109

video display
short-wave pocket radio
5 amp switching power supply
mini crescendo

contents

The most obvious thing about this month's front cover is the colour. It is a long time since we used a black background on a cover but we felt that this would show off the colours of the analytical video display to the best effect. Although primarily intended for use with the real-time analyser (whose LED display is seen in the background) this display can be used for a multitude of applications. Your own imagination is the only limiting factor as there are plenty of situations where a circuit could benefit from having a colourful bar-graph This month's contents also includes something for those who have no interest in displays, coloured or otherwise. How about assulting your aural senses with a short wave pocket radio or the mini crescendo power amplifier. Not interested? What about a switching power supply, or an EPROM copier, or why not just read the magazine anyway. It is interesting; you can take our word for that.

A selection from next month's issue:

HAPPY BIRTHDAY ELEKTOR INDIA.

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SATISH MENON
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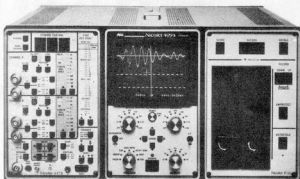
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BOOST TO COMMUNICATION INDUSTRY

Manufacture of telecommunication equipment, telephones and other gadgets at subscribers' premises, reserved for the public sector companies so far, has been now thrown open to the private sector. This significant policy decision announced by the government recently has overwhelmed not only the industry but the consumer too. The monotony of black telephone instruments will be wiped out with a plethora of modern telephones of different shapes, shades and sizes. Competition in the indigenous industry will result in maintaining the high quality of products.

Though the government has to decide on the specifications, test procedures and the type of apparatus to ensure their compatability with the Indian telephone systems, private sector entry in the field of telecommunication will give a filip to the electronics industry, for there is no communication without electronics.

The Association of Indian Engineering Industries has already presented a proposal to the government for establishing a network of earth stations throughout the country and if the government allows the private sector to set up earth stations also, satellite communication is bound to leap-frog. As it were the association has sought some channels from the INSAT-1B as only 1,400 channels out 8,000 were being used for telecommunication.

CIT-ALCATEL—FURTHER DEVELOPMENTS

The Indian electronic industry has more reason to be delighted. A high-level French delegation has visited the country and identified a number of components to be purchased for the CIT-Alcatel digital trunk automatic exchanges. CIT-Alcatel has become a common name in India, thanks to the Indian Telephone Industries' tie up with the French firm for the manufacture of digital automatic exchanges. While the CIT-Alcatel is expected to procure 16 per cent of the components from the French industries, Indian contribution can be as much as 12 per cent, the rest going to Japan, USA and other countries. The complete list of items to be procured from India for the project will be ready by October and two more French delegations are scheduled to visit India to assess the quality and study the capacity of the Indian firms to meet the French requirements. The Thomson group of France also represented in the delegation, has shown interest in offering sophisticated technologies to Indian firms with in-house R & D facilities.

TELEMATICS – PROGRESS

The proposed Centre for the Development of Telematics, to be funded jointly by the Union ministry of communications and the department of electronics, is expected to develop an indigenous digital switching technology within 36 months at an investment of Rs. 35 crores.

Since there was no need for updating the CIT-Alcatel technology, the centre has to develop an original digital technology for the third electronic switching factory to be set up in the seventh plan period.
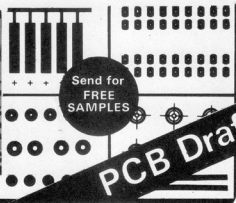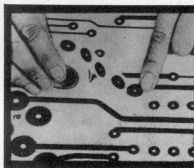
The government has clarified that the telematics centre will be vested with total authority and flexibility to function outside the government norms. It will be on the lines of a scientific society. In addition to 250 engineers and scientists in the country, the government might call upon expatriate Indians to come and work for the centre.

The centre would attempt to develop entirely an indigenous technology for augmenting the production of various telematic services like data transmission, fascimile telex, videotex and electronic mail, in a phased manner, according to Dr. M.S. Sanjeevi Rao, deputy minister for electronics.

INDIA CAN DO IT

India can match Japan in electronics in 10 years if suitable duty concessions are given on raw materials and production equipment, says Dr. U. Venkataswaralu, managing director of the state-owned Central Electronics Limited.

The 40 per cent duty on electronic products and raw materials and the 100 per cent duty on manufacturing equipment should be withdrawn. The duty on components should also be withdrawn gradually, according to Dr. Venkataswaralu.

The CEL has developed a private automatic branch exchange (PMBX) which runs on solar energy. This computerised system can be operated on mains as well.

The system with 48 or 96 subscriber lines can be extended up to 192 lines. A 100-line system working on solar energy approximately costs Rs. 3.25 lakhs and each additional line may cost Rs. 2,500, according to the CEL spokeman. The system working on mains costs about Rs. 2.50 lakhs.

BEL TO MAKE LCDS.

Bharat Electronics Limited, Bangalore, will manuacture five lakh Liquid Crystal Displays per year, based on indigenous technology,for use in electronic watches and computers. After a long wait, the company has procured the letter of intent for manufacturing the LCDs. Its existing pilot plant with a capacity to produce a million LCDs a year will be expanded for commercial production.

ELCOT CAPACITOR UNIT

The Electronic Corporation of Tamil Nadu (ELCOT) has set up an aluminium electrolytic capacitors plant at the industrial complex in Hosur, Dharmapuri district. The Rs. 1.68 crore plant employing sophisticated machinery from Japan has a licensed capacity to produce 50 million capacitors per annum. These precision capacitors, claiming conformation to international standards, can be used in radios, transistors, television sets, public address systems and various electronics industrial testing equipment.

ECIL'S FUTURE PLANS

The Electronics Corporation of India Limited will manufacture computer main frames based on foreign technology, following a government decision to this effect. The country now spends at least Rs. 30 crores on the import of computer main frames. A French and an American firm have been short-listed for buying the technology.

The ECIL, while maintaining its 10 per cent share in consumer electronics, would henceforth concentrate on three areas, namely control systems, computers and communication equipment, according to Mr. S.R. Vijayakar, chairman and managing director of the ECIL.

Mr. Vijaykar said the corporation would spend about Rs 10 crores on R & D during the seventh plan period. He agreed that the investment in R & D had been stagnant, affecting the growth of the industry.

The ECIL would manufacture nearly 30,000 colour TV sets this year, though it claims to have procured ordrs for 40,000 sets. The Tirupati unit of the ECIL has since begun production of B & W TV sets.

The corporation has been engaged in the development of computerised telex echange systems and supply of hardware for freight information system for wagon movement. Another futuristic project of the corporation is the development of roof top terminals for satellite communication, which may cost Rs. 2 lakh per terminal. Data acquistion systems will be installed by the ECIL in thermal power plants at Korba, Ramagundam, Obra and Raichur.

Noise is by definition any undesired sound. While an f.m. receiver has less
noise on a station than an a.m. receiver, it is noisy between transmissions.
In the absence of an incoming signal, a loud hissing sound is heard: this
is generated by the noise voltages in the receiver. A received signal is
amplitude-modulated by the noise and the amount of modulation is a
function of the amplitude of the noise voltage. It is therefore desirable
that f.m. receivers have a noise-squelch circuit which mutes the a.f. amplifier
when no signal is being received. If your receiver is not fitted with a squelch,
the circuit described here may be just what you've been waiting for . . .

noise squelch . . .

. . . for f.m.
receivers

To keep the circuit simple, our design is
based on an old, proven principle: that of
controlling the squelch with a carrier-
dependent signal. To make a suitable control voltage
is available in virtually all f.m. receivers. In
the absence of an incoming carrier, this
voltage drops and actuates the squelch
which breaks the audio path. A good
squelch circuit should react fast and not
produce any switching noise. These are
contradictory requirements, because rapid
on/off switching of the a.f. signal causes
clicks in the loudspeaker. Slow on/off
switching on the other hand leads to loss
of information at the time of switch-on
as well as to noise trails on switch-off. A
practical squelch circuit is therefore a
compromise whereby a small amount
of switching noise is accepted.
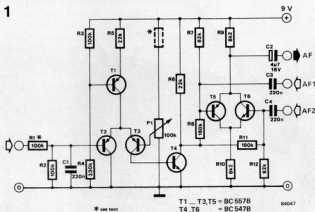
The circuit

A quick glance at the figure will show that
the circuit is pleasantly uncomplicated.
Transistors T1 . . . T3 form a differential
amplifier which functions as a comparator to
provide an adjustable threshold voltage.
When no voltage is present at the control
input, T2 conducts irrespective of the
setting of squelch sensitivity control P1.
Transistors T3 . . . T5 block so that emitter-
follower T5 does not pass the a.f. signal
present at input AF1. Transistor T6 con-
ducts, however, so that a second signal at
AF2 is applied to the AF output. If, for
instance, a cassette recorder is connected to
AF2, music may be listened to during
intervals in reception*. This is akin to the
facility in certain de luxe car radios which
causes cassette playback to be automatically
interrupted by traffic broadcasts. Note that
resistors R9 and R10 must be of the same
value to ensure balanced operation.
When a signal is present at the control
input that causes the voltage at the base of
T2 to be higher than the squelch sensitivity
level set by P1, T2 blocks. Transistors
T3 . . . T5 then conduct so that T5 connects
the a.f. signal from the receiver (AF1) to
the AF output. Transistor T6, however, is
cut off and prevents the AF2 signal from
reaching the AF output.

Practical hints

The signal at AF1 is the a.f. output from the

1

T1 . . . T3, T5 = BC 557B
T4, T6 = BC 547B

84047

receiver or tuner. In receivers it is best to
unsolder the a.f. input line from the volume
control and connect it to AF1. The AF
output of the squelch circuit is then con-
nected to the freed terminal on the volume
control.
Where the signal for the control input of the
squelch circuit is taken from depends on
the receiver. If this has a field-strength
indicator or automatic gain control (a.g.c.),
the control signal may be taken from these.
If not, most current a.f. amplifier ICs
provide a carrier-dependent voltage: the
TBA 120, TBA 120S, SO 41P, and TCA 420,
for instance, at pin 8, and the CA 3089E and
CA 3189E at pin 13. In receivers built
from discrete components, the signal may
conveniently be taken from the relevant
points in the discriminator or limiter circuits
as appropriate.
The squelch sensitivity may be set over a
wide range of control voltages, 0.2 . . . 15 V,
which should be adequate for most purposes.
If you want a sensitivity better than 200 mV,
take a smaller value for R1 and connect a
resistor of corresponding value in series
with P1 (as shown in dashed lines on the
circuit diagram).
The current consumption of the squelch
circuit is about 3 mA and this should present
no difficulties even in a battery operated
receiver.

A noise-squelch circuit
mutes the a.f. output of
a receiver unless a signal
of predetermined strength
is received.

A tone-squelch circuit
mutes the a.f. output of a
receiver until a signal is
received which is continu-
ously modulated by a pre-
determined tone.

A muting circuit auto-
matically actuates a noise-
suppression circuit when
the noise exceeds a pre-
determined level.

Squelch sensitivity is a
measure of the minimum
input voltage required
for suppression of the
receiver's internally
generated noise.

* In CB operation, for
instance.

...

Broadcasting on short waves is confined to a number of relatively narrow bands, usually called wavebands. Although quite a number of high-quality channels (frequency separation = 9 kHz) can be accommodated in each band, there are so many short-wave stations that satisfactory reception is frequently impossible unless you have a very selective receiver.

The use of specific frequencies for long-distance transmission is determined largely by ionospheric conditions which, in turn, depend on the eleven-year sunspot cycle. Schedules of transmission times and frequencies are worked out with reference to the ionospheric conditions at particular times of the day and year and are published by most broadcasting organizations.

The receiver is a double superhet with preselector. In superheterodyne radio reception the incoming signal is mixed with the signal from a local oscillator. This results in a so-called intermediate frequency (i.f.) signal which is equal to the difference between the locally generated signal and the carrier signals and containing all the original modulation. The i.f. signal is then amplified and demodulated. The demodulated signal is fed to an audio frequency (a.f.) amplifier. A double superhet employs two intermediate frequencies which improves the overall performance. In this, the first i.f. signal is mixed with a second locally generated signal before amplification and demodulation.

A preselector improves the sensitivity and selectivity of the receiver: it usually is a tuned radio frequency (r.f.) amplifier which amplifies the incoming signal before it is mixed with the signal from the local oscillator.

The BBC, in common with many other broadcasting organizations, operates in the 49-metre band on a dozen or so different frequencies. A signal on one of these frequencies, say 6090 kHz, is picked up by the aerial (see figure 1). The aerial and following r.f. amplifier together form the preselector which is tuned with a variable capacitor.

The r.f. signal is then mixed with the signal from the first local oscillator, 16.8 MHz. The two products at the mixer output, 22 890 kHz and 10 710 kHz, are applied to a 10.7 MHz band-pass filter which suppresses the higher frequency.

The 10.7 MHz signal is amplified in the first i.f. amplifier and then mixed with the signal from the second oscillator, which in this case is tuned to 10 245 kHz. The resulting second i.f. of 455 kHz is applied to a 455 kHz band-pass filter and then amplified in a second-i.f. amplifier. The gain of the i.f. amplifiers is controlled by the automatic gain control (a.g.c.) which holds the level of the input signal to the demodulator substantially constant despite variations in the received signal strength.

The 455 kHz signal, which still contains the carrier and two sidebands containing the original modulation, is demodulated
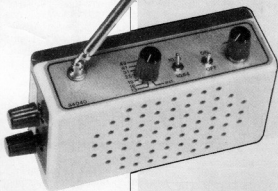
short-wave receiver . . .

. . . the world
in your pocket

(that is, the carrier and one of the sidebands — usually the lower — are removed). The resulting a.f. signal is amplified and then applied to the loudspeaker.

Tuning to a different station is effected by changing the frequency of the second local oscillator.

The circuit

As in most pocket (or portable) radios the aerial is a telescopic rod which is coupled to tuned input circuit L1a/C1 to form the preselector. The preselector tunes the receiver coarsely to the required frequency band (see table 1). The active part of the input stage consists of source follower T1 which passes the aerial signal to the first mixer, IC1, with unity gain. Integrated circuit 1 is a symmetrical mixer for frequencies up to 200 MHz and is driven by an on-chip oscillator. Mixing is carried out by two differential amplifiers the characteristics of which ensure that none of the original signals appears at the output (pins 2, 3, and 5) of IC1.

The internal oscillator is driven by one of a number of crystals, X1 . . . X7, which are switched in by BAND SELECTOR S1, or by an EXTernal VFO signal. Although our design offers seven crystal-controlled bands, it can be expanded with the use of the external source, or by adding suitable crystals, to cover other short-wave bands (see table 1). Note, however, that the printed-circuit board in figure 3 only caters for seven crystals.

The output of the mixer is inductively coupled (L2b and L3b) to IC2 via four 10.7 MHz filters, F1 1 . . . F1 4. It may be seen from table 1 that the bandwidth required varies from 50 kHz to 500 kHz. Unfortunately, we have not been able to source a 10.7 MHz filter with 500 kHz bandwidth. The use of change-over switch S2 offers a solution to this problem, however, because the filters given in the parts list have a centre frequency of 10.76 MHz (F1 1 and F1 2) or 10.64 MHz (F1 3 and F1 4) respectively. Furthermore, the filter tolerance is ±30 kHz. These deviations from 10.7 MHz have no effect whatever on the overall performance of the receiver. As the filters have a 3 dB bandwidth of 280 kHz, the total bandwidth covered is 460 kHz (10.47 . . . 10.93 MHz). The upper and lower part of this bandwidth must, of course, be selected with S2.

Integrated circuit 2 is a monolithic a.m. receiver for operation up to 50 MHz. It includes an r.f. amplifier, a balanced mixer, an oscillator, and an i.f. amplifier. The 10.7 MHz input signal is amplified in the r.f. amplifier and then applied to the mixer, which is also fed with the signal from the (internal) second local oscillator. The oscillator is driven by tuned circuit L5/D1. The diode is a varactor (acronym for variable reactor) which is operated with reverse bias so that it behaves as a voltage-dependent capacitor. The variable voltage emanates from potentiometer P1. The 455 kHz output of the mixer is applied to tuned inductor L6 and then to bandpass filter F1 5: the inductor ensures correct impedance matching to the filter. The ceramic second-i.f. filter, which has a 6 dB bandwidth of 6 kHz, removes the last traces of spurious signals. The 455 kHz signal is then applied to a four-stage amplifier in IC2 from where it is taken to a further filter, L7. The signal at L7 consists of the 455 kHz carrier with superimposed on it the two a.f. sidebands.

The signal from L7 is then demodulated by diode D2. The demodulation process is, firstly, the rectification of the carrier to eliminate the negative half-cycles (essential because the positive and negative half-cycles tend to cancel one another), and, secondly, the removal of the carrier-frequency variation in order to leave only the a.f. modulation. The result is that only a (weak) a.f. signal is present across C20. Part of the rectified voltage at the cathode of D2 is fed back to IC2 (pin 9) via filter R10/C18 for use as automatic gain control. An internal a.g.c. amplifier controls the gain of three of the four i.f. stages and of

Figure 1. The block diagram of the short-wave pocket receiver which shows its configuration as a double superheterodyne with intermediate frequencies of 10.7 MHz and 455 kHz.
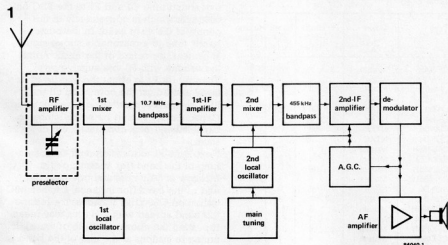

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Table 1

Short-wave band (m)	Frequency (kHz)	Band-spread (kHz)	Crystal frequency (fundamental) (kHz)
120	2300 — 2495	195	13 150
90	3200 — 3400	200	14 050
75	3950 — 4000	50	14 650
60	4750 — 5060	310	15 700
49	5950 — 6200	250	16 800
41	7100 — 7300	200	17 950
31	9500 — 9775	225	20 350
25	11 700 — 11 975	275	22 550
19	15 100 — 15 450	350	25 950
16	17 700 — 17 900	200	28 550
13	21 450 — 21 750	300	32 300
11	25 600 — 26 100	500	36 750

the r.f. amplifier (the latter via external filter R8/C17).

The a.f. signal is amplified in IC3 to a level sufficient for driving the loudspeaker.

As is usual with pocket radios, power is derived from a 9-volt PP3 battery. Only IC3 is supplied directly from the battery: the other stages are powered by 5 V regulator IC4. Current consumption amounts to about 25 mA in the absence of an incoming signal so that a PP3 allows 24 hours continuous service as long as the volume is not set too high.

The design already incorporates provisions for those who want to use the receiver in a stationary location or add certain facilities. An additional turn on the toroidal core of the aerial inductor will allow a long-wire aerial to be used. The possibility of driving the first local oscillator from an external source has already been mentioned. There is a facility to tap off the 455 kHz second-i.f. from C19: if this is mixed with a BFO (beat-frequency oscillator) signal, SSB (single-sideband) reception becomes possible. If the 455 kHz signal is applied to a phase discriminator, reception of morse and other telegraphy signals becomes possible.

The specifications of the filters shown in the table relate primarily to the shape factor. The best filter is that which has the widest 6 dB bandwidth (BW) coupled with the narrowest 60 dB bandwidth. Make sure, however, that the 6 dB bandwidth is suitable for the type of transmission: broadcast signals require a bandwidth of about 6 kHz, communications about 3 . . . 4 kHz, and SSB signals around 3 kHz. The 60 dB bandwidth is a measure of the adjacent-channel (received carrier frequency ± 9 kHz) rejection which for best selectivity should be at least 60 dB. The insertion loss of these filters is 5 . . . 7 dB.

Table 2

Type	Maker	6 dB BW (kHz)	60 dB BW (kHz)	Housing (figure 4)
CFW455HT	Murata	6	12	a
CFW455IT	Murata	4	18	a
LF-H6S	NTKK	6	22	a
LF-H4S	NTKK	4	18	a
LF-H6	NTKK	6	22	b
LF-H4	NTKK	4	18	b
CLF-D6	NTKK	6	12	c
CLF-D4	NTKK	4	8	c
CFK455H	Murata	6	12	c
CFK455I	Murata	4	8	c
CFK455H	Murata	6	12	c
CFK455I	Murata	4	8	c
CFK455I	Murata	4	8	d
SFK-D6	NTKK	6	11	d
SLF-D6	NTKK	6	11	d
SLF-D4	NTKK	4	8	d
CFG455H	Murata	6	7	d
CFG455I	Murata	4	7	d
CFK455H	Murata	6	10	d
CFK455I	Murata	4	7	d

Construction

Before you switch on your soldering iron, consider first which bands you want to receive.

Note, however, that the pc board has provision for seven crystals only.

Secondly, there's F1 5 to be considered. Our design uses a compromise between selectivity and tone quality, but if you want better selectivity (at the cost of sound quality) you may choose a filter with a different shape factor from table 2. Thirdly, the case: the parts give a specific one, which accommodates the pc board nicely. It's left to your own preference how and where (within reason!) to fit the operating controls, the loudspeaker, telescopic aerial, and battery. In our prototype we have situated C1 and P1 in the side of the lower moulding, and the remaining controls in the top moulding (see photographs). The connectors for EXT. ANT, EXT. VFO, and IF OUT must, of course, also be located in the case, if required.

Before you start wiring up the pc board, wind inductors L1 . . . L4. Although initially you may need only L1a, L2a/b, and L3a/b, it is prudent to wind and fit them all, so that if at a later date you want to expand the receiver, you don't have to remove the pc board.

For the same reason, it is wise to fit all seven crystal holders.

Two important points: first, do not forget to solder the earth terminals of the appropriate components nor the metal cans of inductors L5 . . . L7 to the earth plane at the component side of the pc board; second, if you use a 455 kHz filter with metal case (housings c and d in figure 4), the lug on the case must be kept away from the board, for instance, by bending it upwards.

Calibration

Pull out the telescopic aerial to its fullest extent and tune C1 and P1 to the BBC on 6090 kHz which is conveniently in the centre of the 49 m band. In the other bands tune to a reasonably strong signal at or near the centre of the band. Adjust the volume with P2 to a suitable level. Then return P1 to about the middle of its travel and adjust the core of L5 till the BBC, or whatever station, is tuned in again. Capacitor C14 should be set to minimum capacitance (rotor completely out of stator).

Next, tune P1 to a station near the beginning of the band (for instance, radio Moscow on 5950 kHz) and one near the end of the band (for instance, another BBC station on 6180 kHz). If necessary, reduce the band spread with C14 (this may mean repeating the above a couple of times). If tuning to stations at the end of the band is not possible, the 10.7 MHz filter used is
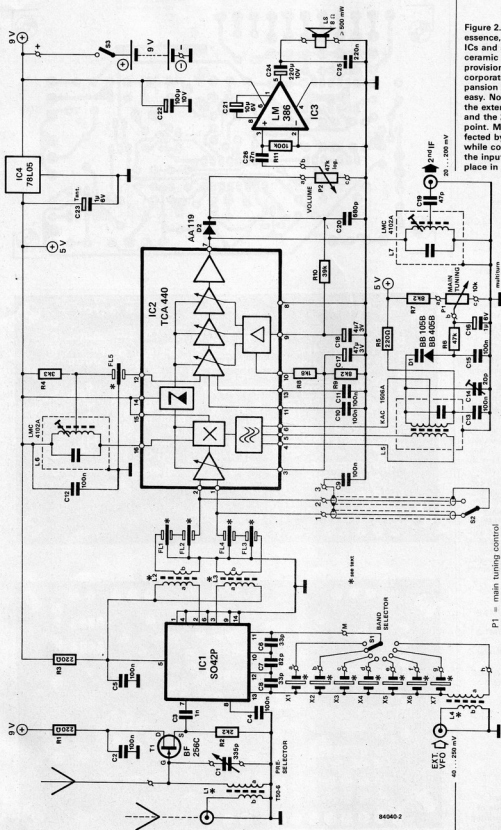
Figure 2. The circuit is, in essence, based on three ICs and a number of ceramic filters. Various provisions have been incorporated to make expansion at a later stage easy. Note particularly the external VFO input and the 2nd-i.f. take-off point. Main tuning is effected by a varactor, while coarse tuning of the input signal takes place in a preselector.

P1 = main tuning control
P2 = volume control
C1 = coarse tuning control
S1 = band selector
S2 = band-spread switch
S3 = on/off switch

* see text

short-wave receiver . . .

3

Parts list

Resistors:
R1,R3,R5 = 220 Ω
R2 = 2k2
R4 = 3k3
R6 = 47 k
R7,R9 = 8k2
R8 = 1k8
R10 = 39 k
R11 = 100 k
P1 = 10 k, 10-turn potentiometer
P2 = 47 k logarithmic potentiometer possibly incorporating S3

Capacitors:
C1 = 335 p variable capacitor
C2,C4,C5,C10 . . . C13,C15 = 100 n
C3 = 1 n
C6,C8 = 33 p
C7 = 82 p
C9 = 100 n only if S2 is used
C14 = 20 p trimmer
C16 = 1 μ/16 V
C17 = 1 μ/3 V
C18 = 4μ7/3 V
C19 = 47 p
C20 = 680 p
C21 = 10 μ/6 V
C22 = 100 μ/10 V (recommend new component here!)
C23 = 1 μ/6 V tantalum
C24 = 220 μ/10 V (recommend new component here!)
C25 = 220 n
C26 = 47 n

Inductors:
L1 = 25 + 1 turn (separate!) enamelled copper wire SWG 30 on toroidal core type T50/6 (available from Ambit)
L2,L3,L4 = 4 + 2 turns (separate!) enamelled copper wire SWG 34 on ferrite bead size 3.5 x 3.5 mm
L5 = KAC1506A (Toko)
L6,L7 = LMCS4102A (Toko)

Filters:
FI1,FI2 = CFSH10.7M1-E (white dot) (Toko)
FI3,FI4 = CFSH10.7M1-D (black dot) (Toko)
FI5 = CFG465H (Murata) or SLFD6 (NTKK) (see table 2 if you require a filter with a different shape factor)
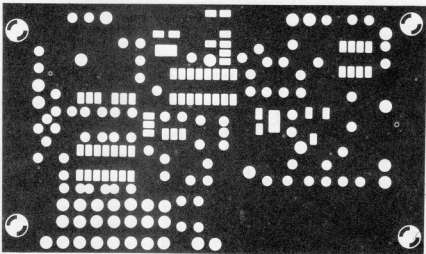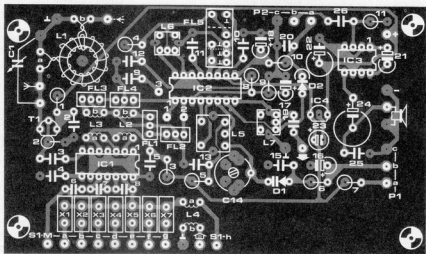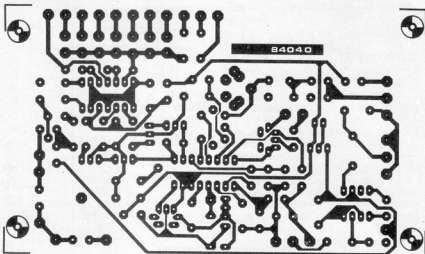
Figure 3. Component layout and track side of the double-plated printed-circuit board. Try not to commence construction until you have read the text!

all views from underneath
all dimensions in mm

84040-4

Figure 4. Dimensions of housings and pin-outs of 455 kHz filters shown in table 2. All these filters can be accommodated on the printed-circuit board but make sure that they are positioned correctly.

Quartz crystals:

(fundamental frequencies)
X1 = 16 800 kHz
X2 = 17 950 kHz
X3 = 20 350 kHz
X4 = 22 550 kHz
X5 = 25 950 kHz
X6 = 28 550 kHz
X7 = 32 300 kHz

too narrow, or S2 has to be switched over. Alignment of L6 is a little tricky. With a spectrum analyser it is quite easy to adjust the core until the two peaks of the response curve just become one without a trough. But who has a spectrum analyser? Experienced short-wave listeners can hear the separate peaks and tune accordingly (push aerial fully home). Incorrect adjustment of the core of L6 results in a deterioration of the sound quality and inexperienced users may therefore align the core for best sound quality.

A tuning indicator may be provided by connecting a 370 µA meter (internal resistance 1500 Ω) between pin 10 of IC2 and earth.

The core of L7 is aligned by adjusting it until the received station is loudest.

Good luck and good reception! ◄

Semiconductors:

D1 = BB 105 or BB 405B
D2 = AA 119
T1 = BF 256C
IC1 = S042P
IC2 = TCA 440
IC3 = LM 386
IC4 = 78L05

Miscellaneous:

S1 = rotary wafer switch, for number of positions see text
S2 = change-over switch, single pole
S3 = single-pole on/off switch (may be combined with P2)
LS = miniature loudspeaker, 8 Ω, 0.5 W
PP3 battery complete with connector
Case 150 × 80 × 50 mm
Telescopic aerial
Printed-circuit board 84040
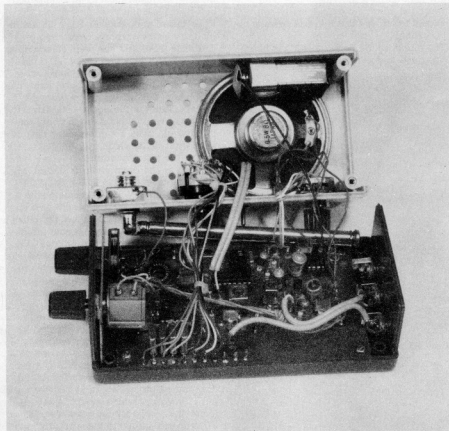
Floppy-disk drives are fast becoming commonplace with the home computer. They are of necessity a fairly complex and finely engineered piece of equipment and, thankfully, they are reasonably reliable. However, in keeping with today's world, they do fail on the odd occasion. The problem is that, when a failure occurs, it can be difficult to establish whether the fault lies with the floppy-disk drive or somewhere else in the system — be it hardware or software. The circuit here provides a quick and simple means of checking the floppy-disk drive for all operating modes.

floppy tester

fault finder for floppy-disk drives

A floppy-disk drive is probably the most expensive single item of hardware (with the possible exception of the computer itself) that will be purchased for many home computer systems. It is of necessity a rather delicately engineered piece of equipment and should therefore not be dismantled without due cause. None the less, problems can and do occur but, before taking a screwdriver to the disk drive, it is important to establish exactly what the fault symptom is and if, indeed, any fault exists at all. This is one time when the computer cannot really help — except to tell you that the disk drive is not operating!

The circuit of the floppy tester has been designed to provide all the operating con-

ditions the disk drive unit requires and, at the same time, to monitor the responses the disk drive makes to the interface board. All the operating conditions.are under manual control so that a thorough check of the disk drive electronics and mechanism can be carried out very quickly.

The circuit diagram may come as a pleasant surprise to readers who are expecting an extensive array of ICs. In this case, simple is best — and effective. Initially, the tester must determine which of a possible number of disk drives are to be tested and switches S3 . . . S5 are included for this purpose. Once the appropriate drive unit has been selected, its motor can be switched on by S6. If the diskette currently in position has a write protect tab in place LED D3 will light. The purpose of this LED is to test the write protect circuit in the floppy-disk drive. It will be as well at this stage to verify that the diskette does not contain any data of significant value which may be destroyed during the test procedure.

As soon as the drive motor is started, LED D1 will light — or, to be more accurate, flash at the rate of 300 per minute with a 5¼ inch drive or 360 per minute with an 8 inch drive. This indicates that the index marker in the disk drive is functioning correctly. If, on the other hand, this LED emits a steady light (or doesn't light at all), the index circuit in the disk drive is at fault. This may be caused by 'foreign bodies' obscuring the photodetector.

LED D2 lights to indicate that the read/write head is positioned above track 0Ø. Movement of the head is effected by two switches, S1 and S7. It will be seen that switch S1 controls the pulse generator (a monostable multivibrator) MMV1 via a debounce circuit consisting of an RS flip-flop formed by gates N1 and N2. The output of MMV1 provides pulses which are fed to the head-position stepper motor drive control circuit in the disk drive unit.

N1 . . . N4   = IC1 = 74LS37N
MMV1,MMV2 = IC2 = 74LS123N
N5 . . . N9   = ¼ IC3 = 74LS05N

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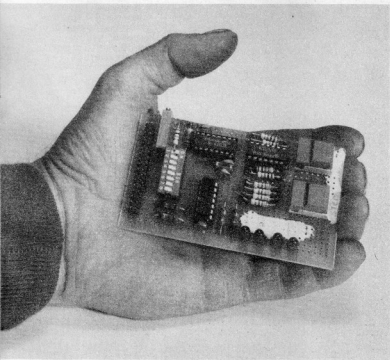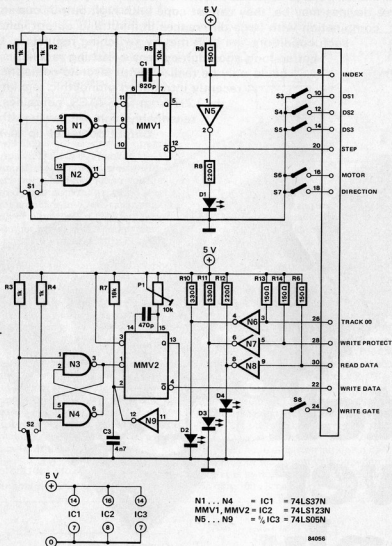
Figure 1. Although simple, the circuit of the floppy tester is able to simulate all the conditions required to operate a disk drive. Head control is provided by MMV1 while MMV2 forms a data generator.

D1 = index
D2 = track 00
D3 = write protect
D4 = read data
S1 = step selector (grey)
S2 = write data control (red)
S3 . . . S5 = drive unit selector
S6 = drive motor switch
S7 = head direction switch
S8 = read/write switch

Each pulse moves the read/write head one step on to the next track. The position of switch S7, open or closed, will dictate whether the head movement is inwards or outwards respectively. These two switches thus provide the means by which a complete check of the head movement mechanism can be carried out.

The rest of the circuit concerns the reading and writing of data into or out of the disk drive. Switch S8 creates the read or write command that would normally originate in the interface board. Data from the diskette is 'read' by LED D4 which will flicker if data is present. For this test it is necessary to have some data on the diskette or D4 will remain off with possibly misleading conclusions.

The floppy tester would not be complete without some means of writing data into

the diskette. The 'data' generator consists of an oscillator formed by monostable multivibrator MMV2 with inverter N9 in the positive feedback loop. The 'data' itself is a train of pulses with a pulse width of about 500 ns and a pulse spacing of approximately 8 µs. The pulse repetition frequency can be adjusted by preset P1. The data flow is switched on or off by switch S2 via another RS flip-flop formed by gates N3 and N4. This data generator has proved invaluable for fault finding with an oscilloscope inside the disk drive unit itself.

No parts of the circuit should give constructional problems. The only critical component is the output connector which must of course be compatible with the floppy-disk drive to be tested.

The design and construction of good-quality, regulated mains power supplies became virtually simplicity itself with the arrival of the now well-known voltage regulators series 78XX and 79XX. Good though these devices may be, they cannot cope with high output currents in combination with large differences in input and output voltage: such conditions are best met by switching regulators. Until not so long ago, high-current switching regulators could only be realized with discrete components. A recently introduced monolithic regulator, the L296 from SGS-ATES, combines a reasonably high efficiency with an output power of up to 160 W.
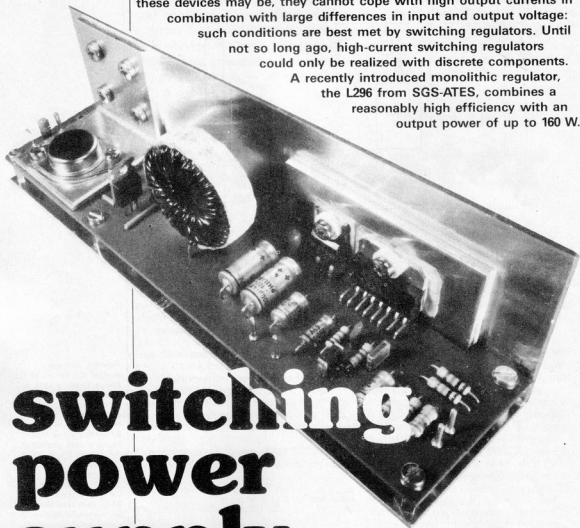
switching power supply...

. . . with high-current regulator

A voltage regulator produces a very stable output voltage from a fluctuating input voltage. The principle is fairly simple: the output voltage is fed back to the input and compared with the required value and any difference between real and required values drives a control circuit. The big drawback of linear voltage regulators is that they dissipate considerable power. The reason for this is that the difference between input and output voltage is dropped across the regulator. Multiply this with the load current (which, of course, flows through the regulator) and you have a nice little heatsource!

Switching power supplies generally function as shown in figure 1. An error signal is produced by comparing the output voltage with a precise 5.1 V on-chip generated reference voltage (U_{ref}). The error signal is then compared with the 20 ... 200 kHz output of the sawtooth generator. The output of the comparator is a pulse-width modulated rectangular wave which is applied to the driver and output stage. The pulse width is dependent on the direct voltage output of the error amplifier.

The collector of the output transistor is connected periodically with the supply voltage, U_s. This voltage is arbitrary as long as it is higher than the required output voltage plus the saturating voltage across the collector-emitter junction of the transistor.

The smoothing capacitor customarily found in linear regulators is replaced in switching regulators by a choke. This inductor smoothes the output current of the transistor by rhythmically storing and releasing energy (in the shape of a magnetic field). Energy is released via free-wheeling diode D. If the inductance is sufficiently large, the direct output current is virtually free of ripple at constant load. Capacitor C smoothes the output voltage and minimizes the effects of load variations. Because the switching frequency is relatively high, comparatively small values of inductance and capacitance will suffice for satisfactory operation. The output voltage is fed back to the loop error amplifier which compares it with a reference voltage. If the output voltage tends to deviate from the required level, the error amplifier in combination with the sawtooth oscillator varies the duty factor of the output signal of the comparator which pulls the output voltage back to the nominal level.

Theoretically, a switching power supply has an efficiency of close to 100 per cent, which means that, in such ideal conditions, neither the inductor, nor the capacitor, nor the switching transistor dissipate any power. Heat sinks can therefore be kept to a minimum. But, of course, in practice losses occur in all these elements. However, the internal dissipation in a switching supply is constant for a given load, whereas in a linear supply it increases linearly with the input voltage. Because of the large difference between input and output voltage, the switching supply therefore has a clear advantage over the linear one.

The practical side

The foregoing discussion shows that a switching supply is in principle fairly simple. As always, in practice it is not! The reason for this is that relatively large powers are switched at frequencies which are high by power supply standards. The L296 power switching regulator introduced by SGS-ATES removes many of the practical difficulties.

The block diagram of the L296 and some external components to complete the regulator are shown in figure 2. An on-chip zener regulator provides a precise 5.1 V reference voltage. Feedback is direct to pin 10 when the output voltage is 5.1 V, but via voltage divider R_{f1}/R_{f2} at higher voltages (as shown).

The frequency of the sawtooth oscillator is set by choosing the correct time constant $R_{osc}C_{osc}$; this will be described in detail a little later on.

Soft starting, that is slowing down of the rise time of the output voltage when the power is switched on, is provided by an on-chip 100 μA current source and exter-

switching power supply . . .

Figure 1. The basic circuit of a switching power supply.

Figure 2. Simplified block diagram of the monolithic power switching regulator L296.

3

Figure 3. The oscillator
frequency is determined
by the time constant
$R_{osc}C_{osc}$.

4

Figure 4. Onset of current
limiting is determined by
the value of a resistor
connected between pin 4
and earth. If pin 4 is left
open-circuited, the maxi-
mum output current is
5 A. Programmable
current-limiting is poss-
ible by using a 500 k
logarithmic potentiometer
instead of a fixed resistor.

nal capacitor C_{SS}. The rise time is approx-
imated by t = $5.1 \times 10^4 \times C_{SS}$, where t is in
seconds and C_{SS} in farads.
The onset of current limiting can be pro-
grammed by connecting a resistor or
preset between pin 4 and earth (see
figure 4). Should this resistor be omitted,
the maximum output current will be 5 A.
When the output current exceeds the
preset value, the regulator switches off,
and C_{SS} discharges. The output voltage is
then reset under the control of the soft-
start circuit until the current-limiting cir-
cuit again causes the regulator to switch
off. This cycle repeats itself until the cause
of the excessive current is removed.
The reset circuit is intended primarily for
use with microprocessor systems. The
reset input on pin 12 is normally con-
nected to the voltage to be monitored via
a voltage divider. When the level on pin
12 rises above 5 V, the reset output re-
mains logic low for a short time to enable
a system reset to be initiated. On expir-
ation of the delay time, the reset output
becomes logic high. When the reset input
level drops below 5 V, the reset output is
instantly switched back to low (see
figure 5).
The inhibit input on pin 6 enables remote

switching of the regulator with a TTL logic
level (1 = off).
A further on-chip facility is the thermal
shutdown which switches the regulator off
when the temperature of the chip rises
above 150°C. The circuit is automatically
switched on again when the temperature
has dropped to 130°C.
A voltage sensing input (pin 1) and SCR
(silicon-controlled rectifier) drive output
(pin 15) are provided for optional crowbar
overvoltage protection with an external
SCR. The crowbar input is normally con-
nected to the feedback input (pin 10) so
that the SCR is triggered when the output
voltage exceeds the preset value by
twenty per cent. When the SCR is trig-
gered, it short-circuits the output of the
regulator to earth. When the crowbar
facility is not used, pin 1 should be con-
nected to earth.

The circuit

The circuit diagram shown in figure 6 is a
basic concept: you may determine your
own requirements as regards level of out-
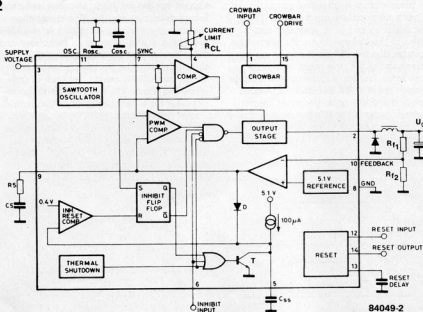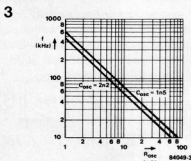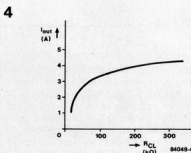put voltage, onset of current limiting,
stability, and so on.
Note that the L296 is connected as a step-
down regulator (output = 8.5 V, followed
by a 78H05 five-volt regulator (IC2). The
78H05 is necessary particularly when the
supply is used with a microprocessor
because the L296 crowbar facility does
not come into action until the output
voltage exceeds the nominal value by 20
per cent. Any microprocessor connected
to the power supply would have given up
the ghost by the time the SCR is
triggered.
The oscillator frequency is 100 kHz: such a
high frequency is advantageous because
both the choke, L1, and capacitors of the
LC filter can be kept relatively small. On
the other hand, the oscillator frequency
should not be too high, because switching
losses then rise rapidly. The dissipation in
the output switching transistor in the L296
is highest at the moment of switching and
the efficiency of the regulator is therefore
dependent on the switching frequency.
Other losses may occur because of free-
wheeling diode D1: this may continue to
conduct for a very short time (that is, the
recovery time) after the output transistor in
the L296 has been turned on. During this
time, pin 2 of the regulator is conse-
quently virtually shorted to earth, which
causes a very large output current. It is

5

Figure 5. A delay in the
L296 reset circuit enables
a system reset to be in-
itiated in microprocessor
systems.
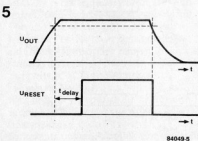
Table 1

Output voltage, U_O, as
function of resistors R7 and
R8: U_O = 5.1 (R7 + R8)/R8
(U_O in volts; R in ohms)

U_O	R8	R7
12 V	4k7	6k2 (4k7 + 1k5)
15 V	4k7	9k1 (100 k // 10 k)
18 V	4k7	12 k
24 V	4k7	18 k
5 . . . 24 V	6k8	25 k (potentiometer)
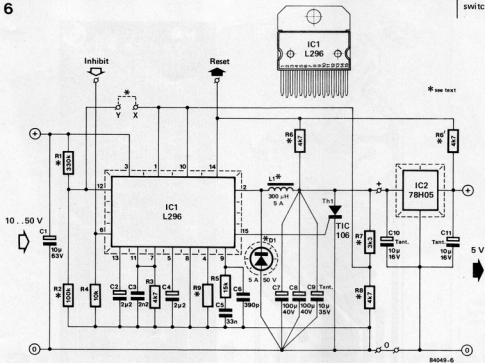
Figure 6. The complete circuit diagram of the switching power supply based on the L296.

therefore essential that the recovery time of the diode is as short as possible. The diode shown, a type UES1402, has a recovery time of only 35 ns and is therefore ideally suited to our purpose, but it may, unfortunately, not always be available.

The choke should be 250 . . . 330 μH at a load current of 5 A minimum. A choice of various makes is given in the parts list (see also figure 7).

The reset input (pin 12) may be connected to either of two voltage dividers: R1/R2 or, via wire bridge x-y, R7/R8. The former has the advantage of reacting faster to short-circuits, but the disadvantage of not re-acting to the L296 switching itself off. If the latter method is chosen, resistors R1 and R2 MUST be omitted.

The pull-up resistor is normally connected

to the output of IC2: if this regulator is not used, the resistor should be connected to the output terminal of the choke.

Construction

First, it should be determined what level of output voltage is required (table 1). A variable output is superior to the others as regards efficiency. If IC2 is used, make certain that the voltage drop across it is 3.5 V minimum. The printed-circuit board shown in figure 8 is designed so that if IC2 is not used, it may be cut as appropriate. Alternatively, a wire bridge as shown near C10 may be fitted.

Cooling requirements in switching power supplies are generally modest, although a distinction must be made between the design as shown and one in which IC2 has been omitted. In the latter case,

Figure 7. A selection of suitable chokes, each of which is suitable for use with the printed-circuit board shown in figure 8.
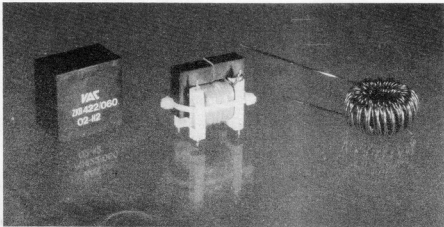
Parts list

Resistors:
R1 = 330 k
R2 = 100 k
R3, R6, R8* = 4k7
R4 = 10 k
R5 = 15 k
R7* = 3k3
R9 = *
* see text

Capacitors:
C1 = 10 μ/63 V
C2, C4 = 2μ2/16 V
C3 = 2n2
C5 = 33 n
C6 = 390 p
C7, C8 = 100 μ/40 V
C9 = 10 μ/35 V tantalum
C10 = 10 μ/16 V tantalum
C11 = 10 μ/16 V tantalum

Semiconductors:
D1 = Schottky diode,
  5 A/50 V, recovery time
  35 ns
Th1 = TIC 106
IC1 = L296

IC2 = 78H05

Miscellaneous:
L1 = choke,
  250 . . . 330 μH, 5 A
  for instance,
  Renco toroidal
  type 1386-3-330;
  VAC type
  ZKB 422/060-02-H2

or Thomson core
GUP 20 x 16 x 7 with
50 turns enamelled copper
wire SWG21 with air gap
of 0.7 mm
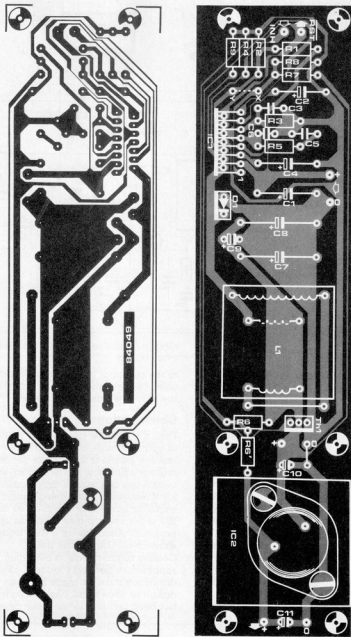
Figure 8. The printed-circuit board of the switching power supply. If no regulator is used in the IC2 position, the corresponding part of the board may be cut off.

strip of aluminium (SWG 14) about 130 x 40 mm will suffice. The L296 and diode D1 are both fitted to this onto mica washers smeared with silicone grease substitute.

When IC2 is used, a rather more elaborate heat sink is needed, because this device dissipates no less than 17.5 W at maximum output current. The photograph at the heading of this article shows how the cooling is effected in this case. The 78H05 is fitted to an angle-piece of aluminium which in turn is mounted onto an angle-section running along the length of the pc board. The angle-section may be chosen to allow the pc board to be attached to it by means of spacers.

A note on heat. We often hear from worried readers that 'it', meaning an IC or heat sink, gets 'hot'. But, of course, heat is a very subjective experience: for instance, 55°C is hot, but can be touched, whereas 80°C is too hot to handle. Bear in mind, therefore, that you can burn your fingers 2 . . . 3 times over before the situation with the IC and heat sink gives cause for alarm. Also, heat sinks function more efficiently at high temperatures because they then lose heat not only by conduction but also by radiation. Heat sinks (and other parts in electronic circuits) are frequently given a far higher rating than required because 'feel' rather than technical considerations was used in their selection.
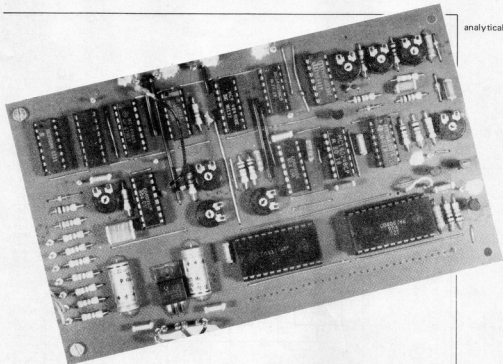
analytical video display

The third, and last, part of the real-time analyser is dealt with elsewhere in this issue. That project is now complete but it can be enhanced with the addition of this display which shows the output of the analyser in colour on a normal TV set or monitor. The bar for each output channel is sub-divided into 32 sections each of which represents a step of 1 dB. The colour of the bar changes per dB so that the value of each bar is easily read. This circuit is also extremely suitable for any application where it is useful to see a read-out in bars on a screen.

The real-time analyser, which is now complete, provides its read-out on a LED display built up from 330 discrete LEDs. This is more than sufficient for 'personal' use. It can sometimes be useful to have a larger display for some applications, such as for demonstration purposes or to be able to see the result at a greater distance from the analyser. This circuit forms the output of the 30 rectifiers into a video signal so that the display can be shown on any monitor or TV set (via a modulator). The LED display of the real-time analyser can simply be left connected, but it is also possible to leave out the LED display completely and build an analyser that has only this video display. As stated in the associated article, the complete display board and the 8 V supply on the base board are then no longer needed.

The analytical video display circuit can also be used for other applications, principally where d.c. voltages are to be displayed on a screen. In the basic version the circuit works with 30 (possibly 31) inputs but this number can be reduced even to one.

The block diagram

The drawing of figure 1 shows the block diagram for the circuit. A 32-channel analogue multiplexer combines all the input signals (the maximum number possible is 31) into one complex signal that must then be processed synchronously with the video signals. The signals needed for the synchronizing are provided by the video sync box published in the March 1984 issue of Elektor.

The output signal from the multiplexer is sent to a fast comparator (IC3) whose switch-over point is defined by the reference voltage applied to the inverting input. This voltage follows an exponential sawtooth characteristic, which is an easy way to achieve quite an accurate ($\pm$ 1%) logarithmic division.

The sawtooth is synchronized to the raster frequency of the video signal (50 Hz) by the field display gate (FG) generated from the synchronization signal (CS) and the blanking signal (CBLK). The total period of the ramp is about 256 lines as this is the number of lines available vertically for

turns a TV set into a colourful bar-graph display with a full scale sensitivity of 1 V d.c.
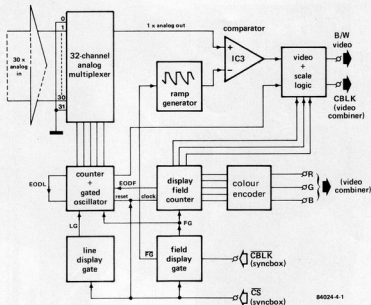
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Figure 1. The functioning of the circuit can be readily seen from the block diagram here.

displaying the signal levels. The display field counter (DFC) ensures that information is only shown on the screen during these 256 lines. As soon as this number is reached the end of display field (EODF) signal stops the analogue multiplexer along with the circuits around it. The display field counter also provides the base signals for the scale logic and the colour encoder. Due to the fact that the DFC is clocked by the CS signal (line frequency) each line can be provided with colour information. Careful use of colour makes the display more attractive and the steps easier to read.

The input signals are multiplexed in the same way as was used in the LED display, namely by using a pair of 16-channel multiplexers connected 'in series'. Now, however, the switching frequency is much higher (about 666 kHz). Switching occurs synchronously with the line frequency so all thirty channels must be examined in one line time. For one complete raster all 30 analog inputs must be examined 256 times. To ensure that this happens correctly the start moment counter and the gated oscillator must be controlled by the CS signal. The LG (line gate) and FG signals are taken from this CS signal. The oscillator is stopped by means of the EODF and EODL (end of display line) signals.

The circuit

The complete circuit for the video display is shown in figure 2. The layout of the multiplexer section has already been dealt with in the article on the LED display so we do not need to spend any time looking at IC1 and IC2. The next section shows a marked difference from the LED

display as a single comparator (IC3) is made to do all the work in this case. This is possible because a varying reference voltage is used here along with a different sort of display (a TV screen).

The varying reference voltage is provided by the external synchronized exponential sawtooth generator. This generator, in fact, is no more than an RC circuit consisting of C10, R7, R8, P2 and P3. The discharging pulse for the sawtooth simultaneously synchronizes to the image window available. The pulse is taken from the composite sync and blanking signal.

If there is a raster blanking signal present this is noted by MMV1 (because no trigger pulse then arrives within 80 μs) and MMV2 is then triggered by the CS signal. This monostable therefore supplies one pulse per raster and this pulse not only synchronizes the sawtooth signal, FG, but also defines the start of the field display gate (FG) as IC11 (the display field counter) can only start counting at the end of this pulse. The vertical image position can be set by changing the width of the FG pulse. The image height is fixed at 256 lines because after this number of CS pulses the gated oscillator (N13) is disabled via output Q8 of IC11 (EODF) and N1. At the same time N1 receives the FG signal so that the gated oscillator is only enabled during the (vertical) visible (not black in other words) part of the image. The build-up and colour of the scale division is entrusted to IC11, as we will see later when we come to describing the layout of the display.

The horizontal image division is handled in much the same way as the vertical. It is done by means of the CS signal, MMV3, MMV4, the gated oscillator (N13) and the address counter of the multiplexer (IC4).

FF1, FF2 = IC5 = 4013
MMV3, MMV2 = IC6 = 4098
MMV3, MMV4 = IC7 = 4098
N1...N3 = IC8 = 4025
N4...N9 = IC9 = 4049
N10...N13 = IC10 = 4093

EODL = end of display line
EODF = end of display field
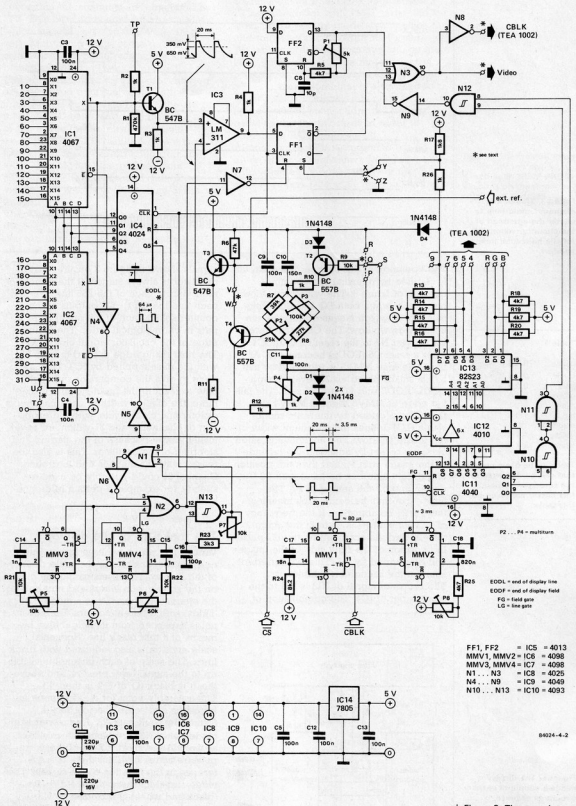FG = field gate
LG = line gate

F2...F4 = multiturn

Figure 2. The complete
circuit diagram, as shown
here, relates very closely
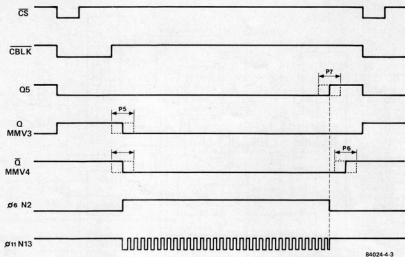to the block diagram of
figure 1.

3

Figure 3. This timing diagram should help to clarify the operation of the circuitry responsible for the horizontal image window.

Table 1.

colour	R	G	B
black	0	0	0
blue	0	0	1
green	0	1	0
cyan	0	1	1
red	1	0	0
magenta	1	0	1
yellow	1	1	0
white	1	1	1

Table 1. The basic colours that can be displayed on the screen with the aid of this circuit are given in this table.

The operation of this section of the circuit is clarified by referring to the timing diagram of figure 3. The waveforms shown only apply when both FG and EODF signals are '0', which means during the vertical image window. The $\overline{CS}$ signal travels via inverter N5 to the reset input of IC4. After a short time Q5 (EODL) becomes '0'. At the same time MMV3 is triggered so that N13 is still disabled via N2 even though EODL, EODF, and FG are all zero. The falling edge of MMV3 triggers MMV4 and the pulse provided by this latter monostable (LG) defines the maximum width of the image window. The horizontal position can now be set by adjusting the left side of the image with P5 and then the position of the right side can be fixed with P6 depending on the width of P7. The image can thus be centred by means of this P5/P6 combination. If the frequency of N13 set with P7 is so high that more than 32 periods occur within the LG pulse the position of the right side of the image will be defined by Q5. This is referred to again in the section on calibration.
The oscillator signal of N13 is fed to the clock input of IC4. The combination of

signals shown in figure 3 ensures that the oscillator is always started at the same moment during each of the 256 lines. During each line the Q0 . . . Q4 outputs of the counter select each of the multiplexer inputs in chronological order. The falling edge of the oscillator signal clocks IC4. The next rising edge is used to load FF1 with the data supplied by IC3. This is done to give the comparator sufficient time to react to the input signal. Furthermore, the display width of each channel is exactly one clock cycle, with the exception of channels 1 and 31 which are never totally visible because of the starting and stopping of the oscillator. This is also the reason why channel 0 is tied to ground. Channel 31 could possibly be used to specify, for example, a certain reference level.

The make-up of the display

Various ICs which we have ignored up to now are responsible for the scale division of the display. The levels cannot be given on the screen with letters and numbers so the alternative chosen was a number of 'fields' with different colours. Each channel is separated from its neighbours by means of a thin black line. Horizontal scale division is also indicated with black lines. The scale of each channel is visible up to the signal level present, and above this it is black ($\overline{Q}$ of FF1 is then '1'). The assembly point for measurement information and scale division data is formed by NOR output N3. An inverter is included after N3 for colour transmissions, this we will see shortly. Both of these outputs are active high but there is a difference in the function of the outputs. The *video* signal for monochrome displays (black and white) is provided by N3 whereas output N8 gives the *blanking* signal for the colour image.
For the time being we will confine ourselves to the blanking signal. It has
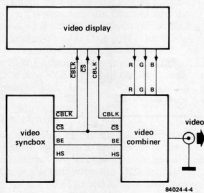
4

Figure 4. The display needs a couple of extra circuits in order to operate and this diagram shows how the video sync box and the video combiner are linked to the display.

already been mentioned that a '1' at the $\bar{Q}$ output of FF1 gives a black image. This can also be seen as blanking. The lines separating the channels are generated by FF2. This flip-flop is connected as a MMV which supplies a pulse of 200 . . . 300 ns with each rising edge of the gated oscillator. Each pulse results in a short blanking period.

The horizontal lines for the scale division are generated from the count on the display field counter (IC11). Working from the top of the image (or the measuring field) every eighth line is blanked. There is a total of 32 black lines resulting from the Q0 + Q1 + Q2 function realized with N9 . . . N12. The thirtysecond line is not visible as it is right at the bottom of the image.

The display now has a scale division but the readability is not all that good yet. A marked improvement can be made by adding colour.

One bit is taken for each of the primary colours, red, green, and blue, so that with three bits a maximum of seven colours can be made, as table 1 shows (black is not usable). The steps in the scale division, each consisting of eight lines, can also contain colour information. The colours chosen are given in table 2. The video display/real-time analyser combination uses a scale in which each of the steps represents 1 dB. Colour is used as follows: between −1 and 0 dB the display is white, above that the steps up to +6 dB are magenta and red, below the white and down to −6 dB the colours are green and yellow. The range from −6 dB to −26 dB (the lowest screen section) is cyan and blue.

The colour information is coded by means of a PROM (IC13) which is addressed by the display field counter. The PROM operates at TTL level (+5 V) so a level adapter (IC12) is needed between this and IC11. A number of resistors, R13 . . . R16 and R18 . . . R20, are included because IC13 has open collector outputs. The output signals are suitable for connecting to the video combiner publisher in the March 1984 issue of Elektor.

Four of the eight PROM outputs are not used but these could possibly be brought into service to make other colour patterns. Output D7 makes a base visible on the screen at the expense of 2 dB of measuring range. This is done by controlling a two-divisions high bar at the bottom of the display via the set input of FF1. This then gives an indication that the display is 'on' even when there are no signals at the input.

Data bits D3 . . . D6 in the PROM are programmed, as table 2 clearly shows. One or more RGB bits can be exchanged with the other data bits if desired. Also, because the outputs here have open collectors it is possible to connect a number of outputs in parallel. If one of the colour bits is exchanged for D6, for example, the display gives a scale with 6 dB steps.

Colour or black and white

The colour display can be shown on a colour monitor with TTL RGB inputs or a TV set or monitor with a PAL input, the latter possibly via a modulator (such as the one published in the April 1984 issue.) We will concentrate, from now on, on the PAL system as this is the more common. Besides the video sync box already mentioned a video combiner is also needed to make a suitable signal. It is clear from figure 4 which connections are needed between the three boards in order to generate a PAL video signal.

A monochrome TV set requires the addition of the circuit shown in figure 5 but without colour the readability of the display suffers. The scale sections will

Table 2.

level	PROM 82S23 address	hex	D6 D5 D4 D3	R G B / D2 D1 D0	hex	colour
+6	00000	00	0 0 1 1	1 0 0	3 C	red
+5	00001	01	0 1 1 0	1 0 0	6 C	magenta
+4	00010	02	0 1 1 0	1 0 0	6 C	magenta
+3	00011	03	0 1 1 0	1 0 1	6 5	magenta
+2	00100	04	0 1 0 1	1 0 1	6 D	red
+1	00101	05	0 1 0 1	0 1 0	5 5	white
0	00110	06	0 0 0 0	1 1 1	0 F	white
−1	00111	07	0 1 1 1	1 1 0	7 E	green
−2	01000	08	0 1 1 1	1 1 0	7 E	green
−3	01001	09	0 1 1 0	0 1 0	6 2	green
−4	01010	0A	0 1 1 0	0 0 1	6 1	yellow
−5	01011	0B	0 1 1 0	0 0 1	6 1	green
−6	01100	0C	0 1 1 0	0 1 0	6 2	green
−7	01101	0D	0 1 0 0	0 0 1	4 1	cyan
−8	01110	0E	0 1 1 1	0 1 1	7 B	blue
−9	01111	0F	0 1 1 1	0 1 1	7 B	blue
−10	10000	10	0 1 1 1	0 1 1	7 B	blue
−11	10001	11	0 1 0 0	0 0 1	4 1	blue
−12	10010	12	0 1 1 0	0 1 1	6 B	blue
−13	10011	13	0 1 1 0	0 0 1	6 1	blue
−14	10100	14	0 1 1 0	0 0 1	6 1	blue
−15	10101	15	0 1 1 0	0 1 1	6 B	blue
−16	10110	16	0 0 0 1	0 1 1	1 B	cyan
−17	10111	17	0 0 0 1	0 1 1	1 B	cyan
−18	11000	18	0 0 0 1	0 1 1	1 B	blue
−24	11110	1E	0 0 0 1	0 1 1	1 B	cyan
−25	11111	1F	0 0 0 1	1 0 0	1 C3	cyan

5

84024-4-5

Figure 6. Constructing this project is greatly facilitated by using the printed circuit board whose layout is shown here.

Parts list

Resistors:
R1 = 470 k
R2 . . . R4,R10 . . . R12,R26 = 1 k
R5,R13 . . . R16,
R18 . . . R20,R25 = 4k7
R6 = 47 k
R7 = 39 k
R8 = 27 k
R9,R21,R22 = 10 k
R17 = 1k8
R23 = 3k3
R24 = 8k2
P1 = 5 k preset
P2 = 25 k ten turn preset
P3 = 100 k ten turn preset
P4 = 1 k ten turn preset
P5,P7,P8 = 10 k preset
P6 = 50 k preset

then simply have to be counted or the colour signal could be displayed in black and white which would at least give a number of different shades of grey. The monochrome version also has one small plus point. If the base for the display is not considered important then IC12, IC13, R13 . . . R20, R26, and D4, as well as the video combiner in its entirety, can be left out. Now, instead of connecting X to Y, X must be connected to Z. The output of N3 (video) is connected to the video input of the circuit in figure 5, as is the CS signal. The output of this circuit then provides a good monochrome signal that can be connected directly to a TV set or monitor with a sensitivity of 1 V_{pp} at 75 Ω.
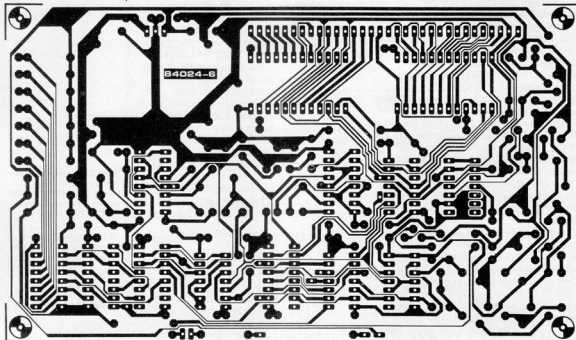
Construction

The whole circuit shown in figure 2 can be constructed on the printed circuit board shown in figure 6. Mounting the components is just a matter of following the parts list and component overlay. Two things to watch out for are that the PROM is programmed correctly (it is also available pre-programmed) and that no wire links are forgotten. Link T-U is inserted if output 31 is not used. To select colour or monochrome display either link XY is made (colour) or X is connected to Z (black and white). Points P, Q, R, and S remain open for the moment.

The connection points for the video sync box and video combiner are on one long side of the printed circuit board and the inputs are at the other side. Inputs 1 . . . 30 are connected to the filter outputs of the real-time analyser by means of a suitable length of ribbon cable. One short side of

the board has the PROM outputs and normally these are only connected to the RGB inputs of the video combiner. A symmetrical + and −12 V supply is needed for the circuit, but if used with the real-time analyser the power can be tapped off the appropriate point on the input or base board. Otherwise a separate supply with two voltage regulators (7812 and 7912) must be built. The current consumption will not exceed about 300 mA.

Calibration

All presets are first centred. The screen should now show at least a part of the base bar (occupying two scale divisions across the screen) and probably also a few undefined vertical bars (for the channels).

The screen format

Rotate preset P2 completely anti-clockwise and then turn P4 anti-clockwise until the 'bars' grow into a rectangular block that stretches the whole height of the screen. The image can now be correctly positioned vertically by means of P8. This preset must be set so that one complete line is visible at the top of the display. The monostable time of MMV2 is then such that the end of the FG signal is in the black part of the screen. The width of the image is decided with P6 by adjusting this preset anti-clockwise until the display fills the whole width of the screen. If this proves impossible the frequency of the oscillator based on N13 must be reduced by rotating P7 clockwise. Now the adjustment of the screen format is finished by setting the horizontal centering with
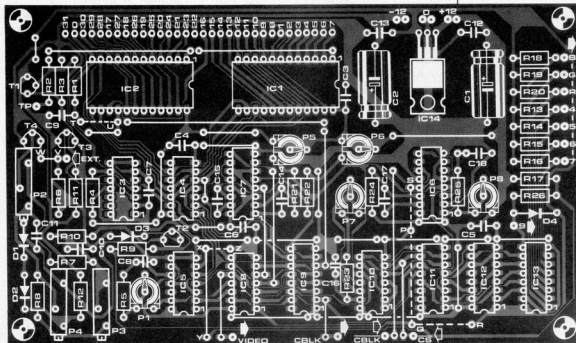
preset P5.

The screen is now set to display 30 (or 31) channels. A smaller number of channels may be selected by reducing the frequency of N13 with P7. In this way the number can be lowered to 25. A further reduction, to 15, for example, can be achieved by increasing the value of C16 to 180 pF. If only 15 or less channels are used IC2 is superfluous and can be removed.

The reference sawtooth

A few 'extra' links have been included on the board in order to enable the sawtooth waveform to be adjusted. These are the already mentioned P, Q, R, and S, with Q being the central point.

Connect Q to R (+5 V) and all the inputs of IC1 and IC2 to ground. As a result of this, capacitor C10 is completely charged and the zero level of the sawtooth can be set with P4. A millivolt meter must be connected from the emitter of T1 (—) to the emitter of T3 (+) and P4 is then adjusted until the meter shows a reading of zero millivolts. This will also be clearly visible on the screen. A negative voltage difference will fill the screen whereas with a positive voltage only the base bar will be visible. The correct setting is when the image is at the (unstable) switching point between full and empty screen.

Connect Q to P instead of to R. The voltage across C10 now drops to about 0.75 V. The millivolt meter remains connected to T1 and T3 but is switched to a range where it can measure up to 2 V. The reading on the meter is adjusted to +1 V by means of P2. A better alternative,

actually, is to apply exactly 1 V to all the inputs and trim P2 so that the meter reads zero volts, but this requires two accurate meters. The upper limit of the measurement range is now set (+6 dB = 1 V). On to the lower limit of the measuring range next. The (non-visible) absolute lowest limit is at —26 dB. The 0 dB level represents a voltage of 0.5 V d.c. so —26 dB corresponds to an input of 25 mV. For practical reasons the bottom of the display is not a very suitable adjustment point. A more usable point, which is also easier to locate on a colour display, is —6 dB (= 250 mV). This is the exact point on the screen where the blue/cyan section meets the green/yellow part (ie. the black line between cyan and green). Taking this as an adjustment point has the added advantage that the error introduced by the fact that the reference voltage is not 100% accurately logarithmic is kept to a minimum. A very accurate millivolt meter is needed for this adjustment.

Apply 250 mV d.c. to all the inputs and connect Q to S (having first removed the Q-P link). Adjust preset P3 (sawtooth frequency) until all the bars come exactly to the cyan/green border. Then, if necessary, the sawtooth adjustment can be repeated. The width of the dividing lines between the channels can now be set with preset P1.

The adjustment of the real-time analyser is dealt with elsewhere in this issue and these two circuits will normally fit together without further ado. Any differences in level can be compensated by adjusting the value of resistor R12 on the input board of the analyser.

K

Capacitors:
C1,C2 = 220 μ/16 V
C3 . . . C7,C9,C11 . . . C13
= 100 n
C8 = 10 p
C10 = 100 p
C14,C15 = 1 n
C16 = 100 p
C17 = 18 n
C18 = 820 n

Semiconductors:
D1 . . . D4 = 1N4148
T1,T3 = BC 547B
T2,T4 = BC 557B
IC1,IC2 = 4067
IC3 = LM 311 (14 pin DIL)
IC4 = 4024
IC5 = 4013
IC6,IC7 = 4098/4528
IC8 = 4025
IC9 = 4049
IC10 = 4093
IC11 = 4040
IC12 = 4010
IC13 = 82S23*
IC14 = 7805

*IC13 programmed according to table 2.

The greatest joy for a zoo-keeper is when one of the animals in the zoo gives birth. It is all the more so if the animal in question is particularly rare or exotic. A great deal of time and effort goes into making all the animals feel as natural and 'at home' as is possible in captivity so it is hardly surprising that the zoo-keeper is overjoyed when the efforts are rewarded by the birth of new offspring. On a much smaller scale there are many individuals interested in breeding animals. Most of the animals (especially rabbits) need little prompting but in some cases, such as birds and fish, a certain amount of zoological know-how is essential. In these cases many factors are important, particularly heat and (what interests us) light.

aviary illumination

an economical alternative for when Mother Nature needs a helping hand

We all know what the birds and the bees get up to in spring but did you ever stop to think why it is at this particular time of the year that a bird's thoughts turn to building nests and rearing families? It is triggered by many prompts, especially the gradual increase in temperature and the lengthening of the days. Bird breeders try to simulate the bird's natural breeding conditions as closely as possible but there are times when they feel the need to give nature a hand.

The circuit here came into being as a result of a bird breeder's request for an auxiliary lighting system that could provide 'natural' illumination for his aviary. It was to have a gradual progression between light and dark (sunrise and sunset), which should be adjustable, and it was also to take account of the ambient light level. The 'sunrise' and 'sunset' durations are of particular importance for several reasons. Apart from making the birds feel 'at home' they also fulfil a very simple function: they warn the birds that it is about to become dark (or light) and that it is time to return to the nest. If a bird stays off the nest for even one night the eggs will not hatch!

Sunrise and sunset

Most of the circuit of figure 1 is concerned with controlling the light at the transition periods, when switching from light to dark or *vice versa*. The timing diagram of figure 2 shows the signals seen at some important points in the circuit. The 50 Hz frequency of the mains supply is sensed by zero-crossing detector IC12, which outputs a 50 Hz square wave signal. This is fed via N17, N16, and N6, which convert it to 100 Hz, to the TR input of MMV1.
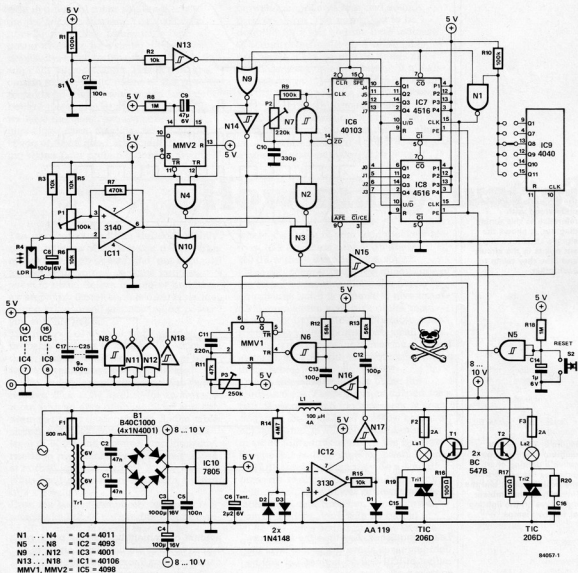
The output of IC12 is also fed to the clock input of binary counter IC9 which divides the mains frequency by a certain factor depending on which output is connected

to the rest of the circuit. Output Q1 is intended as a quick test output, and the division factor can be selected between 256 (2^8) and 4096 (2^{12}). The sunrise and sunset times are therefore adjustable roughly from ten minutes to three hours. The output of the 4040 feeds the clock inputs of binary counter IC8/IC7 via N1. This counter is clocked every 256 . . . 4096 mains cycles and counts up from 0000 0000 or down from 1111 1111. Whether it counts up or down is determined by the position of time-switch S1. If the switch is open the +5 V is inverted by N13 and fed to the U/$\overline{D}$ inputs of IC7 and IC8. The counter then counts down, with the result that the light gradually increases. If S1 is closed the light slowly decreases as the counter counts up.

This count appears at the Q1 . . . Q4 outputs of the two 4516s which are linked to the J0 . . . J7 inputs of IC6. This 40103 is an eight-bit binary down counter. Its single output, ZD (Zero Detect) goes low when the count reaches zero.

While all this has been happening MMV1 has also been busy. The pulse fed via N6 to its TR input triggers the monoflop, causing its Q output to go high for a certain length of time, T. This time is, in fact, very important because it determines when the $\overline{\text{APE}}$ input of IC6 is activated by being taken low. By means of P3, T should be adjusted so that pin 9 of the 40103 goes low just before the zero-crossing point of the mains supply (see figure 2).

The oscillator around N7 provides the clock signal for counter IC6. Its frequency is adjustable with preset P2 which should be trimmed to the value that enables IC6 to count from the maximum possible output of IC8/IC7 (1111 1111) to zero in 10 ms. We will deal with this adjustment later. When the $\overline{\text{APE}}$ line goes low the value present on the J0 . . . J7 inputs is taken as a preset value from which IC6 begins to count down. At the end of the count ZD goes low, disables oscillator N7 and drives transistor T2 via N2, N3 and N15. This in

N1 . . . N4 = IC4 = 4011
N5 . . . N8 = IC2 = 4093
N9 . . . N12 = IC1 = 4001
N13 . . . N18 = IC6 = 40106
MMV1, MMV2 = IC5 = 4098

84057-1

turn triggers triac Tri2 and lights the in-
candescent lamp La2. The timing diagram
of figure 2 shows the effect of the ZD
signal on La2. The width of the ZD pulse
depends on the count on IC8/IC7 so this
defines how brightly La2 lights.
When the incandescent lamp is fully lit
the CO output of IC7 will be low. This
does two things. It disables the clock
inputs of IC7 and IC8 so the counter is
stopped. Provided S1 is closed,
monostable MMV2 is now triggered and
this keeps La2 lit for about 10 seconds. At
the same time the 'high' at the output of
N9 is fed via inverter N14 to NOR N10. If
the ambient light intensity, which is
sensed by LDR R4, is below the value
preset by means of P1 the output of IC11
will be low. The high output of N10 then

drives Tri1 via T1 to light fluorescent lamp
La1 continuously.

A bird's eye view . . .

. . . of how to use this circuit is about all
that is needed now.
The timing diagram of figure 3 should
help clarify the situation. During the day
the light intensity will normally be greater
than the level set with P1. This preset may
be adjusted so that the output of IC11
goes low at the light intensity at which you
want the fluorescent lamp to light.
When S1 opens the brightness of the in-
candescent lamp, La2 (which was fully off),
will increase gradually. When La2 is at
maximum intensity the fluorescent lamp
lights. A short time (5 . . . 10 seconds) later

Figure 1. The circuit
diagram for this aviary
illumination controller may
seem rather complicated
for something that just
turns lights on and off but
constructing the
circuit, on a piece of
veroboard or similar, and
using it are perfectly
straightforward. If a
metal case is used this
should be earthed from
the earth line in the
mains cable used but
MUST NOT be connected
to the circuit. This situ-
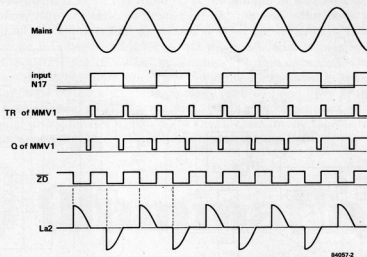ation does not arise with
a plastic case.

2

Figure 2. The operation of the circuit is more easily understood with reference to this timing diagram. It shows the signals at various important points in the circuit and how they relate to each other.
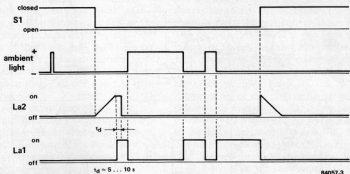
84057-2

3

Figure 3. This timing diagram could also be called a lighting chart. It indicates the effect that the position of S1 and the intensity of the ambient light have on the lighting of the two lamps.

84057-3

La2 extinguishes. Any time the ambient light increases above the preset level the output of IC1l will go high so La1 will be switched off.

When 'sunset' is to begin S1 closes. The fluorescent lamp immediately switches off as $\overline{CO}$ is no longer low and the incandescent lamp lights fully initially and then gradually dims.

As already mentioned, S1 is a time-switch and this should be open during the day and closed at night. Alternatively, it could also be operated manually. Push button S2, on the other hand, acts as a reset and is also used when adjusting the circuit. First P2 is centred and P3 set to maximum resistance and then slowly turned anti-clockwise until the incandescent lamp is completely off (but only just). Keeping the reset button pressed, P2 is then adjusted until the lamp is just on the point of lighting. These adjustments are best done with a link inserted at the Q1 output of IC9.

A few final points. The light dependent resistor, R4, should be mounted in a position where it receives roughly the same amount of ambient light as the birds, but where none of the light from either incandescent or fluorescent lamps falls on it. This suggests some sort of packaging with one translucent side that faces the ambient light and all other sides being opaque.

Throughout this article we have referred to a single incandescent and a single fluorescent lamp. This does not mean that no more than one of each may be used. If the number used is large (more than about 400 W) it may be necessary to use a different type of triac, such as TIC 216D (1200 W) or TIC 226D (1600 W).

Finally, as regards the use of fluorescents as the main light source, there is a good reason for this. It is true that a fluorescent tube does not light instantaneously as an incandescent bulb does. On the other hand, it is far more economical and it gives more scope for the breeder to select what colour light he wants. Incandescent lamps are essential during the 'sunrise' and 'sunset' times, however, as fluorescents are notoriously reluctant to dim properly.

These days if you ask T.C. Mits (The Celebrated Man in the street) how many watts he thinks an amplifier needs to provide in a normal room you are likely to get the wildest answers. They will vary from 20 to 200 watt. Lately, fairly high power seems to be gaining more popularity, much to the joy of many amplifier manufacturers. This is due, to a certain extent, to the efforts of the propaganda specialists who would have us believe that our lives are unfulfilled unless we have the latest 'extra-super-special-ultra-high-power' amplifier taking up the space of three good armchairs in our living rooms (and probably costing more!).

How many watts?

The price of an amplifier depends mainly on the number of watts it can deliver. This means that 'high power' and 'good quality' are not synonymous, as some people seem to think. Before buying or building an amplifier, therefore, it is a sensible idea to first ask yourself just how much power you actually need.

The number of watts actually needed will probably turn out to be a lot less than you think. This is borne out by the table here which gives various sound pressure values and the associated amplifier power needed. This table applies for an average room with an area of about 30 m² and two average-power loudspeakers. As the table shows, soft background music plays at a level of 60 dB and a stereo amplifier only needs to provide 2 x 3 mW for this! Music at normal level needs only 2 x 0.3 watt while loud music is possible with a power of 2 x 3 watt!

Even the largest symphony orchestra in a concert hall does not produce more than 95 dB, a sound pressure that can be reproduced in a living room with a 10 watt amplifier. Why then, you may wonder, do we need any more than 10 watts from an amplifier. Actually it is not simply for impressing the Joneses (or the Mitses). More power is needed, in fact, to reproduce the transients in the music; the very short peaks that can come out at a good 10 dB above the average level. If you want to play symphony music at home at an average level of 90 dB (which would have you sitting somewhere between the strings and the woodwinds — without any ear-plugs) the peaks will be at about 100 dB, which in our average table requires an amplifier producing 2 x 30 to 40 watt.

A few calculations

So it would seem that a 2 x 30 watt amplifier will always be sufficient? Yes and no. The power needed really depends on three things:
■ the maximum sound pressure level that is desired
■ the dimensions of the room

■ the output from the loudspeakers.
Let us start with the first. As the table shows, there is a logarithmic relationship between amplifier power and sound pressure level. The power has to be doubled for a just audible 3 dB increase in sound pressure level. A reasonably noticeable increase of 5 dB needs an amplifier three times as 'heavy'. If a maximum level of 105 dB is required instead of 100 dB, for instance, the amplifier power must be increased from 2 x 30 watt to 2 x 100 watt. A variation of one or two dB is so indistinguishable that there is almost no difference between a good 30 watt and a good 40 watt amplifier. Take note of where the limits of your loudspeakers are; most hi-fi speakers cannot go above 100 dB without distorting the sound.

On to the second point: the dimensions of the room. These also affect the power requirements but it is difficult to give exact dB figures as the acoustics of the room are also of importance. In principle large rooms need more power than small ones. In very large rooms there may be a loss of as much as 5 dB with respect to the values in the table.

Finally, the most important point: the loudspeakers. The table is calculated for a pair of average hi-fi loudspeakers with an output of about 86 to 87 dB. This output is not measured in a living room but rather in an echo-free room where the speaker is supplied with an input of 1 watt. Two 86 to 87 dB loudspeakers provide a sound pressure of about 85 dB in a living room with an amplifier power of 1 watt, as the table bears out. However, if you have a pair of loudspeakers with an output of 90 dB the same sound pressure level requires only half the power stated in the table. So for a sound pressure level of 100 dB a power of 2 x 15 watt is all that is needed instead of 2 x 30 watt. A 2 x 60 watt amplifier, such as the mini crescendo, thus makes levels of at least 106 dB possible.

Loudspeakers with an output of 93 dB only need 2 x 7.5 watt for 100 dB. Loudspeakers with a very high output can produce extremely high pressures of much more than 100 dB with no more than 2 x 5 to 10 watt.

To sum up

The conclusion is clear. The idea of a 'heavy amplifier' is a relative one. Combined with powerful loudspeakers the mini crescendo is a fairly heavy hi-fi amplifier. Depending on the speakers, it is even enough to deafen most people in a large hall. On the other hand if the loudspeakers used have an output of only 81 to 82 dB the 2 x 60 watt do not seem so loud after all so the 'big' Crescendo would be a better choice.

Table 1

amplifier power in watts	sound pressure level in dB	
0.00003	40	background murmur in a quiet room
0.0001		
0.0003	50	damped traffic noise
0.001		
0.003	60	soft background music
0.01		
0.03	70	conversation
0.1		
0.3	80	medium volume music
1		
3	90	loud music
10		
30	100	very loud music
100		
300	110	prolonged exposure to this level can cause permanent ear damage

Table 1. A summary of the sound pressure levels that most people can relate to. The audible threshold is at 0 dB and the pain barrier at about 130 dB.

When we published the design for the 'Cresendo' power amplifier in December 1982 (Elektor U.K.) it proved very popular with the audio enthusiasts among our readers. According to the reactions it was just what the doctor ordered: a real top hi-fi d.i.y. amplifier. The only criticism levelled against it was its price, thus the request for a lower power, lower price, version with the same quality.
No sooner said than done! This slimmed down version of the Crescendo follows the exact same design practice, provides the same quality, produces 2 × 50 (70) watts, but costs less than half the price of its big brother.

mini crescendo

high class
medium power
amplifier

There is no question that the Crescendo is a high-class amplifier but, of course, quality has its price. As d.i.y. amplifiers go it was not particularly cheap although it was much less expensive than a similar quality ready made unit. Furthermore a lot of people have little use for the Crescendo's 2 × 140 watt output (and many examples of the amplifier gave even more than this).
Unless the loudspeakers used are very inefficient (82 dB or lower) 50 or 60 watts per channel will be enough to provide a realistic sound level for the vast majority of domestic installations. A 50 to 60 watt version is also very interesting for anybody who has plans of building his own active loudspeaker system. If you were to provide each of the speakers in a three-way active system with a 140 watt amplifier it would prove not only completely unnecessary but also very costly. A smaller version of the Crescendo should therefore be very suitable for an active loudspeaker system.

The same quality . . . at lower cost

A quick glance at the most costly items in the Crescendo will show that a large saving is possible by accepting a reduction in power.
For an output power of 50 to 70 watts per channel the supply voltage can be reduced from 2 × 70 . . . 75 V to 2 × 45 . . . 50 V. The maximum current also reduces so the power supply becomes significantly cheaper. A single 2 × 35 V/4 A transformer can easily do the job in a stereo system. The electrolytic capacitors in the supply need only be half

1

T1...T7,T9

T8,T10

T11,T12

T1,T2,T6 = BC546B
T3,T4,T5 = BC556B
T7 = BC560C
T8 = BF 470
T9 = BC550C
T10 = BF 469
T11 = 2SK135/134
T12 = 2SJ50/49
D5,D6 = 1N4148

84041 1
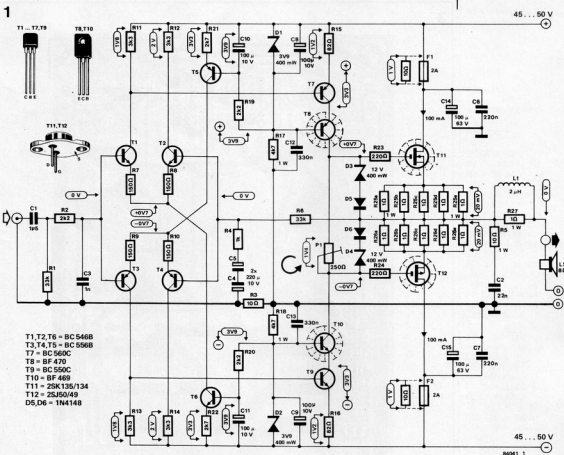
the size and with a lower voltage rating so they are, again, cheaper. The third saving comes from the fact that, for a power of about 50 to 70 watts, it is no longer necessary to connect MOSFETs in parallel. The MOSFET count is thus halved and, as these are relatively expensive components, this saving is very noticeable. Additionally, of course, a saving can be made on the heat sinks and the circuit can be housed in a smaller (and cheaper) case. Taking everything into account we see that, without any compromise as regards quality, the price of the mini crescendo is about half that of its big brother. This then begins to seem much more acceptable.

The circuit diagram

The circuit diagram for the mini crescendo is shown in figure 1, although it can hardly be considered all that 'mini' with an output power of 2 × 70 watt. Clearly it is almost identical to the original version and very little difference will be seen at first sight. The input is fed to a double differential amplifier (T1 . . . T4) with current sources T5 and T6. Then come the cascode driver stages (T7 . . . T10) followed by the MOSFETs (T11, T12) that have to do the hard work. The theory behind this MOSFET set-up is the same as the original and as we dealt with the theory, layout, and background in the

article on the Crescendo we will not spend any space on that here. Instead we will concentrate on the changes that were made.

Returning to the new circuit diagram we now notice that two of the MOSFETs have gone and the supply voltage is reduced to 2 × 45 . . . 50 V. Naturally some component values have been changed here and there to ensure that the circuit works properly at the lowered supply level. In addition to this a few other minor changes have been made, the most important of which is that
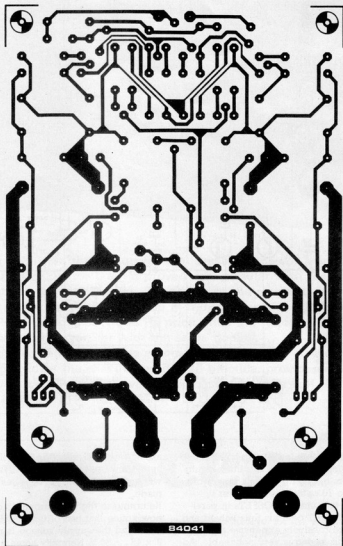
2

84041-2

3

B4041

Figure 3. The dimensions of the printed circuit board, shown here along with the component overlay, have been kept to the popular eurocard format.

wirewound resistors R25 and R26 (Crescendo) are replaced by five ordinary 1 W carbon film resistors. This gives a very low-inductance 5 W resistance and reduces the chance of (undesired) local oscillation to zero.

Dc. freaks will probably note with regret that there is a decoupling capacitor at the input. If this power amplifier is fed from an a.c. coupled pre-amplifier (with a capacitor at the output) then C1 in the mini crescendo can be replaced by a wire link.

One of the most important parts of any power amplifier is the supply, especially as it affects the quality of the final sound. The layout for the stereo version is shown in figure 2. It is a simple combination of transformer, bridge rectifier and electrolytics. The transformer must have a secondary voltage of 2 × 35 V and be able to provide a current of 4 amps. A toroidal transformer is the most compact solution and provides the best appearance but, in principle, any good transformer is suitable. Capacitors C17 and C18 are 10,000 μF electrolytics and these are quite sufficient for our purposes.

The printed circuit board

The number of MOSFETs used is now two instead of four so the size of the printed circuit board can be reduced. We eventually managed to squeeze it down to euro-card format. This is illustrated with the component overlay in figure 3. Mounting the components is quite straightforward and the only important point to note is that it is imperative that coil L1 makes good contact.

We use the same method of mounting the MOSFETs here as was used in the Crescendo. The board is connected to the heat sink by means of a 40 × 40 mm
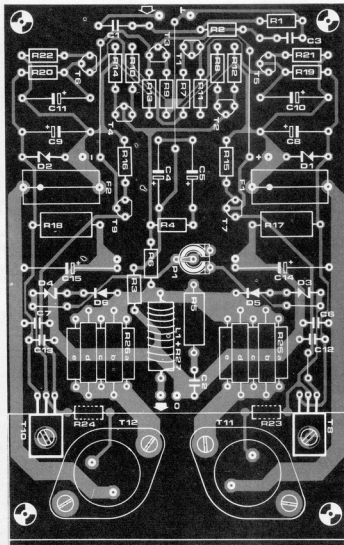
aluminium bracket and the transistors are mounted on this bracket. As a result of the board's smaller size and the lower heat dissipation a single heatsink, about 100 × 200 mm, is sufficient. Drivers T8 and T10 are also mounted on the bracket. Both drivers and MOSFETs must be electrically isolated from the heat sink so mica or ceramic washers must be used when they are mounted and care must also be taken with the mounting bolts. A number of holes have to be drilled in the bracket for the transistors and the printed circuit board could be used as a template for this. As regards the other side of the bracket, this should be fixed to the heat sink with as many bolts as is feasible. It is a good idea to use silicone grease between bracket and heat sink and this is essential when mounting transistors T8, T10, T11, and T12.

Wiring and case

By now you should have two amplifier board fitted to a heat sink, as shown in figure 4. The parts for the power supply given in figure 2 should also be ready so all that remains is to connect the whole lot together to form the complete amplifier. The choice of a case is fairly open. There are some vital wires which must not be too long, of course, but apart from this the only requirement is that the case be sturdy enough. Sensible use of space will probably lead to the choice of a wide, fairly flat, case with the amplifier boards at one side and the supply at the other. This is what we did with our prototype and the result is shown in the photograph at the start of this article; the heat sink, connectors, and mains cord are at the back and the power switch and fuse holder may be at the front.
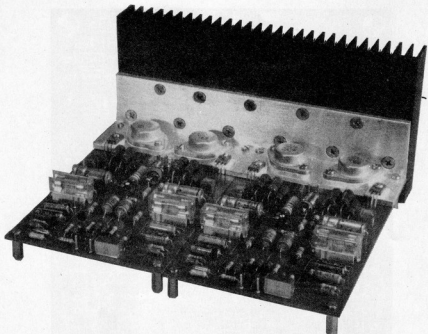More important than the choice of case is

4

Figure 4. The complete amplifier module consists of two printed circuit boards fixed to the heat sink by means of an aluminium bracket.

that the wiring is correctly laid out. By and large, the same guide-lines apply here as in the Crescendo so it is advisable to read the relevant paragraphs in that article. The complete wiring layout of the mini crescendo is given in figure 6 and if this is copied the circuit should certainly work.

A few important points:
Use a single central ground point, preferably the junction of smoothing capacitors C17 and C18. All the ground connections for the amplifier are then taken to this point. These are the earth connection to the (phono) input sockets, the zero volts supply to both boards and the ground connections for the loudspeaker outputs. The central ground point is also connected to the metal case. The input sockets must be mounted in such a way that they are completely insulated from the case and the wiring between these inputs and the boards must be screened cable. All the wiring should be kept as short as possible.

Adjustment and testing

Check once again that the whole circuit is assembled correctly. Pay particular attention to the power supply as the type of electrolytics used here can quite literally blow up if connected without due regard to polarity.

For the sake of testing and adjustment we will consider each half of the stereo amplifier separately. The instructions given here must therefore be followed twice.

■ Remove fuses F1 and F2 and replace them temporarily with 10 Ω/¼ W resistors.

■ Set P1 to minimum resistance by turning it completely anti-clockwise.

■ It is strongly advisable to use a variable a.c. power source when initially switching on power amplifiers. In this way the mains supply can be increased from zero while monitoring to detect any problems that may occur. This is the safe way. The a.c. power supply published in the May 1984 edition of Elektor will be ideal for this purpose since it also includes current limiting. End of commercial break!

■ It will be obvious that if the temporary resistors start to smoulder there must be a fault somewhere in the wiring or in the construction so — hit the panic button — and carefully recheck everything.

■ If nothing untoward happens connect a multimeter across one of the resistors and switch to a low (2 . . . 5 V) d.c. range. If all is correct the meter should read zero volts.

■ Slowly adjust P1 upwards until the voltage drop across the 10 Ω resistor is exactly 1 volt. The quiescent current through the MOSFETs is then 100 mA which is exactly what is required.

■ Switch off the power and replace fuses F1 and F2. When it is switched on again the voltage at the output must not be more than 15 mV with respect to ground.

In principle the amplifier is now finished and ready for use. A final check can be made referring to the test points shown in figure 1.

A tip

The advantages of this sort of symmetrical, complementary, amplifier over an old-fashioned semi-complementary design are
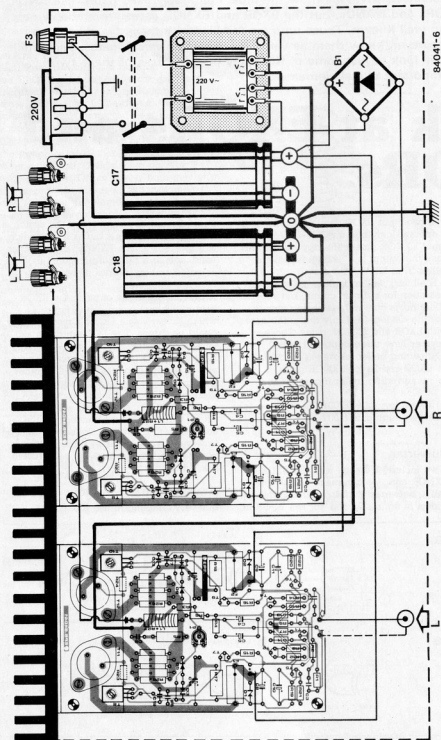
84041-6

numerous but there is one practical disadvantage. The lack of an electrolytic capacitor at the output means that if the amplifier should become faulty there is a possibility of d.c. voltages being fed to the loudspeakers. To prevent that we advise that the amplifier be fitted with a switch-on delay and d.c. protection such as was described in Elektor in January 1983. This circuit protects the loudspeakers against possible damage.

Most electronics hobbyists are now quite familiar with digital technology. We all know the most commonly used gates, like ANDs, ORs and NANDs, but the EXOR and EXNOR gates are probably not so well known. Those who are familiar with these two functions, however, know them as versatile, almost universal, gates. This article will look at just some of the many possible uses of these two important digital elements.

a look at EXORs and EXNORs

the exclusive gates

In a two-input EXOR gate, the output is '1' if one and only one input is '1'. A two-input EXNOR gate is the opposite; in this case the output is '1' if both inputs are either '1' or '0'.

It is all very well to look at the standard definition for a digital circuit, but very often this is little help in seeing how to use the device. Certainly this is the case with EXOR and EXNOR gates. These devices have three important uses:

(a) inverting buffer (shown in figure 1a for EXOR and 2a for EXNOR gates)
(b) non-inverting buffer (figure 1b and 2b)
(c) always low gate (EXOR, as in figure 1c) always high gate (EXNOR, figure 2c).

Bufverter

If we include a switch with the EXOR or EXNOR gate, we can make what we will call a 'bufverter' (buffer/inverter). This is shown in figures 3a and 3b. The EXOR in

figure 3a acts as an inverter if the switch is in position 1, and as a buffer if the switch is in position 2. The EXNOR, on the other hand, acts as a buffer if position 1 is selected and an inverter if switch position 2 is selected.

A practical example of using the EXOR in this mode is given in figure 4, which shows how a Liquid Crystal Display could be driven. The LCD needs an a.c. voltage to operate. This is generated by N8 and N9 and fed to the display common and to one input of each gate. The other gate input then controls the segment. If there is a '0' at the control input, the square wave at the segment will be in phase with the display common so the segment will be visible. These two will, however, be out of phase if the control input is '1', so the segment will be invisible (i.e. 'dimmed').

Always high/always low

It is very tempting to use EXOR or EXNOR

Figure 1. EXOR gate configurations. Inverting buffer (a), non-inverting buffer (b) and always low gate (c).
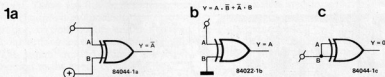
1a $Y = \overline{A}$

b $Y = A \cdot \overline{A} + \overline{A} \cdot A$ $Y = A$

c $Y = 0$

Figure 2. EXNOR gate configurations. Inverting buffer (a), non-inverting buffer (b) and always high gate (c).

2a $Y = \overline{A}$

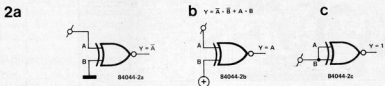**b** $Y = \overline{A} \cdot \overline{B} + A \cdot B$ $Y = A$

c $Y = 1$

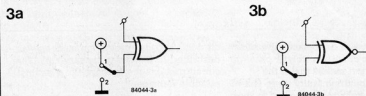Figure 3a. A buffer/inverter based on an EXOR gate.

Figure 3b. Implementing the buffer/inverter with an EXNOR gate.

3a

3b

gates in this mode to block a certain sort of data traffic around a random access memory. This is illustrated in figure 5a for an EXOR gate. When the switch is in position 2 the normal situation pertains and the level on the Read/Write line decides whether the memory is being written to or read from. Switching to position 1 makes the EXOR act as an always low gate, with the result that the memory becomes a write only memory. The data in the memory is then protected from undesired access. A similar application with an EXNOR gate is illustrated in figure 5b. If the switch is in position 2, things simply work as normal. Changing to position 1 makes the EXNOR an always high gate so the memory becomes a read only memory. This time the memory is protected from having anything else written to it.

The next step is, of course, to replace the switches. This is quite easily done by using electronic switches instead of the manual ones shown. The function of the EXOR or EXNOR gate is then more easily controlled, either electronically or by means of the appropriate software.

And there we will leave our short study of exclusive OR and exclusive NOR gates. The one important feature of these devices, which we have left until last, is the pin designations and type numbers of the relevant ICs. They are, in fact, given at the bottom of this page in both TTL and CMOS versions. ◼
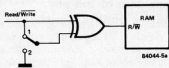
4

N1 . . . N9 = 1/4 CD 4030/
1/4 CD 4070

84044-4
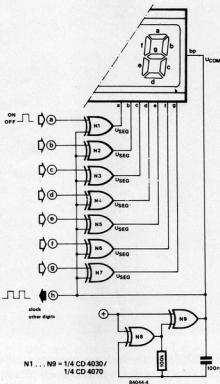
Figure 4. A practical example of how EXOR gates are used in the circuitry to drive a Liquid Crystal Display.

5a

Read/Write

RAM
n/W
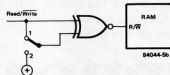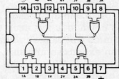
84044-5a

b

Read/Write

RAM
n/W

84044-5b

Figure 5a. Using the EXOR to implement a write only memory.
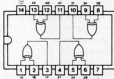
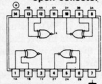Figure 5b. A read only memory can be achieved with the aid of an EXNOR gate.

6

4077: quad 2-input EXNOR

4030 quad 2-input EXOR
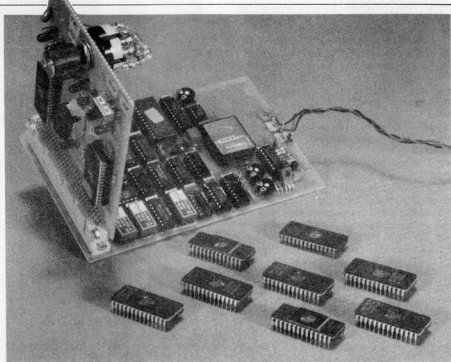4070: quad 2-input EXOR

54/74(LS)266: quad 2-input EXNOR, open collector outputs

54/74(LS)86: quad 2-input EXOR
54/74(LS)136: quad 2-input EXOR, open collector outputs

from an idea
by R. Hasse

EPROMs with various memory capacities and/or from different manufacturers are not standardized as regards pin designations, programming voltage, and programming algorithm. The purpose of the circuit proposed here is to enable all the most commonly used EPROMs (from 16 Kbit . . . 128 Kbit) to be copied by the same universal circuit. The user need only specify the type of EPROM to be copied and the circuit automatically takes care of the pin designations, the programming voltage, and the control signals needed by the EPROM.

EPROM copier

universal
EPROM
duplicator and
verifier

Programming or duplicating EPROMs is not, in itself, such a complicated procedure. A programming voltage is needed (21 V or 25 V) but apart from that it is simply a matter of providing the right addresses, data, and control signals to the EPROM in question. Any method of automation that can speed up the programming process is welcomed, of course. Automatic programming can only be seriously considered, however, if the data required is already stored somewhere, in this case in the (master) EPROM that is to be copied.

One of the most important requirements of this design is that it should be able to copy all the most common types of EPROMs. The memory capacity is, of course, given by the last two or three numbers in the type number but this is not the only difference as the pin designations are also different. In a number of cases this is unavoidable because more address lines are needed to address more

memory. The various manufacturers have also (in the best spirit of 'Murphy Management') avoided standardizing the control signals and programming voltage. What this all boils down to is that each type of EPROM requires a separate programming module or an expensive universal programmer is needed. We were not satisfied with either of these solutions so we set out to find a better alternative.

Different sizes and types

A summary of the most commonly used EPROMs and their pin designations is given in table 1. The 2708 is not included in the series as it differs too much from the norm in needing three supply voltages (-5 V, $+5$ V and $+12$ V). Fortunately this same 2708, which is in any case more expensive than the 2716, is not used very much any more.

As table 1 indicates, EPROMs come in either 24 or 28 pin packages. If we place a 24-pin EPROM on top of a 28-pin device

TYP
TYPE

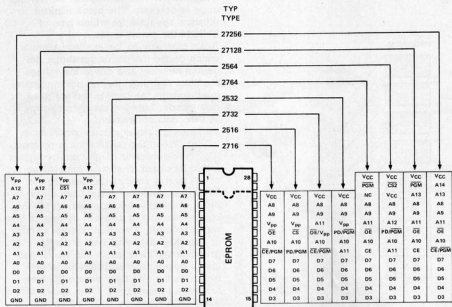|									EPROM							

27256
27128
2564
2764
2532
2732
2516
2716

Table 1. The pin designations vary with the type of EPROM. The 27256 cannot be programmed with this circuit as a result of the different programming method required.

so that pin 1 of the upper chip is above pin 3 of the lower IC we see that most of the pin designations remain the same. There are three pins of the 24-pin package and six of the 28-pin package whose designations change, however, and our copier must adapt the signals on these pins for both master and copy.

The control signals needed are summarized in table 2. The pin numbers in the first column refer to a 28-pin package; the pin number of a 24-pin IC is simply two less than the number stated. The signals are given for both read mode and program mode.

The 27256 is outside the scope of this circuit because it requires a different programming technique. This type of EPROM is programmed by means of repeated writing and checking of the data. If the data is stable the programmer advances to the next byte. This, in fact, comes down to conditional jump commands in the programming algorithm which is something the circuit proposed here cannot deal with.

The circuit

The block diagram for the circuit is shown in figure 1. The heart of the system is the block with the master and copy. The zero insertion force IC sockets into which these mount are not, in fact, connected directly to the address, data, and control signal buses. There are a few auxiliary circuits to ensure that the pin designations are correct for the type of IC plugged in. These circuits are controlled by the control memory, a pre-programmed EPROM. The control memory has a general coordinating purpose. Consisting of a single 2716, it is programmed to ensure that the correct control signals are generated for each type of EPROM. The type of

EPROM, one of those from table 2, is set with type-selection switches. A specific part of the control memory is then addressed, namely the section containing the control signals for that particular type of EPROM. An indication of the approximate contents of the control memory is given in figure 2. A second switch

Figure 1. This is the block diagram of the EPROM copier. The control memory that takes care of the signals for driving the circuit is also an EPROM.
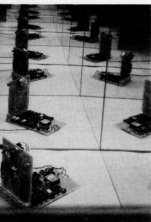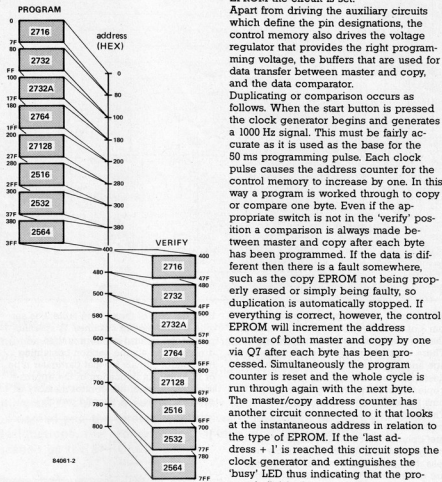
1

2
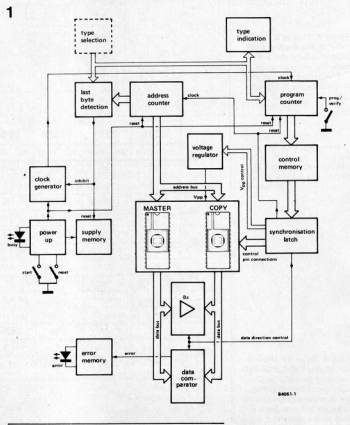
Figure 2. This is a summary of the memory locations where the software for programming and verifying the various EPROM types is stored.

PROGRAM

address (HEX)

Program block	Address
2716	0 – 7F / 80
2732	80 – FF / 100
2732A	100 – 1FF / 180
2764	180 – 2FF / 280
27128	280 – 2FF / 300
2516	300 – 37F / 380
2532	380 –
2564	– 3FF

VERIFY

Verify block	Address
2716	400 – 47F / 480
2732	480 – 4FF / 500
2732A	500 – 57F / 580
2764	580 – 5FF / 600
27128	600 – 67F / 680
2516	680 – 6FF / 700
2532	700 – 77F / 780
2564	780 – 7FF

84061-2

(prog./verify) defines whether we want to copy or compare. The block marked 'indication' specifies for which type of EPROM the circuit is set.

Apart from driving the auxiliary circuits which define the pin designations, the control memory also drives the voltage regulator that provides the right programming voltage, the buffers that are used for data transfer between master and copy, and the data comparator.

Duplicating or comparison occurs as follows. When the start button is pressed the clock generator begins and generates a 1000 Hz signal. This must be fairly accurate as it is used as the base for the 50 ms programming pulse. Each clock pulse causes the address counter for the control memory to increase by one. In this way a program is worked through to copy or compare one byte. Even if the appropriate switch is not in the 'verify' position a comparison is always made between master and copy after each byte has been programmed. If the data is different then there is a fault somewhere, such as the copy EPROM not being properly erased or simply being faulty, so duplication is automatically stopped. If everything is correct, however, the control EPROM will increment the address counter of both master and copy by one via Q_7 after each byte has been processed. Simultaneously the address counter is reset and the whole cycle is run through again with the next byte. The master/copy address counter has another circuit connected to it that looks at the instantaneous address in relation to the type of EPROM. If the 'last address + 1' is reached this circuit stops the clock generator and extinguishes the 'busy' LED thus indicating that the process is finished.

Now we will move on to the circuit

Table 2. Here we show a summary of the pins that can change and the signals needed for them. Pin numbers for a 24-pin package are two less than those stated.

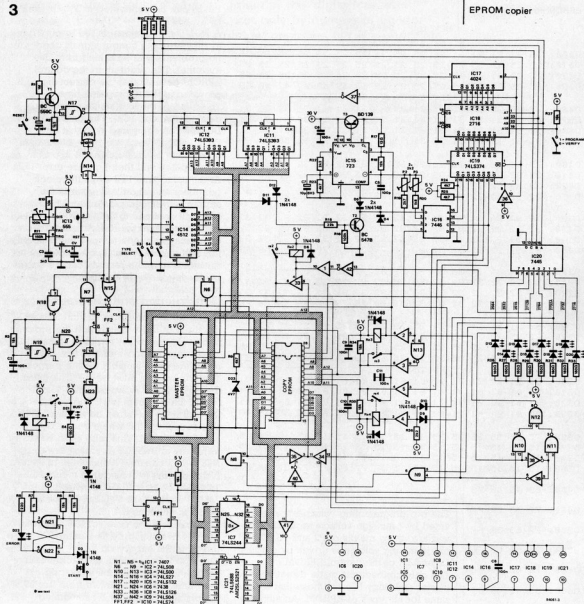| Pin	2716		2732		2732A		2764		27128		2516		2532		2564	
	RD	PGM	RD	PGM	RD	PGM	RD	PGM	RD	PGM	RD	PGM	RD	PGM	RD	PGM
2	NC		NC		NC		A12		A12		NC		NC		CS1	
	*	*	*	*	*	*	*	*	*	*	*	*	*	*	"0"	"0"
20	CE/PGM		CE/PGM		CE/PGM		CE		CE		PD/PGM		A11		A11	
	"0"	⎍	"0"	⎍	"0"	⎍	"0"	⎍	"0"	⎍	#	#	*	*	*	*
22	OE		OE/Vpp		OE/Vpp		OE		OE		CS		PD/PGM		PD/PGM	
	*	*	*	*	*	*	*	*	*	*	*	*	#	#	#	#
23	Vpp		A11		A11		A11		A11		Vpp		Vpp		A12	
	"1"	+25	*	*	*	*	*	*	*	*	"1"	+25	"1"	+25	*	*
26	Vcc		Vcc		Vcc		N.C.		A13		Vcc		Vcc		Vcc*	
	+5	+5	+5	+5	+5	+5					+5	+5	+5	+5	+5	+5
	NC		NC		NC		PGM		PGM		NC		NC		CS2	
	*	*	*	*	*	*	"1"	⎍	"1"	⎍	*	*	*	*	"0"	⎍
Vpp	25 V		25 V		21 V		21 V		21 V		25 V		25 V		25 V	

* = don't care
# = dynamic input

diagram of figure 3. Most of the sections from the block diagram are easily recognized so we will only deal with the sections that have been ignored up to now.

To prevent damage to any EPROMs when fitting them into a socket only IC6 (N21...N24) and IC 20 will be powered up on initial switch on. Power is applied to IC20 in order to select the type of EPROM to be programmed. This is done with S3...S5 and the indication LEDs driven by IC20 then show the type of chip selected. The three selection switches may also be replaced by a single rotary wafer switch as shown in figure 4. After the EPROM type is selected the master and copy are placed in their sockets and the start button, S1, can be pressed. This energizes relay Re1 and the rest of the circuit is then powered up. The address counters and clock generator are started

at the correct time by the power-up circuit. This clock generator is started via FF2 and will later also be stopped by this flip-flop.
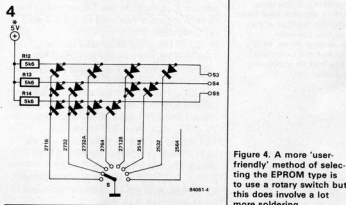
EPROM copier

```
        0  1  2  3  4  5  6  7  8  9  A  B  C  D  E  F

D000:  63 21 21 21 21 21 25 25 25 25 25 25 25 25 25 25
D010:  25 25 25 25 25 25 25 25 25 25 25 25 25 25 25 25
D020:  25 25 25 25 25 25 25 25 25 25 25 25 25 25 25 25
D030:  25 25 25 25 25 25 25 25 63 63 63 63 43 43 40 E3

D080:  77 35 35 35 35 35 31 31 31 31 31 31 31 31 31 31
D090:  31 31 31 31 31 31 31 31 31 31 31 31 31 31 31 31
D0A0:  31 31 31 31 31 31 31 31 31 31 31 31 31 31 31 31
D0B0:  31 31 31 31 31 31 31 31 77 77 77 77 53 50 53 F7

D100:  77 36 36 36 36 36 32 32 32 32 32 32 32 32 32 32
D110:  32 32 32 32 32 32 32 32 32 32 32 32 32 32 32 32
D120:  32 32 32 32 32 32 32 32 32 32 32 32 32 32 32 32
D130:  32 32 32 32 32 32 32 32 77 77 77 77 53 50 53 F7

D180:  7B 3A 3A 3A 3A 3A 32 32 32 32 32 32 32 32 32 32
D190:  32 32 32 32 32 32 32 32 32 32 32 32 32 32 32 32
D1A0:  32 32 32 32 32 32 32 32 32 32 32 32 32 32 32 32
D1B0:  32 32 32 32 32 32 32 32 7B 7B 7B 7B 58 58 58 FB

D200:  7B 3A 3A 3A 3A 3A 25 25 25 25 25 25 25 25 25 25
D210:  32 32 32 32 32 25 25 25 25 25 25 25 25 25 25 25
D220:  32 32 32 32 32 25 25 25 25 25 25 25 25 25 25 25
D230:  32 32 32 32 32 25 25 25 7B 7B 7B 7B 58 58 58 FB

D280:  63 21 21 21 21 21 25 25 25 25 25 25 25 25 25 25
D290:  25 25 25 25 25 25 25 25 25 25 25 25 25 25 25 25
D2A0:  25 25 25 25 25 25 25 25 25 25 25 25 25 25 25 25
D2B0:  25 25 25 25 25 25 25 25 63 63 63 63 43 43 40 E3

D300:  63 21 21 21 21 21 01 01 01 01 01 01 01 01 01 01
D310:  01 01 01 01 01 01 01 01 01 01 01 01 01 01 01 01
D320:  01 01 01 01 01 01 01 01 01 01 01 01 01 01 01 01
D330:  01 01 01 01 01 01 01 01 63 63 63 63 43 43 43 E3

D380:  77 35 35 35 35 35 15 15 15 15 15 15 15 15 15 15
D390:  15 15 15 15 15 15 15 15 15 15 15 15 15 15 15 15
D3A0:  15 15 15 15 15 15 15 15 15 15 15 15 15 15 15 15
D3B0:  15 15 15 15 15 15 15 15 77 77 77 77 53 50 53 F7

D400:  63 43 40 E3
D480:  77 53 50 F7
D500:  77 53 50 F7
D580:  7B 5B 58 FB
D600:  7B 5B 58 FB
D680:  63 43 40 E3
D700:  63 43 40 E3
D780:  77 57 54 F7
```

Table 3. The software stored in the control memory is given in this hexdump. Quite large safety margins have been allowed, particularly as regards the relay switching times, so if desired the execution time can be shortened significantly.

The two programming voltages are provided by precision voltage regulator IC15 and are set by means of P2 and P3. The programming voltage is driven by outputs D0 and D1 of the control EPROM.

The only connections made directly to the master and copy ICs are those that are the same for all types of EPROM. An auxiliary circuit is required to enable pins 2, 20, 22, 23, and 26 to be switched.

Pin 2 is connected to A12 via N6 and if a 2564 is programmed this point is taken low.

Pin 20 is connected to A11, the master via N8, the copy via N35, or this pin receives control signals from the D2 output of the control memory (variants of the CE and PGM signals).

Pin 22 of the copy is treated similarly, except that here it can be provided with either V_{pp} (via Re4) or a control signal via N5. The open collector output of N5 can handle up to 30 V so there should be no problems when Re4 closes. Pin 22 of the master does not have to be connected to V_{pp} so it is connected to ground.

The signal for pin 23 is either V_{pp} (via Re3) or one of address signals A11 or A12 which are transmitted via N3 and N10 . . . N13. When Re3 is closed D23 and R9 ensure that pin 23 of the master has a logic '1' rather than V_{pp}.

Pin 26 can be connected to V_{cc} by Re2 or to A13 via N33.

A PGM or a CS signal is fed to pin 27 and, as they are both control signals, there are no problems with switching from one to the other. The program in the control EPROM ensures that the correct signal is provided at the appropriate time.

The heart of the verification circuit is a data comparator, IC21. The data at inputs D0 . . . D7 is compared with that on D0' . . . D7'. If these two bytes of data are different a '1' will be clocked into FF1. This flip-flop will then light the error LED. At the same time FF2 stops the clock generator and shortly afterwards the supply is switched off via Re1. The error LED will continue to light as N21 . . . N24 are not supplied via Re1. A second attempt may be made to copy the data but this will generally result in a repeated error indication. The remedy can be sought by erasing the copy (again).

Construction and calibration

There are four relays in this circuit and three of these that switch the programming voltage are not required to handle heavy current. Small dual in line reed relays are quite suitable for these purposes. The coil voltage must be 5 V. The gates that drive these relays can sink 40 mA (when the output is low). The resistance of the coil must therefore be at least 125 Ω. The supply for the whole circuit is switched by Re1. This relay must be able to handle at least 0.5 A, and is driven by N23 which can sink a maximum of 24 mA. If this current is considered too small a 7438 can be used for IC6 instead of the 74LS38 as standard TTL can be loaded twice as much as LSTTL.

Take care only to connect IC6 and IC20 directly to the supply. The other ICs are linked to the supply via Re1.

The +30 V from which the programming voltage is taken must be smoothed d.c. Very little current is drawn from this 30 V supply so a small (100 μF) smoothing capacitor is all that is needed.

The circuit must be adjusted at three places. The clock generator must be set to 1000 Hz by means of P1. A frequency meter is essential for this adjustment. If the circuit is set to program a 27128 by means of S3 . . . S5 the circuit operates for about 16 minutes giving us plenty of time to trim the frequency.

Next the programming voltage must be set. To do this we temporarily remove IC19 from its socket and also ensure that there is no master or copy EPROM in their sockets. Make points 9 and 19 of the IC19 socket low. This simply involves connecting them with a piece of wire to pin 10 (ground). The programming voltage of 21 V, measured at pin 3 of IC15, can now be set with P3. The second voltage is now set in the same way. Pins 2 and 19 should now be earthed on the IC19 socket and P2 trimmed to give 25 V at pin 3 of IC15. Then IC19 can be replaced in its socket and the circuit is ready for use.

Because of the extensive interest shown in the 'digital cassette
recorder' (Elektor – february 1984), not only by owners of Elektor
computers, but also by users of 'foreign' ones, we felt it appropriate
to pass on to you the following practical tips.

# digital cassette recorder revisited

The circuit of the digital cassette recorder described in the February 1984 issue of Elektor was designed primarily for storing data originating in the Junior computer. The circuit will work with other computers also, but some of these use such a different tape format that the results are not always satisfactory. If you are one of the sufferers, here are a few hints on how to adapt the circuit better to the computer and cassette recorder used.

In some computers, the hysteresis of the input stages is so long that the data being loaded is not passed correctly to the recorder head. This can be prevented by increasing the value of R6 to 82 k which makes the hysteresis shorter. Furthermore, it is advisable to use high-stability (1%) resistors in the R4/R5 and R12/R13 positions to make the diminished hysteresis as symmetrical as possible. It is also possible to shorten the hysteresis by replacing D3 and D4 by two series-connected zener diodes of 2.7 V or 3.3 V (see figure 1).

The next point concerns the current in the recording head, which is determined by the values of R32 and R33, and is perfectly suitable for most normal cassette recorders. But of course, there may be exceptions to this rule, which manifests itself in incomplete magnetization of the tape (check on a 'normal' recorder how far the meters deflect when you play back the relevant tape). If incomplete magnetization is suspected, the current may be enlarged by reducing the values of the two resistors until the recorded signal no longer increases in strength. It is better to take a slightly too low than a too high value, but it is important that the two values are the same.

Depending on the type of playback/recording head used in the recorder, it may happen that the gain of the playback amplifier is too high. This can be seen from LED D12 which then lights continuously. The gain may be reduced by decreasing the value of R21 to 10 k. Furthermore, a capacitor of 470 p in parallel with R21 will help reduce the higher frequencies which may distort the correct operation of the amplifier. It is recommended to connect the head to the pc board via a coaxial cable as shown in figure 2.

The next tip concerns the two relays, Re1 and Re2. Many of you appear to have used different types than indicated in the parts

1

2

3
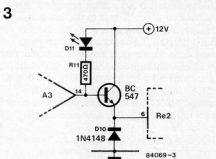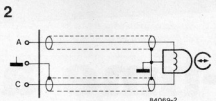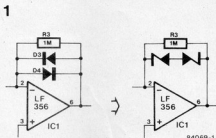
list and consequently found that these did not switch satisfactorily (poor attraction on the one hand, and poor release on the other). This deficiency can be cured by adding an emitter-follower as shown in figure 3.

The problem child of old, the ZX81, continues to misbehave from time to time in spite of several modifications we have tried. The problems appear to arise on the one hand from the level at the cassette output and on the other by the presence of video signals at this output. We continue our search for a definitive solution to these problems . . .

Our real-time analyser still lacks one important printed circuit board, namely that for the pink noise generator. This generator is indispensable for making measurements with the real-time analyser. Using pink noise we can actually show a complete frequency characteristic from 25 Hz to 20 kHz at the same time on the analyser's display. This article will also deal with some practical tips about construction and calibration. And, of course, the finishing touch for the instrument is the front panel, which is now available.
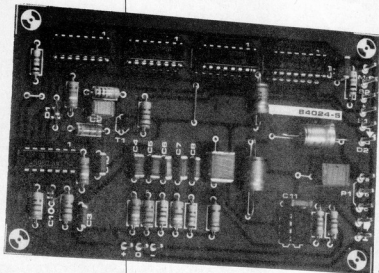
with a contribution from B. König

(part 3)
real-time analyser

pink noise generator, front panel, and final remarks

As of last month the real-time analyser is operative but it is far from complete. Any signal that is presented to the input is analysed and then output to the display (LED or video). For actual measurement purposes, however, a signal is needed that contains all the frequencies in the audio band that is to be analysed. In this way a direct read-out is given of all thirty frequency bands at the same time. The most suitable signal for this measurement is pink noise.

Noise is a signal with a random frequency spectrum. If all the frequencies are equally represented in the noise signal this is called white noise. This might seem like an ideal measuring signal for an analyser but it is not, in fact, possible to use it with our design. In octave and ⅓ octave filters (the latter are used in the real-time analyser) the Q factor of all the filters is the same. This means that the bandwidth of each filter depends on the centre frequency. If white noise is used the output voltage of each filter increases with the centre frequency because the bandwidth is larger at higher frequency bands. The measured frequency would then increase by 3 dB per octave. The white noise generator must therefore be followed by a low-pass filter with a slope of 3 dB/octave in order to get a straight

read-out. This filtered white noise is then known as pink noise.

The noise generator

There are two principal ways of electronically generating noise. One of these uses a noisy transistor junction, the other is a digital noise source. The second method is used here as it produces more consistent results.

The noise generator here is based on a shift register with a period such that the pseudo-random output signal generated has a fairly long repetition time and also appears 'random'. The layout for the circuit is shown in figure 1. Oscillator N1/N2, which has a frequency of about 1.5 MHz, supplies the clock signal for the 31-bit shift register made up of IC2 . . . IC5. Outputs Q28 and Q31 are fed back to the input of the shift register via EXOR gate N3 with the result that one cycle takes $2^{31} - 1$ (= 2 147 483 647) clock cycles. With the clock frequency of 1.5 MHz used here one shift register cycle is about 25 minutes long, so we really can talk of the noise being random.

The disadvantage of such a shift register is that the situation can never arise that the register contains all zeroes, because the circuit would then stay at zero. This is easily solved here by the addition of two

push buttons to start and stop the shift register. The START button (S1) is pressed to read a number of '1's into the register until the contents are somewhere in the region of two milliard. The generator is stopped by pressing S2. The complete shift register is then filled with zeroes and remains like this.

An indication circuit has been included to show when the generator is producing pink noise. Whenever a '1' appears at output Q31 capacitor C2 is charged via D1 and R4. Transistor T1 is then caused to conduct by N4 and the LED then lights. The charging and discharging times of C2 are chosen so that the LED lights continuously when the generator is 'on'.

The white noise signal appearing at output Q31 (pin 11 of IC5) is filtered by a pink noise network consisting of R8 . . . R13 and C3 . . . C8. This is a six stage Tchebychev filter with a theoretical deviation of less than 0.14 dB from the −3 dB/octave line between 12.3 Hz and 31.5 kHz. In practice this means that the deviation from the ideal line is dependent only on the tolerances of the components used in the network.

The filter is followed by a buffer amplifier (IC6) whose gain is set at eleven times. The potentiometer at the output is used to control the output signal. Capacitors C13, C15, and C16 ensure that this signal contains no d.c. components.

Building and installing the pink noise generator

The printed circuit board for the noise generator is shown in figure 2. The layout is fairly simple and mounting the components should not pose any problem. Only three connections to the base board have to be made, namely +, −, and 0. Making these connections is facilitated by using soldering pins in both boards. Soldering the pins together is much more solid than simply using a few pieces of wire. The other connection points on the side are needed for links to control elements and to the output bus for the pink noise section. Note that switch S1 is a changeover push button.

The pink noise generator can also be used as a separate unit. All that is then required is a symmetrical supply of 8 . . . 12 V and, as the current consumption is very modest, this could even be supplied from a battery.

The filters (again)

Last month we dealt with (among other things) the thirty active rectifiers. The charging time constants were then adapted to the centre frequency of each band. This method is quite satisfactory for analysing audio (music) signals but it is not very good when measurements are to be made with the help of the pink noise

Figure 1. In the circuit for the pink noise generator, shown here, the noise is generated by means of a digital pseudo-random noise generator consisting of a clock oscillator and a 31-bit shift register.

1
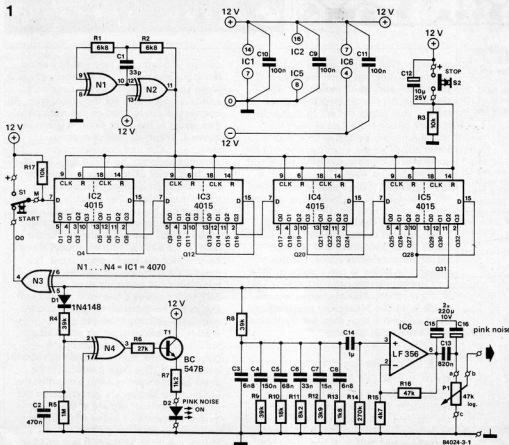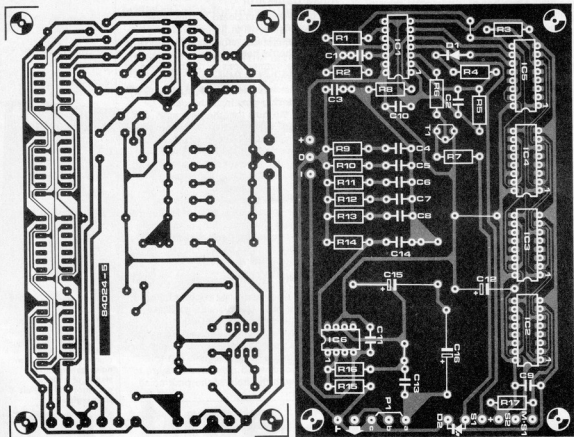
Figure 2. The printed circuit board layout and component overlay are shown here. This is mounted onto the base board in the manner shown in figure 5 of last month's article.

generator. The output on the display will, in fact, be different when measuring with pink noise and when measuring with a sine wave generator (even though the latter will very rarely occur). In higher frequency bands more frequencies fall within a certain period of time than at lower bands so the characteristic will rise somewhat if we compare pink noise with continuous sine waves. It is not such a great problem if the analyser is adjusted with reference to the pink noise generator and no continuous signals are to be measured (the deviation is only about 3 dB). In most cases, however, it would be better if the pink noise and constant signal measurements agree. For this reason it is better to use a value of 150 k for the charging resistors (R1, R3 . . . R59) in the rectifiers. This also gives a more stable read-out. Measuring music signals does take more time then but this could be a blessing in disguise.

The complete layout and the front panel

By this stage you should have a base board with the other boards (except for the display) mounted on it. Now a suitable case must be found or made into which the whole unit will fit. It has to be big enough for the transformer as well, of course, although this may just fit behind the input and noise boards, where there is some room. The back of the case only needs a single hole for the mains wire. All connectors are located on the front panel. The layout of the front panel is given in figure 3. When all potentiometers, switches, LEDs and connectors are mounted they can be wired up. One of the best types of connectors to use are phono sockets as these are commonly used in amplifiers. The input is connected to the input board with screened cable. This is the only place where the ground of the circuit may be connected to the case. A 6.3 mm (¼") stereo socket should be used for the microphone input. For measuring purposes an electret microphone is generally used. Its built in FET buffer needs a few volts supply and this could be transmitted on the spare wire in the microphone cable.
A sheet of red perspex, or something similar, should be fitted behind the display window. For best effect this should be dark red. The base board must be fixed into the case in such a way that the LEDs in the display are just touching the perspex and are at the right height with respect to the scale division on the front panel.
Before using the analyser it is a good idea to check the connections to the transfor-

mer again. The 10 V terminals should be connected to the a.c. voltage points on the base board and the 15 V terminals should be linked to the a.c. points on the input board. The earth terminal of the transformer should be connected to the relevant point on both boards. If the LED display is not to be used and only the video display is desired the symmetrical 8 V supply is not needed. If, on the other hand, both LED and video displays are to be used it is advisable to feed all the lines going to the LED display (the filter outputs and ground) also to the outside via a connector.

Calibration

Last month we dealt with calibration very briefly but this must be changed as the rectifiers now have 150 k charging resistors and everything should be adjusted with reference to the pink noise generator. This generator will in any case be used later as a signal source for almost all measurements made with the analyser. First the frequency bands must be set. Adjust the presets for the rectifiers (P1 . . . P30) to minimum amplification by turning the wiper of each towards the IC. Connect the output of the noise generator to the input of the analyser, with the pink noise output level potentiometer at maximum, the input switch at line, resolution switch at low, and the analyser level potentiometer at maximum. Then the power can be switched on. A number of points on the display will now light and slowly move down. When all these points have disappeared from the display the start button for the pink noise generator is pressed. From the range switch until a series of 'restless' lights appears on the display. Change the range switch and the level potentiometer until the highest level of the whole band is about 0 dB. Make a note of which band is the highest and then set the potentiometers in the rectifiers so that all frequency bands are at about this same level. It may prove impossible to adjust one or two of the bands enough with the potentiometers and if this is the case the resistor in series with the potentiometer will have to be reduced to 180 k in place of the existing 220 k. If all this is done the resolution switch can be set to high and everything can be adjusted a bit more accurately. At the lower frequency bands this adjustment will be a bit difficult as the level varies slowly due to the small number of signals which occur. These bands should be set by taking the average value given on the display. The range switch must now be calibrated. A 1 kHz sine wave with a value of 775 mV a.c. is needed for this. Disconnect the pink noise generator from the input and turn the level potentiometer of the analyser to CAL. Feed the 1 kHz signal into the input, set the range switch to 0 dB, and see what the LED display shows. Only the 0 dB LED in the 1 kHz band should light and R12 must be set so that this is the case. This resistor could be temporarily replaced by a 50 k potentiometer to find the value of the fixed resistance that should be used. This last adjustment is not absolutely essential unless you are interested in measuring absolute voltage values, which only apply for continuous sine wave signals. Otherwise, R12 can be simply left at the stated value. The real-time analyser is now ready for use.

Measurements and measuring microphones

The real-time analyser is quite simple to use. Pink noise is fed to the input of the equipment that is to be analysed and the output of the equipment is connected to the input of the analyser. When the range switch and level potentiometer are properly set the frequency characteristic appears on the display. The resolution switch is then used to set the display range. Some of the most common applications are:

— Measuring the frequency characteristic of an amplifier. This curve is generally fairly uninteresting as it is arrow straight but it is nonetheless handy to be able to see the effect of a tone adjustment, for instance.

— Measuring the frequency characteristic of a tape recorder. In this case the pink noise should be recorded at a low level, such as —20 or —30 VU, to prevent driving the tape into distortion.

real-time analyser (part 3)

Parts list
Resistors:
R1,R2 = 6k8
R3,R17 = 10 k
R4,R8,R9 = 39 k
R5 = 1 M
R6 = 27 k
R7 = 2k7
R10 = 18 k
R11 = 8k2
R12 = 3k9
R13 = 1k8
R14 = 270 k
R15 = 4k7
R16 = 47 k
P1 = 47 k log. pot.

Capacitors:
C1 = 33 p
C2 = 470 n
C3,C8 = 6n8 5%
C4 = 150 n 5%
C5 = 68 n 5%
C6 = 33 n 5%
C7 = 15 n 5%
C9 . . . C11 = 100 n
C12 = 10 μ/25 V
C13 = 820 n
C14 = 1 μ
C15,C16 = 220 μ/10 V

Semiconductors:
D1 = 1N4148
D2 = LED, green
T1 = BC 547B
IC1 = 4070
IC2 . . . IC5 = 4015
IC6 = LF 356

Miscellaneous:
S1 = push button
changeover switch
S2 = push to make push
button
Sennheiser KE 4-211-2
loudspeaker capsule

Figure 3. The front panel for the real-time analyser is larger than one of our pages so it is shown here somewhat reduced. The actual dimensions are 480 × 132.5 mm. Before fitting it remember to remove the plastic protective covering.

3
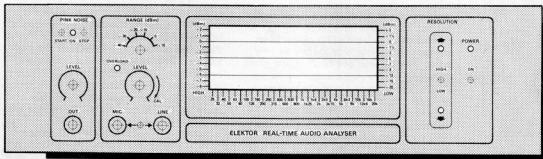
84024
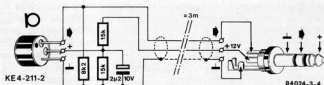
Figure 4. This sketch shows how an electret microphone is connected to the analyser. The supply voltage needed for the built-in FET buffer is taken from the 12 V supply of the analyser via two resistors.

— Studying the acoustics of a room. This last point brings us to a particularly important application of the real-time analyser. In this case a measuring microphone is used. Then, working with a ⅓ octave equaliser and the analyser, the frequency characteristic in the listening area can be made almost completely flat. We cannot forget, of course, the hobbyist with a penchant for building his own loudspeakers. The analyser and a measuring microphone provide an ideal method of examining a loudspeaker and we expect many real-time analysers will find themselves doing just that.

In all applications it is important that the amplifier is not overdriven by the noise. The peaks in the lower bands are about 10 dB greater than at 1 kHz. In these bands

and 20 kHz with a deviation of no more than + or −2.5 dB. A 'case', such as an old microphone body or a suitable piece of plastic tubing, must be found for the capsule but this is well worth the trouble considering the finished result will be about 1/10 of the price of some professional measuring microphones. This microphone is an electret type and also contains a FET buffer. If a stereo socket is used for the microphone input one of the connections can be used to carry the voltage supply for the buffer. The +5 V needed for the FET is taken from the stabilized +12 V on the input or base board. How this is done is clarified by figure 4 which also gives the connection details for the microphone mentioned. Other manufacturers also supply suitable microphones but it is
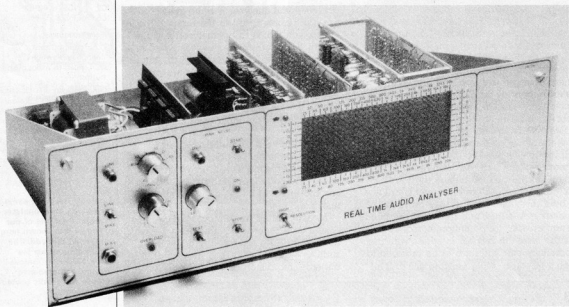
the display will never give a 'calm' read-out but will always be a bit jumpy due to the fact that the bandwidth of the filters are proportionately very narrow so very few signals appear in a band in any given duration.

Finally, a few words about the measuring microphone, which is an essential part of the analyser. This should have a good frequency characteristic and should, ideally, not be too expensive. One suitable possibility would be the Sennheiser KE 4-211-2 capsule which is about the size of a BC 547 and whose frequency characteristic is straight between 40 Hz

essential, of course, to check the data sheet to ensure the frequency characteristic and tolerance are good enough. The level of the sound measured can be adjusted by means of P1 (and if necessary by changing R2) on the input board. This adjustment cannot really be done without a sound pressure meter but the adjustment is only needed if absolute values are to be measured. ◼

Since its introduction at the end of the 1960s the RS 232 norm has become a solid telecommunication standard. More recently, however, other norms have appeared as practice has shown the original one to be lacking in some respects. There is little to recommend going through all these standards with a fine-tooth comb but it is, on the other hand, interesting to trace the lines along which they have evolved as a method of better evaluating their merits. That is what this article will attempt to do for the RS 423 norm which several personal computers have already adopted.

the RS423 interface

A standard, by definition, is fixed. It cannot evolve with the environment it normalises. Some of them resist radical change, such as, for instance, the layout of the keys on a typewriter which was decided by purely practical mechanical reasons but has not been changed for computer keyboards. Others, however, quickly become obsolete.

The authorities on matters of telecommunications are the United Nations' Consultative Committee for International Telegraph and Telephone (CCITT) and the American Engineering Industries Association (EIA). It should be noted that while the American body actually establishes norms the CCITT only makes recommendations, the reason for this being a conflict of interests among the member nations. As far as the norms mentioned above are concerned both bodies are in agreement.

In the field of communication between computers and peripherals (modem, printer, console, etc.) the RS 232C standard (the 'C' simply signifies that it is a revised and corrected norm) is the best known and one of the most used. It is inconceivable that we should examine any newer standard without referring to this archetype.

RS 232C — the reference point

A standard is not simply the correct pin numbering of a connector and an indication of the precise voltages. Certainly it does define electrical and mechanical characteristics but it also contains a detailed description of the signals including their functions and duration. These considerations are often the object of what could be called satellite norms. The CCITT's V24 norm has V28 and V25 as its satellites while the EIA's equivalent, the RS 232C, has its own satellite in the RS 366. The pin layout of the RS 232C interface has been published several times in Elektor, notably on infocard 84. We have never, on the other hand, dealt with the 21 signals officially defined by the standard

because, as a general rule, when a manufacturer provides an RS 232C interface for some equipment it should not be taken as read that all the official signals can be emitted or received by this equipment. Nonetheless manufacturers do ensure that nothing untoward can happen with the signals they do not use.

The RS 232C norm guarantees transmissions up to 20 kilobaud (20 000 bits per second) through a line of 15 metres at most. The voltages used for the logic levels are not the handiest as they do not comply either to TTL or CMOS levels as they are at present.

To understand the limitations inherent in this standard it should be remembered that one logic level is given by a voltage greater than +5 V and the other by a voltage more negative than —5 V (generally $\pm$ 12 V). As far as received signals are concerned the limits are not spaced as far apart ($\pm$ 3 V). The duration of the bit may not lose more than 4% (2 μs if the transmission rate is 20 kbaud) of its total duration during its transmission. It is easy

Figure 1. Among serial telecommunications standards the RS 422A is distinguished by one thing: its speed. Its transmission rate of 10 Mbaud is achieved by the use of an expensive symmetrical system with two wires per signal, giving 46 cables instead of the RS 232C's 25.

1

RS 422

84059-1
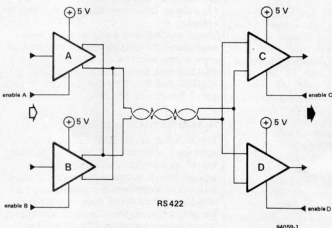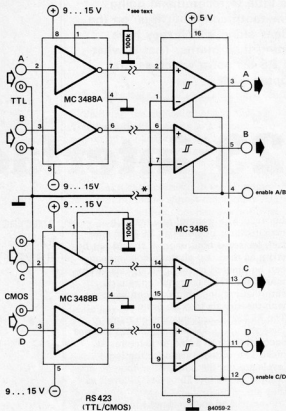
Figure 2. The RS 423A norm is compatible with RS 232C but is less exacting. Its use is facilitated by the availability of special ICs which are compatible with the normal logic families.

to see how the parasitic capacitance of the cable used cannot meet these demands for more than about 15 metres before causing the accuracy of the edges to suffer. The standard is asymmetrical as it has a single common ground line and two-directional 'traffic' so it is inevitable that there will be a difference of potential along the length of the ground line. A current flows in this line with the result that the voltage levels are distorted to some extent.

These can be seen as quite a few limitations and they justify the introduction of new norms (such as the RS 422A and RS 423A with their CCITT equivalents V11/X27 and V10/X26) and attempts to improve the classic RS 232C.

A symmetrical alternative

The RS 422A (or V11/X27) standard, which appeared in the middle of the 1970s, uses two wires per signal and an optional ground line. It is thus a symmetrical (balanced) transmission standard and permits high transmission rates over links that are proportionally very long: 10 Megabaud over a dozen metres or 100 kbaud over 1200 metres. The principle of transmission of a single signal is shown in figure 1. Only a single voltage supply (+5 V) is

needed so there are no problems either with parasitic capacitance or with a current flowing in the ground line. But . . . (there is always a 'but') two wires per signal are needed with the result that this electro-mechanical system is relatively expensive.

In spite of its performance, this symmetrical norm is still far from superceding the good old asymmetrical RS 232C standard. The RS 423A (V10/X26) norm, which conforms to the interface suggested in figure 2, looks to the cheaper procedure for inspiration. This RS 423A is an asymmetrical standard, therefore it is slower, and is an attempt to find the golden mean between the RS 232C and the RS 422A. The maximum transmission rate is about 100 kbaud up to a dozen metres and 1 kbaud up to 1200 metres. The principal characteristic of note of the RS 423A is the use of a *single common ground line which is not connected at the receiver end*. The logic levels are defined by means of a differentiator (see figure 2) whose output is LSTTL-compatible (it could even be put into high-impedance three-state-mode). The common ground line serves as a reference connected to the inverting input of each differentiator but it is isolated from the ground line of the receiver. This completely avoids the problems caused by a current flowing in the ground line. The RS 423A norm tolerates edges that are much less steep than the RS 232C demands. The time taken to determine the logic level may be as much as a third of the length of the total bit (say 300 μs at 1 kbaud) whereas the 232C requires a much steeper slope. The transition zone (about ± 4 . . . 7 V) is compatible with the RS 232C standard but again demands special supply voltages, at least at the transmitter. So we start going round in circles.

As figure 2 shows there are even special ICs available to make the job easier for the budding RS 423 user. They come in eight-pin packages, each of which contains two RS 423 inverting buffers whose transfer characteristics can be modified by changing a single discrete resistor (0.14 μs/kΩ). As we have already indicated, the RS 423A is less exacting than the RS 232C as regards the rise time of the signal. The A suffix designates ICs whose input levels are compatible with the TTL family, whereas the B suffix indicates CMOS compatibility. The buffers' output impedance is 50 Ω and the short-circuit current is 150 mA, although for the almost equivalent MC 1488 it is only 10 mA. ◄

analogue frequency meter

To measure frequency one does not immediately have to 'go digital'. The analogue approach will invariably prove simpler and cheaper, in particular when the analogue readout (the multimeter) is already to hand. All that is needed is a plug-in device, a 'translator', that will give the meter an input it can 'understand'. This design is based upon an integrated frequency-to-voltage converter, the Raytheon 4151. The device is actually described as a voltage-to-frequency converter; but it becomes clear from the application notes that there is more to it than just that. The linearity of the converter IC is about 1%, so that a reasonably good multimeter will enable quite accurate frequency measurements to be made.
Because the 4151 is a little fussy about the waveform and amplitude of its input signal, the input stage of this design is a limiter-amplifier (comparator). This stage will process a signal of any shape, that has an amplitude of at least 50 mV, into a form suitable for feeding to the 4151. The input of this stage is protected (by diodes) against voltages up to 400 V p-p. The drive to the multimeter is provided by a short-circuit-proof unity-gain amplifier.

The circuit

Figure 1 gives the complete circuit of the frequency plug-in. The input is safe for 400 V p-p AC inputs only when the DC blocking capacitor is suitably rated. The diodes prevent excessive drive voltages from reaching the input of the comparator IC1. The inputs of this IC are biased to half the supply voltage by the divider R3/R4. The bias current flowing in R2 will cause the output of IC1 to saturate in the negative direction. An input signal of sufficient amplitude to overcome this offset will cause the output to change state, the actual switchover being speeded up by the positive feedback through C3. On the opposite excursion of the input signal the comparator will switch back again — so that a large rectangular wave will be fed to the 4151 input.
The 4151 will now deliver a DC output voltage corresponding to the frequency of the input signal. The relationship

A true 'universal meter' should be able to read not only voltage, current and resistance — but also other quantities. The usual multimeter will only do this when used as a 'mainframe' in conjunction with 'plug-ins', that convert the input quantity into a form suitable for feeding into the actual meter. This article describes a 'frequency plug-in' that will enable any multimeter (or voltmeter) to read frequencies between 10 Hz and 10 kHz.

between voltage and frequency is given by:

$$\frac{U}{f} = \frac{R_9 \cdot R_{11} \cdot C_5}{0.486 \ (R_{10} + P_1)} \ (V/Hz)$$

The circuit values have been chosen to give 1 V per kHz. This means that a 10 volt f.s.d. will correspond to 10 kHz. Meters with a different full scale deflection, for example 6 volts, can, however, also be used. There are two possibilities: either one uses the existing scale calibrations to read off frequencies to 6 kHz, or one sets P1 to achieve a 6 volt output (i.e. full scale in our example) when the frequency is 10 kHz. The latter choice of course implies that every reading will require a little mental gymnastics! With some meters it may be necessary to modify the values of P1 and/or R10; the value of R10 + P1 must however always be greater than 500 Ω. The output is buffered by another 3130 (IC3). This is an accurate voltage follower, so that low frequencies can be more easily read off (without loss of accuracy) by setting the multimeter to a lower range (e.g. 1 V f.s.d.). The output is protected against short-circuiting by R12. To eliminate the error that would otherwise occur due to the voltage drop in this resistor, the voltage follower feedback is taken from behind R12. To enable the full 10 volt output to be obtained in spite of the drop in R12 (that has to be compensated by the IC) the meter used should have an internal resistance of at least 5 kohm. This implies a nominal sensitivity of 500 ohm/volt on the 10 volt range. There surely cannot be many meters with a sensitivity lower than that. If one has a separate moving coil milliammeter available, it can be fitted with a series resistor that makes its internal resistance up to the value required of a voltmeter giving f.s.d. at 10 volt input. This alternative makes the frequency meter independent of the multimeter, so that it can be used to monitor the output of a generator that for some reason may have a dubious scale- or knob-calibration.

Construction

No trouble is to be anticipated if the

1

2

9869
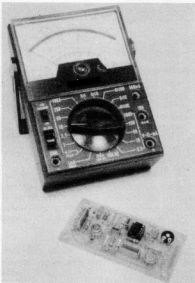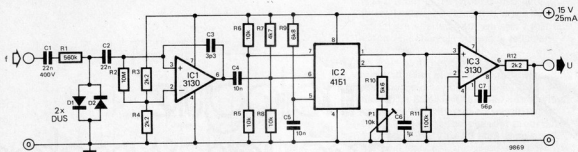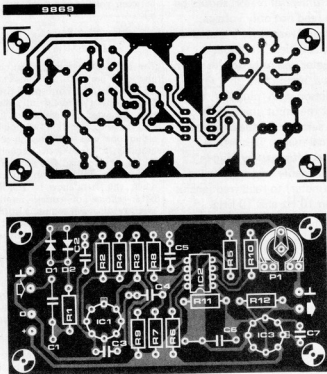
circuit is built up using the PC board layout given in figure 2. Bear in mind that the human body will not necessarily survive contact with input voltages that may not damage the adequately-rated input blocking capacitor. If one contemplates measuring the frequency of such high voltages the circuit should be assembled in a well-insulated box!

The power supply need not need to be regulated, so it can be kept very simple. A transformer secondary of 12 volts, a bridge rectifier and a 470 μ/25 V reservoir electrolytic will do the job nicely. Although a circuit that draws 25 mA is not too well suited to battery supply, one may need or wish to do this. In this case the battery should be bridged by a low-leakage (e.g. tantalum) 10 μ/25 V capacitor to provide a low AC source impedance.

A few specifications:

frequency range:	10 Hz ... 10 kHz
input impedance:	> 560 k
sensitivity:	50 mV p-p
max input voltage:	400 V peak
minimum load on output:	5 k (if 10 V out required)

Figure 1. The input frequency is passed via a comparator (limiter) IC1 to a frequency-to-voltage converter (IC2, 4151), that delivers a DC voltage via buffer IC3 to a normal multimeter.

Figure 2. Printed circuit board and component layout for the frequency 'add-on' (EPS 9869).

Calibration

The calibration can really only be done with an accurate generator.

A 10 kHz signal is fed to the input and P1 is set to bring the multimeter to full scale deflection (e.g. 10 V). That completes the calibration — although it is wise to check that the circuit is operating correctly by using lower input frequencies and observing whether the meter reading is also (proportionately) lower.

DIGITAL MULTIMETER

Micronic Devices market a hand-held digital multimeter, model 130-A, which has a 3½-digit liquid-crystal-display and is suited for field service applications. It features 0.25% basic DCV accuracy and measures upto 10A, A/C or D/C, and performs five functions on 26 ranges. It has also diode check capability, overload protection and both polarity and low-battery indicators.

Further details from

Micronic Devices
54/1, Old Rajinder Nagar
New Delhi 110 060

INSULATING SPACERS

Plastic insulating spacers are manufactured by Instrument Control Devices; they have a variety of applications such as spacers, stand offs, sleeves, etc. Moulded from high grade, low-moisture-absorption plastic, they are insoluble in common solvents. They are available in ¼, ½, ¾ and 1.0 inch lengths. The 3.5 round hole can accommodate ⅛ inch, 6BA or M3 screws.
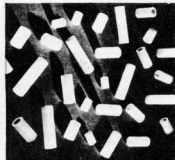
Further details from

Instrument Control Devices
14, Manorama Niwas, Datar Colony
Bhandup, Bombay 400 078

DIODE & TRANSISTOR TESTER

Spectron market a diode and transistor breakdown voltage tester to measure reverse breakdown voltage of diodes and collector-emitter and collector-base breakdown voltage of transistors.

Two modesl are available, the popular one being the direct constant current type of measurement on analogue panel meters. This tester measures breakdown voltage upto 1000 volts at constant currents settable from 0 to 10 mA D.C.

Detailed information from

Spectron Sales & Service Pvt. Ltd.
63, Bharatkunj No. 2
Erandvane, Pune 411 038

CABLE BINDING CORD

Novoflex has introduced their cable binding lacing cord for use in cablings in electronic instruments, electro-technical controls, etc. Being made of soft elastic PVC, it grips the cable bundle firmly. It has high tensile strength, is resistant to chemicals and weather and can withstand temperatures upto 90°C.
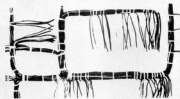
For more information, write to

Novoflex Cable Care Systems
Post Box No. 9159
Calcutta 700 016

LCR METER

Digital LCR meters, introduced by Vasavi Electronics, facilitate easy measurements unlike the conventional bridges which call for tedious bridge balancing. Salient features of the meters include fast, direct reading of dissipation factor for checking the quality of the component. Vasavi Electronics point out that a guard terminal is provided for low capacity measurement to avoid errors caused
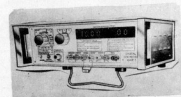
by stray efects and to eliminate errors due to leads. Four terminals are provided for large capacitance and low resistance measurements.

Contact for details

Vasavi Electronics,
162, Vasavi Nagar,
Secunderabad 500 003.

POWER SUPPLY

Rashmi Electronics have developed a "Regulated power supply" in DC range O to 15V for one amp and two amps current capacity. Protected against short circuit and over load, the instrument is designed to yeld continuously variable, regulated output with ripple in output less than 1mV RMS. Fu

Further details can be had from

Rashmi Electronics,
2-15-34, Kadrabad,
Polas Lane Corner,
Jalna 431 203.

INSECT KILLER

Electronics in insect control is replacing the role played by toxic and hazardous chemicals.

"Ventura" is an electronic unit comprising blue light which attracts the insects, electrified grid which electrocutes insects, coming in contact, instantaneously and a collection tray where the dead insects fall. Green hammertone paint is given for the top cover while the protective cage is chrome-plated.

Two models of Ventura with two blue ultra violet tubes (model 150− 485x350 135 mm and model 200− 635 x 500 x 200 mm) cover an effective area of 1,000 sq. ft. and 1,200 sq. ft. respectively, claim the manufacturers.
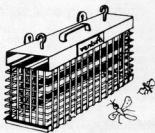
More information from

K.D. Chambers,
Near H.L. Commerce College,
6 Road Junction, Navrangpura,
Ahmedabad 380 009

market

RF CONVERTERS

Viprof Electronics have introduced a video converter, model VCX 81, used for interfacing UHF VCRs TVs (Indian make). The instrument is also applicable for personal computers like Sinclair ZX spectrum, ZX 81, TRS 80, APPLE II, VIC 20 ATARI video games, etc. Specifications: Power – AC 220V/50Hz, 2W. Input - Video: 1 VP-P/10k, audio: 100 mV/100 k. Output - 250 mV, 75 COAX. Inputs and outputs through RCA jacks.
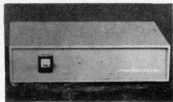
Further information from
Viprof Electronics
63, 1st Floor, Noor Bldg.
J.C. Road, Bangalore 560 002

WIRE STRIPPER

Efficient Engineering have placed on the market their new product, Stripolean' hand-held thermal wire stripper, designed specially for the electronic industry. It removes even the toughest high-temperature insulation with ease according to the manufacturers, and leaves wires free of oxides, nicks or deformations. It has an adjustable strip and wire gauge stop and works on 24V supply.

More particulars can be had from
Efficient Engineering
263, 8th Road, Khar
Bombay 400 052

FIBREGLASS CABLES

Chowdhry Instrumentation manufacture fibreglass insulated cables, consisting of copper stranded wire having cross-sectional area from 0.5 mm² to 400 mm², overlapped with fibreglass, braided with fibreglass yarn and impregnated with high thermal class insulating thermosetting varnish. They resist moisture, chemicals, flames, fungus radiation, acid and ozone attacks. They are recommended by the manufacturers for use where cables have to withstand temperatures upto 300°C.

For further details, contact
Chowdhry Instrumentation Pvt. Ltd.
110, Model Basti
New Delhi 110 005

RADIO CLOCK

Electronics Hobby Centre have developed a radio-clock. The unit is made in white acrylic, with the front panel in black. It works on 3 pen-cells. The manufacturers have developed several other combinations like this one.
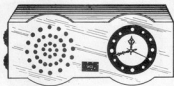
More details from
Electronics Hobby Centre
F-37, Nand-dham Industrial Estate,
Marol
Bombay 400 059

IN/OUT INDICATOR

Subtronics have introduced an electronic digital 'in-out' indicator to replace the convential slide-type indicator. 3" x 3" in size, it works on 230V A.C. mains with very low consumption and can be mounted on door-frames. It has self-illuminated digital display (LED - seven segment), visible easily in darkness and can be remotely operated at long distance by a slide switch.
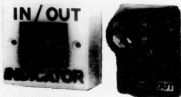
Further particulars from
Subtronics
"Kailandas Udyog Bhavan"
147, Near Century Bazar, Worli
Bombay 400 025

GEARED MOTOR

Vishal Electromag Industries have introduced a miniature reversible synchronous geared motor, claimed by them as most sturdy, compact and requiring no capacitor for starting or running. The motor operates on 240V or 110V (or any other volts), +10%V, 50 HZ supply and consumes less than 2.5 Watts. Basic motor speed is 375 RPM and with suitable geartrain output speed can be available from 1 sec/rev. to 168 hrs./rev. It can be directly used as a drive for motorised potentiometers, etc.
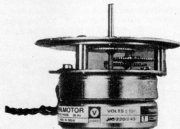
For further information, write to
Vishal Electromag Industries
Unit No. 11, Pavan Nagar, Shukla Compound
Near Morabli Talao,
Western Express Highway
Dahisar East, Bombay 400 068

RF CONNECTOR

"PHP" connector from ITT Cannonis a new high performance RF coaxial connector which can be push mated instantly for blind mate applications. It allows modularised RF interconnections to several RF lines in a very dense environment. Mating is accomplished in a single push mate operation. single or multiple PHPs may be gang mounted in a plate or readily adapted to fit into any Cannon circular or regtangular connector.

"PHP"
HIGH PERFORMANCE
COAXIAL CONNECTOR

For more information, contact
Jost's Engineering Co. Ltd.
60, Sir P.M. Road
Bombay 400 001

PUSH BUTTON TELEPHONE

This push button telephone from Product Promoters is a solid-state, IC-based electronic handyphone. Instead of dialling, the numbers are 'pressed' and are fed into a memory. If the number is engaged, re-dialling is effected bymerely pressing the memory switch. It can be kept on a table or wall-mounted and comes in four colours.

More information can be had from
Product Promoters
P.O. Box 3577, F-41 Lajpat Nagar II
New Delhi 110 024

classified ads

advertisers index

missing link

programmable disco light display
(March 1984, page 3-27)

triac control board
(April 1984, page 4-22)

Where the triac boards are not fitted in the same case as the light display itself (for instance, to prevent mains on the multi-core cable or to enable the use of a relatively thin cable where the lamp current is high) it is best to build the power supply for the zero voltage detector separately. The multi-core cable to the triac boards then carries only the supplies to the LED channels and the common return (all low voltage).

capacitance meter
(March 1984, page 3-52)

Pulse generator
(May 1984, page 5-26)

The front panel foil for these instruments is provided with a very thin protective layer. This layer is not easily seen, but it should be removed after final assembly. Not doing so renders the front panel rather less good-looking owing to small scratches and minute air bubbles.