24 page supplement inside **FREE**

elektor

up-to-date electronics for lab and leisure

tv scope
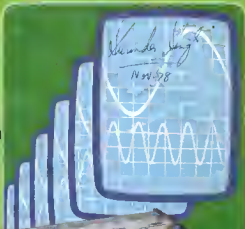converts a tv set into an oscilloscope

reverberation unit
using bucket brigade memories

central alarm system
based on a time division multiplexed bus

resonance filter
for music synthesisers

plus:
· proximity detector
· dc modulator
· getting your bus buffer
· playing all the usual features

UK 4 — elektor october 1978

decoder

elektor 42 decoder

Volume 4   Number 10

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What is a TUN?
What is 10 n?
What is the EPS service?
What is the TO service?
What is a missing link?

Semiconductor types
Very often, a large number of equivalent semiconductors exist with different type numbers. For this reason, 'abbreviated' type numbers are used in Elektor wherever possible
• '741' stand for µA741, LM741, MC641, MIC741, RM741, SN72741, etc.
• 'TUP' or 'TUN' (Transistor, Universal, PNP or NPN respectively) stand for any low frequency silicon transistor that meets the following specifications:

UCEO, max	20V
IC, max	100 mA
hfe, min	100
Ptot, max	100 mW
fT, min	100 MHz

Some 'TUN's are: BC107, BC108 and BC109 families; 2N3856A, 2N3859, 2N3860, 2N3904, 2N3947, 2N4124. Some 'TUP's are: BC177 and BC178 families; BC179 family with the exception of BC159 and BC179, 2N2412, 2N3251, 2N3906, 2N4126, 2N4291.
• 'DUS' or 'DUG' (Diode Universal, Silicon or Germanium respectively) stands for any diode that meets the following specifications:

	OUS	DUG
UR, max	25V	20V
IF, max	100mA	35mA
IR, max	1µA	100 nA
Ptot, max	250mW	250mW
Cd, max	5pF	10pF

Some 'DUS' are: BA127, BA217, BA218, BA221, BA222, BA317, BA318, BAX13, BAY61, 1N914, 1N4148.
Some 'DUG's are: OA85, OA91, OA95, AA116.
• 'BC108', 'BC237B', 'BC547B' will refer to the same 'family' of almost identical better-quality silicon transistors. In general, any other member of the same family can be used instead.
BC107 (-8, -9) families
BC107 (-8, -9); BC147 (-8, -9)
BC207 (-8, -9); BC237 (-8, -9)
BC317 (-8, -9); BC347 (-8, -9)
BC547 (-8, -9); BC171 (-2, -3)
BC182 (-3, -4), BC382 (-3, -4)
BC437 (-8, -9); BC414
BC177 (-8, -9) families
BC177 (-8, -9); BC157 (-8, -9)
BC204 (-5, -6); BC307 (-8, -9)
BC320 (-1, -2), BC350 (-1, -2),
BC557 (-8, -9); BC251 (-2, -3)
BC212 (-3, -4), BC512 (-3, -4),
BC261 (-2, -3); BC416.

Resistor and capacitor values
When giving component values, decimal points and large numbers

of zeros are avoided wherever possible. The decimal point is usually replaced by one of the following abbreviations:
p (pico-) = 10^{-12}
n (nano-) = 10^{-9}
µ (micro-) = 10^{-6}
m (milli-) = 10^{-3}
k (kilo-) = 10^3
M (mega-) = 10^6
G (giga-) = 10^9
A few examples:
Resistance value 2k7: 2700 Ω.
Resistance value 470: 470 Ω.
Capacitance value 4p7: 4.7 pF, or 0.000 000 000 004 7 F . . .
Capacitance value 10n: this is the international way of writing 10,000 pF or .01 µF, since 1 n is 10^{-9} farads or 1000 pF
Resistors are ¼ Watt 5% carbon types, unless otherwise specified.
The DC working voltage of capacitors (other than electro-lytics) is normally assumed to be at least 60 V. As a rule of thumb, a safe value is usually approximately twice the DC supply voltage.

Test voltages
The DC test voltages are measured with a 20 kΩ/V instrument, unless otherwise specified.

U, not V
The international letter symbol 'U' for voltage is often used instead of the ambiguous 'V'. 'V' is reserved for 'volts'. For instance: U_b = 10 V, not V_b = 10 V.

Mains voltages
No mains (power line) voltages are listed in Elektor circuits. It is assumed that our readers know what voltage is standard in their part of the world!
Readers in countries that use 60 Hz should note that Elektor circuits are designed for 50 Hz operation. This will not normally be a problem; however, in cases where the mains frequency is used for synchronisation some modification may be required.

Technical services to readers
• EPS service Many Elektor articles include a lay-out for a printed circuit board. Some — but not all — of these boards are available ready-etched and predrilled. The EPS print service list in the current issue always gives a complete list of available boards.
• Technical queries. Letters with technical queries should be addressed to: Dept. TQ. Please enclose a stamped, self-addressed envelope; readers outside U.K. please enclose an IRC instead of stamps.
• Missing link. Any important modifications, additions to, improvements on or corrections in Elektor circuits are generally listed under the heading 'Missing Link' at the earliest opportunity.

contents

The Central Alarm System (CAS) will relay an alarm indication from a number of remote stations to a central location along a common bus system. Audible indication of the alarm is provided together with a visual display of which station has sent the alarm. The applications for this system are limited only by the ingenuity of the reader.   p. 10-20

As explained in the accompanying introductory article, there are two different versions of the TV scope . This month, the practical circuit and constructional details of the basic version are described; the extended or 'de luxe' scope will be discussed in a second article to be published next month.   p. 10-30
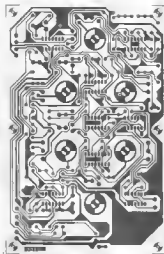
Until comparatively recently the only audio delay units that were within the budget of most home constructors were of the spring line type. Recently, however, completely electronic delays have become a feasible proposition. A design for a digital reverberation unit has already been published in Elektor. The analogue reverberation unit represents an alternative approach.   p. 10-44

An oscilloscope is, without doubt, the single most useful piece of electronic test equipment and is, for many tasks, virtually indispensable. By using a normal, domestic TV set as the display the TV scope allows the construction of an oscilloscope for a very modest outlay.

eps printservice

Many Elektor circuits are accompanied by designs for printed circuits. For those who do not feel inclined to etch their own printed circuit boards. Delivery time is approximately three weeks, U.K./eire mail.
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Way Postal Center, P.O. Box 80689, Los Angeles,
Cal 90080, A/C no. 12324-07007
Prices include post & packing, surface mail. If air mail,
please add $ 0.50

Example
digisplay   (E13) 9376-1   3.10 6.90
              1   2   3   4   5   6   7

1 Circuit title
2 reference to associated article(s)
3 board number
4 p.c.b.'s without solder mask
5 U.K.
6 price (£) inclusive of VAT
   (12.5% VAT) otherwise 8% VAT
   U.S.A. and Canada
7 price in $, not including tax (if any)

		£	£	$
E1: DECEMBER 1974				
equa amplifier	1499*	1.80*		3.95
E3: APRIL 1975				
disc preamp 7631	4040A*	1.25		2.90
E7: NOVEMBER 1975				
ota amplifier	6029*	1.40*		3.10
frequency counter, minitron display board	9312	1.45		3.20
frequency counter, led display board	9313	1.45		3.10
frequency counter, clutch/display driver	9314	1.20		2.60
E9: DECEMBER 1975				
dia	9169	0.85		1.90
730/740 (IC control amplifier)	9191	1.40*		3.10
E12: APRIL 1976				
stylus balance	9343	0.45*		1.05
rhythm generator (M253AA)	9344-1	1.45*		3.20
ic drummer, instruments mother board	9344-2	2.00*		4.40
ic drummer, instruments daughter board	9344-3	0.35*		0.80
UAA170, 270° meter, basic board (E17)	9392-1	1.50		3.30
UAA170, 270° meter front panel (E17)	9392-2	2.00		4.40
UAA170, LED extender, basic board	9392-3	1.15		2.50
UAA170, 16 LED display board	9392-4	0.95		2.10
E13: MAY 1976				
preco, preamplifier	9398	2.40*		5.35
preco, control amplifier	9399	1.40*		3.10
equin	9401	2.65*		5.85
E17: SEPTEMBER 1976				
tachometer, LED board (E12)	9392-1	1.50		3.30
tachometer, front panel (E12)	9392-2	2.00		4.40
tachometer, control board	9460	1.45		3.15
windscreen wiper delay, using SCR	9474-1	0.70		1.50
E19: NOVEMBER 1976				
intercom	9156	1.05		2.30
albar; see E25, May 1977				
albar, power supply	9157	1.15		2.50

		£	£	$
signal horn	9438	1.05		2.30
sensitive metal detector	9750	1.80		3.95
E20: DECEMBER 1976				
perking mortar alarm	9491	0.80		2.00
E21: JANUARY 1977				
stereo audio mixer	9444	6.55*		14.50
marbles	9753	1.55		3.40
E22: FEBRUARY 1977				
local radio AF	9499-1	0.70*		1.55
local radio osc	9499-2	0.70		1.55
E23: MARCH 1977				
OTA-ALA (E7, E24)	6029	1.40*		3.10
scope calibrator	9455	0.30		0.85
run rabbit, run!	9764	4.15		9.15
E24: APRIL 1977				
Digisplay (E13)	9378	3.10		8.90
LED VU/PPM				
rectifier section	9419-1	2.35*		6.15
display section	9419-2	4.50*		9.95
auto slide changer	9743	0.65		1.45
morse injector	9759	7.35		16.25
signal injector	9785	0.50		1.10
E25: MAY 1977				
modal car track (E24)				
multiplex encoder	9796-1	3.10		6.85
infra-red transmitter	9796-2	1.60		3.55
single transducer/receiver (E19)				
receiver	9815-1	1.25		2.80
transmitter	9815-2	0.40		0.85
E26: JUNE 1977				
precision timebase	9448	1.55		3.40
power supply for precision timebase	9448-1	0.60		1.30
formant keyboard interface (E25)	9721-1	2.80*		5.75
levitator	9763	2.35		5.20
active loudspeaker cross over filters (E25)	9786	1.95*		4.30
model car track (E24, E25)				
servo amplifier	9803	0.35		0.80
E27/28: SUMMER CIRCUITS 1977				
knotted handkerchief	9830	0.20		0.45
distortion meter	7005	0.65		1.45
guitar preamp	7020	1.10*		2.45
0 ... 10 V supply	7059	0.95		2.10
x-equaliser	7071	0.75		1.65
automatic NiCad charger	7075	0.65		1.45
TV gamel	7084	2.05		4.50
4 W car radio amplifier	7101	0.65*		1.40
3½ digit DVM	7109	2.15		4.70
E29: SEPTEMBER 1977				
FM mains intercom	9359	4.85		10.75
formant (E25, E26)				
interface receiver	9721-1	1.30*		2.80
power supply	9721-2	4.15*		9.15
keyboard divider (one octave)	9721-4	0.75*		1.70
logarithmic darkroom timer	9797	1.50		3.35
CCIR test pattern generator, motherboard	9800-1	5.95		13.10
modal car track (E24 · E26)				
motor speed control	9829-1	0.55		1.20
proptester	9829-2	1.55		3.45
E30: OCTOBER 1977				
CCIR				
daughter board 1 (E29)	9800-2	3.30		7.35
infrared transmitter	9822	2.50*		5.55
Kirlian camera (two boards)	4523 } 9831	7.25		5.00
model car track (E24 · 26, 29)				
infrared receiver	9848-1	0.80		1.80
multiplex decoder	9848-2	0.85		1.90
E31: NOVEMBER 1977				
formant (E25 · E30):				
VCO	9723-1	8.30*		18.35
CCIR: daughter board 2 (E29, E30)	9800-3	2.55		5.65
video biofeedback.				
display generator	9825-1	2.55		5.60
alpha amplifier	9825-2	2.10		4.80
electrometer	9826-1	0.70		1.50
sensors for electrometer	9826-2	0.65		1.40
magnetiser	9827	0.65		1.45

		£	£	$
experimenting with the SC/MP				
RAM-I/O	9946-1	5.75		12.70
SC/MP board	9946-2	1.80		3.95
model car track (E24 · E30).				
transmitter power supply	9919	1.00		2.20
chassis	9850	2.85		6.35
E32: DECEMBER 1977				
formant (E25 · E31):				
VCF	9724-1	3.40*		7.45
experimenting with the SC/MP (E31): CPU	9851a	8.50		18.80
UHF TV modulator	9844	1.35		2.95
electro thrombolier				
preamplifier	9866	0.80*		1.90
analogue frequency meter	9869	1.10		2.40
heating controller	9877	3.20		7.05
MW radios receiver	9890	0.85		1.85
sensitive lightmeter	8888	0.60		1.30
stylus balance (E12)	9343	0.45*		1.05
signal injector (E24)	9785	0.50		1.10
knotted handkerchief (E27/E28)	9830	0.20		0.45
E33: JANUARY 1978				
simple function generator	9453	2.70		6.00
formant (E25 · E32)				
ADSR	9725-1	3.20		7.05
UAA180 LED voltmeter	9837	1.75		3.90
alaktor equaliser	9832	3.45*		7.65
experimenting with the SC/MP (E31, E32)				
bus board	9857	3.00		5.85
memory card	9858	12.75		28.20
CMOS noise generator	9859	1.45		3.20
peak programme meter	9860	1.30*		2.80
E34: FEBRUARY 1978				
formant (E25 · E33).				
DUAL VCA	9726-1	3.35		7.40
development timer	9840	2.10		4.60
experimenting with the SC/MP (E31 · E33).				
HEX I/O	9893	17.00		37.60
infra-red light gate				
transmitter	9862-1	0.65		1.40
receiver	9862-2	0.85		1.85
E35: MARCH 1978				
formant (E25 · E34):				
LFO module	9727-1	3.55		7.90
noise module	9726-1	3.05		6.75
4 k RAM card	9891	14.70		32.90
colour TV games*				
VHF games board	9892	2.10		4.65
UHF converter	9844	1.35		2.95
SC/MP power supply	9906	3.10		6.80
E36: APRIL 1978				
formant (E25 · E35).				
COM	9729-4	2.80*		6.20
elektornado	9894	2.95*		6.50
cassette interface	9905	2.50		5.50
moving coil preamp	9911	3.30*		7.35
front panels for Formant				
interface	9721-1	1.40		3.05
VCO	8723-2	1.40		3.05
VCF	9724-2	1.40		3.05
ADSR	9725-2	1.40		3.05
DUAL-VCA	9726-2	1.40		3.05
LFOs	9727-2	1.40		3.05
NOISE	9726-2	1.40		3.05
COM	9729-2	1.40		3.05
Complete set of 11 panels (includes 3 × VCO and 2 × ADSR). £ 13.85 or $ 30.55				
E37: MAY 1978				
colour modulator	9873	1.80		3.85
percolator sensors	9902	0.85		1.85
digital reverberation unit, main board	9913-1	4.15*		9.20
axtension board	9913-2	1.20*		2.50
mm short-wave receiver	9920	1.50*		3.35
E38: JUNE 1978				
simple function generator (E33)	9453-6	1.50		3.30
1/4 GHz counter				
timebase and control	9887-1	7.85		17.40
counter and display	9887-2	8.95		19.75
low freq. input amp	9887-3	1.40		3.05
high freq. input amp	9887-4	1.30		2.90
automatic mono/stereo switch	9923	0.90*		2.00
TV sound modulator	9925	0.95		2.05
mini-counter	9927	7.50		5.50

SELEKTOR

Improving radiation treatment planning

The value of whole body X-ray scanning by computerised tomography as an advanced diagnostic tool in oncology is already well established. Now scientists at the Institute of Cancer Research with their medical colleagues at the Royal Marsden Hospital, Sutton, Surrey, are carrying out clinical evaluation trials which they hope will demonstrate that the detailed anatomical data produced by an EMI whole body scanner can also be used to improve the effectiveness of radiation treatment given to cancer patients.

The key factor in planning treatments for individual patients is to ensure that the optimum dose of radiation covers the target area without unduly damaging adjoining healthy tissue.

This task is now being made easier, and tumours pin-pointed with far greater accuracy, by feeding the cross-sectional scan pictures produced by the scanner directly into a radiotherapy planning computer system known as RAD-8.

A patient's treatment plan can then be calculated by adjusting the position of the radiation beams in relation to the CT scan picture displayed simultaneously on a TV-type screen. Isodose curves showing the amount of radiation reaching the tumour and the surrounding areas can then be plotted with a high degree of accuracy.

An on-line computer for Radiotherapy Planning was first programmed at the Royal Marsden Hospital in 1969 by the Institute of Cancer Research's own Computer Group. It has since gone into use in hospitals in many parts of the world. At the Royal Marsden, where the system is in routine clinical use, anatomical data on individual patients has up to now been obtained by using conventional X-ray and ultrasonic scans. The outline of the patient's body is then simulated from measurements and the tumour area sketched in as accurately as possible.

But using CT scans an accurate cross-sectional picture can be fed directly into the RAD-8 system. These show clearly soft tissue and other detail normally not discernible on conventional X-ray film and enable absorption factors likely to vary the treatment dose reaching the malignancy to be accurately calculated. Dr Roy Parker, ICR physicist in charge of the project, says: 'We can now pin-point a tumour in relation to other organs with considerable accuracy. With such data and the carefully calculated X-ray absorption factors for all the parts of the anatomy through which treatment beams have to pass to reach the tumour, we now have for the first time a springboard for really high-precision radiotherapy'.

For the Royal Marsden trials, carefully selected patients who have already had treatment plans drawn up using routine procedures, are having a second plan prepared with a CT scan and the RAD-8 system.

Of the first 50 specially selected patients examined in this way about two thirds needed significant changes to their treatment plans. In about half of these this was due to inaccuracies in patient outlines or because normal internal organs were not precisely where expected. Many organs, like kidneys, are highly 'mobile' and it is only by taking scan X-rays when the patient is breathing normally that they can be correctly located. In the remainder, changes were necessary because the scan showed extensions to the tumour not suspected by other techniques; accurate localisation has also led to reduction of the treatment volume.

In operation the radiotherapy planning system is used like this: The radiotherapy scan slice selected for planning is transferred directly from the scanner computer to a floppy disc in a diagnostic display console (DDC). The image can then be viewed simultaneously on the DDC's screen in a grey-scale mode on a colour TV monitor using specially developed hardware. This TV monitor acts as a link between the RAD-8 system and the CT scan picture. First the outline of the patient is automatically read into the system. Then the shape and position of relevant internal anatomy and the malignant area to be treated are transferred into the computer memory by using a light sensitive pen to trace round the outline displayed on the colour TV screen. As the pen is moved over the screen its path is recorded in red and superimposed on the scan. It is then permanently stored on another floppy disc within the RAD-8.

The arrangement of the treatment beams is then planned interactively. Throughout the process the operator is guided by a series of questions displayed on the TV screen. The system enables the beam positions (in green) and the isodose levels (in red) to be shown on the TV monitor superimposed on the CT grey scale image. This is of great assistance in relating the isodose distribution to the anatomy of the patient whilst actually lying in the same position as they will be treated. Once the best treatment plan has been arrived at, it is printed out, verified by using a radiotherapy simulator, and then actual treatment can start.

Although clinical experience using the system is still extremely limited, it is already evident that these new aids improve the accuracy and ease with which plans can be tailored to suit individual patients. Initial clinical findings on a highly selected group of patients strongly suggest that the system will make a significant contribution to radiotherapy treatment planning.

The trials now taking place at the Royal Marsden Hospital, Sutton, with the whole-body scanner and associated systems are a collaborative effort between the Institute of Cancer Research, the Department of Health and Social Security, the Cancer Research Campaign and EMI Limited.

[360 S]

SELEKTOR

North London Hobby Computer Club

On Wednesday October 5th 1978, at 6.30 p.m., the inaugural meeting of the North London Hobby Computer Club will be held in Room 47 in the Old Building at Holloway Road, just opposite Holloway Road underground station on the Piccadilly line.

The Polytechnic of North London, and its Department of Electronic and Communications Engineering in particular, have made available many resources for this venture. Within the Department there are two PETs (with a third coming), four SWTPc 6800 computer systems with floppy discs, printers and VDUs, and some KIM and Motorola microcomputer systems. Most of these will be available for use, as will some PETs and SWTPc systems in other departments.

As the club is envisaged at present, little 'homebrew' activities are anticipated before Christmas, with any meetings centered around talks by manufacturers and discussions on programming, etc. However, from the new year it is anticipated that three sets of activities will be running concurrently — or sequentially, depending on how many people turn up! These are short courses on programming, Basic and machine level; a 'homebrew' section using the facilities of the Department (up to 35 people can solder and test at the same time); and introductory talks and discussions for those anticipating their own systems. The varied programme should be of interest to a wide variety of people. Obviously, students from the Polytechnic will be coming to the meeting, but the organisers emphasise that this club is open to all those who are interested. The Polytechnic will be providing some backup, especially with expert staff and other facilities. The club is intended as part of the Community Development Programme that has recently been instituted.

Department of Electronic and Communications Engineering,
The Polytechnic of North London

(369 S)

UK phone users dial the world

Telephone subscribers in Britain have become the first in the world to be able to dial directly to 50 countries and can now reach 76 different countries without going through an exchange. Announcing this in its report for the year ended 31 March 1978, the UK Post Office says the country has

850 electronic telephone exchanges in service and is now investing heavily in a computerised telephone exchange of the future known as System X.

The Post Office is placing its first production orders this year for System X which is expected to come into service in 1981. The development of these new electronic exchanges is said to be the most important part of the corporation's research programme which was last year financed to the tune of more than £56 million.

The report reveals that the Post Office invested a total of £842 million in the year under review on new and improved telecommunication buildings and plant. There are now more than 23 million phones in use in Britain and over 17,000 million calls were made last year. The international sector was easily the fastest growing, with calls up more than 27 per cent from 118 million to 150 million.

Capital investment

Chairman Sir William Barlow says the Post Office continues to have one of the largest capital investment programmes in the UK. Last year it invested a total of £870 million and this will rise to £1,000 million this year. He said the £367.7 million profit made in 1977-78 would be used to help finance this capital investment programme.

The British postal system is one of the few such systems in the world now making a profit. Last year's profit was the third successive one for the UK corporation and represented a 26.2 per cent rise over the previous year. System X is not the only new technology being pioneered by UK telecommunications experts. Telephone calls are being carried in Britain through tiny strands of glass called optical fibres. The calls are transmitted as pulses of light.

The Post Office is also planning to introduce next year a system whereby the telephone in the home can be used to call up information that is then transmitted via the domestic television screen.

Britain's postmen also had their successes. Last year letters posted in the UK rose by more than one per cent to 9,484 million and of those sent by first class mail more than 92 per cent were delivered the day after posting. With the help of Concorde, the postmen also offered the fastest mail-run in history — delivering urgent letters from London to New York in three and a half hours. The UK Post Office employs a total of 422,000 people and has an annual turnover of more than £4 billion.

(362 S)

Hot wire!

A new robust cable developed in Britain continues to function - lighting an electric lamp - despite being subjected to over 30 minutes of continuous gas flame at 750°C.

Called the FP200 cable, it has been designed to withstand three hours of continuous burning at this temperature, but in tests has continued to function after 6 hours! In this type of cable the conducting core is surrounded by silicone rubber that is not destroyed when subjected to fire, but burns to a hard white material that is just as effective an insulator as the original silicone. This keeps the wires apart and prevents short circuits.

The cable is particularly suitable for wiring safety circuits such as fire alarms and emergency lighting where high standards of reliability and performance are required in the event of fire. In one striking example a factory wired with FP200 was recently gutted, but all the overhead lights continued to burn. An order has recently been completed for Britain's Central Electricity Generating Board for the Dungeness 'B' nuclear power station, where 50 km of FP200 will be used for the vitally important reactor emergency shutdown control circuits.

Apart from its fire resistant qualities, the new cable is surge resistant, moisture resistant and does not 'age'.

Pirelli General Cable Works Ltd.
PO Box 4
Western Esplanade
Southampton, SO9 7AE

(359 S)

an introduction to the TV scope

An oscilloscope is, without doubt, the single most useful piece of electronic test equipment and is, for many tasks, virtually indispensable. Unfortunately, the high cost of oscilloscopes puts them beyond the reach of many electronics enthusiasts. A major proportion of the cost of an oscilloscope is accounted for by the cathode-ray tube and its associated high-voltage power supplies. By using a normal, domestic TV set as the display the TV scope eliminates this cost and allows the construction of an oscilloscope for a very modest outlay. The principles of the TV scope and a basic version of the instrument are described this month. Next month's issue will discuss the extension of the basic TV scope into a 'deluxe' version.

The basic principle of a TV scope was discussed in Elektor No. 37, May 1978 and has since been developed into a pratical system in the Elektor laboratory. To understand how the system works it will be useful first to take a look at how a normal TV picture is formed on the screen.

The image on a TV screen is built up by an electron beam scanning across the phosphor coated screen of the CRT in the manner shown in figure 1. The scan proceeds from top to bottom of the picture in a zig-zag fashion, each left-right horizontal sweep being known as a *line*. The duration of each line scan is 64 μs, and the line frequency is therefore 15.625 kHz, which is an important figure to remember, as will be seen later. Each complete image or *frame* of the picture is made up of 625 lines, and the picture frequency is therefore 25 Hz. However, this low picture frequency could give rise to noticeable flicker, and to minimise this effect each frame of the picture is scanned, not in a single 625 line scan, but in two *fields* of 312½ lines each. These two fields are fully *interlaced*, i.e. lines of the even field fall between those of the odd field, to build up a complete frame of 625 lines. Field frequency is twice frame frequency, i.e. 50 Hz, another important figure to remember.

The tonal gradation of the picture is, of course, produced by varying the electron beam current and hence the brightness with which the phosphor glows. Maximum beam current produces the brightest (white) areas of the picture, whilst zero beam current produces the black areas.

To understand how the TV scope works the TV set it is best to imagine that the TV set is turned on its side so that the electron beam scans, not from left to right, but up and down, as shown in figure 2. In practice it may or may not be possible actually to turn the set on its side, depending on the cabinet design, ventilation etc.

In figure 2 a sinewave signal is displayed
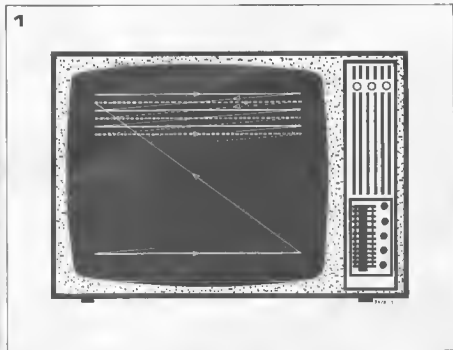
Figure 1. How an image is formed on the TV screen. The frame is composed of two interlaced fields, each consisting of 312.5 lines. One field is shown in bold, the other in dotted lines.

on the TV screen. This is achieved by taking a sample of the instantaneous amplitude of the signal every 64 μs, i.e. during each line scan. Each sample is then displayed on the screen as a white spot whose position along the line scan is proportional to the instantaneous amplitude of the signal. In this way the display is built up from a number of such white spots, the rest of the screen being black.

Figure 3 shows the video waveform for one line of such a display, which consists of nothing more than a single white level pulse in an otherwise black level signal, plus, of course, line sync pulses which are below black level. The position of the white level pulse along the line sweep, i.e. the time at which it occurs after the sync pulse, must be proportional to the sampled, instantaneous amplitude of the input signal. The circuit required to achieve this is a voltage-time converter.

Figure 4 illustrates how the input signal is sampled and converted into a white pulse. A ramp waveform is generated having the same repetition rate as the TV line frequency (15.625 kHz) and synchronous with it. This signal is fed to a voltage comparator along with the input signal. Whenever the ramp voltage exceeds the input voltage the output of the comparator will change state, this change being used to trigger a monostable multivibrator that produces the white level pulse. If the input voltage is low it will quickly be exceeded by the ramp voltage and the spot will appear low down on the screen. Conversely, if the input voltage is high it will not be exceeded by the ramp voltage until near the end of the line scan, and the spot will appear high on the screen. The voltage-time converter circuit will be discussed in detail when the circuit for the basic version of the TV scope is described. In addition to the sawtooth generator and comparator, which are the essential features of the basic TV scope, the TV scope must also be equipped with input amplifiers and attenuators to vary the sensitivity, as with a normal oscilloscope, and also with a synchronising pulse generator to provide the line and field sync pulses required by a TV set. These will also be discussed in the circuit description of the basic TV scope.

Possibilities and limitations of the basic TV scope

The facilities offered by the basic version of the TV scope are limited principally by the 'timebase'. Since the image of the displayed waveform is repeated during each field scan the timebase frequency is equal to the field frequency of 50 Hz. It is important here to note the difference between this and a normal TV picture. Whereas one 625-line frame of a normal TV picture is made up of two interlaced, 312½ line fields, successive fields of the TV scope
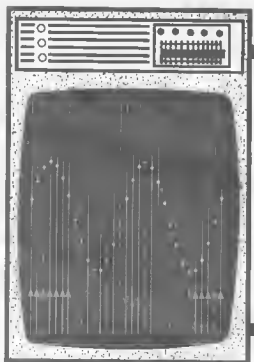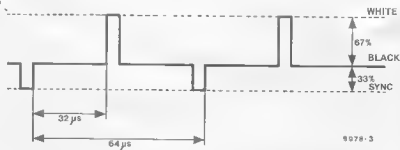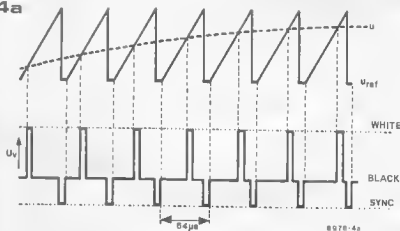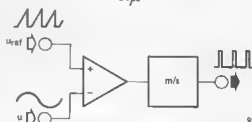
Figure 2. The TV scope can best be explained by imagining the TV es turned on its side, so that the spot is swept vertically up and down the screen. The signal trace is composed of a series of discrete spots, one per line of the frame. The point on the line where the spot appears is proportional to the amplitude of the input signal. For the sake of simplicity, only a few lines are shown.

Figure 3. The video signal for the first line scan of figure 2. Halfway along the line scan is a white level pulse.

Figure 4. This figure shows how the input signal, u, is converted into a video signal u_v with the aid of a sawtooth reference voltage u_{ref}.

Figure 5. One of the most important features of the basic version of the TV scope is the fixed timebase. These photographs show how this effects the traces obtained from signals of differing frequency.

Figure 6. The block diagram of the extended version of the TV scope. This in actual fact consists of the basic version of the scope preceded by two bucket brigade memories. It is possible to vary the frequency at which signals are read into these memories independently of the frequency at which they are read out; electronic switches are used in the circuit proper.

5a

50 Hz

250 Hz

5b

1 kHz

3 kHz

6

TV SCOPE, BASIC VERSION

Y-amplifier | Voltage-to-time converter | video section

EXTENSION

clock
memory
memory
clock

f_s | 15.625 kHz

9978-6
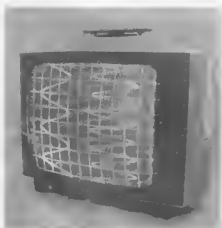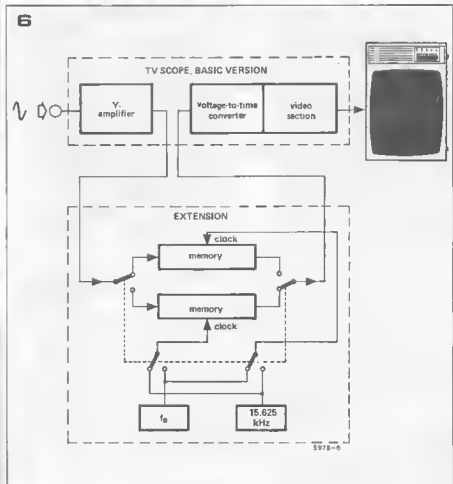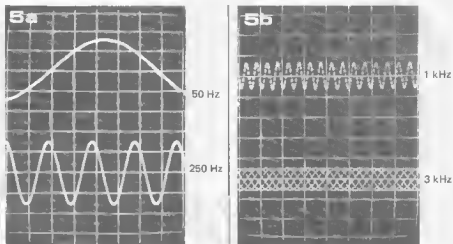
picture are not interlaced but are superimposed, and consist of 312 lines. Each field therefore makes up a complete image of the displayed signal.

The field frequency of the TV set is fixed at 50 Hz; it cannot be varied, (except by a small amount to adjust the 'vertical hold' of the set) nor can the field timebase be triggered. This fixed timebase naturally places limitations on the basic version of the scope. Firstly, a stable display will result only if the input signal frequency is a multiple of 50 Hz, e.g. 100 Hz, 150 Hz etc., so that it can then be synchronised with the timebase. This makes the basic TV scope less useful for applications where the signal frequency is outside the control of the user. However, for many

applications, where the signal source is the lab signal generator, test signals can often be chosen as multiples of 50 Hz.

The second limitation of the basic TV scope is its restricted frequency range, which is illustrated in the photographs of figure 5. The lowest input frequency that can be displayed is 50 Hz. One complete cycle of this signal is displayed, apart from some ten lens of the picture that are lost during the field blanking interval of the TV set. The single cycle of the 50 Hz signal which is displayed is made up of some 300 spots, and the picture is quite detailed.

As the signal frequency is increased, however, so the number of cycles appearing on the screen increases,

whilst the number of spots per cycle decreases. Figure 5b shows a 1 kHz signal (upper trace) and a 3 kHz signal (lower trace), displayed on the basic TV scope. In the case of the 3 kHz signal, not only are some 60 cycles of the waveform crammed onto the screen, but each cycle consists of only 5 spots. Clearly the highest input frequency that can be displayed is very much less than 3 kHz.

Deluxe version

The limitations of the basic version of the TV scope are a direct consequence of the fact that the signal is sampled at TV line frequency (15.625 kHz). The TV scope would be considerably more

7a

200 Hz

10 kHz

7b

4 kHz

10 kHz

8a

100 Hz

500 Hz

8b

100 Hz

500 Hz

Figure 7. The maximum input frequencies of both versions of the TV scope. The basic version (7a) will produce an acceptable trace at frequencies up to roughly 1.5 kHz. As can be gathered from figure 7b the extended version can handle much higher frequencies without any problems.

Figure 8. A comparison of the rise and fall times of the two versions of the scope based on their response to squarewave input signals show that the performance of the basic version is superior in this respect.

Figure 9. Displaying tone-burst signals is only possible with the extended version of the scope, since the basic version does not have the necessary trigger facilities.

Figure 10. This figure illustrates one possible application for the basic version of the TV scope: adjusting an amplifier for symmetrical clipping.

Figure 11. Another application for the (two-channel version of the) basic scope is comparing DC voltage levels.

Figure 12. A final example of a possible application for the basic scope is its use in conjunction with a sweep generator to display the frequency response of a circuit or system under test (x).

flexible if the sampling frequency could be varied independently of TV line frequency.

The solution is to store the waveform to be displayed in some form of memory, after which it can be read out of the memory at a suitable rate to be displayed on the basic TV scope. For example, if a 10 kHz signal were to be displayed then, say, one cycle of the waveform could be stored in the memory. By reading out the contents of the memory at TV line frequency, i.e. cycling through the contents of the memory in 20 ms, the stored signal would appear as a 50 Hz signal which could then be displayed on the TV scope with a resolution of 300 dots per cycle.

The use of a memory immediately removes the two main limitations of the basic TV scope, namely limited frequency range and lack of a trigger facility. Use of a memory allows any portion of the input signal to be stored (from a small part of one cycle to many cycles). Whatever the frequency of the original input signal, the stored signal can be read out of the memory at a 50 Hz rate in order to synchronise with the TV field timebase. The only limitation is the maximum input frequency that can be handled by the memory.

The term 'memory' has, until now, been used somewhat vaguely to describe a method of storing the input signal, so it is perhaps a good idea to look at what the memory entails in terms of hardware. One type of memory which could

be used is a digital memory such as a random access memory (RAM) or shift register. This type of memory can store only logic levels (0's and 1's), so it would obviously be necessary to sample the input signal, digitise it and store it in the memory as some form of binary code. This would require a sample-and-hold circuit and an A/D converter. Conversely, reading out of the memory would require a D/A converter to reconstitute the analogue signal. The maximum input frequency that such a system could handle would be limited mainly by the conversion rate of the A/D converter, and high conversion rates are not obtained cheaply.

The type of memory finally chosen for the deluxe TV scope was a bucket-brigade memory or analogue shift register. This type of memory has the advantage that it will accept an analogue input signal directly, sample it and transfer it from input to output as a sequence of charge packets. This removes the need for sample-and-hold circuits, A/D and D/A converters.

Little external circuitry is needed to operate an analogue shift register and devices are available that will accept input signals at relatively high frequencies — well beyond the upper limit of the audio spectrum. These advantages make analogue shift registers an excellent choice for this application.

Block diagram

A block diagram of the deluxe TV scope

is shown in figure 6. It will be seen that this in fact contains two memories, which may seem a little odd. However, it should be remembered that when information is being read out of the memory to be displayed then the memory will be continuously in use and furthermore will be tied to the timebase frequency of the TV set, that is to say it will never be available for storing the signal in the first place. The way out of this dilemma is to use two memories, so that while one is storing the signal the previously stored signal can be read out of the other. By switching between the two memories the signal displayed on the TV set will be continuously updated. This switching is, of course, performed electronically.

Since the deluxe version of the TV scope consists of the basic version plus an add-on memory unit it is perfectly feasible to build the basic version of the scope and to extend it at a later date. This allows the constructor to investigate the possibilities of the TV scope at a modest cost before deciding whether to opt for the more expensive version.

Calibration graticule

A conventional oscilloscope has a calibrated graticule that allows the amplitude of signals to be estimated. This may be engraved onto a perspex mask, printed onto the inside face of the CRT, or in very sophisticated scopes, generated electronically. The first method is not suitable for the

350 mV

0 V
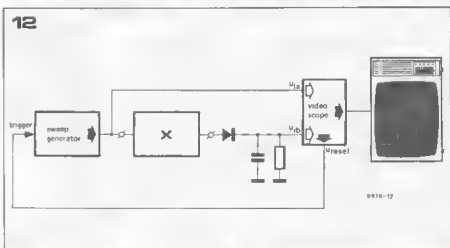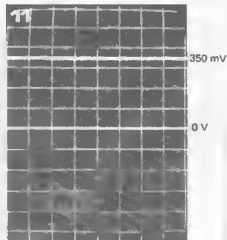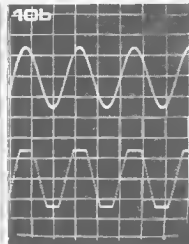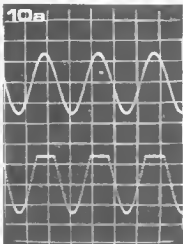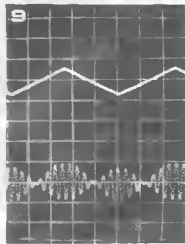
TV scope if the TV set is also to be used for its intended purpose, the second method is obviously impossible, so the TV scope is equipped with an electronically generated graticule. This is nothing more than a grid of horizontal and vertical lines generated in precisely the same way as the crosshatch pattern produced by a test pattern generator. Both the basic and deluxe versions of the TV scope are equipped with this facility.

Comparison of the basic and extended versions of the TV scope

An idea of the comparative performance of both versions of the TV scope can be gained from the photographs in the following figures. The respective frequency ranges of the two versions are shown in figure 7. It is apparent that the basic TV scope can produce a satisfactory image of a 200 Hz signal, but that signals of 10 kHz or higher merely give rise to an interference pattern. Nevertheless the basic scope can still display the amplitude of the 10 kHz signal (and even higher frequencies). The de luxe version, on the other hand, displays the 10 kHz signal quite clearly.
The photographs in figure 8 show how the two versions cope with squarewave input signals. The obvious point here is that the rise and fall times of the squarewaves are much shorter in the case of the basic version. The superior

Specifications		
	basic version	extended version
input impedance	1 M // 30 pF, AC/DC	1 M // 30 pF, AC/DC
Y-channel		
bandwidth	100 kHz (−3 dB)	100 kHz (−3 dB)
highest input frequency at which trace is still recognisable	> 1 kHz	100 kHz
sensitivity/div.	10 mV/100 mV/1 V/10 V	10 mV/100 mV/1 V/10 V
timebase/div.	2 ms (± 1%)	40 μs/100 μs/250 μs/500 μs/1 ms/2 ms above this continuously variable
trigger	none	variable level AC/DC
graticule	generated electronically	generated electronically
output	video, VHF/UHF	video, VHF/UHF

performance of the basic version when processing squarewaves is due to the fact that, in the extended version of the TV scope, the signal is subject to extra filtering after it has been read out of the memory. It should be noted however, that, in contrast to a conventional oscilloscope, the rise and fall times of squarewave signals vary with the timebase setting: the shorter the timebase the shorter the rise and fall times. Thus one should not draw any hard and fast conclusions as to the rise time of the extended version of the scope on the sole basis of figure 8.

The photo of figure 9 shows one application of the extended version of the scope which, because of its lack of a trigger facility, is not possible with the basic version. In this photograph the scope is being used to display a tone burst, such as might be derived from, e.g. the Formant music synthesiser.

From the above comparison of the two versions of the TV scope, one might be inclined to think that the basic version of the scope comes an extremely poor second. Although there is no doubt that the extended version offers considerably more possibilities and can be used in a much wider range of applications, it nevertheless would be wrong to regard the basic version as not worth the bother of building, particularly in view of the fact that it can later be extended to provide the 'de luxe' version.

For all its obvious limitations, there are however a number of applications for which the basic version of the TV scope will prove quite adequate. For example, setting the quiescent output current of an amplifier so that it clips symmetrically (see figure 10) or measuring DC voltages (figure 11).

A further possible application for the basic TV scope is illustrated in figure 12. Together with a sweep generator, the basic scope can be used to directly measure the frequency response of a particular component or circuit. The sweep generator provides a test signal consisting of a sinewave with either linearly or logarithmically increasing frequency. The signal is periodic, returning to a low initial frequency, before once more sweeping up the frequency range over which the response of the circuit in question is to be measured. Adding a simple rectifier circuit produces a DC voltage level which is proportional to the portion of signal passed by the circuit under test. If this is then fed to (the basic version of) the TV scope one can obtain a direct representation of the resultant frequency response.

A more detailed comparison of the two versions of the TV scope is contained in the specifications listed in table 1. Further details on the TV scopes are contained in the respective constructional articles.

Literature:
Elektor 37, May 1978. p 5-20 Elektor, 'TV scope using bucket brigade memory'.

missing link

Modifications to
Additions to
Improvements on
Corrections in
Circuits published in Elektor

master tone generator

Elektor 41, p. 9-09. The component layout shown in figure 2 contains an interesting error: the indications for notes 1 and 12 are transposed for all octaves except octave 4! Furthermore, subsequent testing has shown that the stability of the clock generator can be marginally improved by *omitting* R5 and C5. Since the inputs to the unused inverters in IC16 ought really be connected to positive supply or supply common (as described in the article buffered/unbuffered CMOS . . .), it has been decided to modify the board accordingly before supplying it through the EPS service. This has, of course, led to some delay in the supply.

electronic piano

Elektor 41, p. 9-12. There are a few minor inconsistencies in the article as published, and these have led to a few queries.
— in figures 9, 10 and 11, the position indications 'octave 1' . . . 'octave 5' are transposed: '5' should be '1', '4' should be '2', etc. Note that the indications on the filter board as supplied through the EPS service have been corrected.
— the same error appears in the total wiring diagram shown in figure 17. The output from octave board 5 is shown connected to what is actually the octave 1 input on the filter board, and so on.
— a few of the component values shown in the circuit of the filter (figure 10) differ from those shown in the parts list. The correct values are as follows: C24 = 22 n; R72 = 12 k; R82, listed in the parts list, does not exist . . . S1 . . . S3 are shown as a three-deck four-position switch in the circuit, but in the parts list they are given as three separate single-pole on-off switches. Both options are possible, according to personal taste.
— in the power supply (figure 12) and the corresponding parts list, D7 is shown as a 12 V/400 mW zener diode. This power rating is just on the borderline, and to play it safe it is advisable to use a 1 W type.
— also in the interest of playing it safe, it is advisable to include a 27 Ω/1 W resistor and a 400 mA fast-blow fuse in series with the negative supply rail to the keyboard contacts. In figure 8, this is the connection (in the lower right-hand corner of the diagram) between $-U_2$ and the normally-closed contacts of the keyboard switches.

coming soon

- ASCII keyboard

- Tag

- Ring the bell and win a prize

- Cackling egg-timer

- Digiscope

- Mastermind on the SC/MP

AND

the first ESS recording, containing the following programmes:

— reaction timer
— SC/MP as clock
— Mastermind
— Kojak siren
— RAM diagnostic

proximity detector

There are many methods of detecting the presence of a person within a specified area, for example by using ultrasonic or microwave Doppler techniques, which are the methods frequently employed in intruder alarms. The approach adopted in this article is based on the fact that a person moving about in a room alters the geometry and strength of the electric field that invariably exists. The circuit detects changes in the electric field and produces an audible warning.

Natural and artifical electric fields exist practically everywhere. Their geometry and strength is influenced by the presence of objects, particularly conductors, that are in the field, but in a static situation, i.e. with no moving objects, field patterns will change only slowly, over a period of some hours. If a large conducting object such as a human body moves through an electric field then it will distort the field pattern. Due to the electric charges generated on clothing by friction these variations in the electric field can be very large. In a carpeted room, particularly if the carpets are of man-made fibre, the changes can be even more pronounced.

An electric field can be monitored by a sensor electrode connected to the input of a high impedance amplifier. The electrode will acquire a potential which is dependent on the field strength at the point where the electrode is mounted. Changes in field strength can also be detected very easily by using an analogue voltage comparator.

If the output of the sensor electrode amplifier is connected to one input of a comparator then the voltage at that input will consist of that due to the normal electric field with a changing voltage due to any variations in the field superimposed upon it. If the same signal is connected to the second input of the comparator via a lowpass filter with a very low cutoff frequency (~ 0.2 Hz) then the signal appearing at this input will consist only of the voltage due to the static component of the electric field. Whilst it can follow slow changes due to natural variations in the field over a period of time it will be unable to

follow voltage changes due to objects moving in the field. The voltage on the second input of the comparator thus provides a reference against which to measure changes in the field. Normally the voltage on both inputs of the comparator will be the same, but if the field changes then the voltage on the first input of the comparator will vary and the comparator output will change state.

Two problems must be solved before a practical field variation proximity detector can be built. The first is caused by the 50 Hz AC field which is invariably present in any building where there is mains wiring, and which the sensor would see as a rapid change in field strength. This problem can be overcome by using a second lowpass filter to remove the 50 Hz component from the signal picked up by the sensor plate. The cutoff frequency of the filter (1.8 Hz) is chosen so that the 50 Hz component is completely suppressed, but is still sufficiently high to pass the somewhat slower changes in voltage caused by movements in the field.

The second problem is that, since the amplifier connected to the sensor plate has a high input impedance (which it must have to detect electric fields) the voltage on the sensor plate cannot discharge. The sensor plate will therefore simply charge up to the highest voltage that it sees and any drop in voltage will not register. This problem is solved by periodically discharging the sensor plate through an (electronic) switch. To avoid possible spurious signals caused by beating between the 50 Hz AC voltage and the signal that controls the dis-

Figure 1. Block diagram of the proximity detector.

1

9574 - 1

Figure 2. Complete circuit of the proximity detector.

Figure 3. Printed circuit board and component layout for the proximity detector (EPS 9974).
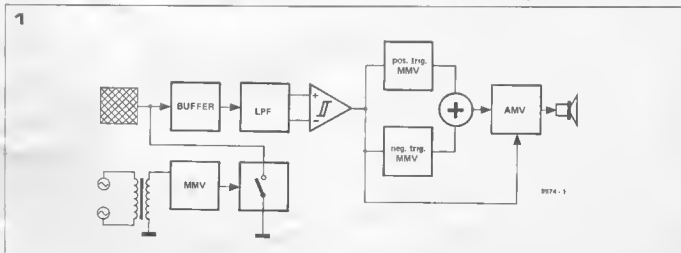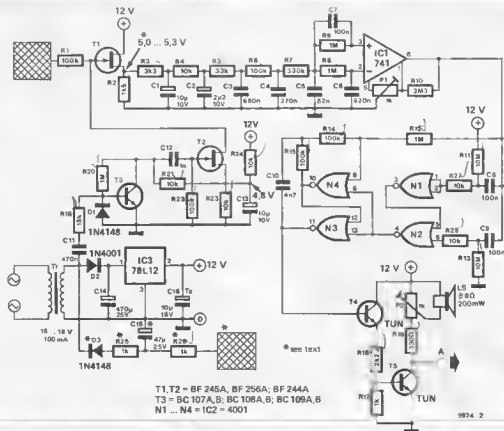
charge switch, it is essential that the plate should be discharged in synchronism with the mains frequency. This is achieved simply by having a 50 Hz mains signal control the switch.

Block diagram

Figure 1 shows a block diagram of the proximity switch. The sensor plate is connected to the input of a high impedance buffer amplifier. This is followed by a lowpass filter, which consists of two sections. The first is the 50 Hz filter; the output of this section connects to the first input of the comparator, i.e. the 'signal' input. A second filter section with a much lower cutoff frequency precedes the second input to the comparator, the 'reference' input. The signal arriving at the signal input of the comparator will thus consist of the total voltage picked up by the sensor plate, i.e. the static reference plus any variations caused by objects moving in the field, whilst only the (practically) unchanging reference voltage will get through the second filter section to the reference input of the comparator.

At the output of the comparator are connected two monostable multivibrators, one of which is positive-triggered and the other negative-triggered, so that either positive- or negative-going transitions of the comparator output can be detected. The outputs of the two monostables are used to control an astable multivibrator, which drives a loudspeaker to give an audible warning. By using the output of the comparator to vary the frequency of

the astable a two-tone signal is provided, the frequency depending upon whether the comparator output is high or low.

Complete circuit

The complete circuit of the proximity detector is given in figure 2. The sensor plate is connected to the gate of T1, which is a FET connected as a source follower. This stage has an extremely high input impedance and a low output impedance. The gain is slightly less than unity. Resistors R3 to R7 and their associated capacitors form the lowpass filter which removes 50 Hz signals. The output of this filter is connected to the non-inverting input of the comparator, a 741 op-amp, via R9 and C7. A lowpass filter section with a very long time constant (R8·C6, approximately 800 ms) removes all but very slow variations from the voltage applied to the inverting input of IC1.

To obtain clean switching of the comparator output a small degree of hysteresis is introduced by applying positive feedback to one of the offset inputs via R10. Negative-going transitions of the comparator output cause the input of N1 to be pulled low via C8. The output of N1 therefore goes high and the output of N2 goes low. Positive-going transitions of the comparator output take the input of N2 high via C9, so that in this case also the output of N2 goes low. The length of time for which the output of N2 remains low depends on the time constant C8·R11 (or C9·R13). N3 and N4 are connected as an astable multivibrator, which drives a small audio amplifier consisting of T4 and T5. When

3

the output of N2 is low the multivibrator will oscillate. An input to the multivibrator from the comparator, via R12, alters the multivibrator frequency depending on whether the comparator output is high or low. The sensor plate is discharged every 20 ms by FET T2. Transistor T3 turns off at each negative-going zero-crossing of the mains waveform, at which point T2 conducts briefly and discharges the sensor electrode.

Power supply

Power for the circuit is obtained from a mains transformer with a 15 V or 18 V secondary rated at 100 mA or greater. The output voltage of the transformer is half-wave rectified by D2 and smoothed by C14 before being fed to a 12 V IC regulator. The mains transformer also provides the 50 Hz signal to switch T3 and T2.

For optimum sensitivity the 0 V rail of the circuit must be connected to an earth point such as mains earth or a metal water pipe. If no such earth is available then an 'artificial earth' must be used consisting of a second electrode connected to a negative supply voltage as shown in figure 2. However, if a true earth is used then R25, R26, C16 and D3 can be omitted.

Construction and use

A printed circuit board and component layout for the proximity detector are given in figure 3. All the components, with the exception of the loudspeaker and mains transformer, are mounted on this board. The electrode(s) may be

made from copper laminate board approximately 15 cm square. If two electrodes are used, they should be mounted about 1 metre apart. The sensor plate must be well-insulated from surrounding objects. Probably the best method is to mount it on the outside of the box in which the circuit is housed using nylon spacers. The unit should function immediately when switched on, and the only adjustments required are to vary P1 for the best sensitivity and to set the volume of the audible warning using P2.

Although intended mainly as demonstration of the principle, the circuit can also be used for practical applications such as intruder alarms, provided its limitations are known. The proximity detector is much less prone to false alarms than ultrasonic or microwave Doppler alarms, which can be triggered by flapping curtains or rattling doors and windows. However, the circuit may be falsely triggered by changes in field strength caused by switching on and off of electrical equipment. This is not such a problem if the unit is intended to protect unoccupied premises, provided care is taken not to mount it in the vicinity of equipment that switches on and off automatically, such as a refrigerator or freezer.

To trigger an external alarm or other circuit the signal from point (A) may be used. This is normally high, but when the audible warning sounds point (A) goes alternately high and low at the same frequency. ∎

resonance filter module

In addition to an almost limitless variety of non-natural, wholly 'electronic' sounds, the Formant music synthesiser can, of course, be used to imitate the voicing of conventional (mechanical) musical instruments. The filter module described in this article is designed to allow more realistic simulation of natural musical instruments by providing the fixed bandpass resonances which are an important determining factor in the timbre of machanical tone generators.

Figure 1. The fundamental frequency range of a number of traditional musical instruments, with reference to that of a grand piano. *(From: 'Elektronik Taschenbuch, Band 1', Ferd. Dümmlers Verlag, Bonn, with kind permission from the publishers.)*

Although music synthesisers are capable of producing the most 'wierd and wonderful' electronic effects, it is a fact that they are frequently employed to imitate the sound of traditional acoustic instruments. Many commercially available synthesisers, for example, are provided with preset facilities for various common instrumental voices, whilst special units such as 'string-synthesisers', which are designed solely to reproduce the sound of a string section, are becoming increasingly popular.

The basic factors influencing the characteristics of a musical note are pitch, dynamic amplitude, and dynamic harmonic content. As the reader will be aware, pitch and dynamic amplitude characteristics are controlled by the VCO (Voltage Controlled Oscillator) and VCA + ADSR (Voltage Controlled Amplifier and Attack - Decay - Sustain - Release) modules in the Formant synthesiser, whilst the VCF (Voltage Controlled Filter) is used to vary the harmonic content of the signal.

However, in the case of mechanical tone generators, for example brass and woodwind instruments, an additional consideration is the existence of resonant areas in the instrument which possess free vibration periods of their own. These resonances, which are known as *formants* (whence the name for the Elektor music synthesiser!) are determined by the shape and mechanical construction of the particular instrument (the wooden back and belly of a violin, the pipes of an organ etc.). Unlike the variable pitch of, say, a violin string, they tend to reinforce the same harmonics, whatever the pitch of the note being played. The nature of the formants in an instrument is in fact one of the factors which govern its quality.

It should thus be apparent that, in order realistically to simulate the tonal characteristics of traditional instruments, one must be able to tailor the static harmonic content of the note accordingly. What is required is a number of resonant filters with independently variable centre frequency, gain and Q-factor. These features are present already in the state variable VCF of the Formant; however, that is only one filter, and more to the point, in this particular application there is no need for the filters to be voltage-controlled, since the filter parameters will be preset to suit whatever musical instrument is being imitated. This explains the reason for the separate manually-controlled resonance filter module described in this article.

The uses of resonance filters

The effect of resonance filters can best be heard on 'bright' sharp VCO waveforms which have a high proportion of fairly intense upper harmonics. The effect on vocal sounds can be illustrated by taking a suitable signal with a frequency of around 200 Hz, setting the Q of the filter to a mid-value, and varying the centre frequency from minimum to maximum. At first 'dark' sounding tones, largely devoid of higher harmonics will be obtained; as the centre frequency is increased however, one by one the various vowel sounds can be distinguished until, at high

centre frequencies, reedy flute-like sounds are produced. The higher the Q of the filter, the more pronounced the above effects – and vice versa

All bandpass resonances of musical importance lie between roughly 100 and 2000 Hz. Table 1 lists the main fixed resonances of a number of common musical instruments and also indicates which VCO waveform is best suited to imitate the instrument in question. This table is, of course, merely intended as a rough guide, in the final instance the decision should rest with one's own ears. Unless otherwise indicated, the Q-control should be set to the mid-position. As a further aid, figure 1 shows the fundamental frequency ranges of various traditional instruments, with reference to a piano keyboard.

Circuit

The basic requirements of the filter circuit are, independently variable centre frequency, Q and gain. Since the function of the filter is essentially to enhance a particular band of frequencies (corresponding to the formants of the instrument in question), the circuit is of the boost-only type, i.e. provides selective gain. Without the need to provide a selective cut (below the 0 dB line) the circuit design is considerably simplified. A total of three resonant filters forms

an acceptable compromise between the number of settings required for reasonably realistic imitation and the constraints of space and economy. Of course, it is quite possible to double the range of control facilities by connecting a second filter module in cascade with the first.

Block diagram

The block diagram of the resonant filter module is shown in figure 2. The figures in brackets indicate which components in the final circuit are associated with the different sections of the circuit. Signals can be fed in via the panel-mounted socket (ES) or via the hard-wired input (1S). A portion of the signal is fed direct to the output summing amplifier via R (R5) in the complete circuit) and the input signal is also fed to three bandpass filters whose gain, centre-frequency and Q can be varied. The outputs of these filters are also summed in IC5 via resistors R_0. The output of the filter module will thus consist of a portion of the input signal plus signals boosted around the centre frequencies of the three filter stages. Two outputs are provided from the filter module, an internal hardwired output (IOS) and an output to a front panel socket (EOS). A bypass switch is provided, which allows the three filter

sections to be switched out, in which case only the original signal appears at the output, and the gain is frequency independent, being unity.

The amount of boost that can be provided by a filter section relative to the gain obtained in the 'bypass' condition is determined by the gain of the filter sections and the ratio R/R_0. If it is assumed that the filter gain can be varied between zero and one then the maximum amount of boost (in dB) is $20 \log \left(1 + \frac{R}{R_0}\right)$.

The frequency response of a filter section is shown in figure 3. The figures in parentheses indicate which controls in the complete circuit vary the different parameters of the filter.

The complete circuit of the filter module is shown in figure 4. IC1 sums and inverts the two input signals, whilst the three filter sections are of the (to Elektor readers by now familiar) state-variable type. The resonant gain of the filters is set by means of P1, P4 and P7 respectively. One gang of the pots is connected at the input, the other at the output of filter This has the effect of improving the dynamic range, since it means reduced noise and less chance of overloading. Finally, there is the inverting summing amplifier round IC5, which also cancels the phase shift introduced by IC1.

1

percussion			kettle drum				
plucked string instruments			bass guitar	guitar			
woodwind instruments			bassoon	bass saxophone	clarinet	flute	soprano saxophone	piccolo
brass instruments			tuba	trombone	horn	trumpet		
bowed string instruments			double bass	violin cello		violin		
vocal			bass	tenor	alto	soprano		

| A_2 | C_1 | A_1 | A | c | c' d' e' f' g' a' b' c² d² | a² | c³ | c⁴ | c⁵ |
| 27.5 Hz | | 55.0 Hz | 110.0 Hz | 220.0 Hz | 440.0 Hz | 880.0 Hz | 1760 Hz | 3520 Hz | |

octave

9961-1

2

3

Table.	instrument	main resonance at	VCO signal
	flute	approx 800 Hz	fairly asym, squarewave
	clarinet	1 - 2 kHz*	sym, squarewave
	oboe	1300 - 1700 Hz*	heavily asym, squarewave (pulse)
	bassoon	approx 440 Hz!	heavily asym, squarewave (pulse)
	trumpet	approx 1500 Hz	'spaced' sawtooth
	bugle	approx 1000 Hz*	sawtooth
	trombone	approx 600 Hz	'spaced' sawtooth
	French horn	approx 400 Hz*	sawtooth
	tuba	approx 250 Hz	sawtooth
	violin	approx 4000 Hz*!	'spaced' sawtooth
	cello	approx 200 Hz**	sawtooth or heavily asym
	double bass	approx 100 Hz**	squarewave (pulse)

NB.
* with increased Q
** if possible, use several resonant filters for a comb filter

Figure 2. Block diagram of the resonant filter module. As can be seen, it possesses three independently variable filter sections.

Figure 3. The frequency response of one of the three filter sections contained in the resonant filter module. The figure illustrates how the filter parameters can be independently varied by means of the control potentiometers.

Figure 4. Detailed circuit diagram of the filter module.

Figure 5. Track pattern and component layout of the filter module p,c,b, (EPS 9951-1).

4

IC1 = LF 356 MINIDIP
A4 = IC2 = TL 074, TL 084
A5 ... A8 = IC3 = TL 074, TL 084
A9 ... A12 = IC4 = TL 074, TL 084
IC5 = LF 357 MINIDIP
C10 ... C16 = 100 n
P1a,P1b,P2,P4a,P4b,
P5,P7 = 10 k log
47 k 150 k) log
P3a,P3b,P6a,P6b,P8a,P8b =
10 k log

5

6

With the values for R and R_0 given in
the circuit diagram, the maximum gain
of the filter is approx. +15 dB. The
quality factor, Q, can be varied by P2
(P5, P8) between roughly 0.8 and 5.
The centre frequency can be varied
between approx. 50 and 2300 Hz,
which is more than sufficient for normal
use. The frequency range can, however,
be modified by altering the value of a
number of components; the necessary
changes are detailed in the appendix.
Maximum Q is obtained for the mini-
mum resistance of the Q-potentiometer.
The maximum Q can therefore be
increased by reducing the value of R7
(R22, R37); in this way a Q of between
20 and 30 can easily be obtained. A
high Q is useful when processing wave-
forms such as squarewaves, which have
very steep edges. These tend to set the
filters 'ringing' at their resonant fre-
quencies, and produce percussive ef-
fects. For R7 (R22, R37) = 470 Ω,
a Q of 11.3 is obtained; R7 = 330 Ω
gives a Q of 15.8, and R7 = 220 Ω a Q
of 23.4. The higher the Q, the more
pronounced the percussive effect.

Figure 6. Because of the large number of controls, the front-panel for the resonant filter module is clearly different from the other Formant front-panels (EPS 9951-2).

Figure 7. Wiring diagram for the components mounted on the front-panel.
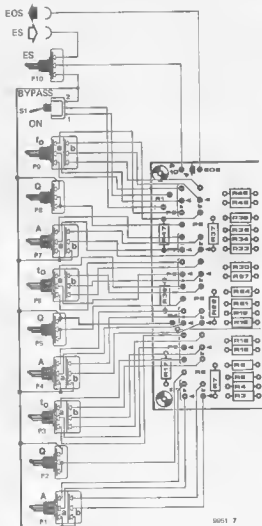
Construction

The printed circuit board for the resonant filter module is shown in figure 5.

As far as the selection of components is concerned, the criteria which applied in the case of the Formant are also applied here. The only difference is that in view of the large number of front-panel controls (10 potentiometers) it is strongly recommended that miniature components (miniature pots with 4 mm diameter spindles) be used. In this way the controls can be arranged in functional groups of three to a row.

The front panel for the filter module is shown in figure 6, and the details of the wiring for the front-panel controls are illustrated in figure 7. In contrast to the other Formant modules, the resonant filter module requires no calibration or adjustment procedure. The operation of the circuit can be checked by feeding in a white noise input from the noise module. Varying the three filter parameters should produce clearly audible changes in the resulting sound. It will

also be apparent that rapid variation of the Q- and f_O controls produces effects similar to phasing, thus the filter module can be used to provide manual phasing.

The scale on each of the f_O potentiometers on the frontpanel is calibrated with five nominal frequencies. The three middle settings in particular should be viewed as rough guidelines, since the resistance curve of logarithmic potentiometers can exhibit fairly wide tolerances.

The filter module should be placed between the COM-module and the power amp. However, if one wishes to use the headphone output on the COM-module, the resonant filter module can be connected directly before the latter.

Appendix

With the component values given in the circuit diagram, the centre frequency of the filters can be varied between roughly 50 and 2300 Hz. To calculate the correct values for higher frequencies, this the procedure is as follows:

Firstly, the desired maximum frequency of f_O can be used to calculate the value of C2 = C3 = C4 = C5 = C6 = C7 = C from the following equation:

$$C = \frac{16}{f_O \text{ max}}$$

where C is in nanofarads and f_O in kHz. Secondly the value of resistor R (see figure 2) can be determined on the basis of the desired minimum centre frequency f_O min:

$$R = \frac{16}{C \cdot f_O \text{ min}}$$

where C is in nanofarads, R is in kΩ, and f_O in kHz

The value of R_O = R11 = R14 = R26 = R29 = R41 = R44 can be calculated from:

$$R_O = \frac{10}{R - 2}$$

where R and R_O are in kΩ. These equations can be used to check the values of figure 4.   ∎

databus buffer
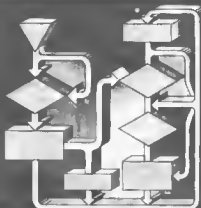
Users of the Elektor SC/MP system may eventually wish to expand the system memory. If any significant extension of the memory is contemplated then it will be necessary to buffer the data bus, which can be accomplished with the simple data bus buffer described in this article.

The SC/MP microprocessor is capable of addressing up to 65 k of memory. The memory capacity of the Elektor SC/MP system can be expanded by the addition of extra 4 k RAM cards (see Elektor 37, March 1978) or by adding other types of memory (e.g. ROM or PROM). The address bus of the Elektor SC/MP system is already buffered by tristate buffers on the CPU card and can easily drive this amount of memory. The data bus however, is unbuffered, i.e. the data lines of the SC/MP are connected direct to the data bus, and the limited drive capability of the SC/MP means that the data bus will be unable to handle large additional sections of memory.

A large portion of page '0' of the memory is accomodated on the CPU card and is therefore connected direct to the SC/MP data lines. It is obviously not possible (or necessary) to buffer this section of memory. However, the load on the SC/MP can be kept well within its drive capability by leaving page '0' of the memory unbuffered and buffering the remainder of the data bus. This is done by connecting a data bus buffer between the first busboard and the second busboard. The CPU card, memory extension card and HEX I/O card are then plugged into the first busboard whilst the additional memory cards are plugged into the second and subsequent busboards. This is shown in the block diagram of figure 1.

Since the data bus is bi-directional, i.e. the SC/MP can write information onto the data bus or read information from it, the data bus buffer must also be bi-directional. In other words, when the SC/MP is sending data along the bus then the buffer must present a high impedance to the SC/MP and a low impedance to the bus. Conversely, when the SC/MP is reading data from the bus then the buffer must present a high input impedance to the bus and a low output impedance to the SC/MP. If, however, the SC/MP is reading data from page '0' of memory then the output of the buffer must assume a high impedance, otherwise the buffer would load the memory output.

The circuit of the bi-directional data bus buffer is given in figure 2. It utilises two, dual-four-bit tri-state buffers, each of which is connected in 'reverse-parallel' to form a bi-directional buffer. The NWDS and NRDS lines are used to control the buffers so that when SC/MP is writing onto the data bus the buffer is active in one direction and when SC/MP is reading from the bus the buffer is active in the opposite direction. When the SC/MP is addressing page '0' of the memory the chip enable output for the address decoder ($\overline{CE}$) is low and this is used to inhibit both IC1 and IC2 via N1 and N4 so that their outputs are in the high impedance state.

1

2
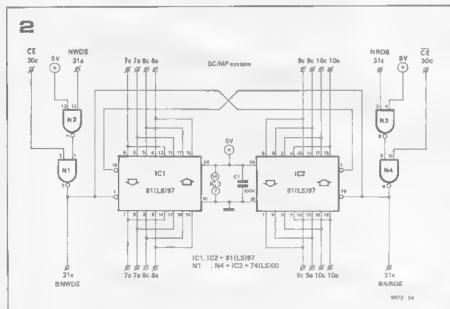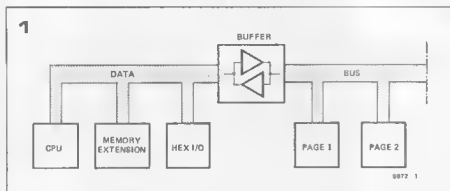
Parts list for figures 2 and 3

Capacitors:
C1 = 100 n

Semiconductors:
IC1,IC2 = 81 (LS) 97
          (National Semiconductor)
IC3 = 74 (LS) 00
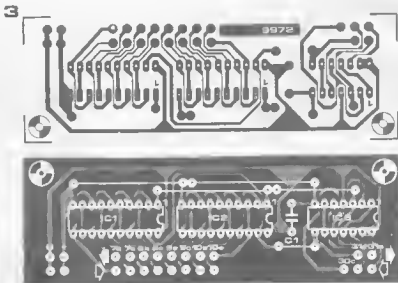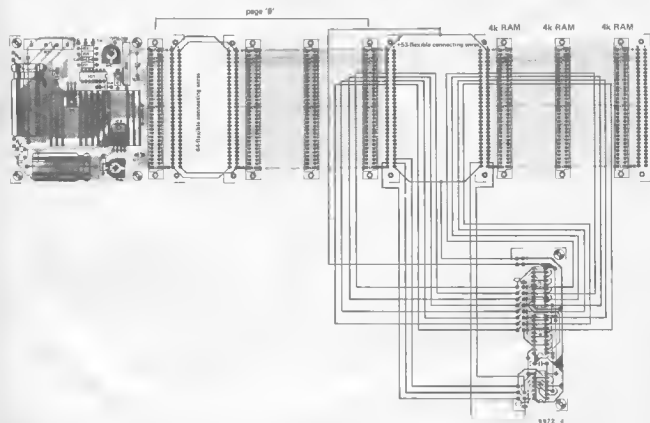
Printed circuit board

A printed circuit board and component layout for the data bus buffer are shown in figure 3. This board is designed to be mounted between two busboards as shown in figure 4. The buffer board can be mounted at right angles to the plane of the busboards and joined to them by short wire links although, for clarity, it is shown detached in figure 4. All those connections on the first busboard not linked to the buffer board should be connected direct to the corresponding points on the second busboard.

Figure 1. Block diagram showing how the buffer is connected into the bus system.

Figure 2. Circuit of the data bus buffer.

Figure 3. Printed circuit board and component layout for the data bus buffer (EPS 9972).

Figure 4. Wiring diagram showing how to interconnect the buffer board with the busboards. For clarity the buffer board is shown detached from the busboards but in practice it is mounted at the intersection of the busboards, at right angles to them and is connected by short wire links.

Once the buffer has been installed all cards containing page '0' memory addresses (CPU card, memory extension card and HEX I/O card) must be plugged into the busboard on the left of the buffer, whilst cards containing other memory pages may be connected to the busboard(s) on the right of the buffer.
Finally, it should be remembered that each additional 4 k RAM card draws an extra 1 A from the power supply, so it will be necessary to uprate the supply accordingly if the memory is expanded.

central alarm system

The Central Alarm System (CAS) will relay an alarm indication from a number of remote stations to a central location along a common bus system. Audible indication of the alarm is provided together with a visual display of which station has sent the alarm. The applications for this system are limited only by the ingenuity of the reader.

Elektor 19, December 1976 featured a Domestic Early Warning (DEW line) system, which could also relay alarm signals from remote locations to a central post. Alarm sensors such as telephone bell repeater, freezer alarm, water level alarm etc. could be linked up to the DEW line. However, the big disadvantage of the DEW line was that it gave no clear indication of which sensor was sending the alarm, a deficiency which is remedied in the Central Alarm System.

A block diagram of the CAS is given in figure 1. It will be seen that each post is simply connected to a three-wire bus. There is no need to connect each station to the central station by separate wires, which greatly simplifies wiring of the system and can also save cost if a long cable run is required to a number of remote stations. Additional stations can also be added at a later date simply by linking them into the existing bus system.

It is obviously a simple matter to send an alarm indication along three wires, as was done in the DEW line, but how, using only three wires is it possible to indicate which station has sent the alarm?

The answer is to use time division multiplexing of the signals. As shown in the more detailed block diagram (figure 2), each station contains a decade counter which is made to count in synchronism with a counter in the master station. By this means each station is allocated a time slot during which it can send back an alarm to the central station. The first of the three wires of the system bus is simply a ground, the second wire is used as the alarm line to send signals back to the central station and the third wire is used to transmit sync and clock pulses to remote stations. An ingenious feature is that this third wire is also used to transmit power to the remote stations by allowing the sync and clock pulses to charge up a capacitor in the remote station. The remote station then operates from this stored voltage.

In addition to the alarm indication at the master station provision is also made for slave indicator stations which may be mounted anywhere along the system bus. These are similar to the master station except that their synchronisation and power are provided along the system bus in the same manner as the alarm stations.

Sync and clock waveform

The synchronising and clock waveform is shown in figure 3. The clock waveform is derived from the mains frequency, 50 Hz or 60 Hz depending on country. At the beginning of each clock pulse train is a 300 ms synchronising pulse which is used to reset all the counters in the system to zero, ensuring that every counter stays in synchronism with the master station counter. This is followed by a sequence of 9 clock pulses, which step every counter through 1 to 9. Each alarm station is allocated a number, and the corresponding output of its counter is connected via the alarm sensor circuit to the alarm line. Thus, for example, at station 5, output 5 of the counter will be connected to the alarm line and station 5 can therefore send an alarm signal only when output 5 of the counter is high. In practice a saving may be made by connecting two alarm sensors to one station, as will be seen later.

Alarm station

To understand the operation of the complete circuit it is perhaps best to begin with a description of the alarm station, whose circuit is shown in figure 4. Sync and clock pulses are picked off the bus at point (S). The positive-going pulses charge up C13 via D25, the voltage across C13 being used to power the station. As CMOS devices are used the power consumption of the station is small and a value of 68 μ for C13 is adequate in most cases. However, power for the alarm sensor may also need to be derived from this rail, and if this additional loading causes excessive supply ripple then the value of C13 must be increased.
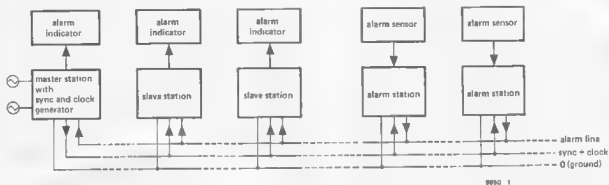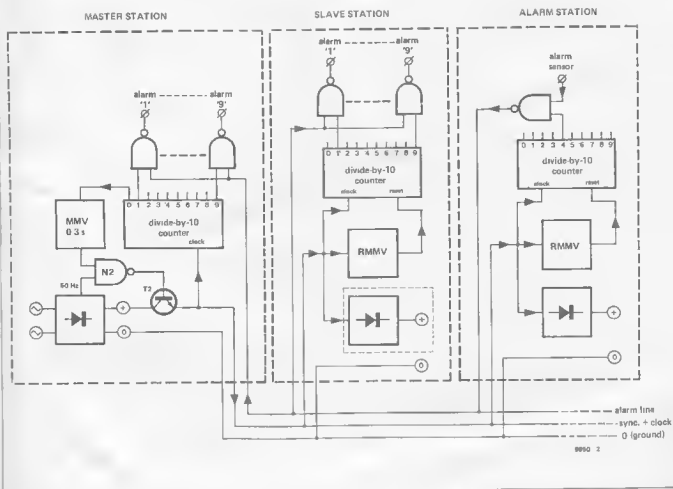
Figure 1. Block diagram of the CAS, showing the three different types of station connected to the system bus.

Figure 2. Detailed block diagram showing the master station, one slave station and one alarm station.

Figure 3. The clock/sync waveform.

4
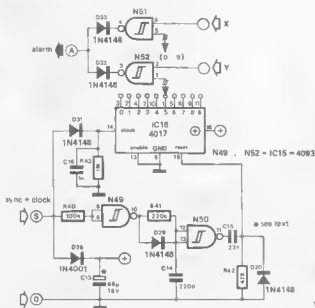
The sync/clock waveform is also fed to the clock input of the counter IC16, and to a sync pulse detector, which is a retriggerable monostable multivibrator comprising Schmitt triggers N49 and N50. When clock pulses are present the output of N49 goes alternately high and low at 50 Hz and C14 is charged via D29, holding the input of N50 high. During this time IC16 counts clock pulses. During the 300 ms sync-pulse the input of N49 is held high, its output is low and C14 discharges via R41. When the lower threshold of N50 is reached the output of N50 goes high and a reset pulse is applied to IC16 via C15.

Two alarm inputs are provided at points X and Y. This allows two alarm sensors to be connected to a single alarm station, which can be useful if two sensors are located fairly close together, as it saves the cost of a station. Each sensor is allocated a number and one input of the appropriate gate is

5
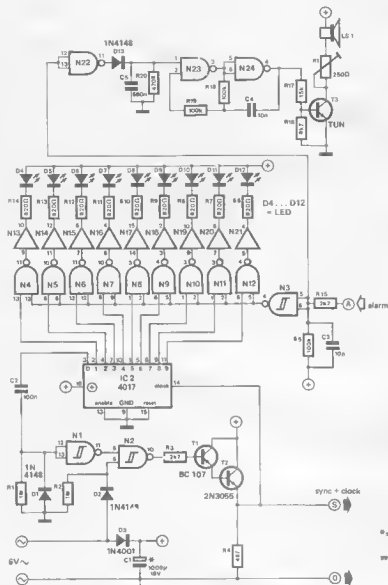
N1,N2,N3,N11 = IC1 = 4093
N4,N5,N8,N9 = IC3 = 4011
N6,N7,N10,N12 = IC4 = 4011
N13 . . N18 = IC5 = 4010
N19 . . N21 = IC6 = 4010
N22 . . N24 = IC7 = 4011

D4 . . . D12 = LED

linked to the corresponding output of IC16. In the example shown the input of N52 is linked to output 5 of IC16. The output of N52 is normally high. If an alarm signal takes point Y high then when output 5 of IC16 also goes high the output of N52 will go low. The result is that a negative-going alarm pulse will be sent back down the alarm line each time the clock pulse sequence reaches count 5.

Master station

The circuit of the master station is given in figure 5. Clock pulses are derived from a 6 V AC supply, which is half-wave rectified by D2 and squared up by Schmitt trigger N2. T1 and T2 buffer the output of N2 to provide a low impedance drive to the bus. The master station counter, IC2, also counts clock pulses from the emitter of T2. Each time IC2 reaches count zero the

Figure 4. Circuit of an alarm station.

Figure 5. Circuit of the master station.

Figure 6. Circuit of a slave station.

'0' output of IC2 goes high, taking the input of N1 high via C2. The output of N1 goes low, taking pin 9 of N2 low and inhibiting the clock pulses.
C2 charges until, after about 300 ms, the voltage at the input of N1 has fallen to its lower threshold, when the output of N1 goes high and clock pulses are again through N2. Each output of IC2 (except '0') is connected to one of the inputs of N4 to N12. The other inputs of these gates are joined and connected to the output of N3, whose input is connected to the alarm line. Normally the alarm line is high, so the output of N3 will be low and the outputs of N4 to N12 will be high.
If there is an alarm from, say, sensor 4 then the alarm line will go low when the clock sequence reaches count 4. The output of N3 will thus go high at this time. Output 4 of IC2 will also be high, so both inputs of N7 will go high and its output will be high. Via buffer N16 LED D7 will therefore be lit.

Figure 6.

N26 . . . N36 = IC8 = 4093
N29, N28, N32, N33 = IC10 = 4011
N30, N31, N34, N36 = IC11 = 4011
N37 . . . N42 = IC12 = 4010
N43 . . . N45 = IC13 = 4010
N46 . . . N48 = IC14 = 4011

* see text

9660 6
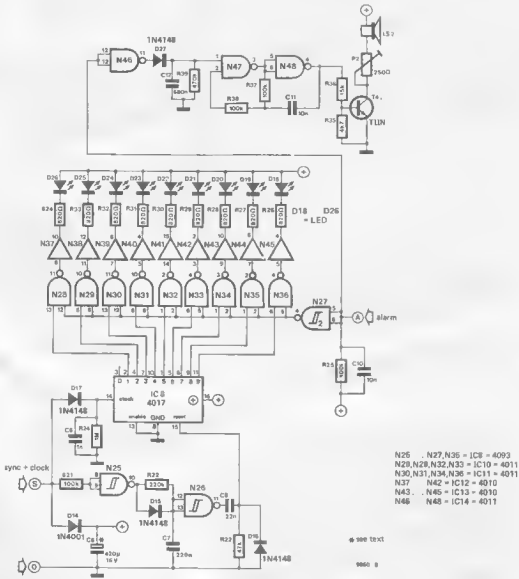
7a
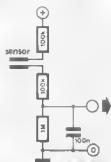
7b

7c
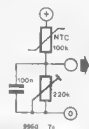
Parts list for figure 8

Resistors:
R40 = 100 k
R41 = 220 k
R42 = 47 k
R43 = 1 M

Capacitors:
C13* = 68 µ/16 V
C14 = 220 n
C15 = 22 n
C16 = 1 n

Semiconductors:
D28 = 1N4001
D29 . . . D33 = 1N4148
IC15 = 4093
IC16 = 4017

7d
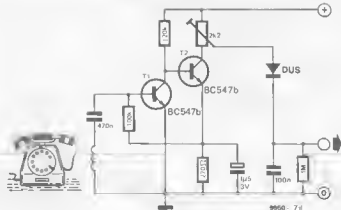
8
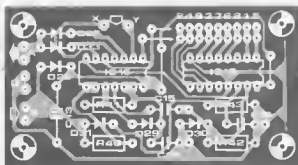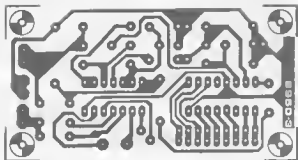
Parts list for figure 9

Resistors:
R1,R2 = 1 M
R3 = 2k7
R4,R16 = 4k7
R5,R18,R19 = 100 k
R6 . . . R14 = 820 Ω
R15 = 2k2
R17 = 15 k
R20 = 470 k
P1 = 220 Ω (250 Ω) preset

Capacitors:
C1 = 1000 µ/16 V
C2 = 100 n
C3,C4 = 10 n
C5 = 680 n

Semiconductors:
T1 = BC107
T2 = 2N3055
T3 = TUN
D1,D2,D13 = 1N4148
D3 = 1N4001
D4 . . . D12 = LED
IC1 = 4093
IC2 = 4017
IC3,IC4,IC7 = 4011
IC5 IC6 = 4010

Miscellaneous:
LS1 = Loudspeaker, 15 Ω
or greater

9

Of course the LED lights for the duration of only one clock pulse during the complete clock sequence, and since the clock sequence repeats about twice a second the LED will flash at a 2 Hz rate.

An audible alarm warning is provided by an astable multivibrator N47/N48, which drives a loudspeaker via T4. Normally pin 1 of N47 is low and the astable is inhibited, but when pulses appear on the alarm line the output of N46 goes alternately high and low, charging up C12 via D27 and activating the audible alarm.

Slave station

Figure 6 shows the circuit of the slave station, which might be described as a cross between the master station and the alarm station. The audible and visual alarm sections of this circuit are identical to those of the master station, but it derives its sync, clock and power from the bus in the same manner as the alarm station, and uses an identical sync pulse detector.

Alarm sensors

Circuits for the alarm sensors may vary

Figure 7. A selection of alarm sensors. a. Water level alarm. b. Voltage failure alarm. c. Over-temperature alarm. d. Telephone alarm.
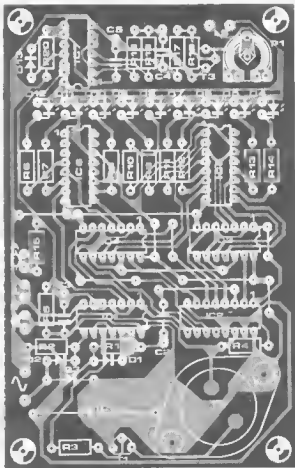
Figure 8. Printed circuit board and component layout for the alarm station. (EPS 9950-3).

Figure 9. Printed circuit board and component layout for the master station. (EPS 9950-1).

from the very simple to the complex. Whatever the circuit, it must take the X or Y input of the alarm station high when an alarm condition is sensed. A selection of alarm sensors are shown in figure 7. Figure 7a is a water level sensor. Normally the input of the alarm station is pulled low by the 1 M resistor, but when the probes are immersed in water, or some other conductive liquid, the input to the alarm station goes high. Figure 7b shows a voltage failure detector using a relay. While the supply is present the relay is pulled in and the contact is closed; should the supply fail the relay will drop out and the contact will open, causing the alarm station input to be pulled up by the 1 M resistor. Figure 7c shows an over-temperature alarm sensor. As the temperature rises the resistance of the NTC thermistor falls until the input voltage to the alarm station rises until it exceeds the upper threshold of Schmitt trigger N51 or N52. The alarm temperature is adjusted by the 220 k potentiometer. An undertemperature alarm sensor can be constructed by transposing the positions of the thermistor and potentiometer. Finally, figure 7d shows a telephone bell alarm sensor. The signal from the telephone pickup coil attached to the base of the 'phone is amplified by T1 and T2 and

Figure 10. Printed circuit board and component layout for the slave station. (EPS 9950-2).
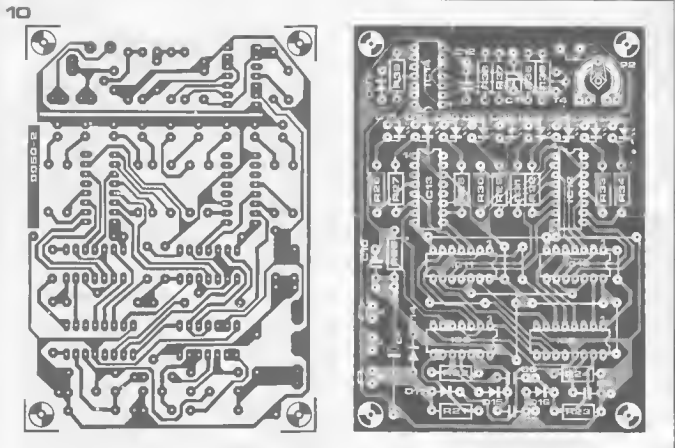
Parts list for figure 10

Resistors:
R21,R25,R37,R38 = 100 k
R22 = 220 k
R23 = 47 k
R24 = 1 M
R26...R34 = 820 Ω
R35 = 4k7
R36 = 15 k
R39 = 470 k
P2 = 220 Ω (250 Ω) preset

Capacitors:
C6* = 470 μ/16 V
C7 = 220 n
C8 = 22 n
C9 = 1 n
C10,C11 = 10 n
C12 = 680 n

Semiconductors:
T4 = TUN
D14 = 1N4001
C15,D16,D17,D27 = 1N4148
D18...D26 = LED
IC8 = 4093
IC9 = 4017
IC10,IC11,IC14 = 4011
IC12,IC13 = 4010

Miscellaneous:
LS2 = Loudspeaker, 15 Ω
   or greater

rectified by the diode to give a DC voltage which takes the alarm input high.

These are just a few examples of the types of alarm sensor that can be used and the possibilities are limited only by the ingenuity of the constructor.

Construction

Printed circuit boards and component layouts for the alarm station, master station and slave station are given in figures 8, 9 and 10 respectively. The master station of course requires a 6 V

transformer to provide power and clock pulses for it and the rest of the system. Since the slave stations consume around 50 mA when indicating an alarm it may be necessary to increase the value of C1 in the master station if more than one slave station is used. It may also be advisable to increase the value of C6 if supply ripple is a problem at the more remote slave stations. The current rating of the transformer should be adequate to supply all the stations included in the system, allowing about 50 mA for the master station and each slave station and a few mA for each alarm station.

If a large number of slave stations are to be used then it is perhaps best to equip each one with its own mains power supply. This simply means disconnecting the anode of D14 from the board and connecting a 6 V AC supply between the anode of D14 and 0 V.

For wiring up the system any type of three-core cable may be used, for example light-duty (3 A) mains flex. Screening of the cable is not necessary in a normal domestic environment, but for a neat appearance twin screened (stereo) audio cable could also be used, in which case the screen should be connected to 0 V. ◀

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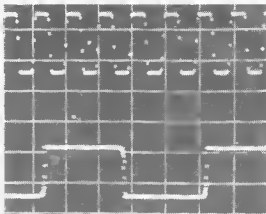
To illustrate the principle of the TV modulator it is useful to look at a typical video waveform and the corresponding modulated r.f. signal, both of which are illustrated in figure 1.
Figure 1a shows one line of a video waveform. The maximum positive excursion of the signal is known as white level,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

Designed principally for use with the TV Scope described elsewhere in this issue, this easy-to-build circuit will modulate a video signal onto an r.f. carrier to give a signal that may be fed direct to the aerial socket of a VHF or UHF television receiver.

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

Modulator circuit

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

Construction and adjustment

A printed circuit board track pattern and component layout are given in figure 4. This board is available from the Elektor Print Service, EPS No. 9967. Two alternative mounting positions are provided for the crystal, allowing for two different pin spacings.
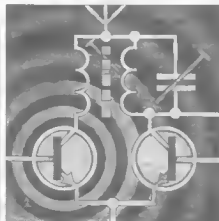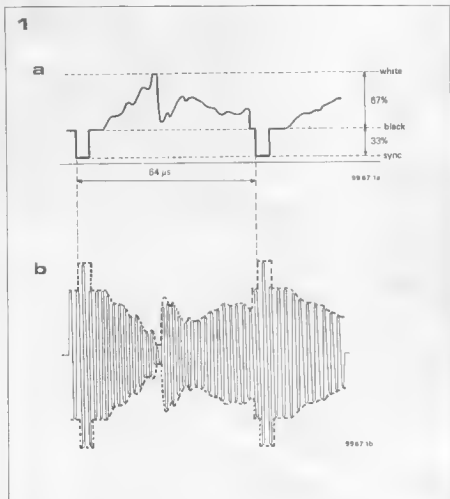
Figure 1. a. One line period of a typical video signal, showing picture information and line sync pulses. b. An r.f. carrier modulated with the signal of 1a, using negative modulation.

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

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

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

T1,T2 = BF 494
T3 = BFY 90
D1 = 1N4148

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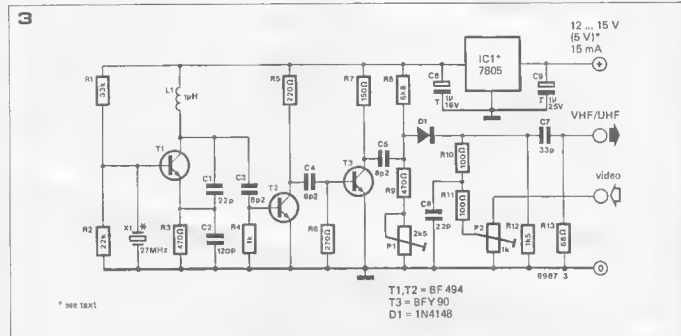
Because of the high frequencies involved the board is designed with a generous earth plane for stability. In addition a screening plate, made of tinplate or a piece of copper laminate is connected between the oscillator and modulator. The completed board *must* be mounted in a metal box for screening,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

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

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

A video signal may now be fed in, and

Parts list to figure 2.

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

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

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

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

2

a
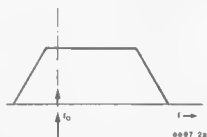
b

4
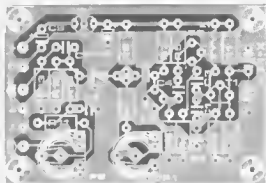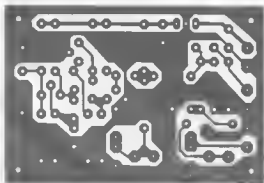
P2 should be adjusted so that the video signal level does not exceed 3 V peak-to-peak at its wiper.
The TV set may now be tuned to the sideband which gives the best picture. If tuned to the wrong sideband the picture will tend to appear negative. If the picture lacks vertical synchronisation (i.e. rolls) it will be necessary to adjust P1 until it stabilises. P2 is used to adjust the contrast by varying the video input level, but should not be turned up too much or the modulator will overload, causing the picture to appear negative on highlights.
Finally it should be noted that, when using the modulator, the output should always be connected direct to the TV set via a length of coaxial cable and must never be connected to any un-screened wire or other conducting object that could act as an aerial, otherwise the user could receive an unwanted visit from the Post Office Radio Interference Officer!

TV scope-basic version
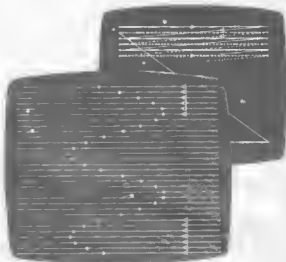
As explained in the accompanying introductory article, there are two different versions of the TV scope. The following article describes the practical circuit and constructional details of the basic version; the extended or 'de luxe' scope will be discussed in a second article to be published next month.

Since the fundamental principles of the TV scope have already been discussed, we can proceed straight to the block diagram of the basic TV scope shown in figure 1. Although the design shown is for a two-channel scope, it can of course be adapted for single-channel operation simply by omitting the Y_B input amplifier (shown in dotted lines).

The input amplifier Y_A allows either continuous or stepwise adjustment of the input sensitivity of the scope. The maximum gain of the input amplifier is x 23, and this corresponds to the maximum sensitivity of 10 mV/div. The output voltage, u_{ya}, of the input stage is fed to a comparator circuit, where it is measured against a sawtooth reference voltage, u_{ref}.

The moment u_{ya} equals the sawtooth voltage, u_{ref}, the comparator triggers a monostable which provides a 'white-level' pulse, u_{pa}.

Exactly the same process occurs in the case of the other input signal, u_{yb}, on the second channel. The white-level pulses from both channels are then summed, and the resulting signal is fed to a video mixer circuit where it is provided with the necessary sync pulses and a signal which generates a graticule on the screen. The circuits for generating both the sawtooth reference voltage and the graticule signals are synchronised by a central timebase.

The timebase signal is derived either from a fixed crystal oscillator or else an oscillator which can be varied over a small frequency range. Finally, the block diagram includes a modulator, which in the vast majority of cases will be required to modulate the video output signal of the TV scope onto an r.f. carrier wave, thus allowing it to be received on a conventional TV set. The modulator is shown in dotted lines however, since it forms the subject of a separate article contained elsewhere in this issue.

Input amplifier

The complete circuit diagram of the input amplifier is shown in figure 2.

As in the other circuit diagrams contained in this article, a number of voltages are shown; those in brackets, however, refer not to the present explanation of the basic TV scope, but rather to the description of the extended version of the scope which will be published next month.

The input signal, u_i, is fed via an AC coupling capacitor, C3, to a switched input attenuator (R1 ... R7). The input capacitor can be switched out of circuit by means of S2, so that the scope can also accept DC input signals. From the attenuator the signal is fed to the input of the amplifier stage based on op-amp A1. The gain of this stage can be continuously adjusted with the aid of potentiometer P1, the fine sensitivity control.

A2 and A3 together form a unity-gain non-inverting amplifier; P2 varies the DC bias level on the non-inverting input of A3, which in turn varies the Y-position of the trace.

A4, which is necessary for DC bias balancing of the analogue shift registers, will be discussed in greater detail in the article on the extended version of the scope.

A printed circuit board (the 'Y-board') has been designed to accommodate the circuit of figure 2. The track pattern and component layout of the board are shown in figure 3. As can be seen, the switch and potentiometers are mounted directly on the board, to facilitate construction. Normal 5% components are used in the voltage divider network, and these should prove quite adequate for most applications; if desired, however, closer tolerance resistors can be used. Two Y-boards are required for a two-channel version of the basic TV scope.

Main board

The main board accommodates a number of circuits which perform a variety of different functions. There is the crystal oscillator, the timebase circuit, which is also responsible for generating the

graticule, and the white-level pulse generator which produces the actual trace on the screen.

For the sake of clarity, each of these circuits is granted a separate diagram, i.e. figures 4a, 4b and 4c respectively.

Figure 4a shows the circuit diagram of the crystal oscillator. A 4.433 MHz crystal is used, and since this type is commonly found in a number of colour TV sets it is both reasonably cheap and easy to obtain.

The output of the actual oscillator circuit round T1 is fed via a NAND-buffer to a frequency divider, IC1. This is a CMOS decade counter, which is in fact connected as a nine-counter. The output of the frequency divider, which forms the output signal, u_f, of the oscillator circuit, is one-ninth of the crystal frequency, i.e. 492.5 kHz. The circuit also has an input for a control

Figure 1. Block diagram of the basic version of the circuit (i.e. the second input amplifier and the sync circuit) and the TV-modulator, which is discussed elsewhere in this issue, are shown in dotted lines.

Figure 2. The circuit diagram of the input amplifier. Those voltages which are shown in brackets refer to the description of the extended TV scope to be published next month. For a two-channel version of the scope two such circuits will obviously be required.

voltage, u_{xtal}; if u_{xtal} equals the supply voltage of 15 V and a signal is present at the output of the circuit, the oscillator functions normally and a signal is present at the output of the circuit; however, if u_{xtal} equals 0 V the oscillator is inhibited and no output signal is produced. This control input functions in conjunction with the sync input, which will be discussed later. If, however, the sync circuit is not incorporated in the basic scope, u_{xtal} should simply be connected to +15 V.

The timebase circuit is shown in figure 4b. As far as the function of the basic TV scope is concerned, only three of the timebase outputs are of importance: the composite sync pulse, u_{sync}, the line sync pulse, u_{line}, the period of which corresponds to the line period of the TV receiver, and u_{cal}, which is responsible for generating the graticule or calibration lattice. A fourth timebase

3
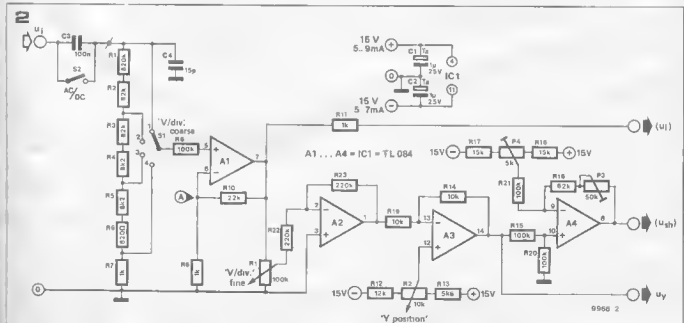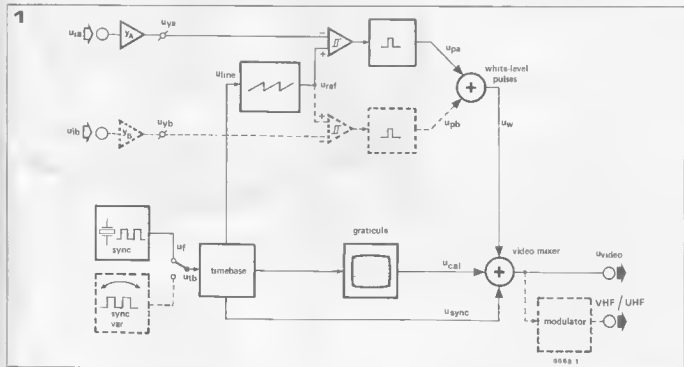
Parts list for input amplifier
(figures 2 and 3)

Resistors:
R1 = 820 k
R2,R3,R16 = 82 k
R4,R5 = 8x2
R6 = 820 Ω
R7,R8,R11 = 1 k
R9,R15,R20,R21 = 100 k
R10 = 22 k
R12 = 12 k
R13 = 5k6
R14,R19 = 10 k
R17,R18 = 15 k
R22,R23 = 220 k
P1 = linear potentiometer, 100 k
P2 = linear potentiometer, 10 k
P3 = preset potentiometer,
50 k (47 k)
P4 = preset potentiometer
5 k (4k7)

Capacitors:
C1,C2 = 1 µ/25 V tantalum
C3 = 100 n*
C4 = 15 p

Semiconductors:
IC1 = TL 084

Miscellaneous:
S1 = single-pole 4-way switch
S2 = single-pole single-throw*
switch
*Note: C3 and S2 are not
mounted on the p.c. board
(see figure 16).
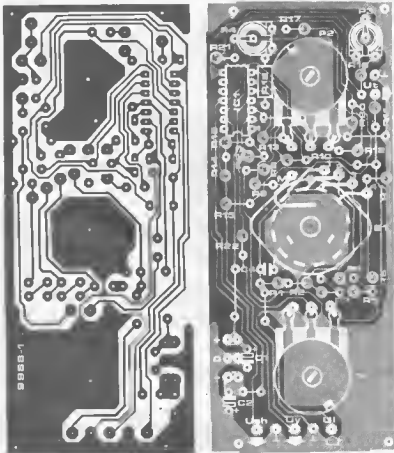
Figure 3. The p.c.b. for the input amplifier
(EPS 9968-1). With the exception of the
AC/DC switch on which is mounted C3, the
controls are mounted directly on the board,
thereby facilitating construction. The board is
provided with a generous copper earth plane,
however it should nonetheless ba carefully
screened.

Figure 4. The various circuits mounted on the
main board: figure 4a shows the crystal
oscillator and 4b the timebase circuit, which,
as can be seen, has a large number of outputs,
although only four of these are utilised in the
basic version of the TV scope.
Figure 4c shows the circuit which generates
the white-level pulses for the trace of the
input signal. In the case of the basic TV scope,
u_{gate} output should be connected to the
+ supply rail, regardless of whether it is a one-
or two channel version.

4a
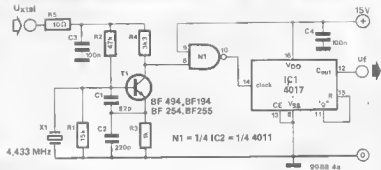
output signal, u_{reset}, although it does
not affect the operation of the basic
scope, is brought out as an external
trigger signal for other devices (e.g. see
figure 12 of 'TV Scope' — contained
elsewhere in this issue).
All the timebase output signals are
derived from a single input signal, u_{tb}.
For most applications (with the excep-
tion of those which require the variable
sync circuit) u_{tb} is equal to the output
signal, u_f, of the crystal oscillator circuit.
The input signal is fed to a number of
dividers, the first two of which are
formed by flip-flops FF1 and FF2. The
following 12 divider stages are all con-

tained in IC5, and the various divider
outputs are combined using logic gates
to produce the desired timebase output
signals. These do not conform exactly
to the CCIR-norm for television signals,
but the differences are so small as to
have virtually no effect in practice. The
u_{sync} signal is the conventional com-
posite sync signal, containing the line-
and field sync pulses. This signal is fed
straight to the video mixer.
The u_{line} signal produces only line
pulses, and does so even during a field
pulse in u_{sync}. The line pulses are used
to trigger the sawtooth generator (see
figure 1).

4b

N2 ... N5 = IC7 = 4011
N6 ... N7 = IC6 = 4012
N8 ... N11 = IC4 = 4011
N12 = 1/4 IC8 = 1/4 4071
N13 = IC9 = 4068
FF1, FF2 = IC3 = 4013

4c

IC10 = 741
IC11 = 709
IC12 = 709
N14 = 1/4 IC8 = 1/4 4071
N15,N16 = 1/2 IC2 = 1/2 4011

* see text
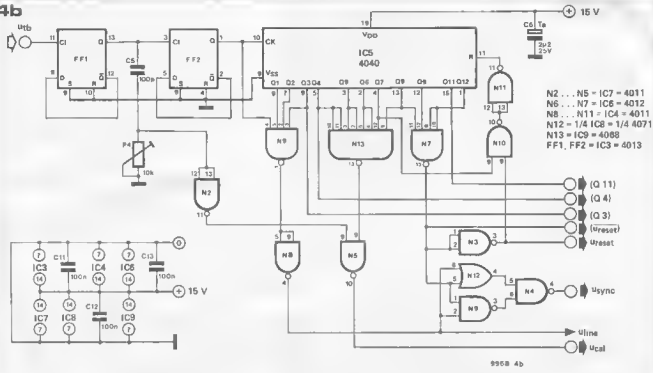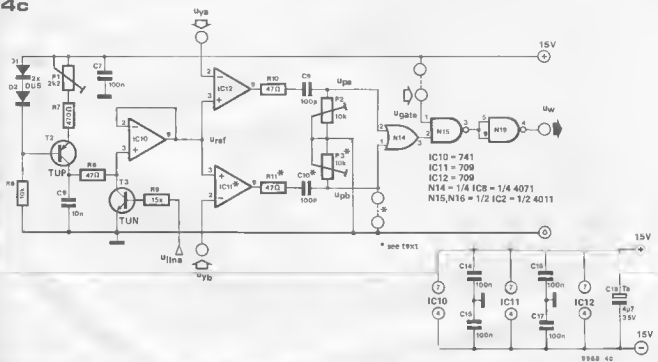
The trace of the input signal is not the only image displayed on the screen of the TV (-scope). There is also a graticule which can be used as a calibration grid. How this graticule is formed is illustrated in figure 5, where the TV screen is shown turned on its side. The vertical lines of the graticule (which define the time-axis) are generated by driving the picture signal to white-level at regularly recurring intervals, whilst the horizontal lines (which define the voltage-axis) are produced by a regularly spaced series of white-level pulses on each of the picture lines.

The signal used to generate these white-level pulses is derived from the output of flip-flop FF1 (see figure 4b), and consists of a squarewave with a frequency of 246 kHz. When fed to the monostable round C5, P4 and N2 the result is a train of narrow pulses with a repetition rate of roughly 4 µs. This means that each picture line, which has a period of 64 µs, contains 15 such pulses, i.e. the graticule has 15 horizontal lines. However not all of these lines are visible on the final picture displayed by the TV scope, since some occur during the line blanking interval, and also the extreme edges of the picture fall outside the screen of a conventional TV receiver.

The pulse width of the monostable output can be varied by means of preset P4. This has the effect of varying the width of the graticule lines.

The vertical graticule lines are generated by a 64 µs white-level signal every 32 lines. This is derived from IC5 via NAND-gate N13. The interval between successive horizontal graticule lines is thus 32 x 64 µs = approx. 2 ms (see figure 19). The vertical- and horizontal graticule signals are combined in NAND-gate N5.

The logic gating for the line sync pulses is provided by N6 and N8, whilst the

field sync pulses are derived via N7. These are combined by N4, N9 and N12 to produce the composite sync signal, u_{sync}.

The white-level pulses which constitute the trace of the input signal are generated by the circuit shown in figure 4c. T2 and T3 form a linear sawtooth generator, which in fact consists of a constant current source (T2) which is used to linearly charge capacitor C8. The capacitor is discharged via T3 by the line pulse u_{line}, which causes this transistor to saturate. The resulting sawtooth, u_{ref}, is thus in sync with each picture line. Via a buffer stage, IC10, it is fed to the comparator circuit formed by IC11 and IC12, where it is compared with the output signals, u_{ya} and u_{yb}, of the two input amplifiers. The white-level pulses are then generated by simple differentiating networks before being
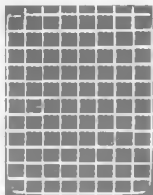
combined in the OR-gate, N14, N15 and N16 function as buffers. In the basic version of the TV scope the u_{gate} input should be connected to +15 V by means of the wire link adjacent to P1 (marked *).

If a single-channel version of the basic

Figure 5. This photo illustrates how the graticule is generated on the TV screen. The horizontal lines (with the TV turned on its side) consists of a regularly spaced series of white-level pulses in each picture line, whilst the vertical lines are produced by driving the picture signal of every 32nd line into the white level.

Figure 6. Track pattern and component overlay of the main p.c.b. (EPS 9968-2). The outputs are grouped so as to simplify the connections to the other boards.

Figure 7. Circuit diagram of the video mixer. P1 and P2 are the intensity controls for the signal trace and graticule respectively.

Figure 8. The operation of the video mixer is clearly illustrated in this pulse diagram.

Figure 9. The p.c.b. of the video mixer (EPS 9968-3).
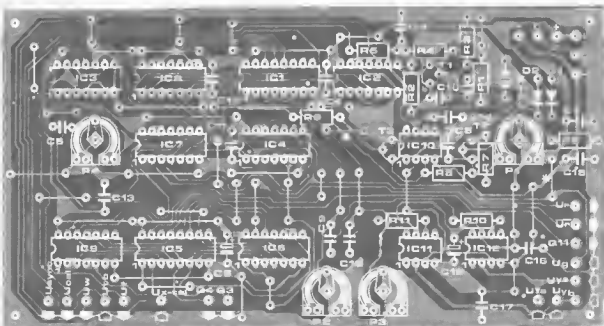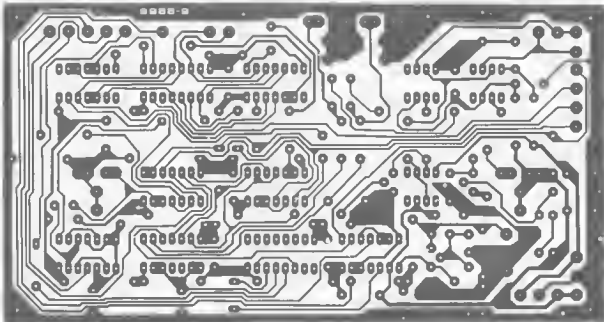
Parts list for video mixer (figures 7 and 9)

Resistors:
R1 = 33 k
R2 = 47 k
R3 = 10 k
R4 = 1 k
R5,R6 = 18 k
R7 = 2k2*
P1 = 250 k (220 k)
linear potentiometer
P2 = 500 k (470 k)
linear potentiometer
P3 = 1 k preset potentiometer

Capacitors:
C1 = 100 n

Semiconductors:
T1,T2 = TUN
D1,D2 = DUS
D3 = LED*
*Note: R7 and D3 are not actually mounted on the board proper (see figure 16)

7
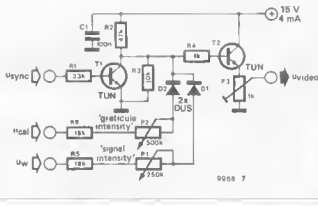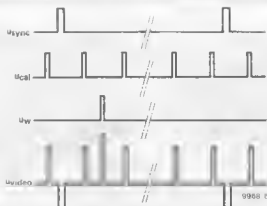
Parts list for main board (figures 4 and 6)

Resistors:
R1,R9 = 15 k
R2 = 47 k
R3 = 1 k
R4 = 3k3
R5 = 10 Ω
R6 = 10 k
R7 = 470 Ω
R8,R10,R11 = 47 Ω
P1 = preset potentiometer, 2k5 (2k2)
P2,P3,P4 = preset potentiometer, 10 k

Capacitors:
C1 = 82 p
C2 = 220 p
C3,C4,C7,C11 . . . C17 = 100 n
C5,C9,C10 = 100 p
C6 = 2µ2/25 V tantalum
C8 = 10 n
C18 = 4µ7/35 V tantalum

Semiconductors:
IC1 = CD 4017
IC2,IC4,IC7 = CD 4011
IC3 = CD 4013
IC5 = CD 4040
IC6 = CD 4012
IC8 = CD 4071
IC9 = CD 4068
IC10 = 741
IC11,IC12 = 709
T1 = BF 194,BF 195,BF 254,
BF 255,BF 494,BF 495
T2 = TUP
T3 = TUN
D1,D2 = DUS

Miscellaneous:
X1 = crystal, 4.433 MHz
(colour TV crystal)

8

9
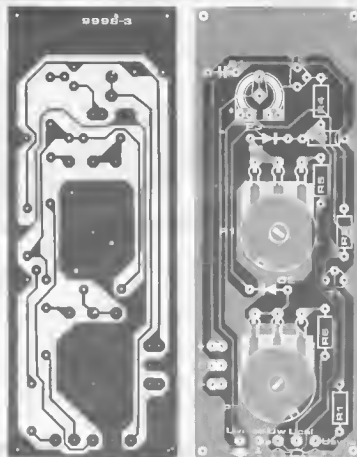
10
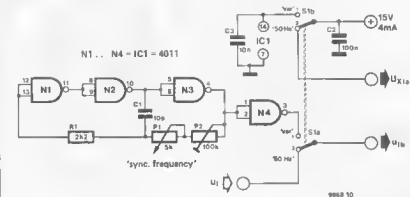
Parts list for sync circuit
(figures 10 and 11)

Resistors:
R1 = 2k2
P1 = linear potentiometer,
5 k (4k7)
P2 = preset potentiometer, 100 k

Capacitors:
C1 = 10 p
C2 = 100 n
C3 = 10 n

Semiconductors:
IC1 = CD 4011

Miscellaneous:
S1 = DPDT switch

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13

Figure 10. The circuit diagram of the (variable) sync circuit, which basically consists of a CMOS squarewave generator and a switch. This circuit, although not essential, increases the possibilities of the basic TV scope; however it is completely superfluous in the extended version of the scope.

Figure 11. Track pattern and component overlay of the p.c.b. for the sync circuit shown in figure 10 (EPS 9968-4). The generous spacing between components, is due to the desire to standardise the size of the subsidiary boards.

Figure 12. Suggested design for a TV scope front panel. The various controls are arranged for ease of operation. The dimensions of the p.c.b.'s have been deliberately tailored to accommodate a front panel of this type.

Figure 13. This photograph shows how the front panel and boards carrying the controls can be mounted together in a compact sandwich-construction. The screening plates round the boards can clearly be seen.

scope is required, the upb input should be connected to earth by means of the wire link adjacent to T3 (marked *). Strictly speaking, IC11, R11 and C10 can also be omitted, however, in view of the possibility of subsequent extension of the circuit to two channels, it is well worth the minimal cost of these components to keep one's options open.
The size (i.e. line width) of the resultant trace can be varied by means of potentiometers P2 and P3. Preset P1 is used to adjust the sawtooth oscillator.

The circuits of figures 4a, b and c are all mounted on the one board, the track pattern and component overlay of which are shown in figure 6. In order to keep the interwiring between boards as simple as possible, a fairly large number of wire links are used. In the basic version of the scope u̲gate is connected to the positive supply rail.

Video mixer

All the necessary components of the

13

(front panel layout — y-channel, y-channel, synchronisation, intensity)

composite video signal have now been generated: u_{sync} for the line- and field sync pulses, u_{cal} for the graticule, and $u_{\overline{}}$ for the actual trace of the input signal. All that remains is to sum these three in the relatively simple video mixer circuit shown in figure 7.

The operation of this circuit is illustrated by the timing diagram of figure 8, and needs little explanation. The u_{sync} signal is inverted by T1, and after being summed in the correct proportion with the other signals, is fed to the base of the

emitter follower, T2, which functions as an output buffer for the composite video signal.

The intensity or brightness of the input signal waveform and of the graticule can be independently adjusted by means of potentiometers P1 and P2 respectively. P3 controls the level of the video output voltage, the maximum value of which is approx. 6.5 V p-p. The video signal can either be fed direct to a TV-receiver equipped with a video input or to a modulator circuit such as the VHF/UHF

TV modulator described elsewhere in this issue. The video mixer circuit (including the 'power' indicator LED also shown in figure 7) is mounted on the p.c.b. shown in figure 9.

Sync board

The circuits discussed so far form the most basic version of TV scope — more about which later. With the addition of a simple variable sync circuit, however,

Figure 14. Circuit diagram of the TV scope power supply, which is suitable for both the basic and extended versions.

Figure 15. The p.c.b. for the power supply (EPS 9968-5).

Figure 16. Details of the wiring between the various boards used in the basic scope. Screened wire should be used where indicated; otherwise normal, reasonably thick, connecting wire can be used. The AC coupling capacitor, C3, of the input amplifier is mounted on the AC/DC switch.

Parts list for power supply (figures 14 and 15)

Capacitors:
C1,C2 = 470 µ/35 V
C3,C4 = 100 n
C5,C6 = 1 µ/25 V tantalum

Semiconductors:
IC1 = 7815
IC2 = 7915
D1 ... D4 = 1N4001

Miscellaneous (not on p.c. board proper, see figure 16)
Tr1 = mains transformer, 2 x 18 V/250 mA
S1 = double-pole mains switch
F1 = fuse, 100 mA
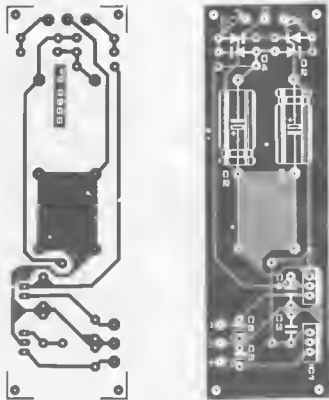
it is possible to extend the range of applications of the scope. The sync circuit is in fact an oscillator which can be varied over a small range of frequencies and which replaces the crystal oscillator on the main board. The TV scope will continue to work satisfactorily with most TV sets at frequencies slightly different to that of the crystal oscillator signal. This fact can obviously be utilised to obtain a stable trace of input signals whose frequencies are not quite an exact multiple of 50 Hz by correspondingly adjusting the frequency of the sync oscillator.

The circuit diagram of the variable sync oscillator is shown in figure 10, and as can be seen consists of nothing more than a CMOS squarewave generator, a switch and three capacitors. The output voltage of the sync circuit is u_{th}, i.e. the input voltage of the timebase circuit.

Depending upon the position of the switch, this voltage is equal either to the output voltage, u_f, of the crystal oscillator circuit, which is also mounted on the main board, or derived from the squarewave generator. The switch also has a second pole; in one position it provides the crystal oscillator with supply voltage, and in the other position it does the same for IC1 of the sync circuit. This precautionary measure is necessary to prevent the two oscillators, which operate at almost the same frequency, from influencing one another.

Readers who intend constructing the extended version of the TV scope, should note that the variable sync oscillator is the only circuit of those discussed so far which will not be used when the scope is upgraded.

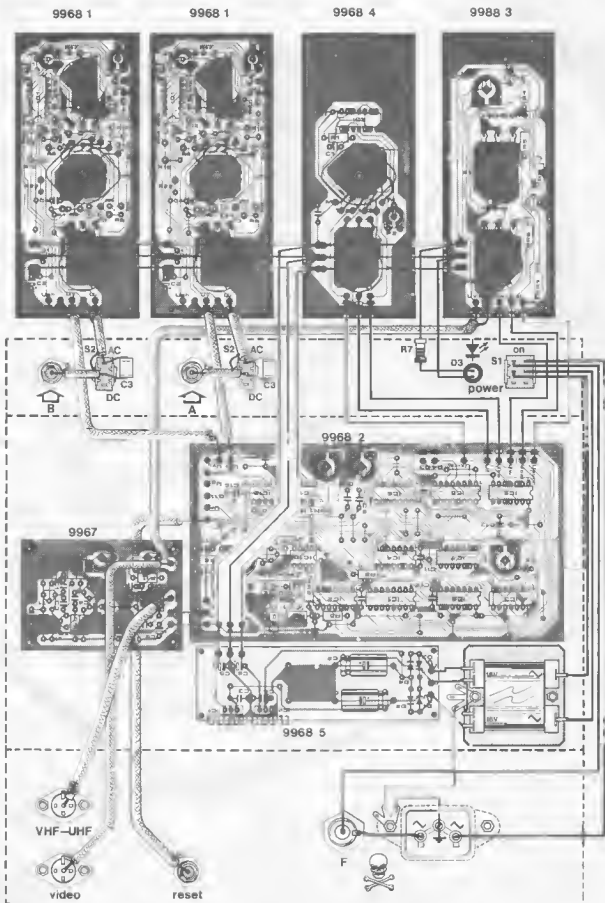The track pattern and component layout of the sync circuit p.c.b. is

shown in figure 11. The component spacing is fairly generous since it was decided to standardise the dimensions of all the subsidiary boards.
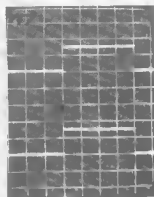
The complete TV scope

The various circuits described above together form the complete basic TV scope. Thanks to the uniform dimensions of those boards which accommodate the controls, it is a simple matter to fit them to the front panel shown in figure 12.

In view of the sensitivity of the input amplifier (10 mV/div), it should be well screened. Although the board is already provided with a large copper earth plane, it is recommended that it be further shielded by strips of copper laminate board which can be soldered to the earth plane. The other boards carrying controls can also be screened in this

16

9968 1   9968 1   9968 4   9988 3

B   A

S2 AC   S2 AC
C3   C3
DC   DC

R7   D3   S1 on
power

9968 2

9967
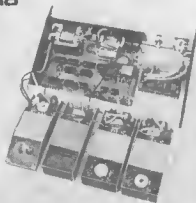
9968 5

VHF—UHF

video   reset

F

9968 16

18

Figure 17. If P1 on the main board is adjusted correctly, this should be the signal which appears on the screen when the Q11 output signal is fed to the input of the TV scope — see text.

Figure 18. A finished prototype of a two-channel basic TV scope, with the sync circuit.

Figure 19. Analysing a signal on the TV scope. Range: 10 mV/div. Assuming the dotted line coincides with a voltage of 0 V (to be set with the aid of the Y-position control under quiescent conditions), the sinewave displayed on the screen has an average value of 30 mV, and measures 40 mV peak-to-peak. The period of the signal is approx. 6 ms, the frequency is therefore roughly 167 Hz.

Figure 20. In some (exceptional) cases this offset nulling circuit may have to be added to the input amplifier (figure 2).

way (in the case of the sync board, for example, it is important to prevent r.f. radiation). A piece of copper laminate mounted over the range switches completes the screening precautions. This is illustrated in figure 13.

The construction of the completed scope is illustrated by the photograph in figure 18.

A special power supply was designed for the TV scope, and this is shown in figure 14; the corresponding p.c.b. is given in figure 15. To prevent mains hum etc., the transformer should be mounted as far away as possible from the input amplifier(s).

Details of the wiring between the different boards are given in figure 16. Screened wire should be used where indicated; the remaining connections can be made with normal — reasonably thick — connecting wire.

If the sync board is not used, points u₀b and u₀f on the main board should be joined together, and the u_xtal connection, which is also on the main board, should be wired to the 15 V plus supply rail.

If a one-channel version of the scope is desired, u₀b should be connected to earth, thereby rendering the u_yb input on the main board inoperative.

Constructional hints

In order to keep the size of the completed unit to reasonable proportions, the p.c. boards are designed to accommodate miniature (20 mm diameter) potentiometers. Larger diameter pots, should not be used as these will not fit the board. Care should be taken when mounting the pots that the lockwashers do not short out any of the copper tracks, especially on the Y boards.

Bulk parts list for basic version of TV scope			
(two input amplifiers, main board, video mixer, sync circuit and power supply)			

Resistors:

Number	value
1	10 Ω
3	47 Ω
1	470 Ω
2	820 Ω
8	1 k
2	2k2
1	3k3
4	5k6
2	8k2
6	10 k
2	12 k
6	15 k
2	18 k
2	22 k
1	33 k
2	47 k
6	82 k
8	100 k
4	220 k
1	820 k
1	5 k (4k7) linear potentiometer
2	10 k linear potentiometer*
2	100 k linear potentiometer
1	250 k (220 k) linear potentiometer
1	500 k (470 k) linear potentiometer
1	1 k preset potentiometer
1	2k5 (2k2) preset potentiometer
2	5 k (4k7) preset potentiometer
3	10 k preset potentiometer
2	50 k (47 k) preset potentiometer
1	100 k potentiometer

*Note: maximum diameter 20 mm.

Capacitors:

number	value
1	10 p
2	15 p
1	82 p
3	100 p
1	220 p
2	10 n
16	100 n
6	1 μ/25 V tantalum
1	2μ2/25 V tantalum
1	4μ7/35 V tantalum
2	470 μ/35 V

Semiconductors:

number:	type:
4	CD 4011
1	CD 4012
1	CD 4013
1	CD 4017
1	CD 4040
1	CD 4068
1	CD 4071
2	741
1	709
2	TL 084
1	7815
1	7915
3	TUN
1	TUP
1	BF 194, BF 195, BF 254, BF 255, BF 494, BF 495
4	DUS
4	1N4001
1	LED

Miscellaneous:

number:	type
1	crystal 4.433 MHz (colour-TV-crystal)
2	single-pole switches
1	double-pole mains switch
1	double-pole double-throw switch
2	single-pole 4-way switches
1	fuse 100 mA
1	mains transformer, 2 x 18 V/250 mA

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20
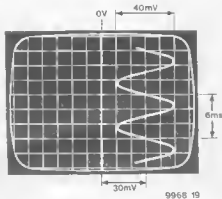
To avoid pins or component leads shorting to the front panel the boards should be mounted at least 3 mm behind the panel, using insulated spacers to avoid shorting out any of the copper tracks. The boards should be screened using pieces of copper laminate board or tinplate, soldered around the edges of the p.c.b.'s to form boxes, as shown in figure 13.

Because of the compact construction and the large amount of metalwork mounted on the panel it is essential that all the mains wiring should be well insulated to avoid the possibility of short-circuits. Take particular care with the mains switch, which is mounted on the front panel.

During experiments with the prototype of the TV scope it has been noticed that problems may occur due to the offset voltage of some TL084 ICs, which can result in a DC shift at the output of A1. The effect of this is to cause the position of the trace to change when the fine Y-gain control is operated, even with zero input voltage. The amount of shift depends on the offset voltage of A1, which varies from IC to IC. If the shift is unacceptable then an offset nulling circuit may be incorporated as shown in figure 20. The preset should be adjusted until the output voltage of A1 is zero with no input signal. The capacitor prevents spurious AC signals from reaching A1 via the 1 M resistor.

Calibration

Even the relatively simple basic version of the TV scope contains a fair number of preset potentiometers; however despite this fact, the adjustment and calibration procedure is not particularly complicated, and (of course) does not require an oscilloscope!

Before the power supply is connected to the rest of the circuit, the supply voltages of + and − 15 V should be checked. The first thing to be adjusted is the TV modulator, and the appropriate procedure is described in the separate

article on this circuit, which is contained elsewhere in the issue. When tuning the modulator, the 'graticule intensity' control should be turned fully clockwise, and the 'signal intensity' control fully anticlockwise; presets P3 and P4 on the main board should likewise be turned to their right-hand stops. In addition, if the sync board is used, the sync selector switch should be in the 50 Hz position.

Once the modulator has been aligned, the brightness of the graticule can be adjusted as required by means of the 'graticule intensity' control. The thickness of the graticule lines can be varied by means of preset P4 on the main board.

The 'signal intensity' control should now be turned fully clockwise and the two presets, P2 and P3, on the main board which determine the thickness of the resultant trace should also be turned fully clockwise.

P1 on the main board (which controls the sawtooth generator) should be set to the mid-position. Under quiescent conditions (i.e. in the absence of any input signal) it should now be possible to adjust the Y-position controls such that two vertical (i.e. perpendicular to the line scan of the receiver) white lines appear on the screen.

By means of presets P2 and P3 respectively (both on the main board) the width of the trace can be varied for both channels (i.e. adjusted to be slightly broader than the graticule). The brightness of the trace(s) can be varied by means of the 'signal intensity' control.

The next step is to set the range switch (V/div) of the Y-amplifier(s) to the 10 V position and turn the range potentiometer fully clockwise ('cal'). A signal from the TV scope itself, namely from the 'Q11' terminal on the main board (this is one of the connections which is not used in the basic version of the scope) is then fed to the input of the Y-amplifier. A squarewave signal, four divisions of the graticule in length (i.e. along the time-axis), should now appear
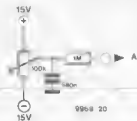
on the screen. The amplitude of this signal should be adjusted to exactly one and a half graticule divisions (this corresponds to 15 V) by means of preset P1 on the main board, so that the resulting trace looks like that shown in figure 17. During this procedure the trace will shift across the screen to the left; this should be corrected by means of the Y-position control.

The calibration of the TV scope is now completed. It of course goes without saying that the values of the range switch (V/div) thus obtained are only valid for the extreme clockwise setting ('cal') of the range potentiometer.

Literature: 'An introduction to the TV scope' (elsewhere in this issue) 'VHF/UHF TV Modulator' (also elsewhere in this issue) ◼

APPLIKATOR

National Semiconductor LM 1890
light sensing chip

The LM 1890 is a general purpose building block for use in visible light and infra-red applications whether they be analogue or digitally switched. Included on a single bipolar chip are a linear light-to-current converter, a voltage comparator biased with light-derived currents, and a voltage reference (figure 1).

Photocurrent is produced by an on-chip photodiode that is ion-implanted to produce a shallow junction depth. The resultant spectral response is enhanced in the visible light range (figure 2) as compared with conventional silicon photodiodes. The sensor is commonly referred to as 'blue-enhanced'. Careful processing and a unique geometry will help to reduce the diode's dark current leakage. In addition, active circuitry is employed to set its voltage bias very close to zero volts, further minimising its leakage. Quality silicon photodiodes are well known for their excellent linearity and wide dynamic range. The LM 1890 photodiode is followed by a high gain current amplifier that is specially designed to maintain the sensor's excellent characteristics yet provide a larger, and therefore more useful, linear photocurrent output, IX. This gain cell acquires its dynamic range from unique circuit techniques that rely on a second, 'active load' photodiode imbedded within the area of the larger, sense photodiode. This second photodiode is accurately scaled in area and other parameters to the sensor diode; a natural outcome of their monolithic construction. It is used as an active load in a differential amplifier that is

the heart of the current gain cell. The gain factor of the cell can be adjusted at the wafer sort level in order to guarantee the absolute value of light-to-current conversion to smaller tolerances than are commonly available. The light-to-current conversion is held to a linear range of operation over 5 decades of illumination with excellent linearity observed over a 3 decade range. The current mode output avoids the dynamic range limitations of voltage mode operation but handles IX from 1 nA to 10 mA. Since the current sink output is cascaded, its output impedance extends into the gigaohm region while operating from 1 V to full supply.

The voltage comparator is specially designed to sense the IX node with minimum loading even in extremely low light where IX may only be a few nanoamps. Comparator input bias current is held to typical 1% of the IX output, independent of the illumination, by allowing all of the comparator's bias currents to track with the light. This extreme form of adaptive bias tailors input bias current and speed to the incident light level, trading the latter for the former at low power light levels. The voltage reference is of the semiconductor bandgap variety. It is referenced to V+ because the IX output is frequently converted to a voltage with a supply referred resistor or capacitor for use with the comparator.

The device is housed in an 8 pin, Dual-In-Line, clear plastic package that has a recess in the top for insertion of an optical filter or lens. Some typical applications of the LM 1890 are given in the following paragraphs.

Simple light-sensitive switch

Only one external component is needed to create a sensitive, stable light-sensitive switch (figure 3). That component, R1, sets the switching threshold, which occurs when the voltage across R1 reaches the internal reference potential of 1.3 V. Simple hysterasis can be added in the form R2, which changes the effective trip point slightly as a function of output swing by working against the 600 Ω internal series resistance of the voltage reference. Although the circuit is phased such that the load is energised when the illumination exceeds the set threshold, the opposite phase can be used just as easily. The circuit works equally well for any supply voltage between 2.5 V and 25 V.

In this example the load is a relay driven via an external transistor.

Burglar alarm

The circuit (figure 4) looks for quick changes, like path interruptions, in the ambient light level. It can be used in intrusion alarms or any back-lit object detection. VREF is used as a level shift to put the inputs of the comparator within their common-mode range. The total received light appears at the non-inverting input. Except for the drop across R1, the average of the total received light appears at the inverting input. R1 is used to control the dead band so that peak AC variations of the ambient light do not trip the comparator. Choose

$$R1 > R3 \frac{\text{peak AC ambient illumination}}{\text{DC ambient illumination}}$$

Switching occurs when

$$IX = \frac{R_3 + R_1}{R_3} \quad IX \text{ (avg)}$$

Light meter

This relatively simple circuit (figure 5) offers high performance. The on-chip comparator is used as an op-amp by frequency compensating it with a large tantalum capacitor at the output. This capacitor should be connected directly from pin 1 to pin 5 or small amplitude high frequency oscillations may occur. IX is converted to a light-dependent voltage via selection of the appropriate resistor for the desired range. The op-amp will then current drive the meter, M1, with IX (R2/R5). The LM 1890 is internally calibrated, but more accurate calibration can be accomplished by making R5 a potentiometer. The four light ranges are given below:
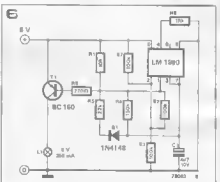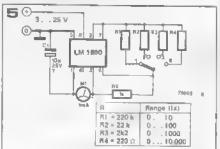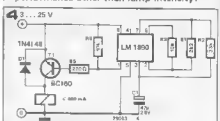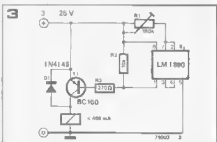
R1 = 200 k	0	1fc
R2 = 20 k	0	10 fc
R3 = 2 k	0	100fc
R4 = 200 Ω	0	1000 fc

Roadworks barrier

This circuit (figure 6) delivers about 500 mA pulses to an incandescent lamp after the sun has set. The comparator runs as a 1.4 Hz oscillator with approximately a 13% duty cycle, set by R4, R5, D1, and C. As the sun begins to rise, so does IX, effectively stealing the charging current from C. The light threshold is given

approximately by $\frac{\frac{1}{2} V+}{(R_4)(5 \mu A/fc)} = 4fc.$

R8 provides bias for the comparator during darkness when little or no light bias is present. Quiescent supply current is ~ 3 mA and the circuit works down to 2.5 V with little change in performance other than lamp intensity.

1

2

3

4

5

R	Range (fc)
R1 = 200 k	0 ... 1
R2 = 22 k	0 ... 100
R3 = 2k2	0 ... 1000
R4 = 220 Ω	0 ... 10,000

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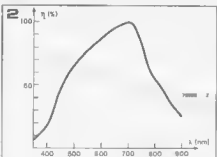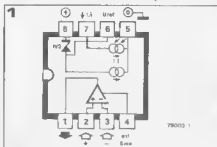
Table 1.

ELECTRICAL CHARACTERISTICS PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
LM 1890 as a whole (25°C, V+ = 10 V, and pin 4 open unless otherwise specified)					
Supply voltage		2.5	10	25	V
Supply current	H¹ = 0 mW/cm²		1.4	4	mA
LIGHT-TO-CURRENT CONVERTER (V$_{pin\,7}$ = 10 V)					
Photocurrent (Iλ)	H¹ @ 660 nm wave length (see fig. 2)		0.32		mA (mW/cm²)
Photocurrent (Iλ)	H¹ @ 2854° K color temperature		5		μA/fc
Photocurrent (Iλ)	H¹ @ 4700° K color temperature		2.5		μA/fc
Linearity	0.01 to 100 fc referenced to 1 fc		± 4		%
Dark current	H¹ = 0 mW/cm²		< 1		nA
Photodiode active area	'sensor' photodiode		720		mils²
Photodiode Responsivity	'sensor' photodiode²		0.36		A/W
Output resistance	@ pin 7, Iλ = 10 μA		2		GΩ
	$\frac{\Delta I\lambda}{\Delta V_{pin\,7}}$ for any value Iλ		0.005		%/V
Power supply rejection ratio	$\frac{\Delta I\lambda}{\Delta V+}$ for any value Iλ		0.03		%/V
Frequency response	−3 dB³,		70		kHz/ (mW/cm²)
Step response	from t = 0 to 90% of final value⁴,		30		μS (mW/cm²)
Step response	from t = 0 to 90% of final value⁵,		8.5		μS (mW/cm²)
Output voltage compliance	range of linear operation of Iλ output @ pin 7	1		25	V
VOLTAGE COMPARATOR (R$_{load}$ = 1 k, input common mode = 5 V, H¹ = 15 μW/cm² @ 660 nm)					
Input offset voltage	note 6		1		mV
Input bias current (I$_B$)	note 6		50		nA
Input offset current (I$_{os}$)	note 6		± 5		nA
			0.001		A/A
	I$_{os}$/Iλ				
Small signal DC voltage gain	note 6		100		dB
Input common-mode rej. ratio	note 6		100		dB
Input common-mode voltage range	note 6	1.2		V+−0.5 V	V
Input differential voltage range	either polarity	V+			V
Output saturation voltage	I$_{load}$ = 10 mA @ pin 1		0.5		V
Output saturation resistance	pin 1 to GND		60		Ω
Output sink current	pin 1 @ 1 V	10	30		mA
Output leakage current	pin 1 @ 25 V		5		nA
Response time	100 mV step, 5 mV overdrive, inverting mode,		1.1		μS
VOLTAGE REFERENCE					
Reference voltage	referred to V⁺		1.33		V
Dynamic impedance	@ 1 kHz		600		Ω
Power supply rej. ratio			70		dB

NOTES:
1. 'H' is used to designate either illumination (fc) or irradiance (mW/cm²)
2. Determined by uniformly irradiating the entire die @ 700 nm and then calculating the ratio of the actual photodiode output (nominally Iλ/200) to the light power (flux) incident on that photodiode's active area.
3. Small sinusoidal AC light signal superimposed upon DC light level measured in mW/cm² @ 600 nm.
4. Excitation consists of LED @ 660 nm stepping from absolute zero (worst case) to a given irradiance, Ho, in mW/cm² @ t = 0.
5. Excitation consists of LED @ 600 nm stepping from Ho/3 to a given irradiance, Ho, in mW/cm² @ t = 0.
6. At output switch point where V$_{pin\,1}$ = 5 V.

analogue reverberation unit

Until comparatively recently the only audio delay units that were within the budget of most home constructors were of the spring line type, which suffer from a number of disadvantages such as fixed delay time, uneven and limited frequency response, and susceptibility to mechanical vibration. Recently, however, completely electronic delays have become a feasible proposition, with the result that high-quality reverberation and other audio effects are now within economic reach of the amateur. A design for a digital reverberation unit has already been published in Elektor. The circuit published here represents an alternative approach using analogue techniques.

As explained in the article on the digital reverberation unit (Elektor 37, May 1978) a digital delay line is an elegant method of producing reverberation and other time-related audio effects. In a nutshell, the analogue input signal is converted to a digital code using an A/D converter. This code is then fed through a shift register of the desired length to produce the delay, and the analogue signal is reconstituted at the output by a D/A converter. This method has a number of advantages. Firstly, since it is a digital signal that is being passed through the shift register, the signal that comes out will be identical to that which goes in irrespective of the length of the shift register. Any noise and distortion in the retrieved analogue signal are due only to deficiencies in the A/D and D/A conversion processes.

Secondly, once the initial investment in A/D and D/A converters has been made, the digital delay line can be extended to any length, simply by the addition of inexpensive digital shift registers. These two factors make the digital delay line an ideal choice for long delays such as those required for echo effects.

An alternative approach to a digital delay line is an analogue delay line using analogue shift registers (bucket brigade memories) such as those used in the Phasing and Vibrato Unit (Elektor No. 20, December 1976). These accept an analogue signal directly and transfer it from input to output as a sequence of charge packets, of which more later. Analogue shift registers are an attractive

proposition for short delay times, since the cost of a 1024-bit analogue shift register (between £12 and £18) is less than the cost of an equivalent digital shift register plus A/D and D/A converter. Furthermore, the analogue shift register does not suffer from 'quantisation noise' which is inherent in the A/D conversion process.

The analogue shift register is thus ideal for producing effects such as phasing, flanging and vibrato and for the modest reverberation times required for enhancement of room ambience. However, the analogue shift register is not such an attractive proposition for longer delay times, since noise and distortion increase as the analogue shift register is made longer.

Analogue shift registers

Analogue shift registers are commonly referred to as 'bucket-brigade memories', since their operation is analogous to that of a chain of men passing buckets of water from hand to hand, the 'buckets' being capacitors and the 'water' being electric charge.

The basic principle of an analogue shift register is illustrated in figure 1. It consists of a number of capacitors and (electronic) switches. The switches are opened and closed alternately by a two phase clock generator, i.e. an oscillator which generates two squarewaves in antiphase. When $S1a$, b, c etc. are closed then $S2a$, b, c etc. are open, and vice versa. The input signal is applied to $S1a$. When this switch is closed then

---

Specification

|	
Signal-to-noise ratio at maximum output level:	> 60 dB
Bandwidth of reverberation signal:	2.5 kHz, 5 kHz or 15 kHz (see text)
Maximum delay time:	200 ms, 100 ms or 33 ms (see text)
Monitor output bandwidth:	25 Hz to 100 kHz.
Input sensitivity:	variable; most sensitive setting gives maximum output for 30 mV RMS (100 mV peak-to-peak) input.
Maximum output:	2.5 V peak-to-peak
External clock input:	15 V p-p, 5 kHz to 500 kHz.
Supply voltage:	+15 V/75 mA, −15 V/25 mA
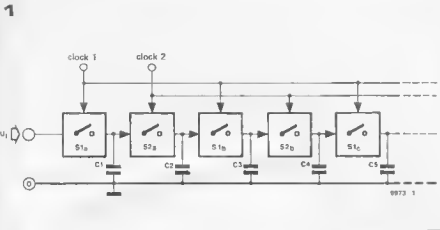
Figure 1. Illustrating the principle of the bucket-brigade memory.

Figure 2. The analogue input signal is sampled during each clock pulse.

Figure 3. Internal circuit of an analogue shift register, the Reticon SAD 1024.
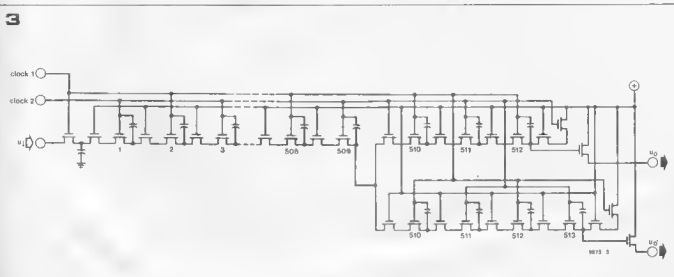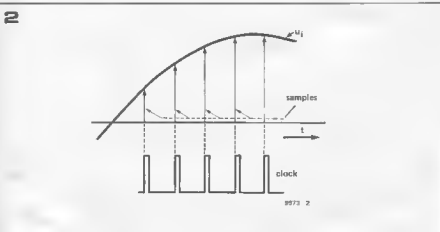
C1 charges to the instantaneous value of the input signal, i.e. the input signal is sampled.

When S1 opens and S2 closes then some of the charge on C1 is transferred to C2 via S2a. When S1 again closes C1 takes a new sample of the signal, whilst C2 transfers some charge to C3 via S1b and so on.

In this way a number of samples are taken at various points along the input waveform as shown in figure 2, and these are stored in the shift register as a sequence of charge packets. The actual operation of an analogue shift register is somewhat more complex than this simple explanation would

suggest, but the basic principle involved is that described above. In a practical shift register IC the switches are MOSFETS and the capacitors are also fabricated on the chip. An abridged internal circuit of an analogue shift register is given in figure 3.

The output signal of the shift register will appear as a series of pulses whose envelope follows that of the original input signal. The original signal can be recovered quite simply by lowpass filtering to remove the clock frequency component. The sampling theorem tells us that the clock frequency must be twice the maximum

signal frequency. In fact, it is fairly obvious that the clock frequency must be greater than the maximum signal frequency, otherwise it will be impossible to filter it out. Furthermore, if the limitations imposed by the sampling theorem are not observed, an objectionable effect known as 'foldover distortion' can occur. This is caused by the signal and clock frequencies interacting to produce spurious products within the audio spectrum, which can occur even if the clock frequency is above the audio range and therefore inaudible.

The delay time produced by a bucket-brigade memory is dependent upon two
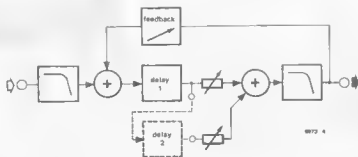
Figure 4. Principle of a reverberation unit. The input signal is delayed be feeding it through the bucket-brigade memory and a portion of the delayed signal is fed back to the input. Lowpass filters at the input and output limit the bandwidth of the signal to avoid foldover distortion and remove clock frequency components.

Figure 5. A more detailed block diagram of the analogue reverb unit.

Figure 6. Complete circuit of the reverb unit. Extensive use is made of FET op-amps.
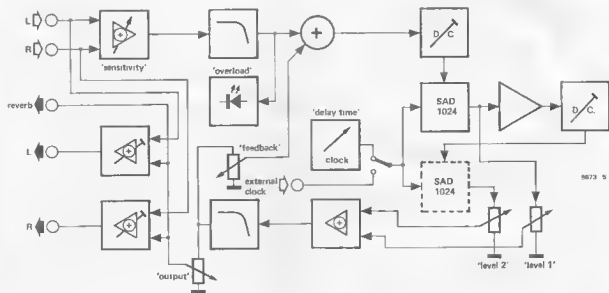
factors, the number of stages in the memory and the clock frequency. Since the signal is shifted through two stages for each clock pulse it is apparent

that the delay time is $t = \dfrac{n}{2 \cdot f_c}$.

where n is the number of stages and f_c is the clock frequency.

Since the clock frequency must be at least twice the maximum signal frequency, it follows that the maximum

delay obtainable is $t = \dfrac{n}{4 \cdot f_s (max)}$.

In other words a compromise must be adopted between delay time and signal bandwidth. Increase one and the other must be decreased. In practice this means that the bandwidth of the reverb signal must be limited to somewhat less than the full audio bandwidth, if adequate delay times are to be obtained with reasonably short shift

registers. This means band limiting the input signal using a lowpass filter at the input of the memory to prevent foldover distortion.

Reverberation unit

The basis of the reverberation unit is shown in figure 4. The input signal is fed through a lowpass filter and thence through the bucket-brigade memory. An attenuated portion of the delayed signal is fed back and summed with the input signal. Each time the delayed signal goes round the loop it is attenuated further and so gradually decays, thus giving rise to the characteristic reverberation effect. For longer delays a second memory may be added as an optional extra.

The SAD 1024

The analogue shift register chosen for the reverberation unit is the Reticon

SAD 1024. This IC contains two, completely independent 512-stage bucket-brigade memories, which may be used separately or together. The compromise chosen between the delay time and maximum signal frequency was 100 ms and 2.5 kHz. With a 1024 stage memory and a bandwidth of 2.5 kHz it should theoretically be possible to achieve a delay of 102.4 ms at a clock frequency of 5 kHz. However, in practice the clock frequency must be set slightly higher than that demanded by the sampling theorem in order that it can be filtered out without attenuating the highest signal frequency. Even so, the output lowpass filter must have an exceedingly sharp cutoff and an astounding 48 dB/octave is used in this design.

A signal bandwidth of 2.5 kHz may seem rather small, but in fact it is quite adequate for a convincing reverb effect. For those who require a longer delay

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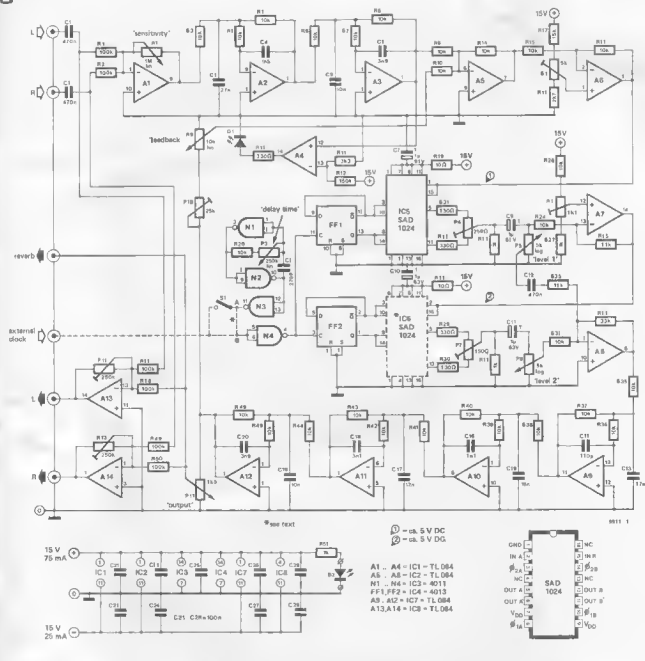
time or a wider bandwidth there is the option of adding a second SAD 1024 and/or raising the clock frequency.

Since the SAD 1024 has two sections of 512 stages the question arises of how to connect them to give a 1024 stage delay. It is, of course, possible to connect them in cascade, but this would not give the optimum signal-to-noise ratio and distortion, since passing through an additional 512 stages would further degrade the signal. The question of clock suppression also arises. With a clock frequency only slightly more than twice the maximum signal frequency it is not possible completely to filter out the clock component, even with a very steep cut filter. The solution to both these problems is to operate the two sections of the memory in 'parallel multiplex'. This means feeding the input signal to the parallel-connected inputs of both sections of the memory whilst

clocking the two sections in antiphase, the result being that the signal is sampled twice per clock pulse, alternately by each shift register. The outputs of the two memory sections are then mixed, with the result that the clock frequency components, which are in antiphase, tend to cancel.

The clock cancellation effect can, of course, be achieved by summing the outputs of the last two stages of a single memory section, since these too are in antiphase. This was done in the Phasing and Vibrato Unit, which used a bucket-brigade memory with only one section.

It may appear that parallel multiplexing provides only a 512-stage delay. This is indeed the case; however, since the signal is sampled twice per clock pulse the sampling rate is actually twice the clock frequency. The clock frequency can therefore be lowered to 2.5 kHz

whilst still achieving a 5 kHz sampling rate. This combination of a parallel-multiplexed 512-stage delay and 2.5 kHz clock of course gives the same delay as a 1024-stage memory (two cascaded 512-stage registers) and a 5 kHz clock.

Block diagram

A more detailed block diagram of the reverberation unit is given in figure 5, which is a stereo version of the reverberation unit. The left and right input signals are summed in a variable gain mixing amplifier and the resulting mono signal is fed to the input lowpass filter, which removes all frequencies above 2.5 kHz. The output signal from the filter is then fed to an offset circuit which sets the DC bias at the input of the SAD 1024. This is necessary as the SAD 1024 will only accept positive input signals, so the symmetrical AC

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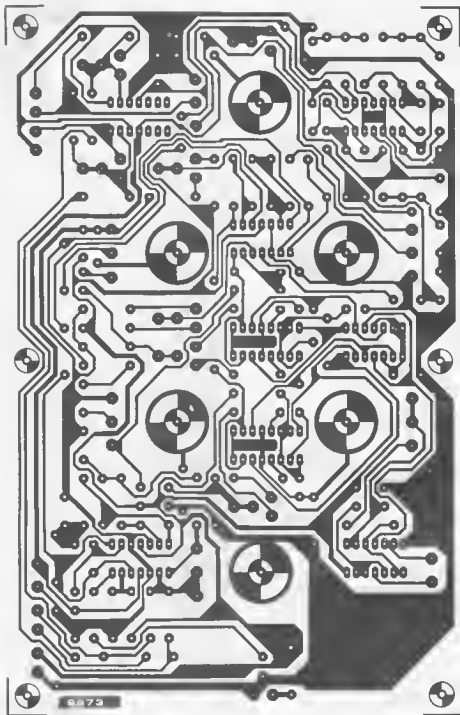
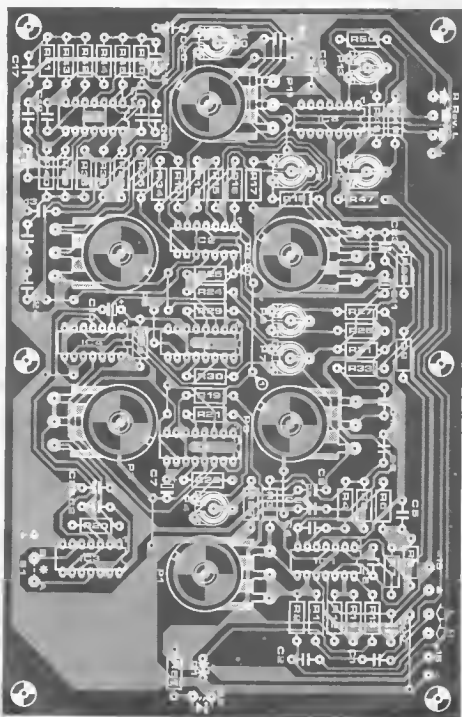output of the lowpass filter must be offset by adding a positive DC bias.
The signal is then fed through the first SAD 1024. If a second SAD 1024 is to be used then the output of the first IC is fed through an amplifier to make up for the attenuation of the first IC. The outputs of both SAD 1024s are equipped with level controls.
The output of the two memories are fed to a mixer and thence to the output lowpass filter. Part of the output signal from this filter is fed back to the input of the first SAD 1024 via a feedback level control which determines the reverberation time. The remainder of the signal is mixed equally with both

Figure 7. Printed circuit board and component layout for the reverberation unit (EPS 9973).

the original (undelayed) left and right signals, so that it appears as a mono image when fed to left and right loudspeakers. A separate output for the reverb signal only is provided. The output control varies the proportion of reverb signal in the output signal.
It may seem a little odd to add a mono reverb signal to a stereo signal, but in fact this simulates what happens in, say, a concert hall. Reverberation is the result of multiple reflections from the walls of the room and therefore conveys no directional information, i.e. it is mono. It appears more or less equally at both ears of the listener, superimposed on the direct left and right sounds.

There is nothing to be gained by having completely separate reverb channels for left- and right signals.

Complete circuit

Figure 6 is the complete circuit of the reverberation unit. The input signals are summed by op-amp A1, whose gain can be varied by P1. The output signal is then fed to the input lowpass filter comprising A2 and A3, which consists of two, cascaded, second-order Butterworth filters, giving a total slope of 24 dB/ octave above the cutoff frequency of 2.5 kHz. Since there is no clock frequency component to remove the slope of this filter is only half that of the output filter.

The output signal from A3 is summed with the feedback signal from the bucket-brigade memory by A5. The output of A3 is also fed to peak overload indicator A4. When the voltage on the non-inverting input of A4 exceeds that set on the inverting input by R11 and R12, the output of A4 will go positive and D1 will light.

The signal from the output of A5 is fed to A6, which is a unity-gain inverting amplifier with a variable DC offset at the non-inverting input. P2 is used to set the quiescent output voltage of A6 and hence the DC bias at the input of the first SAD 1024, IC5.

The output of IC5 is fed via P5, the 'level 1' control, to the input of A8, and thence to the output lowpass filter, A9 to A12. This consists of four, cascaded, second-order Butterworth filters and has a slope of 48 dB/octave.

The output of the filter is fed to P5, the reverb output control, which varies the proportion of reverberation in the final output signal. The reverb signal is mixed with the left and right direct signals in A13 and A14, the gain of these amplifiers being adjustable by P12 and P13 to suit the output level required by succeeding equipment.

If a second SAD 1024 (IC6) is included in the circuit then the output of IC5 is also fed to it via a second DC offset circuit, A7. This stage also has a gain of two to compensate for the attenuation introduced by IC5. The output of IC6 is fed, via the 'level 2' control P8, to the input of A8 and thence to the output filter.

The clock generator is an astable multivibrator based on two CMOS NAND gates N1 and N2. The clock output is buffered by the two remaining gates in the 4011 IC and is then fed to two flip-flops FF1 and FF2, whose Q and Q̄ outputs provide the two-phase clock for IC5 and IC6. The actual clock frequency fed to IC5 and IC6 is half the clock generator frequency since FF1 and FF2 function as divide-by-two counters.

Construction

A printed circuit board and component layout for the circuit are given in figure 7. The six main control potentiometers are mounted on the p.c.b. to simplify wiring. The whole assembly can then be mounted on spacers behind a fascia panel through which the potentiometer spindles protrude.

If the component values given in the circuit diagram are used then the filters will have a turnover frequency of 2.5 kHz. If a higher turnover frequency is required for greater signal bandwidth then table 1 should be referred to, which gives values for 5 kHz and 15 kHz turnover frequencies.

Adjustment and use

The circuit contains six control potentiometers and seven control presets. P12 and P13 simply set the gain of A13 and A14, and hence the output level of the reverb unit, to suit subsequent equipment.

The adjustment procedure for the remaining presets, and the operation of the controls, is as follows. P1 should be set so that D1 just lights on the loudest passages of the input signal. The optimum signal-to-noise ratio will then be obtained without overloading the circuit. P1 should not be used as a volume control, as overloading of the

circuit or a poor s/n ratio may result.

The feedback control, P9, should be turned fully anticlockwise and the output control, P11, fully clockwise, after which the reverb output should be fed to an amplifier and loudspeaker so that it is clearly audible. The level 1 control, P5, should be turned fully clockwise and the level 2 control, P8, anticlockwise and the clock frequency should be lowered until it becomes audible. The balance control, P4, should then be adjusted until the clock noise has been reduced to a minimum. This should occur with P4 approximately in its mid-position.

The D.C. bias voltage of the shift register may now be adjusted. A signal should be fed in of sufficient amplitude to cause D1 to light up and P2 should be adjusted until there is no audible distortion. Alternatively, if an oscilloscope is available, the input level may be increased until the output signal clips and P2 can then be adjusted so that the clipping is symmetrical.

If IC6 is included then the adjustment procedure for clock nulling and offset must be repeated for this IC, using P7 and P6 respectively. In this case the level 2 control, P8, should be fully clockwise, whilst the level 1 control, P5, should be anticlockwise.

Finally the feedback preset, P10, should be adjusted to give maximum decay time. This adjustment is carried out with P5, P8 if fitted, and P9, all turned fully clockwise. P10 is then adjusted so that the reverb signal decays gradually when the input signal ceases. If P10 is incorrectly set the system will be unstable and the reverb signal will swell to a cacophony of noise. This adjustment should be repeated at several settings of the delay time control, P3.

As already mentioned, the reverb unit has three outputs: left channel plus reverb, right channel plus reverb and reverb only. The unit can be used with an existing stereo system simply by using the tape socket(s) of the amplifier. The signal to the reverb unit is taken from the tape output and the signal from the reverb unit is taken to the tape input.

Alternatively, the left and right outputs of the reverb unit need not be used. Instead the reverb output may be fed to a separate amplifier and loudspeaker. This gives a more spacious effect to the sound.

It is important to note that the clock frequency will become audible if it is set too low. If the 5 kHz or 15 kHz bandwidth option is adopted then this will obviously occur at higher settings of P3. P3 may be fitted with a frequency-calibrated scale to reduce the chance of setting the frequency too low. Alternatively, its range may be reduced by connecting 'padding' resistors in parallel with P3. These should be chosen by experiment such that with P3 set to its maximum resistance the clock frequency is inaudible.

Table 1		
turnover frequency (−3 dB)	5 kHz	15 kHz
C3	12 n	3n9
C4	820 p	270 p
C5	5n6	1n8
C6	1n8	680 p
C8	150 p	47 p
C13	27 n	8n2
C14	390 p	120 p
C15	8n2	2n7
C16	1n8	390 p
C17	5n6	1n8
C18	1n8	560 p
C19	4n7	1n5
C20	2n2	680 p

M

Wherever possible in Elektor circuits, transistors and diodes are simply marked 'TUP' (Transistor, Universal PNP), 'TUN' (Transistor, Universal NPN), 'DUG' (Diode, Universal Germanium) or 'DUS' (Diode, Universal Silicon). This indicates that a large group of similar devices can be used, provided they meet the minimum specifications listed in tables 1a and 1b.

TUPTUNDUGDUS

Table 1a. Minimum specifications for TUP and TUN.

	type	U_{ceo} max	I_c max	h_{fe} min	P_{tot} max	f_T min
TUN	NPN	20 V	100 mA	100	100 mW	100 MHz
TUP	PNP	20 V	100 mA	100	100 mW	100 MHz

Table 1b. Minimum specifications for DUS and DUG.

	type	U_R max	I_F max	I_R max	P_{tot} max	C_D max
DUS	Si	25 V	100 mA	1 µA	250 mW	5 pF
DUG	Ge	20 V	35 mA	100 µA	250 mW	10 pF

Table 2. Various transistor types that meet the TUN specifications.

TUN		
BC 107	BC 208	BC 384
BC 108	BC 209	BC 407
BC 109	BC 237	BC 408
BC 147	BC 238	BC 409
BC 148	BC 239	BC 413
BC 149	BC 317	BC 414
BC 171	BC 318	BC 547
BC 172	BC 319	BC 548
BC 173	BC 347	BC 549
BC 182	BC 348	BC 582
BC 183	BC 349	BC 583
BC 184	BC 382	BC 584
BC 207	BC 383	

Table 3. Various transistor types that meet the TUP specifications.

TUP		
BC 157	BC 253	BC 352
BC 158	BC 261	BC 415
BC 177	BC 262	BC 416
BC 204	BC 263	BC 417
BC 205	BC 307	BC 418
BC 206	BC 308	BC 419
BC 212	BC 309	BC 512
BC 213	BC 320	BC 513
BC 214	BC 321	BC 514
BC 251	BC 322	BC 557
BC 252	BC 350	BC 558
	BC 351	BC 559

The letters after the type number denote the current gain

A α' (β, h_{fe}) = 125-260
B α' = 240-500
C α' = 450-900

Table 4. Various diodes that meet the DUS or DUG specifications.

DUS	DUG	
BA 127	BA 31B	OA 85
BA 217	BAX13	OA 91
BA 218	BAY61	OA 95
BA 221	1N914	AA 116
BA 222	1N4148	
BA 317		

Table 5. Minimum specifications for the BC107, -108, -109 and BC177, -178, -179 families (according to the Pro-Electron standard). Note that the BC179 does not necessarily meet the TUP specification (I_c,max = 50 mA).

	NPN	PNP
	BC 107	BC 177
	BC 108	BC 178
	BC 109	BC 179
U_{ceo} max	45 V	45 V
	20 V	25 V
	20 V	20 V
U_{eb0} max	6 V	5 V
	5 V	5 V
	5 V	5 V
I_c max	100 mA	100 mA
	100 mA	100 mA
	100 mA	50 mA
$P_{tot.}$ max	300 mW	300 mW
	300 mW	300 mW
	300 mW	300 mW
f_T min	150 MHz	130 MHz
	150 MHz	130 MHz
	150 MHz	130 MHz
F max	10 dB	10 dB
	10 dB	10 dB
	4 dB	4 dB

Table 6. Various equivalents for the BC107, -108, ... families. The data are those given by the Pro-Electron standard; individual manufacturers will sometimes give better specifications for their own products.

NPN	PNP	Case	Remarks
BC 107	BC 177		
BC 108	BC 178		
BC 109	BC 179		
BC 147	BC 157		P_{max} = 250 mW
BC 148	BC 158		
BC 149	BC 159		
BC 207	BC 204		
BC 208	BC 205		
BC 209	BC 206		
BC 237	BC 307		
BC 238	BC 308		
BC 239	BC 309		
BC 317	BC 320		I_{cmax} = 150 mA
BC 318	BC 321		
BC 319	BC 322		
BC 347	BC 350		
BC 348	BC 351		
BC 349	BC 352		
BC 407	BC 417		P_{max} = 250 mW
BC 408	BC 41B		
BC 409	BC 419		
BC 547	BC 557		P_{max} = 500 mW
BC 548	BC 558		
BC 549	BC 559		
BC 167	BC 257		169/259 I_{cmax} = 50 mA
BC 168	BC 258		
BC 169	BC 259		
BC 171	BC 251		251 ... 253 low noise
BC 172	BC 252		
BC 173	BC 253		
BC 182	BC 212		I_{cmax} = 200 mA
BC 183	BC 213		
BC 184	BC 214		
BC 582	BC 512		I_{cmax} = 200 mA
BC 583	BC 513		
BC 584	BC 514		
BC 414	BC 416		low noise
BC 414	BC 416		
BC 414	BC 416		
BC 413	BC 415		low noise
BC 413	BC 415		
BC 382			
BC 3B3			
BC 384			
BC 437			P_{max} = 220 mW
BC 438			
BC 439			
BC 467			P_{max} = 220 mW
BC 468			
BC 469			
	BC 261		low noise
	BC 262		
	BC 263		

MARKET

Bimdip

Capable of being used as both an insertion and withdrawal tool, the new BIMDIP accepts DIP IC packages with 4 to 18 leads. As an insertion tool it can pick up devices from either a carrier or direct from the bench and eject them into both pcb's and DIP sockets. Similarly it can be used as an extraction tool for withdrawing IC's, again from pcb's or DIP sockets. In all cases the metal jaws clamp over the lower part of the IC leads, not only minimizing strain but in the case of MOS type devices, shorting all leads together.

*Boss Industrial mouldings Ltd.,
Higgs Industrial Estate,
2 Herne Hill Road, London,
SE24 0AU, England.*

(840 M)

Switching regulator subsystem

A recent addition to Fairchild's family of linear integrated circuits is this monolithic regulator subsystem, part number μA 78S40. Included in one 16-pin package are all the usual active building blocks needed to assemble switching regulator systems. The device's already broad operational range can be extended, if required, by the addition of external transistors. The μA 78S40 consists of a temperature-compensated voltage reference, a duty-cycle controllable oscillator which incorporates an active current limit circuit, an error amplifier, a high-current high-voltage output switch, a power diode and an uncommitted operational amplifier.

Depending on the circuit configuration adopted the device can be used for the design of step-up, step-down or inverting switching regulators. Any of these circuits can be produced with the need for additional external components kept to a minimum. Output is adjustable from 1.3 V to 40 V without the use of external transistors. The device will operate from 2.5 V to 40 V input, the low end making it
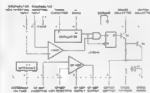
ideally suited to use in battery operated systems. Other device features include low standby current drain and 80 dB line and load regulation.

*Fairchild Camera & Instrument
230 High St, Potters Bar
Herts, EN6 5BU, England*

(853 M)

LED displays

Industrial Electronic Engineers, Inc. (IEE), a supplier of diversified, information display technologies, presents IEE-HERCULES Models 1784/85R .54" dual, alphanumeric LED displays with common cathode and right hand decimal point.

These models consist of two .54" high, red 14-segment characters combined in a single, compact package, which can display alpha and numeric characters plus some symbols. The unique and stackable feature allows designers variable display lengths in accordance with their needs. Composed of GaAsP emitting material, these solid state displays have a typical 600 μcd/segment luminous intensity at 20 mA/1.6 Vp. The 18, horizontal double DIP pins on .100" spacing are set up for multiplex drive for maximum pinout economy.

Models 1784/85R install simply in integral, multi-digit arrays with

IEE-ATLAS Display Mounting Hardware. These LEDs are in stock and are available through IEE stocking distributors.

*IEE, 7740 Lemona Avenue,
Van Nuys, CA 91405, USA*

(854 M)

Digital tong tester

The new Amprobe ACD-1 clamp-on digital volt/amp/ohmmeter from Havant Instruments Ltd is the latest addition to the company's growing range of test instrumentation. It provides instant readings of current of insulated or non-insulated cables with up to a 2 inch o.d. merely by clamping on the tongs. Test leads are provided for taking voltage and resistance readings. In all cases the correct range is selected automatically and the measured

value is then displayed in clear digits 0.43 in. high at the base of the handle.

The 3-digit display caters for AC amps, volts and ohms from 0.1 to 999 but accessories available allow the instrument to read currents from 10 mA to 6000 A. DC voltage can be measured using a multiplying factor. Accuracy for AC amps, volts and ohms is ± 2% based on a 50-60 Hz sinusoidal waveform. Protection for up to 1 minute is provided against 50% current or voltage overload. The Amprobe ACD-1 is powered by a 9 V battery with a life verified on test of 2700 cycles. The case, with a breakdown test voltag of 3000 V AC, is designed for comfortable one-hand operation using either hand. The weight of the instrument is only

15 oz and it is supplied complete with all-weather test leads, alligator clip adaptor, fused ohmmeter attachment and a rugged carrying case.

*Havant Instruments Ltd
Unit 3, Westfields
Horndean, Hants, England*

(857 M)

AM/FM radio system

The μA721 AM/FM radio receiver system from Fairchild is a versatile integrated circuit that meets the requirements of the majority of car, hi-fi, clock and portable radio systems. It is also suitable for other applications such as FM communications systems, CB radio receivers and wireless telephony receivers. A complete AM/FM radio designed around the μA721 requires the addition of only a few other active devices; two or
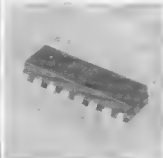
three transistors for the FM tuner front end and one audio power IC.

Versatility of use is ensured by virtue of the fact that the device has relatively independent sections that can be used in a variety of ways depending on the configuration of the external circuitry. Its various blocks consist of a bias circuit, AM oscillator mixer, amplifier 1 and amplifier 2 and the FM IF amplifier-limiter/detector.

Housed in a 16-pin moulded DIP package the small space occupied allows it to be built readily into todays consumer and industrial applications. It will operate over a wide supply voltage range of 3.5 V to 16 V. Quiescent current drain is low at 20 mA.

*Fairchild Camera & Instrument
230 High Street, Potters Bar,
Herts, EN6 5BU, England*

(856 M)

12 W audio power amp

The TDA 2030 is a new class B audio amplifier expressly designed for hi-fi equipment where ruggedness, reliability, compact size and economy are of prime importance.

It is assembled in the easily mounted 5-lead Pentawatt (R) package and it has a typical output power of 14 W (d = 0.5%) at ± 14 V/4 Ω. The guaranteed output power at ± 14 V is 12 W into a 4 Ω load and 8 W into an 8 Ω (DIN 45500).

The main features of the TDA 2030 are high supply voltage, high current, high supply voltage rejection, low noise and low distortion.

In addition, this device has built-in protection systems against load short circuit and excessive chip temperature.

The high degree of integration of the TDA 2030 has minimized the number of external components, reducing space consumption and costs.

SGS-ATES
Via C. Olivetti, 2
20041 Agrate Br., Milan, Italy

(859 M)

Tristate LED

A tristate light-emitting diode which produces red or green emission according to the polarity of the applied voltage is now available from Xciton. Known as the Xciton XC 5491, the device uses a back-to-back double-diode

configuration which produces red light (at 697 nm wavelength) or green light (565 nm) at an intensity of 1.8 mcd for both colours.

Forward voltage is typically 2.2 V at a forward current of 10 mA, dynamic resistance is typically 25 Ω and capacitance 100 pF. Maximum continuous forward current is 25 mA for both colours and peak pulse current for 1 µs (300 pulses per second) is 1 A. Maximum power dissipation is 100 mW, derated by 1.5 mW/degC from 25°C. Operating temperature is −55°C to +85°C. The XC 5491 is supplied in a T-1-3/4 package measuring 0.2 inch diameter x 0.34 inch height, and is supplied with wire-wrappable leads.

Ditronic Limited,
50-51 Burnt Mill, Elizabeth Way
Harlow, Essex, England

(860 M)

IC regulators with diodes

A new range of low cost IC voltage regulators in the 5 lead Pentawatt plastic package, the L-192 series ICs combine a high performance 250 mA regulator with AC rectifier diodes of 5 A surge and 85 V reverse ratings. Only a single smoothing capacitor is required to provide a stabilized, short circuit proof and thermal overload protected supply.
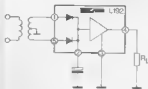
Output voltages are standard 5, 12 and 15 V (± 5%) with an AC input rating of up to 28 V rms. Load regulation is 0.5% for a 200 mA change in load current and line regulation is better than 65 dB with 10 V input ripple.

SGS-ATES, Via C. Olivetti, 2
20041 Agrate Brianza, Milan, Italy

(861 M)

Z80 based microcomputer

SGS-ATES has now made available a series of microcomputer cards based on its own Z80 CPU and peripherals. The card offers a choice of two available RAM sizes, 4 K or 16 K

bytes and sockets for 16 K ROM, PROM and EPROM. It can be expanded by means of additional cards up to a maximum of 64 K of memory.
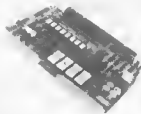
Particularly powerful and flexible are the interface circuits which include 4 bidirectional I/O ports (2 Z80 PIO), a communication interface (USART) compatible with the RS 232 and 20 mA current loop standards and a double interface for low cost audio cassette recorders. Complete ROM resident software support, including a 1 K Monitor and a complete 8 K operating system, consisting of Editor, Assembler and Debugger, is also available.

SGS-ATES, Via C. Olivetti, 2
20041 Agrate Brianza,
Milan, Italy

(862M)

Radiocode clock

The Radiocode Clock, which automatically receives a 60 kHz transmission from Rugby MSF and decodes all of the time and date information, is claimed to provide the most authoritative portable and self-contained time source available.
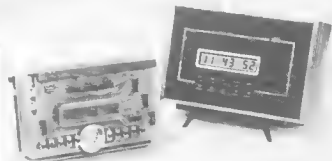
A liquid crystal display shows either hours, minutes and seconds or day, month and year. Because the unit has a crystal backup, the clock will continue to operate even if the transmission stops during a maintenance period. The instrument can also be supplied with an alarm/timer module which enables the clock to control other equipment at certain times for precise periods. No initial or subsequent adjustments are required because the clock sets itself and accounts for leap seconds, leap years and BST. Internal standard batteries allow a years continuous use even with a built-in sounder operating. The estimated range is around 1000 miles, and the receiver delay, after compensation, is quoted as 5 ms. For use on the Continent, the clock has an add-on-hour facility. Alternatively, a modified version can be supplied which receives a signal from the DCF 77 transmitter at Mainflingen, West Germany. This allows the clock to be used in eastern Europe where the MSF transmission may be weak.

Various optional outputs are also available, enabling the connection of a complementary record/replay unit. This interface allows the 'time' to be recorded on one track of a conventional tape machine. On replay the recorded time is displayed by the clock. If very low frequency signals or d.c. levels are to be recorded on the other tracks of the tape recorder, an additional f.m. interface unit can be supplied. For applications which require an accuracy of around 1 µs, details of an NPL system using a Radiocode clock and television sync pulses can be supplied. Prices for standard clocks range from £ 275 to £ 365 (plus VAT).

Circuit Services,
6 Elmbridge Drive, Ruislip,
Middx. HA4 7XB, England

(863 M)