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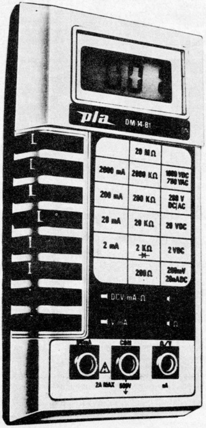

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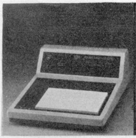

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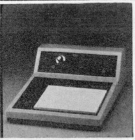

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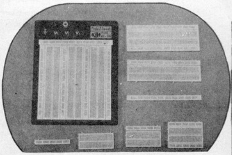

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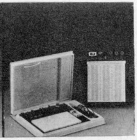

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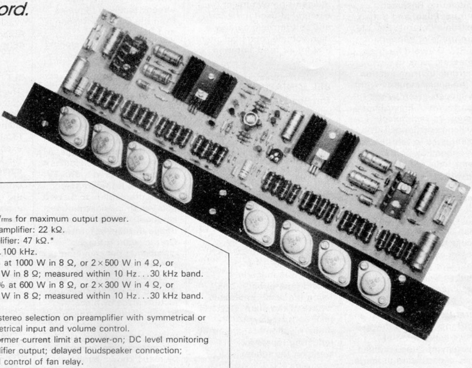
HIGH-POWER AF AMPLIFIER - 1

Here is an amplifier that meets the demand for good quality sound reproduction at very high sound pressure levels. Capable of delivering either 2x500 W in a stereo arrangement, or 1000 W in a bridge configuration, this design may be called powerful in the true sense of the word.



Technical characteristics

Input sensitivity:	775 mV$_{rms}$ for maximum output power.
Input impedance:	power amplifier: 22 kΩ. preamplifier: 47 kΩ.*
3dB bandwidth:	8 Hz...100 kHz.
Distortion:	<0.1% at 1000 W in 8 Ω, or 2×500 W in 4 Ω, or 2×250 W in 8 Ω; measured within 10 Hz...30 kHz band. <0.01% at 600 W in 8 Ω, or 2×300 W in 4 Ω, or 2×200 W in 8 Ω; measured within 10 Hz...30 kHz band.
Damping factor:	>100.
Features:*	mono/stereo selection on preamplifier with symmetrical or asymmetrical input and volume control. Transformer current limit at power-on; DC level monitoring at amplifier output; delayed loudspeaker connection; thermal control of fan relay.

* to be discussed in part two.

The considerable power reserve of the amplifier described in this article will be of definite interest for applications in discotheques as well as in large PA (public address) systems, where a sufficiently high SPL (sound pressure level) for the low and lower middle audio frequency ranges is normally only attainable through a combination of amplifiers and a number of stacked, high-efficiency bass bins.

Apart from presenting an amplifier with outstanding features, both as to performance and reliability, this article is also interesting from a theoretical point of view, since processing small audio signals to ten odd amperes of output current requires quite a lot of attention to overall efficiency and problems pertaining to stability, as well as to optimum transmission of dissipated heat.

General considerations

An amplifier with an output capability of the order of 1000 watts poses problems as to the heat dissipation of the power output stage. In order to shed light on these problems, their theoretical aspects will be briefly discussed below.

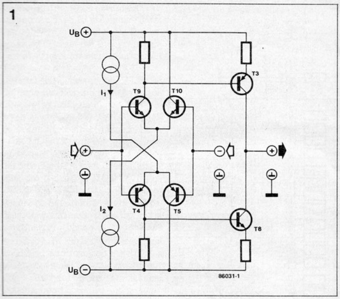
In theory, the output stage has a maximum efficiency of 78.5%; that is, with maximum drive level applied and disregarding the transistors' drain-source saturation voltage of about 2.5 V. At 1000 W output power, therefore, the DC input, P_{in}, to the final stage amounts to

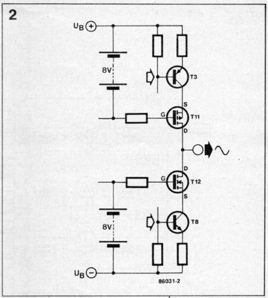
$$P_{in}=1000(100/78.5)=1274 \text{ watts.}$$

The maximum dissipation, however, does not occur at full drive, since the overall efficiency drops with lower drive levels to the output stage, but, theoretically, at a drive level of 64%, and amounts to

$$P_{diss}=0.4P_{out}=0.4\times 500=200 \text{ W per channel.}$$

Since this is a stereo design, we can





expect 400 W in the worst case condition. The losses due to the quiescent current add a further few watts to the total dissipation. Given a quiescent current of 100 mA per transistor, i.e. 400 mA per channel, the additional power demand P_{qc} is calculated from

$P_{qc} = 0.4 \times 75\,V \times 2 = 60\,W$ per channel.

Again, this figure should be doubled, since there are two identical channels. Note that the factor two in the above calculation represents the symmetrical ±75 V supply. In conclusion, it is seen that, theoretically, each power transistor dissipates some 33 W in a worst case condition. Obviously, this calls for a suitable heat-sink with very low thermal resistance, supported by a powerful fan which is switched on automatically when the heat-sink temperature exceeds a safe value.
In order to achieve maximum efficiency and a large signal handling capability, both at instantaneous and continuous operation near the peak output power level, the amplifier input and driver sections have been arranged to operate from a higher supply voltage than the output stage; this ensures full drive reserve in all conditions and thus avoids driver 'pulling' at peak output currents.
Of necessity, several protective measures have been incorporated in the present amplifier design, since its huge power reserve is capable to destroy even the most rugged of high-power loudspeakers in the absence of suitable circuitry to delay both the speaker connection and the presence of the full mains voltage at the power transformer primary winding. Also, the heat-sink temperature and the output DC level are under constant surveillance in order to timely detect amplifier malfunctions are/or gross distortion occurring in overdriving conditions. All of these protections aim at preventing costly and disastrous bangs in the loudspeaker(s) and blown mains fuses when the amplifier is switched on.
This article discusses theory and construction of the main high-power amplifier board, two of which are required for a 2×500 W (4 Ω, stereo setup) or a single 1000 W unit (8 Ω, bridge connection). Next month's issue of *Elektor Electronics* will deal with the power supply for the input and driver sections, a stereo/bridge preamplifier, details for setting up and testing, the protective circuitry, and constructional hints.

Basic section design

The functional division of the present amplifier board into input stage, driver stage, and power stage is a logical consequence of the specific task assigned to each circuit section. All functions have been thoroughly analysed and the resulting basic section designs will be discussed below.
The input section has been devised for optimum characteristics as regards low noise level, stability, and frequency response. Figure 1 shows the basic concept of this section which exhibits outstanding qualities by virtue of its symmetrical arrangement. At the left is the audio input, at the right the input for a portion of the amplifier output signal (feedback). Basically, the circuit consists of two complementary, differential amplifier stages (T_9-T_{10}; T_4-T_5), each with its associated current source. Alternating voltages at the inputs 'see' the differential amplifiers as connected in parallel. The advantages offered by this setup may be summarized as follows: first, the complementary transistor types and the equal currents, supplied by I_1 and I_2, cause the base currents of T_9 and T_{10} to counterbalance with respect to the input; secondly, the four transistors operate at virtually constant collector-emitter voltage, which makes for constancy of capacitive feedback characteristics and, consequently, a further reduction of possible non-linear operation. Furthermore, the constant voltage ensures pure current amplifier operation of the differential configuration so as to obviate the need for the internal transistor capacitances to be charged and discharged at audio speed; this works out to be a great asset as to the quality of amplification at low collector currents, and, therefore, the low-noise and high cut-off frequency properties of the design. In short, the input section achieves a remarkably low TIM (transient intermodulation) distortion figure. Driver transistors T_3 and T_8 must provide clean voltage amplification; however, contrary to the basic arrangement shown in Fig. 1, these transistors have, in fact, been cascaded and connected to driver MOSFETs — see Fig. 2.
The typical advantage of the driver cascade setup is a further improvement upon the already highly linear $I_d = f(U_d)$ curve, relevant to these complementary MOSFET devices. Moreover, the extensive frequency range of this driver design fully matches that of the input stage as de-

Fig. 1. Basic circuit arrangement of the amplifier input section. If correctly dimensioned, it offers excellent performance.

Fig. 2. The driver circuit of the high-power amplifier is basically a symmetrical and complementary cascade configuration. The application of VMOSFET drivers ensures ultra-linear operation over a wide range of audio signal levels.

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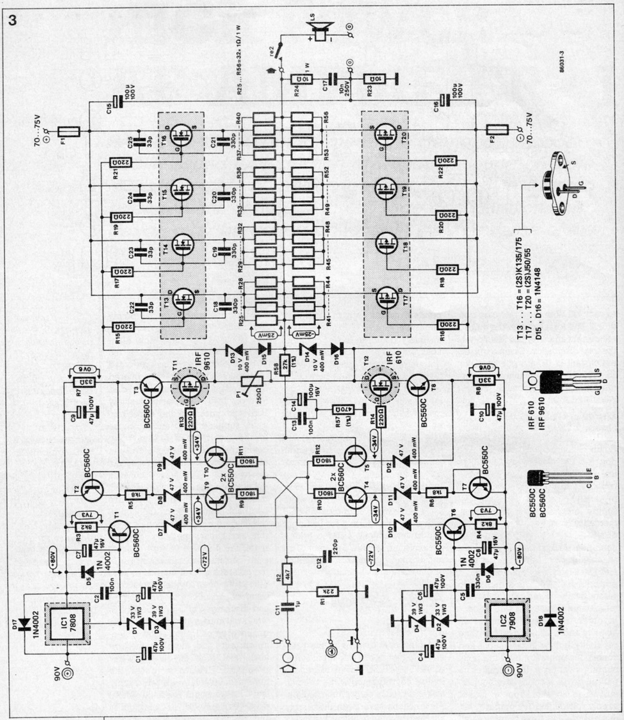

Fig. 3. The basic circuits of Fig. 1 and 2 can easily be spotted in this circuit diagram of the high-power amplifier. Note that this is but one of two identical units!

scribed.

The power output section is basically a conventional push-pull design with complementary N- and P-channel power MOSFETs of the horizontal type, selected for good transient response and linearity at all possible drive levels.

Circuit details

A careful examination of the circuit diagram shown in Fig. 3 reveals the practical realizations of the sections discussed above. Note that the part numbers have been retained for this purpose.

There is a fair number of zener diodes in the circuit; $D_1 \ldots D_4$, together with IC_1 and IC_2, provide the stable ±80 V supply voltage for the input and driver section. $D_7 \ldots D_{12}$ ensure the presence of the correct supply voltage for the complementary, low-noise transistors Types BC550C-BC560C. T_1 and T_6 supply a constant collector current to each differential amplifier; set to about 0.45 mA per transistor, this current constitutes the right compromise between minimum noise level and maximum cut-off frequency of the input stage. T_2 and T_7 have been connected as diodes to reduce the voltage excursion at the collectors of T_9 and T_4, as well as to correct any thermal runaway effects in T_3 and T_8. The quiescent current of the driver stage has been arranged at a fixed 25 mA, which flows through T_3, T_{11}, T_{12}, and P_1. The latter is used to set the quiescent current of 400 mA for the power output stage.

Unfortunately, power MOSFETs of the type used in the present amplifier tend to oscillate quite easily, especially when connected in parallel. In order to combat this tendency, each MOSFET is fitted with a low-value gate resistor. Owing to essential differences in their internal structure, the N-channel MOSFET Types 2SK135 and 2SK175 typically present lower gate-to-source and gate-to-drain capacitances than the complementary, P-channel Types 2SJ50 and 2SJ55. To avoid output stage unbalancing and resultant instability, a number of small ceramic capacitors, C_{18}...C_{25}, are fitted at suitable points around T_{13}...T_{16}.

Diodes D_{13}...D_{16} limit the drain current of each MOSFET to 5 A in case of an output short circuit. This effective protection causes no measurable distortion during normal operation.

Each MOSFET source terminal is connected to the loudspeaker output rail by four parallel connected 1 watt type resisistors. These are used instead of a single 4 watt type, which is typically a wirewound type. This type of resistor cannot be used here since it would present a stray inductance in a highly critical location, causing amplifier instability and a strong tendency to oscillate.

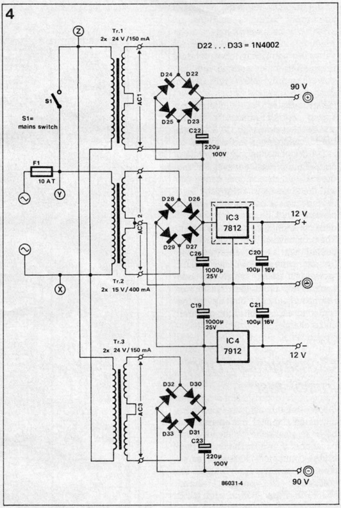

Fig. 4. The driver and preamplifier supply section provides a higher output voltage than the power stage supply to ensure sufficient drive at continuous operation near peak amplifier output. The construction of this supply unit will be reverted to in part 2 of this article.

Power supplies

The circuit diagram of Fig. 4 shows the ±90 V supply for the input and driver sections of two amplifier boards, as well as the necessary supply voltages for the protective circuitry. This combined power supply will be reverted to in next month's second article.

Figure 5 shows the ±75 V, high-current power supply for the two amplifier boards as described in this article. It should be made quite clear at this stage that the final sound quality of the proposed amplifier depends direct and inevitably on the current sourcing capability of this power supply. Any attempt to skimp on this vital section will result in failure of the amplifier to produce good sound quality, especially in the low and lower middle frequency ranges where generally most of the music power is contained. The proposed supply ensures good amplifier response to continuous as well as short-duration signal peaks generated by musical instruments such as electric bass guitars, bass drums, or synthesizers.

To meet the current demand of the amplifier boards, the proposed ±75 V supply incorporates two identical, toroidal 750 VA mains transformers, a 25 A bridge rectifier, and 2×30,000 µF smoothing capacitors. It stands to reason that the construction of such a supply unit deserves the necessary care and attention, and this will also be reverted to in next month's article.

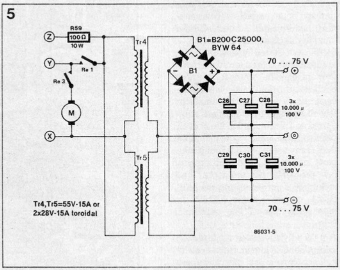

Fig. 5. Vital to the correct operation of the amplifier, this powerful mains supply unit is equipped with two toroidal transformers and a suitably dimensioned smoothing section, capable of catering for the amplifier's high current demand.

Fig. 7. Component mounting plan for the high-power amplifier.

Parts list (relevant to a single amplifier board)

Resistors:

R1 = 22 k
R2 = 4k7
R3;R4 = 8k2
R5;R6 = 1k8
R7;R8 = 33 Ω
R9 . . . R12 = 180 Ω
R13 . . . R22 = 220 Ω
R23;R24 = 10 Ω; 1 W
R25 . . . R56 = 1 Ω; 1 W
R57 = 470 Ω,1%
R58 = 22 k;1%
P1 = 250 Ω preset (good quality!)

Capacitors:

C1;C3;C4;C6;C9; C10 = 47 μ; 100 V; electrolytic
C2 = 100 n
C5 = 330 n
C7;C8;C13 = 47 μ; 16 V; electrolytic
C11 = 1 μ; MKT
C12 = 220 p
C14 = 100 μ; 16 V; electrolytic
C15;C16 = 100 μ;100 V; electrolytic
C18 . . . C21 = 330 p
C22 . . . C25 = 33 p

Semiconductors:

D1;D2 = zener diode 33 V; 1.3 W
D3;D4 = zener diode 39 V;1.3 W
D5;D6;D17;D18 = 1N4002
D7 . . . D12 = zener diode 47 V; 0.4 W
D13;D14 = zener diode 10 V; 0.4 W
D15;D16 = 1N4148
T1 . . . T5 = BC560C
T6 . . . T10 = BC550C
T11 = IRF9610/9612/9620/9622 (International Rectifier)
T12 = IRF610/612/620/622 (International Rectifier)
T13 . . . T16 = 2SK135/2SK175* (Hitachi)
T17 . . . T20 = 2SJ50/2SJ55* (Hitachi)
IC1 = 7808 + finned heat-sink
IC2 = 7908 + finned heat-sink

Miscellaneous:

heat-sinks for T11;T12 (37.5 mm e.g. Fischer SK59)
2 PCB-mount fuse holders

To get the most out of the amplifier, all supply wiring should be of 2.5 mm² cross-sectional area, preferably heat-resistant stranded wire. Do not fail to observe due precautions when working with this power supply; 150 volts is a dangerous level!

Power resistor R59 in the mains supply line prevents the mains and/or the domestic 13 A fuse(s) from blowing when the amplifier, or rather the power supply, is switched on. Without this current limiting device, the discharged capacitors and the absence of a magnetic field in the mains transformers would cause a very high, momentary mains current, enough to blow the fuses. The protective circuitry, which is discussed next month, energizes Re1 (i.e. short circuits R59) after a short 'power-on' delay, which is long enough to allow the transformer magnetic field to be built up and the smooting capacitors to be given an initial charge.

Construction and initial test

Before commencing the construction of the amplifier, it is advisable to be quite clear as to its intended applications. If it is to be used as a bridge-connected 1000-watt mono type, the power supply should be configured as outlined above. MOSFETs Types 2SK175 and 2SJ55 are then preferred to Types 2SK135 and 2SJ50: the former are more rugged and better capable of withstanding high-voltage surges. If the amplifier is intended for use as a 2 x 250 watt stereo type with 8-ohm loudspeakers, the toroidal transformers may be rated at only 7 A each, and the total smoothing capacitance may be halved. Note that all amplifier configurations mentioned so far require **two** items of all parts as indicated in Fig. 3, including the ready-made PCB. Fitting the parts as per Fig. 7 should present few problems, but the eight TO-3 style power transistors and the heatsink require some skill in mechanics; this will be explained later on.

It is strongly advised to use first-class components of known make in **all** locations. Never use cheap, baker's dozen capacitors or resistors, and closely observe tolerance and maximum rating of each and every part before soldering it into place. Also opt for safety where the high-voltage supply rails and the amplifier output are involved.

The MOSFET power transistors are

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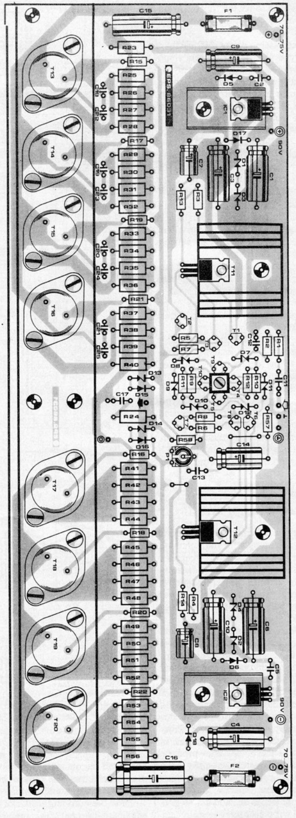
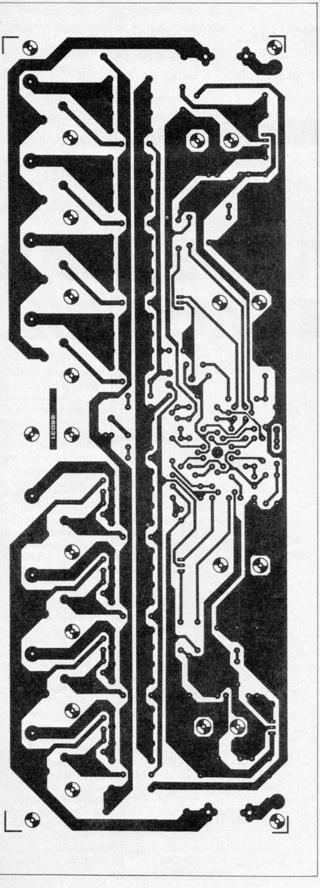




fitted onto the board last, along with a suitably drilled, 5 mm thick aluminium angled bracket; see Fig. 6 for the relevant dimensions. Do not forget to fit the transistors with good quality mica washers; ceramic (Al_2O_2) types are preferred, but more expensive and harder to get. Also remember to use a generous amount of heat conducting paste. Check for any short circuits between the transistors and the bracket once these have been bolted together.

It is strongly suggested to take ample time for a thorough inspection of all parts when they are fitted on the amplifier board; verify the correct polarization of all zener diodes and electrolytic capacitors; make sure that the NPN and PNP transistors have been fitted in the correct PCB positions. Keep in mind that any mistake, however trifling it may appear, may have costly consequences for the output stage and/or the power supply, not to mention the loudspeakers...

If everything appears to be in perfect order, proceed with bolting the amplifier board to a large heat-sink with a thermal resistance of no more than 0.3 K/W. Now consider whether you want to test the board right away, or wait until next month's issue is on your work-bench. It should be noted that testing at this stage of construction involves a number of risks, owing to the fact that the protective circuitry is not present as yet. Therefore, if you feel less sure about taking a risk, wait till next month and have the protective circuits correct any of your mistakes. When in doubt, opt for the safe way!

For an initial test, it is assumed that the amplifier board has been bolted to a heat-sink, and the ±75 V supply has been constructed in an experimental setup. Connect the +75 V to the +90 V, and the −75 V to the −90 V terminals on the amplifier board. Replace the 6.3 A fuses with 1 Ω, 4 W resistors, and solder 5K6, 1 W resistors in parallel with zener diodes D_3 and D_4. Now put the board aside for a moment and test the ±75 V supply.

Temporarily short out R_{59} and insert a 10 A anti-surge fuse in the mains line to the transformers. Make sure that the experimental setup is safe as regards the presence of the mains voltage at several points. Now switch on. Should the 10 A fuse blow, replace the wire across R_{59} with a suitably rated switch. Verify that the switch is open and apply power again. Close the switch as fast as you can; the new fuse should not blow this time. Leave the power supply on for a few minutes and measure the output voltages; these should be of

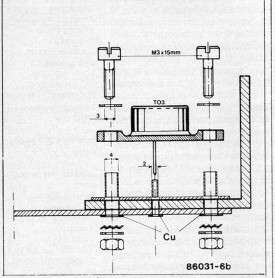
3 car-type terminals for ±75 V and earth connections
2 fuses 6.3 A anti-surge
soldering pins as required
aluminium bracket*
large heat-sink 0.3 K/W* (40 × 15 cm, e.g. Fischer SK39)
PCB 86031
8 insulating washers TO-3 style*

Parts for ±75 V power supply:

(purchase in quantity as listed)
R_{59} = 100 Ω;10 W
Tr_4;Tr_5 = toroid transformer; 55 V-15 A secondary or 2 × 28 V-15 A*
(e.g. ILP Type 9B656)
B = B200C25000;BYW64
C_{26} ... C_{31} = 10,000 μ; 100 V*

* see text and/or relevant Figure.

6a



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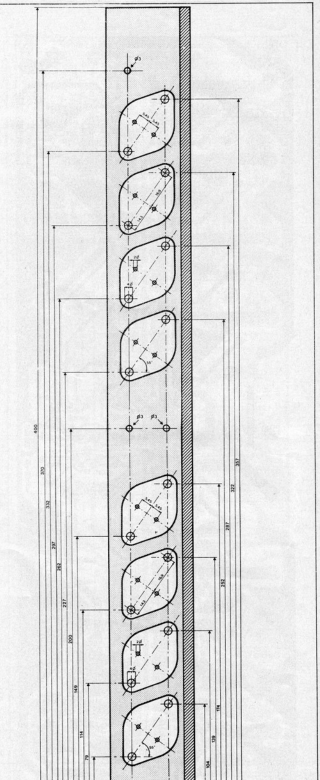

Fig. 6. Dimensional outlines of the support bracket which forms the thermal contact between transistors and heat-sink. Also shown are transistor mounting details.

the order of ±75 V to ±80 V, depending on the exact secondary voltages of the transformers in use. Switch off and slowly discharge the smoothing capacitors with a 500 Ω-10 W resistor. If applicable, set the auxiliary switch to the off position again.

Connect the supply to the relevant terminals on the amplifier PCB and turn P_1 to its minimal resistance position (fully counter clockwise). It is not necessary as yet to have a load connected to the amplifier output; hook up an oscilloscope instead. Switch on as outlined above and carefully measure the voltage drop across the fuse replacements; this should be 0 V. Slowly turn P_1 for a reading of 0,4 V across each 'fuse' to set a quiescent current of 100 mA per output transistor. Observe the measured value for a while and verify that the amplifier does not oscillate at slightly different quiescent currents; neither should there be any tendency to thermal instability. Measure the DC level at the amplifier output; this should not exceed about ±50 mV. If everything appears to be in order, a suitably rated loudspeaker may be connected to verify distortion-free amplification. Do not test for maximum power in this test setup!

Finally, replace the 1 Ω, 4 W resistors with the fuses again, remove the supply wiring, and unsolder the resistors across D_3 and D_4. The test procedure for the other amplifier board is, naturally, entirely identical to that outlined.

NOTE:
The next part of this article will be featured in our October issue.

The single-trace type of oscilloscope is definitely one of the most widespread items of measuring equipment, and as such it is generally appreciated by those who do any kind of testing or repair on (home made) audio circuitry.
However, the single-trace scope has its limitations, which are the more keenly felt when trying to compare, say, an amplifier input to an output signal. Here is an add-on design to achieve double-trace operation from that old, simple scope of yours!

SINGLE-TRACE CRT CONVERTER

The obvious advantages of having a second, simultaneously visible, channel available on an oscilloscope are likely to be so well known to any electronics enthusiast as to obviate the need for any further discussion. However, a close examination of the typical operation principles of the two-channel and dual-trace type of oscilloscope is essential to a basic understanding of the present add-on unit.

As will be generally known, the main circuits in a standard oscilloscope may be represented schematically as shown in Fig. 1. The input signal to the scope is amplified before it can deflect the cathode ray tube (CRT) electron beam in the vertical (*Y*) direction. Also the signal is used to modulate the sawtooth voltage, generated by the timebase section (horizontal or X deflection). The setup as shown allows the displaying on the CRT screen of a single trace (i.e. input signal) only.

Basically, there are two methods of simultaneously displaying two or more curves on a single CRT screen. The **dual-beam** configuration is the rarer and also the more expensive of the two, since it involves a CRT with two independent sets of *X* and *Y* deflection systems and associated electronic circuits. However easy the latter may be built, it will be readily understood that providing a single-trace CRT with an additional electron beam is definitely out of the question as a means for single-to-two-channel conversion of an existing oscilloscope. Contrary to the dual-beam type, the typical **two-channel** oscilloscope has only one CRT electron beam and, consequently, only one *X* and *Y* deflection system. The trigger and timebase sections are also single circuits; the difference with a single channel type lies in the presence of two attenuators and a fast switching channel selector, which operates at a speed, high enough to make both channels appear simultaneously and correctly positioned on the CRT screen. Obviously, such a channel switching unit may be used as a separate add-on item in conjunction with any single-channel oscilloscope to obtain the enhancement as outlined above.

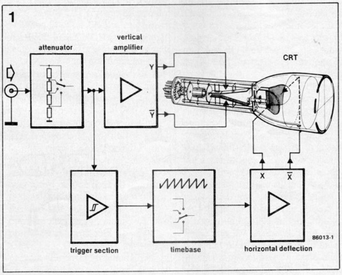

Fig. 1. Functional sections in a single-channel oscilloscope.

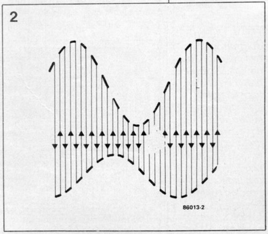

Fig. 2. The chopper mode involves very fast trace switching between the two input channels. If the timing is correct, the curves will appear as smooth and continuous to the observer.

Chopping or alternating?

Most commercially available two-channel oscilloscopes offer two modes of operation: chopping or alternating. Operation in the alternating mode is basically as follows; assuming that the electronic switch circuitry has selected channel 1, then a trigger pulse enables the scope to display the curve relevant to the signal as applied to the channel 1 input attenuator. On completion of the horizontal sweep of the luminous spot, it is arranged to return to the left of the CRT screen again, ready to be set off by the next trigger pulse. However, not only does the trigger pulse start a new horizontal sweep, it

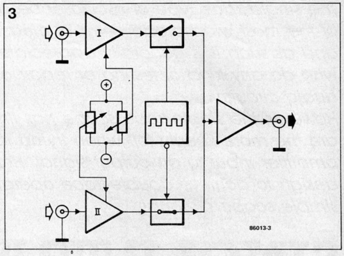

Fig. 3. Block schematic presentation of the two-channel scope add-on unit.

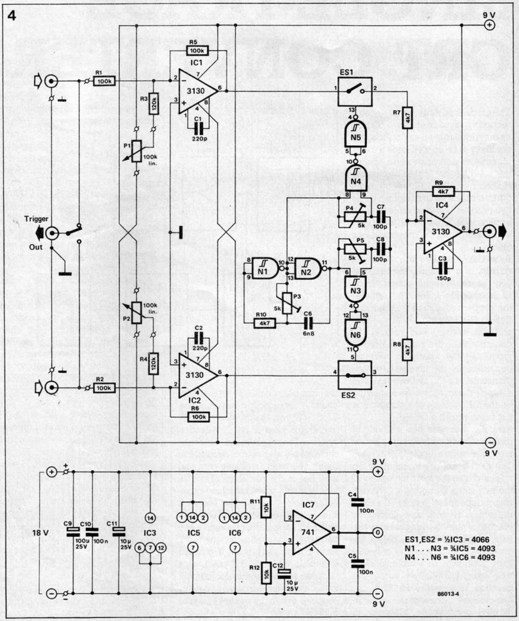

Fig. 4. Circuit diagram of the two-channel add-on unit. Note that only two of the four bilateral switches, contained in IC_3 have been used; the control inputs of the remaining two have been grounded to preclude interference caused by the chopper oscillator N_1-N_2.

also causes the electronic switch to select the other input channel for display on the CRT screen. Therefore, both channels are alternately displayed, but the mode has one distinct disadvantage, which should not be left unmentioned. If, for instance, the scope is to display two complete cycles of a 1000 Hz sinusoidal input voltage, the timebase is set to the 0.2 ms/div. range, given a screen graticule of ten by ten squares. In this setup, the travelling electron beam needs a minimum of 4 ms to display two times two complete cycles of the sine wave. The display frequency relevant to this measurement equals *1/0.004 = 250 Hz*, which is high enough to ensure a stable, flicker-free image on the CRT screen. However, a less favourable situation arises in the case of input signals in the lower than 100 Hz frequency range, since these are displayed at a frequency of 25 Hz or less, which typically causes the display to flicker to the degree of disturbing the visibility of the signal curves.

Chopper operation, on the other hand, is typically devoid of the above disadvantage, since the channel selector is controlled with a relatively high-frequency signal (several kilohertz), independent of the trigger pulse and the input signal frequency. Assuming that the chopper frequency is 50 kHz, and the signal frequency 1000 Hz, the luminous CRT spot is arranged to alternately display tiny (chopped) sections of the curves on both channels; the principle is illustrated in Fig. 2, which shows that the displayed waveforms are, in fact, chopped into some 50 sections each. The switching rate of the CRT beam is so high as to make the gaps in the curves invisible to the human eye; the curves, therefore, appear as smooth and continuously present. If the chopper frequency is well in excess of the signal frequency, as in the above example (50 to 1 ratio), this oscilloscope display mode ensures stable, flicker-free visibility of the applied signals on the CRT screen. In case the input signal frequency exceeds that of the chopper section to the extent of resulting in a ratio of, say, 6 to 1, the situation that ensures is not necessarily dramatic as yet, since the curves on both channels are each displayed three times over. Problems are only anticipated in case the chopper and signal frequencies are either about equal or in some fixed relation to one another; the resulting effect on the CRT screen is comparable to that outlined above in the section on the alternating mode. However, the solution to the problem is relatively simple in this case, since the chopper frequency may conveniently be made variable; in case of display instability, the chopper oscillator is slightly detuned.

The circuit

The block diagram of Fig. 3 aims at offering an insight into the basic operation of the present scope add-on unit. Two input amplifier sections, each with a vertical trace positioning preset, pass the signals to two electronic switches, which are antiphase controlled by a central chopper oscillator section.

All of the above functional blocks can be seen in Fig. 4, the circuit diagram of the add-on unit. At the left are two identical, fast opamps Type CA3130, which amplify the input signals to both channels. Presets P_1 and P_2 are the trace positioning controls; they elevate the AC signal to a certain DC level in order to obtain the correct vertical position of each trace on the CRT screen. Electronic switches ES_1 and ES_2 are contained in a Type 4066 CMOS quad bilateral switch IC. To prevent the input capacitance of the oscilloscope from delaying the steep edges of the chopper signal — this would make them visible on the screen —, IC_4 has been incorporated as a fast output buffer opamp. The chopper oscillator is a conventional design using Schmitt-trigger NAND gates; P_3 provides the tuning control. The necessary phase difference between the output control signals is realized by taking them from the input and the output of N_2. The expected frequency range of the proposed setup should be about 50 to 100 kHz. Gates N_4-N_5 and N_3-N_6 prevent the switching moments of ES_1 and ES_1 from coinciding. Finally, IC_7 creates a virtual earth level in order to enable the circuit to work off a single 18 V supply.

Construction, adjustment and use

In order to preclude undesirable spurious radiation caused by the chopper oscillator from manifesting itself in domestic receiving equipment, the present add-on unit should be fitted in a suitably dimensioned metal enclosure.

After connection of the completed board to the oscilloscope, P_1 and P_2 are adjusted to obtain the correct trace position for each channel on the CRT. Now adjust P_3 to obtain a stable display of the chopper switch signal with the oscilloscope timebase set to 10 μs/div. Presets P_4 and P_5 may now be adjusted for maximum edge steepness of the chopper signal, i.e. it should, ideally, become invisible on the screen. This completes the necessary adjustments.

The use in practice of the present add-on unit is, of course, subject to the limitations brought about by the relative simplicity of the proposed circuit. Given the absence of input attenuator sections, the measured voltages should not exceed 12 V peak-to-peak (4.3 V_{rms}). The use of opamps in the circuit inevitably limits the attainable bandwidth to several hundred kilohertz, but this need not be a drawback if the user mainly intends to measure audio signals. Should the chopper frequency become visible on the screen, then P_3 may be set to a slightly different position to make the signal edges invisible again.

Finally, the present design does not incorporate a power supply; the user must either avail himself of an existing mains supply, or construct a separate unit to this end, capable of delivering 18 V at about 50 mA. Also note that no ready-made PCB exists for this project; the true scale track layout, however, is given in *Make your own PCBs*, elsewhere in this issue, while the component mounting plan is given in Fig. 5. *KD;TS*

Parts list

Resistors:
R_1;R_2;R_5;R_6 = 100 k
R_3;R_4 = 120 k
R_7. . . R_{10} = 4k7
R_{11};R_{12} = 10 k
P_1;P_2 = 100 k linear potentiometer
P_3. . . P_5 = 5 k preset

Capacitors:
C_1;C_2 = 220 p
C_3 = 150 p
C_4;C_5;C_{10} = 100 n
C_6 = 6n8
C_7;C_8 = 100 p
C_9 = 100 μ;25 V; electrolytic
C_{11};C_{12} = 10 μ;25 V; electrolytic

Semiconductors:
IC_1;IC_2;IC_4 = CA3130
IC_3 = 4066
IC_5;IC_6 = 4093
IC_7 = 741

Miscellaneous:
S_1 = single-pole toggle switch
2 knobs for P_1 and P_2
4 sockets for inputs and outputs
metal enclosure
PCB 86013 (not available through Readers Services)
suitable power supply; 18 V; 50 mA regulated

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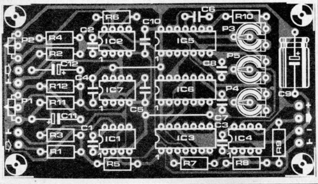

CCD VIDEO MEMORY SYSTEMS

It seems fairly certain that over the next few years more and more video systems will incorporate picture frame memories. With these, the picture quality of monitors and television receivers can be improved, while at the same time the way is opened for a host of new features.

Video memories are used in satellite receivers; in medical scanners; in material testing by infra-red, supersonic, and X-ray techniques; in astronomy and photography; and, last but not least, security equipment. Such memories are, in the main, dynamic RAMs.

CCD (charge-coupled device) memories are inherently slower than RAMs, but also cheaper and more compact. This makes them suitable for applications that are either serial in nature or that do not require the fast operating speeds of RAMs. Now that digital signal processing is used in modern TV receivers, video memories can be incorporated to offer a number of new operational aspects.

- Improved picture quality through more effective noise suppression, greater freedom from flicker, and better colour separation.
- Picture freeze facility, and the possibility of conveying such pictures over telephone networks.

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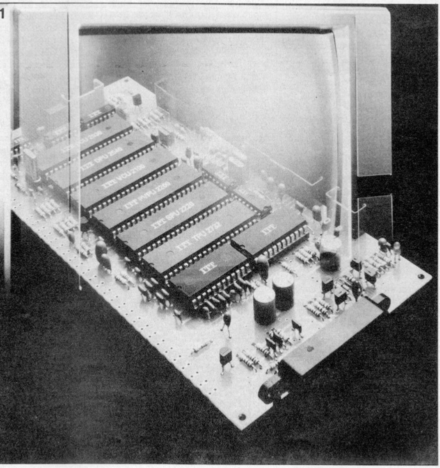
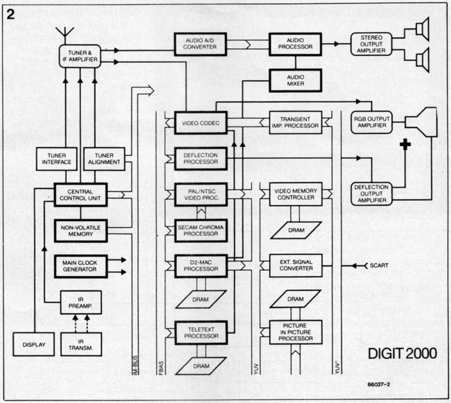
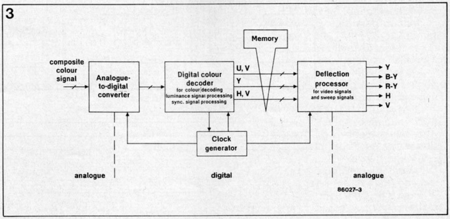


Fig. 1. Digit 2000 prototype board offering complete digital processing of video, audio, and teletext signals. (Photo courtesy of ITT)

Fig. 2. Block schematic of ITT's Digit 2000 digital colour television receiver.

Fig. 3. Illustrating the principle of Valvo's video signal processing. The clock generator is synchronized with the line time-base generator.





- Superimposition of pictures on one another.
- Zoom-in facility.
- Teletext storage with instant access.

It would also be possible to use the video memory in conjunction with a video cassette recorder and microcomputer to obtain an editing facility.

Digital television techniques

Since the early 1980s, a number of semiconductor manufacturers have introduced digital video signal processing devices. International Telephone and Telegraph Corporation—ITT—was the first to put such a device into standard production (in 1983). This Digit 2000 offers complete picture, sound, and teletext processing and is already used in hundreds of thousands TV receivers.

Valvo, in conjunction with Philips and Siemens, have developed another system that is now being used in a number of TV receivers under development.

The main difference between the two approaches lies in the choice of scanning frequency. ITT links the clock frequency to that of the chrominance subcarriers, whereas in the Philips/Valvo/Siemens system the scanning frequency is synchronous with the line frequency.

In the line-based concept the video memory is organized on the basis of picture build-up. This makes it possible for additional signal processing to be carried out by including adjoining picture elements in successive rasters. In this technique, use is made of specially designed CCD memories in which the data is stored line by line. ITT prefers standard RAMs as video memories. Although these are more expensive than CCDs, fewer of them are required: the ITT system requires five 256 K DRAMs, whereas the Philips/Valvo/Siemens set-up needs seven 317 K CCDs.

DRAM system

ITT has had a TV receiver

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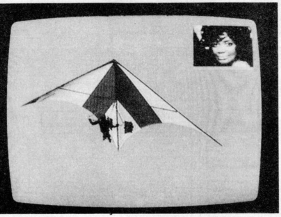

(Digivision) with a 12 Kbyte RAM in production for just over a year. This enables two video signal sources to be displayed simultaneously on the same screen. The video signal to be faded in is taken from one of the SCART connectors via a single chip PAL decoder. The RGB signals at the output of the decoder are converted in a multiplex process by a single digitizer at a scanning rate of 1.5 MHz. A 4:1 data reduction results from the simple process of reading only every fourth line from the RAM that synchronizes the pictures. Because of the small format of the superimposed picture, it is sufficient to store just one raster. This requires only 4 Kbyte per colour, making a total of 12 Kbyte.

Control of the memory as well as addressing the RAM is carried out by two gate arrays, which replace no fewer than thirty standard ICs.

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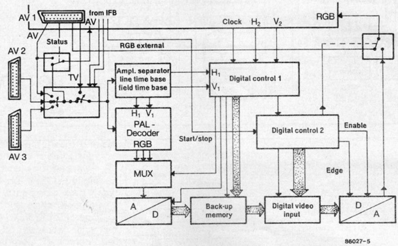

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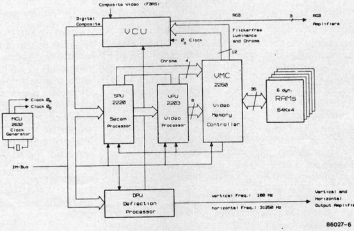

Fig. 4. Photograph showing the display of two different video signals onto the screen of an ITT Digivision TV receiver. The secondary picture is identified by a coloured band at its lower edge. (Courtesy of ITT).*

Fig. 5. Block schematic of an ITT Digivision chassis containing a 12 Kbyte video RAM.

Fig. 6. ITT's Type VMC2260 Video Memory Controller drives a video memory consisting of five 256 K DRAMs. Thanks to the doubling of the frame frequency, the picture is virtually free of any flicker.

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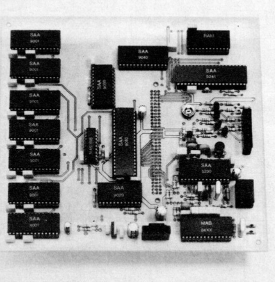

Fig. 7. Prototype CCD video memory board from Philips. Next to the seven Type SAA9001 CCD memories (left-hand side) are four control ICs which provide a number of features, such as still picture, noise reduction, and recall picture.

Fig. 8. Block schematic showing how the various special features are obtained in the Valvo system.

Fig. 9. Block schematic of the Type SAA9001 CCD memory.

Fig. 10. The SAA9001 is arranged in 294 lines of 1080 bits each.

However, gate arrays are not suitable for a complete video memory with five 256 K DRAMs. For that purpose, ITT has developed a special video memory controller, the Type VMC2260. Apart from doubling the frame frequency to 100 Hz, this device also provides the still picture, zoom, superimposition, and teletext memory facilities.

CCD technique

In the Philips/Valvo/Siemens system, the analogue video signal is converted into 7-bit digital words synchronous with the line frequency. The clock frequency of 13.5 MHz results in a fixed scanning rate of the luminance signal (Y-signal) of 720 samples per line. Because of their limited bandwidth, the chrominance signals (U and V signals) at the output of the decoder, however, are scanned at only 3.375 MHz, i.e. 180 samples per line. All together there are, therefore, 720+2×180=1080 samples per line, resulting in a frequency of the multiplexed signal (Y+U+V) of 20.25 MHz.
The video memory, built up in accordance with the scanned frame structure, is based on CCD Type SAA9001. In this device, 317 Kbit can be contained on a small crystal surface, arranged in 294 lines of 1080 bits each.
The visible part of a normal raster (two rasters constitute a complete picture or frame) in the 625-lines-per-frame system is composed of 288 fifty-two-microsecond lines. The SAA9001 is, therefore, able to store a complete raster with one bit per sample. The relevant 1080-bit line of the SAA9001 receives 720 luminance bits and 2×180 chrominance bits from each of the 720 scanned pixels in a raster line. Since each scanned pixel results in seven bits, the memory consists of

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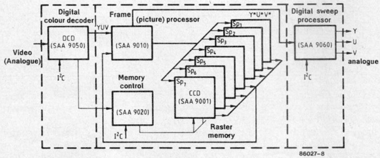

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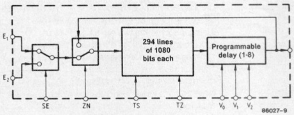

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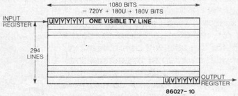


seven CCDs, which together store the information pertaining to 720×288=207 360 pixels. In contrast to other CCD memories, the SAA9001 uses serial-parallel-serial transfer of information. In this method of operation, the first 1080-bit data line is input serially to the first row in the array at high speed. When this row is filled, the bits are transferred in parallel at a slower rate, while the next data line is input. Successive lines are thus transferred through the array. After 294 data lines have been input, the first of them appears at the output from where it is transmitted serially at high speed.
The line shifts in the memory are synchronous with the line frequency. Higher line or field frequencies are not yet planned in this concept. None the less, it offers these features:

- cross-colour reduction;
- noise reduction on noisy signals (particularly from video recorders);
- store and recall picture during normal TV operation;
- picture freeze during normal TV operation;
- storage for up to 252 teletext pages with instant access.

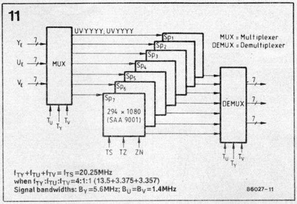


Fig. 11. Multiplex structure of a CCD video memory containing seven Type SAA9001 devices.

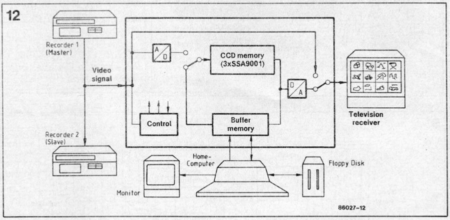


Fig. 12. Possible set-up of a video editing aid using a computer and a CCD video memory.

Other features are possible, but their incorporation will depend largely on consumer demand.
The SAA9001 is also an interesting memory device for other than television applications. Since it has only three control inputs, its use is straightforward and allows the construction of digital video and audio memory units at relatively low cost. Its facility for accessing parts of a video picture via a computer should be of interest to microcomputer and television (slow scan TV) experts and amateurs alike.
Interested readers are also referred to *The Accordion Image Sensor (EE India,* March 1986) EK

NEW PRODUCTS

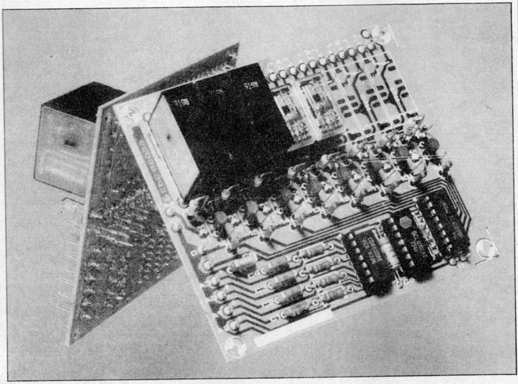
New wideband opamp

National Semiconductor Corporation have recently introduced a wideband, FET-input operational amplifier that can provide 100 mA continuous output current. Designated the Type LH4101, the chip eliminates the need for a buffer to provide the additional current drive not available with other wideband opamps.
The Type LH4101 provides internal compensation for unity gain stability and all the internal gain set resistors for most popular gain settings; also of interest are its 45 MHz bandwidth and capability to drive 50 ohm loads directly.
The new part, as compensated, is claimed to represent an optimum compromise between slew rate, bandwidth, settling time, and gain linearity, at the same time replacing compensation and bypass capacitors, and gain set resistors.
Applications of the hybrid opamp include video distribution, summing amplifiers, fast sample and hold circuits and speed integrator circuits.
The Type LH4101 is the first in a series of opamps from National Semiconductor that will be combining internal bypassing compensation and providing all external components normally found in high speed opamp configurations, and it is currently available in a 24-pin dual-in-line plastic (DIP) package.
National Semiconductor (UK) Limited
301 Harpur Centre
Horne Lane
Bedford MK40 1TR
Telephone: (0234) 47147
Telex: 826209 (3459:12)

EIGHT-WAY RELAY BOARD

by
P C M Verhoosel

Whatever they say, don't believe that computer interfacing is within reach of the average owner of a personal micro equipped with a parallel output port. Always remember that the way from CPU accu to, say, automatic control relays is a mighty long one, and stick to these beliefs until you have constructed this universal board.



Despite its heavy accent on versatility and compatibility with any type of computer having a parallel output port, the present relay controller board comprises only very few components, as can be seen from the circuit diagram shown in Fig. 1. No LSI chips, dedicated I/O controllers, or handshaking hardware; the proposed relay controller along with a few BASIC instructions puts you in control of any of eight DIL type (reed) relays, merely using four databits from the computer's parallel output port.

A self-strobing decoder

IC_3 is a Type 4099 CMOS 8-bit addressable latch which can pass the logic level at the D (data) input to one of eight outputs selected by the combination of bits at the $A_ø$, A_1 and A_2 inputs; latching of databit and address takes place when the enable ($\overline{E}$) input is pulled low. In addition, the Type 4099 has a RESET input to clear the internal latch and pull all chip outputs low.

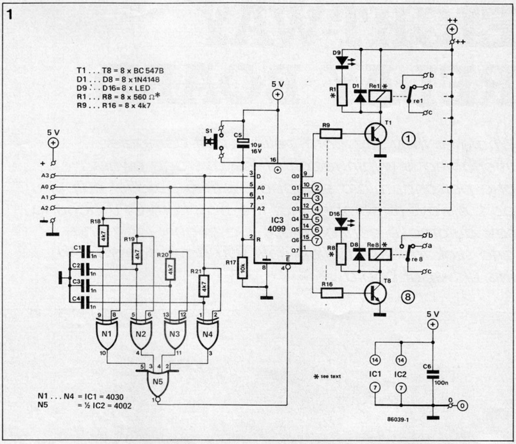
Table 1. Boolean algebra functions of the XOR and NOR type of logic gate.

Table 1.

Exclusive OR		
input 1	input 2	output
Ø	1	1
1	Ø	1
1	1	Ø
Ø	Ø	Ø

4-input NOR				
input 1	input 2	input 3	input 4	output
Ø	Ø	Ø	1	Ø
Ø	Ø	1	Ø	Ø
'	'	'	'	'
'	'	'	'	'
'	'	'	'	'
1	1	1	1	Ø
Ø	Ø	Ø	Ø	1

Fig. 1. Circuit diagram of the universal relay controller board. Several types of DIL relay may be accommodated as explained in the text.



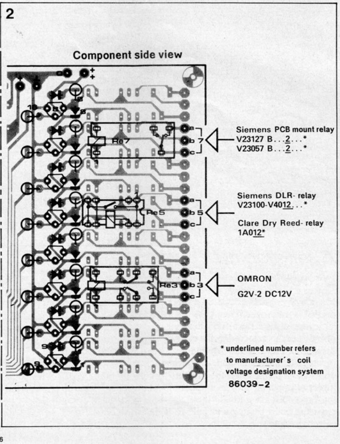


Fig. 2. Showing how different types of PCB-mount relays may be fitted onto the board.

Since the circuit is to be suitable for connection to any computer having a parallel output port, a means other than any kind of output strobe pulse had to be devised for clocking the latch, since many computer manufacturers do not even seem to be bothered by, say, the rules laid down in the Centronics standard.

The present circuit therefore needs no computer generated strobe pulse; it provides its own whenever data is written to the relevant four bits comprised in the output port data.

Table 1 shows that the output of an exclusive-OR (XOR) gate does not go high until the logic levels at its inputs are complementary. IC_1, a Type 4030, contains four XOR gates, each of which has one input driven direct by an address bit $A_0 \ldots A_3$, while the other input receives the same level, but slightly delayed by a R-C network. Therefore, every logic change on any of the $A_0 \ldots A_3$ lines causes the relevant capacitor to be either charged or discharged over the associated resistor, providing a short-duration complementary pulse combination at the XOR gate inputs, which fact causes the gate to produce a high level pulse at its output. Quad input NOR gate Type 4002 receives the output levels of the four XOR gates, executes the the logic function as per Table 1, and supplies IC_3 with an $\bar{E}$ pulse, which causes the databit on A_3 and the relevant channel (relay) number to be strobed into the device, which activates or deactivates the corresponding output $Q_0 \ldots Q_7$, each driving a transistor with a relay and associated indicator LED connected in the collector supply line.

Figure 2 shows how a number of relays by different manufacturers may be fitted onto the board. For types not listed, you may use the spare holes, but check the internal configuration as well as the coil resistance and voltage before using any unlisted type of PCB-mount relay.

Supply voltages

The relay board requires two regulated supply rails; one of +5 V for the CMOS ICs, and another, +12 V, for the relay coils and driver transistors. The latter supply should be connected to point ++ on the ready-made PCB.

The circuit as shown in Fig. 1 has been designed for the incorporation of relays with a 12 V DC coil voltage, but differently rated types may also be used, provided the LED series resistors are dimensioned according to

$$R_{1\text{-}8}=(V_{coil}-V_{LED})/I_{LED} \qquad \langle\Omega\rangle$$

Since the circuit as shown incorporates 12 V type relays and LEDs which draw 20 mA at 2 V, the given resistor value of 560 Ω is accounted for by

$$R_{1\text{-}8}=(12-2)/0.02=500\ \Omega,$$

$R_{1\text{-}8}$ having the next higher value in the E12 series.

Construction

It is suggested to start the construction with fitting the IC sockets and soldering pins, followed by the remaining passive components (Fig. 4). Note that R_1 to R_{16} and protective diodes D_1 to D_8 are fitted vertically to save board space.

The LEDs may be mounted either at the soldering or the component side of the PCB, depending on the type of enclosure you have in mind for the project. Reset switch S_1 is connected to a pair of soldering pins, using two short wires.

Despite the tempting presence of soldering pins for the supply wires to other equipment, it must be strongly advised not to have the relay contacts switch or carry currents or voltages in excess of the manufacturer's specifications, since doing so may cause the PCB tracks to burn out after the relay and possibly the driver transistor have been destroyed internally.

Practical use

Users of the well-known Commodore C64 computer may readily wire the present relay controller board to a parallel output port, whether this is a DIY or ready-made type. The program listing shown in Fig. 3 is intended as an initial test to verify the correct function of the relay board.

Owners of other types of computer having a parallel output port may refer to Table 2 to find the relevant bit combination for each relay as well as the code to turn it on and off (effected with A_3).

Finally, the Reset switch may be pushed at any time while in the process of writing and debugging relay control subroutines, which, as any serious programmer will admit, is usually by way of trial and error as well as frequently occurring computer hangups.

HS;GK

Table 2.

Relay	A2	A1	A0	on*	off*
1	0	0	0	X8	X0
2	0	0	1	X9	X1
3	0	1	0	XA	X2
4	0	1	1	XB	X3
5	1	0	0	XC	X4
6	1	0	1	XD	X5
7	1	1	0	XE	X6
8	1	1	1	XF	X7

*In hexadecimal notation and assuming that $A_0 \ldots A_3$ are connected to port bits $D_0 \ldots D_3$ in that order.

Table 2. Summary of relay addresses and output port data codes.

3

```
10 POKE 56579,15: REM P0 to P3 are outputs
20 POKE 56577,0: REM soft reset for relay board
25 FOR I=0 TO 7: R(I)=0: NEXT
30 INPUT "WHICH RELAY";R$
35 IF VAL(R$) < 1 OR VAL(R$) > 8 THEN 30
40 I=VAL(R$)-1
50 IF R(I)=1 THEN R(I)=0: GOTO 60
55 R(I)=1
60 POKE 56577,I+8*R(I)
70 GOTO 30
```

Fig. 3. This short programme may be keyed in to test the relay board as an extension to the C64 computer's parallel output port.

4

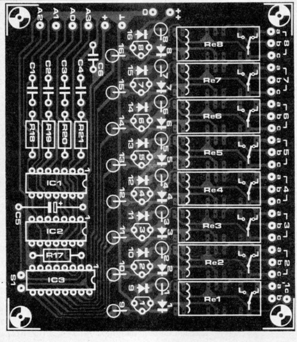


Fig. 4. Track layout and component overlay for the relay controller board.

Parts list

Resistors:
$R_1 \ldots R_8$ = 560 Ω*
$R_9 \ldots R_{16}$; $R_{18} \ldots R_{21}$ = 4k7
R_{17} = 10 k

Capacitors:
$C_1 \ldots C_4$ = 1 n
C_5 = 10 μ; 16 V electrolytic
C_6 = 100 n

Semiconductors:
$D_1 \ldots D_8$ = 1N4148
$D_8 \ldots D_{16}$ = LED
$T_1 \ldots T_8$ = BC547B
IC_1 = 4030
IC_2 = 4002
IC_3 = 4099

Miscellaneous:
S_1 = push to make button
$Re_1 \ldots Re_8$ = PCB mount DIL relay*
2 off 14-way IC sockets
1 off 16-way IC socket
34 off soldering pins
PCB 86039
Suitable enclosure
Sockets for relays, if required
Sockets for computer and relay connections

*see text and/or relevant Figure.

In actual fact, it is not always the fault of (amateur)transmitters that they cause interference on TV sets. As a rule, it is the 'broad-band aerial amplifier' included in the TV set's aerial system which is at the root of the problem. Broad-band amplifiers have the disadvantage of being rather indiscriminate. They pick up and amplify everything, including signals which are not meant for them at all. When powerful broadcast, amateur or mobile transmitters are around, the voltage in the aerial amplifier rises to such an extent that the amplifier becomes completely 'jammed' and this makes a clear reception of TV signals very difficult.

TV interference suppression

Nearly everybody will agree that interference on TV can be extremely annoying. Interference can be caused, among other things, by local transmitters. Usually, however, this can be dealt with in a fairly simple and effective way.

So what do you do? Well, after reading the above, it would seem an obvious conclusion that it is probably better to do without an aerial amplifier altogether. For that matter, very often one is included in the aerial system 'just to be on the safe side', without it being strictly necessary.
It is a much better (and cheaper!) idea to simply use a good TV aerial which is a powerful 'amplifier' anyway (and will have a more accurate directional effect and an improved front-back ratio — both important factors). If, on the other hand, you cannot manage without an amplifier, it is advisable to use tuned aerial amplifiers (also known as channel amplifiers). These, being narrow-band, do not pick up unnecessary signals and so interference is no longer a problem.
However, if you already have an aerial system which is fitted with a broad-band amplifier, it is rather frustrating to talk about the kind of aerial you should really have.
Quite a few interference problems can be dealt with in an inexpensive way by simply inserting a band-stop filter in the broad-band amplifier's input. This eliminates the interfering signal (produced by an amateur transmitter, for example) before it reaches the broad-band amplifier. The so-called ¼ λ-filter is a good choice: it is easy to make — all you need is a piece of coax cable!

The ¼ λ-filter

Figure 1 shows what the filter looks like. In passing, it should be noted that this filter can be used for all kinds of purposes — not only eliminating interference in broad-band amplifiers!
As the drawing demonstrates, the (coax) aerial cable, leading from the aerial to the broad-band amplifier, is stripped at a certain point and connected to one end of a piece of coax. This coax, believe it or not, is the filter. It should be exactly ¼ wave length of the signal that is to be eliminated. The other end of this piece of coax, which is known as ¼ λ (quarter-lambda) stub, remains open. This is how it works:
Radio waves reaching the open end of the ¼ λ stub are reflected. For the unwanted signal, the stub is exactly ¼ λ long, so that the reflected waves have travelled a distance of 2 x ¼ λ = ½ λ by the time they get back to the beginning of the stub. Consequently, the reflected wave is in exact phase-opposition with

1

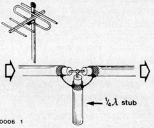
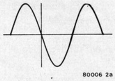


Figure 1. The filter is a piece of coax, connected in the lead from the aerial to the broad-band aerial amplifier. In practice, it is often best to connect the ¼ λ stub at the input of the amplifier.

2a



2b

2c

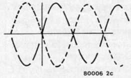
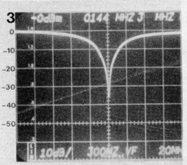


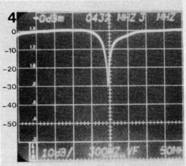
Figure 2. The filter works as follows: The voltage reflected in the stub (2b) is in exact anti-phase to the input voltage (2a), so that the resulting voltage (2c) is nil.

the input signal, so that the resulting voltage is nil. This is illustrated in figure 2. Figure 2a shows the input voltage, figure 2b shows the reflected voltage and figure 2c gives the result.
Everything always sounds marvellous in theory, but often turns out differently in practice. Here too, unfortunately this is the case. What happens is that the ¼ λ stub attenuates the reflected wave, so that the resulting voltage is not completely nil, as shown so optimistically in figure 2c. It doesn't have to be! A reduction by about 30 dB (32 times) is usually achieved with the aid of the filter and nine times out of ten that is enough. Furthermore, the filter not only blocks interference on the wave length which is four times as long as the ¼ λ stub, but it also works for wave lengths corresponding to ¾ λ, 5/4 λ, 7/4 λ etc. The input signal and the reflected wave are in anti-phase at these frequencies as well!



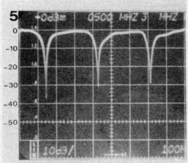
80006 3

Figure 3. A spectrum-analyser photo of a coax ¼ λ-filter for the 2-metre band. The attenuation is approximately 36 dB.



80006 4

Figure 4. The rejection filter intended for the 2-metre band can also be used for the 70-centimetre band, with marginally poorer results.



80006 5

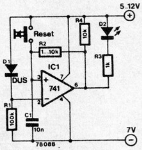
Figure 5. A spectrum-analyser picture over a much wider frequency range (100 MHz per division) shows that there are many more frequencies at which the input signal and the signal reflected by the filter are in anti-phase.

In practice

As far as the exact length of the filter is concerned, simple theory is one thing, practice another. The speed at which radio waves travel along coax is not the same as that in air. For this reason, the wave length inside the cable is shorter than that outside: a radio wave may have a wave length of 3 ft. outside and as little as 2 ft. inside the coax cable. The reduction factor, in that case, is: ⅔ = 0.67.
Let us consider a rejection filter for a 2-metre amateur transmitter. Amateur transmitters on the two-metre and 70-centimetre bands seem to be prime targets for complaints about interference. On the two-metre band ¼ λ corresponds to ¼ x 2 = 0.5 metres. In order to find out what the exact length of the ¼ λ stub should be, this figure must be multiplied by the reduction factor of the coax. Every manufacturer (and reliable retailer) will be able to supply this information. It is advisable to make the cable slightly longer than the calculated length, so that once the stub has been connected, it can be trimmed for maximum suppression of the interfering signal. This can be done by cutting off small bits at a time. When you have found the correct length, the ¼ λ stub can be rolled up. It looks neater, that way.
One of the characteristics of this type of filter, as mentioned earlier, is that it will eliminate several frequencies. This can be an advantage: a filter for the 2-metre band can be used for signals on the 70-centimetre band as well. The spectrum-analyser photo's (figures 3 and 4) illustrate this. Figure 3 shows how the filter attenuates interference at the frequency for which it was originally intended: 144 MHz (the 2-metre band). Figure 4 illustrates the effect at 432 MHz (70-centimetre band).
Since the damping of the coax cable is greater at higher frequencies, the attenuation achieved is less than that at 144 MHz. As the photo's illustrate, the difference is approximately 6 dB. The spectrum-analyser photograph in figure 5 gives an idea of the attenuation over the whole frequency range (horizontally 100 MHz per division). ◀

supply failure indicator

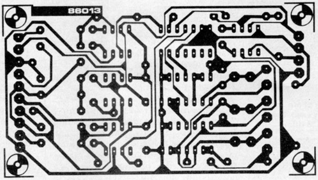
Many circuits, especially digital systems such as random access memories and digital clocks, must have a continuous power supply to ensure correct operation. If the supply to a RAM is interrupted then the stored information is lost, as is the time in the case of a digital clock.
The supply failure indicator described here will sense the interruption of the power supply and will light a LED when the supply is restored, thus informing the microprocessor user that the information stored in RAM is garbage and must be re-entered, and telling the digital clock owner that his clock must be reset to the correct time.



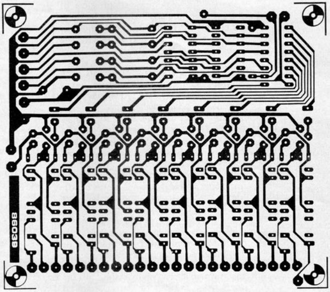
When the supply is initially switched on the inverting input of IC1 is held at 0.6 V below positive supply by D1. Pressing the reset button takes the non-inverting input of IC1 to positive supply potential, so the output of IC1 swings high, holding the non-inverting input high even when the reset button is released. LED D2 is therefore not lit.
When the supply is interrupted all voltages, of course, fall to zero. Upon restoration of the supply the inverting input of IC1 is immediately pulled up to its previous potential via D1. However, C1 is uncharged and holds the non-inverting input low, so the output of IC1 remains low and D2 lights.

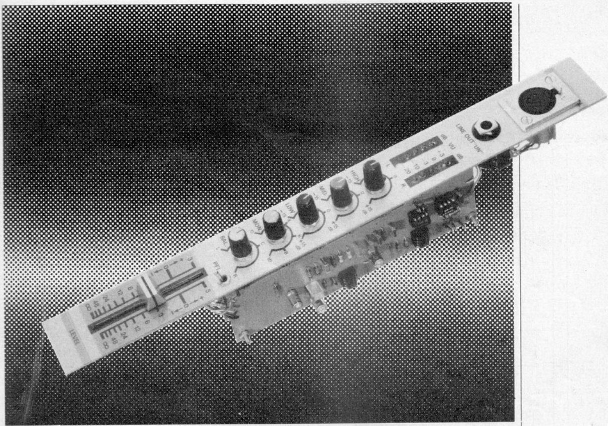
PCB track patterns for
Single-trace CRT converter
8-way relay board

86013
Single-trace CRT converter



86039
8-way relay board





by A Schmeets

PORTABLE MIXER — 2

This second part in the series continues with the construction of the first output module, which incorporates tone controls, an output level indication and a balanced as well as an unbalanced line output.

As explained in the preceding article (*Elektor India*, June 1986), there are two output modules to the portable mixer: one for general usage, and one for more specific applications, incorporating a monitor and effects amplifier, as well as a parametric equalizer section. The former is described below, whereas the latter will be discussed in next month's issue.

The circuit

The circuit diagram of the first output module is shown in Fig. 1. Opamps A_1 and A_1' are summing amplifiers for the left and right channel respectively. The active tone control section of each channel consists of a number of R-C networks in conjunction with an operational amplifier. Note that the tone control potentiometers are stereo types to ensure identical and simultaneous tone setting on both channels. The HP_L and HP_R signals, as well as the mono MONITOR line (P_4), go to the relevant inputs of the second output module, to be described next time. P_6 is the balance control and P_5 the master output slide potentiometer. Provision has been made to connect the LINE output signal to the PFL section by

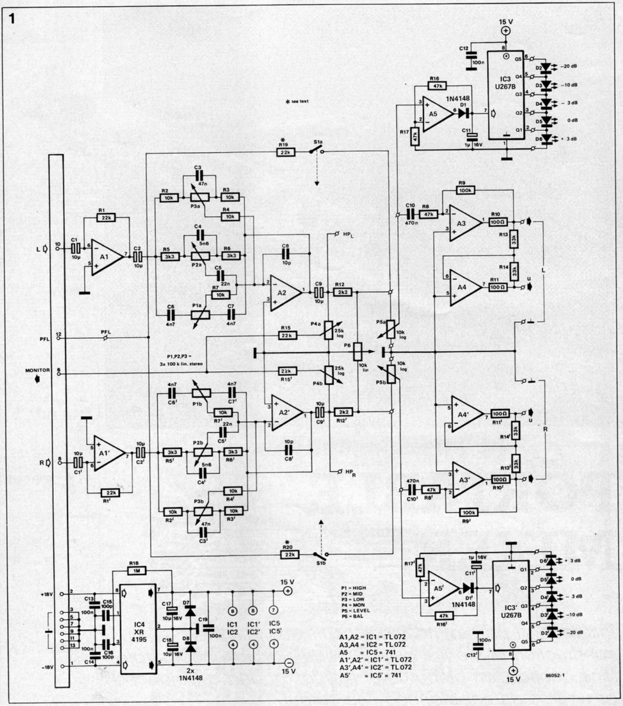


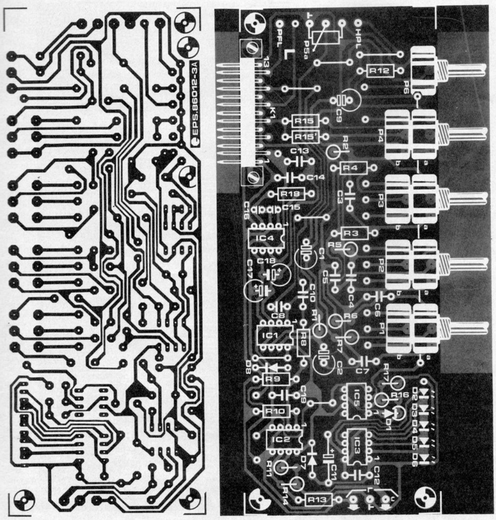
Fig. 1. Circuit diagram of output module 1, which has a 3-way tone control, balanced and unbalanced outputs, and a LED VU meter for each output channel.

means of R_{19}, R_{20}, and S_1.
The LED VU section for each output channel consists of an opamp-diode combination (A_5-D_1) which rectifies the signal level at the wiper of the master fader. The variable DC level is next applied to a special LED VU driver, IC_3. The division in five output signal levels is sufficiently accurate for most purposes; 0dB corresponds to about $1V_{rms}$.
LINE output amplifiers A_3 and A_3' also receive their input signals from the wipers of the master fader. The unbalanced mixer output signal is available at the U terminals. Two additional opamps, A_4 and A_4', provide balanced output signals, which are available across high-stability (1%) resistors R_{13}-R_{14} (L) and R_{13}'-R_{14}' (R).

Construction

Output module number 1 is fitted on the PCB shown in Fig. 2. The sandwich construction of the completed module should be familiar from the photographs illustrating the first part of this series of articles. The only parts common to both PCBs are the stereo potentiometers and the 13-way PCB connector. The compactness of the unit necessitates vertical mounting of some resistors and capacitors; the terminals of $D_2 \ldots D_6$ and $D_2' \ldots D_6'$ should be bent to suit the protruding LED positions in the front panel, which is made to the outlines given in Fig. 3. Fitting the output sockets, the potentiometer spindles and the PFL switch should present

2



Parts list

Resistors:
R1;R1';R15;
R15';R19*;R20* = 22k
R2;R2';R3;R3';
R4;R4';R7;R7' = 10k
R5;R5';R6;R6' = 3k3
R8;R8';R16;R16';
R17;R17' = 47k
R9;R9' = 100k
R10;R10';R11;R11' = 100Ω
R12;R12' = 2k2
R13;R13';
R14;R14' = 33k;1%
R18 = 1M
P1;P2;P3 = 100k stereo linear potentiometer +
P4 = 25k stereo log potentiometer +
P5 = 10k log stereo slide potentiometer; 58mm travel*
P6 = 10k linear potentiometer +

* not mounted on PCB
\+ with 4mm spindle for PCB mounting

Capacitors:
C1;C1';C2;C2';
C9;C9' = 10μ;
40V bipolar electrolytic
C3;C3' = 47n
C4;C4' = 5n6
C5;C5' = 22n
C6;C6';C7;C7' = 4n7
C8;C8' = 10p
C10;C10' = 470n
C11;C11' = 1μ;16V;
electrolytic
C12;C12';C13;
C14;C19 = 100n
C15;C16 = 100p
C17;C18 = 10μ;16V;
electrolytic

Semiconductors:
D1;D1';D7;D8 = 1N4148
D2;D2';D3;D3';D4;
D4' = 3mm LED;green
D5;D5' = 3mm LED;
amber
D6;D6' = 3mm LED;red
IC1;IC1';IC2;IC2' = TL072
IC3;IC3' = U267B*
(AEG-Telefunken)
see Table 1
IC4;IC4' = XR4195*
IC5;IC5' = 741

* see text

Miscellaneous:
S1 = double miniature switch
6.3mm cinch-type socket (stereo)
13-way PCB-type connector to DIN41617
5-way XLR Cannon type socket
PCB 86012-3A;3B*
knobs for potentiometers as required
front panel foil 86012-3F*

2

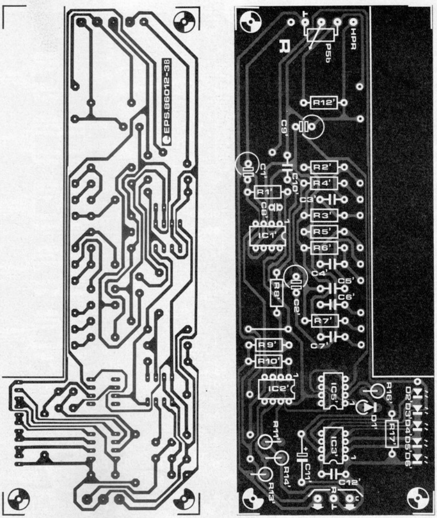


Fig. 2. Track layout and component mounting plans for the output module PCBs.

few problems after a careful look at the relevant drawings and the photograph in this article.

Finally, the on-board voltage regulator, IC_4, may be replaced by regulators Types 79L15 and 78L15 as explained and illustrated in last month's article.

Modules and amplification

Below are a number of useful hints to obtain the correct total amplification of the modules as described so far. Where necessary, some resistor and/or capacitor values may have to be adapted to suit the individual signal levels of equipment connected to the portable mixer.

MIC/LINE-module: the amplification of output opamp A_3 depends on the ratio $R_9/((R_7\text{-}P_1)+R_6)$ so that any of these resistive elements may be given a different value to obtain the desired total amplification. Note, however, that $R_9=R_8$ and $R_{12}/R_{10}=R_{13}/R_{11}$. Alternatively, R_{10} and R_{11} may be changed, but these should keep identical values. The resistance is inversely proportional to the resulting total amplification.

Stereo input module: the amplification of the MD preamplifier is arranged at 35dB at 1kHz. R_3 and R_3' may be given different values; amplification is inversely proportional to the value of these resistors. C_3 and C_3', however, should also be changed in inverse proportion to R_3 and R_3' to ensure the correct cut-off frequency of the preamplifier; the lower the value of R_3, the higher that of C_3, and vice versa.

The total amplification of this module depends on the resistor arrangement around A_2, to the effect that the amplification, α, of this opamp equals

$$\alpha(A_2)=1+R_{13}/R_{12}.$$

The value of R_{12} is inversely proportional to the resulting total amplification of the module. Like C_3, C_9 must also be dimensioned accordingly.

It is even possible to turn A_2 into a variable amplifier; Fig. 4 shows the necessary circuit modification, which may be useful to correct level differences between, for instance, 33 and 45 rpm records.

Output module 1: the amplification of the summation opamps A_1 and A_1' has been arranged at unity (0dB); this value may be changed, if desired, to a maximum of 10dB by suitable dimensioning of R_1 between values of 47k to 100k. The amplification of the output buffers A_3 and A_3' is 6dB. Since this value is determined by the ratio R_9/R_8, R_8 may be given a lower value and C_{10} a higher value to obtain an increase in output amplification.

A final word about the VU indication: a mixer output level of 0dB corresponds to about $1V_{rms}$ at the input of A_5. If A_3 is arranged to have a higher amplification, the amplification of A_5 should be reduced, and vice versa, of course, to ensure that the VU meter indicates the correct output level. The amplification α of opamp A_5 is given by

$$\alpha(A_5)=1+R_{16}/R_{17}.$$

Table 1 shows a number of alternative LED drivers with different input level ranges and linear or logarithmic characteristics.

Current consumption

The typical current consumption at $\pm 18V$ of all types of module in the portable mixer is summarized in Table 2.

NOTE:
The next part of this article will be featured in our October issue.

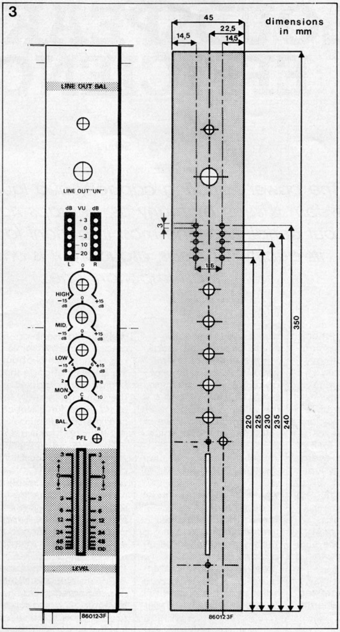


Fig. 3. Front panel foil and drilling template for the output module.

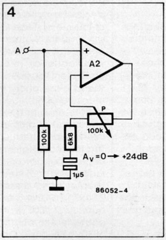


Fig. 4. Showing how the circuit with A_2 may be modified to operate as a variable amplifier stage.

Table 1

Input levels for VU meter ICs

Type	input threshold					units	characteristic
U237B	0.2	0.4	0.6	0.8	1.0	$[V_{rms}]$	linear
U247B	0.1	0.3	0.5	0.7	0.9	$[V_{rms}]$	linear
U257B	0.18	0.5	0.84	1.19	2.0	$[V_{rms}]$	logarithmic
	−15	−6	−1.5	+1.5	+6	[dB]	
U267B	0.1	0.32	0.71	1.0	1.14	$[V_{rms}]$	logarithmic
	−20	−10	−3	0	3	[dB]	

Table 1. Summary of different IC types which may be used in the IC_3 and IC_3' positions.

Table 2

Current consumption of mixer modules [mA].

supply voltage [V]	MIC/LINE	STEREO	OUTPUT 1	OUTPUT 2
+18	20	30	60	80
−18	30	40	25	20

Table 2. Refer to this table when in doubt about the expected total current consumption of the portable mixer.

LOUDSPEAKER EFFICIENCY

by D J Schulz

The power handling capacity of a loudspeaker system is seen by many as, perhaps not the only, but certainly as the most important factor as to its quality, whereas, arguably, it is one of the least important ones.

Loudspeakers convert only about 0.25 to 2.5 per cent of the electrical energy supplied to them into acoustic energy. The remaining 97.5 odd per cent is converted into heat.
The energy efficiency, or simply efficiency, η_o, of a loudspeaker is the ratio of the useful acoustic energy to the signal energy input.

$$\eta_o = 10log_{10}(P_L/P_E) \quad [dB] \qquad [1]$$

where P_L is the total radiated acoustic power in watts, and P_E is the electric power delivered to the speech coil.
The efficiency may, of course, also be expressed as a percentage, when it is

$$\eta_o = 100(P_L/P_E) \quad [\%] \qquad [2]$$

Nowadays, it is customary for producers to state the sensitivity of a drive unit in the relevant data sheet.
The sensitivity is the intensity level in decibels at a distance of 1 metre from the unit (dB m^{-1}), when the electrical signal input is 1 watt, referred to the international standard reference intensity.
The intensity, I, of a plane or spherical "free" sound wave (no reflections) in the direction of propagation is

$$I = p^2/10^2\varrho c \quad [W_a\ cm^{-1}] \qquad [3]$$

where p is the effective sound pressure in pascals, ϱ is the density of dry air at 20 °C (1.205 kg cm^{-3} at an atmospheric pressure of 1.01325×10^5 Pa), and c is the velocity of propagation of a sound wave of small amplitude; its value is

$$c = 330.6 + 0.61\theta \quad [m\ s^{-1}] \qquad [4]$$

where θ is the temperature in degrees centigrade.
The standard reference intensity is 10^{-16} W$_a$ cm^{-2}.
The intensity level in dB of a plane or spherical "free" sound wave in the direction of propagation is

$$L_I = 10\log_{10}(2.42\times10^5 p^2) \quad [dB] \qquad [5]$$

It should be noted that the decibel is not a measure of loudness, since the sensitivity of the human ear to changes in intensity varies with frequency. The unit of equivalent loudness of a sound is the phon: this is a measure of the intensity level relative to a reference tone of defined intensity and frequency.
The internationally accepted standard reference tone has a root-mean-square sound pressure of 2.04×10 Pa and a frequency of 1000 Hz: this is equivalent to an intensity of 10^{-16} W$_a$ cm^{-2}. One decibel represents an increase in intensity of 26 per cent, which is about the smallest change the human ear can detect.
The standard intensity level of 112 dB at 1 m distance from the sound source (112 dB m^{-1}) is equivalent to a sound pressure of 20 Pa and an acoustic power of 1 watt (W$_a$). This is a very high level for the human ear (about the same as a jet engine at 6 metres distance), which, in an average living room, results in a mean intensity level of 104 dB.
The operating input power, P_W, is a useful characteristic indicated primarily on enclosures and loudspeaker system test sheets. It is the electrical power required to produce an intensity level of 90 dB/m (formerly 96 dB/m).
If a loudspeaker system produces an intensity level of 112 dB m^{-1} when the electrical input power is 1 watt, its efficiency is 100 per cent. It follows from formula [1] that for each decibel the actual intensity level is lower than 112 dB the efficiency is reduced to 0.7944 of its previous value. In other words, if the intensity level for an electrical input power of 1 watt is 102 dB m^{-1} (10 dB m^{-1} below the standard of 112 dB m^{-1}) the efficiency is only 10 per cent of that at the standard ($0.7944^{10} = 0.10$).
The reference efficiency, η_o, of a drive unit may be expressed as

$$\eta_o = 9.7\times10^{-8} f_s^3 V_{AS}/Q_{ES} \quad [\%] \qquad [6]$$

where f_s is the resonant frequency stated by the manufacturer; V_{AS} is the volume compliance in litres; and Q_{ES} is the electrical Q(uality) factor.
The values obtained with formula [6] pertain to a hemispherical space subtended onto an infinite baffle (see Fig. 4). Typical values of a popular 25 cm drive unit are: f_s=19 Hz; V_{AS}=310 litres; Q_{ES}=0.28.
Entering these into formula [6] gives an efficiency of 0.737 per cent. Calculating this percentage in decibels ($10\log_{10}0.00737$)—the so-called electro-acoustic index—gives a value of −21.325 dB. The negative sign indicates a loss.
The electroacoustic index added to the standard intensity level gives the Sound Pressure Level (SPL), so that in the above example

SPL=112+(−21.325)=90.675, or, rounded off, 91 dB W^{-1} m^{-1}.

Another example, a polypropylene drive unit with a smaller diaphragm, has the following characteristics: f_s=50 Hz; V_{AS}=13 litres; Q_{ES}=0.93. Entering these into formula [6] yields

$$\eta_o=9.7\times10^{-8}\times50^3\times13/0.93 =0.169\%$$

Electroacoustic index is $10\log_{10}0.00169= -27.7086$ dB

SPL=112+(−27.7086)= 84 dB W^{-1} m^{-1} (rounded off).

A comparison of these two examples shows how the SPL varies with the diaphragm area, when the electrical input is kept constant. Here, the SPLs differ by 7 dB, which is a power ratio of 5:1.

The efficiency is no yardstick for the maximum obtainable loudness level (in phons), but the power handling capacity is. It is, however, necessary, to differentiate between the electrical and mechanical power handling capacities. The former indicates the maximum electrical power in watts that may be applied to the speech coil before this burns out. The maximum loudness level, particularly from a bass unit, depends primarily on the ability of the unit to produce large cone displacements (amplitudes), and this in turn depends on the construction.

The diaphragm should, of course, only move backwards and forwards, not sideways, and stiff, hard materials are better in this respect than soft, pliable ones. However, if the cone material is too stiff, it may actually impede the free movement of the diaphragm. As so often in life, a suitable compromise has to be arrived at.

A further criterion is the difference between the length, h, of the speech coil and the height of the annular air gap, H_E (see Fig. 2). In modern bass drive units this difference lies between six and ten millimetres. Provided that the diaphragm is seated centrally, the speech coil can, therefore, move from ±3 mm to ±5 mm before it leaves the uniform field of the ring magnet. When high electrical power inputs cause the coil to move outside the magnetic field, distortion of the sound produced is the inevitable result. Generally, the spider supports that maintain the speech coil at the centre of the gap allow a free movement of about ±2 mm, outside which they decelerate the diaphragm. It is because of this that many woofers produce distortion at even medium input powers.

Drive units with relatively small cone areas need a greater speech coil displacement to produce the same sound intensity as units with a larger diaphragm. Clearly, these smaller units also reach the limits of their mechanical capabilities sooner. Such compression factors are among the most troublesome in the design of compact loudspeaker systems.

Moreover, frequency modulation occurs when a single diaphragm moves with large amplitude at low frequencies, while simultaneously radiating high frequencies, which causes the high frequencies to be altered because of the Doppler effect.

The following example makes this all a little clearer. The maximum intensity level, L_m produced by a loudspeaker (fitted in a closed box) is

$$L_m=112+10\log_{10}P_L \quad [\text{dB}] \qquad [7]$$

where P_L is as defined before (formula [1]), and may be calculated from

$$P_L=50Z_r v^2 \quad [\text{W}_a] \qquad [8]$$

where Z_r is the radiation impedance and v is the root-mean-square diaphragm velocity in m/s.

The radiation impedance, Z_r is calculated from

$$Z_r=2\pi\rho r^4 f^2/c \quad [\Omega] \qquad [9]$$

where ρ and c are as designated in formulas [3] and [4] respectively; r is the effective radius of the diaphragm in metres; and f is the operating frequency in hertz.

The diaphragm velocity, v is determined by

$$v=H_d f \quad [\text{m s}^{-1}] \qquad [10]$$

where H_d is the difference between the length of the speech coil, h, and the height of the air gap, H_E, in metres, i.e.,

$$H_d=h-H_E \quad [\text{m}] \qquad [11]$$

Again, the 25 cm and 13 cm drive units encountered previously will be compared (fitted in a closed box).

The operating frequency shall be 60 Hz throughout. The 25 cm unit has an effective diaphragm radius, r, of 0.107 m; the length of the speech coil is 0.016 m; and the height of the air gap is 0.008 m. This gives a value for H_d of 0.008 m.

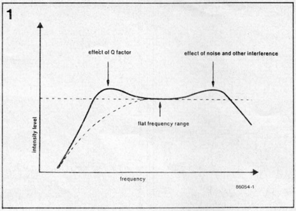


Fig. 1. Intensity level vs frequency characteristic.

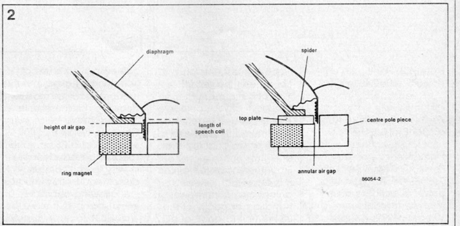


Fig. 2. Cross section of a drive unit. Maximum speech coil displacement without distortion is equal to the difference between the length of the speech coil and the height of the air gap.

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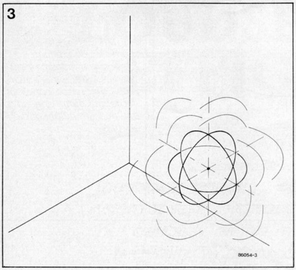

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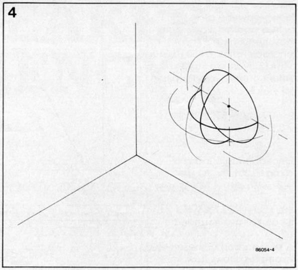

From [10]:
$v=0.008\times60=0.48$ m s^{-1}.

From [9]:
$Z_r=2\times3.142\times1.205\times0.107^4\times60^2/342.8=0.010415974\ \Omega$.

From [8]:
$P_L=50\times0.010415974\times0.48^2=0.11999$ W_a. This acoustic power is equal to an intensity level of $10\log_{10}0.11999=-9$ dB.

From [7] :
$L_m=112-9=103$ dB.

The 13 cm unit has an effective diaphragm radius of 0.05 m; the length of the speech coil, h, is 0.012 m; and the height of the air gap is 0.006 m.

From [11] : $H_d=0.006$ m.

From [10] :
$v=0.006\times60=0.36$ m s^{-1}.

From [9] :
$Z_r=2\times3.142\times1.205\times0.05^4\times60^2/342.8=0.000496945\ \Omega$.

From [8] :
$P_L=50\times0.000496945\times0.36^2=0.00322$ W_a. This acoustic power is equal to an intensity level of $10\log_{10}0.00322=-25$ dB.

From [7] :
$L_m=112-25=87$ dB.

The 25 cm unit requires a signal input of only 16 W to produce the maximum intensity level of 103 dB. Any higher electrical input will lead to distortion.
None the less, this particular unit is rated at 110 W by the manufacturers.
The 13 cm unit reaches its maximum intensity level of 87 dB at a signal input of only 2 W.
From these considerations, it is clear that the mechanical maximum power handling capacity of the 25 cm unit (a good-quality, reputable make) is about 16 watts at a frequency of 60 Hz, while that of the 13 cm unit (also from a first-class manufacturer) is of the order of 2 watts at 60 Hz.
The maximim intensity level may be increased by the use of a bass reflex or horn enclosure. The bass reflex box increases the effective diaphragm area, while a horn enclosure causes a substantial increase in the radiation impedance.
A simple way of increasing the radiation impedance (and thus the efficiency) is placing the bass loudspeaker in a corner of the room (see Fig. 3 to 6 incl.). In practice, this will only work well, however, with loudspeakers that have a small electrical Q factor (Q_{ES}). Such units have a high driving force which ensures that the frequency response rises smoothly into the middle frequencies as shown in Fig. 7.
The best reproduction of bass frequencies is achieved by the use of horn loudspeakers, but this is impractical for most indoor uses as these units are very large.
Finally, a detailed example of a 38 cm loudspeaker intended for use in very large rooms or discotheques. This unit has the following characteristics.

- $f_s=30$ Hz
- $Q_{ES}=0.43$
- $Q_{MS}=2.3$ (mechanical Q factor)
- $Q_{TS}=0.36$ (total Q factor)
- $S_D=0.0780$ m^2
- $V_{AS}=330$ litres
- $h=0.014$ m
- $H_E=0.008$ m
- $P_E=250$ W maximum

To obtain the optimum overall quality factor, Q_{TC}, of 0.6 in normal operation, the enclosure should have a net volume of not less than 160 litres. The resonant frequency of the system then lies around 50 Hz. On the basis of these data, it seems natural to choose a bass reflex enclosure. It should be noted, however, that a net volume of 160 litres would give rise to a poor step response.
A volume of about 250 litres is, therefore, chosen, which lowers the overall resonant frequency, f_c, to around 33 Hz, and gives a clear Chebishev response, i.e. 0.26 dB ripple. The −3 dB frequency is 34 Hz.
The reference efficiency of the drive unit, calculated from formula [6] is 2.01 per cent, equivalent to an electroacoustic index of −17 dB. The SPL is thus 95 dB W^{-1} m^{-1}.
The efficiency of the system at 33 Hz, 100 Hz, and 300 Hz will be different from the reference, because at 33 Hz the bass reflex enclosure will effectively double the area of the diaphragm to 0.1560 m^2. At 100 Hz, the effective area, because of phase shift, is about 0.1170 m^2. At 300 Hz, the reflex aperture has no effect, and the system behaves as a closed box.
At 33 Hz, the effective radius is 0.22284 m, and the radiation impedance is 0.057974816 Ω. The acoustic power is 0.11364 W_a, and the maximum intensity level is 102.6 dB. Since the resonant frequency (33 Hz) is very nearly the same as the −3 dB frequency of 34 Hz, the reference SPL at the resonant frequency is 95−3=92 dB. The maximum intensity level is thus 10.6 dB above the reference level. The maximum power handling at 33 Hz is, therefore, 10 dB above 1 W, i.e. 10 watts.
At 100 Hz, the effective radius is 0.193 m, and Z_r is 0.299549182 Ω. The acoustic power is 5.39189 W_a, corresponding to a maximum intensity level of 119.32 dB, which is 24.32 dB above the reference level of the drive unit. To obtain 5.39189 W_a

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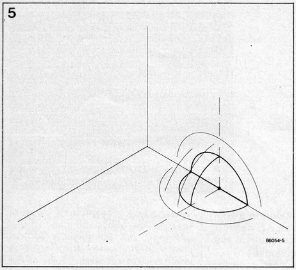

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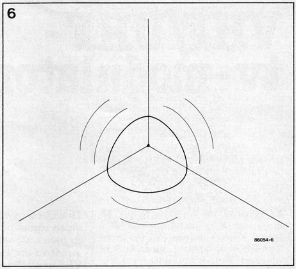

Fig. 3. A freely suspended sound source propagates the sound equally in all directions (spherical).

Fig. 4. A sound source fitted to a closed baffle propagates the sound hemispherically.

Fig. 5. A sound source located at the junction of two baffles propagates the sound in the shape of a quarter sphere. Certain horn loudspeakers operate in this way.

Fig. 6. The radiation impedance of a loudspeaker is increased by placing the unit at the junction of three baffles.

Fig. 7. Typical frequency response of a bass drive unit with a strong driving force (low Q_{ES}).

of acoustic power, therefore, an electrical signal input of around 250 watts is required, i.e. the maximum rated power.
At 300 Hz, the effective radius is 0.1576 m, and the radiation impedance is 1.198688295 Ω. The acoustic power amounts to 194.1875 W_a, which is equivalent to a maximum intensity level of 134.88 dB, or very nearly 40 dB above the reference level of 95 dB. To achieve this, the electrical signal input would have to be an enormous 10 000 watts. This shows that at frequencies above around 200 to 250 Hz the only limitation is the electrical power handling capacity.
Music power handling is of the order of 370 W, corresponding to an intensity level of 121 dB, or some 26 dB above the reference level of the unit. It should be noted that the required electrical power input as calculated pertains only at one frequency. With the amplifier operating over the whole audio range, it has to provide higher powers than calculated to ensure faithful step response.

Conclusions

The foregoing considerations and calculations lead to the following conclusions.

- High efficiencies are only possible with large effective diaphragm areas.
- Large cone areas result in lower distortion than small diaphragms.
- The efficiency cannot be improved by more than 6 dB however much the electrical input is increased. The main reason for this is that, particularly at low frequencies, the mechanical power handling capacity becomes the limiting factor.
- The electrical power handling capacity, because of modern construction methods and improved speech coils, has become one of the least important parameters of a loudspeaker system.

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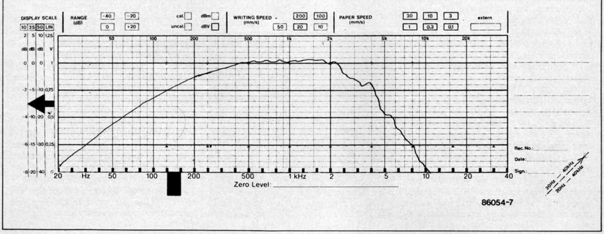
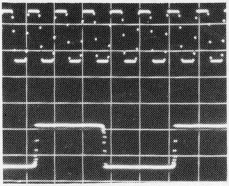
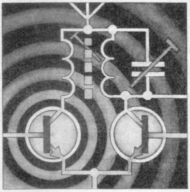


VHF/UHF - tv - modulator



To illustrate the principle of the TV modulator it is useful to look at a typical video waveform and the corresponding modulated r.f. signal, both of which are illustrated in figure 1.
Figure 1a shows one line of a video waveform. The maximum positive excursion of the signal is known as white level, since it is the signal obtained from white areas of the picture. Line sync pulses are, of course, present at the beginning of each line, and are distinguished from picture information by the fact that they are negative-going pulses from 33% of white level down to zero (sync level). Picture information, on the other hand, extends from 33% (black level) up to 100% (white level). This description of a video signal is necessarily rather brief, and the various levels, etc. for broadcast video signals are, of course, defined much more rigorously.
An r.f. signal amplitude-modulated with this video signal is shown in figure 1b. It will be noted that the type of modulation employed is *negative modulation,* i.e. minimum video signal level (sync level) corresponds to peak r.f. signal level and vice versa. This type of modulation is used in the practical modulator circuit, which means that it is unsuitable for use with British, VHF, 405-line TV sets, which use positive modulation. In the UK the modulator must be used with UHF, 625-line sets, which are designed for negative modulation.
The VHF output capability of the modulator is principally intended for use in countries outside the UK which use VHF systems employing negative video modulation.
In a broadcast TV transmitter great care is taken to ensure that the carrier is a pure sinewave, otherwise spurious signals could occur around harmonics of the carrier frequency. Steps are also taken to reduce wastage of transmitter power by partial suppression of the carrier, and one of the sidebands of the signal is also partially suppressed to minimise the bandwidth of the transmitted signal. This is illustrated in figure 2.
In a TV modulator for domestic use none of these criteria apply, since the signal is not going to be broadcast (and care must be taken to ensure that it is

This circuit will modulate a video signal onto an r.f. carrier to give a signal that may be fed direct to the aerial socket of a VHF or UHF television receiver.



not broadcast). There is no need to suppress the carrier or one of the sidebands, and the presence of harmonics of the carrier frequency is a positive advantage since (if the carrier fundamental is in the VHF band) it allows TV sets to be tuned to these harmonics right through from the VHF band to the UHF band. This means that a single modulator can supply signals to both VHF and UHF sets and makes tuning easier, since the set can be tuned to a signal at one of several frequencies throughout its tuning range.

Modulator circuit

The fundamental carrier frequency is derived from a 27 MHz crystal in an oscillator circuit based on T1 in figure 3. For domestic use, crystal stability is not always required. In that case the crystal, X1, can be replaced by a 10 n capacitor. The output signal of this oscillator is amplified by T2 and T3 and differentiated by the three RC networks C3/R4, C4/R6 and C5/(R9 + P1). The resulting waveform at the junction of R8 and R9 is a sequence of short spikes containing harmonic multiples of 27 MHz up to around 1 GHz.
The video signal is fed in via P2 and modulates the carrier by varying the forward bias on D1 and thus changing its impedance. This causes the level of the r.f. signal appearing across R10 to vary in sympathy with the video input signal, i.e. the carrier signal is amplitude modulated. The signal is coupled out via C7 to a coaxial output socket. R13 matches the output impedance of the modulator to that of the coaxial cable.
Potentiometer P1 can be used to set the carrier level by varying the static forward bias on D1, whilst P2 adjusts the video input level and hence the modulation depth.

Construction and adjustment

A printed circuit board track pattern and component layout are given in figure 4. This board is available from the Elektor Print Service, EPS No. 9967. Two alternative mounting positions are provided for the crystal, allowing for two different pin spacings.

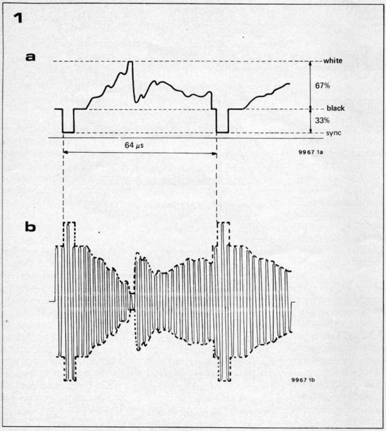
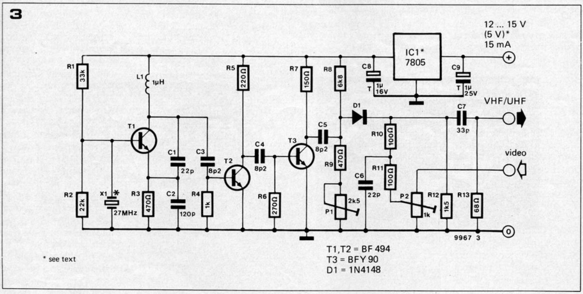


Figure 1. a. One line period of a typical video signal, showing picture information and line sync pulses. b. An r.f. carrier modulated with the signal of 1a, using negative modulation.

Figure 2. a. Spectrum of a broadcast TV signal with partially suppressed lower sideband and vestigial carrier. b. Spectrum of a TV modulator for domestic use, in which both sidebands and the carrier are retained. This spectrum is also repeated at multiples of the carrier frequency.

Figure 3. Complete circuit of the TV modulator. The precise frequency of the crystal is not critical and any radio control crystal around 27 MHz will be suitable.

Figure 4. Printed circuit board and component layout for the circuit of figure 3. (EPS 9967).



Because of the high frequencies involved the board is designed with a generous earth plane for stability. In addition a screening plate, made of tinplate or a piece of copper laminate board is connected between the oscillator and modulator. The completed board *must* be mounted in a metal box for screening, to avoid the possibility of stray radiation.

The modulator may be powered from +12 V to +15 V unstabilised DC supply, which is stabilised at +5 V by the IC regulator on the board. Alternatively, the unit may be powered direct from an existing stabilised +5 V supply, in which case IC1 should be omitted and the holes in the board for its two outer pins should be bridged by a wire link.

Setting up the modulator is extremely simple. Connect the modulator to the aerial input of the TV set using 75 Ω coaxial cable, then switch on the modulator and the TV set. Set P1 to its mid-position and tune the TV set to one of the harmonics of the carrier. This will be around channel 7 (189 MHz) in the VHF band and at a number of frequencies in the UHF band. When the carrier is picked up the screen of the TV set will darken and noise (snowstorm effect) will disappear.

A video signal may now be fed in, and

Parts list to figure 2.

Resistors:
R1 = 33 k
R2 = 22 k
R3,R9 = 470 Ω
R4 = 1 k
R5 = 220 Ω
R6 = 270 Ω
R7 = 150 Ω
R8 = 6k8
R10,R11 = 100 Ω
R12 = 1k5
R13 = 68
P1 = 2k5 (2k2) preset potentiometer
P2 = 1 k preset potentiometer

Capacitors:
C1,C7 = 33 p
C2 = 120 p
C3,C4,C5 = 8p2
C6 = 22 p
C8,C9 = 1 µ/16 V tantalum

Semiconductors:
T1,T2 = BF 194, BF 195, BF 254, BF 255, BF 494, BF 495.
T3 = BFY 90
D1 = 1N4148
IC1 = 7805 (see text)

Miscellaneous:
L1 = 1 µH
X1 = crystal, 27 MHz approximately. (or X1 = 10 nF, see text)

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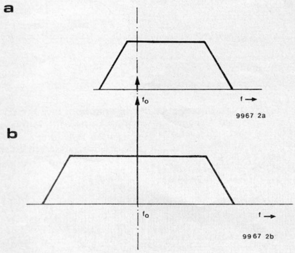

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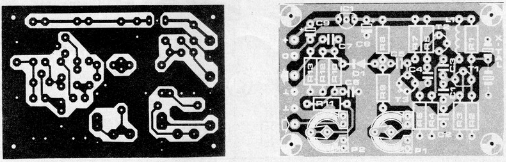


P2 should be adjusted so that the video signal level does not exceed 3 V peak-to-peak at its wiper.
The TV set may now be tuned to the sideband which gives the best picture. If tuned to the wrong sideband the picture will tend to appear negative. If the picture lacks vertical synchronisation (i.e. rolls) it will be necessary to adjust P1 until it stabilises. P2 is used to adjust the contrast by varying the video input level, but should not be turned up too much or the modulator will overload, causing the picture to appear negative on highlights.
Finally it should be noted that, when using the modulator, the output should always be connected direct to the TV set via a length of coaxial cable and must never be connected to any unscreened wire or other conducting object that could act as an aerial, otherwise the user could receive an unwanted visit from the Post Office Radio Interference Officer! ■

A compact radar for helicopters

by A W Pressdee, BSc, CEng, MIEE

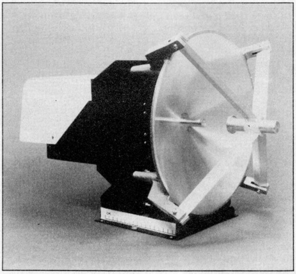
The pilot of a Hiller UH12 helicopter engaged on crop-spraying in Lincolnshire failed to see the spur of a power line to a farm. The helicopter hit the wires which damaged the control rods and caused it to climb out of control until it crashed into a field.
Over the last six years the number of United Kingdom registered helicopters sustaining damage from hitting power cables has run into double figures.
Helicopter accidents occur during a diversity of tasks. Many happen while crop-spraying, but others happen on surveys, low-level photographic work, and military operations.
Such examples of accidents to helicopters are repeated in international air accident statistics.
Wherever this type of aircraft operates at low altitude, it is vulnerable to the hazards of poor visibility and unseen obstacles, particularly power lines, or both. Even when power lines have been clearly visible to the pilot, accidents have happened because he has been unable to estimate accurately his distance from them.
For the majority of helicopters, payload is of prime importance. For most small and medium types, weight and space considerations rule out the carrying of bulky radar or obstacle-detection equipment. Instead, the pilot has to rely almost entirely on his visual acuity and good sense. Accidents caused by collision with unseen or undetected obstacles are unfortunately common and, even if not fatal, are expensive in terms of repairs to the aircraft and compensation for the damage it causes.

Electronic solution

Such problems may be mitigated by a new radar system under development by Philips Research Laboratories at Redhill, Surrey. It is small, light in weight, and compact. It is also extremely accurate and has high definition, operating at millimetric wavelength, and employing a technique known as frequency modulated continuous wave (FMCW).
The ability of a radar system to detect an object depends directly on the "illumination" of that object by electromagnetic waves. Conventional pulsed radar systems employ high power pulses, at say 10 kV, of very short duration, perhaps one microsecond, at a pulse repetition frequency of possibly 1000/s, giving a relatively low mean target illumination of one microsecond pulse every thousand microseconds.
An FMCW radar illuminating the target continuously —for 1000 microseconds every 1000 microseconds —can achieve comparable target illumination and hence equivalent or better target detection with considerably lower power. The low-voltage system enables millimetric wave solid-state oscillators to be used. Lower voltages and solid-state techniques mean considerable reduction in space and weight.
The FMCW radar operates at 94 GHz, about ten times the frequency of most standard radars, which enables a very compact front end unit to be assembled. This incorporates a 10 mW bias-tuned Gunn oscillator for the transmitter and a balanced mixer in the receiver, both items developed by Philips Research Laboratories. The aerial reflector dish is 300 mm in diameter and produces a beam width of 0.7 degrees.

Fast frequency sweeps

In the FMCW radar system the transmitter is modulated with a continuous linear sweep. Consequently, the frequency of a returning echo will differ from the instantaneous frequency of the transmitter by a beat frequency proportional to the target range. The use of fast frequency sweeps allows small range differences to produce large frequency differences so enhancing the range resolution.
The received signal generally will contain several frequencies, corresponding to targets at different ranges, so a means of frequency analysis is necessary. This is achieved by a mathematical technique known as fast Fourier transform (FFT). The technique analyses data over a fixed period, made conveniently equal to the transmitter sweep time, so frequency flyback does not affect operation of the system.
In the past, FFT has required several hundred medium scale integrated (MSI) logic chips, but the application of high speed very large scale integrated (VLSI) techniques to digital signal processing has increased the speed and reduced the size of such systems. For the FMCW radar, a Texas Instruments TMS 320 programmable single chip processor implements a single Euro-card sized board comprising the FFT processor.
A dual processor system is used to maximize the duty factor for each set of data samples. One processor is always inputting data for the next FFT and outputting data for the previous FFT, while the other processor is executing the current FFT. At the completion of this sequence,



The front end aerial dish is just 300 mm in diameter.

the two processors exchange functions.

Field tests

The specification of the FFT was determined by the maximum range and range resolution requirement coupled with the bandwidth needed for the FMCW receiver. All the FFT software has been written with a high level assembler and, to maximize program execution speed, straight line coding was used; in other words, to remove all time consuming program loops, all executable instructions were placed in consecutive locations.

The FFT has been connected to an experimental display system and various field tests of the development system conducted to obtain radar views of sites which would present hazards to helicopter operation. Radar pictures in plan and elevation have been obtained of aerial towers and power cable lines. On the display used, which is colour-coded to show incremental height, strong signals were recorded from the cables, the conductor spacers and the top of the supporting pylon. The droop of the catenary can be clearly seen. As the system can measure targets of approximately 20 m² at 400 m with signal-to-noise ratio of 34 dB, such results can be confidently expected.

It is evident from the tests that the viability of FMCW radar as a future aid to reduce helicopter accidents has been established and that the system can be made sufficiently light and compact for smaller helicopters. As the system nears production model stage, careful consideration is needed for the development of a display system which will give the pilot the information he needs in an easily understandable form.

An additional bonus of the FMCW system is the fact that it is very difficult for electronic systems to detect it and even then it exhibits an excellent electronic counter-counter measures (ECCM) performance.

There are uses for the system, or components of it, other than in helicopters. Its high target definition, coupled with its compactness and portability, suggests a variety of possible applications such as in weapon guidance for small munitions systems.

(LPS)

Philips Research Laboratories, Cross Oak Lane, Redhill, Surrey, England, RH1 5HA.

Monitoring highways electronically

by K W Dickinson

Techniques for the automatic detection and counting of vehicles as they pass an overhead video camera, have been developed from research at the University of Sheffield[1] and the University of Manchester Institute of Science and Technology (UMIST)[2]. Individual vehicle speed and length can also be estimated from a sequence of video images.

During recent years there has been a growing need for greater traffic monitoring.

Traffic data can be collected automatically by equipment installed in the carriageway. However, such installations are generally only suitable at permanent or semi-permanent sites.

Traffic planning and surveillance engineers are now becoming increasingly interested in the concept of a wide area vehicle detector, based for example on a video camera rather than the present highway point sensors such as axle or inductive loop detectors.

Although highway authorities are using more video cameras for single traffic surveys, manual analysis of the video tapes is time consuming. It can take up to five hours to analyse a 30 minute recording. Because of such problems there is an obvious need for a completely automatic video system to monitor traffic at permanent sites and provide automatic data abstraction from video tapes of short-term traffic surveys.

During the early 1980s several computer-based video image processing systems were constructed by engineers from Sheffield University and UMIST. Over the past three years the British Transport and Road Research Laboratory[3], as part of its research programme, has provided financial support for a project, the aim of which was to assess the feasibility of using image processing to collect traffic data for various purposes and later to develop techniques for traffic monitoring on motorways.

Accurate system

This has resulted in the design and construction of the traffic research and image processing (TRIP) system, a flexible development tool based on a powerful Intel microcomputer linked to purpose-built hardware.

The fully automatic system is capable of accurately counting vehicles, measuring their speed and length, and calculating lane occupancy and the gaps between the vehicles.

In its original version, TRIP takes the video signal from a solid-state camera, converts it to digital form, and presents the video picture to the computer system as a two dimensional array of numbers, each representing the average image brightness (grey value) in that picture element (pixel). Vehicles are counted by analysing the different shades of grey whithin each image and by filtering out extraneous non-moving features such as carriageway markings and parked vehicles.

There are several ways of interpreting a sequence of digitized images and detecting moving objects in the scene. The TRIP system uses both background frame and inter frame differencing techniques.

Essentially, background frame differencing is a method of storing a grey value reference image, which does not contain any vehicles, and subtracting it from each incoming frame or image. This causes all non-moving features of the image to disappear, leaving moving grey value objects which can be represented as binary images after applying a suitable threshold. Counting vehicles then becomes the simpler task of counting white shapes. However, since ambient light levels can radically change within seconds, it is periodically necessary with this method to update the stored background frame for satisfactory detection of moving vehicles over a length of time.

Overcoming daylight variations

Inter frame differencing usually overcomes the problem of changes in ambient light. A background frame is again subtracted from the incoming image, but then the incoming frame becomes the background for the subsequent frame.
Such systems can suffer from problems associated with matching edges from frame to frame; stationary vehicles disappear and random noise in pairs of frames becomes cumulative.
Trials were undertaken during 1985 to evaluate the feasibility of the TRIP system to monitor traffic passing a point on the road network. During trials with an earlier system, a high threshold was applied to overcome noise in the binary image caused by changes in lighting conditions. Therefore twelve per cent of vehicles were missed and, because of limitations in the available computing power, the system was only able to capture images at 4 frames/s. Also, it was impossible to estimate vehicle speeds because a fast vehicle might appear in one frame but move out of the scene before the next frame.
Performance of the TRIP system has been improved by splitting each image and concentrating attention on a few small but important areas (windows) within each scene. This can be considered as a method of projecting simple light sensors on to the carriageway.
Both background frame and inter frame differencing techniques can be applied to the whole image or windows within the image. If processing is confined to windows, the image frame rate can be increased, giving a better measure of the time at which an event occurs.

Site trials

During site trials, the solid state video camera was mounted at heights between 8 m and 24 m above the carriageway and several windows were superimposed across the image of each traffic lane.
Images of the scene were sampled at 50 frames/s and the time that each vehicle was observed/detected at a window was later compared with its time of arrival at the next window. By knowing the space between the road elements corresponding to the window positions, an estimate of vehicle speed, vehicle length, and lane occupancy could be derived.
Trials were carried out in a range of weather conditions and ambient light levels. The results indicated a miss of less than 1 per cent of vehicles.
Estimates of individual vehicle speed were found to be within 10 per cent although no systematic error in speed was observed.
It is now possible to use the TRIP development system to automatically detect and measure the speed of vehicles as they pass through a typical highway environment in which ambient light levels change. However, wider traffic engineering applications such as surveillance throughout a 1 km stretch of motorway for automatic incident detection purposes, vehicle tracking through a junction, and classification of traffic by vehicle type are awaiting further investigation by the Sheffield University/UMIST TRIP group.
Nevertheless, much effort will be required to provide suitable applications software before general purpose traffic data collection systems based on video image analysis are readily available. (LPS)

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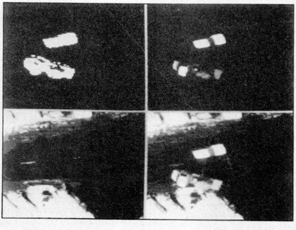


Background or reference frame differencing images. Top left: input. Top right: reference. Bottom left: difference. Bottom right: binary.

The role of plastics in communications

The use of plastics in the world's expanding information technology industries will be examined by leading world experts at the Plastics and Rubber Institute's fourth international conference, to be held in London from 17 to 19 September this year.
More than 100 delegates are due to attend from outside Britain, including a large delegation from Japan, which will be fielding six conference speakers. Leadings professionals are also expected from the polymer and telecommunications industries in the United States and Australia.
The conference, to be held at the Institution of Electrical Engineers, will feature thirty-three lecture papers, reinforced by displays in the conference hall.
Apart from a detailed examination of the role of plastics in telecommunications equipment, the delegates will be briefed on the latest testing methods by British Telecom's Materials and Components Centre, which is organizing a reception for delegates aboard a motor vessel on the Thames. (LPS)

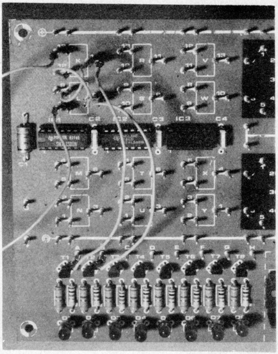
Plastics and Rubber Institute
11 Hobart Place
London SW1W 0HL

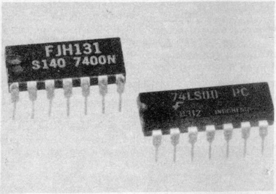
Digi Course II

Chapter 8

In the last few chapters of Digi Course II we saw how combinations of individual building blocks available in form of integrated circuits are achieved.

The basic idea is to use standard building blocks like the NAND gates, to create different logical functions. The ICs which can be combined in such a way without any problems of matching (compatibility) are said to be of the same family. ICs of the TTL and LS TTL families are examples of such grouping.

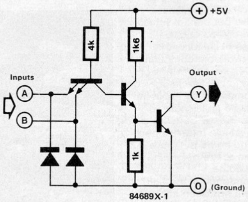




TTL

The TTL family is characterised by their type numbers starting with 74. The LS TTL family also has similar numbers starting with 74 LS. Both TTL and LS TTL families are very similar, most of the individual ICs are interchangeable, and pin to pin compatible. Only the output loading capacities differ. As LS TTL output should not be loaded with more than 5 TTL inputs. The 74 LS series ICs are as fast as the 74 series ICs but consume less current than the 74 series ICs, contrary to the concept that faster the circuit, greater is the current consumption. This has been made possible by the low power Shottky devices used in these ICs.

1

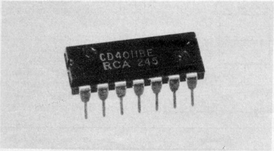


Outputs of these ICs can not be coupled arbitrarily. Only when the outputs are always logically identical, at the most 2 such outputs can be coupled together to enhance the output loading capacity.

With how many inputs can one output be loaded? This can be calculated from the specified loading factors of the ICs, which are given in the data sheets of the ICs. When connecting more than one inputs to an output of another IC, care should be taken not to exceed the specified loading capacities.

For example, consider a 7404 inverter which has an output loading capacity of 10. We can connect 4 clear inputs of 7476 ICs and 2 gate inputs of 7400 ICs. Without overloading the 7404 inverter.

TTL and LS TTL ICs can be operated at 5V (I 0.25 V). This makes it difficult to adapt them into circuits operating from other sources voltages. To overcome this difficulty to some extent, some devices have been designed with open collector outputs as shown in figure 1. In case of these ICs, the output does not switch between 0 and 5 Volts but it has a built in driver transistor with emitter tied to ground and collector brought out on the output pin. Such ICs can work with voltages upto 30 V at the output pin which then switches between 0 and 30 (if the output is tied to 30 V through a pull up resistor).

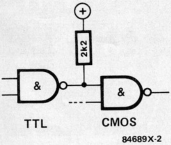


CMOS

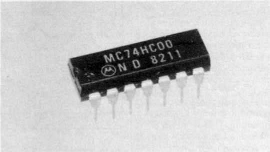
CMOS technology is totally different from the conventional TTL technology. Though the logic of all gates must be same whether they are TTL, LS TTL or CMOS, the electrical characteristics are different. CMOS ICs consume very low power, as the operating current currents are very low. However, the speed is sacrificed tosome extent. A TTL NAND gate draws almost 20 times more current compard to the CMOS counterpart. As the current drawn is dependent on frequency of switching, the CMOS operated at low speeds consumes still less power.

The CMOS logic ICs are generally identified by type numbers starting with 40 or 4. A 4001 contains 4 CMOS - NOR gates. The 40 series ICs are not pin compatible with 74 series. The supply voltage for CMOS ICs can be between 3 and 15 Volts. The TTL and CMOS ICs are difficult an arrangement similar to the one shown in figure 2 does work. CMOS ICs may drive LS TTL inputs but never a TTL input.

2



Another point to remember about CMOS is that no input pins should be left floating; to reduce effect of interference. Even CMOS ICs can be used onthe Digilex Board provided that all precautions are taken to avoid any damage to them. CMOS inputs are sensitive to electrostatic discharges and can be damaged even during handling. CMOS ICs are to be stored on conductive foam or in an aluminium foil.



74 HC

A more advanced development in the Integrated Circuit Technology is the high speed CMOS ICs. These ICs combine merits of both the TTL and CMOS technologies. Their speeds are as fast as TTL and the current drawn, as low as the CMOS ICs. However, these are still out of the reach for the hobbyist due to their high cost. This family is characterised by type numbers starting with 74 HC

With this chapter, our Digi Course comes to an end. The theme "Digital Technology" is of course not finished — there is much more to learn. More about it in the next issue.

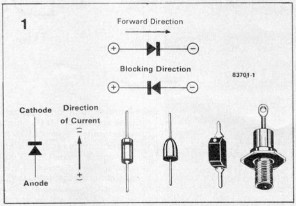
1 nS = Nanosecond = 1 billionth of a second.

Threshold Voltage and the LED.

We already know that the diode conducts current only in one direction, like a "Value of current". Figure 1 shows the directions in which current flows or blocks. The direction in which current flows is called the forward direction. The blocking direction is called the reverse direction. Current flows when the Anode is more positive than the Cathode. In the diode symbol, the bar represents the cathode. Physically, the diode has coloured ring or a dot marked on the body to indicate the cathode.

Diode can be compared also to a switch which depends on the polarity. Just as a mechanical switch requires pressure of our fingers to close it, the diode requires electrical pressure (potential difference). Only difference is that once the mechanical switch is closed it remains closed. Diode requires continuous energy to keep it conducting. This is more similar to a mechanical key switch, which remains closed only as long as the pressure is applied, and opens as soon as pressure is removed. Applying a pulling force instead of pushing does not close the switch. Electrical pressure (Potential difference) applied to the diode in the reverse direction does not force current through it.

The energy required by the diode to keep it conducting appear as the loss of voltage across its terminals. This drop in voltage is about 0.6 to 0.7 V in Silicon diodes. This is also called the threshold voltage because the diode cannot start conducting unless a voltage more than this value is given in the forward direction across the diode. Below this threshold the diode cannot conduct even if the voltage across it has the correct polarity.



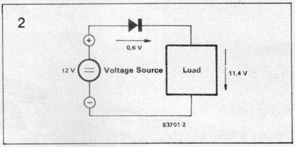
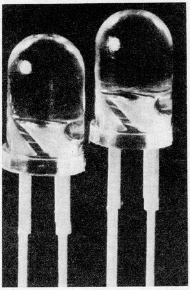


Figure 1

Blocking and conducting directions of a diode. The cathode is marked as a coloured ring or dot on the body of the diode.

Figure 2

There is a drop of about 0.6 V in the forward direction across the diode. The difference between the supply voltage and the diode voltage appears across the load.



A Germanium diode requires around 0.2 to 0.4 V for conducting. This voltage does not fluctuate very much with change in current through the diode and can be used as a reference voltage. However, it should be remembered that the threshold voltage is dependent on temperature. Figure 2 shows how the voltage is distributed between the diode and the actual load. Figure 3 shows a simple arrangement to obtain a reference voltage of 0.6 V. The limiting resistor ensures that the diode current does not become too high. As the voltage across the diode is fixed, energy dissipated as heat in the diode is directly porportional to the current flowing through it. A diode which carries high currents must therefore be cooled by providing a heat sink.

LED

The energy that is dissipated as heat in ordinary diodes has been exploited by the inventive scientists to design a very useful device called LED!

LED is a Light Emitting Diode, which is similar to an ordinary diode except that the energy required by the diode to keep it in conduction is given out in from of light rather than heat. LEDs are made of materials like Gallium Arsenide or Gallium Phosphide; and can have different colours depending on material. Threshold voltage for an LED can be between 1.6 V to 2.2 V (See table 1). The intensity of glow depends directly on the current flowing through the LED. Commercially available LEDs have current ratings upto maximum 50 mA. A safe value to operate the LED without damage is between 15 to 20 mA.

The cathode of an LED can be seen through the transparent casing and is a broad dish shaped electrode. The cathode terminal is made shorter than the anode terminal as a physical indication of polarity.

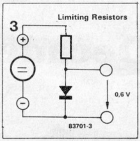
The current limiting resistor can be calculated using the Ohm's law, taking into consideration the threshold voltage. For example: A green LED being opeated from a 12 V supply and having a threshold voltage of 2.2 V will need a limiting resistor given by the following calculation.

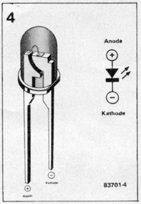
$$U_R = 12\,V - 2{,}2\,V = 9{,}8\,V$$

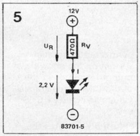
$$R = \frac{U_R}{I} = \frac{9{,}8\,V}{20\,mA} = 490\,\Omega$$

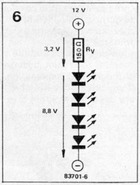
The nearest available standard value is 470Ω, which can be used with LED.

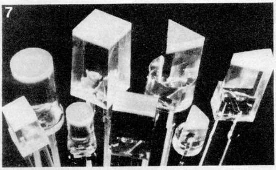
When more than one LEDs are connected in series, their threshold voltages will add up. If we connect 4 LEDs in series, having an individual threshold voltage of 2.2 V then the total drop across them would be 8.8 V, thus leaving only 3.2 V to be taken by the limiting resistor. A resistor carrying 20 mA with 3.2 V across its terminals must be about 150 Ω











Parallel combination of LEDs is not a sensible application because depending on individual characteristics they will carry different currents and thus give varying intensities.

The light emitting property of LEDs affects their blocking properties in the reverse direction. LEDs can tolerate at the most 3V in the reverse direction, and should never be operated with reverse polarity. The leakage current through an LED with reverse voltage can go upto 0.1 mA whereas a regular diode like 1N 4148 typically conducts about 25 nA in the reverse direction (one nanoampere = One billionth of an ampere) when supplied with 20V reverse voltage.

LEDs should be used only for indication purpose and not as ordinary diodes, so that their defective blocking properties do not become significant - LEDs are available in various sizes and shapes.

The most commonly available colours are red yellow and green. Blue LEDs are also being offered by some manufacturers but they are very expensive.

Figure 3

A reference voltage of 0.6 V canbe obtained from any standard voltage source.

Figure 4:

LEDs are always connected in the forward direction. The cathode canbe recognised by three features: the shorter terminal, flatterned body and wider of the two inside electrodes.

Figure 5:

The limiting resistor takes the difference between the supply voltage and the threshold voltage of the LED. Value of the limiting resistor decides the current passing through the LED.

Figure 6:

Threshold voltages add up in a series connection of LEDs.

Figure 7:

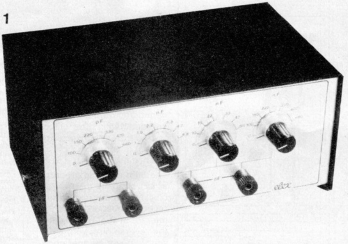
LEDs are available in various shapes, sizes and colours. A wide range of LEDs characterises the appearance of modern electronic apparatus.

Table 1

Colour	Light Intensity	Beam Angle	LED Current (mA)	Threshold Voltage (V)
Red	1	90°	20	1,6
Red (High glow)	3	90°	20	2,2
Yellow	2,5	90°	20	2,2
Green	2,5	90°	20	2,2

Capacitance Decade Box

1



You already know what a resistance decade box is. A capacitance decade is almost same - except for the fact that it uses capacitors!

The circuit is very simple, and very useful for experiments. Our decade box has two pairs of output terminals, one for values from 0 to 680 pF and the other far values from 0 to 750 nF. A photograph of the decade box is shown in Figure 1. There are four rotary switches. The leftmost switch controls the available capacitance at the leftmost pair of terminals which gives 0 to 680 pF. Remaining three switches control the capacitance value available at the second pair of terminals. The connection diagram is shown in figure 2.

It can be seen from figure 2 that switch 2 gives capacitance values from 1 nF to 6.8 nF, switch 3 gives from 10 nF to 68 nF and switch 4 gives 100 to 680 nF. As all the three are connected in parallel, what we get on the second pair of output terminals in the sum of the three capacitance values selected by S_2 S_3 and S_4.

Figure 3 illustrates the exact operation of these three switches. It is possible to get a combination of maximum 3 capacitors from the three individual groups and the result is their sum, because of the parallel combination. If any one or two switches are used, it will either give

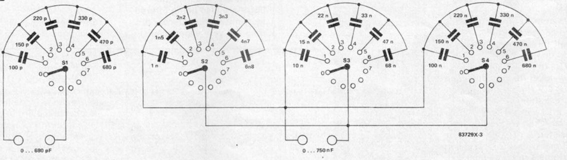
Figure 1 :

A sturdy housing with a properly laid out front panel gives a professional look and ease of operation.

Figure 2 :

Total 24 capacitors are divided into 4 groups with 6 pieces each. Switch S1 controls the first group, of 6 switches S2, S3, S4 control the remaining three groups.

2



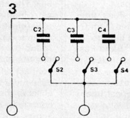


Figure 3 :

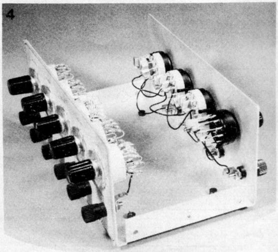
Various combinations of effective capacitance are possible.

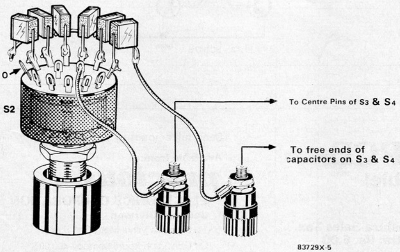
Figure 4 :

Capacitors are directly soldered on to the lugs of the switches. Those who have already constructed the resistance decade can use its rear panel to serve as the front panel of the capacitance decade.

Figure 5 :

All soldering of capacitors should be carried out before mounting the switches on the front panel. Including the Zero position, the switch requires 7 positions.





one capacitor across the ouput or a sume of two capacitors which are selected. If S_2, S_3, S_4, are all closed to select a capacitance, the effective value is

$$CE = C_2 + C_3 + C_4$$

The characteristic of a parallel combination of capacitors is that the individual values add up to give the effective value. This is in contrast to the characterisc of resistors in parallel combination. In case of resistors the series combination gives the total of individual resistance values.

We shall take a few examples to see the usefulness of the decade box. Say, we need a capacitor of 270 nF. We can set S4 to 220 nF and S_3 to 47 nF so that the result is 267 nF. It is certainly not exactly 270 nF, but if we consider the tolerance range of capacitors available, the value of 267 nF calculated theoretically will fall within the tolerance range of $\pm$ 10%. We could have used switch S2 at 3.3 nF to get the effective value of 270.3 nF, but practically it would make no difference. Now, let us take another example where we need 39 nF. In this case we set S2 to 6.8 and S3 to 33 so as to get 6.8 + 33 = 39.8 nF.

Using S1 we can obtain 0 to 680 pF and using S2, S3, S4 we can obtain 0 to 754.8 nF.

If we need a value little higher than 650 pF, we can connect both pairs of terminals in parallel to get a maximum addition of 750 nF to the 680 pF capacitance.

Construction

While assembling this useful decade box, the main job consists of soldering the 24 capacitors in place over the roatry switches, as shown in figure 4 and figure 5.

6 capacitors are soldered on the lugs of the rotary switch S1 as shwon in figure 5. The free ends of the capacitors are connected together and then to one of the output terminals of the left most pair. The centre pin of the switch (the common point connected internallyto the wiper contact) is connected to the remaining output terminals. S2, S3, S4 are also connected similarly as shwon in figure 2. All free ends of the 18 capacitors coming to one of the output terminal and all three centre: pins of S2, S3, S4 coming to the remaining terminal. Once all 4 switches are soldered they canbe fitted onto the front panel.

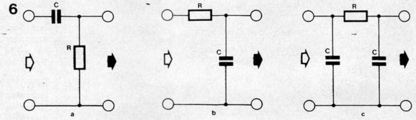
The type of capacitors to be used depends on your application, required accuracy and your budget. High accuracy will always cost more.

Application

This capacitance decade box can be used with the resistance decade as a very useful test unit for R.C. cirucits. When working with AC cirucits, various RC combinations can be used as filter cirucits. High-Low or Band Pass filters. These are nothing but frequency dependent impedances. A 'High Pass' filter is one which allows only frequencies above the desired limit to pass through. A 'Low Pass' filter allows frequencies from 0 upto the desired limit to pass through. A 'Band Pass' filter allows frequencies between a lower and a higher limit to pass through. In case of a practical filter cirucit, even frequencies which are not intended to pass through are allowed to pass through the filter but they are considerably attenuated.

Figure 6 shows three basic filter circuits using RC network.

Figure 7 shows how you can start experimenting with the capacitance decade box. The cirucit shown is an astable multivibrator which is "almost" complete — just a capacitor between points A and B is missing. You can use the decade box to connect different values of capacitors across A B and observe the effect. With change in capacitance value across A B, the frequency and duty cycle of the astable multivibrator changes. First use the leftmost pair of output terminals to select a capacitor from 100 to 680 pF. Then use the second pair to select 1 nF to 750 nF. You will get a complete series of audio frequencies.



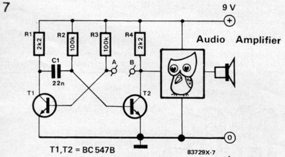


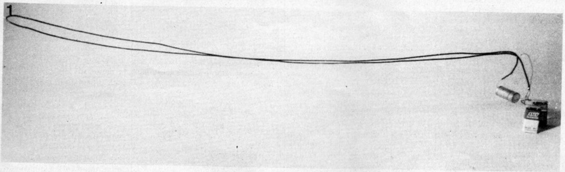
Figure 6 :
A resistance decade and a capacitance decade together can create an RC circuit which plays an important role in AC applications.
Three basic types of RC filter circuits are shown here: (a) High Pass, (b) Low Pass and (c) Band Pass.

Figure 7 :
By connecting the capacitance decade between points A and B you can create different audio frequencies by changing the effecting capacitance values.

High current and magnetic fields

Electric current and magnetic fields are very closely related. Any current flowing through an electrical conductor produces a magnetic field around itslef. The higher the current flowing, the higher is the magnetic field surrounding it. Ordinary electrical wiring carrying currents also has a magentic field surrounding itslef, but as the currents

1



are not very high, the magenetic field will need very sensitive equipment to detect it.

Existance of such a field can be proved with a simple experiment, by producing a very high current.

For the experimental construction, about 1.5 meter insulated thin copper wire, an electrolytic condensor of 4700 μF/25V, two 9V batteries and a changeover switch are required. The copper wire is suspended as shown in figure 1 without tension. The distance between the outgoing wire and the returning wire should be as small as possible, not more than one or two millimeters. One end of the suspended wire is connected to the negative terminal of the battery and the capacitor. The changeover switch is connected as shown in figure 2. The experimental set up is now ready.

As the switch connects the capacitor across the battery, the capacitor gets charged to 18 Volts. Once the capacitor is charged the switch is thrown over to the other position, where it connects the wire directly across the capacitor. This forces the stored charge in thie capacitor to discharge quickly through the wire, producing a very high current for a moment. This momentary current goes almost upto 45 A in the above set up.

This high current flowing through the conducting wires produces a magnetic field around itself. This creates a momentary jerk, which can be observed. The duration of this jerk can become more visible if larger values of capacitors or higher charging voltages are used. (Be careful about the the rated Votage of the electrolytic condensor.)

2

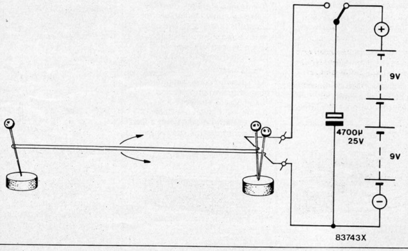


Figure 1 :

The conducting wire loop must be suspended without tension and the gap between them should be very small.

Figure 2 :

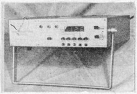
The circuit of the experimental set up. The electrolytic capacitor is charged from the batteries to 18 V. After the changeover switch is thrown to othe position, the capacitor quickly discharges through the wire, driving a very high current through the loop.

NEW PRODUCTS

DIGITAL MEGGER

ARUN has introduced a Digital Megger for measurement of insulation resistance. This is designed for 0.1% linearity throughout the range. The instrument has various ranges, which are manually selected. The test voltage is generated with the help of a DC to DC convert or eliminating the need for a hand driven mechanism.

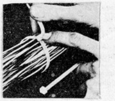
This model has four ranges viz. 20,200,2000 and 10000 M ohms with test voltages of 250 V DC, 500V DC and 1000V DC. The unit operates on 230V DC.



For further details contact:
M/S. ARUN ELECTRONICS PRIVATE LIMITED
B-125/126 Ansa Industrial Estate, Saki Vihar Road, Saki Naka, Bombay 400 072.
Phones: 583354/581524

CABLE-TIES

Novoflex have introduced RE-USABLE CABLE TIES which provide positive holding and permanent locking. The Ties remain securely locked until released by pushing the projection hear the locking head. Available for cable bundle diameter upto 106 mm. They maintain holding strength over a temperature from -40°C to +135°C.



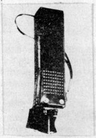
For further details please contact:
NOVOFLEX CABLE CARE SYSTEMS
Post Box No. 9159
Calcutta-700 016

WALKIE - TALKIE/ TRANSCIEVER KITS

Fiji Electronics offer complete know how and literature of assembling a walkie-talkie/Transceiver set with indigenous components.

The literature includes an application form and procedural details on how to obtain an experimental licence from the government to operate the transceiver.

The kit supplies PCB and only those components for the transceiver which are usually difficult to get like quartz crystals, filters, IFT, hardware etc.



Contact for further information:-
M/S. FIJI ELECTRONICS,
Mail Order Sales,
(WIT) Puthencurichy,
Trivandrum - 695 303.

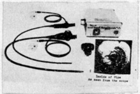
FIBERSCOPE

Mecord introduces Industrial Fiberscopes of Japan make.

The Fiberscope transmits a bright and clear image over a long distance in vivid colours.

The images can be observed not only by the naked eye but also through a TV camera.

The probe is flexible and small in diameter permitting inspection of normally difficult areas.



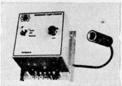
For Details Contact:
MECORD MARKETING PVT. LTD.
304, Hill View Industrial Estate, Off. L.B.S. Marg, Ghatkopar (W)
Bombay-400 086

TWILIGHT SWITCH

'IEC' Twilight Switch is an automatic light control device for switching lights "ON" automatically when the intensity of light falls below a preset level and conversely it switches "OFF" when the ambient light intensity exceeds a preset threshold.

The Twilight Switch is available in two models: OUTDOOR and INDOOR, with a choice of load ratings varying from 10 A to 50

Typical applications can be found in street/factory/courtyard lighting, housing societies, railway stations, airport runways, neon displays, cinema houses, etc.



For further information, Contact:-
INDIAN ENGINEERING COMPANY,
Post Box 16551,
Worli Naka,
Bombay 400 018

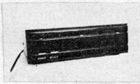
PRINTER DATE SWITCH

A new data switch, introduced for the first time in India by Kit Enterprises, allows one Centronics-type printer to be switched between two computers or one computer between two printers.

The switch supports the 36 pin Centronics Parallel interface used by many leading printer manufacturers. It is equipped with three 25 pin D shell connectors and is operated via a single front-panel switch.

Applications:

(a) To switch one computer between two printers. For example, to alternate a computer between a Daisy wheel (letter quality) and a Dot Matric printer (for high speed printing).



(b) To share one printer between two computers so that while one computer is printing the other can be used to run a different programme.

It eliminates the need to plug and unplug delicate printer cables which may cause damage to connectors if done too often.

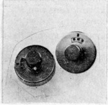
For enquiries contact
Kit Enterprises
18, Rebello Road,
St. Sebastian Colony,
Bombay - 400 050.
Tel: 642 9064

COUNTING DIALS

* 10-Turns Counting Precision Dials
* 100 dial div per turn
* Shaft diameter ¼" (⅛" and 3/32" - non-standard) available upon request.
* Two types 1¼" and 1⅞" outer diameter
* Available both locking/non-locking type

These dials are claimed to be highly accurate and easy to read with a long life. They can be mounted directly to the potentiometer shaft.

Dial backlash has been eliminated by locking the knob to the shaft and non-rotating base to the panel.



For further details please contact:
UNIROYAL
15, B/6 Silveria house, L.J. Road, Mahim
Bombay - 400 016

NEW PRODUCTS

SPIRALWRAP

MICROSIGN Spiralwrap (Flexible Protective Sheath) provides abrasion protection to wires - cables and tubings. Available in five different sizes it can be quickly installed and is re-useable too. It is used to bundle, harness, protect or insulate wires-cables tubings or house. Versatile by design and material it can be used in many applications and can be reused as and when revisions are required.

For further information contact:
MICROSIGN PRODUCTS
"Mehta Terrace"
Satyanarayan Road
Bhavnagar: 364 001.

SERVO CONTROLLED VOLTAGE STABILIZER

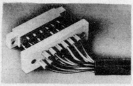
The Jivan is a Stepless Servo Controlled Voltage Stabilizer with output accuracy of + 1% against wide input range. The output level can be set between 220 to 240 V or 400 to 420 V. Manual operation is also possible. High and low voltages are idicated by the neon provided on the front panel. MCB is provided to protect against over load (Upto 10 KVA, 1 phase & 30 KVA, 3 phase). It is recommended for computer systems, laboratories, various industries, communication systems, CNC Machines etc.

For further details write to:—
M/S. JIVAN ELECTRO INSTRUMENTS,
394, G.I.D.C.
Makarpura,
BARODA - 390 010.

INTERCONNECTION GUARD

Nectar Electronics offer moulded polythene Guard for interconnections. The guard can be assembled with the connector or detached from it for tests and repairs with the help of nuts and bolts. The assembled guard improves reliability and appearance of electronic equipment. The number of interconnections and the pitch between them can be selected to suit various types of standard electronic connectors.

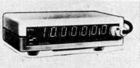
For more critical requirements, further protection from environmental effects caused by moisture, gases, vapours, fine dust particles or living organisms etc. can be provided by using more complex guarding systems incorporating rubber gaskets, grommets and other types of sealing.



For further information, please write to:
NECTAR ELECTRONICS
P.B. No. 5009
GPO Bangalore
Karnataka-560 001

DIGITAL FREQUENCY COUNTER VDC18

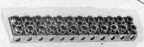
VDC18 is the smallest size over made in India. Features include BATTERY OPERATION cum mains operation through adaptor, 7digit 0.5 inch LED display,30 MHz frequency range, light weight, resolution selection etc.etc. VDC18 incorporate latest L.S.I. circuitry. Model VDC19 has frequency range upto 500 MHz and PERIOD, FREQUENCY modes.



for details contact:
VASAVI ELECTRONICS
(Marketing Division)
630,Alkarim Trade Centre,
Raniganj
Secunderabad-500 003
Phone: 70995

STRIP CONNECTORS

IEC Strip Connectors are available in wide range, from 5 Amps to 30 Amps in 12 ways, moulded in Bakelite & PVC. The metal parts are made of brass and screws of M.S. duly plated to prevent corrosion. The strip connectors are tested to withstand High Voltage for 2 K.V.



For further information, please contact:-
ASIA ELECTRIC COMPANY
Katara Mansion
132A, Dr.A.B. Road, Worli Naka,
Bombay 400 018.

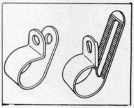
DIGITAL TACHOMETER

RC offers digital tachometers in two models - DT 4 (0-9999 RPM) and DT 5 (0-19999 RPM). At an accuracy of × 1 count and speed of one reading per second, it works within the temperature range of 10° C to 50° C. It has input impedence of 100 K Ohms. Readings are displayed on seven segment display. Measurement methods are magnetic pick-up (variable reluctance method, non contact type) and optical pick-up (contact type). Magnetic pick-up models have an input voltage protection for + 250 Volts.

For further details, contact:-
RC INFORMATION TECHNOLOGY SYSTEMS PVT. LTD.,
1413, Dalamal Tower,
Nariman Point,
Bombay 400 021.

P-CLIPS

This product can be used to secure a wide range of cables diameters because of its adjustability. Apart from this, a range of sizes is available in adjustable and non adjustable forms. The P-Clips are made from Nylon and all expected to withstand temperature from 35° C to 135°C.

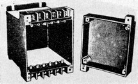


The P-Clip edges are radiused to prevent damages of cable insulation and hence find particular use in instrumentation and electronic equipment & appliances.

For further information, please contact:-
STARLITE ENTERPRISES
124B, Vivekananda Road,
Calcutta - 700 006.

PLASTIC INSTRUMENT BOX FOR BACK MOUNTING

Comtech T-77' is an elegently designed plastic moulded instrument box suitable for back mounted instruments such as Timers, & various other control instruments, having overall dimensions of 110 mm. l x 77mm W x 100 mm. B. It consists of a moulded box, a cover, & a M.S. plate for back mounting. The box has an inside space of 73 mm. x 71 mm. for various components. A six way terminal strip fixed at the top & bottom, infornt of the box provides a easy access for the terminals. The cover can accommodate a PCB of 77 mm x 72 mm. from inside & has a 1.2 mm deep recess in front to take an Aluminium plate of 65 mm x 66 mm. for control indications. The box offered in Black & Grey colour with either Glossy or Matt finish, is most suitable for small instruments to be mounted side by side from the back, like e.g.. counters, controllers & timers etc.



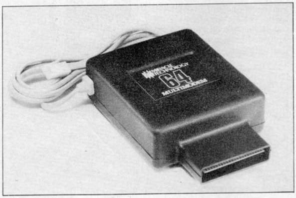
For further details contact:
COMPONENT TECHNIQUE
8, Orion Appartment
29-A Lallubhai Park Road,
Andheri (West)
Bombay - 400 058.

NEW PRODUCTS

Plug-in modem for Commodore 64 and 128 computers

Miracle Technology Ltd have introduced their new 64 MULTIMODEM, which gives Commodore 64 and 128 owners access not only to Prestel, Micronet, Microlink and viewdata services, but also to databases, bulletin boards, electronic mail, telex and user-user communications. The modem features autoanswer, autodial, and on-board software in ROM. Menu-driven and multi-speed, it supports the CCITT V21/23 and Bell 103 standards, handling baud rates of 300/300, 1200/75 and 75/1200. Functions include save and print frame, automailbox with edit and save and telesoftware downloading. The unit fits in the computer's cartridge port, and has only one external connection — the telephone lead.
At £98.50 exc. (£116.15 inc VAT & UK delivery), the 64 MULTIMODEM puts comprehensive data communication within the reach of C64 and C128 owners. BABT approval is expected shortly.

Miracle Technology (UK) Ltd.
St Peters Street
Ipswich IP1 1XB.
Telex: 946240 CWEASY G 19002985 (3417:4:F)



A safe and highly reliable DMM

Harris Electronics have announced the availability of a new hand held Digital Multimeter, designed with safety and reliability in mind. TMK model G44 is housed in a rugged plastic case with integral tilt stand and safety sockets.
A large 0.5 inch LCD display clearly indicates the 0.25% basic DC voltage accuracy to 1000V and AC voltage range from 100μV to 750V. AC and DC current ranges are specified as 100mA to 20A and resistance can be measured from 100 milliohms to 20 megohms.

The G44 gives approximately 2000 hours of use when fitted with a single 9V alkaline battery. A low battery warning has also been incorporated, along with full overload protection on all ranges.
The G44 is priced at £49 and comes complete with battery, safety style test lead set and a comprehensive operator's manual.

Quiswood Ltd
21 Eastbury Court
Lemsford Road
St Albans
Herts. AL1 3PS (3417:10:F)

New lithium battery

Venture Technology has launced a new 3-volt lithium-manganese dioxide battery, the LiM3512E. This battery has a capacity of 27 000 mAh when discharged at 275 mA at 25 °C. It has a diameter of 35.8 mm and a height of 128 mm.

Venture Technology Limited
18 Nuffield Way
Abingdon
Oxon OX14 1TG
Telephone: (0235) 20502
Telex: 837887 (3382:13:F)

New versatile switching regulator chip

National Semiconductor has recently introduced the LM1578, a switching regulator that generates a positive or negative voltage from one positive supply. The LM1578 can be set up for dc-to-dc voltage conversion circuits such as the step-down, boost, and inverting configurations. The output can switch up to 750 mA while output pins for its collector and emitter have been provided to promote design flexibility. Also the LM1578 has a 1% on-chip oscillator and an external current limit terminal.

National Semiconductor (UK) Ltd.
301 Harpur Centre
Horne Lane
Bedford MK40 1TR.
Telephone (0234) 47147
Telex 8 26 209 (3417:7)

classified ads

CORRECTIONS

High-power AF amplifier — 1

in this issue

Resistor R_{58} should be a 27 k, 1% type, as indicated in the circuit diagram Fig. 3.

Telephone exchange

(January 1986)

Capacitors C_{21} and C_{22} have been shown with the wrong polarity in the component overlay, Fig. 3. Also R_6 in the parts list should read R_{62}.

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