Fig. 12. Simplified internal diagram of the LMC688.

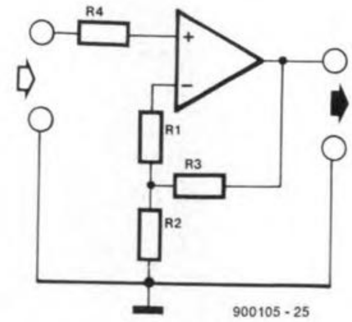


Fig. 13. Resistor R4 is normally superfluous, but it is fitted here to ensure a thermal balance at the input of the circuit.

Next problem: intermodulation

A further problem with chopper opamps may not be noticed until you are dealing with alternating voltages. Unfortunately, an alternating input voltage may cause unwanted sum and difference frequencies because it is mixed with the clock signal. The cause of this annoying effect, called intermodulation, can be traced back to the fact that the voltage between the inverting and the non-inverting inputs of the opamp corresponds closely to the off-set voltage. It should be noted, however, that this is valid for direct voltages only, when the main opamp has a very high open loop gain, and equation 1 may be replaced by equation 3. As soon as an alternating voltage is applied to the opamp, the open-loop gain drops rapidly, as shown by the graph in Fig. 11.

Equation 1 allows us to deduce that the limited value of A_{ol} in $(u_p - u_n)$ also includes a part of the input signal:

$$\frac{(1-E)\cdot R_2}{R_1+R_2}\cdot u_i$$

Furthermore, this part increases with frequency since variable E deviates more and

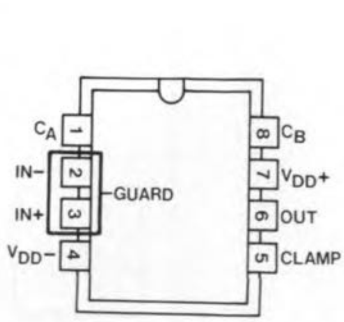


Fig. 14. Leakage currents may be kept to a minimum by providing a guard area around the opamp inputs.

Table 1. Electrical specifications at T = 25 °C

TYPE	U_{os} (μV) max.	dU_{os}/dT (nV/K) typ.	INPUT BIAS (pA) typ.	NOISE(1) (mV_{pp}) typ.	SUPPLY CURRENT (mA) typ.	SUPPLY VOLTAGE(3) (V) max.
ICL7650	5	100	1.5	2.0	2.0	18
TLC2652C	3	3	4	2.8	1.5	16
TLC2654C	20	4	50	1.5	1.5	16
LMC668	10	50	20	2.0	2.5	18
MAX420C	10	20	10	1.1	1.3	36
TSC900BC	15	100	80	4.0	0.2	18
LTC1049C	10	20	15	3.0	0.2	18
LM741C	6000	5000	80,000	—	1.7	36
OP177B	55	100	2400	0.33(2)	2.0	44

Notes: (1) 0 – 10 Hz (2) 1 – 100 Hz (3) V+ to V–

Table 1. Overview of the most popular chopper opamps, and their main technical characteristics. The 741 and the OP177B are not choppers — they are included here for reference.

more from the ideal value of 1 when the open-loop gain becomes smaller (see equation 2). Hence, this alternating voltage component appears also at the output of the auxiliary amplifier and at the input of S&H-2 (see Fig. 7). These components are generated as a result of the sampling operation, which causes sum and difference frequencies. To prevent these frequencies rising to an unacceptably high level in the output signal, the chip contains a special suppressor circuit. As shown in Fig. 8, the anti-intermodulation circuit injects a compensation signal into the null amp. This also results in additional suppression of harmonics of the chopper frequency.

Unfortunately, the suppressor circuit is not capable of resolving all problems. When the input frequency approaches the chopper frequency, a low-frequency beat signal is generated. This component is inevitably treated as off-set during the nulling of the main amp, and thus causes a complete disruption of the chopper amplifier. This annoying problem may be solved to a large extent by using a chopper frequency which is at least twice as high as the highest anticipated frequency in the amplified signal.

In many applications that rely on high d.c. accuracy (e.g., thermocouples) the bandwidth of the input signal is no more than a few hertz. It will be clear that such low frequencies prevent interference problems with the chopper frequency beforehand. In a number of cases, however, the signal bandwidth will have to be limited by a low-pass filter. When it is not possible, for whatever reason, to limit the bandwidth, the designer still has the possibility to apply another chopper amplifier rated for a higher clock frequency. The ICL7650, for instance, 'chops' at 200 Hz, the LMC688 (National Semiconductor) at 400 Hz, the TLC2652 at 450 Hz, and the TLC2654 at 10 kHz. In some cases, it is possible to apply an externally generated clock signal to the chip.

Practical notes

Chopper-stabilized opamps usually have the same pinning as standard types. This allows them to be used as upgrades in existing circuits, replacing opamps with worse d.c. specifications. The only components to be added are the two external capacitors, C_A and C_B. This is not required, however, with some amplifiers. The LTC1049 and LTC1050 from Linear Technology, for instance, have on-chip capacitors. Unfortunately, production techniques limit the maximum capacitance of such integrated capacitors to about 450 pF, which gives these opamps a low performance in regard to noise. The usual values of the external capacitors lie between 0.1 μF and 1.0 μF. In al cases, high-grade capacitors are required to bring out the specific qualities of a chopper opamp. Film capacitors like polystyrene and polypropylene types are well worth using.

Unfortunately, the use of high-grade capacitors is no guarantee that a d.c. amplifier is obtained with a small off-set and a low drift. There is another factor, which has not been mentioned so far: thermovoltages. Thermovoltages occur where two different metals are in contact. As indicated by the name of the phenomenon, the voltage is temperature-dependent. In practice, a thermovoltage readily amounts to a few microvolt per kelvin. The average drift of a good chopper-stabilized opamp is of the order of 10 nV/K. However, this value is not usually achievable in a practical amplifier without paying attention to thermoelectric effects in and around the circuit. Components which form connections without soldering, such as switches, relays and connectors, must not be used in the input circuit. Where parts are soldered, it is best to use solder tin with a low thermoelectric specification, such as a tin-cadmium alloy. Errors brought about by thermoelectric effects may also be kept to a minimum by arranging a symmetrical circuit at the opamp inputs. The most sensitive part of the amplifier is thermally balanced by using the same components in the two branches (even if they are really superfluous for the function of the circuit, see Fig. 13), and by forcing an equal number of solder joints. Furthermore, temperature differences as a result of, say, ventilation or power dissipation, must be kept as small as possible.

Guard !

An additional advantage of chopper-stabilized opamps is the extremely low input currents. The TLC2652, for instance, has an average input bias current of 4 pA at an ambient temperature of 25 °C. In practice, however, little use is made of this characteristic because the external leakage currents are much higher. Nonetheless, these leakage currents are fairly easily kept in check. The necessary measures may already be taken during the printed-circuit board design phase. For instance, the solder spots near the inverting and the non-inverting inputs of the opamp can be surrounded by a screening copper area, called a guard. The principle is illustrated in Fig. 14. It is desirable that the guard be held at about the same potential as the inputs of the opamp. Thus, the guard is connected to ground in an inverting circuit, and connected to the –input of the opamp in a non-inverting circuit. It will be clear that guards must be provided at both sides of the PCB. Finally, the PCB is cleaned with alcohol before fitting the components.

The differences

From the above discussion you will have gathered that there are many types of chopper opamps available. A selection of the most popular types, along with their main specifications, may be found in Table 1. The good old 741 opamp, which is *not* a chopper, is also included for your amusement. The OP177B at the end of the list represents the latest in bipolar technology, and is a competitive alternative to chopper opamps, according to the manufacturer, PMI.

Finally, a word of warning to those of you who want to start immediately replacing standard opamps by chopper types: as yet, these devices are quite expensive (expect to pay around £10 per amplifier) and difficult to obtain as one-offs. ■

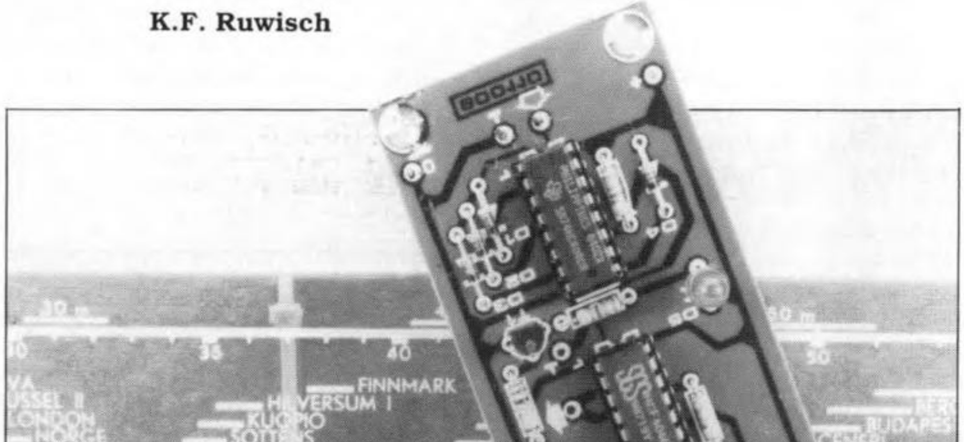
Reference:

1. "Introducing OP-series opamps". *Elektor Electronics* February 1990.

DROITWICH TIMEBASE

Roughly two years ago, the carrier frequency of the 400-kW long-wave Droitwich transmitter was changed from 200 kHz to 198 KHz. This was done by the BBC to comply with the internationally agreed 9-kHz spacing for broadcast stations in the medium- and long-wave bands. The frequency change of 2 kHz was largely unnoticed by thousands of listeners of the Radio Four and BBC World Service programmes. Not so, however, by the many users of frequency standards and timebase circuits which derived their stability from the 200 kHz carrier. All these circuits became useless overnight since the new frequency, 198 kHz, can not be divided to give multiples of 10 Hz. Fortunately, there is a way to get your timebase ticking again. An update for our own Droitwich receiver, an immensely popular project which goes back as far as 1977, is described here.

K.F. Ruwisch



THE BBC Radio-4 and BBC World Service programmes from the 198-kHz transmitter at Droitwich (near Birmingham) can be received throughout Western Europe. The programmes are not our main concern here, however, since these can be listened to with almost any MW/LW radio. As with many stations in the long-wave band, the stability of the carrier transmitted by Droitwich is derived from an atomic reference, and can be used for building a precision timebase at a small outlay. How this is done with simple means is explained in Ref. 1. Basically, the carrier is picked up with an aerial, amplified and subsequently digitized. Next, the output signal is fed to a divider cascade which supplies the commonly used timebase frequencies of 1 Hz, 10 Hz, 100 Hz, etc., to 100 kHz. The stability of each of these timebase frequencies is, in principle, the same as that of the carrier from Droitwich, which, up to a two years ago, was accurately maintained at 200 kHz. Over the years, the 200-kHz carrier from Droitwich has served thousands of hobbyists and professional workers in electronics laboratories all over Europe by providing a reference frequency of a stability that is not achievable with any affordable circuit. Traditionally, Droitwich receivers, including our own, are of a charming simplicity, and the signal is strong and freely available.

Just retuning?

Although the difference of 2 kHz is hardly noticed on the tuning scale of the vintage radio in the introductory photograph, the output frequencies supplied by an unmodified Droitwich timebase are useless for most, if not all, digital circuits. This is because they are no longer exact multiples of 10 Hz.

The problem is obvious: we can no longer use our receiver plus timebase because the Droitwich transmitter is at 198 kHz instead of 200 kHz. All is not lost, however. The good news is that the stability of the Droitwich carrier is still just as good as before the change from 200 kHz to 198 kHz. So, the solution to the problem is also obvious: to enable us to use our timebase circuits, we must convert the 198-kHz output signal of our Droitwich receiver to 200 kHz.

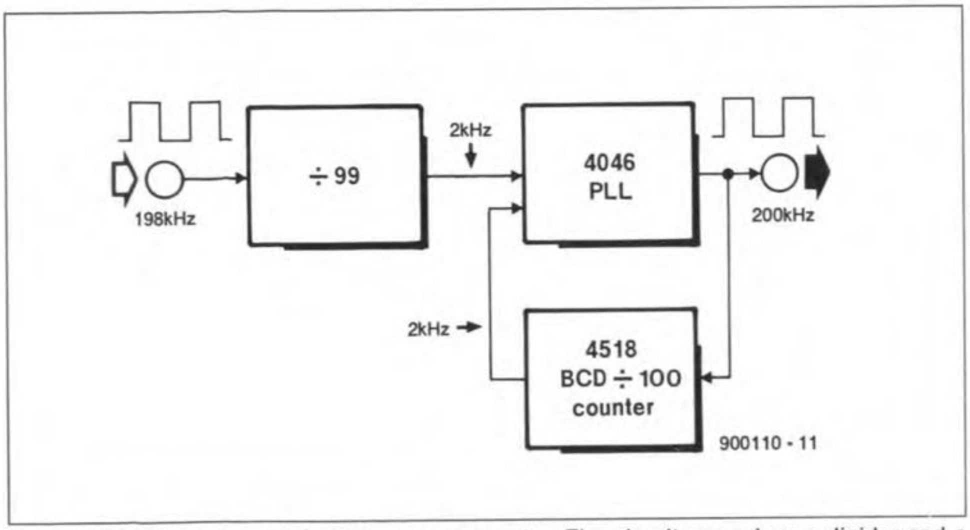


Fig. 1. Block diagram of the frequency converter. The circuit comprises a divider and a phase-locked loop.

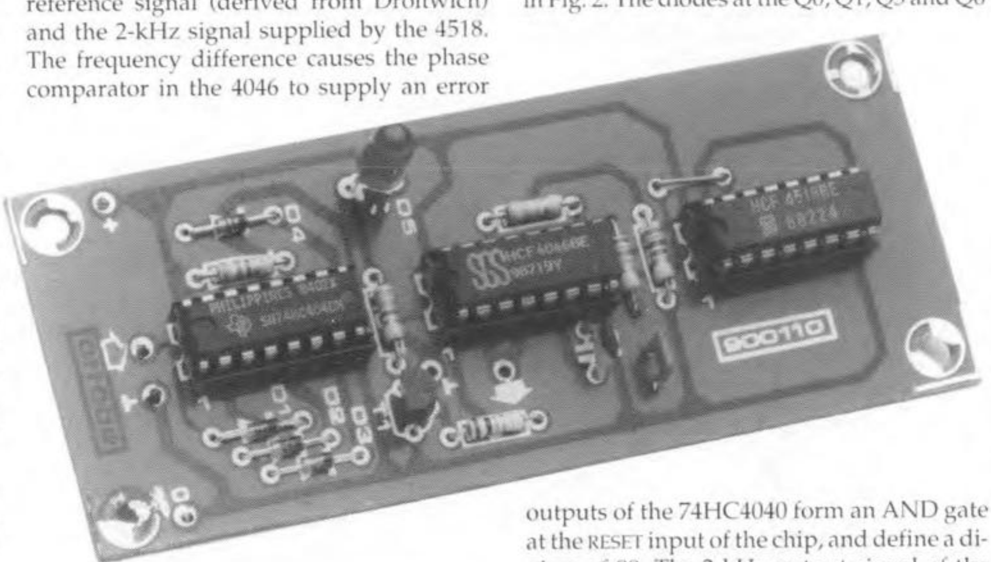
Up by 2 kHz

The block diagram of the circuit we have in mind is shown in Fig. 1. Assuming that the unit is provided with the 198-kHz digital output pulses from a Droitwich receiver, it supplies a rock-steady 200-kHz output signal. No changes are required to the existing Droitwich receiver.

At the input of the upgrade circuit we find a special divider around a 4040, wired for a divisor of 99. Its output signal has a frequency of 2 kHz and is used as a reference for a phase-locked loop (PLL) circuit based on the well-known 4046. The voltage-controlled oscillator in the PLL is set to operate at 200 kHz. Its output signal is divided by 100 by a 4518 dual decade counter to give 2 kHz.

To understand how the output frequency of the circuit is kept stable, let us assume that the VCO drifts from the nominal frequency of 200 kHz. This drift, however small, causes a frequency difference between the 2-kHz reference signal (derived from Droitwich) and the 2-kHz signal supplied by the 4518. The frequency difference causes the phase comparator in the 4046 to supply an error voltage. In this way, the VCO is automatically retuned to minimize the frequency difference. The upshot is that the VCO output frequency of 200 kHz is 'locked' to the carrier received from Droitwich.

It could be argued that the PLL is not required because the 2-kHz signal from the 4040-based divider may be fed, at a suitable point, into an existing divider cascade to give the previously mentioned decade timebase frequencies. We feel that the up-conversion to 200-kHz is required, however, to make sure that the formerly available frequencies of 100 kHz and 10 kHz are retained without any change to the existing divider cascade. In other words, all the functions of the Droitwich timebase you were just about to throw away are restored simply by installing the proposed up-converter.



Circuit description

The circuit diagram of the upgrade is shown in Fig. 2. The diodes at the Q0, Q1, Q5 and Q6 outputs of the 74HC4040 form an AND gate at the RESET input of the chip, and define a divisor of 99. The 2-kHz output signal of the HC4040 is fed to the 4046 PLL, whose internal organization is shown in Fig. 3. The VCO frequency is defined by external parts R_2 and C_1. Here, phase comparator 2 is used. Network R_6-C_2 forms the PLL loop filter at the control input of the VCO. A LED indicates that the PLL is locked to the Droitwich signal.

The 200-kHz VCO signal is divided by 100 in a 4518 dual BCD counter. The 2-kHz output signal at pin 14 of this IC is fed to the Cin (phase comparator in) input of the 4046.

The 200-kHz output signal of the upgrade circuit is digitally compatible with a swing of 5 V_{pp}, and can be fed to any existing divider cascade based on TTL ICs or CMOS ICs operating at a supply voltage of 5 V.

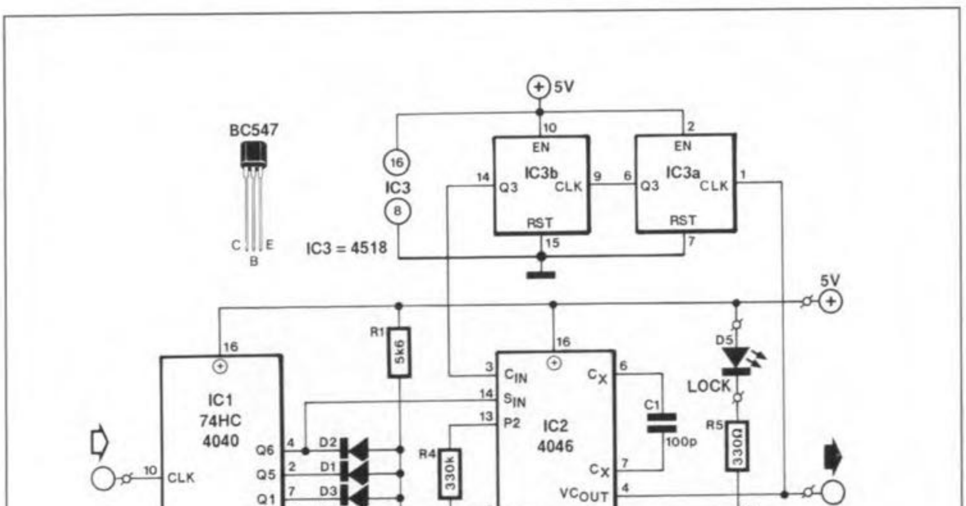


Fig. 2. Circuit diagram of the timebase upgrade.

Construction

Construction of the upgrade circuit is straightforward on the small PCB shown in Fig. 4. The input of the board is connected to output A of the Droitwich receiver (see Ref. 1). The output of the board is connected to the existing 200-kHz output socket of your frequency standard, and to the input of any divider cascade you may have built into the

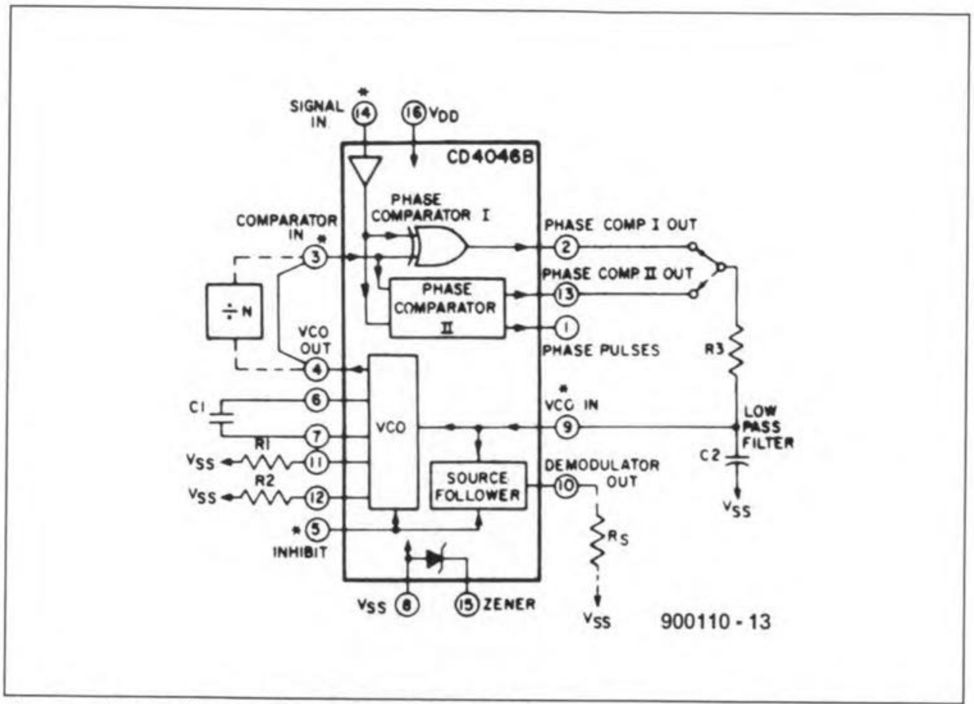


Fig. 3. Block diagram of the 4046 phase-locked loop used in the upgrade (illustration courtesy RCA/Harris Semiconductor).

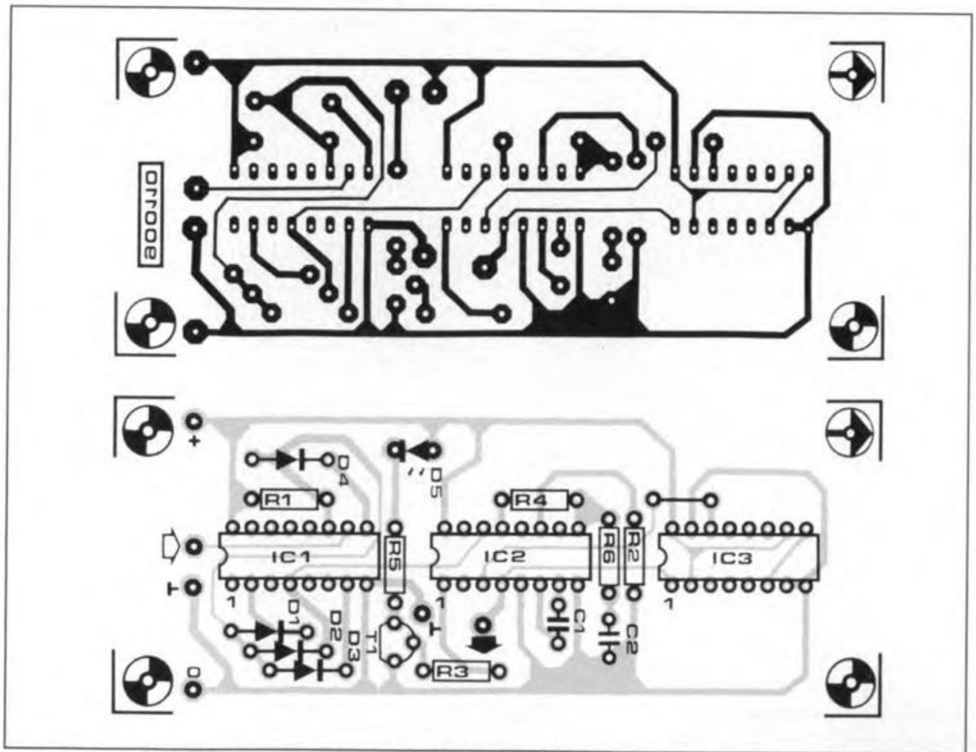


Fig. 4. Single-sided printed-circuit board for the timebase upgrade.

enclosure (a suggested divider based on TTL ICs is included in Ref. 1).

Assuming that you use the Droitwich receiver described in Ref. 1, carefully adjust the aerial, and then preset P1, until the LOCK LED on the receiver board lights. Use the earpiece, and check that you are tuned to Droitwich by listening to the programme. Next, check that output A of the receiver supplies digital pulses to the upgrade board. If the pulse train is steady, the LOCK LED on the upgrade board lights, and the output should supply a stable 200-kHz signal. ■

Reference:

1. "Precision timebase for frequency counter". *Elektor Electronics* June 1977.

COMPONENTS LIST

Resistors:		
1	5kΩ6	R1
1	2kΩ2	R2
1	10kΩ	R3
1	330kΩ	R4
1	330Ω	R5
1	15kΩ	R6
Capacitors:		
1	100pF	C1
1	100nF	C2
Semiconductors:		
4	1N4148	D1 – D4
1	LED	D5
1	BC547B	T1
1	74HC4040	IC1
1	4046	IC2
1	4518	IC3

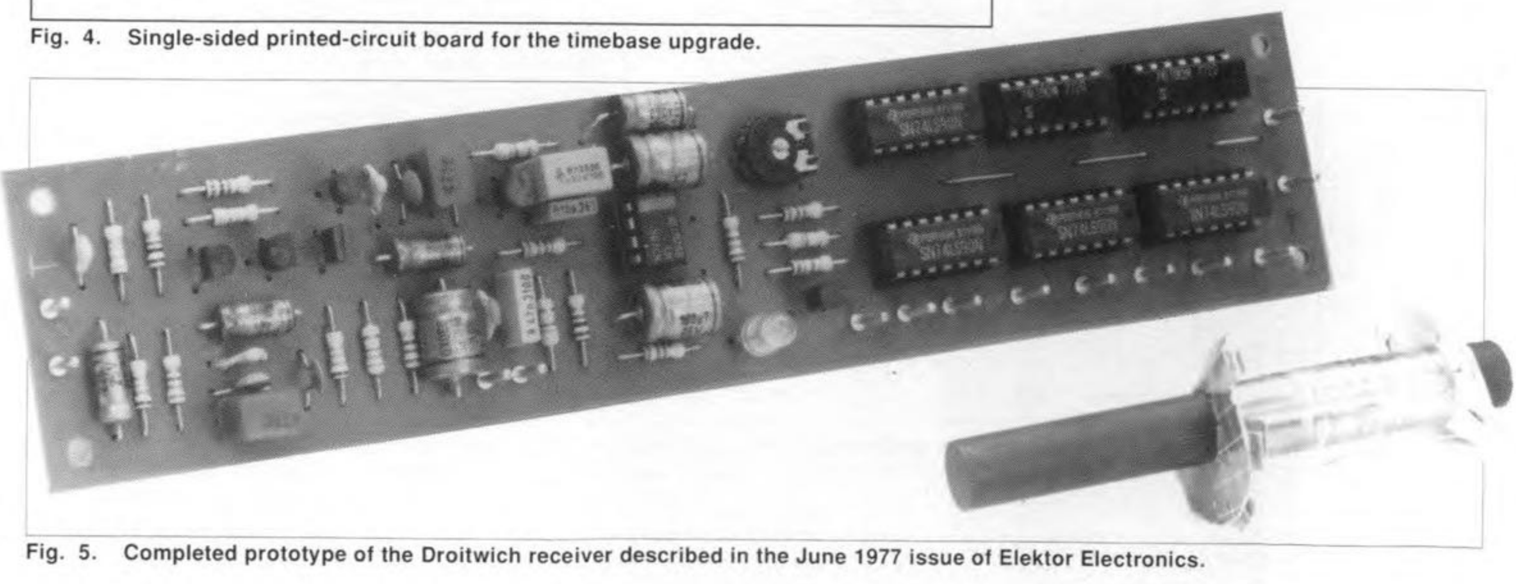
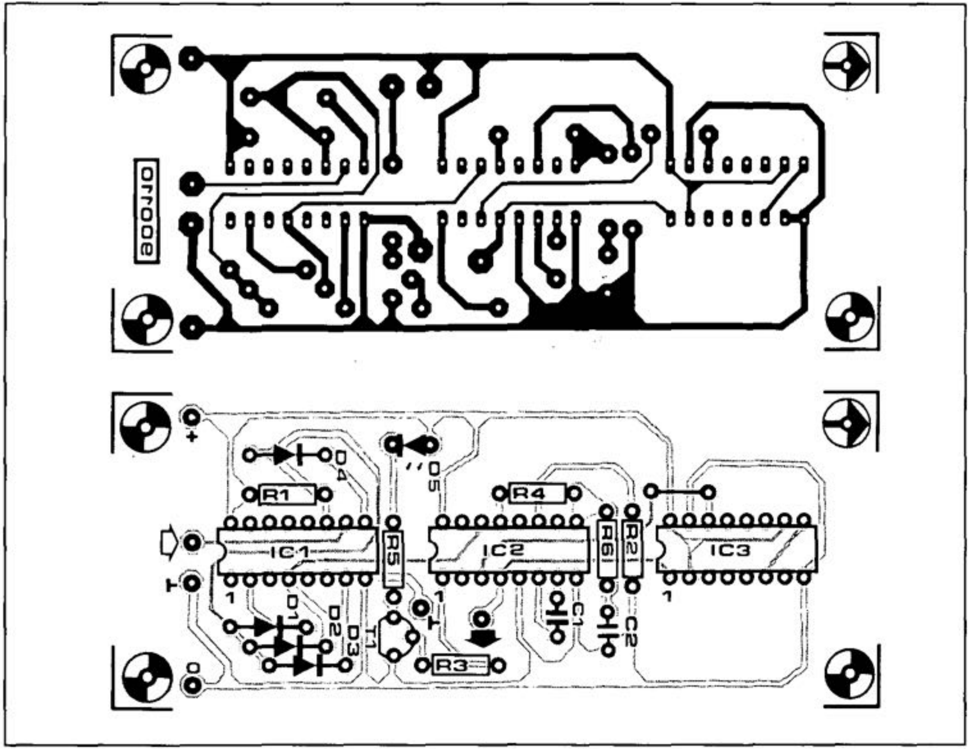


Fig. 5. Completed prototype of the Droitwich receiver described in the June 1977 issue of Elektor Electronics.



COMPONENTS LIST

Resistors:

1	5kΩ6	R1
1	2kΩ2	R2
1	10kΩ	R3
1	330kΩ	R4
1	330Ω	R5
1	15kΩ	R6

Capacitors:

1	100pF	C1
1	100nF	C2

Semiconductors:

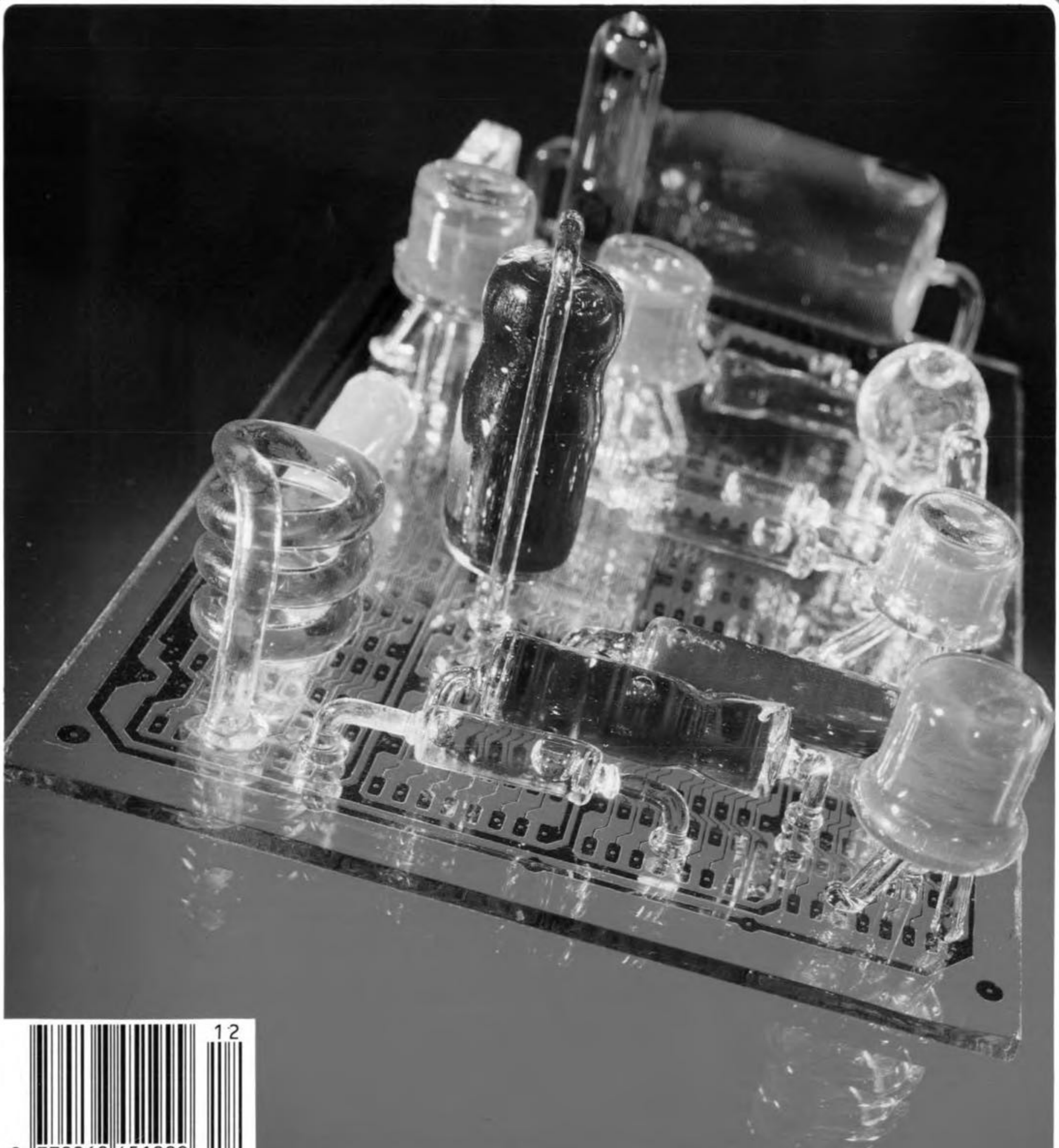
4	1N4148	D1 – D4
1	LED	D5
1	BC547B	T1
1	74HC4040	IC1
1	4046	IC2
1	4518	IC3

THE INTERNATIONAL MAGAZINE FOR ELECTRONICS ENTHUSIASTS
December 1990

Elektor Electronics

SUPPLEMENT

36 pages of construction projects on Audio & Hi-fi, Computers & Microprocessors, General Interest, Power Supplies, Radio & Television, and Test & Measurement



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Elektor Electronics

December 1990
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Supplement

CONTENTS

Front cover
Crystal glass blown specially for our front cover photograph is symbolic of the way we try to make electronics colourful and clear as crystal!

001 LOW CONSUMPTION MOVEMENT DETECTOR

This highly sensitive movement detector is designed from bipolar transistors and draws a current of only 0.3 mA during quiescent operation. It is intended primarily for use as a protection device, but may also be used in certain games.

Mechanical movement detectors react only to large changes in velocity or vibrations that set a metal leaf provided with a suitable counter weight into motion. The present detector is much more sensitive: moving an object that is protected by it is a real challenge as even the most careful attempt at doing so is punished by the sounding of a buzzer.

Yet, the principle is simple: a magnet is suspended by a thin thread 20–30 mm long a few millimeteres above the coil of a relay (whose contacts are not used). Even a minute movement of the protected object will disturb the magnet. The resulting changes in the magnetic field above the relay coil will induce a tiny varying voltage across the coil.

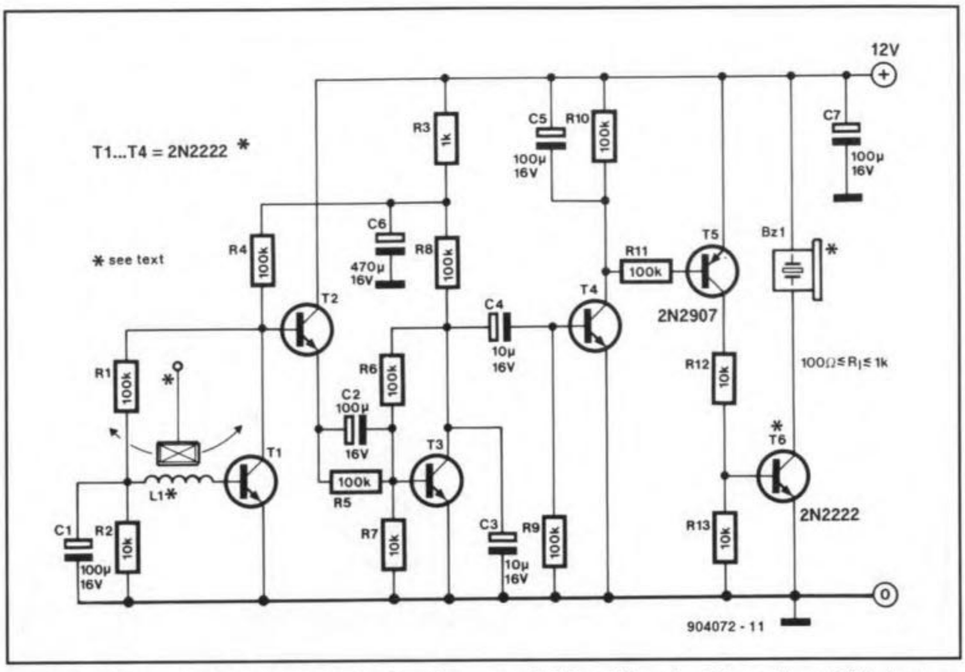
Although an opamp could be used for the amplification of this tiny voltage, types that combine low consumption with single supply voltages are rare and expensive. The present design therefore uses discrete bipolar transistors that are easily available, draw little current and are not expensive.

The first stage consists of a common emitter design with automatic regulation. The collector resistors and the resistors in the regulation bridge have unusually high values.

Feedback from the bridge ensures stability of operation of T1. Each increase in collector voltage will be opposed by an increase in base-emitter current. Conversely, each reduction in collector voltage will be opposed by a decrease in base-emitter current. Consequently, the collector voltage will stabilize at a value that corresponds to a base voltage of about 0.6 V. Capacitor C1 delays the immediate effect of the feedback when the collector voltage changes rapidly.

The small varying voltage induced in the relay coil is magnified appreciably by T1 because C 1 prevents automatic regulation. The output impedance of the first stage is very high, which is, of course, the price to be paid for low consumption. It would not make sense to follow this stage by one with a low output impedance, because this would adversely affect the overall amplification,

Because of that, T1 is followed by an



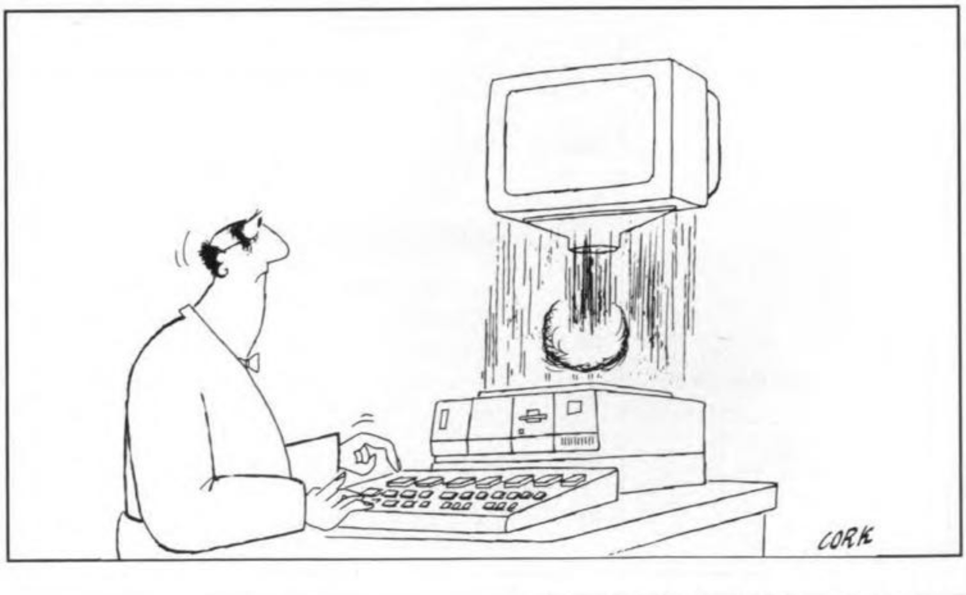
emitter follower, T2, which provides the coupling between T1 and T3. Resistor R5 allows a partial discharge of C2 if T2 is switched off by a reduction in the output of T1. Since this resistor, because of the low- consumption requirement, has a high value, the circuit will attain its maximum sensitivity some ten seconds after the last movement detection. This is the time required for the charge on capacitor C2 to stabilize.

The detection proper is carried out by T4, which switches on when the voltage variations in the amplifier, passed on by C4, reach a level of 0.6 V. Saturation of T4 leads to the instant charging of C5. This capacitor will discharge partly via R10 and R11 to the base of T5 when T4 switches off again. When C5 discharges, T5 is thus on and this will make T6 conduct. This in turn will actuate a load, for instance, a buzzer, in the collector circuit of T6.

The sensitivity of the detector depends to a large extent on the distance between the magnet and the relay and the length of the 'pendulum'.

If the circuit is powered by a battery, there is a little problem: batteries have a large internal resistance. This means that the supply voltage may vary by some tenths of a volt if a sudden, large current is drawn. If the buzzer has stopped after a detection, such a situation can lead to a retriggering of the circuit and this may cause undesired oscillations. To prevent this happening, the supply of the amplifier stage is decoupled by R3 and C6. ■

(O. Bailleux)



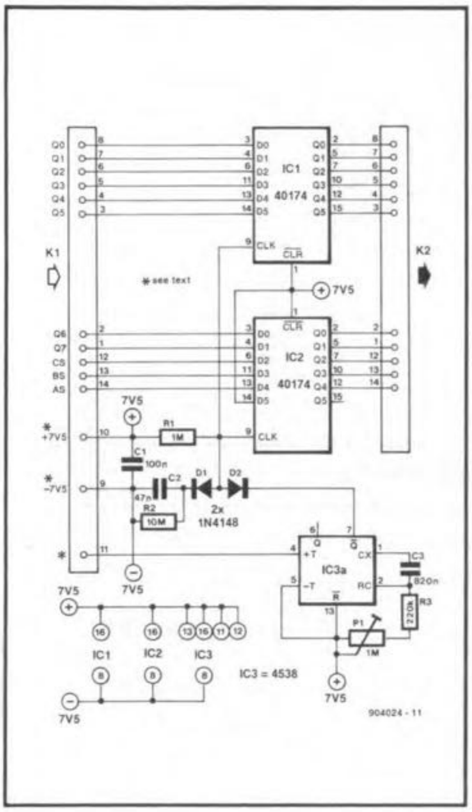
SIGNAL SUPPRESSOR FOR ALL-SOLID-STATE PREAMPLIFIER

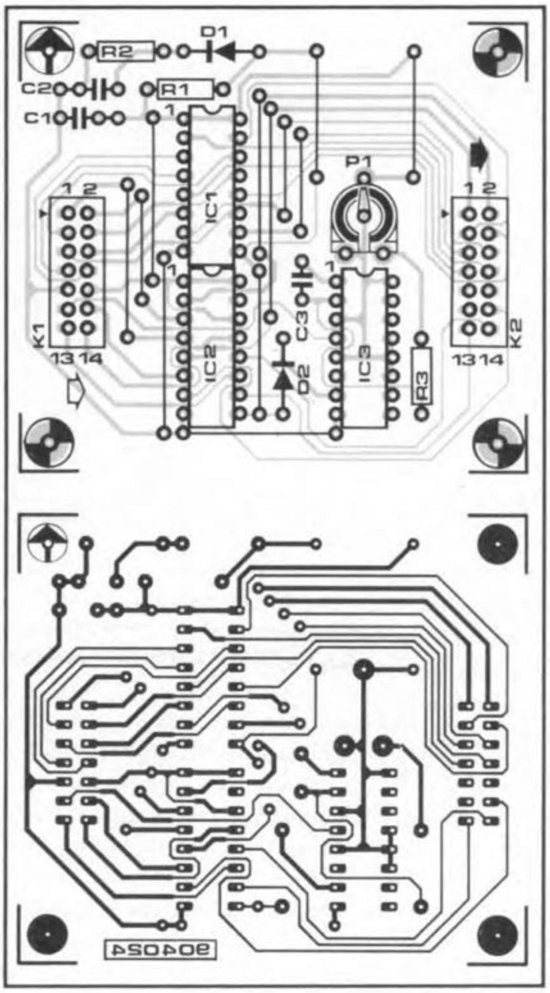
002

The all-solid-state preamplifier we published some months ago (Ref. 1) is controlled entirely electronically, including the switching of the inputs. When several inpinput sources are active, and switching takes place between two of them, the signals of the sources between them may be heard, admittedly for a very short time. Nevertheless, this may be inadmissable in certain circumstances.

The remedy for this is fairly simple: connect a buffer between the control lines of the basic PCB and the switching inputs to the multiplexer. This buffer retains the data of the currently selected channel until a definite code is present for the newly selected channel.

The circuit presented here acts as that buffer and is inserted between connectors K14 and K17. Three pins of those connectors are not used in the original circuit and these are therefore available to provide a symmetrical supply voltage and the necessary clock signal to the present circuit. The ±7.5 V supply is taken from pins 8 and 16 of IC37 and fed to pins 9 and 10 of K17 while the clock signal is taken from pin 11 of IC35 and applied to pin 11 of this connector. The connections are simply made with short lengths of insulated circuit wire. ■





Reference: "All-solid-state preamplifier", *Elektor Electronics*, December 1989.

PARTS LIST

Resistors
R1 = 1 M
R2 = 10 M
R3 = 220 k
P1 = 1 M preset

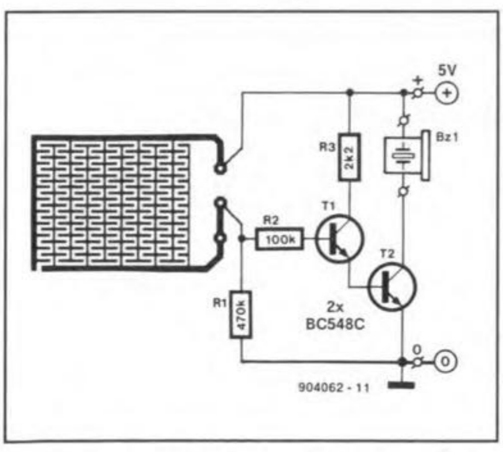
Capacitors
C1 = 100 n
C2= 47 n
C3 = 820 n

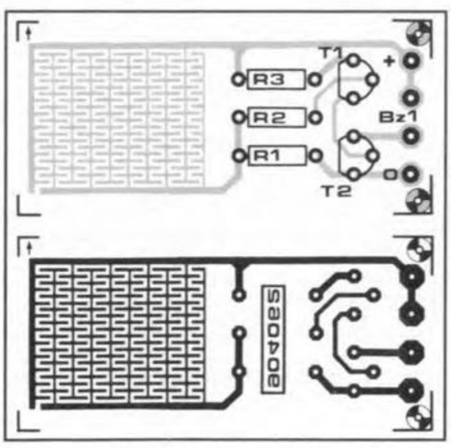
Semiconductors
D1, D2 = 1N4148
IC1, IC2 = 40174
IC3 = 4538

Miscellaneous
K1, K2 = header FC14VB
14-way flat cable, 10 cm
2×female header, 14-way for cable mounting

"BATH FULL" INDICATOR

003





Running a bath can end in a minor domestic disaster if you forget to turn off the taps in time. The indicator presented here actuates an active buzzer to provide an audible warning when a given water level is reached.

Since the water sensor and the driver circuit for the buzzer are contained on one PCB, the indicator, together with the 9-V battery and the buzzer, may be built into a compact case. Obviously, the sensor, which is etched on the PCB, must not be fitted in cast-iron or steel bath, the indicator is secured to it with the aid of a magnet glued on to the case. To prevent scratching the bath, the magnet may be covered in plastic or rubber. If you have a polypropylene bath, the indicator may be stuck on to it with blue tack or double-sided adhesive tape.

When the water reaches the sensor, the base of T1 is connected to the positive supply line. As a result, T1 and T2 are switched on so that buzzer Bz1, a self-oscillating type, is actuated. The current drawn by the circuit in that condition is about 25 mA.

In case the circuit is actuated by steam, its sensitivity may be reduced by increasing the value of R2. It is recommended to tin the PCB tracks to prevent corrosion. ■

(D. Lorenz)

PARTS LIST

Resistors;
R1 = 470 k
R2 = 100 k
R3 = 2k2

Semiconductors:
T1, T2 = BC548C

Miscellaneous
Bz1 = active piezo-ceramic resonator

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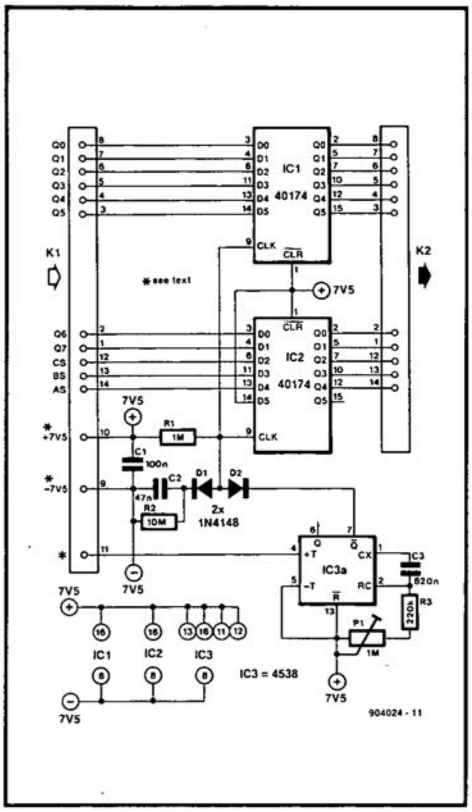

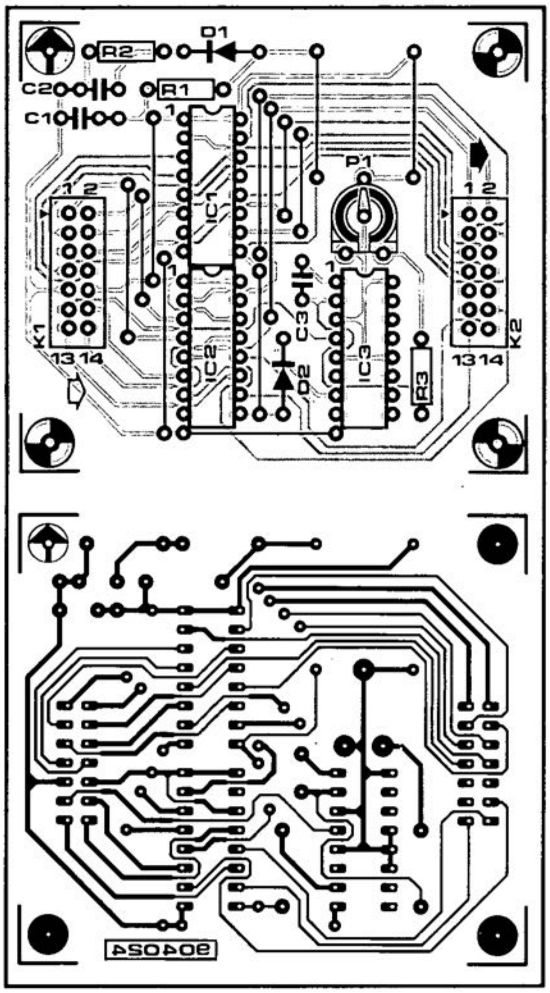

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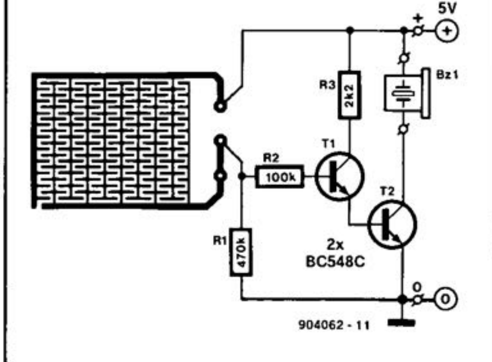

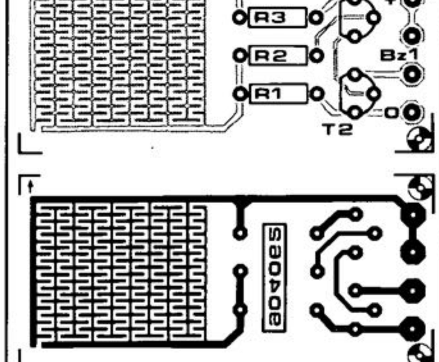

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004 SAWTOOTH CONVERTER

Simple function generators normally provide sinusoidal, rectangular and triangular waveforms, but seldom a sawtooth. The circuit in Fig. 1 derives a a sawtooth signal from a rectangular and a triangular signal. Its quality depends on the linearity of the triangular signal, the slope of the edges of the rectangular signal and the phase relation between the rectangular and triangular signals.

The conversion is carried out in IC1. Whether the triangular signal at input A is converted or not by IC1 depends on the state of T1. This FET is controlled by the rectangular signal at input B.

The signal at the output of the opamp is a sawtooth—see Fig. 2—whose trailing edge is inverted. The frequency of this signal is double that of the input signals.

If in this state the d.c. level of each inverted edge is raised sufficiently to make the lower level of that edge coincide with the higher level of the preceding edge, a sawtooth signal of the same frequency but double the peak value of the input signals is obtained. The d.c. level is raised by adding input B to the output of IC1 via R7 and P1. The preset should preferably be a multi-turn type.

Resistors R2 and R4 are 1% types.

If a rectangular signal is not available, or its peak value is too small, the auxiliary circuits shown in Figures 3 and 4 will be be found useful. That in Fig. 3 amplifies the triangular input at A by 10. Differentiating network C1-R10 derives rectangular pulses from the amplified triangular signal and these are available at F.

The pulses at F are shaped by the circuit in Fig. 4 to rectangular signals that have the same peak value as the supply voltage. Capacitor C2 increases the slope of the edge and may be omitted for low-frequency signals.

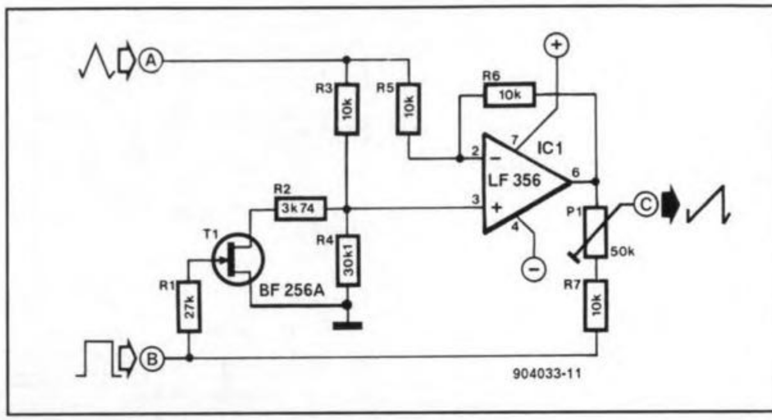


Fig. 1. Circuit diagram of the basic sawtooth convertor.

The convertor provides sawtooth signals over the frequency range of 15 Hz to 15 kHz. If the auxiliary circuits are used, capacitor C1 must be compatible with the frequency of the sawtooth signal (its value lies between 2 nF and 100 pF).

The supply for all circuits may be between ±10 V and ±15 V. Each opamp draws a current of 4–6 mA. ■

(*A. Ferndown*)

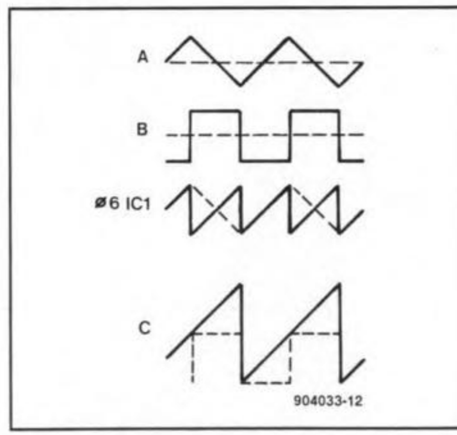


Fig. 2. Signals at various points in Fig. 1.

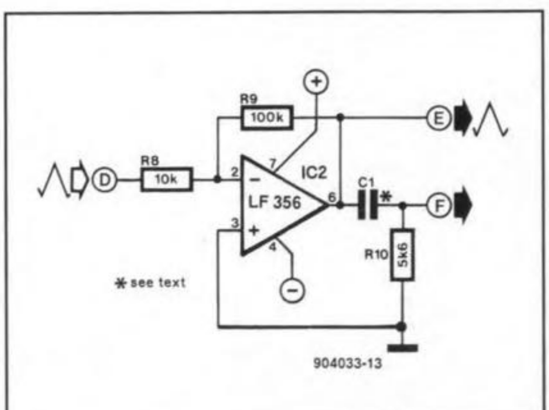


Fig. 3. Circuit for amplifying the input at A in Fig. 1.

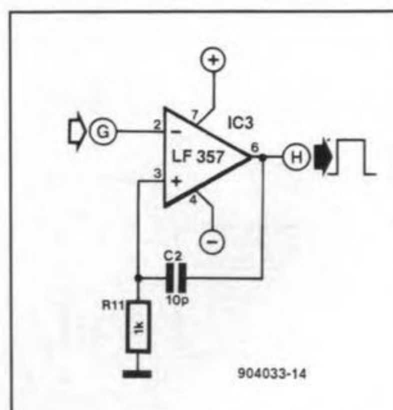


Fig. 4. Circuit for shaping the rectangular pulses at output F in Fig. 3.

ELECTRONIC IGNITION 005

The electronic ignition circuit presented here is intended to be inserted into a car's conventional ignition system. In effect, it replaces the original 12 V switching circuit in the primary winding of the coil by one generating more than 100 V. It thereby converts a current circuit, which is upset by lead and stray resistance, into a voltage circuit that is much more efficient.

The pulses emanting from the contact breaker, shown at the extreme lower left-hand side of the diagram, are applied to transistor T1 and subsequently differentiated by R3-C1. This causes a negligible ignition delay. The current through the contact-breaker points is determined by the value of R1. This value has been chosen to ensure that the points remain clean.

Transistor T1 is followed by two monostables, IC1a and IC1b, which are both triggered by the output pulses of T1. However, whereas IC1a is triggered by the trailing edge, IC1b is by the leading edge.

Monostable IC1a passes a pulse of about 1.5 ms—determined by R4-C2—to NAND gate IC2a. This gate switches off high-voltage darlington T3 via gates IC2b, IC2c and IC2d, and driver T2, for the duration of the pulse. Gate IC2 ensures that T3 is switched on only when the engine is running to prevent a current of some amperes flowing through the ignition coil.

As long as pulses emanate from the contact breaker, IC1b is triggered and its Q output remains logic high. The mono time of this stage is about 1s and is determined by R5-C3.

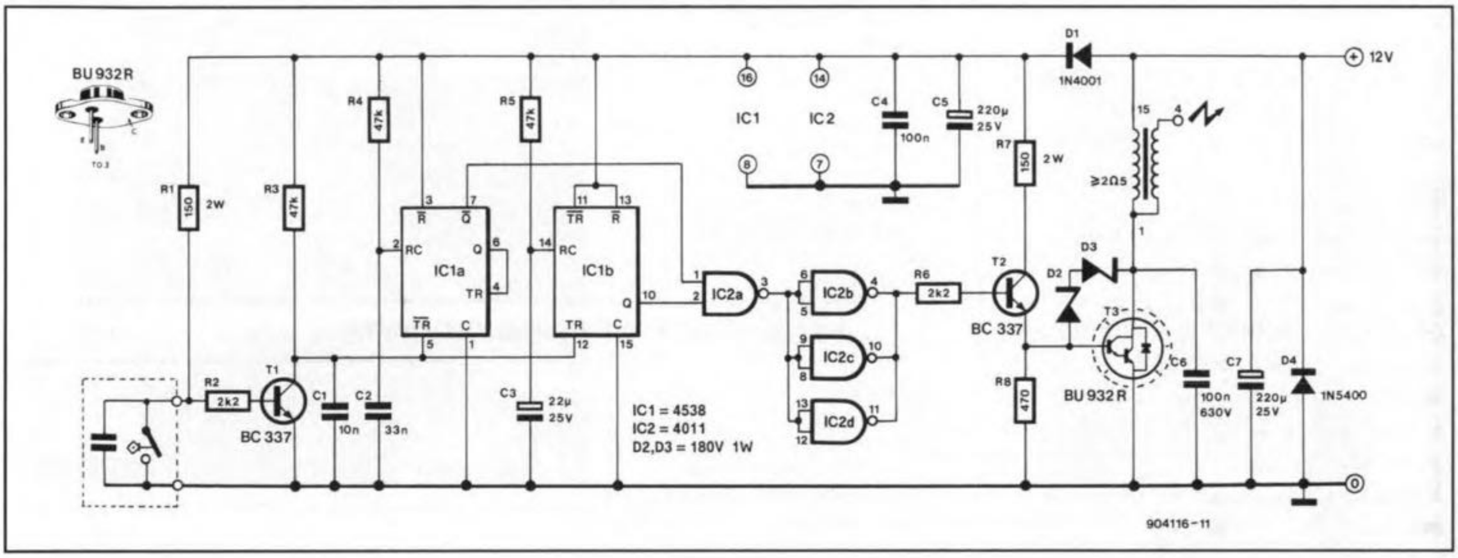
Darlington T3 is switched on via T2 and IC2a–IC2d as long as IC1a does not pass an ignition pulse.

When the engine is not running, the Q output of IC2b goes low after 1 s and this causes T2 and T3 to be switched off.

The two series-connected 180-V zener diodes protect the collector of the BU932R against too high a voltage.

The darlington must be fitted on a suitable heat sink. ■

(H. Döpfer)



WIDEBAND UHF AMPLIFIER 006

The construction of a UHF amplifier frightens most people, unless they are experienced radio/TV enthusiasts. They should, therefore, appreciate the circuit presented here, which is as straightforward as can be. It offers 10–15 dB gain over the frequency range 400–850 MHz and is therefore eminently suitable for situations where the television signal is on the weak side. Moreover, the filters may be adapted to the individual needs of users.

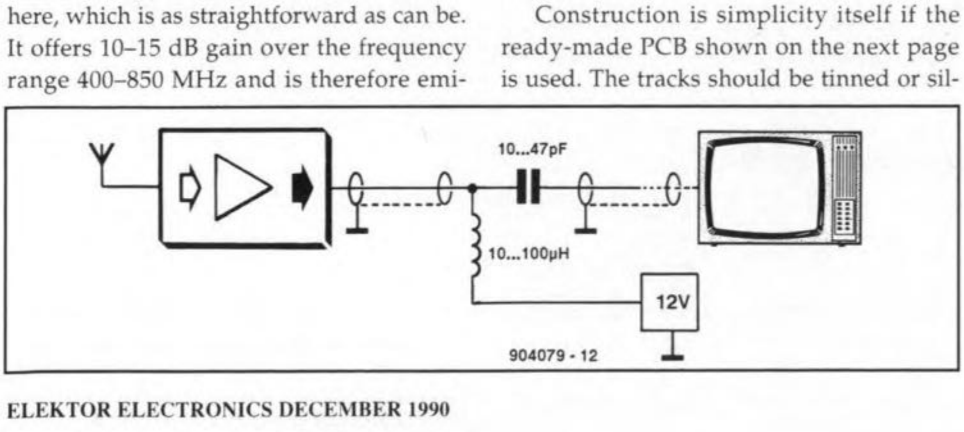
Construction is simplicity itself if the ready-made PCB shown on the next page is used. The tracks should be tinned or silvered for optimum performance and long life.

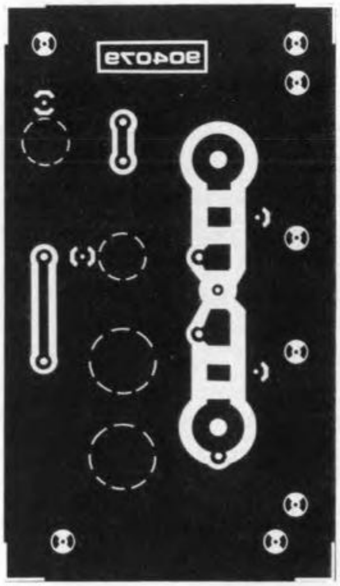
The opening at the centre of the board is intended to accommodate the transistor. This device has two emitter pins, both of which should be connected to ground.

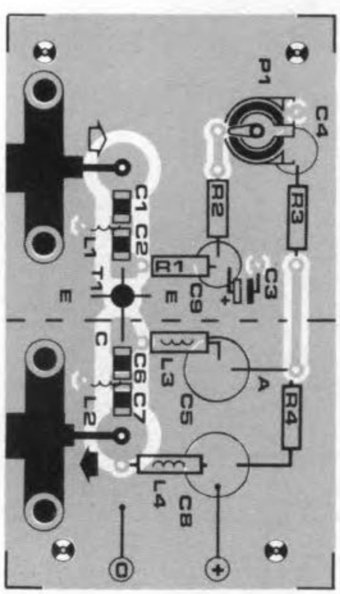
The drawings show that the board is divided into two by a small piece of tin plate, which should have a small cut-out for the transistor.

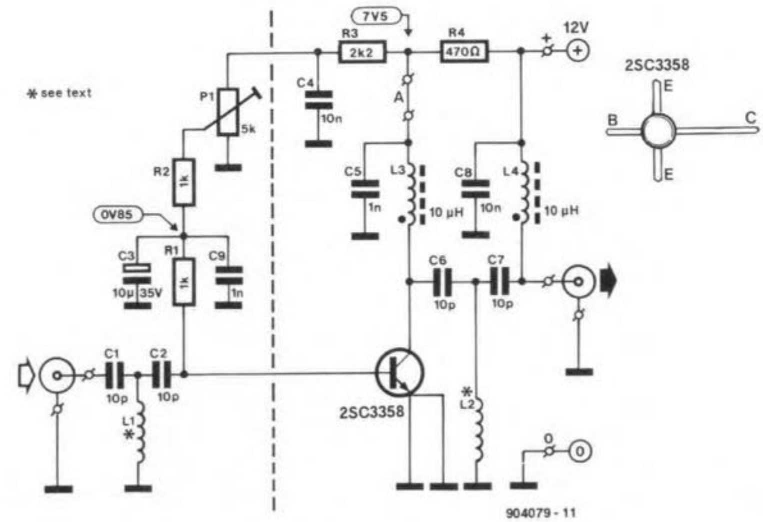
The input and output terminals are made from small cable clamps and M3 nuts and bolts.

One side of disc capacitors C4, C5, C8 and C9 is soldered direct on to the board









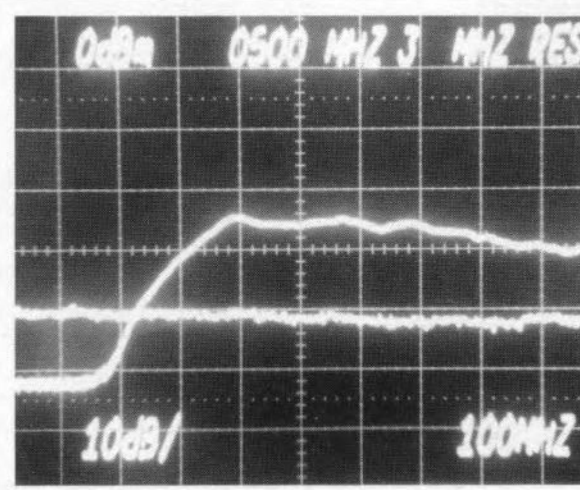
PARTS LIST

Resistors:
R1, R2 = 1 k
R3 = 2k2
R4 = 470 Ω
P1 = 5 k preset pot meter

Capacitors:
C1, C2, C6, C7 = 10 p surface-mount
C3 = 10 µ; 35 V
C4, C8 = 10 n disc type
C5, C9 = 1 n disc type

Inductors:
L1, L2 = air cored, 2 turns of 3 mm dia enamlled copper wire
L3, L4 = 10 µH choke or 10 turns of 0.2 mm dia enamelled copper wire on a ferrite bead.

Semiconductors:
T1 = 2SC3358



for which a fairly large soldering iron should be used.

All remaining components should be fitted with their terminals cut as short is feasible.

Input and output capacitors, C1 and 2, and C6 and C7 respectively are surface-mount types. C1-C2-L1 form an input filter and C6-C7-L2 an output filter. The value of the capacitors may have to be lowered to 3.9 pF to obtain the correct frequency range.

The overall frequency characteristic is shown in the second photograph.

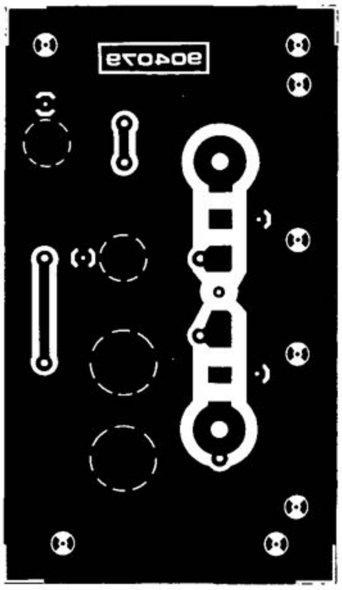
The amplifier may be housed in a watertight case and then mounted near the antenna at the top of the mast (if used).

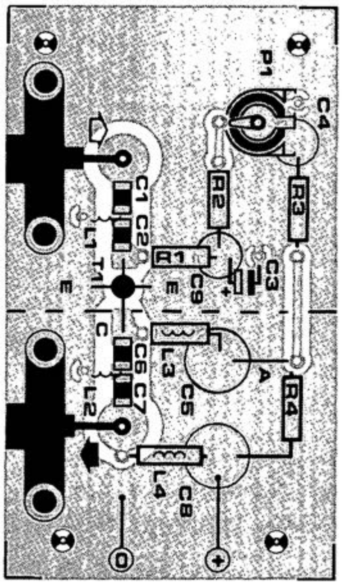
The power is obtained from a simple stabilized 12 V supply: a mains adaptor with a 78L12 will do nicely. This may be kept indoors, of course. The amplifier may be powered via the coaxial feeder cable, for which purpose a 10–100 µH choke is inserted in the supply line.

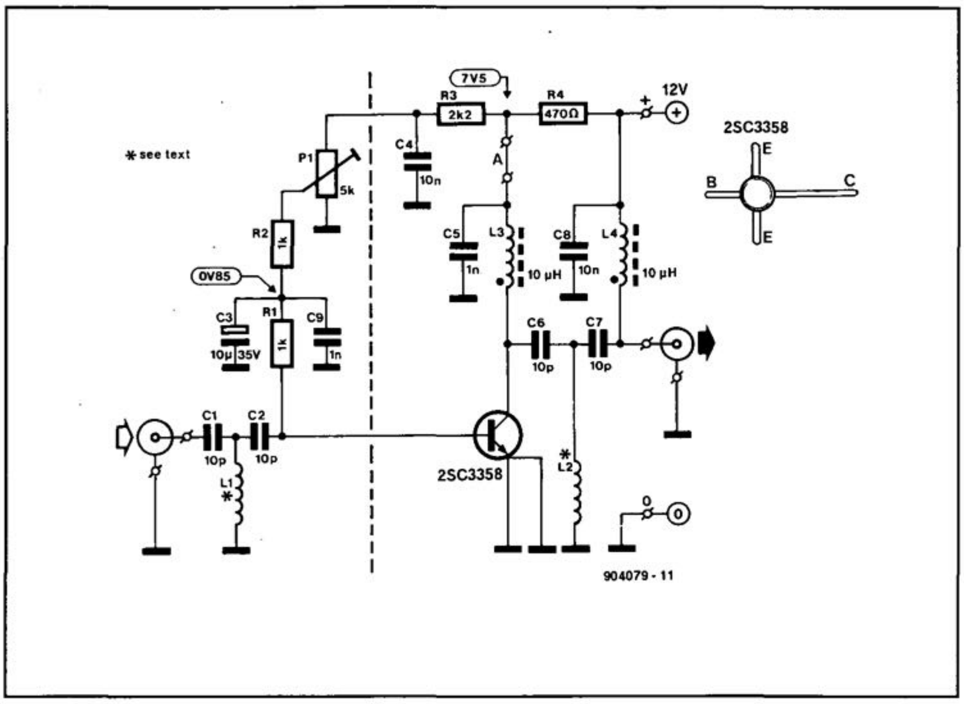
The television receiver is connected to the amplifier via a small coupling capacitor as shown on the previous page.

Calibrating the amplifier is straighforward: set P1 to the centre of its travel and then adjust it for optimum picture quality. In practice, the collector current of the transistor is then 5–15 mA. This may be checked by temporarily replacing jump lead A by a milliammeter. ■

(*K. Kraus*)







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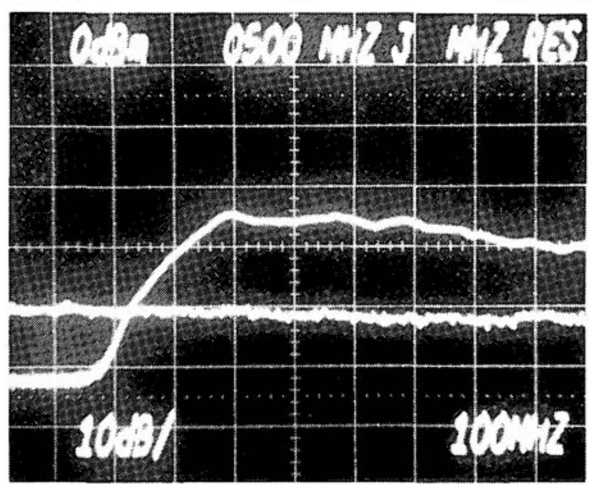

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(K. Kraus)

007 LIGHT GUARANTEED

The circuit presented here guarantees that if bulb La1 gives up the ghost, bulb La2 will take over its task, so that there is always light.

In series with La1 is triac Tri2. Resistor R3 and C2 form a delay network. As soon

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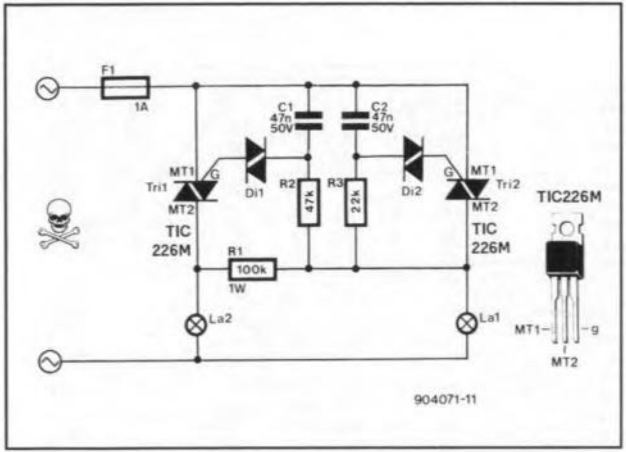


as the voltage across C2 rises above about 30 V, diac (= gateless triac) D2 is switched on, which causes Tri2 to conduct, so that La1 lights.

The control circuit of La2 is parallel to that of La1, but because R2-C1 has twice the delay of R3-C2, Tri1 will not be triggered when Tri2 conducts. When Tri2 conducts, C1 discharges, so that Tri1 can not be triggered.

When, however, La1 is open-circuited, there is a voltage across both *RC* networks via La2 and R1. Again, Tri2 will be triggered first, but since the current through it is smaller than than its holding current, it will cease to conduct almost immediately. Capacitor C1 will then continue to charge and after a little while Tri1 is switched on.

Because the time constant for La2 is somewhat longer than that for La1, La2 will always be slightly less bright than La1. It is, of course, possible to give La2 a slightly higher wattage than La1 to ensure equal brightness.

Without heat sinks, the triacs can handle up to 100 W each; with heat sinks powers of up to 1000 W may be accommodated. It is not recommended to use bulbs with a wattage below 25 W since these may flicker.

The triacs may be any type that can handle at least 400 V at not less than 5 A. The M types used in the prototype can handle 600 V at 5 A. ■

(*O. Bailleux*)

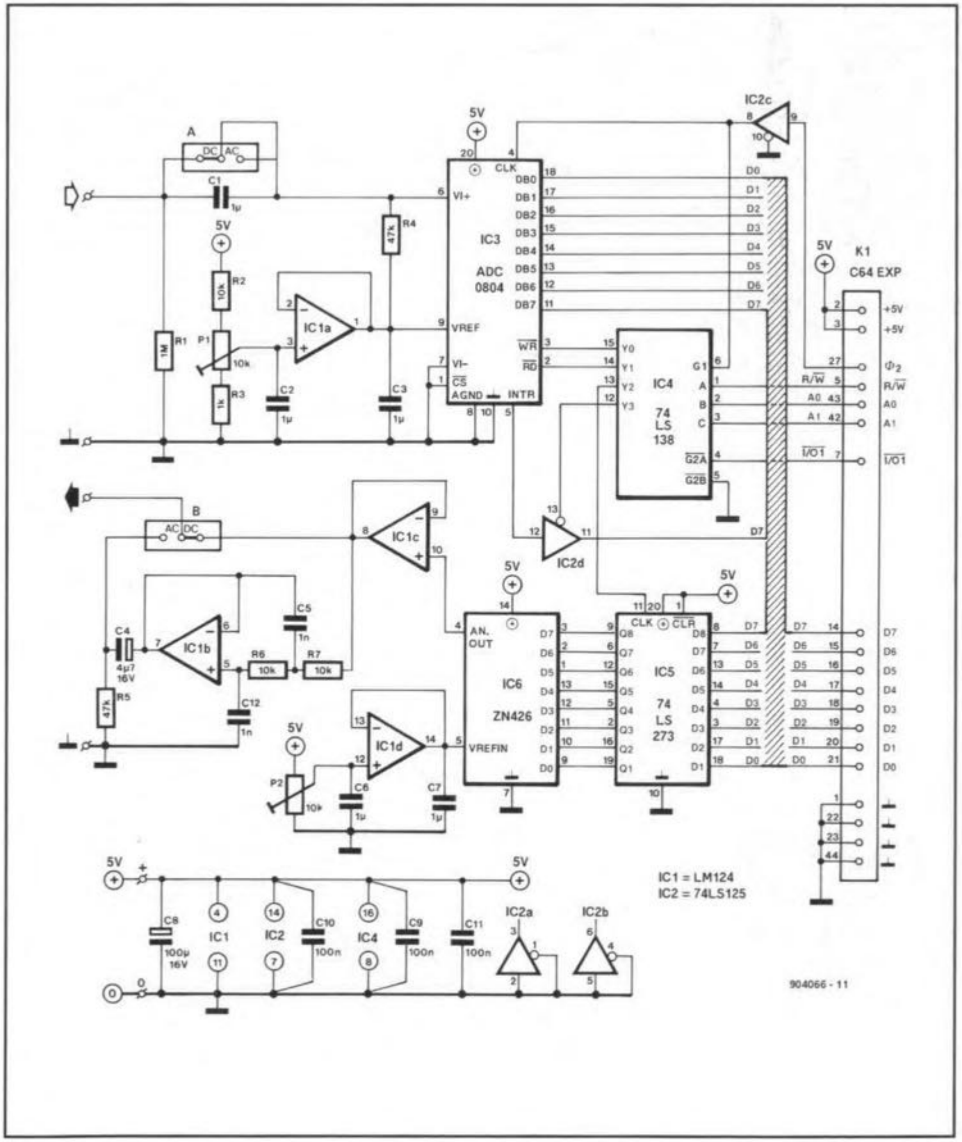
For processing analogue signals, virtually all computers need add-on units. The one presented here was designed for use with C64 machines, but it may be used with other computers with little or no change.

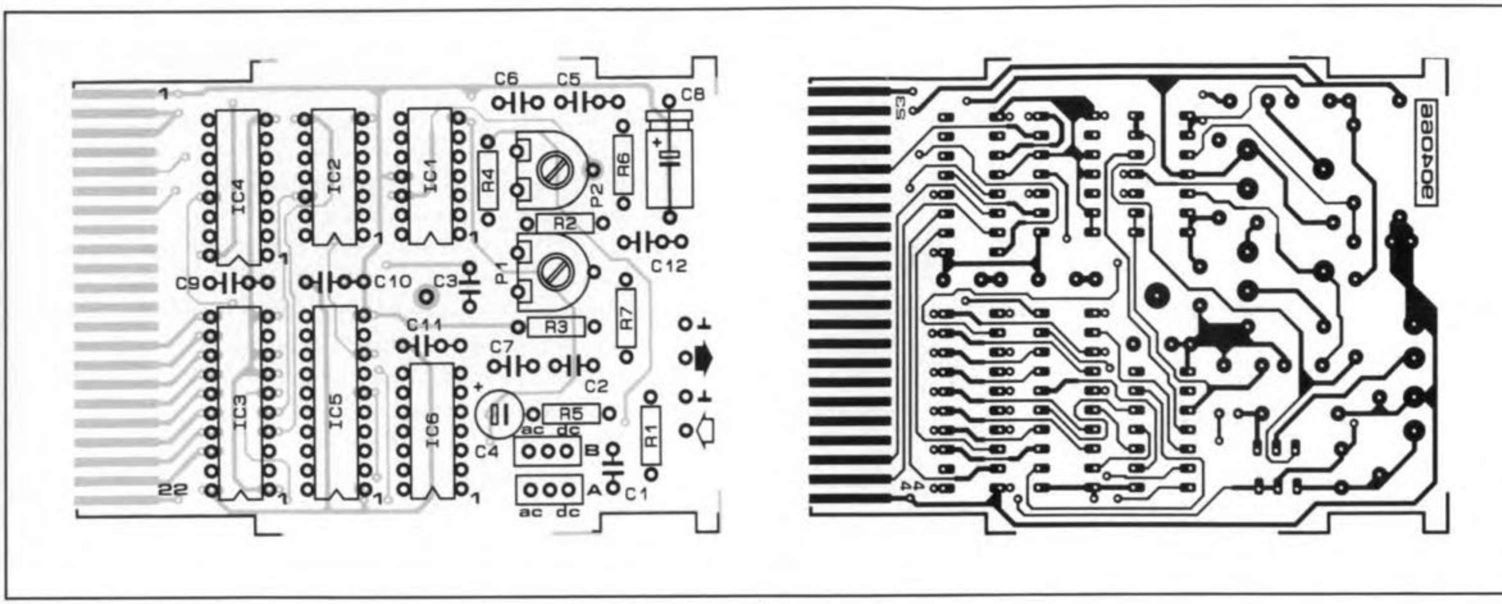
The card is intended for sampling and reproducing of sound or for measurements and production of test signals. It is an 8-bit design. This may sound simplistic in these days of 'more than 16 bits' CD players, but in practice an 8-bit design gives perfectly acceptable sound reproduction. As far as testing and measuring is concerned, accuracy is better than 0.5%.

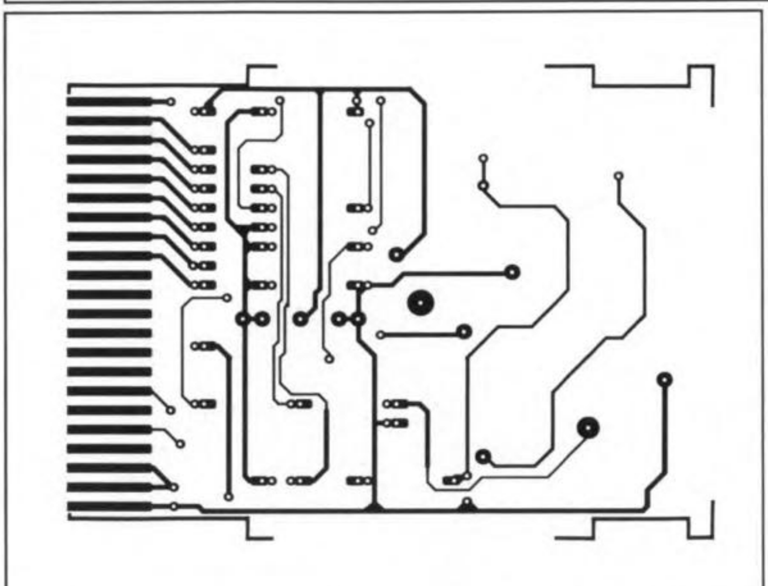
The circuit is based on address decoder IC4. Fed with signals $\overline{\mathrm{I/01}}$, A0, A1, R/$\overline{\mathrm{W}}$ and ϕ2, this IC ensures that read and write instructions to addresses DE00 and DE01 of the C64 fulfil the functions indicated in the table.

DE00	write	start A-D conversion
DE00	read	read result of A-D conversion
DE01	write	write data for A-D conversion
DE01	read	read status of A-D conversion (bit 7)

The analogue signal may be d.c. (for test and measurements) or a.c. (for audio signals) coupled to the input of the A-D convertor. The reference voltage of the converter is set with P1. This voltage must not be greater than 2.5 V: this level gives a range of 0–5 V for d.c. coupled signals and ±2.5 V for a.c. coupled ones. The reference voltage may be reduced proportionally for small input signals: this ensures that for







PARTS LIST

Resistors:
R1 = 1 M
R2, R6, R7 = 10 k
R3 = 1 k
R4, R5 = 47 k
P1, P2 = 10 k preset

Capacitors:
C1, C2, C3, C6, C7 = 1 μ
C4 = 4μ7, 16 V, radial
C5, C12 = 1 n
C8 = 100 μ, 16 V, axial
C9, C10, C11 = 100 n

Semiconductors:
IC1 = LM124
IC2 = 74LS125
IC3 = ADC0804
IC4 = 74LS138
IC5 = 74LS273
IC6 = ZN4226

those smaller signals the full 8-bit resolution is retained.

Analog-to-digital conversion is started by writing to address DE00. When that is finished, output INTR will go low: this state may be checked by reading address DE01 and ascertaining that bit 7 is 0. The result may be read by the computer at address DE00.

Digital-to-analog conversion is even simpler: the data are written to address DE01 and that's all.

The D-A convertor may also be d.c. or a.c. coupled.

With jumper B in position AC, the analog output signal is passed through a low-pass filter with a cut-off frequency of 15 kHz. This ensures that the sample frequency and its harmonics are suppressed during the reproduction of audio signals.

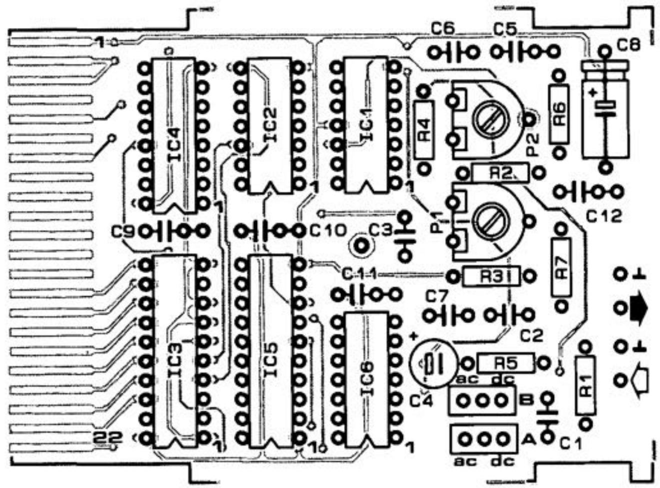
When test signals are produced, it is better not to use filtering: jumper B must then be set to position DC.

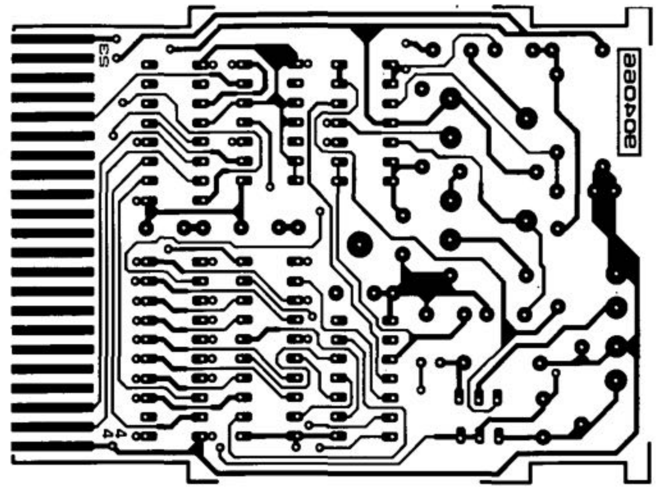
The reference voltage in the D-A convertor should also not be greater than 2.5 V: the level is set with P2.

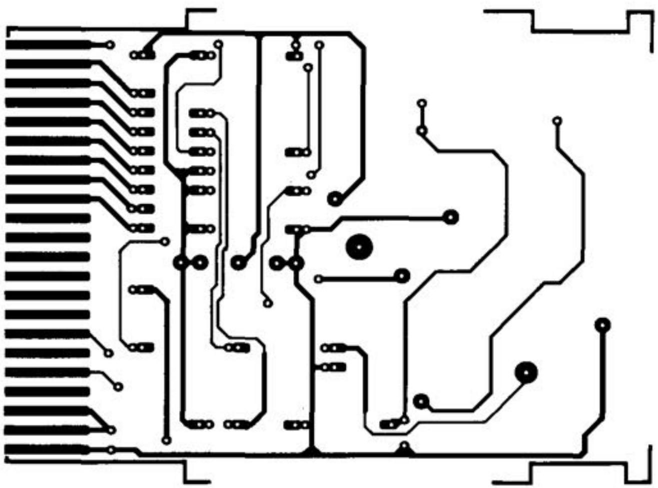
To enable the PCB to be 'through-plated', additional pads have been provided. Before any components are fitted, short lengths of bare circuit wire should be soldered to both sides of these pads. Once that is done, the components need soldering only at the track side of the board. ■

(*C. Kuppens*)

S 10







PARTS LIST

Resistors:
R1 = 1 M
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R3 = 1 k
R4, R5 = 47 k
P1, P2 = 10 k preset

Capacitors:
C1, C2, C3, C6, C7 = 1 μ
C4 = 4μ7, 16 V, radial
C5, C12 = 1 n
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C9, C10, C11 = 100 n

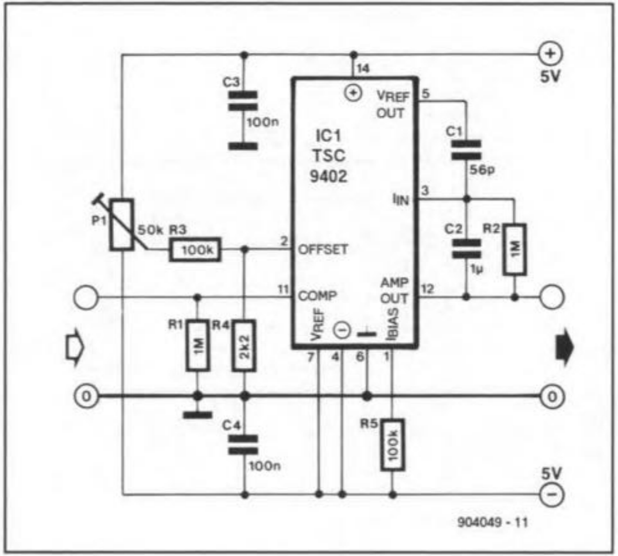
Semiconductors:
IC1 = LM124
IC2 = 74LS125
IC3 = ADC0804
IC4 = 74LS138
IC5 = 74LS273
IC6 = ZN4226

009 FREQUENCY TO VOLTAGE CONVERTER

Teledyne Semiconductor's Type TSC9402 is a versatile IC. Not only can it convert voltage into frequency, but also frequency into voltage. It is thus eminently suitable for use in an add-on unit for measuring frequencies with a multimeter. Only a few additional components are required for this.

There is just one calibration point to set the centre of the measuring range (or of that part of the range that is used most frequently).

The frequency-proportional direct voltage at the output (pin 12 – AMP OUT) contains interference pulses at levels up to 0.7 V. If these prove to have an adverse



fect on the multimeter, they may be suppressed with the aid of a simple RC network. The output voltage, U_o, is calculated by:

$$U_o = U_{ref}\,(C_1+12\ \text{pF})\ R_2\, f_{in} \qquad [\text{V}]$$

Since the internal capacitance often has a greater value than the 12 pF taken here, the formula does not yield an absolute value.

The circuit has a frequency range of d.c. to 10 kHz. At 10 kHz, the formula gives a value of 3.4 V.

The circuit draws a current of not more than 1 mA. ■

(T. Giffard)

A line amplifier is always a useful unit to have around, be it for matching a line signal or raising its level somewhat. This may be needed during a recording session or with a public-address system. Furthermore, a line mixer may be constructed from a number of these amplifiers.

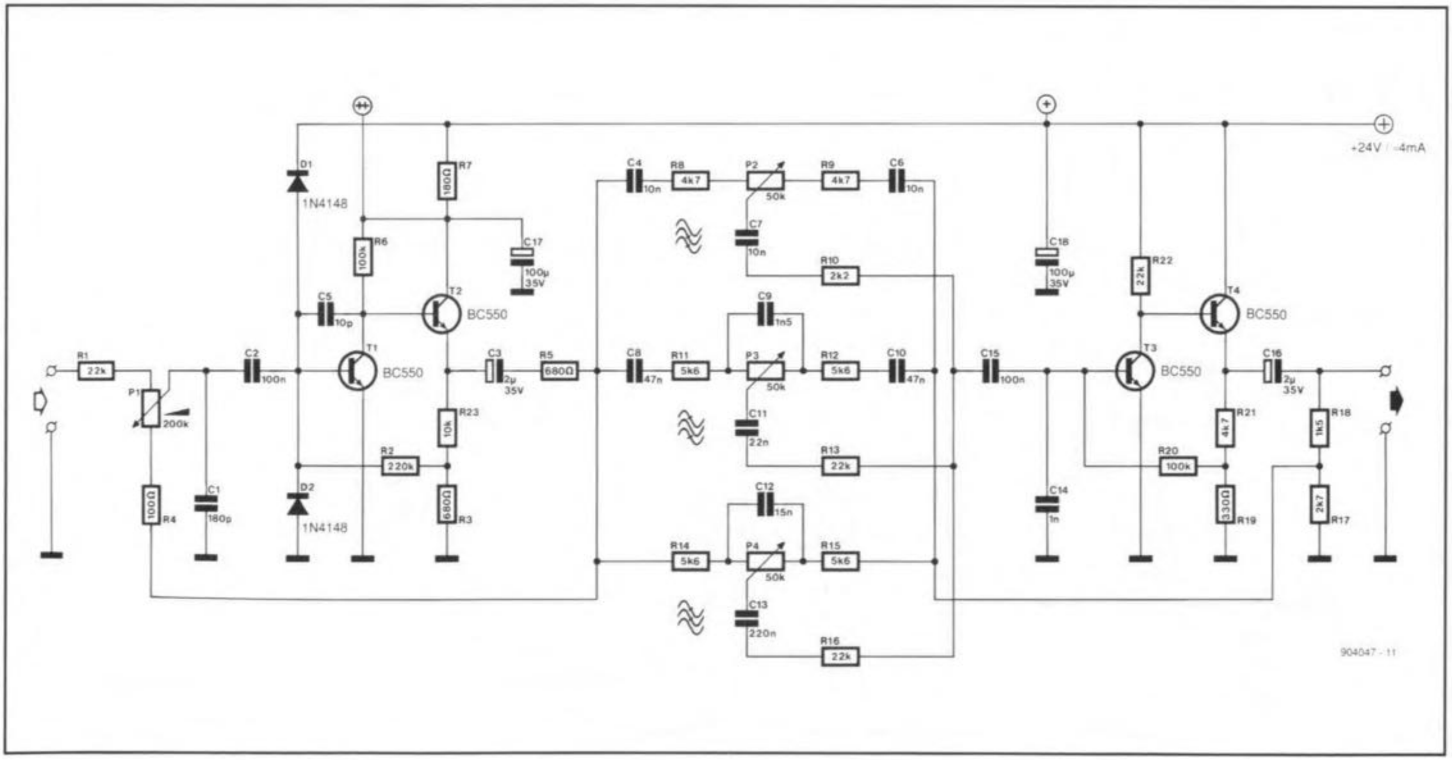
The input of the amplifier is proof against high voltages. The output impedance is low.

The circuit is a conventional design: two d.c. coupled stages of amplification separated by a three-fold Baxandall* tone control system. The volume control at the input is conspicuous by having its 'cold' side connected not to ground but to the output of the first amplifier. Because the signal there is out of phase with the input signal, the amplifier obtains negative feedback via P1. The amplification is therefore inversely proportional to the magnitude of the input signal. This makes it possible for the amplifier to accept a wide range of input levels. It is quite possible to input a signal taken direct from the loudspeaker terminals of a power amplifier.

The supply voltage is 24 V; at that voltage the amplifier draws a current of about 4 mA. If several amplifiers are used in conjunction (as, for instance, in a mixer panel), the various supplies (+ and ++ in the diagram) may be interlinked. Capacitors C17, C18 and resistor R7 need not be duplicated in that case. ■

(*A. Ferndown*)

* P.J. Baxandall, "Negative feedback tone control", *Wireless World*, **43**, 402, October, 1952; **43**, 444, November, 1952.

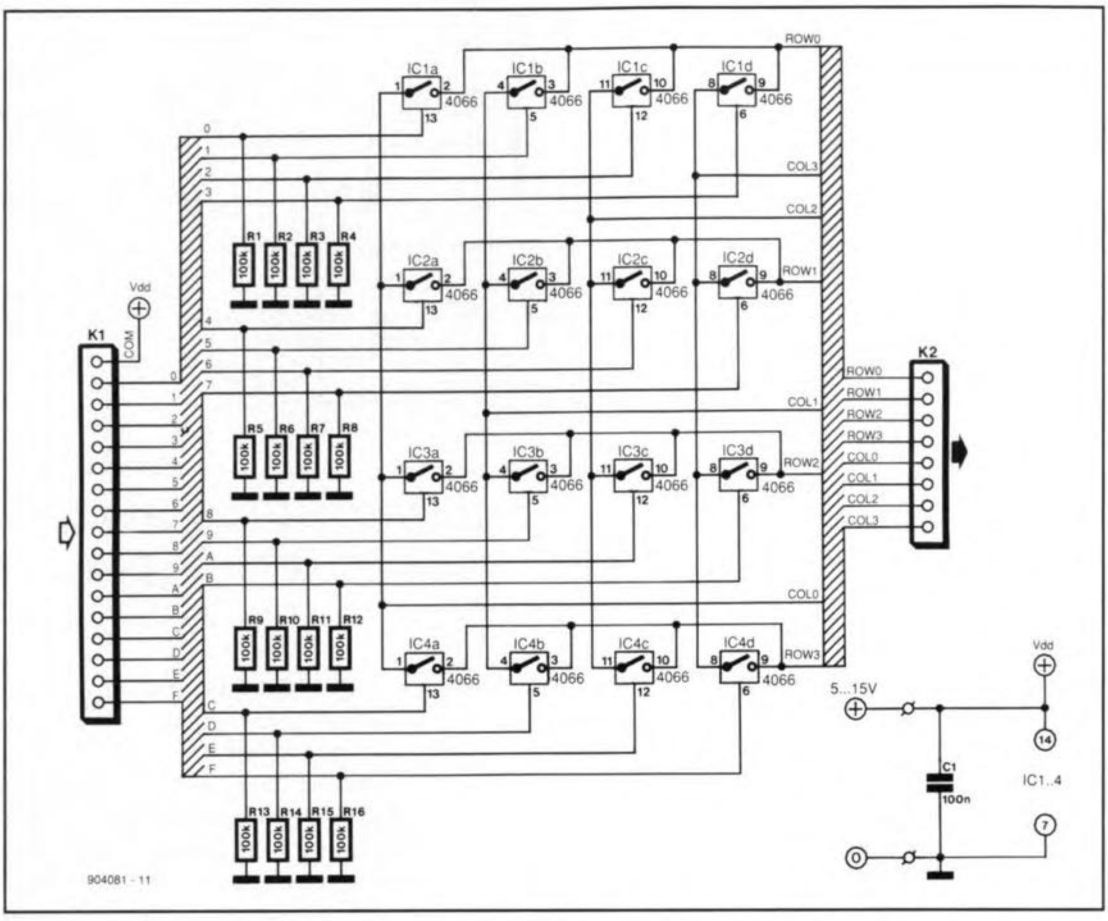


011 MATRIX INTERFACE FOR KEYBOARDS

Keyboards may be slotted into two categories, at least as far as the manner in which the switches are connected is concerned: those with a common connection and those with the switches arranged in a matrix.

The matrix type has the important advantage that the number of connections is an absolute minimum. Such an arrangement is ideal for ICs and many of these are therefore designed for use with a matrix keyboard.

However, there are many keyboards available in job lots, for instance, that apart from a common connection also have a connection for each key. Such keyboards may be connected to ICs that require a matrix type with the aid of a number of electronic switches.



The principle is straightforward: each key of the keyboard controls an electronic switch included in a matrix. As an example, the diagram shows a hexadecimal keyboard that is arranged in a 4×4 matrix. Each of the electronic switches is held in the open position by a pull-down resistor. If a key on the keyboard is pressed, the associated electronic switch closes.

The current drawn by the circuit is very small and is determined mainly by the value of the pull-down resistors and the number of keys being pressed. The CMOS switches draw virtually no current.

(*T. Giffard*)

012 SUPPLY VOLTAGE FROM 0 VOLT

The special characteristic of this regulator is that the output voltage may be adjusted down to 0 volt. The regulation is provided by an integrated regulator Type LM317. As is normal in supplies that can be adjusted to 0 V, this IC is used in conjunction with a zener diode. This diode provides a reference voltage that is equal but of opposite sign to the reference voltage U_r of the reg-

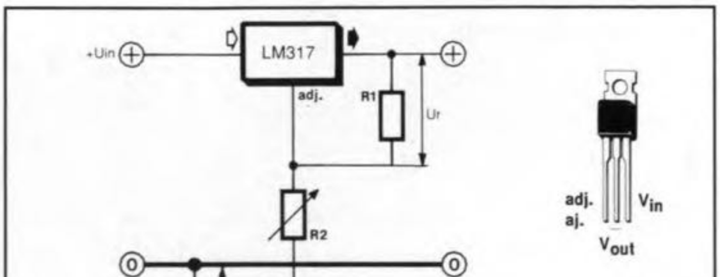


Fig. 1.

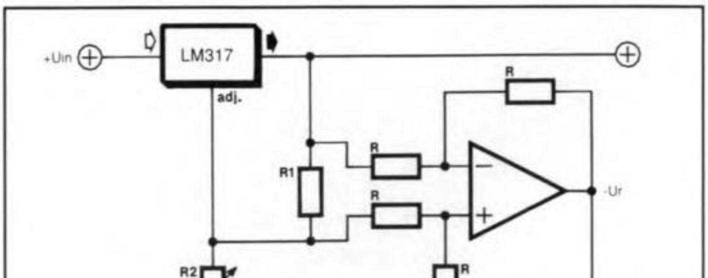
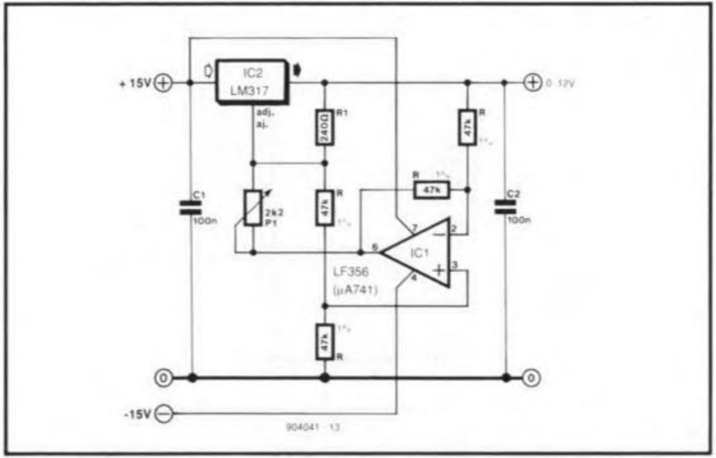


Fig. 2.

Fig. 3.



ulator, as shown in Fig. 1. Potential divider R1-R2 enables adjustment of the output voltage.

In the present circuit, the negative reference voltage is derived in a different manner: from the regulator with the aid of an opamp as shown in Fig. 2. The opamp is connected as a differential amplifier that measures the voltage across R1 and inverts this voltage to U_r. An additional advantage of this method is that at low output voltages a change in the reference voltage has less effect on the output voltage than the circuit in Fig. 1. The prototype, constructed as shown in Fig. 3, gave very satisfactory results.

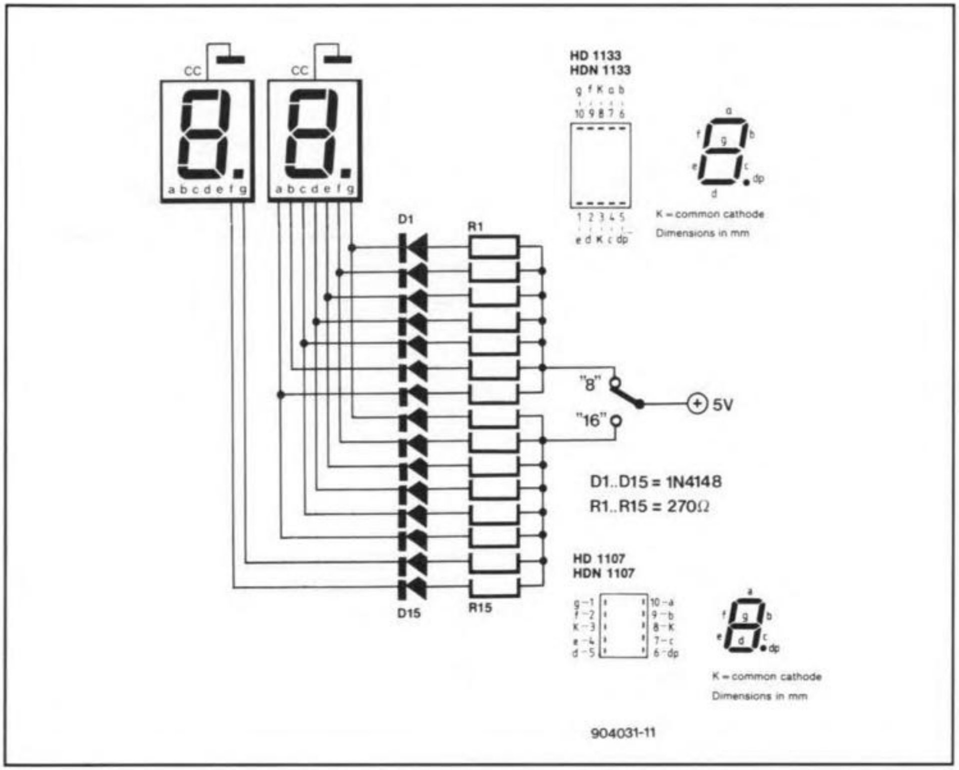
The opamp need not meet any special requirements: a Type μA741 works fine, although an LF356 gives a slightly better performance.

The negative supply for the opamp may be obtained with the aid of a centre-tapped mains transformer. ■

(*L. Nunnink*)

CLOCK FREQUENCY INDICATOR

013



The indicator is intended to display the clock frequency of a personal computer (PC) in megehartz. It consists of two common-cathode displays (Type HD1107 for 10 mm high figures; Type HD1133 for 13.5 mm high figures), a two-position switch and a number of Type 1N4148 diodes to control the lighting of the displays. Furthermore, to limit the current through the displays, a 270 Ω resistor is connected in series with each diode.

With the switch in position '8', the displays show the normal speed of the PC and in the lower position the'turbo' speed. With some dexterity, it is possible to use the turbo switch on the computer instead of the switch shown in the diagram. ■

(*A. Ferndown*)

CURRENT TO FREQUENCY CONVERTER

014

Teledyne Semiconductor's Type TSC9402 IC is eminently suitable for use as an inexpensive current-to-frequency converter. The maximum input current of the design shown in the diagram is 10 µA (input voltage range is 10 mV to 10 V), while the output frequency range extends from 10 Hz to 10 kHz. The conversion factor is exactly 1 kHz/µA. The factor may be altered by changing the value of R1, as long as the maximum input current of 10 µA is not exceeded.

The circuit has two outputs. That at pin 8 is a short-duration pulse whose rate is directly proportional to the input current, while that at pin 10 is a square wave of half the frequency of the pulse at pin 8.

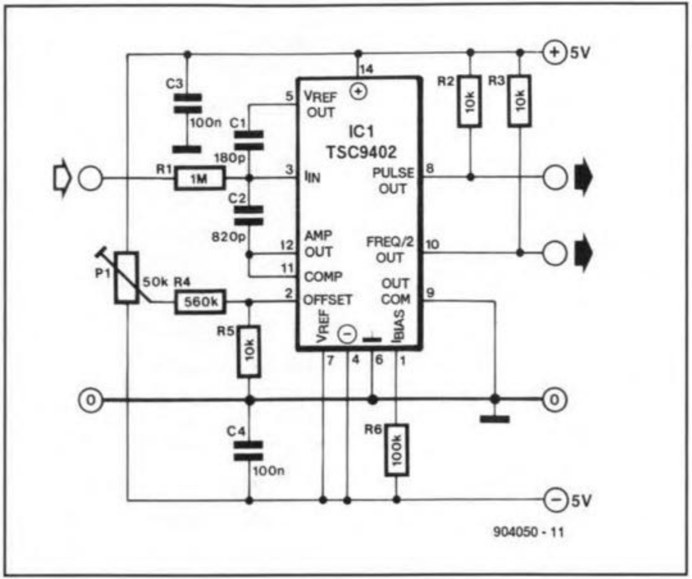
Calibrating the circuit is fairly simple. Connect a frequency meter to pin 8 (preferably one that can read tenths of a hertz) and connect a voltage of exactly 10 mV to the input (check with an accurate millivoltmeter). Adjust P1 to obtain an output of exactly 10 Hz. Next, connect a signal of exactly 10 V to the input and check that the output signal has a frequency of 10 kHz. If this frequency can not be attained, shunt C1 with a small trimmer or replace R1 by a resistor of 820 kΩ and a preset of 250 kΩ.

The circuit may be adapted to individual requirements with the aid of:

$$f_{out} = I_{in} U_r (C_1 + 12 \text{ pF}) \qquad \text{[Hz]}$$

The reference voltage, U_r, here is –5 V. ■

(T. Giffard)



015 MATCHING AMPLIFIER FOR ELECTRIC GUITARS

It often happens that an electric guitar has to be connected to a mixing panel, a tape deck or a portable studio. As far as cabling is concerned, that is no problem, but matching the high impedance of the guitar element to the low impedance of the line input of the mixing panel or tape deck is. Even the so-called high impedance inputs of those units are not suitable for the guitar output. When the guitar is connected to such an input, there is hardly a signal left for the panel or deck to process.

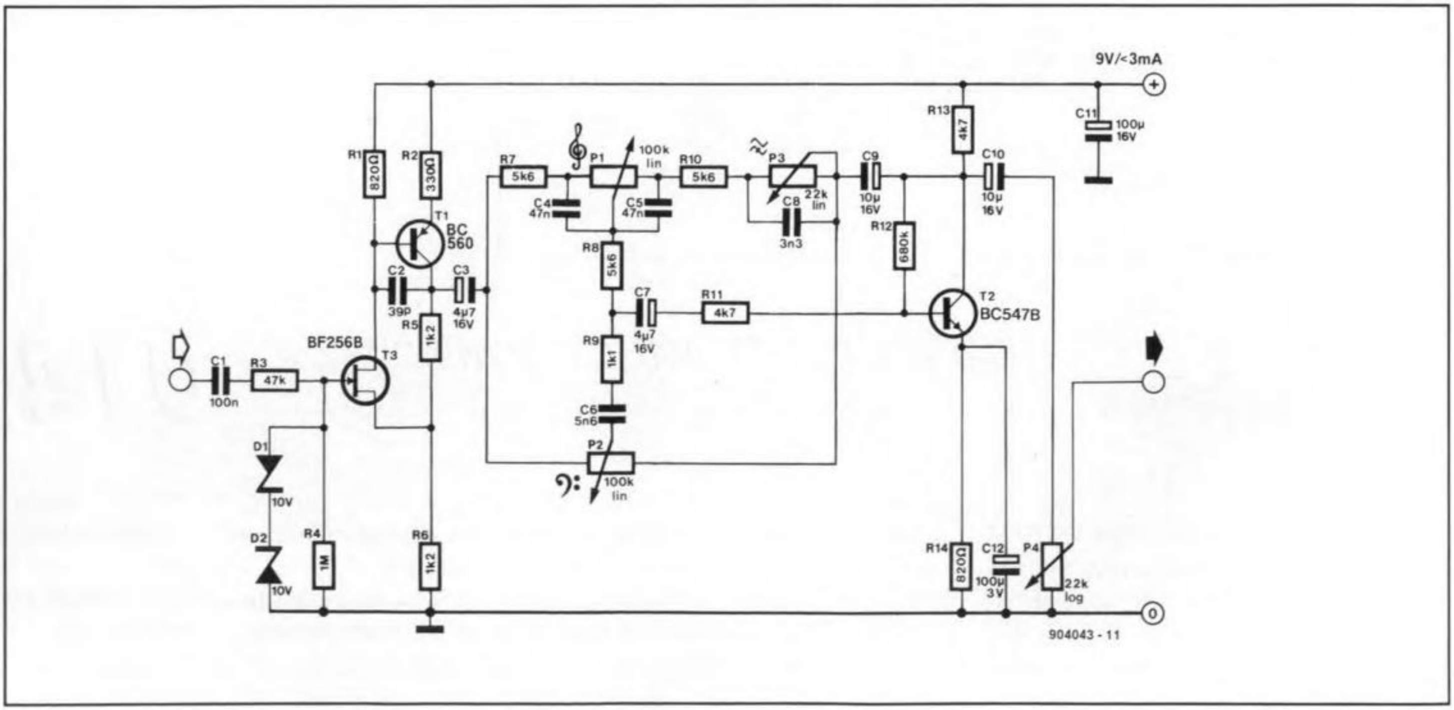
It would be possible to connect the guitar to the (high-impedance) microphone input, but that is normally far too sensitive for that purpose, so that clipping of the guitar signal occurs all too readily.

The matching amplifier presented here solves those problems: it has a high-impedance (1 MΩ) input that can withstand voltages of over 200 V. The output impedance is reasonably low. Amplication is ×2 (6 dB). Dual tone control, presence control and volume control are provided.

The circuit can handle input levels of up to 3 V. Above that level distortion increases, but that is, of course, a good thing with guitar music. Real clipping of the input signal does not occur until much higher levels than obtainable from a guitar are applied.

Power is supplied by a 9-V (PP3) battery from which the circuit draws a current not exceeding 3 mA. ■

(*A. Ferndown*)



DIGITALLY CONTROLLED MUTE CIRCUIT 016

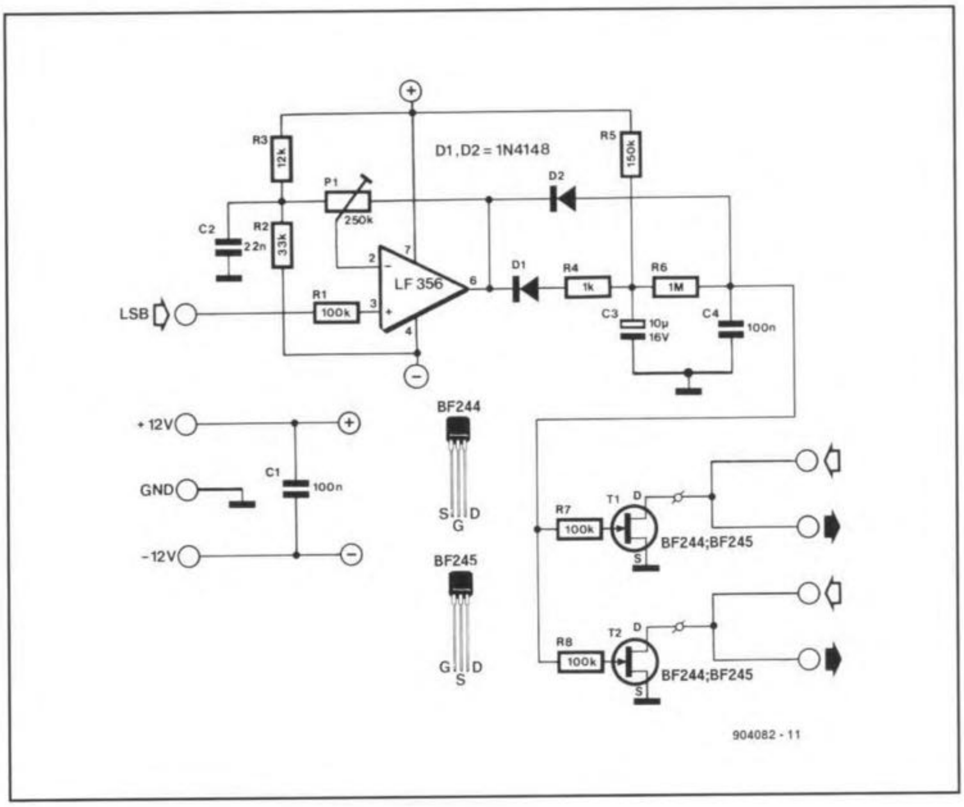
The mute circuit presented here is specially designed for use with the Roland MT32 module, although with some small alterations it should be suitable for use with other makes of expander or synthesizer. It is intended to eliminate the noise that the expander produces after a note-off. This noise, which remains audible, becomes pretty irritating after a while when the expander is used at home. For studio use a noise gate is, of course, used.

The circuit is intended to be fitted inside the MT32, for which there is ample space.

Muting proper is effected by two field-effect transistors (FETs) Type BF244 or BF245. These devices short the analogue output of the expander to ground when there is no signal.

The circuit is triggered by the data on the databus immediately preceding the digital-to-analogue (D-A) converter. The data are active low.

Data is taken from dataline D0 and compared with a 5-V reference voltage, provided by potential divider R2-R3, in IC1. When D0 is high, the circuit is inoperative and the output of the opamp is about +5 V.



The FETs obtain their gate voltage from the junction R6-C4-D2 via R7 and R8. Since that voltage is also around +5 V, the FETs conduct and short the output of the expander to ground.

When D0 goes low, the output of the comparator will also go low (negative). How low depends on the setting of P1. At that instant, C4 is discharged at once via D2 and the gate voltage of the FETs becomes negative. The FETs then switch off and the output signal of the expander is present. If this is a short, percussive signal, C3 will discharge only partially via D1. When D0 goes high again, the FETs will gradually begin to conduct. The rate of change of the gate voltages is determined by R6 and C4.

When the output signal of the expander is of longer duration or has considerable reverberation, the output of the opamp remains low long enough for C3 to discharge almost completely. This means that when D0 finally remains high, the rate of change of the gate voltages is much lower, because C3 must charge first via R5. This results in a gradual attenuation of the expander signal, so that a reverberation is not just cut off. In practice, the prototype performed very satisfactorily.

The circuit is powered by the ±12 V supply of the MT32, and draws a current of about 6 mA.

Preset P1 must be adjusted empirically to individual taste. ■

(A. Ferndown)

TWO-WIRE INTERCOM 017

With today's mains and FM duplex intercoms, the traditional circuit presented here creates an almost old-fashioned image. Nevertheless, it works very well, is easy to build and uses only standard parts and components.

The design consists of an amplifier, a double-pole change-over switch and two loudspeakers: one for the master station and one for the slave. More than one slave unit may be used, but each requires an additional change-over switch.

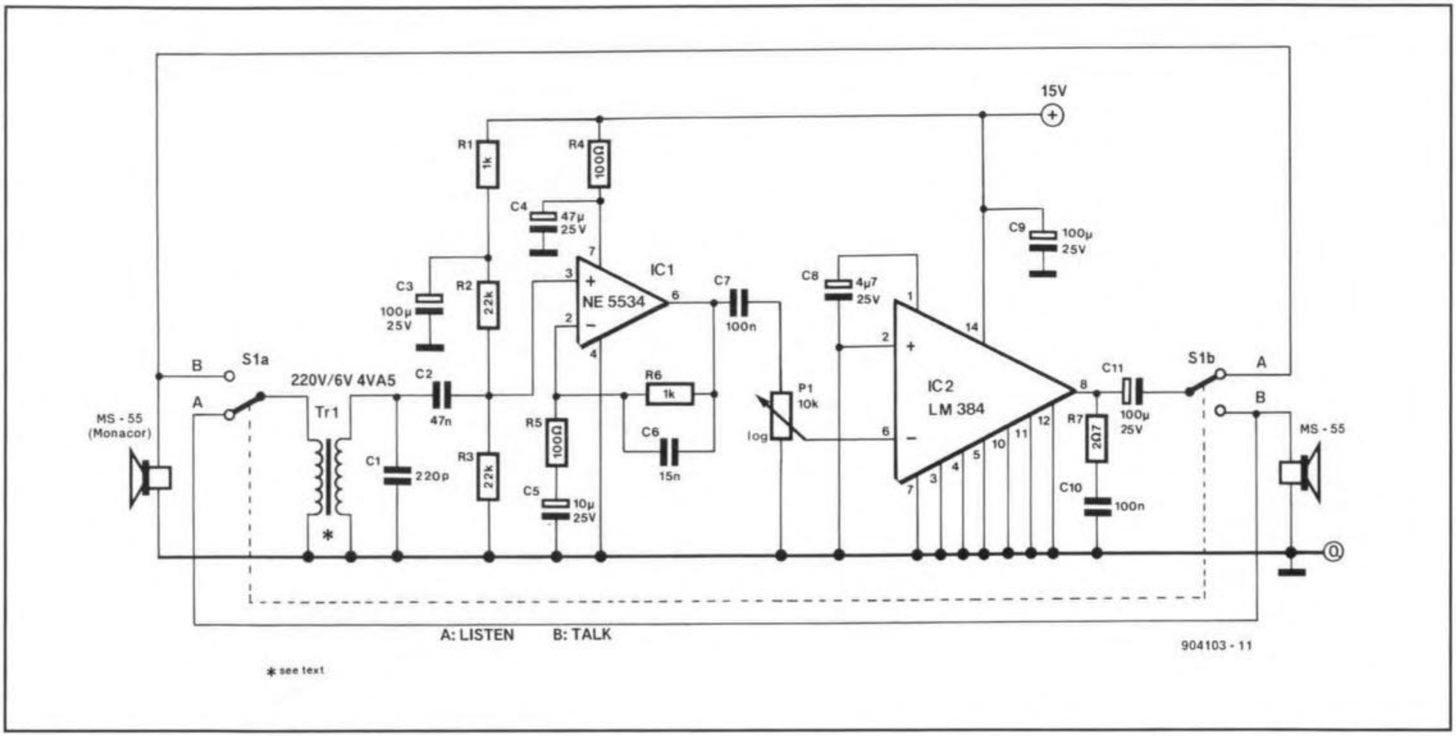
The power amplifier is a Type LM384, which can provide almost 2 watts output at a supply voltage of 15 V. Pins 3, 4, 5, 10, 11, and 12 are connected to ground and at the same time afford some cooling of the device. Because of that, the IC should not be fitted in a socket, but be soldered direct to the circuit board.

The LM384 processes signals with respect to earth so that an asymmetric supply suffices. The amplification has been set internally to ×50 (34 dB). The IC's supply line is decoupled by C9.

To ensure adequate input sensitivity, a preamplifier, IC1, is provided and this has an amplification of ×11 (21 dB). Because this stage is intended for speech only, its bandwidth is limited to 160 Hz to 10 kHz. Divider R2-R3 at the input of the opamp is decoupled by C3.

Special loudspeakers that can also serve

S 16



as microphones are readily available: in the prototype MS-55 units from Monacor were used, but there are a number of other makes that will do just as well. The bandwidth of the MS-55 used as loudspeaker extends from 150 Hz to 20 kHz and used as a microphone from 20 Hz to 20 kHz. The MS-55 can handle up to 5 W output.

To ensure satisfactory operation, particularly as a microphone, the loudspeaker must be fitted in a closed box.

Although it is advantageous that the 'microphone' has a low internal resistance, it makes it necessary for a transformer to be used at the input of the circuit. This has, however, the advantage that long cables may be used. The present circuit uses a standard mains transformer instead of a special microphone transformer. For this purpose, the secondary (6 V) winding is connected to the 'microphone'. The microphone impedance is thereby magnified from about 8 Ω to around 10 kΩ. The power handling of the transformer has been chosen quite high to ensure that signal losses in the primary winding are kept to a minimum. Capacitor C1 suppresses HF interference.

If the mains transformer and the 'microphone transformer' are housed in the same enclosure, some trial and error and screening are necessary to eliminate hum.

It may also happen that the 'microphone transformer' itself causes hum in the remainder of the circuit. In that case, the preamplifier stage must also be screened.

In the prototype, the speech bandwidth was limited to 400 Hz to 4 kHz and this proved perfectly acceptable for good speech transfer.

Most of the current drawn by the circuit flows through the power amplifier. At worst this amounts to 210 mA (680 mA peak), when the amplifier delivers 1.8 W output.

The LM384 can deliver a power of up to 5 W. The supply voltage should then be raised to 22 V and a heat sink for the device will be necessary. ■

(T.Giffard)

018 AUDIO INPUT SELECTOR

The design described here enables the selection of up to eight inputs of a preamplifier without any switch clicks or other noises. It may be used with virtually any preamplifier, provides individual switching of tape and line outputs, and enables monitoring of a tape recording. Morever, it needs relatively few components and is so compact that together with a high quality preamplifier it takes no more space than a typical car radio/cassette player.

Typical inputs and outputs of a preamplifier are shown in Fig. 1. There are seven line inputs that are switched into a signal bus by relays Re1–Re7. Tape OUT is also contained in the signal bus and may be switched by relay Re10.

Relay Re9 enables either the signal bus or the tape IN(put) to be connected to the preamplifier.

Relay Re8 switches the line OUT(put) on or off. To ensure that no switching clicks will be audible from the loudspeaker(s), each switching action at the input causes Re8 to switch off the line OUT(put). As soon as the switching action is completed, Re8 switches the line OUT into circuit again.

It must, of course, be possible, when a tape recording is being monitored, to

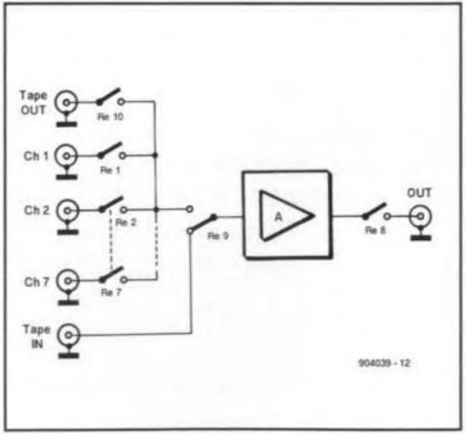
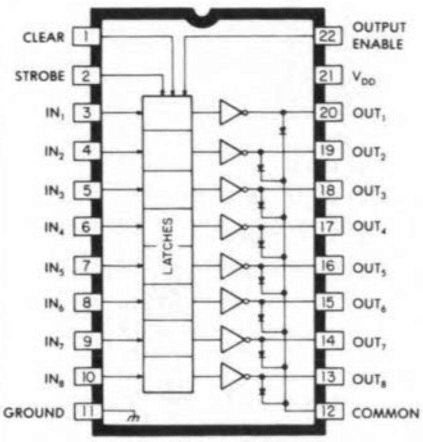
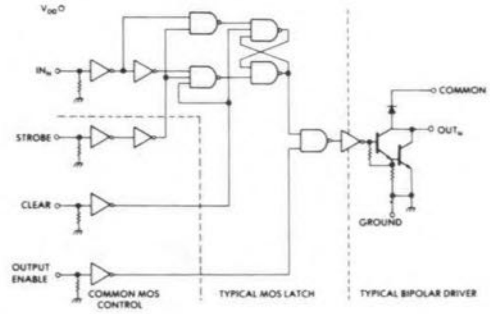


Fig. 1.





IN_N	STROBE	CLEAR	OUTPUT ENABLE	OUT_N t-1	OUT_N t
0	1	0	0	X	OFF
1	1	0	0	X	ON
X	X	1	X	X	OFF
X	X	X	1	X	OFF
X	0	0	0	ON	ON
X	0	0	0	OFF	OFF

904039 - 13

Fig. 2. Internal circuit, truth table and pinout of UCN5801A.

PARTS LIST

Resistors:

R1, R6, R7, R9 = 10 k
R2, R3 = 100 k
R4 = 27 k
R5 = 56 k
R8 = 2k2
R10–R19 = 180 Ω

Capacitors:

C1, C2, C3 = 100 n
C4 = 10 μ, 25 V

Semiconductors:

D1–D10, D19 = 1N4148
D11–D18, D20, D21 = 3 mm LED (in press-button switches)
T1, T2 = BC517
T3 = BC547
IC1 = UCN5801A (Sprague)
IC2 = 4027

Miscellaneous:

S1–S10 = press-button switch (with LED)
K1 = 11-way right-angled PCB edge connector
PCB 904039

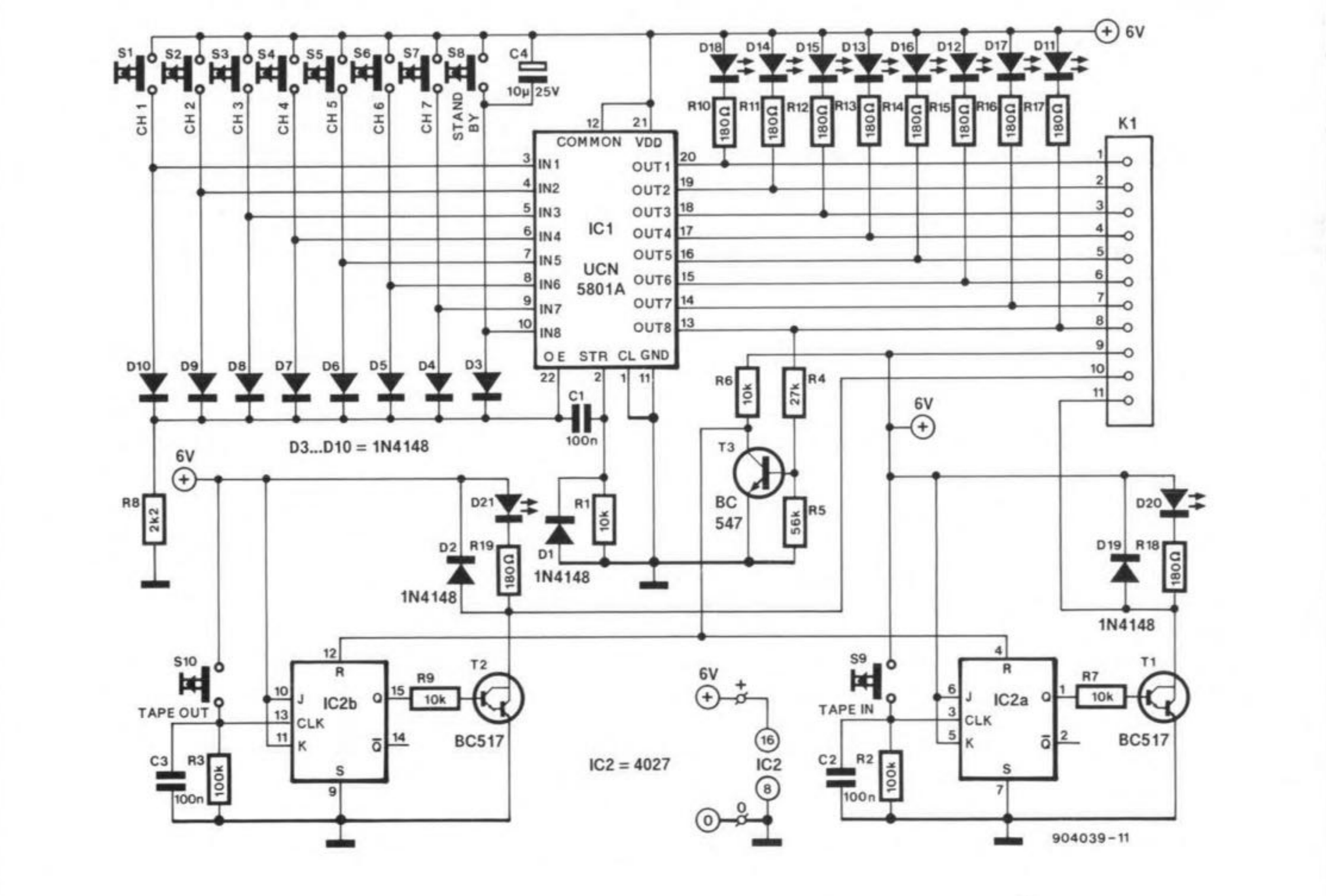


Fig. 3. Circuit diagram of the audio input selector.

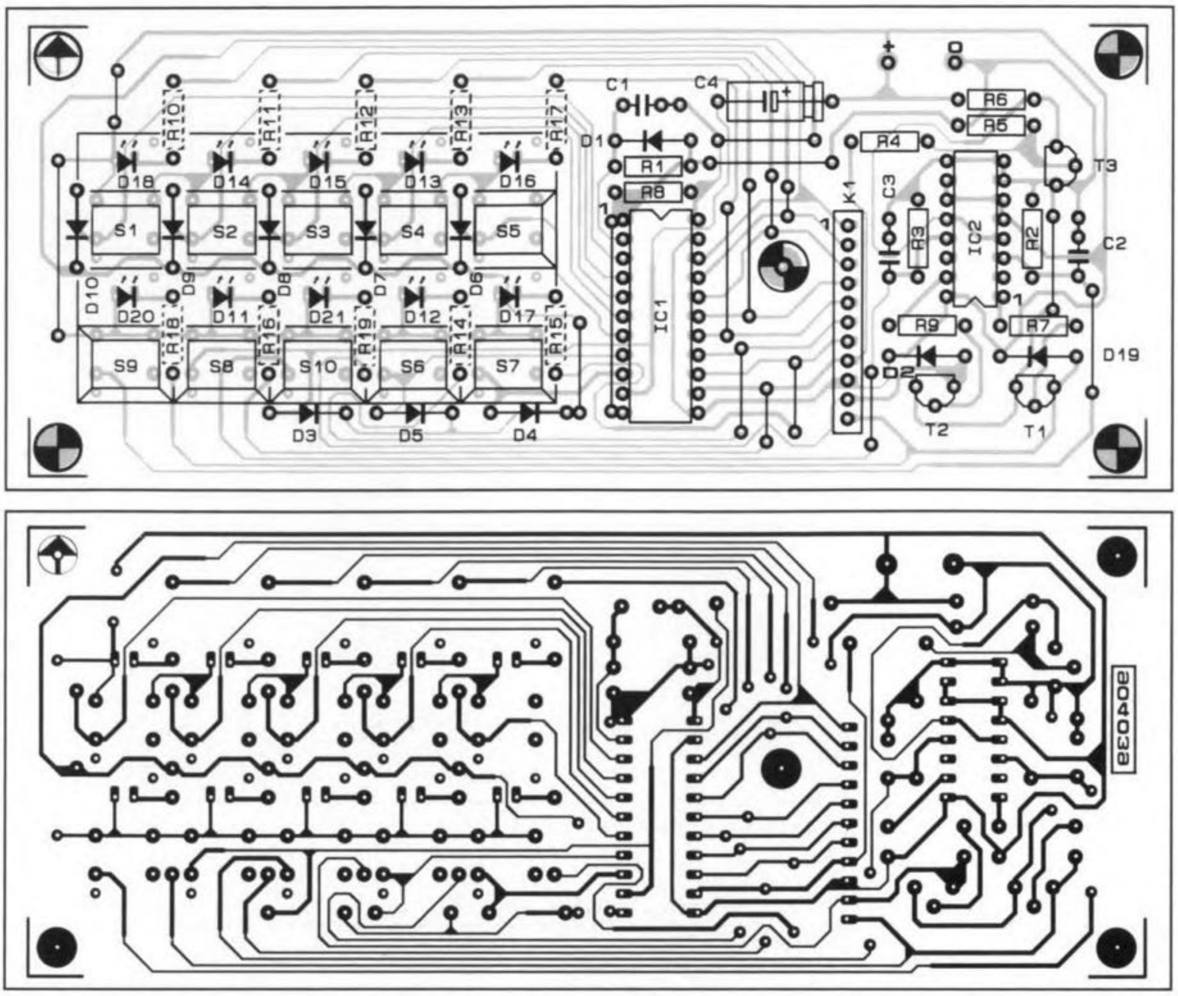


Fig. 4. Printed circuit board for the audio input selector.

switch between the recorded and recording signals without any delay.

The line output may be switched off manually. When the equipment is next switched on, the line output is also switched on but with a short delay.

The whole selection process is made possible by IC1, which is a special driver IC from Sprague, Type UCN5801A. This device contains eight identical latches with one individual (IN1–IN8) and three common inputs (CLEAR, STROBE and OUTPUT ENABLE). The latches are connected to darlington power drivers with open-collector outputs that can handle a continuous current of up to 400 mA. All inputs are provided internally with pull-down resistors and all power drivers with protection diodes. The internal circuit, truth table and pinout of the IC are shown in Fig. 2.

When no signals are being selected, all inputs of IC1, as well as OUTPUT ENABLE and STROBE are connected to ground via the pull-down resistors. CLEAR is permanently linked to earth.

When one of the switches S1–S8 is pressed, OUTPUT ENABLE goes high via the relevant diode and internal pull-down resistor. This level deactuates the NAND gate at the output of all latches, so that the driver transistors are switched off and all connected relays, including the LINE OUT, are deenergized. At the same time, the state of the input of the latch associated with the key being pressed changes; that input then waits for a '1' at the STROBE which will enable the input information to be written into the internal bistable.

Because STROBE always goes high after OUTPUT ENABLE, it suffices for a short, delayed pulse to be produced with the aid of network R1-C1.

When the key is released, the NAND gate at the output of the latch passes the logic state of the latch to the power driver: the relay of the associated signal source will then be energized.

When the selector key is pressed, the STAND BY key must be pressed at the same time, otherwise, although the input relay would be energized, the output relay would not. This is a protective arrangement that may be omitted by replacing D3 by a jump lead. If there is a requirement for independent on/off switching of the output relay, capacitor C4, NOT D3, should be replaced by a jump lead. Note, however, that in that case the default mode is lost, that is, when the supply is switched on, the logic state of the circuit will then be arbitrary.

Diode D1 at the STROBE input obviates a negative potential when the key is released and is therefore an essential component.

Diodes D11–D18 are integrated in the selector switches to show which input has been selected.

The UCN5801A does not arrange the actuation of relays Re9 and Re10: that is done by two identical J-type bistables contained in IC2, which provide a conventional on-off switch function under the control of switches S9 and S10. These circuits serve to switch the TAPE OUT(put) and the TAPE IN(put) terminals.

The pulses caused by the closing of the switches are applied to the CLOCK input of the relevant bistable via networks R3-C3 and R2-C2 respectively. SET (earthed), RESET (also to ground via T3), J and K (both at U_b) are switched in such a manner that each leading edge at the CLOCK input results in a change of state at the Q output. Darlingtons T1 and T2 are power drivers for the relevant relay; LEDs D20 and D21 indicate whether the associate relay is energized.

Inverter T3 arranges for both outputs to be switched to logic 0 when the LINE OUT relay has been switched off by the STAND BY selector switch.

The printed-circuit board has been designed for fitting immediately behind the front panel of the relevant preamplifier. The hole next to IC1 is intended for the spindle of P1—take care that if this is a metal one it cannot make contact with the track surrounding the hole. Note that the terminals of the three transistors should be bent at right angles before these devices are fitted to the board. It is best to fit the ICs in appropriate sockets. The bias resistors for the LEDs should be fitted on appropriate solder pins at the track side.

The power supply is 6 volt to ensure smooth operation of the 5-V relays. Both ICs can stand up to 15 V, but if the supply voltage is altered, the value of the LED bias resistors should also be changed. The circuit should not be powered by the supply of the preamplifier to prevent current pulses caused by switching operations penetrating into signal lines and thus causing unwanted noise in the speaker(s).

If eight inputs are not enough, the circuit may be doubled. Apart from the supply voltage lines, the STROBE and OUTPUT ENABLE lines on the two boards should be interlinked. Except for IC1, only S1–S8, D3–D18 and R10–R17 need to be used on the second board: all other components may be omitted. ■

(*P. Coster*)

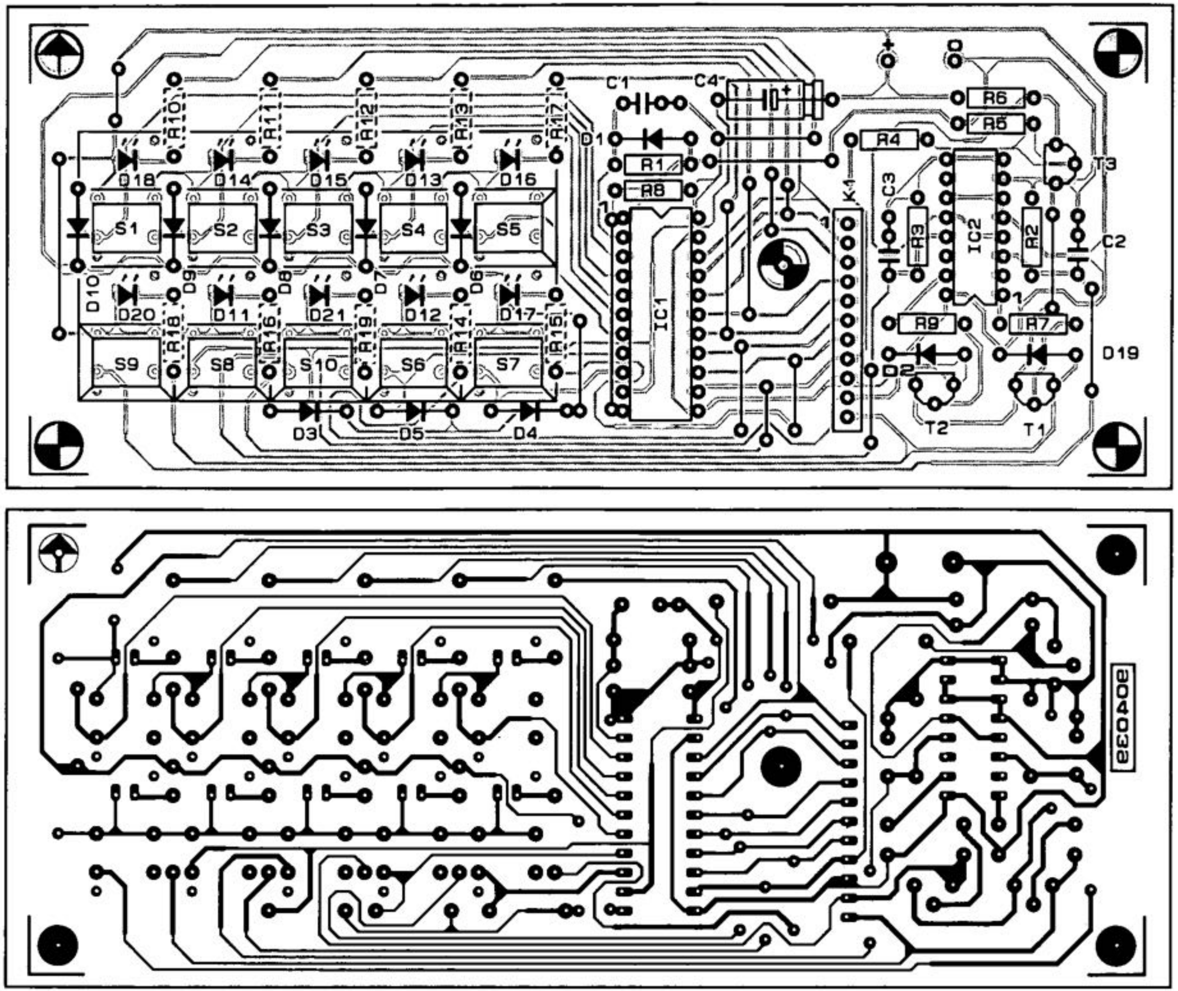


Fig. 4. Printed circuit board for the audio input selector.

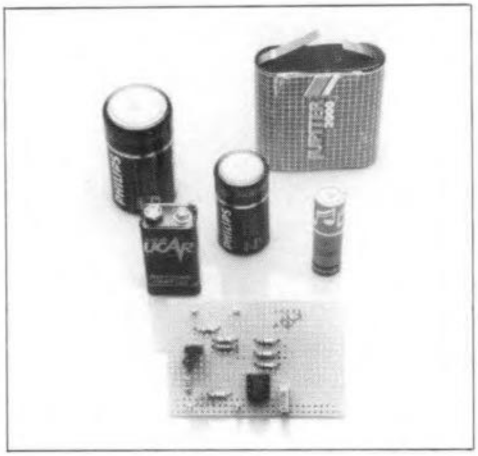
There are many times that a designer needs to know the value of the internal resistance of a battery. There are quite a few testers that give a relative indication of the value, but this is seldom in ohms. The present tester can, in principle, provide that information.

The basic idea behind it is to load the battery with a varying current so as to cause an alternating-voltage drop across the internal resistance that can be measured at the battery terminals. Provided the current variations are regular and constant, the voltage drop is directly proportional to the internal resistance.

By choosing the variation of the current carefully, it becomes possible to read the value of the internal resistance directly on the scale of an a.c. voltmeter.

The load current is varied with the aid of a current source, T1 in the diagram, which is switched on and off by square-wave generator IC1. The chosen switching frequency of 50 Hz ensures that the a.c. component at the battery terminals can be measured by a standard a.c. voltmeter (universal meter).

The battery is loaded constantly by R8, which has a value of 1.5 Ω for 1.5 V batteries, shunted by the a.c. voltmeter. The indicated voltage times ten is the value of the internal resistance of the battery. When the battery under test is flat, or if the supply battery is flat, no current flows and the meter will read zero. It would then appear as if the battery under test is an ideal type without internal resistance.

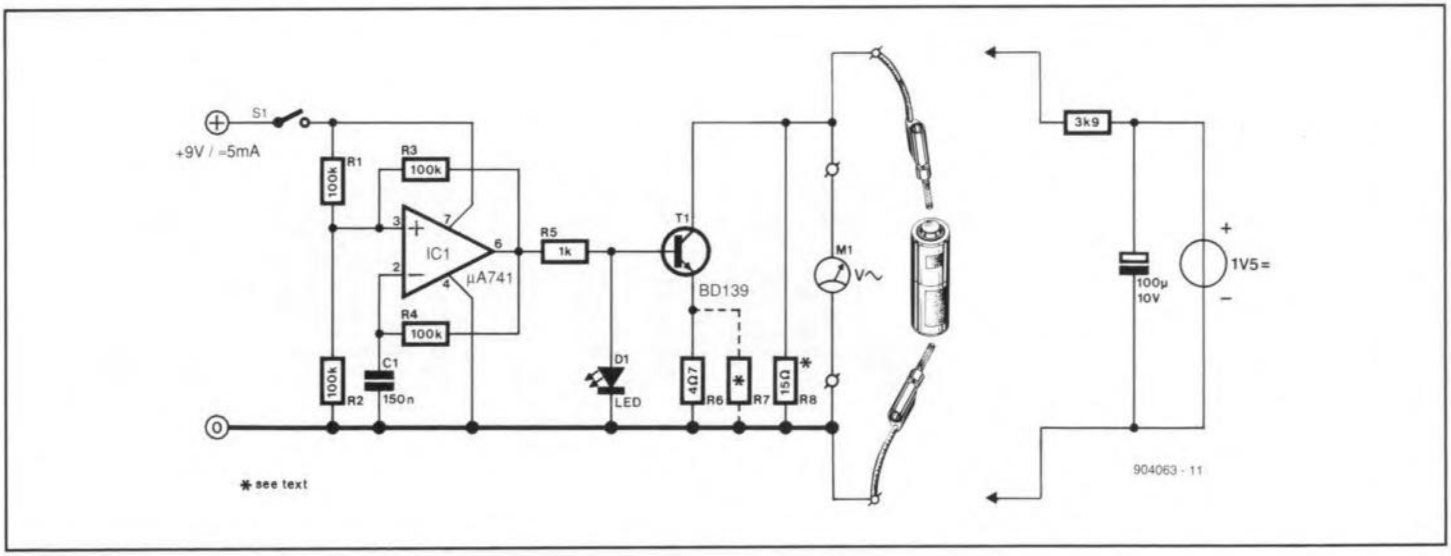


A flat supply battery is indicated by the not lighting of D1. Whether the battery under test is flat may be ascertained by measuring the direct voltage across its terminals. The load must be left connected, of course, otherwise the e.m.f. is measured and this may well be 1.5 V even if the battery is flat.

The tester is calibrated with the aid of the auxiliary circuit shown at the extreme right in the circuit diagram. The 1.5 V supply and electrolytic capacitor form a virtually ideal voltage source, of which the 3.9 Ω resistor forms the internal resistance. With this source connected across the output terminals of the tester, a suitable value should be ascertained for R7. That value is found when the a.c. voltmeter shows 0.39 V. Note that this procedure is not the same for all measuring instruments: the alternate use of a digital and a moving coil meter, for instance, is not feasible.

The tester is intended for 1.5 V batteries. The load current is fairly high: about 100 mA through R8 and around 170 mA through T1. For 9-V batteries that is rather too much: the current should then be reduced by taking greater values for R6–R8.

(K. Walters)



020 DIRECTION DETECTOR

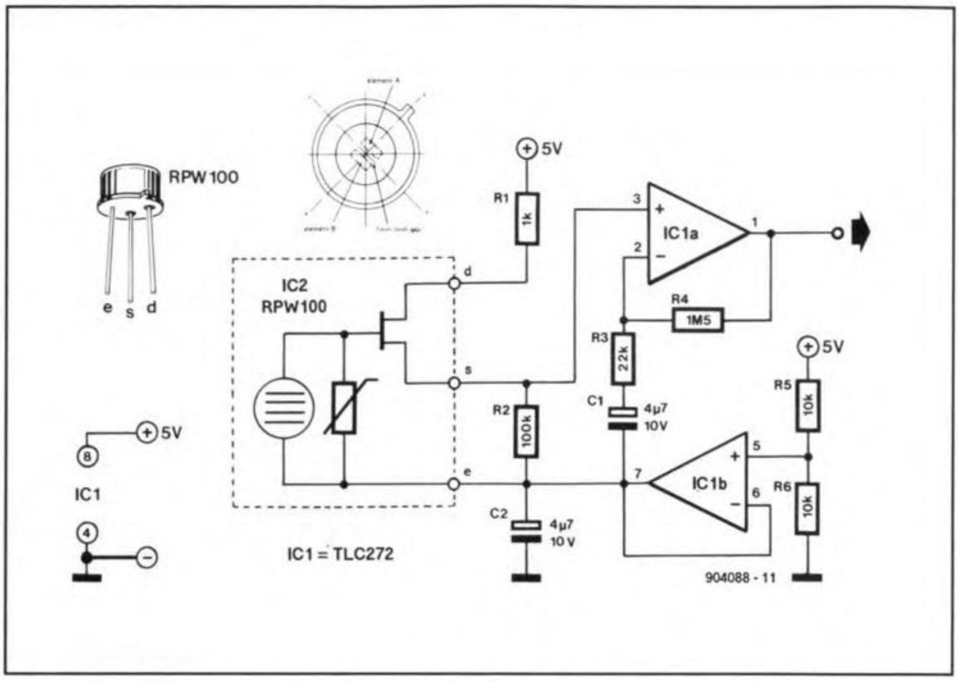
A heat-sensitive sensor may be used to construct a direction detector. Such a sensor reacts to all animal heat. The one used in the present design has a sensitive surface that has been divided into two, and it makes a difference therefore whether the heat approaches from the left or the right. The indication for cold objects is, of course, exactly the opposite.

Circuit IC1b forms a symmetric supply. Terminal 's' of the sensor is its output. The signal at 's' is amplified in IC1a by a factor of about 70 before it is available at the output of the detector.

To obtain good directivity, it is best to place the sensor behind a single narrow slit rather than behind the usual raster or multi-facetted mirror.

The circuit draws a current of only a few milliamperes from a 5 V supply ■

(*K. Walters*)

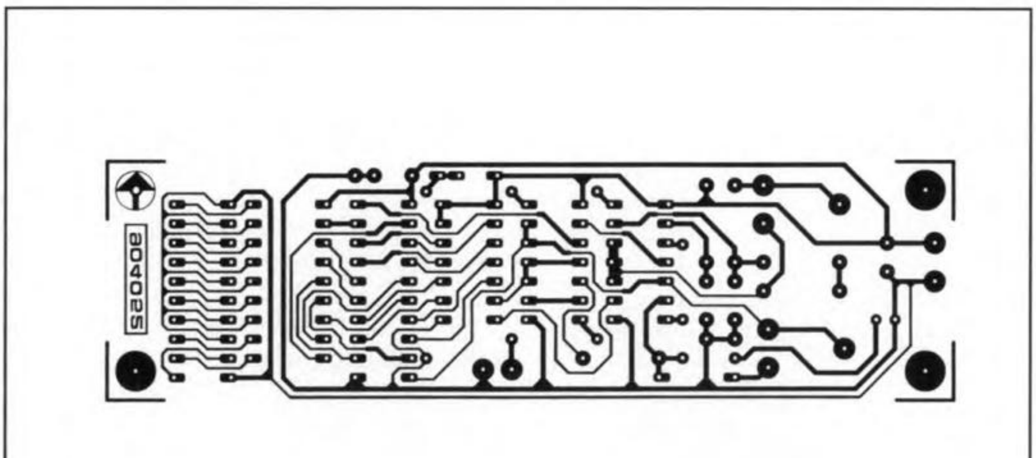


021 SEARCH TIME MONITOR

On hard disks with an ST506 compatible interface, there is a 'seek complete' ($\overline{SC}$) signal available. This signal is inactive high when the hard disk seeks new data. The duration of the high interval is thus directly proportional to the wasted search time. The length of that search time is determined primarily by the time required to shift the heads to the desired cylinder. Hard disks with an access time of 68 ms use search times of between 5 ms and 200 ms. By measuring the search time at each access and make this visible with the aid of an LED bar, an impression is gained of the performance of the hard disk. The length of the bar increases appreciably with fragmenting of the disk. If the bar is 'in the red' often, it's getting time to run a disk optimalization program!

Signal $\overline{SC}$ is connected to pin 8 of the ST506 interface. Each odd-numbered pin is connected to earth.

When the heads begin a search, that is, after $\overline{SC}$ has gone high, the clock input of buffer/register IC1



goes low. At the same time, because of the rapid discharge of C2 (via D1) and the going low of pin 1 of IC4c, the clear state of IC4 is removed. Because of the delay in the gates, the clear input of shift register IC2 will still be inactive when oscillator IC4a begins to operate. Dependent on the oscillator frequency and the time during which $\overline{SC}$ is high, IC2 will be clocked a couple of times. At each clock pulse, a '1' is shifted into this IC, so that at the end of the cycle the number of actuated outputs is a measure of the search time.

One gate-delay after the trailing edge of $\overline{SC}$, the buffer/register appropriates the data from the shift register. Another gate-delay later the shift register is cleared. It is then ready to measure the duration of the next search cycle. Only when each search cycle has been completed is the measured time indicated on the LED bar. This method obviates any irritating flickering of the display.

When $\overline{SC}$ has gone low, C2 is charged slowly via P2 and R3. If no new search cycle is started, the output of IC4c toggles from high to low after a short delay and this resets IC1. The setting of P2 thus determines the period during which the last measured search time is shown on the display (max. 1.5 s).

Resistor R13 and capacitor C6 provide a power-up reset for IC1.

The oscillator frequency may be set between 15 Hz and 400 Hz with P1. This preset should be adjusted in such a manner that during the maximum search time all LEDs just light. This is done with the aid of the assembler program shown. This program reads the first and last sector of disk C one after the other. A counter has been provided to limit the shifting around of the heads to 200 times. That gives ample time to adjust P1 as required. ■

(*J. Ruffell*)

```
; NAME: SEC.ASM       ***   SEEK-TIME MONITOR    ***
; USE :               *** P1 ADJUST AND TESTDEMO ***
; ID  : ELEKTOR/JR
; DATE: 8-2-'90

PROGSEG GROUP   CODE_SEG, DATA_SEG
        ASSUME  CS:PROGSEG, DS:PROGSEG

CODE_SEG        SEGMENT
 *      ORG     100H

MAIN            PROC NEAR
        LEA     SI,STRMSG       ;WRITE START MESSAGE
        CALL    MESSAGE
        MOV     DX,00H          ;READ BOOT SECTOR (SECTOR 0)
        CALL    READSEC
        MOV     BX,13H          ;DX:=TOTAL NUMBER OF DISK SECTORS
        MOV     DL,SEC[BX]
        MOV     DH,SEC[BX+1]
        DEC     DX              ;DX:=LAST SECTOR NUMBER
        CALL    READSEC         ;CHECK LAST SECTOR
WHILE:  CMP     DI,00H          ;WHILE ERROR FLAG=SET DO
        JZ      EWHILE
DO:     LEA     SI,ERRMSG       ;WRITE ERROR MESSAGE
        CALL    MESSAGE
        DEC     DX              ;LAST SECTOR NUMBER:=LAST SECTOR NUMBER-1;
        CALL    READSEC         ;READ LAST SECTOR
        JMP     WHILE
EWHILE: LEA     SI,ADJMSG       ;WRITE ADJUST MESSAGE
        CALL    MESSAGE
        MOV     CX,200.         ;LOOP COUNTER:=200 (TIME-OUT VALUE)
NEXT:   PUSH    DX              ;SAVE LAST SECTOR NUMBER
        MOV     AH,06H          ;IF ANY KEY PRESSED THEN END
        MOV     DL,0FFH
        INT     21H
        JNZ     END
        MOV     DX,0            ;READ SECTOR 0
        CALL    READSEC
        POP     DX              ;READ LAST SECTOR
        CALL    READSEC
        LOOP    NEXT            ;LOOP COUNTER:=LOOP COUNTER-1
                                ;IF NOT LOOP COUNTER<>0 THEN NEXT
END:    LEA     SI,ENDMSG       ;WRITE END MESSAGE
        CALL    MESSAGE
        INT     20H             ;RETURN TO DOS
MAIN            ENDP

;*******************************************************************

READSEC         PROC NEAR       ;INPUT :DX = SECTOR TO READ
                                ;OUTPUT:DI = ERROR FLAG
        PUSH    AX
        PUSH    BX
        PUSH    CX
        MOV     AX,2            ;SELECT DRIVE C (LOAD "3" TO SELECT "D")
        LEA     BX,SEC          ;GET SECTOR DUMP ADDRESS (DS:BX)
        MOV     CX,1            ;ONLY ONE SECTOR
        INT     25H             ;READ SECTOR (DX)
IF1:    JNC     ELSE1           ;IF READ ERROR
THEN1:  MOV     DI,01           ;SET ERROR FLAG
        JMP     EIF1
ELSE1:  MOV     DI,00           ;CLEAR ERROR FLAG
EIF1:   POPF                    ;RESTORE FLAGS
        POP     CX
        POP     BX
        POP     AX
        RET
READSEC         ENDP

;*******************************************************************

MESSAGE         PROC NEAR       ;INPUT: SI=MESSAGE ADDRESS
        CALL    CRLF
PUSH    DX
        PUSH    AX
        MOV     DX,SI
        MOV     AH,09
        INT     21H             ;WRITE MESSAGE
        POP     AX
        POP     DX
        RET
MESSAGE         ENDP

;*******************************************************************

CRLF            PROC NEAR
        PUSH    AX              ;SAVE REGISTERS
        PUSH    DX
        MOV     AH,02           ;CR
        MOV     DL,0DH
        INT     21H
        MOV     DL,0AH          ;LF
        INT     21H
        POP     DX              ;RESTORE REGISTERS
        POP     AX
        RET
CRLF            ENDP

CODE_SEG        ENDS

;*******************************************************************

DATA_SEG        SEGMENT
STRMSG  DB      ">>$"
ADJMSG  DB      "PLEASE ADJUST P1 (LED BAR TO FULL SCALE)."
        DB      "        Quit = Any key$"
ERRMSG  DB      "DISK READ ERROR: AUTOFIXING!!$"
ENDMSG  DB      ">> Program terminated$"
MARK    DB      "SECTOR  DUMP    "
SEC     DB      "512       BYTES   "
DATA_SEG        ENDS

;*******************************************************************

        END     MAIN
```

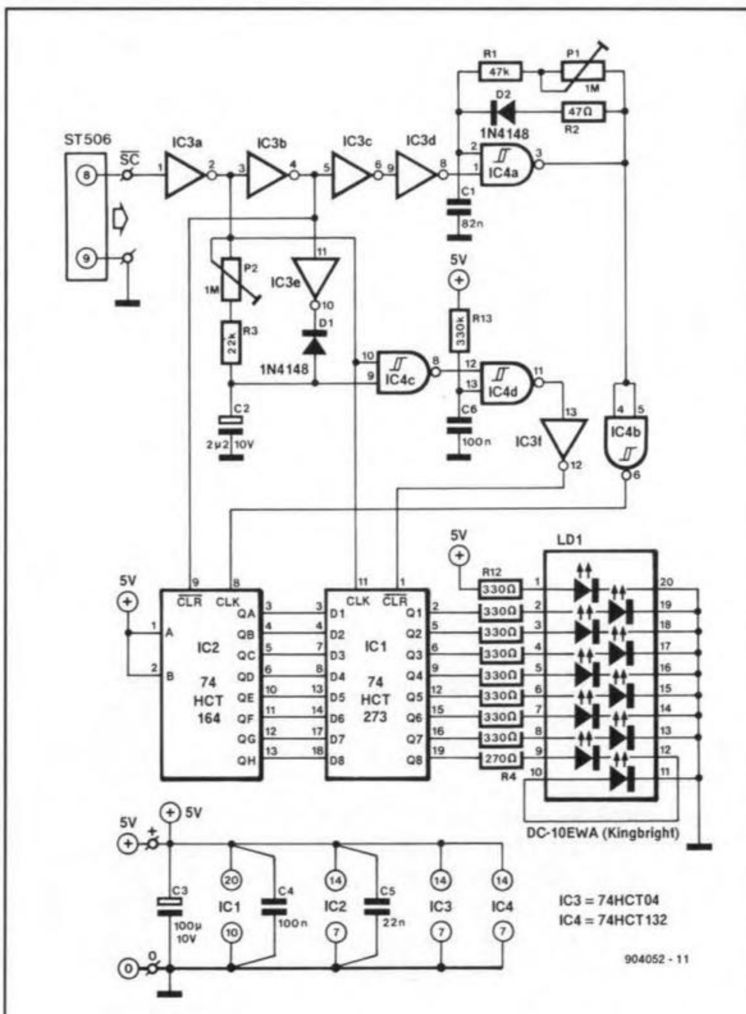
904052 - 12

PARTS LIST

Resistors:
R1 = 47 k
R2 = 47 Ω
R3 = 22 k
R4 = 270 Ω
R5–R12 = 330 Ω
P1, P2 = 1 M preset

Capacitors:
C1 = 82 n
C2 = 2µ2, 10 V, radial
C3 = 100 µ, 10 V, radial
C4, C6 = 100 n
C5 = 22 n

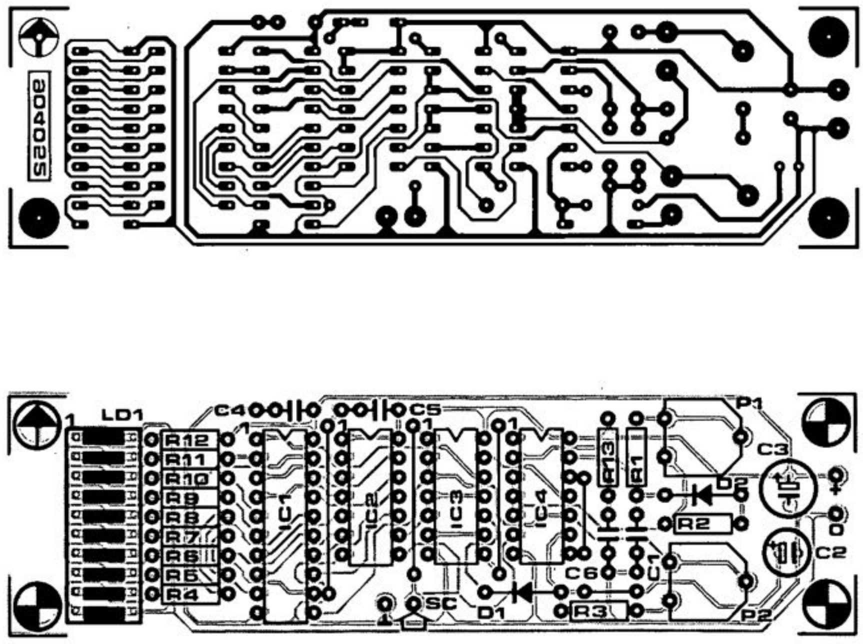
Semiconductors:
D1, D2 = 1N4148
LD1 = LED bar, e.g. super red red DC-10EWA from Kingbright
IC1 = 74HCT273
IC2 = 74HCT164
IC3 = 74HCT04
IC4 = 74HCT132



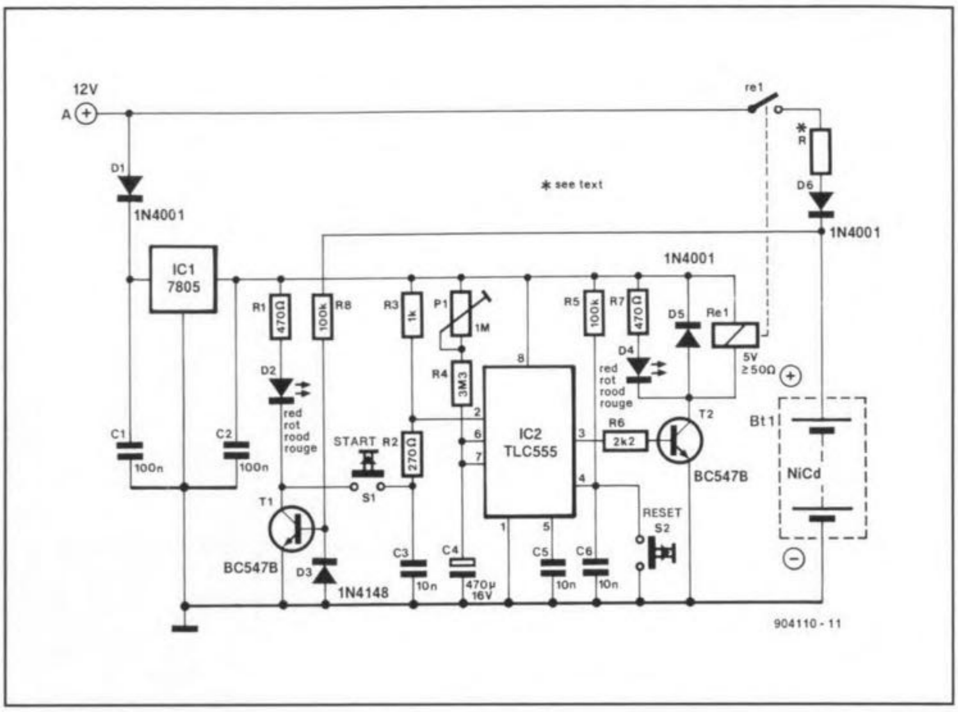
On hard disks with an ST506 compatible interface, there is a 'seek complete' ($\overline{SC}$) signal available. This signal is inactive high when the hard disk seeks new data. The duration of the high interval is thus directly proportional to the wasted search time. The length of that search time is determined primarily by the time required to shift the heads to the desired cylinder. Hard disks with an access time of 68 ms use search times of between 5 ms and 200 ms. By measuring the search time at each access and make this visible with the aid of an LED bar, an impression is gained of the performance of the hard disk. The length of the bar increases appreciably with fragmenting of the disk. If the bar is 'in the red' often, it's getting time to run a disk optimalization program!

Signal $\overline{SC}$ is connected to pin 8 of the ST506 interface. Each odd-numbered pin is connected to earth.

When the heads begin a search, that is, after $\overline{SC}$ has gone high, the clock input of buffer/register IC1



022 PORTABLE NICD CHARGER



The portable charger is intended primarily to give model enthusiast the opportunity of charging their NiCd batteries from a car battery out in the open.

The supply voltage for the circuit is regulated by IC1.

When the circuit is connected to the car battery, D2 lights only if the NiCd to be charged has been connected with correct polarity. For that purpose, the + terminal of the NiCd battery is connected to the base of T1 via R8. Since even a discharged battery provides some voltage, T1 is switched on and D2 lights.

Only if the polarity is correct will the pressing of the start switch, S1, have any effect. If so, the collector voltage of T1 is virtually zero, so that monostable IC2 is triggered by S1. The output, pin 3, of this CMOS timer then goes high, T2 is switched on and relay Re1 is energized. Charging of the NiCd battery, via R5 and D6, then begins and charging indicator D4 lights. During the charging C4 is charged slowly via P1 and R4. The value of these components determines the mono time of IC2 and thus the charging period of the NiCd battery. With values as shown in the diagram, that period may be set with P1 to between 26 and 33 minutes. Note that this time is affected by the leakage current of C4; it pays to use a good quality capacitor here. The charging may be interrupted with reset switch S2.

The charging current through the NiCd battery is determined by the value of *R*, which may be calculated as follows:

$$R = \{12-(0.7 + 1.3 \times \text{no of cells})/I_c\} \ [\Omega]$$

where I_c is the charging current, which is here, because of the chosen charging period, twice the nominal value of the capacity of the NiCd battery.

Resistor *R* must be able to dissipate a power of $I_c^2 R$ watts.

Finally, make sure that the NiCd battery is suitable for fast charging and never charge for longer than half an hour! ■

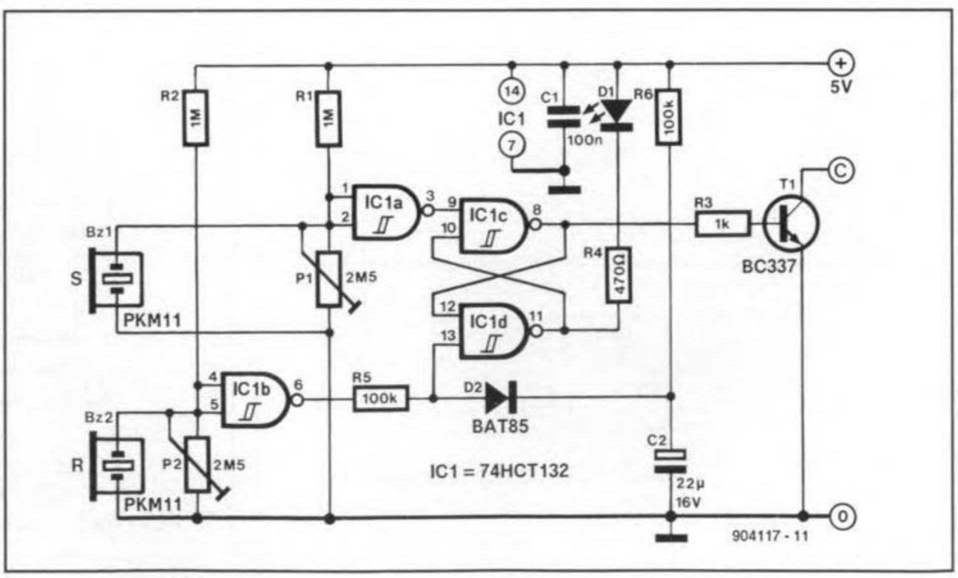
(*G. Boddington*)

023 MECHANICALLY CONTROLLED BISTABLE

Applications for this mechanically set and reset bistable are found, among others, in anti-theft devices and model railway crossings.

The transducers are formed by buzzers Bz1, which sets the bistable, and Bz2, which resets it. Their sensitivity is set with P1 and P2 respectively. The presets are adjusted correctly if the output of buffers IC1a and IC1b just toggles from high to low or vice versa.

If all has been set correctly, a slight tap on Bz will set the bistable. This causes T1 to switch on, which enables, for instance, a relay to be energized. At the same time, D1 lights. A tap on Bz2 or on its mounting resets the bistable, whereupon D1 goes out and T1 is switched off.



The bistable draws a current of about 12 mA only, the larger part of which flows through the LED.

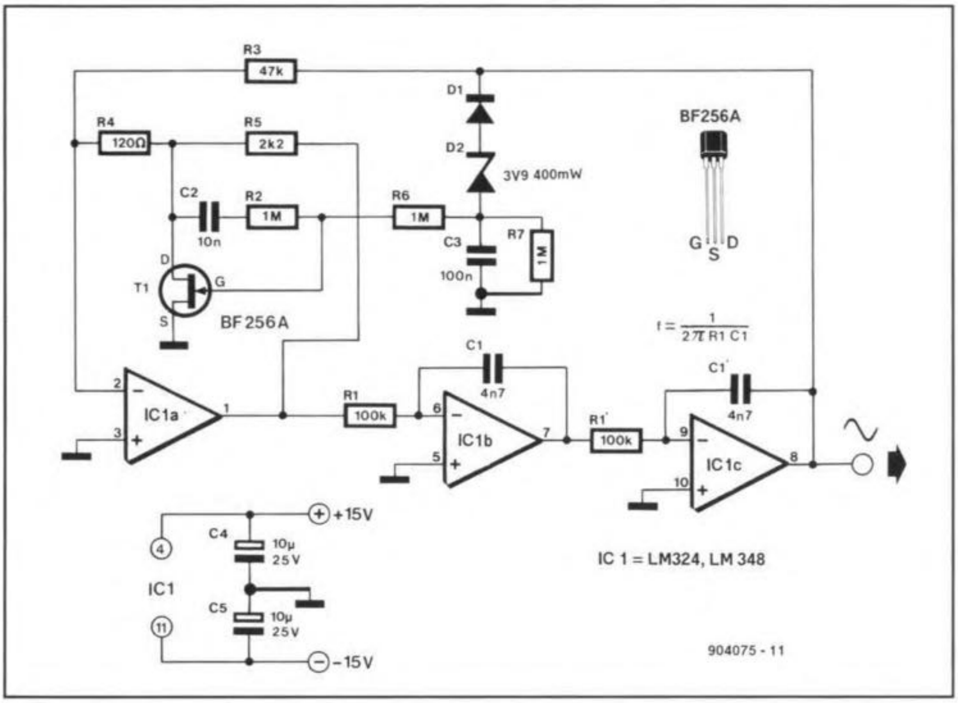
Capacitor C2 ensures that the bistable is reset when the supply is switched on: after that, the LED must thus be out. ■

(J. Ruffell)

SINE WAVE GENERATOR 024

The frequency of the generator presented here is determined by integrators IC1b and IC1c. An integrator has two properties that are used in this design. Firstly, there is a phase shift of 90°between input and output (ignoring for the moment the non-ideal behaviour of the opamp), and secondly, its amplification is –1 (i.e., inversion of signal), provided the frequency,

$f = 1/2\pi R1C1$.



Cascading two identical integrators will thus result in an overall phase shift of 180° and an amplification of unity (provided the frequency is $1/2\pi R1C1$): an ideal basis for an oscillator.

The two integrators are connected in the feedback circuit of an amplifier whose gain is determined by the amplitude of the output signal. Consequently, the generator has a reasonably stable output voltage (at a level of about 4.5 V p-p).

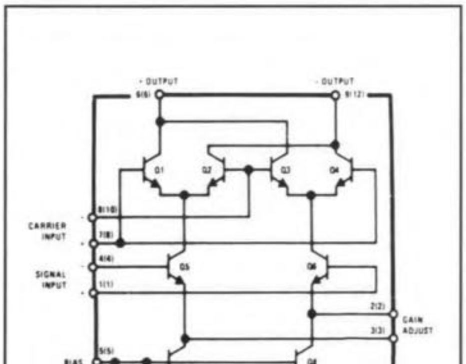
With the values of C1 (C1') and R1 (R1') as shown in the diagram, the output has a frequency of about 300 Hz. The frequency may be varied by replacing R1 and R1' by a stereo potentiometer. To keep the frequency setting within bounds, the overall range of this potentiometer should not exceed a decade.

The maximum attainable frequency is about 5 kHz. Distortion is not greater than 0.1%. The current drawn by the generator is only a few milliamperes.

Finally, the LM348 is a quadruple 741; it is thus possible to construct the generator from four 741s. ■

(G. Boddington)

FREQUENCY DOUBLER 025



Internal circuit of the LM1496.

It is often required that the frequency of a signal be doubled: modulator/demodulator chip LM1496 is an ideal basis for this.

From trigonometry it is well known that

$2\sin x\cos x = \sin 2x$

and

$\sin^2 = 1–\cos 2x$.

These equations indicate that the product of two pure sinusoidal signals of the same frequency is one signal of double that frequency. The purity of the original signals is important: composite signals would give rise to all sorts of undesired product.

The LM1496 can process only signals of not greater than 25 mV: above that serious distortion will occur. The design is therefore provided with a potential divider at its input. This makes it possible, for instance, to arrange for a 500 mV input signal to result in a signal of only 25 mV at the input of the LM1496.

To provide a sufficiently high output signal, the output of IC1 is magnified by opamp IC2, which is connected as a non-inverting amplifier. Since the output of IC1 contains a d.c. component of about 8 V, the coupling between the two stages must be via a capacitor, C4.

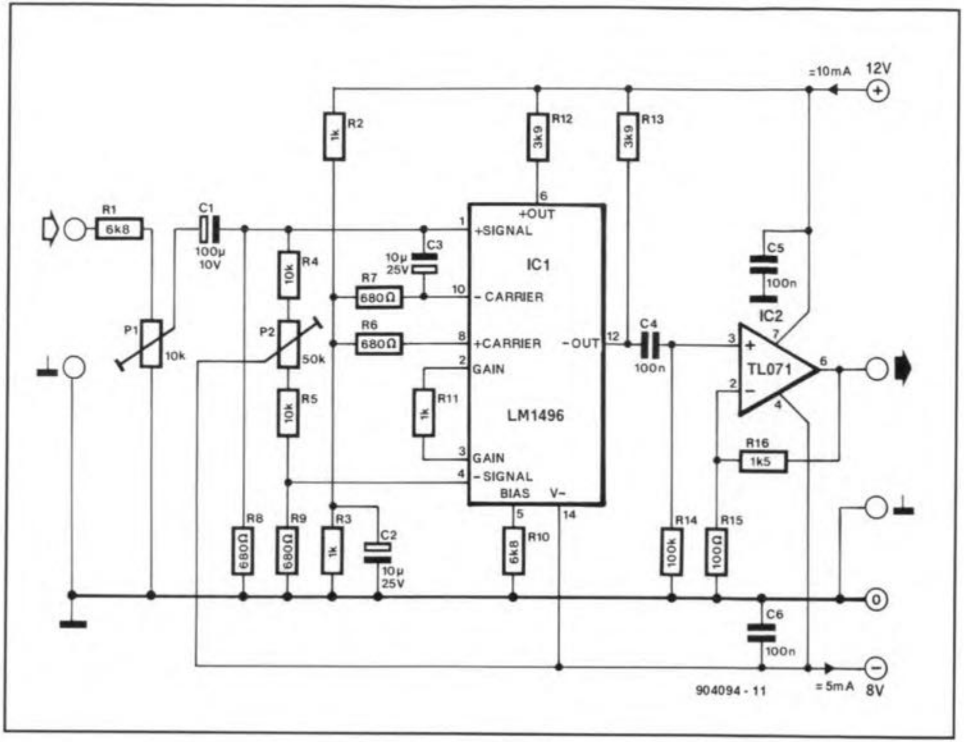
With values of R15 and R16 as shown, IC2 gives an amplification of 16 (24 dB). The overall amplification of the circuit depends on the level of the input signal: with an input of 1.2 V, the amplification is unity; when the input drops to 0.1 V, the amplification is only just 0.1.

The value of the input resistors has been fixed at 680 Ω: this value gives a reasonable compromise between the requirements for a high input impedance and a low noise level.

To ensure good suppression of the input signal at the output, it is essential that the voltages at pin 1 and pin 4 of IC1 are made absolutely identical with P4. It is possible, with the aid of a spectrum analyser, to suppress the fundamental (input) frequency by 60–70 dB.

The output signal at pin 12 is distorted easily since the IC is not really designed for this kind of operation. The distortion depends on the level of the input signal. At a frequency of 1 kHz and an input level of 100 mV, the distortion is about 0.6%; when the input level is raised to 500 mV, the distortion increases to 2.3%, and when the input level is 1 V, the distortion is 6%. The signal-to-noise ratio under these conditions varies between 60 dB and 80 dB

The circuit draws a current of 10 mA from the positive supply line and 5 mA



from the negative rail.

The phase shift between the input and output signals is about 45° (output lags).

Finally, although the normal output is taken from pin 12, there is a similar output, but shifted by 180° (with respect to that at pin 12), available at pin 6. ■

(T. Giffard)

026 SYMMETRICAL LOW-NOISE PREAMPLIFIER

The SSM-2016 differential audio preamplifier from PMI is primarily intended for amplifying signals from low-impedance sources (<1 kΩ), such as a 150-ohm microphone. If higher impedances are used, the SSM-2015 is a better choice.

The circuit diagram of the preamplifier is shown in Fig. 1, while the internal circuitry of the SSM-2016 is given in Fig. 2.

The amplification, α, of the preamplifier is determined solely by resistor R5 and is calculated from:

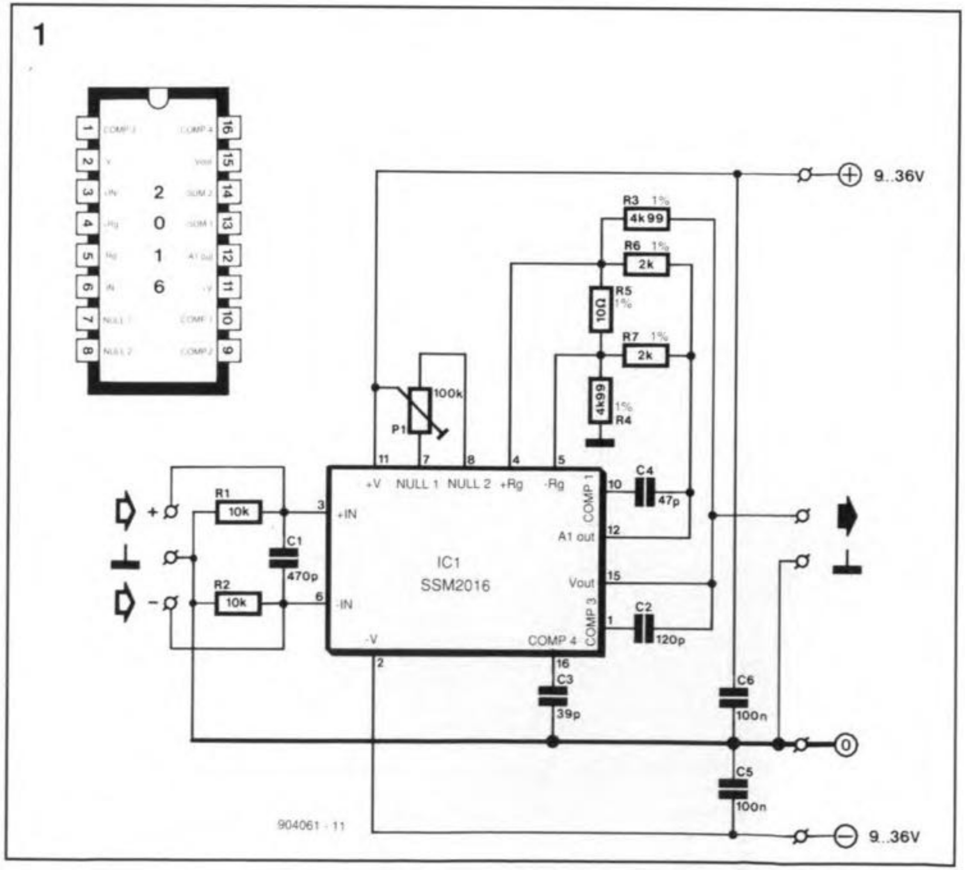
$$\alpha = (R3+R4)/R5 + (R3+R4)/(R6+R7).$$

With values as shown, this may be simplified to:

$$\alpha = 10^3/R5 + 3.5.$$

With R5 = 10 Ω, the amplification is thus 1000 (60 dB). Although the specification of the preamplifier is hardly dependent on the chosen amplification, it shouid be noted that the distortion is slightly lower at smaller amplification factors.

The external resistors have a large bear-



ring on the quality and performance of the preamplifier: class A, 1% metal film resistors are therefore essential.

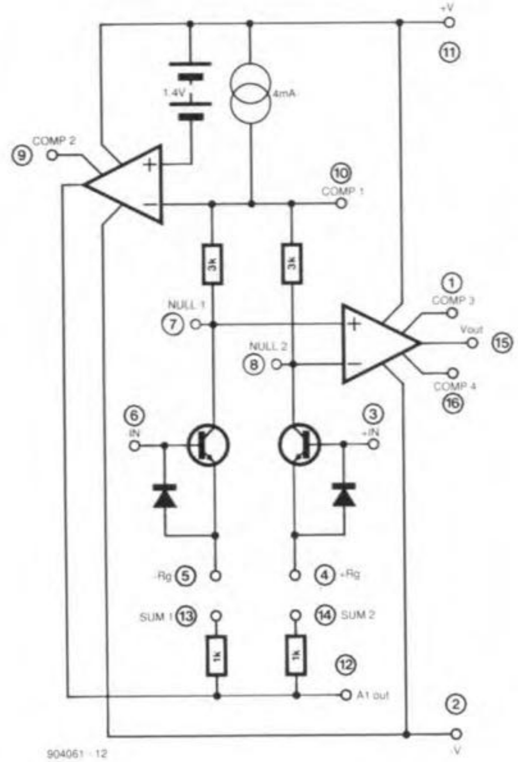
The input referred noise of the IC is very low: 800 pV/√Hz. In view of the common-mode noise, the values of resistors R1 and R2, which determine the bias current, must be chosen with care: they should not exceed 10 kΩ.

Capacitors C2, C3 and C4 are compensating components. Moreover, the value of C2 has a decided effect on the the bandwidth of the amplifier: when it is 120 pF as shown, the bandwidth is around 450 kHz (if the amplification factor is less than 100, the bandwidth may even be as large as 1 MHz).

Since the bandwidth is determined mainly by C2 and the feedback resistor, it is virtually independent of the amplification: with amplification factors between 3.5 and 1000, it varies from 1 MHz to 450 kHz.

Capacitor C1 provides additional decoupling of the inputs and should therefore be mounted as close to the input pins of the IC as possible.

2



The SSM2016 is capable of fairly high output currents (min. 40 mA), so that with a supply voltage of ±18 V, an undistorted signal of 10 V r.m.s. is available across a load of 600 Ω. With higher supply voltages, care should be taken that the maximum dissipation of the IC does not exceed 1.5 W.

The prototype had an harmonic distortion of not more than 0.006% (up to 10 kHz) with a load of 10 kΩ and an output voltage of 1 V r.m.s. When the load was reduced, this figure increased to 0.02% at 1 kHz and 0.035% at 10 kHz.

The slew rate was 10 V/µs. The signal-to-noise ratio at an amplification of 1000 and an output voltage of 1 V was 98 dB with the inputs short-circuited and 88 dB with a source impedance of 600 Ω.

The common-mode rejection ratio (CMRR) is high over the whole audio range: 114 dB at 1 kHz and 108 dB at 20 kHz. This means very effective suppression of hum at the input.

The complete preamplifier draws a current of 12–15 mA.

The offset voltage at the input may be compensated with P1. Because of the high input bias current of the opamp (up to 25 µA max), an extra offset may arise at the input with pseudo-differential or asymmetric use of the inputs that can not or hardly be compensated with P1. The result of this is higher distortion.

Although the power supply suppression is about 100 dB, it is recommended to decouple the supply lines well. ■

(T. Giffard)

LOW CONSUMPTION MONOSTABLE RELAY

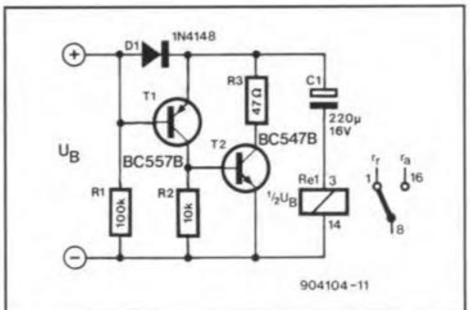
A monostable relay has two states: operative when a large enough current flows through its coil and quiescent when no current flows. A relay contact that assumes a certain position after the supply voltage has been switched on is required in many applications, and, of course, many relays operate in that manner.

However, most of these relays require an energizing current of 50 mA or more and that normally precludes a battery supply. The circuit presented here, which uses a bistable relay, may solve that problem.

The contact of a bistable relay normally remains in the position it is in after the supply is switched off. The present circuit, however, makes the bistable relay behave like a monostable type, and that at a very modest current.

When the supply voltage is switched on, C1 charges via D1 and the relay coil. The current then flowing through the coil causes the relay contact to assume one of two positions. The forward drop across D1 ensures that the base of T1 in this condition is more positive than its emitter, so that T1, and thus T2, is switched off.

When the supply voltage is switched off, the emitter of T1 is connected to the positive terminal of C1, while the base is connected to the negative terminal of the capacitor via R1 and the relay coil. This results in T1, and thus T2, switching on, so that C1 discharges via T4 and the relay. The current rhough the relay coil then flows in an opposite direction and this causes the contact to change over.



The bistable relay thus behaves exactly as a monostable type with the advantage, however, that the operational current is determined by R1, and here amounts to only 130 μA.

To ensure reliable operation, the rating of the relay coil should be 65–75 per cent of the supply voltage. In the prototype, a 9-V relay was used with a battery supply voltage of 12 V. ■

(*F. Hueber*)

028 WINDSCREEN WIPER INTERVAL

The windscreen wiper interval circuit presented here is very compact and is noteworthy for its use of two thyristors instead of a relay. It has only two connections and operates without any problems even in conjunction with multi-stage wiper circuits.

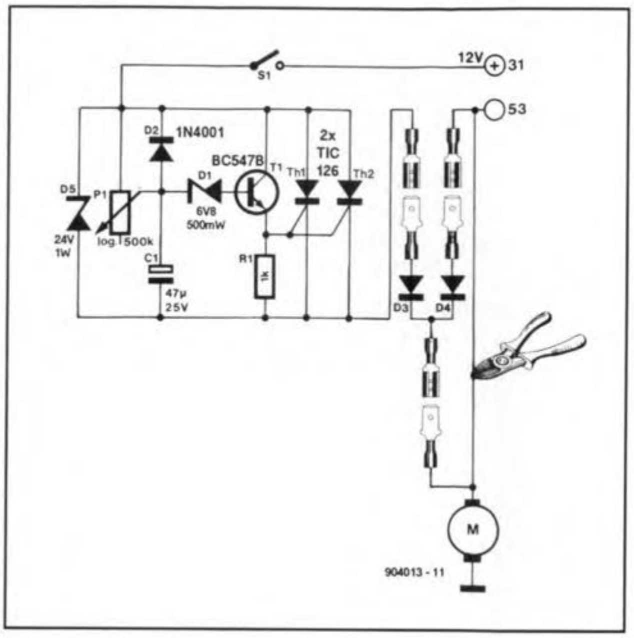
The connecting wire between the wiper motor and terminal 53 is cut and new connections are made as shown in the diagram.

When the interval switch, S1, is closed, capacitor C1 charges via P1 and the wiper motor. After a time set with P1, transistor T1 switches on and triggers the thyristors. The wiper motor is then energized via the thyristors and D3 and sets the wipers into motion. At the same time, C1 discharges via D2 and the thyristors.

After a short time, the wiper stop switch connects terminal 53 to the +12 V line, so that the wiper motor is energized via D4. The thyristors are switched off because the voltage drop across D3 plus Th1-Th2 is then greater than that across D4.

When the wipers reach the end of their travel again, the stop switch connects terminal 53 to ground and this enables C1 to charge again.

(E. Tienken)



029 POWER DRIVER FOR OPERATIONAL AMPLIFIER

It happens frequently that the output current of an operational amplifier is inadequate for the application as, for instance, whent a small motor or loudspeaker has to be driven. Normally, this is resolved by adding an emitter follower to the circuit as shown in Fig. 1. Unfortunately, that circuit does not allow the full supply voltage, *Ub*, to be used, because the output voltage of the opamp must always be 1–2 V smaller than ±*Ub*. To that must be added the drop across the base-emitter junction of transistors T1 and T2.

The circuit shown in Fig. 2 (principle) and Fig. 3 (practical) is a more appropriate solution: it was designed specifically for driving small motors. Since the output current of the opamp flows through its supply lines, the driver transistors may also be controlled over these lines.

The value of base-emitter resistors R4 and R5 has been chosen to ensure that in spite of the quiescent current through the opamp, T1 and T2 are switched off.

Resistor R6 limits the output current of the opamp. If the opamp is a type with guaranteed short-circuit protection, R6 may be replaced by a jump lead.

The output voltage is only 50–100 mV (collector-emitter saturation voltage of the driver transistors) smaller than the supply voltage. When choosing these transistors, it is therefore essential to take into account the saturation voltage in addition to the maximum current amplification and power rating.

The value of the resistors in an inverting circuit are calculated from:

$\alpha = R2/R1$

and

$R3 \approx R2/R1$,

where α is the amplification.

In a non-inverting circuit (R1 between

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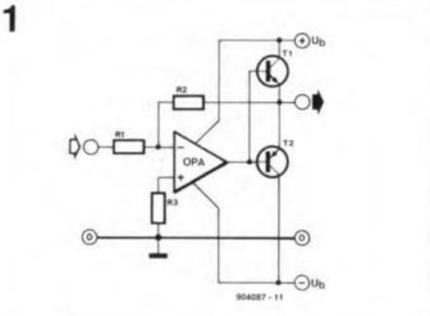

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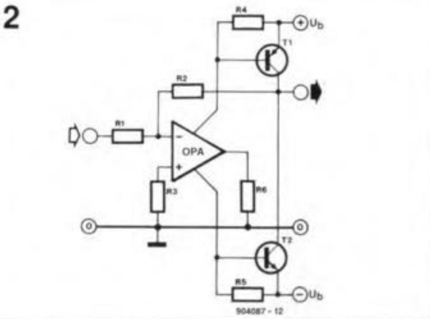

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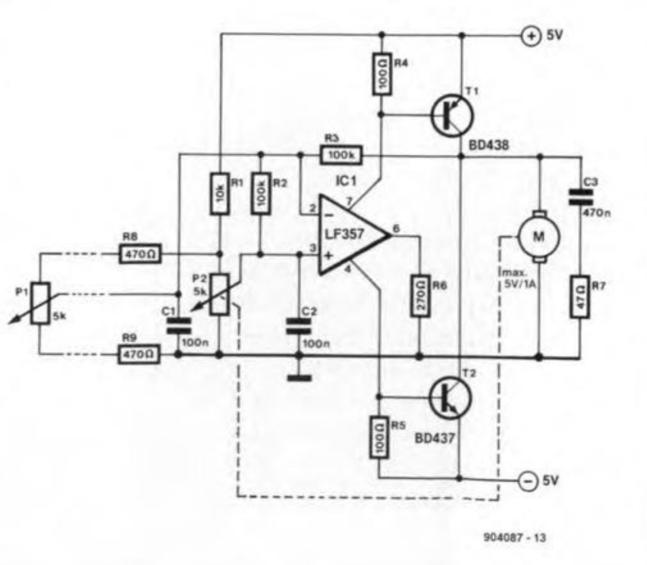


the – input and earth and the input signal connected to the + input of the opamp), the amplification is

$\alpha = (R2/R1) + 1$

and

$R3 << Re$,

$R4 < +\alpha / +Ub$ $R5 < -0.5\alpha / -Ub$

$R6 \approx Ub / Imax$

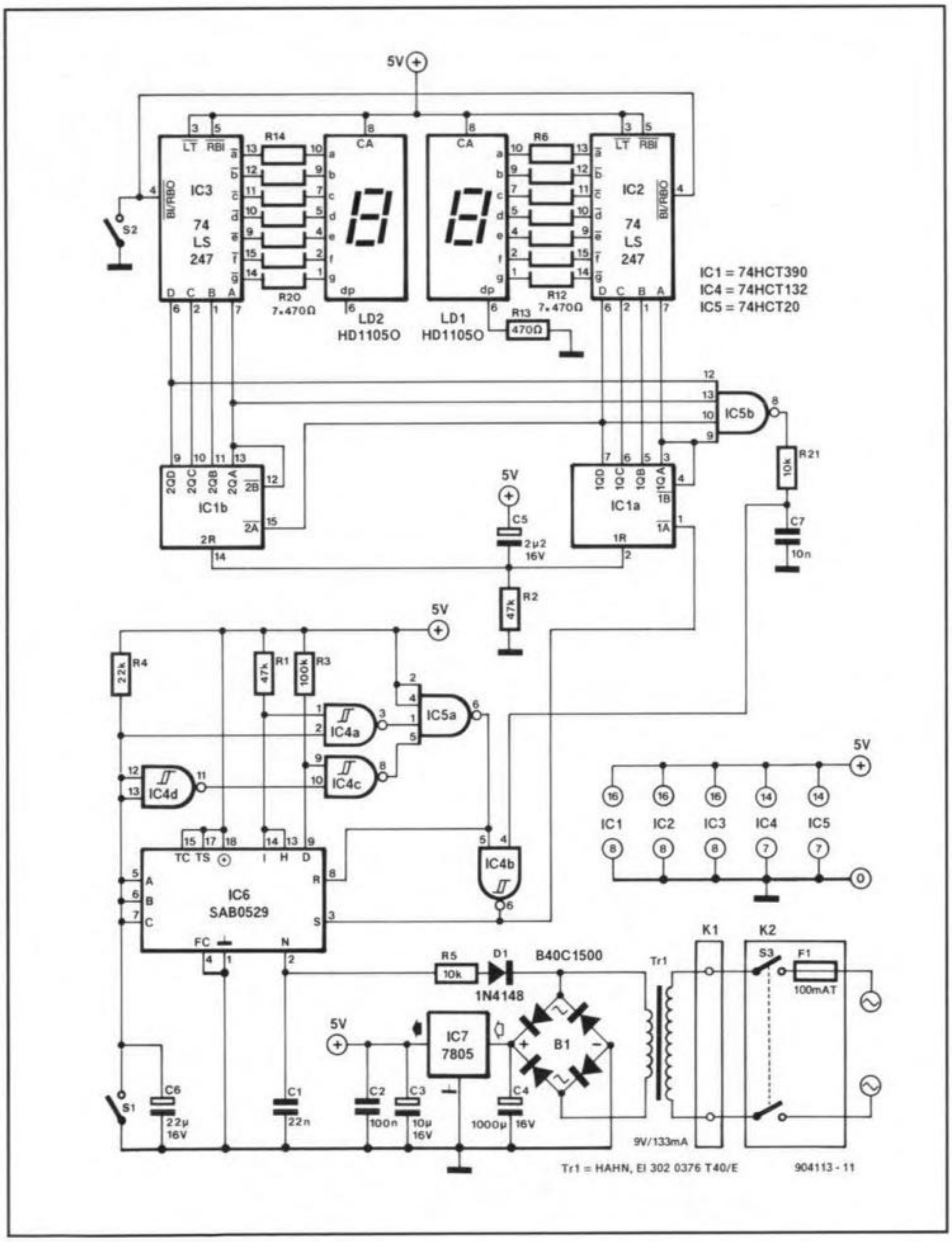
where Re is the input impedance of the opamps.

The circuit can be used with discrete (single) opamps only, because double or quadruple types in one package share the supply voltage pins.

The setting accuracy of the circuit in Fig. 3 is better than 1%. ■

(*G. Pelz*)

COUNT THE DAYS ... 030



When it is necessary, or desired, to count the number of days to a particular event or date, the circuit here, which can count up to 99, may be found a useful aid to keep a check on how many days have passed.

The circuit is based on a Type SAB0529, which counts on a 24-hour day basis. The chip is reset automatically when the supply is switched on: an external *RC* network is thus not necessary.

When switch S1 is open, open-collector output pins I and H of IC6 go logic high 24 hours after the supply is switched on. At that moment, the 24-hour timer is set back by 1 via NAND gate IC5a and the next 24-hour cycle begins.

At the instant the timer is set back, the pulse from IC5a is inverted and applied to the clock input (pin 1) of IC1. Only when counter position 99 has been reached will the output of IC5b go low and thus disable th clock input of IC1 via IC4b. The counter thus stops after 99 days and remains in that position until the circuit is reset.

The circuit may be tested by closing S1 which converts the time base to a seconds clock. In other words, the counter then reaches position 99 after 99 seconds.

With the display on and S1 open, the circuit draws a current of about 100 mA; in the economy position (S1 closed) this reduces to about 10 mA. ■

(*M. Ruitters-Franssen*)

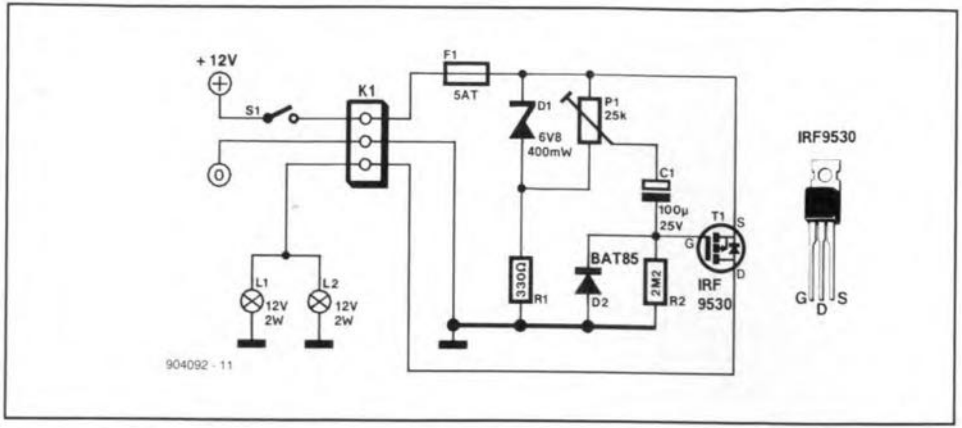
REAR FOG LIGHT DELAY 031

Although we assume that most of our readers are thoughtful drivers who do not switch on their rear fog lights when closely followed by other traffic, since following drivers for an instant think you are braking (although they have seen no reason for your doing so) and thus slam on their brakes as well. This may often give rise to a very dangerous situation, it is better to

S 28

avoid such a potentially dangerous action and install the rear fog light delay circuit presented here.

Switch S1 is the on-off control for rear fog lights L1 and L2. As soon as this switch is closed, the gate-source voltage (*Ugs*) of MOSFET T1 will become more and more negative. This means that the IC will conduct harder and harder, and this in turn reasults in the brightness of the lights gradually becoming brighter. Maximum brightness is reached after a delay of about 20 seconds, which is determined by time constant R2-C1.



The gate of T1 may be given a bias by preset P1. This provides compensation for the initial period after the lights are switched on and the lamps do not light, because they need some hundreds of milliamperes before they can do so. With P1 set correctly, the lamps will light, albeit weakly, immediately the control switch is closed. The gate potential is then equal to the voltage at the wiper of P1 (bear in mind that C1 is then still discharged).

Although the dissipation of T1 is a maximum during the transitional period (between switch on and the lamps lighting brightly), the heat sink required is calculated on the basis of the dissipation when the lamps light brightly. Normally, rear fog lights are rated at 21 W, so that if two of them are fitted, a heat sink Type SK59 (i.e., 36.5×42.7×12.5 mm) provides ample cooling. This type of heat sink is available from DAU (UK) Ltd, 70–75 Barnham Road, BARNHAM PO22 0ES. ■

(*J. Ruffell*)

032 CURRENT SOURCE FOR PORTABLE BATTERY CHARGER

The "portable battery charger" described earlier in this supplement may be extended by a current source that ensures a constant current through the batteries to be charged at all times. The source may, of course, also be added to other NiCd battery chargers not yet so equipped.

Transistors T1 and T2 and resistor R3 form a darlington that obtains a constant base voltage via D3. There is thus also a constant voltage across resistance *R* in the emitter circuit of the darlington, which means that the value of *R* determines the charging current.

Resistor R1 provides the current for voltage reference D3. The LED in series with R1 indicates whether the batteries have been connected properly.

If the current source is used with the portable charger, D2 may be omitted, because that charger already provides a polarity check.

The value of *R* may be calculated from:

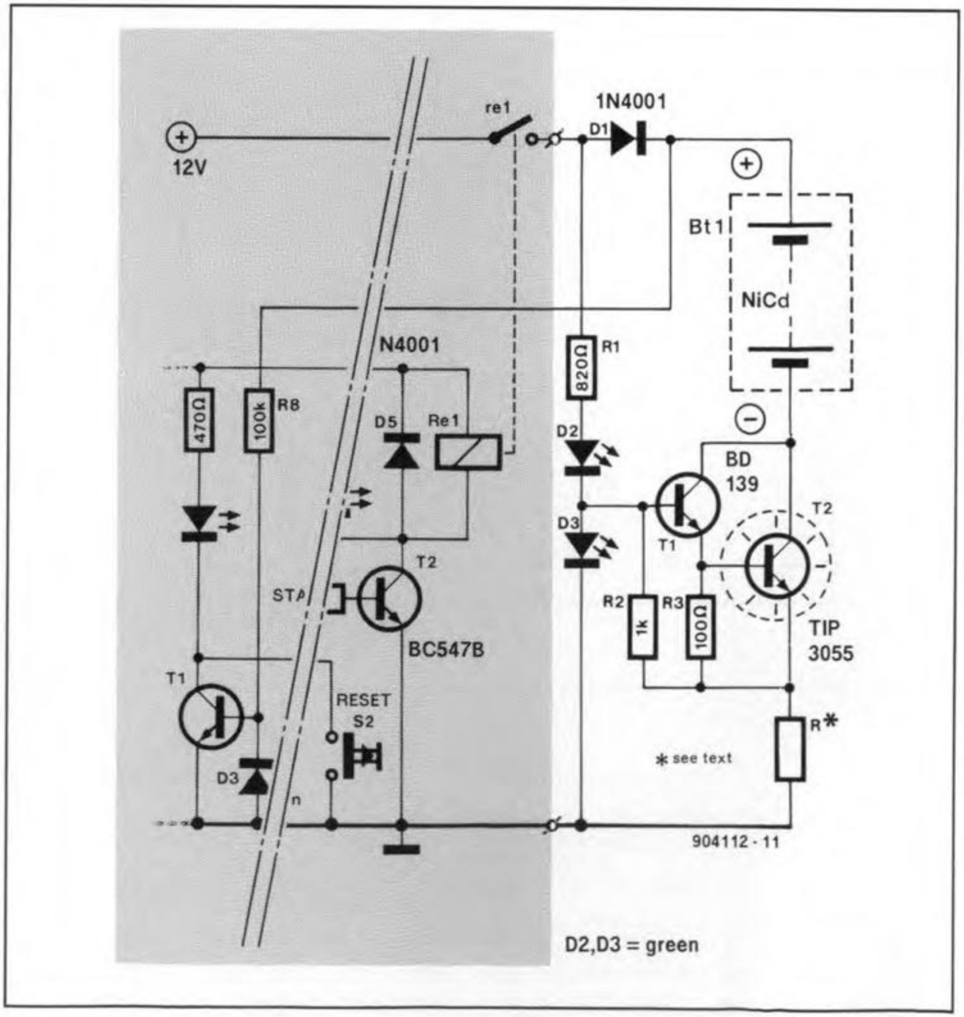
$R = 0.7/\text{charging current}$.

Again, account must be taken of the dissipation of *R*, which is

$P = I_c^2 R$.

Transistor T2 must be fitted on a heat sink, whose size depends on the number of series-connected NiCd batteries and the current flowing through them. ■

(*D. Oberye*)



PREAMPLIFIER FOR OSCILLOSCOPES

033

In many oscilloscopes, the most sensitive range is 2–5 mV, although it is often possible to improve this to 1–2 mV by a variable gain control. To obtain even better sensitivity, the present preamplifier, which has an amplification of about 10 (20 dB), may be found useful.

Because most oscilloscopes have a bandwidth of 20 MHz or more, the amplifier must, of course, have a slightly wider bandwidth and that is achieved with a Type OP260 opamp. This has a slew rate of 550 V/μs (at an amplification of 10) and a bandwidth of 40 MHz that is virtually independent of the amplification. The gain vs frequency response is not so good, however: as may be seen from Fig. 2, where the characteristics are given for a number of loads. The hump in the curves depends on the value of the feedback resistor, whose optimum value appears to be 2.5 kΩ.

The curves in Fig. 3 accord with different values of R2/R8 for an amplification factor of 10. Some experimentation with the value of R2/R8 for different amplification factors may be instructive. Bear in mind, however, that the output impedance increases from 20 Ω to 225 Ω over the frequency range of 10 MHz to 60–70 MHz. It is therefore important to keep all connections on the prototyping board as short as possible and to connect all earth points to a common ground via a separate, heavy track. Also, do not use an IC socket.

An input impedance of 1 MΩ was chosen, which results in a fairly high level of noise at the output (with open-circuit input). This value may be reduced, since otherwise the use of a 1:10 probe will be inhibited, because that would give constant problems with the noise. However, when the amplifier is connected to a suitable source, the noise reduction is normally more than ample to obtain a good trace on the screen.

Presets P1 and P2 serve to provide compensation for the d.c. offset and input offset caused by R1 and R7 respectively.

The input bias current for the non-inverting input is about 10 times lower than that for the inverting input, which makes the OP260 more suitable for non-inverting circuits. The inverting circuit may also give problems because of the low values of R2 (R8) and R3 (R9).

The input bias current is typically 0.2 μA, while the input offset is about 3 mV (max. 7 mV).

In this type of circuit it is important to use a well-regulated power supply. The power supply suppression up to 10 kHz is roughly 70 dB, and this reduces with increasing frequency. Any noise or tiny ripple on the supply lines would make the application of the circuit as a small signal amplifier impossible.

The circuit draws a current of about 14 mA. The slew rate, as with most opamps, is asymmetric and may lead to visible distortion of the signal when the drive to the 560 Ω resistor is high at the higher frequencies. ■

(T. Giffard)

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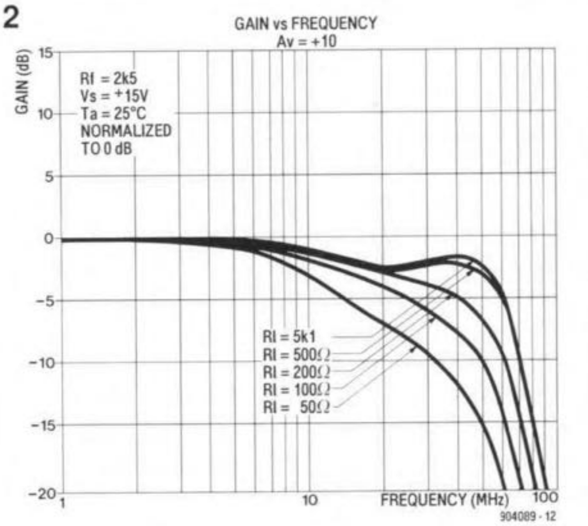

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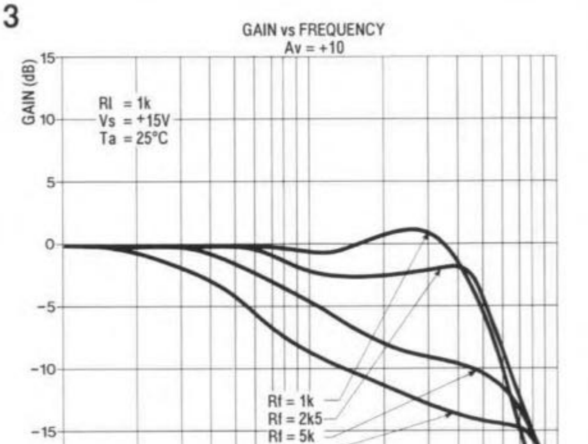

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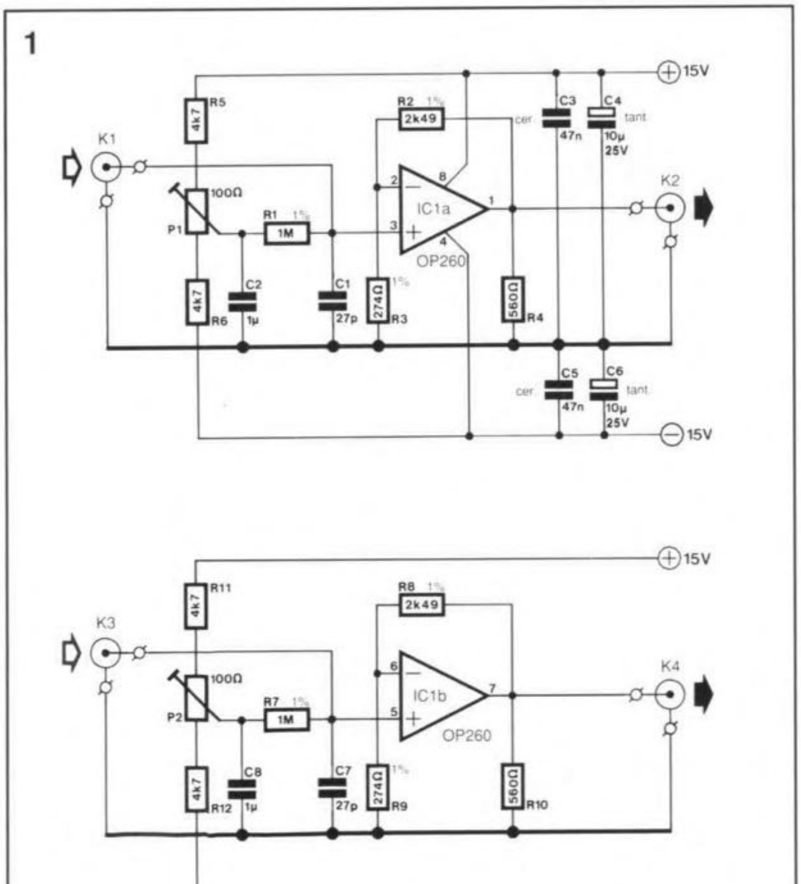

PREAMPLIFIER FOR OSCILLOSCOPES

033

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(T. Giffard)

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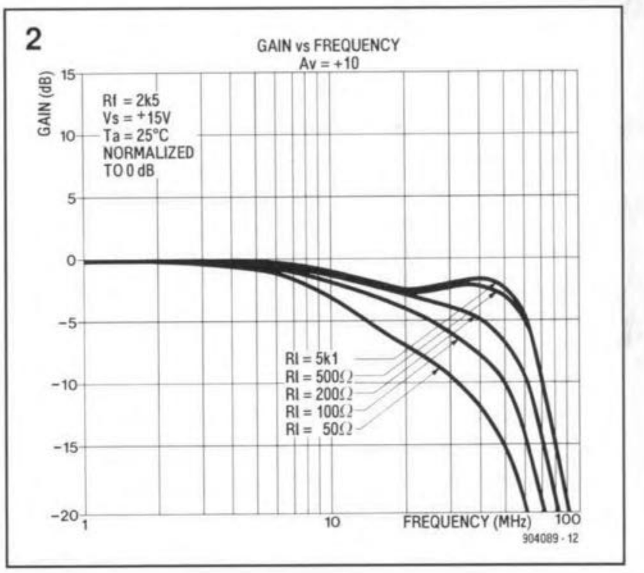

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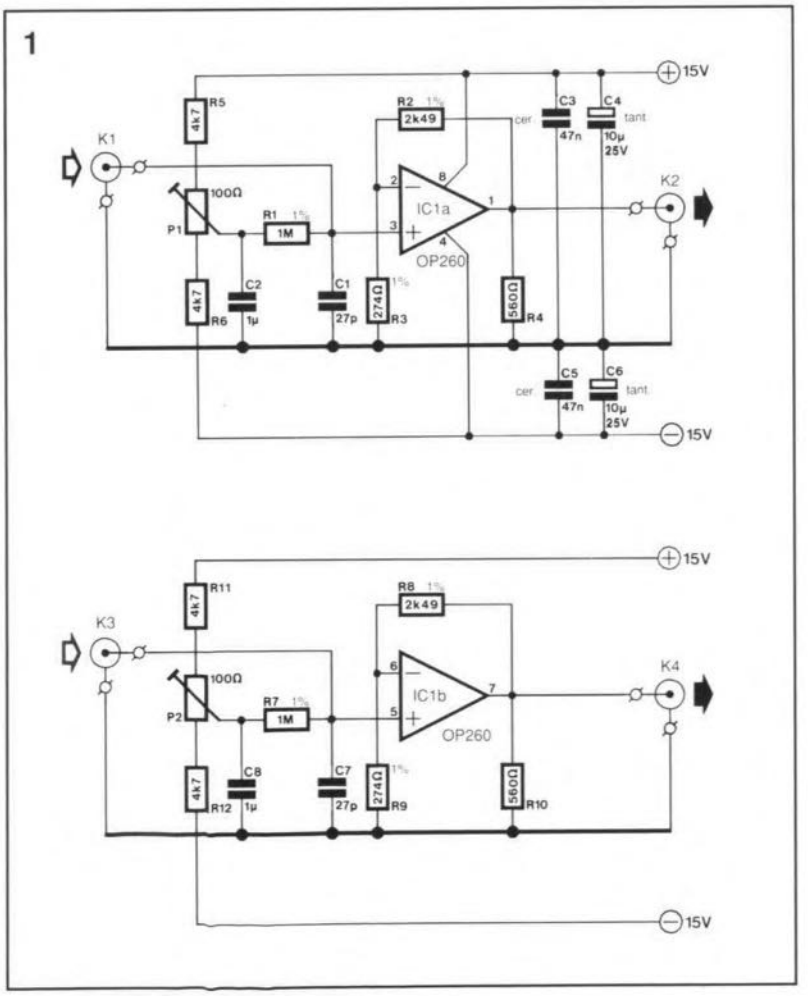

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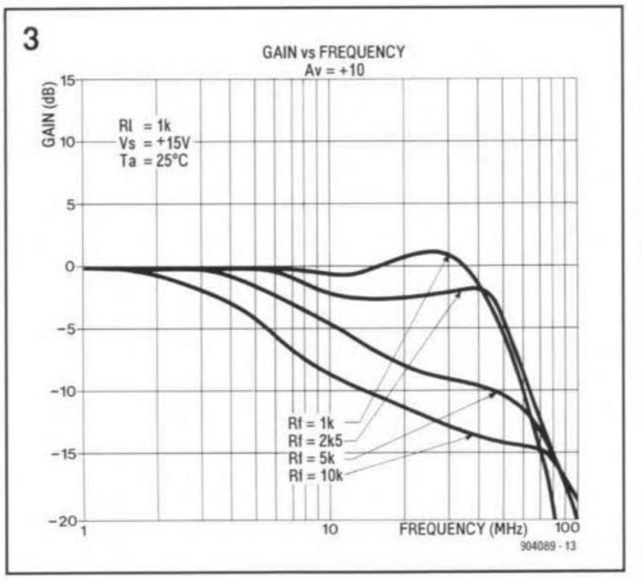


034 RUMBLE FILTER

Although sales of gramophone record may have slumped, there are millions of people who still treasure their record collection. Many record players unfortunately exhibit two undesired side effects: rumble (noise caused by the motor and the turntable) and other low-frequency spurious signals. The active high-pass Chebyshev filter presented here was designed to suppress those noises. The filter has a 0.1 dB ripple characteristic and a cut-off point of 18 Hz. A note for designers: a passive filter with the same characteristics was tried: a sixth order Bessel filter, which was soon rejected when it was found that inductors of 600 H were needed!

The circuit itself is not too exciting; it is the selection of components that makes a filter successful. The choice of a Chebyshev filter may not seem too clever for audio purposes, but because of its 0.1 dB ripple in the pass band it behaves very much like a Butterworth type with the advantage that the response has steeper skirts as shown in Fig. 2 (which is a calculated curve). Frequencies below 10 Hz are attenuated by more than 35 dB. The phase behaviour in the pass band shows a gradual shift, so that its effect on the reproduced sound is inaudible.

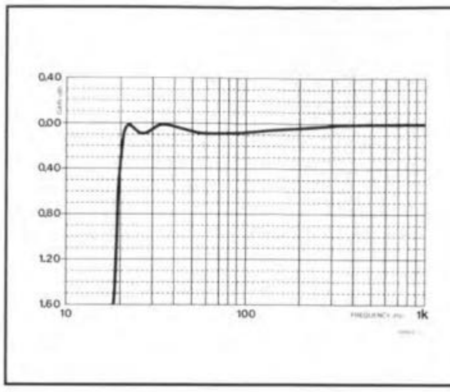
If the filter is used in a stereo installation, it is essential that the characteristics of both filters are identical or nearly so. Phase differences between channels can be heard—perhaps not so much at lower frequencies, but certainly in the mid ranges. To ensure identity and also to obtain the desired characteristics, capacitors C1–C5 must be selected carefully. It does not matter much whether their value is 467 pF or 473 pF: that only causes a slight shift of the cut-off point. What does matter is that they are identical within that 1% tolerance. For symmetry of channels, the capacitors may be paired and then used in either channel at the corresponding position.

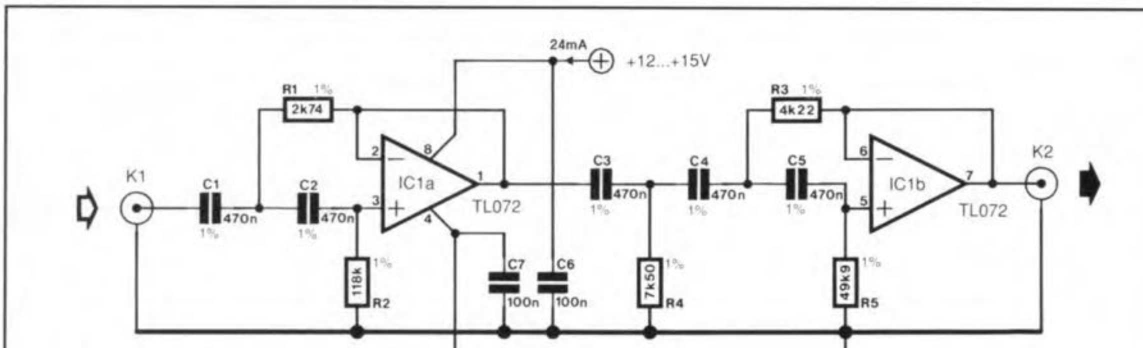
The diagram shows theoretical values for the resistors: their practical values are given in the table. The prototype was constructed with 5% metal film types from the E12 series and these were used without sorting. Their tolerance proved to be perfectly acceptable in practice.

The current drawn by the circuit is purely that through the opamp and amounts to about 4 mA. The high cut-off point is also determined by the opamp and lies at about 3 MHz.

The only problem that cannot be foreseen is a possible coupling capacitor in the signal source. That component will be in series with C1 and this may adversely affect the frequency response. However, if its value is greater than 47 μF, it will have little if any effect; if it is below that value, it is best removed; C1 will assume its function. ■

(T. Giffard)



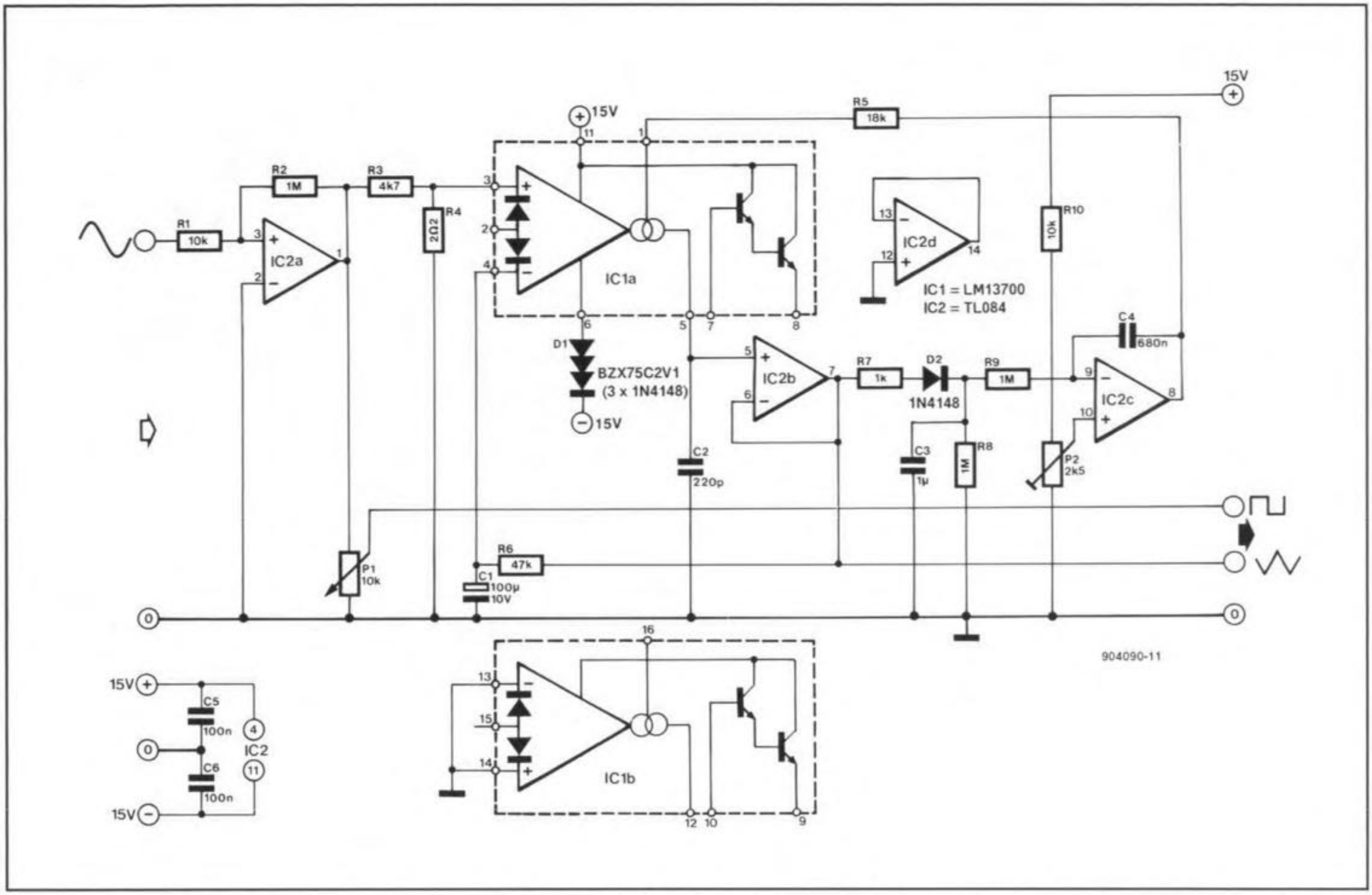


E12	E96	EXACT
R1 = 2k7 + 68Ω	2k74	2k7626
R2 = 120k	118k	119k05
R3 = 3k9 + 330Ω	4k22	4k2314
R4 = 6k8 + 680Ω	7k50	7k4648
R5 = 47k + 2k2	49k9	49k475

035 RECTANGULAR/TRIANGULAR WAVEFORM CONVERTER

Many function generators are based on a rectangular waveform generator consisting of a Schmitt trigger and integrator. The triangular signal produced by the integrator is then used to form a sinuoidal signal with the aid of a diode network. The converter presented here works the other way round. It converts the output of a good-quality sine wave oscillator into a rectangular and a triangular signal.

The sinusoidal signal is converted into a rectangular signal by IC2a. Since the output of this gate varies between –15 V and +15 V, it is reduced to a value suitable for integration by potential divider R3-R4. It is then integrated by transconductance amplifier IC1a and C2. The amplifier has a current output that is controlled by the current through pin 1. The output therefore behaves as a resistance with which it is possible to influence the integration time. The voltage across C2 is available in buffered form at the output of impedance inverter IC2b: this is the triangular signal. The amplitude of this signal is compared with a voltage set by P2 and the difference between these voltages, which is the out-



put of IC2c, is applied to the current source at the output of IC1 via R5. This arrangement ensures that the level of the output voltage is virtually independent of the frequency of the rectangular signal or the sinusoidal input.

One problem with a precision integrator is its being affected by offset voltages and bias currents. Feedback loop R6-C1 ensures that the output follows the potential across R4 accurately, although tiny deviations may be caused by the bias current in circuit IC1, which is not greater than 8 µA at 70° C.

The time constant R6-C1 is large for a purpose: to ensure that the triangular signal, even at low frequencies, can not affect the waveform of the signal to be integrated—the rectangular shape must be retained.

The converter can process signals at frequencies from 6 Hz—where the amplitude is not affected—to 60 kHz—where the amplitude is reduced by 10%.

Because of the long time constants, the time taken for the recovery of the amplitude of the triangular signal at frequencies above 1 kHz is rather long. The peak value of this signal should be set to 1 V.

Diode D1 is a so-called stabistor—three diodes in one package. It may be replaced by three discrete Type 1N4148 diodes.

The current drawn by the converter is of the order of 9 mA. ■

(T. Giffard)

The electronic antenna selector is intended to switch between two FM antennas by means of a logic signal.

Gates IC1 and IC1b ensure a clean switching action and at the same time form the interface between the 5 V logic level (probably available from the receiver) and the 12 V supply voltage for the selector. Depending on the type of gate used, a digital TTL or CMOS control signal is available in direct and inverted form at the outputs of IC1.

When input A is logic high, the output of IC1a is low and that of IC1b is high. Current then flows from the positive supply line to IC1a via T2, R9 and D8; T2 is switched on and D9 lights.

Because direct currents flow through R1-D1-R2 and R5-D3-R4, diodes D1 and D3 conduct and pass the VHF signal from input A to output D. At the same time, a direct current flows through R6-D4 so that D4 conducts. This arrangement ensures that any VHF signal at input C can not reach the output via the parasitic capacitances of the relay contacts and the wiring.

When A is logic low, and IC1b is therefore low, current flows from the positive supply line to IC1b via T1, R7 and D17; T1 is then switched on and D10 lights. At the same time, the two series-connected relays, Re1 and Re2, are energized, their contacts close and the VHF signal at input C is fed to output D. Moreover, a direct current flows through R3-D2 so that D2 conducts. Any signal at input B is then shorted to ground via D2.

All resistors should be carbon film types, because these have a higher parasitic series inductance than metal film resistors, so that the attenuation of the VHF signal caused by them is reduced to a minimum.

The attenuation losses caused by the diode junctions (5–10 dB) are somewhat larger than those caused by the relays. It is thus advisable to connect the antenna that provides the weaker signal (normally the domestic one) to input C.

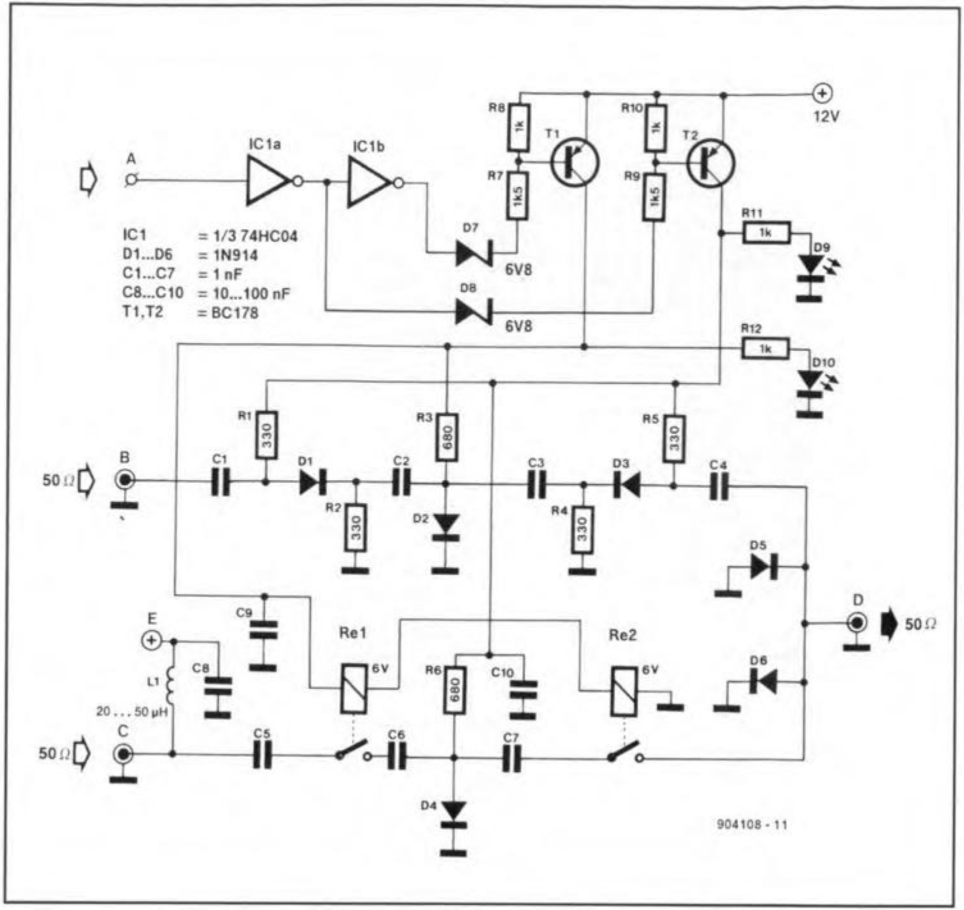
If the domestic antenna is equipped with an antenna amplifier, it may be supplied via terminal E.

Diodes D5 and D6 protect the circuit against high voltage spikes that occur during the on and off switching.

The selector draws a current of around 65 mA. ■

(T. Shaerer)

A = control input – '1' = central antenna system
'0' = domestic antenna
B = central antenna (cable) input
C = domestic antenna input
D = output to receiver
E = supply output to antenna amplifier



037 VOLUME INDICATOR FOR ALL-SOLID-STATE PREAMPLIFIER

The indicator is intended for use with the all-solid-state preamplifier we published some time ago (see reference), but may also be used in other applications where a number of steps or changes must be counted rapidly.

To prevent interference with the audio signal, the circuit is a static design. This means that if the volume control is not adjusted, the circuit does nothing.

The circuit does not need an external clock signal, since this is derived from any changes in the least significant bit—LSB. This is done by two differentiating networks: R9-C1 and R10-C2, which double the frequency of an available LSB signal.

Moreover, to ensure that the counters of the indicator remain in step with the volume control, signals 'up/down' and 'preset' from the preamplifier are used. It may seem rather extravagant to couple the state of the counters in the preamplifier with that of the present counters, but it is a good way of keeping the connections between the two units to a minimum. Furthermore, the present counters operate in 8-bit BCD instead of 6-bit binary as used by those in the volume control (in the preamplifier). All that is required to display the state of the volume control are a couple of BCD-to-seven-segment decoders and seven-segment displays.

The preset in the indicator must be set in BCD code (whereas that of the control in the preamplifier is set in binary code). It is, of course, possible to give the preset in the indicator the same value as that in the preamplifier control to give a display that varies from 00 to 63. It is, however, perhaps rather more realistic to have a display from 01 to 64, because the minimum attenuation is 78.75 dB , not infinity. There is no suppression of leading zeros, so that numbers up to and including 9 are displayed starting with a 0.

The DIP switches and resistors R1–R8 in the diagram may be omitted if only one fixed preset is likely to be used. The resistors should be replaced by jump leads.

The balance control of the preamplifier may also be indicated, but the present circuit should then be duplicated, with the exception of IC5, which has two gates to spare. The LSB connection of one indicator is coupled to IC23 in the volume control stages, while the other indicator is linked to IC25. The current drawn is, of course, doubled to around 220 mA. This makes it necessary to increase the rating of the mains fuses and to change the inscription on the relevant label from 100 mAT to 200 mAT.

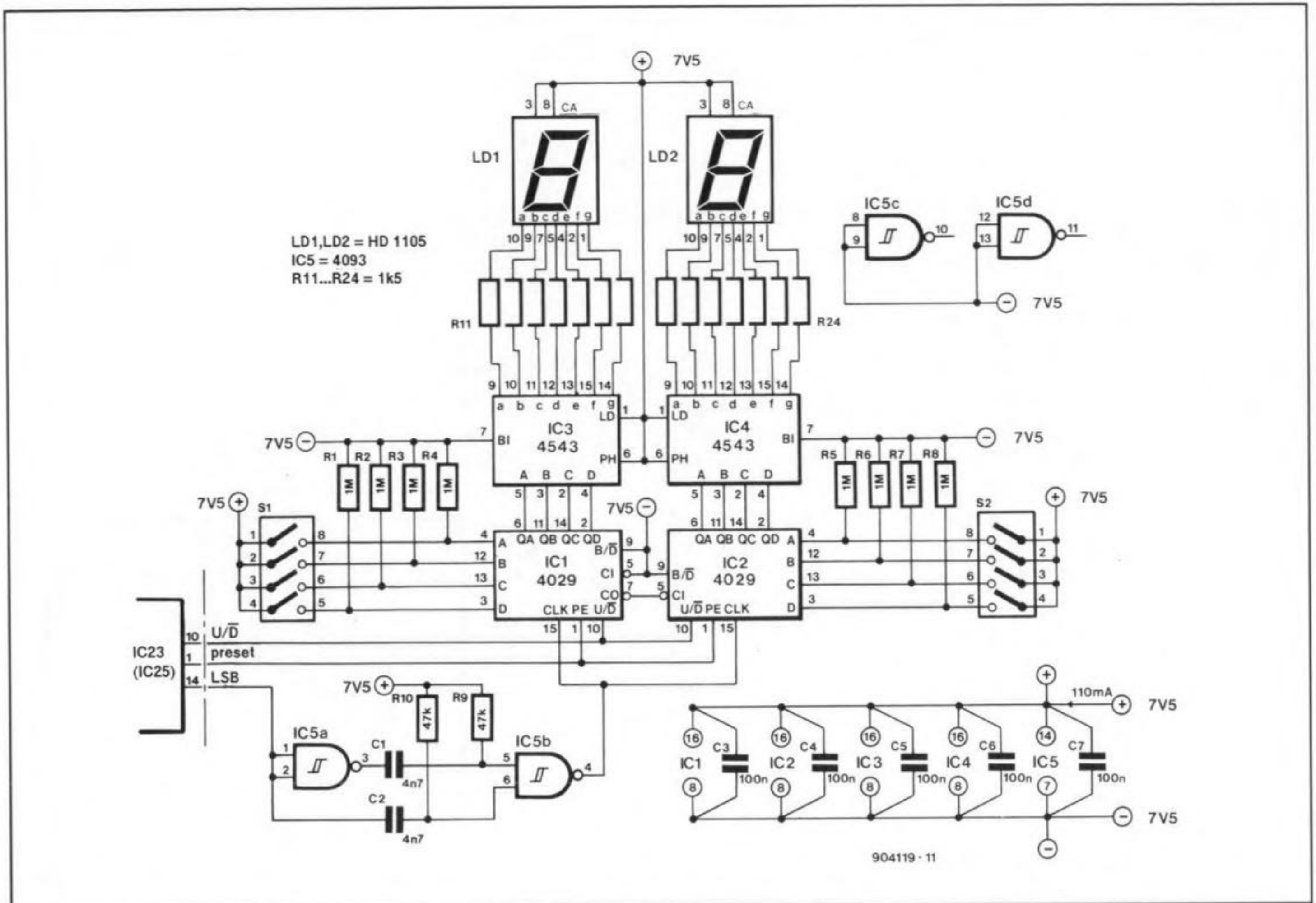
The supply voltage may be taken from the preamplifier, but careful account should be taken of the cooling of the voltage regulators, particularly if two indicators are used. It may be necessary to improve that cooling.

The interference suppression of the regulators, IC33 and IC34 in the preamplifier, may also be improved by additional 10 μF, 10 V electrolytic capacitors at their adjust pins.

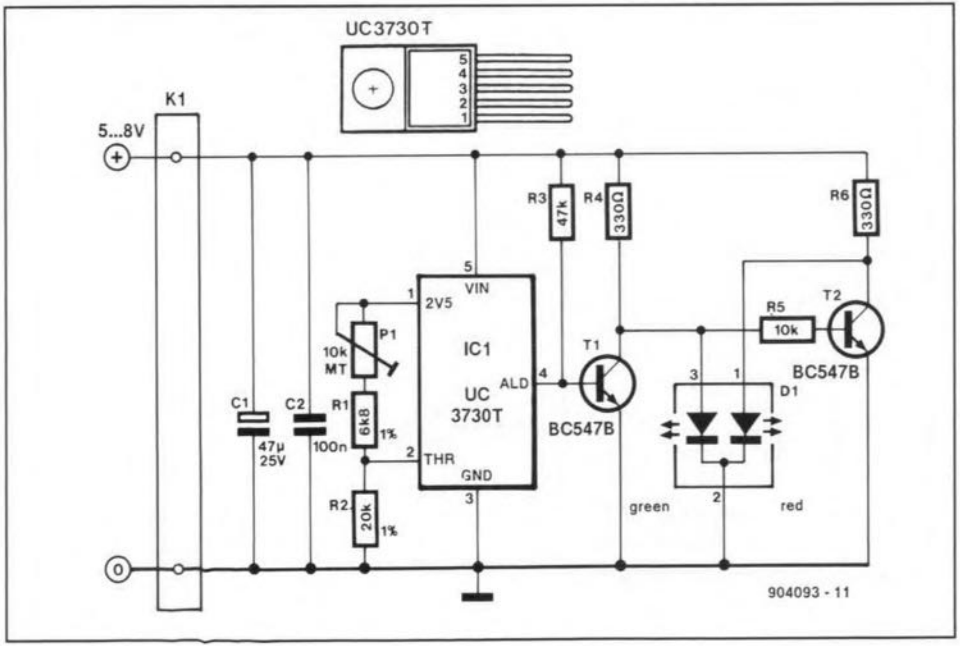
Finally, placing the displays behind red perspex makes them easily readable in all circumstances. ■

(*T. Giffard*)

Reference: "All-solid-state preamplifier", *Elektor Electronics*, December 1989.


7V5
LD1
LD2
CA
LD1,LD2 = HD 1105
IC5 = 4093
R11...R24 = 1k5
IC5c
IC5d
R11
R24
IC3
4543
IC4
4543
LD
PH
BI
R1
R2
R3
R4
R5
R6
R7
R8
1M
S1
S2
IC1
4029
IC2
4029
QA QB QC QD
B/D
CI
CO
CLK PE U/D
U/D PE CLK
IC23
(IC25)
U/D
preset
LSB
IC5a
IC5b
C1
C2
4n7
R10
R9
47k
110mA
C3
C4
C5
C6
C7
100n
IC1
IC2
IC3
IC4
IC5
904119 · 11

Unitrode's UC1730 family of integrated circuits is designed for use in a number of thermal monitoring applications. Each IC combines a temperature transducer, precision reference, and temperature comparator to allow the device to respond with a logic output if temperatures exceed a predetermined level.



The monitor presented here is based on a Type UC3730T and is intended to be fitted to a heat sink. Although the supply to the device can be as high as 40 V, a 5–8 V one is chosen here, because that is normally readily available in the quipment where the monitor may find application: power amplifiers, power supplies, etc.

The threshold temperature, Tt, in °C,is determined by:

$Tt = 2.5R2/0.005(R1+R2+P1) - 273.15.$

The temperature may be preset with P1 to values between −1 °C and +100 °C.

The indicator is formed by a bicolour LED, controlled by transistors T1 and T2. Resistors R4 and R5 limit the current through the LED.

When the temperature of the heat sink is below the threshold temperature, the ALD (alarm delay) output, pin 4, is logic low, so that T1 is switched off and the green LED lights.

S 34

When the temperature of the heat sink exceeds the threshold level, the ALD output goes high, T1 conducts so that T2 is switched off and the red LED lights.

Although the present circuit was designed for use with a heat sink, it may equally well be used for many other thermal monitoring purposes.

The circuit draws a current of about 30 mA from a 5 V supply. ■

(J. Ruffell)

039 UNIVERSAL COMPANDER

Signetics' Type NE575 compander IC is intended primarily for use with battery power supplies of 3–7 V (max. 8 V). It draws a current of 3.5 mA at 3 V and 5 mA at 7 V. The compander process—compression at the input, expansion at the output—significantly improves the signal-to-noise ratio in a communications link.

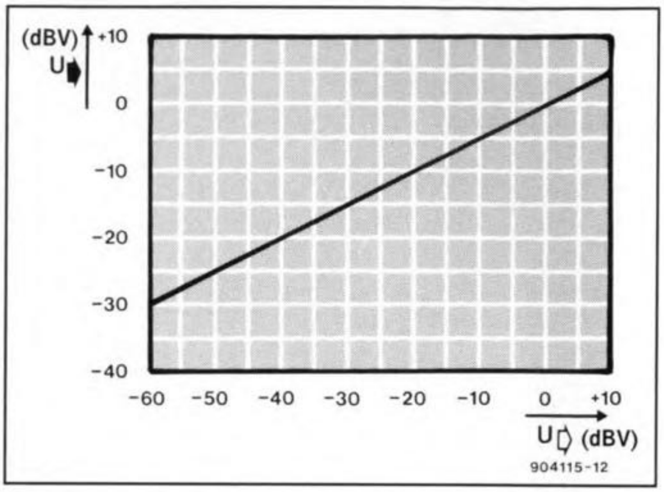
The IC contains two almost identical circuits, of which one—pins 1 to 9—is arranged as an expander. The other —pins 11 to 19—may be used as expander, compressor or automatic load control (ALC), depending on the externally connected circuit. For the compressor function, the inverting output of the internal summing amplifier is brought out to pin 12. This is not the case in the expander section, where a reference voltage is available at pin 8. This pin is interlinked to pins 1 and 19 to enable the setting of the d.c. operating point of the opamps.

The opamp in the expander section, pins 1–3, serves as output buffer—that in the compressor section, pins 17–19, as input buffer.

The IC has a relatively high input sensitivity and is evidently intended for processing small signals (microphone output level). A signal of 100 mV, for instance, is amplified by 1 only.

The present circuit caters for larger input signals (line level): its maximum input level is 1.5 V r.m.s.

With a 1 V input into R13, a potential of about 550 mV exists between compressor output R7 and expander input R5. The



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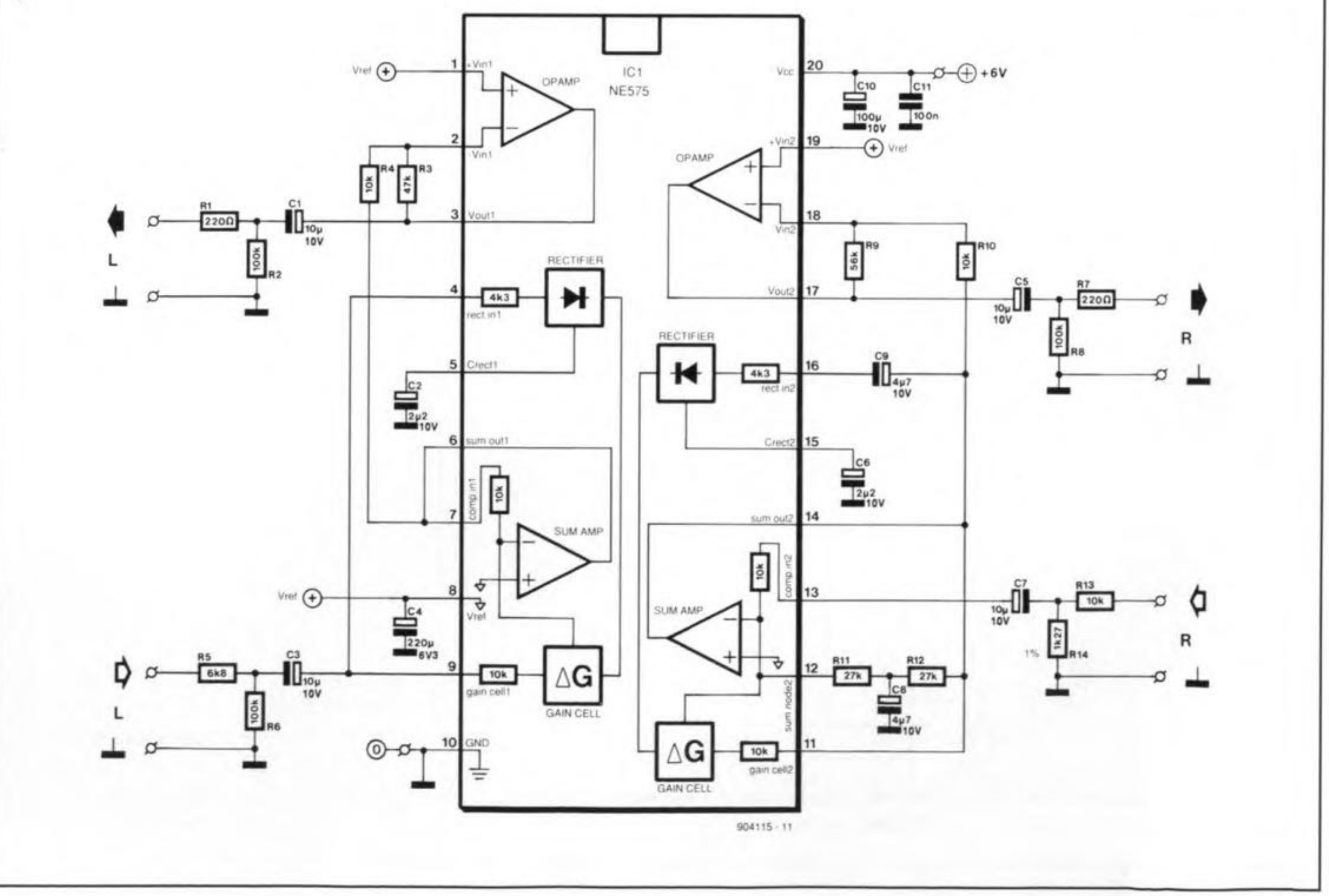


compression characteristic is shown in Fig. 2. The signal range is reduced by about one half at the output, which is doubled in the expander. This means that the range after compression and expansion is the same again, but that is not necessarily the case with the input and output level. The compander may be arranged to provide a constant attenuation or amplification. With the circuit values as shown in the diagram, the input and output levels are the same. The prototype had an overall gain of 0.5 dB when the expander input was connected direct to the compressor output.

To allow acceptance of high input levels, R13, R14 and the compressor input resistance form a 10:1 attanuator. At the expander input, R5 and the expander input impedance of about 3 kΩ form a potential divider. If the compander is to be used with smaller signals, the attenuation may be reduced as appropriate. If the input level lies below 100 mV, R5, R13 and R14 may be omitted.

The compander covers the frequency range of 20 Hz to 20 kHz; the overall distortion is less than 1%; and the signal-to-noise ratio is about 80 dB. ■

(*T. Giffard*)

The display is intended to be added to the address and data bus of the Type 8052 microcontroller. Liquid crystal displays come in a number of varieties: in the prototype a two-row, 16 characters per row type was used, which, moreover, contains two registers.

The signals on the RD and WR lines of the controller are too short to enable data to be written into, or read from, the display registers. The way this problem is resolved consists of using the lowest value address bit, A0, to verify whether a write or a read action is required. The address signals last long enough for completing a data exchange with the display. The next highest address line, A1, is used to differentiate between the data register and the instruction register of the diaplay.

Then:

Basic address: write data into instruction register;

Basic address +1: read contents of instruction register;

Basic address +2: write data into data register;

Basic address +3: read contents of data register.

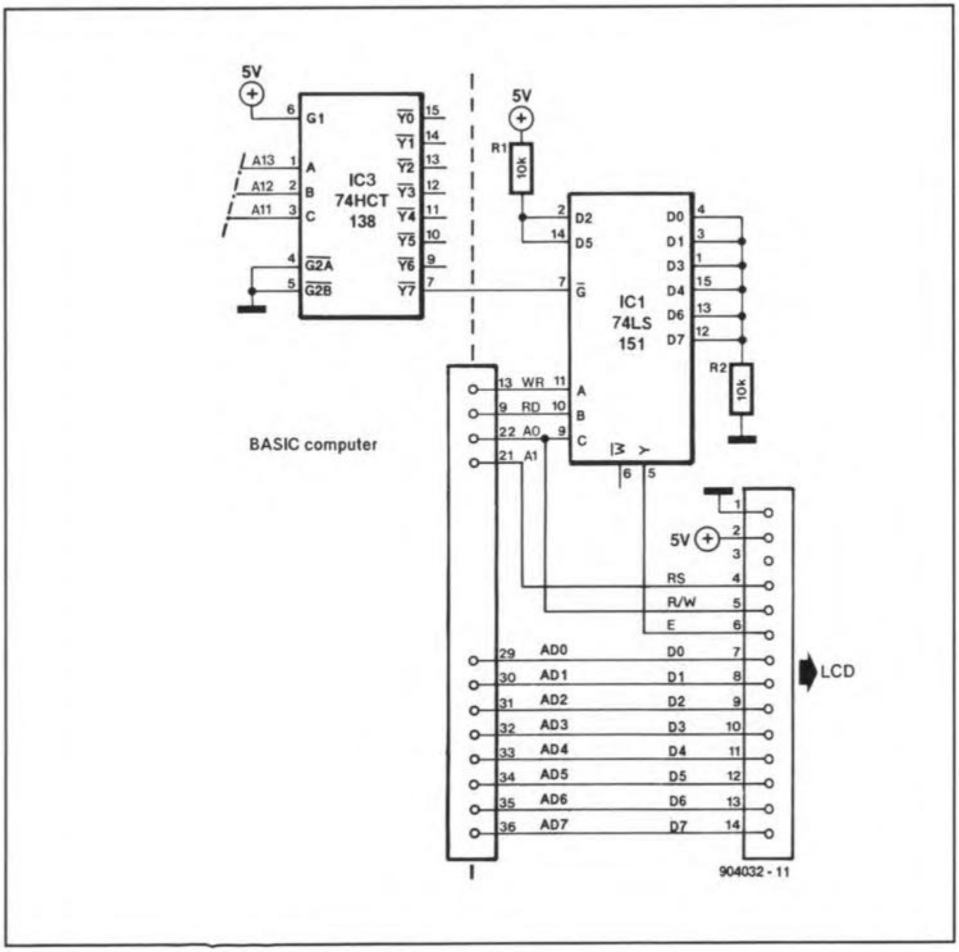
The basic address, which must be a multiple of 4, is determined by the chip select (CS) signal of the controller.

The enable signal for the display is derived from the CS signal, the RD and WR signals, and address signal A0.

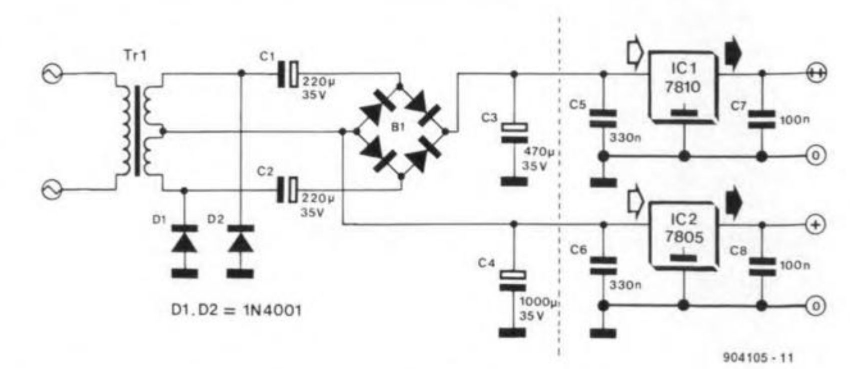
These functions are carried out by a Type 74LS151 IC. This device prevents a spurious address to be read or written and so avoids a conflict between the buses. Only when the display is addressed by CS when either RD or WR are logic low will the address line A0 give an enable signal. The 74LS151 may be replaced by the corresponding HC or HCT type.

If more protection for the controller is required, the data bus may be expanded by a bus driver, for which a bidirectional buffer, such as Type 74LS245, is required. The direction of transfer is determined by the lowest value address line, A0, and the linking of the enable signal with the W signal of IC1. ■

(*J. Romanus*)



041 BRIDGE CIRCUIT FOR ASYMMETRIC LOAD



The bridge circuit is intended for those cases where two unequal supply voltages are required.

The lower voltage is obtained with the aid of a transformer with symmetric windings and half-wave rectification of the potential across one winding.

For the higher voltage, the potential across both windings is rectified. To that end, the output of the transformer is linked to the bridge rectifier via two electrolytic capacitors that provide isolation of the two direct voltages.

A bonus with this type of circuit is that although the two supplies may be loaded unequally, the currents through the two transformer windings are the same. This means that the transformer is loaded symmetrically, so that its full capacity may be used. Moreover, there is no unnecessary dissipation in the voltage regulators.

The load on the lower voltage supply depends primarily on the rating of the transformer. The load on the higher voltage supply is limited by the reactance of C1 and C2 (= $1/2\pi 50C$) and the required minimum output voltage. ■

(A. Rigby)

042 LOW-SIGNAL AF AMPLIFIER

This amplifier is intended to be added to preamplifiers that have no phono input. Such a phono input is, of course, required for normal record players with a dynamic pick-up, of which there are still millions around. Moreover, the amplifier does not only bring the output of the pick-up to line level, it also adds the correction to the frequency response according to RIAA requirements.

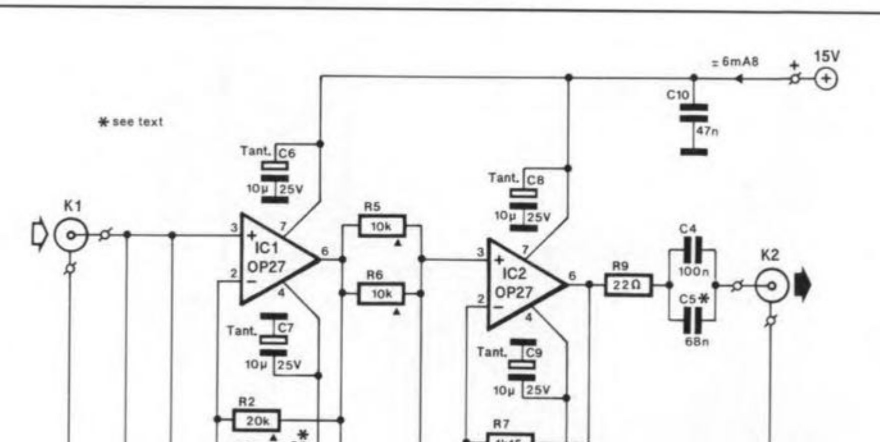
During the recording of gramophone records, the frequency characteristic is lifted at the high end. This lift must be countered in the playback (pre)amplifier. The corrections to the frequency response characteristic are according to a norm set by the Record Industries Association of America (RIAA) and also by the IEC.

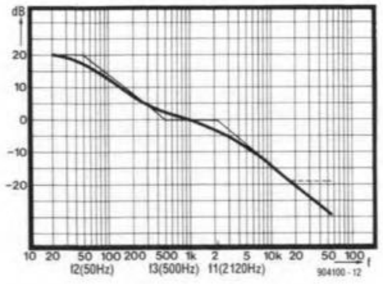
The corrective curve provided by the amplifier is shown in Fig. 2 (bold line). The thin line shows the ideal corrective curve. The sharp bends in this at 50 Hz and 500 Hz are nearly obtained in the practical curve by network R3-C2, while that at just above 2 kHz is approached in practice by filter R5-R6-C3. The arrangement of R3-C2 in the feedback loop of IC1 gives noticeably better results than the usual (passive) filter approach.

Circuit IC1 provides a d.c. amplification of some 40 dB, which drops to about 20 dB when the frequency rises above 500 Hz. To minimize the (resistor) noise and the load of the opamp at higher frequencies, the value of R3 is a compromise. The associated polystyrene capacitor, C2, should have a tolerance of 1–2%.

To raise the 2 mV output of the dynamic pick-up to line level at 1 kHz, linear amplifier IC2 has been added. This stage has a gain of 22 dB, so that a signal of 250 mV is available at its output.

Capacitors C4-C5 at the output, in conjunction with the input impedance of the following preamplifier form a high-pass





filter with a cut-off frequency of 20 Hz: this serves to suppress any rumble or other low frequency noise.

The value of C1 is normally given in the instruction booklet of the dynamic pick-up.

The power supply for the amplifier must be of good quality—particularly, the transformer should be a class A1 type with a small stray magnetic field.

When the amplifier is built into the record player (which is the best way), the power supply should not be included unless this is very well screened; otherwise, hum is the unavoidable result.

In the prototype, Type OP27 opamps were used. A slightly cheaper way is to use a Type OP227 (dual version of the OP27). Opamps from the TL 07X family may also be used. ■

(*T. Giffard*)

The code display is intended as an aid in obtaining a rapid indication as to the available data in an EPROM. It enables up to 13 bits to be read.

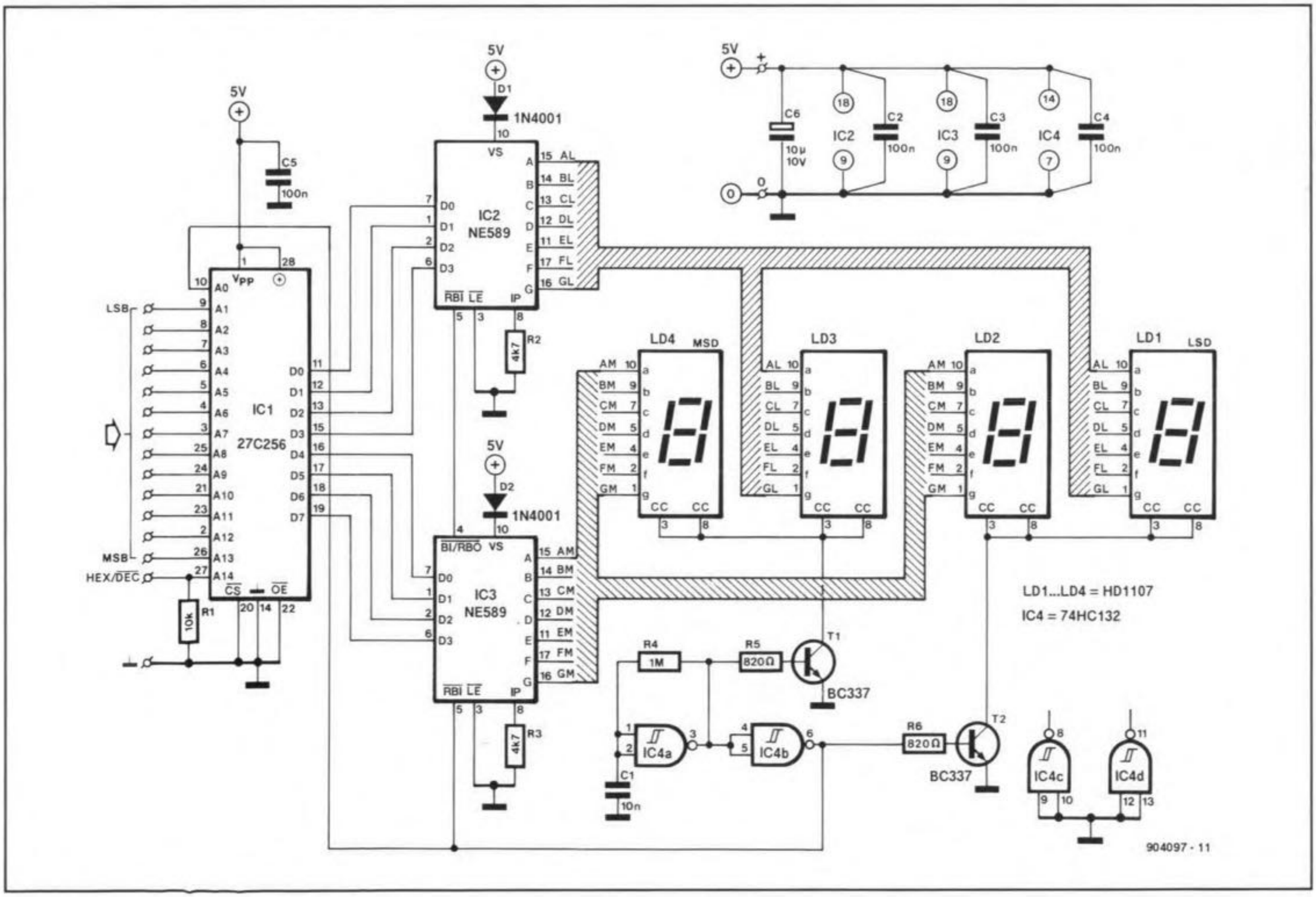
An EPROM will be used to show the application of the circuit as a decimal and as a hexadecimal indication. The contents of the EPROM are shown in the listing in the table. The display will read 00 to 8191 or 00 to 1FFF incl. It is, of course, possible to use a different code. Moreover, by the use of a text tool socket, and changing the EPROM, it is possible to adapt the function of the circuit. Another possibility is using larger EPROMs, which, by switching over the MSB address lines, will immediately make more codes available.

The data output of the EPROM is used to provide data to the two display-decoder/drivers. This arrangement makes it possible to control two displays simultaneously. For instance, four displays may be controlled by just one oscillator and an inverter.

The contents of the EPROM consist of two bytes per 13-bit word. The first is an LSB byte with a nibble for the LSB display and a nibble for the second display. The second byte is an MSB byte that contains a nibble for the MSB display and a nibble for the third display.

The next two successive addresses are used for continuously changing over A0. The arrangement is that when A0 = 0, the MSB and the third display are driven, and when AO = 1, the second and the LSB display. The data can then be read conveniently in the listing.

To minimize the power consumption,



two diodes have been incorporated in the supply to the display section.

The brightness of the display depends on the value of R2 and R3. The output current is about 120×1.3/R2 (or R3). With values as shown, the current is around 30 mA so that the current through the displays is about 15 mA.

Ripple blanking is possible in the case of the two MSB displays only. The ripple blanking is actuated by connecting the clock to RBI of IC3.

The frequency of oscillator IC4a is set to a value where flicker of the displays just stops. It should not be too high otherwise there is a risk of ghosting.

The circuit operates from 5 V; if that is not available on the machine used, the circuit must be extended by level shifters.

The oscillator is designed with HC chips to obtain a symmetrical clock; if HCT devices were used, the hysteresis would not be symmetrical with respect to the supply. This would result in the brightness of the two MSB displays differing from that of the LSB displays. If nevertheless HCT chips are used, a diode and a resistor have to be added to form a symmetrical clock.

The EPROM here forms the input of the circuit. It is worth mentioning that latches may be added to enable the reading of data at a defined moment (possibly with a byte or word recognizer).

The total current drawn by the circuit depends mainly on the display controls and was 500 mA maximum in the prototypes. ■

(T. Giffard)

```
0000:  00 00 00 01 00 02 00 03  00 04 00 05 00 06 00 07
0010:  00 08 00 09 00 10 00 11  00 12 00 13 00 14 00 15
0020:  00 16 00 17 00 18 00 19  00 20 00 21 00 22 00 23
0030:  00 24 00 25 00 26 00 27  00 28 00 29 00 30 00 31
0040:  00 32 00 33 00 34 00 35  00 36 00 37 00 38 00 39
0050:  00 40 00 41 00 42 00 43  00 44 00 45 00 46 00 47
 060:  00 48 00 49 00 50 00 51  00 52 00 53 00 54 00 55
        0 56 00 57 00 58 00 59  00 60 00 61 00 62 00 63
3F         0 65 00 66 00 67  00 68 00 69 00 70 00 71
3F30:  80          0 74 00 75  00 76 00 77 00 78 00 79
3F40:  80 96 80           83  00 84 00 85 00 86 00 87
3F50:  81 04 81 05 81            0 92 00 93 00 94 00 95
3F60:  81 12 81 13 81 14 81         1 01 01 02 01 03
3F70:  81 20 81 21 81 22 81 23  81           1 10 01 11
3F80:  81 28 81 29 81 30 81 31  81 32 81           19
3F90:  81 36 81 37 81 38 81 39  81 40 81 41 81
3FA0:  81 44 81 45 81 46 81 47  81 48 81 49 81 50 81
3FB0:  81 52 81 53 81 54 81 55  81 56 81 57 81 58 81 59
3FC0:  81 60 81 61 81 62 81 63  81 64 81 65 81 66 81 67
3FD0:  81 68 81 69 81 70 81 71  81 72 81 73 81 74 81 75
3FE0:  81 76 81 77 81 78 81 79  81 80 81 81 81 82 81 83
3FF0:  81 84 81 85 81 86 81 87  81 88 81 89 81 90 81 91
4000:  00 00 00 01 00 02 00 03  00 04 00 05 00 06 00 07
4010:  00 08 00 09 00 0A 00 0B  00 0C 00 0D 00 0E 00 0F
4020:  00 10 00 11 00 12 00 13  00 14 00 15 00 16 00 17
4030:  00 18 00 19 00 1A 00 1B  00 1C 00 1D 00 1E 00 1F
4040:  00 20 00 21 00 22 00 23  00 24 00 25 00 26 00 27
  50:  00 28 00 29 00 2A 00 2B  00 2C 00 2D 00 2E 00 2F
         30 00 31 00 32 00 33  00 34 00 35 00 36 00 37
7F3          39 00 3A 00 3B  00 3C 00 3D 00 3E 00 3F
7F40:  1F          42 00 43  00 44 00 45 00 46 00 47
7F50:  1F A8 1F            4B  00 4C 00 4D 00 4E 00 4F
7F60:  1F B0 1F B1 1F            54 00 55 00 56 00 57
7F70:  1F B8 1F B9 1F BA 1F           5D 00 5E 00 5F
7F80:  1F C0 1F C1 1F C2 1F C3  1F           66 00 67
7F90:  1F C8 1F C9 1F CA 1F CB  1F CC 1F           6F
7FA0:  1F D0 1F D1 1F D2 1F D3  1F D4 1F D5 1F
7FB0:  1F D8 1F D9 1F DA 1F DB  1F DC 1F DD 1F DE 1F
7FC0:  1F E0 1F E1 1F E2 1F E3  1F E4 1F E5 1F E6 1F E7
7FD0:  1F E8 1F E9 1F EA 1F EB  1F EC 1F ED 1F EE 1F EF
7FE0:  1F F0 1F F1 1F F2 1F F3  1F F4 1F F5 1F F6 1F F7
7FF0:  1F F8 1F F9 1F FA 1F FB  1F FC 1F FD 1F FE 1F FF
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904097 – 12

044 PURE CONCORDS

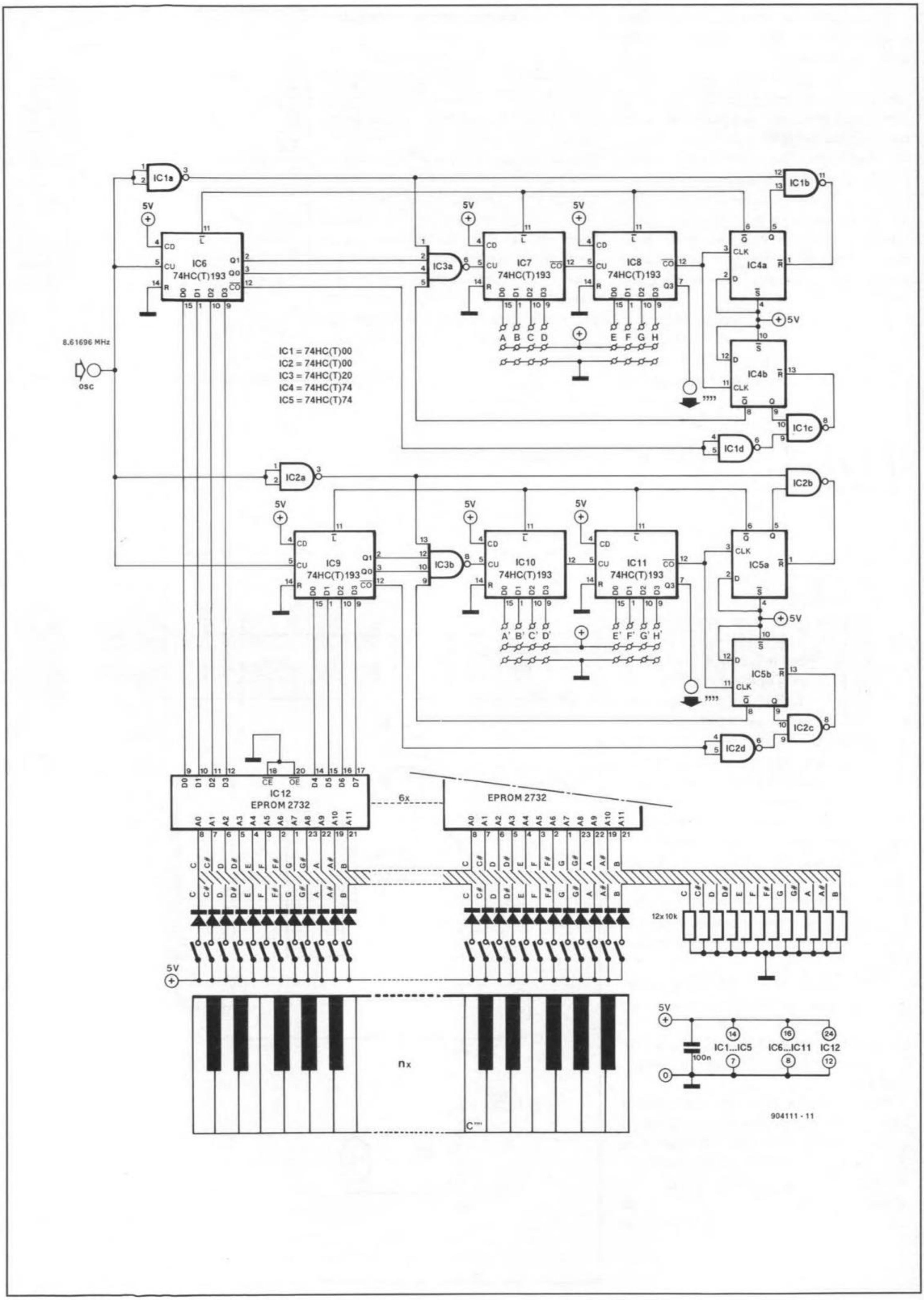
The piano, modern organ, and other fixed-pitch modern instruments are tuned to equal temperament, which means that each semitone is made an equal interval. In other words, the twelve tones in an octave are equi-distant on a logarithmic frequency scale, that is, each tone has a frequency that is 1.059 times greater than the preceding frequency. The advantage of this type of tuning is that the instrument may be played in any key. The disadvantage is, however, that the tones are not tuned 'naturally', which is especially noticeable when concords are played. A perfect fifth should have a ratio of 2:3, but in equal temperament tuning that is 2:2.9966. That is a tiny difference, but it is audible. In a major third it is even worse: 4:5 instead of 4:5.0397.

The circuit presented here offers a remedy. Equal temperament tuning remains the basis, but as soon as a chord is played, the circuit detunes the notes in such a manner that pure concords result. This is accomplished in a way different from that in some modern synthesizers where the tuning is switched to a different key by means of presets.

The present circuit is in principle a main oscillator as found in an electonic organ. The oscillator signal has a frequency of 8.61696 MHz which is divided into twelve for the highest octave (here the fourth above middle C). The signals for the other octaves are obtained from these with the aid of binary scalers.

A total of twelve identical dividers is required, one for each tone. Only two are shown in the circuit diagram: IC6–IC8 and IC9–IC11. All of them are preset. Presetting is used on the one hand to obtain a given pitch (with the aid of jumpers as shown in the table), and on the other to detune the tones in accordance with the chord being played.

Information for the detuning is contained in an EPROM that can control two dividers at a time. A total of six EPROMS is therefore needed: one each for C and C#, D and D#, E and F, F# and G, G# and A, and A# and B.


IC1 = 74HC(T)00
IC2 = 74HC(T)00
IC3 = 74HC(T)20
IC4 = 74HC(T)74
IC5 = 74HC(T)74
8.61696 MHz
osc
IC6 74HC(T)193
IC7 74HC(T)193
IC8 74HC(T)193
IC9 74HC(T)193
IC10 74HC(T)193
IC11 74HC(T)193
IC12 EPROM 2732
EPROM 2732
6x
nx
12x10k
100n
IC1...IC5
IC6...IC11
IC12
904111 - 11

The address inputs of the EPROMS are connected to the make contacts that are added to the keyboard. The octaves available on the keyboard are then applied to the circuit via OR gates formed by diodes. Bear in mind that the only action of importance for the detuning process is which of the twelve keys has been struck, not in which octave the resulting note belongs. This is fortunate, because that means that the size of the design is independent of the number of octaves available on the keyboard, apart from the number of additional make contacts and diodes, of course.

In principle, all information for whatever note combination may be stored in the EPROMS. The present design is limited to that for 2-, 3- and 4-note chords; in all other cases, the equal temperament tuning is maintained. It is unfortunately not possible to list all the data that need to be stored in the EPROMS or the necessary calculations. A package of six programmed EPROMS plus a 5.25 in. (PC) disk containing all the data is, however, available through our Readers Services.

Finally, when during the shifting from one chord to another one or more notes are sustained, the detuning of the tones may become audible. In that case, a different way of playing, that is, waiting for the notes to die down before the next note is played, is the price for a natural sound. ■

(J. Peereboom)

Where to connect the jump leads.

	A	B	C	D	E	F	G	H
C	⊥	⊥	⊥	⊥	⊥	⊥	⊥	⊥
C#	⊕	⊕	⊕	⊕	⊥	⊥	⊥	⊥
D	⊥	⊥	⊕	⊕	⊕	⊥	⊥	⊥
D#	⊕	⊥	⊥	⊕	⊥	⊕	⊥	⊥
E	⊕	⊥	⊕	⊥	⊕	⊕	⊥	⊥
F	⊕	⊥	⊥	⊥	⊥	⊥	⊕	⊥
F#	⊕	⊕	⊥	⊕	⊥	⊥	⊕	⊥
G	⊥	⊕	⊕	⊥	⊕	⊥	⊕	⊥
G#	⊕	⊕	⊕	⊕	⊕	⊥	⊕	⊥
A	⊥	⊥	⊥	⊕	⊥	⊕	⊕	⊥
A#	⊕	⊥	⊥	⊥	⊕	⊕	⊕	⊥
B	⊕	⊥	⊥	⊕	⊕	⊕	⊕	⊥

045 DIGITAL VOLUME CONTROL

The heart of the digitally operating volume control is IC2, a Type 4067 16-channel analogue multiplexer.

Depending on the logic state on pins A, B, C and D of the multiplexer, one of its 16 inputs or outputs is connected to pin 1, which is the 'wiper' of the control.

Since a 1 kΩ resistor has been connected between each input and output, the multiplexer may be considered a linear potentiometer with 16 fixed steps. Its overall resistance is 15 kΩ. It is, of course, possible to use a different value for each of the resistors to obtain a different characteristic, for instance, a positive logarithmic one.

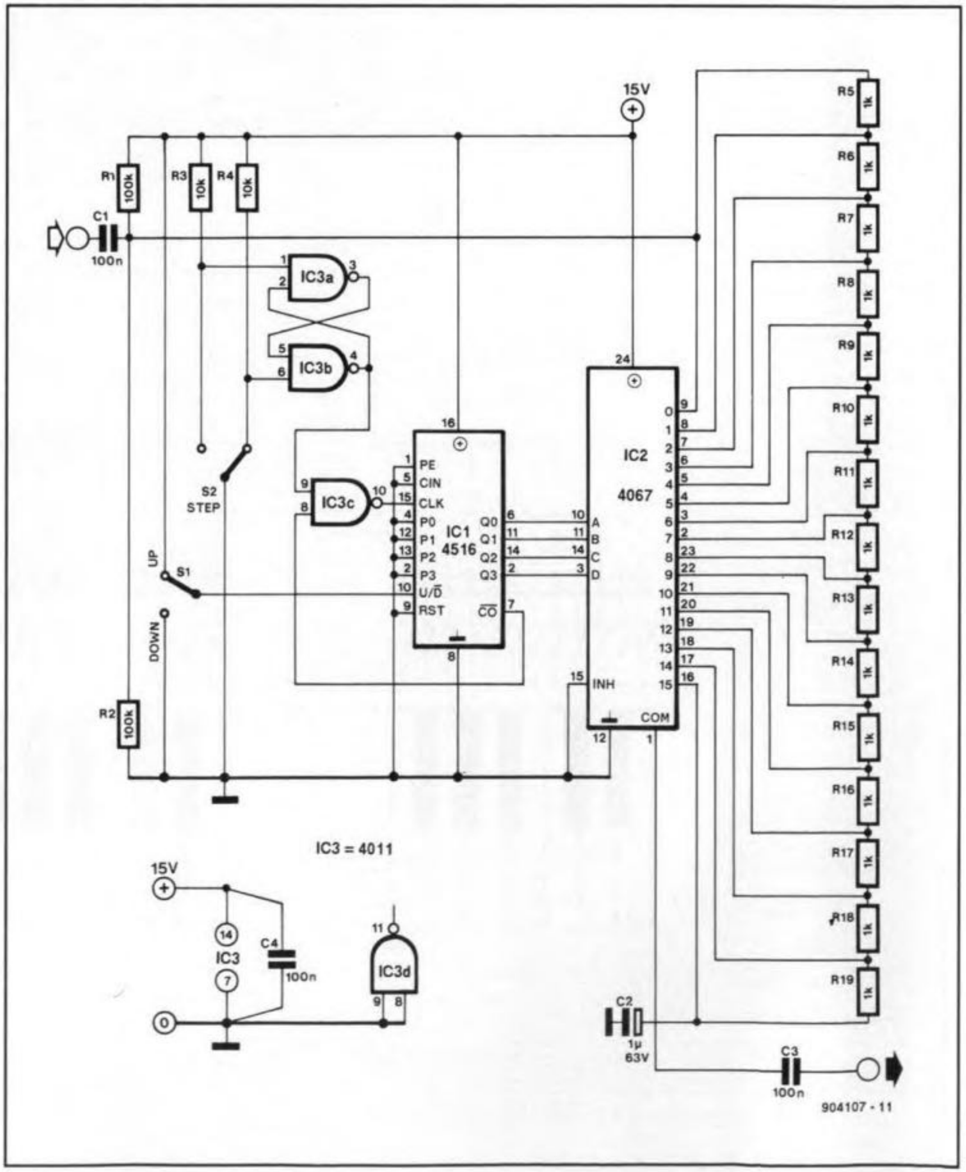
The setting of the potentiometer is controlled by counter IC1. Dependent on the position of switch S1, the counter moves one step up or down when switch S2 is changed over. Circuits IC3a and IC3b provide debouncing of S2.

A jump from 0000 to 1111 or the other way around is not possible, because further count pulses are suppressed with the aid of the $\overline{CO}$ line. This line is logic low when both the counter state and signal U/$\overline{D}$ are 0.

When U/$\overline{D}$ is high and the counter state is 15, $\overline{CO}$ again becomes logic low. It is then necessary to reverse the logic state at U/$\overline{D}$ and thus the direction of counting.

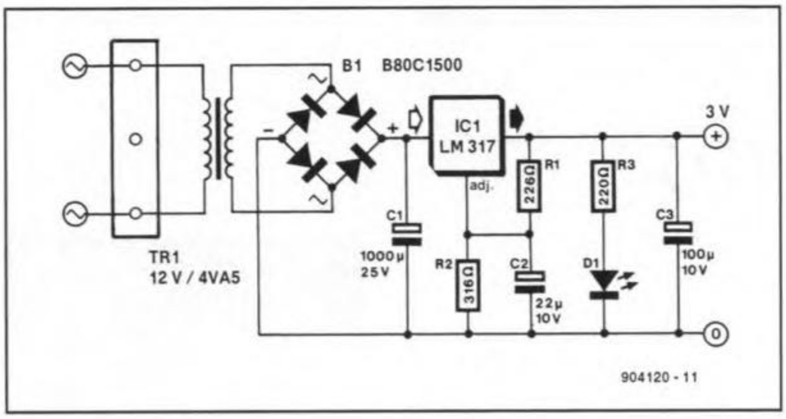
The volume control draws a current of around 1 mA. ■

(*A. Ferndown*)



3 V MAINS SUPPLY FOR PORTABLE RADIOS

046



Most small portable radios require a 3 V supply, normally provided by two size AA or AAA batteries. Since rechargeable batteries are an option with many of these radios, most of them are fitted with a charger socket. When such radios are used in a stationary condition, e.g. in the kitchen or in the office, it is useful (and economical) to use the mains operated supply described here.

The supply is small enough to be fitted inside the radio or in a mains adaptor case (less the transformer).

Voltage regulator IC1 is adjusted for an output of 3 V by resistors R1 and R2, which are decoupled by C2. Capacitor C3 provides addtional filtering. Diode D1 indicates whether the unit has been connected to the mains. The diode also provides the load necessary for the regulator to function properly; in its absence, the secondary voltage of the transformer might become too high when the unit is not loaded.

The transformer should be a short-circuit-proof miniature type rated at 12 V and 4.5 VA. The secondary voltage is slightly higher than needed for a radio, but this reserve is useful when the unit is used with a cassette or CD player.

It is advisable to check the output voltage of the unit when it is switched on for the first time before connecting it to a radio or cassette player. ■

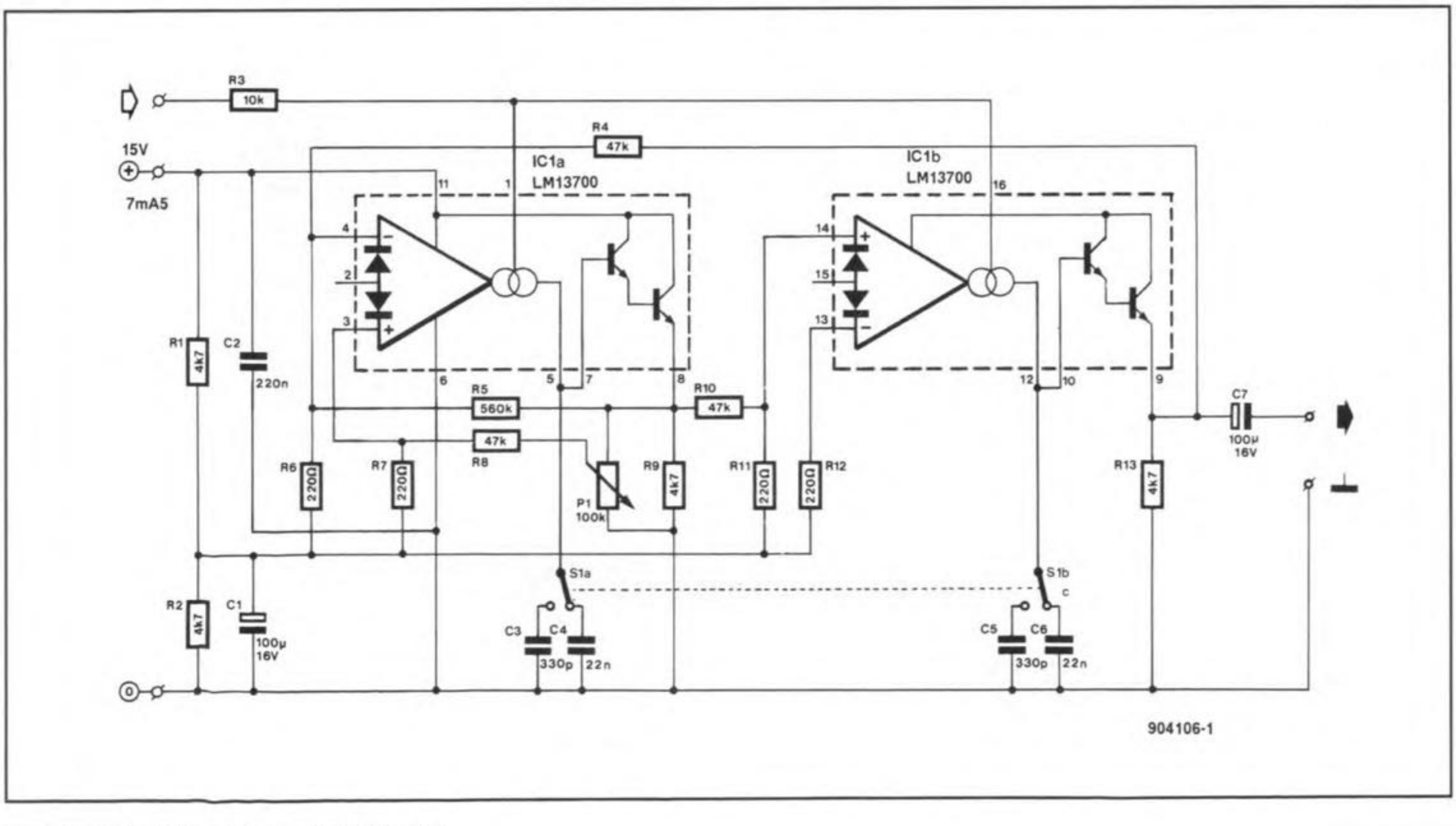
(T. Giffard)

SIMPLE VCO

047

The frequency of the sine wave oscillator shown here is determined by a direct voltage, *Uc*, of 0–15 V. The distortion on output signals of up to 10 V p-p is not greater than 1%; when the output is reduced with the aid of P1 to 1 V p-p, the distortion drops to below 0.1%. It is not recommended to use output signals below 1 V p-p, because the oscillator then become unstable and temperature-dependent.

The oscillator consists of two operational transconductance amplifiers (OTAs)



contained in one package. Their AMP-BIAS inputs, pins 1 and 16, are connected in parallel. These inputs can drive the output currents at pins 5 and 12 to a peak value of up to 0.75 mA.

Switch S1 enables the oscillator output to be set to two ranges: 6.7–400 Hz and 400 Hz to 23.8 kHz. The overall range needs a control voltage varying from 1.34 V to 15 V. When the frequency is changed by a variation of *Uc* and the setting of P1 is not altered, the output signal may be distorted. In other words, the amplitude of the signal must be adapted to the frequency. ■

(*T. Giffard*)

048 LOGIC TESTER

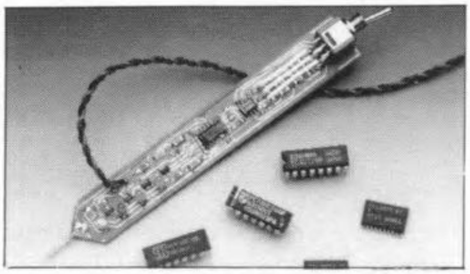
The logic tester described here is designed in surface mount technology, which makes it very compact indeed, as may be seen from the printed circuit boards.

The input consists of two comparators that operate with different reference voltages supplied by separate potential dividers. Divider R3-R4-R5 provides a voltage of about 40% of the supply voltage, *Ucc*, to pin 6 of IC1b and one of about 16% of *Ucc* to pin 3 of IC1a. When *Ucc* = 5 V, these voltages are exactly the thresholds (0.8 V and 2.0 V) of TTL comparators.

Similarly, divider R6-R7-R8 provides voltages of 23% of *Ucc* and 73% of *Ucc* to pin 3 of IC1a and pin 6 of IC1b respectively; these levels correspond to the standard threshold for CMOS comparators.

The voltage to be measured, *Ua*, is applied to pin 5 of IC1b and pin 2 of IC1a and compared with the respective reference. The output of comparator IC1b goes high when *Ua* exceeds the reference, whereas the output of IC1a goes high when *Ua* lies below the voltage at pin 3.

The comparators are followed by driver stages, T1 and T2, for the LED display—D1 for 'high' and D2 for 'low'—and also NOR gate IC2a that switches on T3 when the output of both comparators is low, that is, when it is undefined. This state is indicated by D3.



The remaining three gates in IC2 form a monostable. During quiescent operation, *Ucc* is present at the input of inverter IC2c. The output of the inverter is then low, T4 is off and D4 is out. Pin 4 of IC2b is also high, but this state changes when a pulse arrives at pin 5. The output of IC2b then goes low, C2 discharges, the inverter toggles, T4 is switched on and D4 lights. This state is unstable, however, because C2 recharges via R13. Although the pulse at pin 5 may be very short, the time constant R13-C2 lengthens it to about 100 ms.

The supply voltage may lie between 5 V and 15 V. At 5 V, the circuit draws a current of about 15 mA.

The input impedance of the tester is of the order of 330 kΩ. ■

(*J. Ruffell*)

PARTS LIST

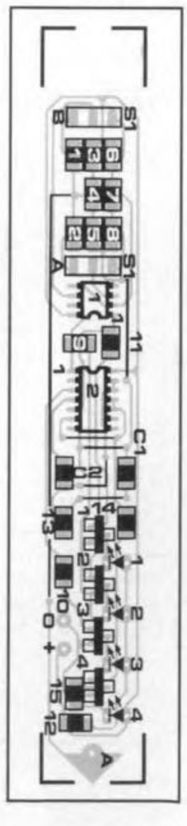
Resistors:
R1, R13 = 1 M
R2 = 470 k
R3 = 39 k
R4, R6, R8 = 15 k
R5, R9, R10, R11, R14 = 10 k
R7 = 27 k
R12, R15 = 1 k

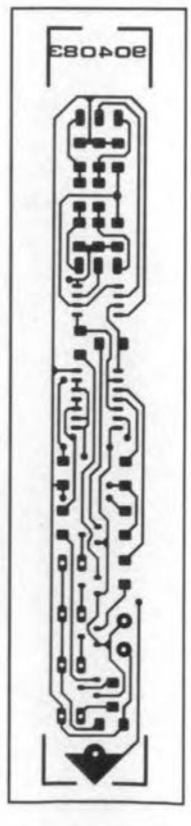
Capacitors:
C1, C2 = 100 n

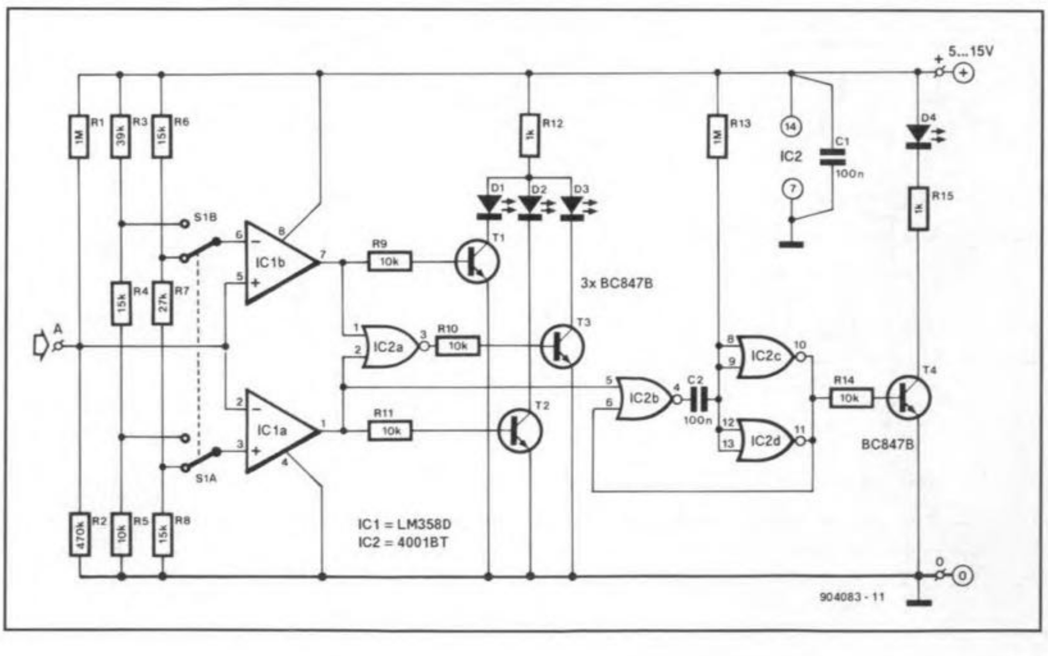
Semiconductors:
D1, D2 = LED, 3mm, green
D3 = LED, 3 mm, red
D4 = LED, 3 mm, yellow
T1, T2, T3, T4 = BC847B
IC1 = LM358D
IC2 = 4001BT

Miscellaneous:
S1 = sub-miniature switch, 2 make before break contacts

NOTE: all components must be for surface mount technology.







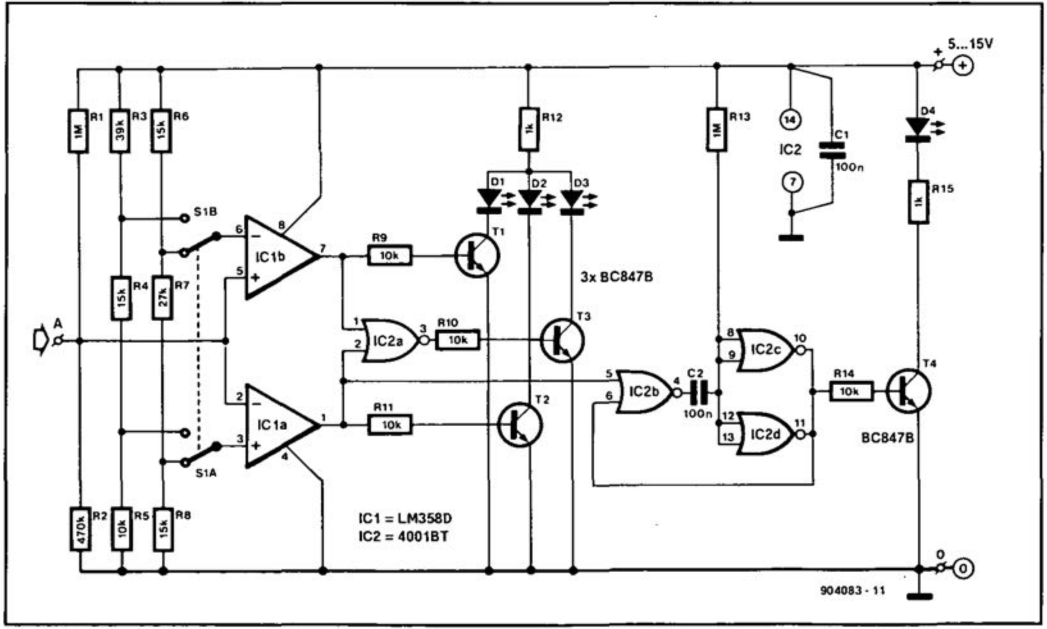
when *Ua* lies below the voltage at pin 5.

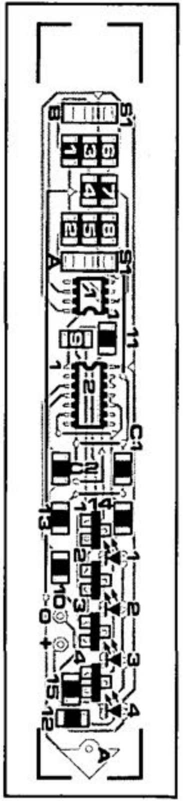
The comparators are followed by driver stages, T1 and T2, for the LED display—D1 for 'high' and D2 for 'low'—and also NOR gate IC2a that switches on T3 when the

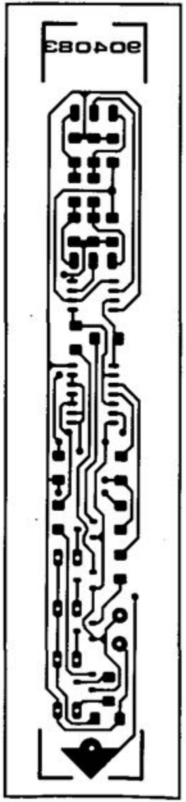
5 V and 15 V. At 5 V, the circuit draws a current of about 15 mA.

The input impedance of the tester is of the order of 330 kΩ. ■

(J. Ruffell)

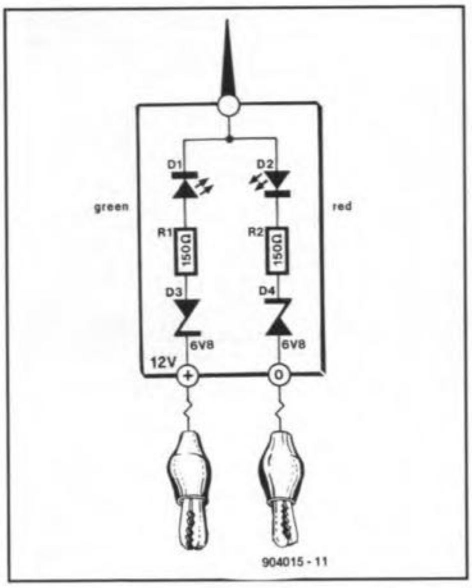






TESTER FOR GOODS VEHICLE ELECTRICS

049



This is a useful little tester for use in testing and checking the electric circuits of a goods vehicle.

Two LEDs indicate whether one of the clips is connected to the positive supply line (red) or to mass (green).

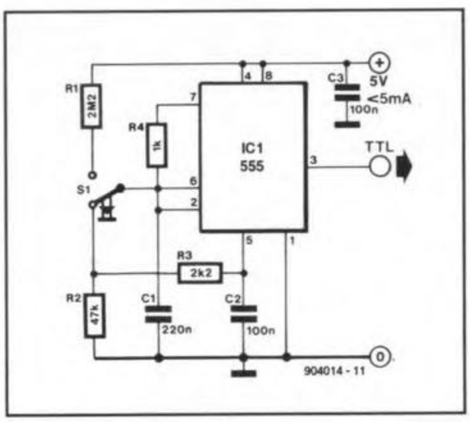
The unit is powered by the vehicle battery. It is advisable to terminate the unit into two insulated heavy-duty crocodile clips. These enable connection to be made direct to the battery or to terminals on the fuse box. It is also possible to terminate it into a suitable connector that fits into the cigarette lighter socket.

If a sharp needle is soldered to one of the terminals, it is possible to check insulated wiring—but only that carrying 12 V. Although the needle pierces the insulation, it does not damage it. ■

(*D. Folger*)

BOUNCE-FREE AUTO REPEAT SWITCH

050



A switch that keeps on giving pulses as long as it pressed is often required. The circuit here uses the well-known Type 555 for this purpose. Its output is a TTL compatible signal.

At pin 5 of the timer exists a potential of 67% of the supply voltage, *Ucc*. In the quiescent condition (switch not pressed), C1 charges via R2 and R3 to a voltage that is lower than that at pin 5 and thus also lower than the toggle voltage.

When the switch is pressed, C1 is rapidly charged via R1 to the toggle voltage upon which the timer emits a pulse. At the same time, the capacitor is discharged again via R4.

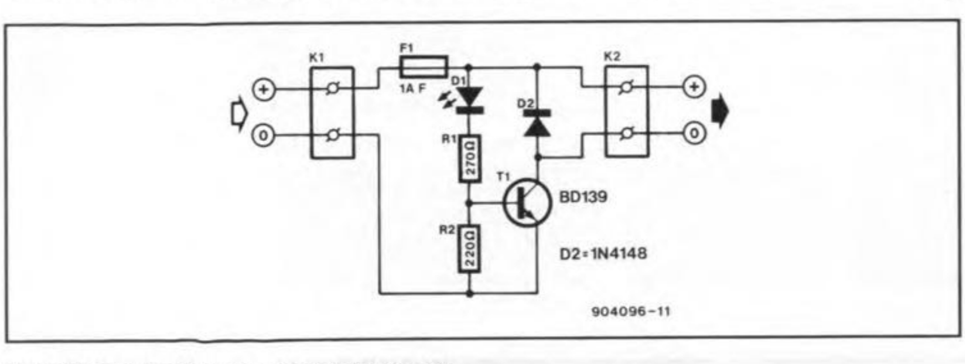
As long as the switch is pressed, the circuit functions as an astable toggle and produces pulses. When it is released, the capacitor cannot charge to the toggle voltage. ■

(*B. Krien*)

FLASHING LED CONTROLLER

051

The light-emitting diode with integrated flasher is connected in series with the base emitter junction of transistor T1. This results in a load connected to K2 being switched on and off in rhythm with the flash rate. This load may be a relay or a lamp.



It is essential that the maximum collector current of the transistor (of the BD139 = 750 mA) is not exceeded. If that is not sufficient, a power darlington may be used, which will give some amperes.

The current drawn by the circuit under no-load conditions amounts to 20 mA. ■

(*J. Ruffell*)