ELECTRONICS TODAY

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MARCH 1976

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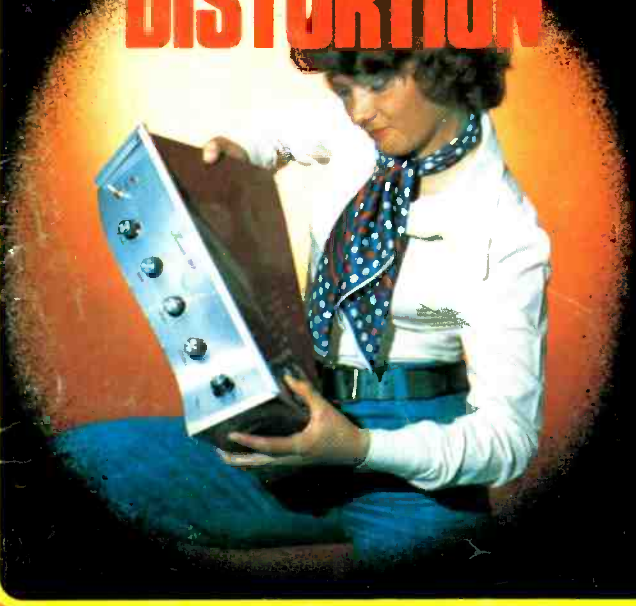

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AUDIO LEVEL METER

INDEX APRIL 1972–MARCH 1976

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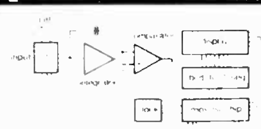
CD4528AE	£1.30 *	£1.08 *	87p *

3½ DECADE DVM I.C.



This state-of-the-art MOS LSI chip contains all the logic necessary for a 3½ decade, dual slope integrating, automatic polarity detecting DVM. Supplied with free data and circuit booklet.

OUR PRICE ONLY £6.59 *

(Data and Circuit Booklets alone 20p.)

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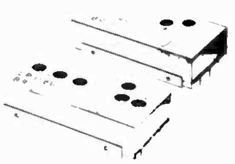

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LP1185	I.F. Strip	£6.95
LP1400	Multiplex decoder	£9.02

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Pin Sockets, pins in strips of 100. Just snip off what you need. 65p per strip *

Dual in line

8 pin 13p	24 pin 20p	TO5
14 pin 15p	28 pin 30p	8 pin 31p
16 pin 15p	36 pin 39p	10 pin 35p

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Our 8 Page A4 Audio I.C. Booklet is supplied FREE with purchases of Linear I.C.'s worth £1. or more (35p. if sold alone) Contains circuits and pin connections

- 15 Amplifiers 250 mW to 20 watts
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- 1 Tape Pre-Amplifier
- 1 Power Driver
- 1 Instrument Amplifier (Bifet)
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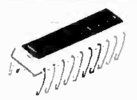
- 1 D.C. Controlled Gain Control
- 1 Micro-mini radio (2 I.C's)

PHOTO-DARLINGTON

Vceo 25v 2N5777
Vcbo 25v
Vebo 8v
I1 250 mA
Pd 200 mW
Hfe 2500

35p

LINE-O-LIGHT



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74 TTL

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7404	16p*	13p *	11p *
7408	16p*	13p*	11p *
7409	16p*	13p*	11p *
7410	16p*	13p*	11p *
7413	29p*	24p*	20p *
7417	27p*	22½p*	20p *
7420	16p*	13p*	11p *
7427	27p*	22½p*	18p *
7430	16p*	13p *	11p *
7432	27p*	22½p*	18p *
7437	27p*	22½p*	18p *
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7442	65p*	55p*	43p *
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7447	81p*	75p*	65p *
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74121	34p*	28p*	23p *
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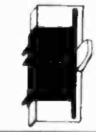

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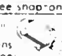

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Editor
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COIL FORMERS & CORES

NORMAN ¼" Cores & Formers 0.07p
⅜" Cores & Formers 0.09p

SWITCHES

DP/DT Toggle 0.28p
SP/ST Toggle 0.22p

FUSES

20mm, 100mA, 200mA, 250mA, 500mA, 1A, 1.5A, 2A QUICK BLOW *0.05p each
Anti-surge 20mm only *0.8p each

VEROBOARDS*

VB 1 containing approx. 50sq. ins. various sizes all 0.1 matrix *0.60p
VB 2 containing approx. 50 sq. ins. various sizes all 0.15 matrix *0.60p

DECON-DALO 33PC Marker*

Etch resistant printed circuit marker pen. Full instructions supplied with each pen *0.92p

BATTERY HOLDERS*

Takes 6 H.P. 7's complete with terminal clip and lead *0.31p each

V.A.T.

ALL PRICES EXCLUDE V.A.T.

Please add 8% to all prices marked *. Remainder add 25%. Do NOT add V.A.T. to prices marked †.

INSTRUMENT CASES

(In 2 sections, Black Vinyl covered top and sides and bezel)

No.	Length	Width	Height	Price
BV1	8"	x 5¼"	x 2"	*£1.25
BV2	11"	x 6"	x 3"	*£1.62
BV3	6"	x 4¾"	x 1¾"	*£0.92
BV4	9"	x 5¼"	x 2½"	*£1.39

ALUMINIUM BOXES

No.	Length	Width	Height	
BA1	5¼"	x 2¼"	x 1½"	*0.45
BA2	4"	x 4"	x 1½"	*0.45
BA3	4"	x 2¼"	x 1½"	*0.45
BA4	5¼"	x 4"	x 1½"	*0.54
BA5	4"	x 2½"	x 2"	*0.45
BA6	3"	x 2"	x 1"	*0.39
BA7	7"	x 5"	x 2½"	*0.79
BA8	8"	x 6"	x 3"	*£1.02
BA9	6"	x 4"	x 2"	*0.65

(Each complete with ½" deep lids & screws)
PLEASE ADD 20p POSTAGE AND PACKING FOR EACH BOX

COMPONENT PAKS

Pak No.	Qty	Description	Price
C1	200	Resistors mixed values approx. count by weight	.60
C2	150	Capacitors mixed values approx. count by weight	.60
C3	50	Precision Resistors mixed values	.60
C4	75	⅛th width Resistors mixed preferred values	.60
C5	5	Pieces assorted Ferrite Rods	.60
C6	2	Tuning Gangs. MW/LW VHF	.60
C7	1	Pak Wire 50 metres assorted colours	*.60
C8	10	Reed Switches	*.60
C9	3	Micro Switches	*.60
C10	15	Assorted Pots & Pre-Sets	.60
C11	5	Jack Sockets 3 x 3.5m, 2 x standard Switch Type	.06
C12	30	Paper Condensers preferred types mixed values	.60
C13	20	Electrolytics Trans. types	.60
C14	1	Pack assorted Hardware-Nuts/Bolts,Grommets, etc.	*.60
C15	5	Mains Slide Switches, 2 Amp	.60
C16	20	Assorted Tag Strips Panels	.60
C17	10	Assorted Control Knobs	.60
C18	4	Rotary Wave Change Switches	.60
C19	2	Relays 6-24V Operating	*.06
C20	20	Sheets Copper Laminate. approx. 200 sq. ins.	*.60

Please add 20p post and packing on all component packs, plus a further 10p on pack nos. C1, C2, C19 & C20.

AVDEL BOND

SOLVE THOSE STICKY PROBLEMS! *with*

CYANOACRYLATE G2 ADHESIVE

The wonder bond which works in seconds. Bonds plastic, rubber, transistors, components, permanently, immediately!

OUR PRICE ONLY 69p*
for 2 gm phial

CABLES		Per Metre
CP 1	Single lapped screen	*0.08
CP 2	Twin Common Screen	*0.11
CP 3	Stereo Screened	*0.12
CP 4	Four Core Common Screen	*0.21
CP 5	Four Core individually screened	*0.28
CP 6	Microphone Fully Braided Cable	*0.11
CP 7	Three Core Mains Cable	*0.11
CP 8	Twin Oval Mains Cable	*0.08
CP 9	Speaker Cable	*0.06
CP 10	Low Loss Co-Axial	*0.14

P&P

Postage and Packing add 25p unless otherwise shown. Add extra for airmail. Minimum order £1.00

ANTEX EQUIPMENT

SOLDERING IRONS

x25. 25 watt	*£2.45
Model G. 18 watt	*£2.70
CCN 240. 15 watt	*£2.90
SK2.Soldering Kit	*£3.90

BITS AND ELEMENTS*

Bit No.		
102 for model CN240	3/32"	*42p
104 for model CN240	3/16"	*42p
1100 for model CCN240	3/32"	*42p
1101 for model CCN240	3/8"	*42p
1102 for model CCN240	¼"	*42p
1020 for model G240	3/32"	*42p
1021 for model G240	1/8"	*42p
1022 for model X25	3/16"	*42p
50 for model X25	3/32"	*44p
51 for model X25	1/8"	*44p
52 for model X25	3/16"	*44p

ELEMENTS*

Model ECN	£1.10*
Model EG 240	£1.35*
Model ECCN 240	£1.55*
Model EX 25	£1.20*

SOLDERING IRON STAND

ST3 Suitable for all models *£1.10
Antex heat shunt *10p

PLUGS		PRICE
PS 1	D.I.N. 2 Pin (Speaker)	0.10
PS 2	D.I.N. 3 Pin	0.11
PS 3	D.I.N. 4 Pin	0.14
PS 4	D.I.N. 5 Pin 180°	0.15
PS 5	D.I.N. 5 Pin 240°	0.15
PS 6	D.I.N. 6 Pin	0.16
PS 7	D.I.N. 7 Pin	0.17
PS 8	Jack 2.5mm Screened	0.17
PS 9	Jack 3.5mm Plastic	0.11
PS 10	Jack 3.5mm Screened	0.17
PS 11	Jack ¼" Plastic	0.14
PS 12	Jack ¼" Screened	0.20
PS 13	Jack Stereo Screened	0.33
PS 14	Phono	0.09
PS 15	Car Aerial	0.14
PS 16	Co-Axial	0.14

INLINE SOCKETS		
PS 21	D.I.N. 2 Pin (Speaker)	0.13
PS 22	D.I.N. 3 Pin	0.19
PS 23	D.I.N. 5 Pin 180°	0.19
PS 24	D.I.N. 5 Pin 240°	0.19
PS 25	Jack 2.5mm Plastic	0.15
PS 26	Jack 3.5mm Plastic	0.15
PS 27	Jack ¼" Plastic	0.28
PS 28	Jack ¼" Screened	0.32
PS 29	Jack Stereo Plastic	0.28
PS 30	Jack Stereo Screened	0.35
PS 31	Phono Screened	0.17
PS 32	Car Aerial	0.20
PS 33	Co-Axial	0.20

SOCKETS		
PS 35	D.I.N. 2 Pin (Speaker)	0.07
PS 36	D.I.N. 3 Pin	0.09
PS 37	D.I.N. 5 Pin 180°	0.09
PS 38	D.I.N. 5 Pin 240°	0.10
PS 39	Jack 2.5mm Switched	0.11
PS 40	Jack 3.5mm Switched	0.11
PS 41	Jack ¼" Switched	0.19
PS 42	Jack Stereo Switched	0.28
PS 43	Phono Single	0.07
PS 44	Phono Double	0.09
PS 46	Co-Axial Surface	0.09
PS 47	Co-Axial Flush	0.19

P.C.B. KITS & PENS

PROFESSIONAL D.I.Y. PRINTED CIRCUIT KIT

Containing 6 sheets of 6" x 4" single sided laminate, a generous supply of etchant powder, etching dish, etchant measure, tweezers, etch resistant marking pen, high quality pump drill with spares, cutting knife with spare blades, 6" metal ruler, plus full easy to follow instructions. *£7.80 per kit

Spare container of etchant for above, complete with instructions *70p

P.C.B. MARKING PENS

2 x quality market pens, specifically designed for drawing fine etchant resistant circuits on copper laminate. Complete with full instructions *£1.53 per pair

LOW-NOISE CASSETTES

C60	*33p
C90	*44p
C120	*56p

SLIDER PAK

Containing a range of slider pots.

SP1	6	mixed values sliders	0.60
SP2	6	47OR Lin. sliders	0.60
SP3	6	10K Lin. slider	0.60
SP4	6	22K Lin. sliders	0.60
SP5	6	47K Log. sliders	0.60
SP6	6	47K Lin. sliders	0.60

AUDIO LEADS

S221	5 pin DIN plug to 4 phono plugs length 1.5m	£1.08
S222	5 pin DIN plug to 5 pin DIN socket length 1.5m	.68p
S237	5 pin DIN plug to 5 pin DIN plug mirror image length 1.5m	£1.20
S238	2 pin DIN plug to 2 pin DIN socket length 5m	.68p
S268	5 pin DIN plug to 3 pin DIN plug 1 & 4 and 3 & 5 length 1.5m	£1.00
S270	2 pin DIN plug to 2 pin DIN socket length 10m	.80p
S271	5 pin DIN plug to 2 phone plugs connected to pins 3 & 5 length 1.5m	.70p
S275	5 pin DIN plug to 2 phono sockets connected to pins 3 & 5 length 23cm	.68p
S318	5 pin DIN socket to 2 phono plugs connected to pin 3 & 5 length 23cm	.68p
S404	Coiled stereo headphones extension cord extends to 7m	£1.40
S217	3 pin DIN plug to 3 pin DIN plug length 1.5m	.80p
S219	5 pin DIN plug to 5 pin DIN plug length 1.5m	.80p
S474	3.5mm Jack to 3.5mm Jack length 1.5m	.68p
S600	5 pin DIN plug to 3.5mm Jack connected to pins 3 & 5 length 1.5m	.80p
S700	5 pin DIN plug to 3.5 jack connected to pins 1 & 4 length 1.5m	.80p

CROSSOVER NETWORK

K4007 1/P Impedance 8 ohms.
(2-way) Insertion Loss 3dB.
Crossover Frequency 3 KHz.
PRICE £1.12

3-WAY-STEREO H/PHONE JUNCT BOX

H 1012 Enables change-over from loudspeaker to headphone listening. Also has a centre position for both outputs. PRICE £1.73

HANDBOOKS

TRANSISTOR DATA BOOK. DTE 2 227 Pages packed with information on European Transistors. Full specification including outlines. Price † £2.95 each

TRANSISTOR EQUIVALENT BOOK BPE 75 256 Pages of cross references and equivalents for European, American and Japanese Transistors. This is the most comprehensive equivalents book on the market today and has an introduction in 13 languages. Price † £2.68 each

DIODE EQUIVALENT BOOK DE 74 144 Pages of cross references and equivalents for European, American and Japanese Diodes, Zeners, Thyristors, Triacs, Diacs and L.E.D.'s. Price † £1.98 each

MULLARD DATA BOOK 1974/75 MDB 74 The latest edition of this popular handbook contains information on Semiconductors, Integrated Circuits, Television Picture Tubes, Valves, Capacitors and Resistors. Included in the 161 informative pages are 21 pages on Semiconductor Comparables. Price † £0.40 each

TTL DATA BOOK. DIC 75 Now complete Data book of 74 series TTL (7400-74132). Covering 13 main manufacturers in U.S.A. and Europe, this book gives full data as well as equivalents. Price † £3.74

THE WORLD'S BROADCASTING STATIONS WBS 75 An up-to-the-minute guide for those interested in DX-ing. Contains all the world's broadcasters on SW, MW and LW, as well as European FM/TV stations. Price † £3.56

A full range of technical books available on request.

INDICATORS *

3015F Minitron 7 Segment Indicator *£1.11

SIL G.P. DIODES

300 mW 40 PIV(min) SUB-MIN
FULLY TESTED
Ideal for Organ builders
30 for 50p, 100 for £1.50, 500 for £5
1,000 for £9

CERAMIC PAKS

Containing a range of miniature ceramic capacitors in mixed values, unrepeatable value.

MC1 24 ceramic capacitors: 22pF, 27pF, 33pF, 39pF, 47pF, 56pF, 68pF, and 82pF 0.60

MC2 24 ceramic capacitors: 100pF, 120pF, 150pF, 180pF, 220pF, 270pF, 330pF, and 390pF 0.60

MC3 24 ceramic capacitors: 470pF, 560pF, 680pF, 830pF, 1000pF, 1500pF, 2200pF, and 3300pF 0.60

MC4 21 ceramic capacitors: 4700pF, 6800pF, 01μF, .015μF, .022μF, .033μF and .047μF 0.60

MAMMOTH I.C. PAK *

APPROX. 200 PIECES

Assorted fall-out integrated circuits, including: Logic, 74 series, Linear, Audio and D.T.L. Many coded devices but some unmarked — you to identify.
OUR SPECIAL PRICE £1.00

WORLD SCOOP *

JUMBO SEMICONDUCTOR PAK

Transistors — Germ. and Silicon. Rectifiers — Diodes — Triacs — Thyristors — I.C.'s and Zeners. ALL NEW AND CODED.

APPROX. 100 PIECES

Offering the amateur a fantastic bargain PAK and an enormous saving — identification and data sheet in every pak. **ONLY £1.85 each**

BIB HI-FI ACCESSORIES

REF		PRICE
'D'	2 Hi-Fi Cable & Flex Tidy	34p
'J'	Tape Head Cleaning Kit	72p
'P'	Hi-Fi Cleaner	30p
	Model 9 Wire Stripper	£1.00
23	¼" Tape Editing Kit	£1.80
24	⅛" Cassette Editing Kit	£1.84
29A	Salvage Cassette	44p
32A	Stylus Balance	£1.28
33	Splicing Tape	38p
36A	Record & Stylus Cleaning Kit	32p
41	8 Track Cartridge Head Cleaner	88p
	Model 42 Groove-Kleen	£1.84
42/S	Roller & Brush for REF 42 & 2000	24p
43	Record Care Kit	£2.76
45	Auto Changer Groov-Kleen	98p
46	Spirit level	72p
48	Record Dust-Off	26p
52A	Cassette Tray	54p
53	Hi-Fi Stereo Test Cassette	£2.40
56	Hi-Fi Hints & Tips Book	48p
	Model 60 Groov-Kleen	£1.72
160/S	Replacement Brush Velvet Pad and Base Sticker for Model 60	24p
62	Cassette Head Cleaner (Liquid)	48p
71	Record 'Dust Off' (Displays of ten)	66p
71A	Record 'Dust Off' (Bubble Pack)	70p
75	Indexa Record	£1.50
76	Stylus Cleaner	36p
78	Cassette Fast Hand Winder	98p
83	Cassette Title & Container Labels (20 & 10)	36p

UNTESTED LIN PAK

Manufacturers "Fall Outs" which include Functional and part Functional Units. These are classed as 'out-of-spec' from the maker's very rigid specifications, but are ideal for learning about I.C.'s and experimental work.

PAK NO.		CONTENTS			PRICE
ULIC709	=	10	x	709	0.60
ULIC710	=	7	x	710	0.60
ULIC741	=	7	x	741	0.60
ULIC747	=	5	x	747	0.60
ULIC748	=	7	x	748	0.60

G.P. SWITCHING TRANS *

TO18 SIM. TO 2N706/8
BSY27/28/95A. All usable devices. No open and shorts. ALSO AVAILABLE IN PNP similar to 2N2906, BCY 70.
20 for 50p, 50 for £1, 100 for £1.80, 500 for £8, 1,000 for £14.

When ordering please state NPN or PNP

BI-PAK

PO BOX 6 WARE HERTS

BI-PAK

High quality modules for stereo, mono and other audio equipment.

NEW

PUSH-BUTTON STEREO FM TUNER

Fitted with Phase Lock-loop

OUR PRICE ONLY

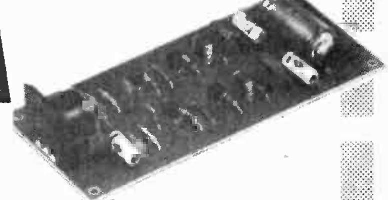
£19.95

The 450 Tuner provides instant program selection at the touch of a button ensuring accurate tuning of 4 pre-selected stations, any of which may be altered as often as you choose, by simply changing the settings of the pre-set controls.
Used with your existing audio equipment or with the BI-KITS **STEREO 30** or the **MK60** Kit etc. Alternatively the **PS12** can be used if no suitable supply is available, together with the Transformer **T461**.
The S450 is supplied fully built, tested and aligned. The unit is easily installed using the simple instructions supplied.

- ★ FET Input Stage
- ★ VARI-CAP diode tuning
- ★ Switched AFC
- ★ Multi turn pre-sets
- ★ LED Stereo Indicator

Typical Specification:
Sensibility 3μ volts
Stereo separation 30db
Supply required 20-30v at 90 Ma max.

MPA 30



Enjoy the quality of a magnetic cartridge with your existing ceramic equipment using the new M.P.A. 30, a high quality pre-amplifier enabling magnetic cartridges to be used where facilities exist for the use of ceramic cartridges only.
It is provided with a standard DIN input socket for ease of connection.
Full instructions supplied.

£2.65

VAT ADD 25%

POSTAGE & PACKING
Postage & Packing add 25p unless otherwise shown. Add extra for airmail. Min. £1.00

STEREO PRE-AMPLIFIER

PA 100

OUR PRICE
£13.50

A top quality stereo pre-amplifier and tone control unit. The six push-button selector switch provides a choice of inputs together with two really effective filters for high and low frequencies, plus tape output.

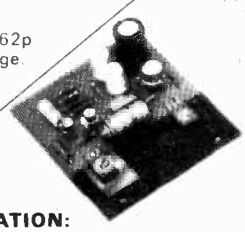
Frequency Response + 1dB 20Hz 20KHz Sensitivity of inputs
1. Tape Input 100mV into 100K ohms
2. Radio Tuner 100mV into 100K ohms
3. Magnetic P.U. 3mV into 50K ohms

P.U. Input equalises to R1AA curve with 1dB from 20Hz to 20KHz
Supply — 20-35V at 20mA
Dimensions — 299mm x 89mm x 35mm

MK. 60 AUDIO KIT: Comprising 2 x SPM80 1 x BTM80. 1 x PA100. 1 front panel and knobs. 1 Kit of parts to include on/off switch, neon indicator, stereo headphone sockets plus instruction booklet. **COMPLETE PRICE £27.55.** plus 62p postage.

TEAK 60 AUDIO KIT: Comprising Teak veneered cabinet size 16¾"x11½"x3¾", other parts include aluminium chassis, heatsink and front panel bracket plus back panel and appropriate sockets etc. **KIT PRICE £9.20** plus 62p postage.

AL10-20-30 AUDIO AMPLIFIER MODULES



The AL10, AL20 and AL30 units are similar in their appearance and in their general specification. However, careful selection of the plastic power devices has resulted in a range of output powers from 3 to 10 watts R.M.S.
The versatility of their design makes them ideal for use in record players, tape recorders, stereo amplifiers and cassette and car-tridge tape players in the home.

SPECIFICATION:
- **Harmonic Distortion Po=3 watts f=1KHz 02.5%**
- **Load Impedance 8-16ohm**
- **Size: 75mm x 63mm x 25mm**
- **Frequency response ±3dB Po=2 watts 50Hz-25Hz**
- **Sensitivity for Rated O/P — Vs=25v. RL=8ohm f=1KHz 75mV. RMS**

AL10 3w R.M.S. £2.30 | **AL20 5w R.M.S. £2.65** | **AL30 10w R.M.S. £2.95**

STEREO 30

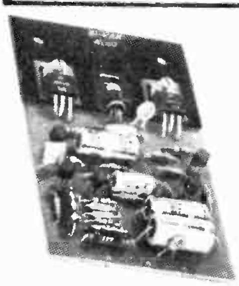
COMPLETE AUDIO CHASSIS
7+7 WATTS R.M.S.

P & P 45p
£15.75

The Stereo 30 comprises a complete stereo pre-amplifier, power amplifiers and power supply. This, with only the addition of a transformer or overwind will produce a high quality audio unit suitable for use with a wide range of inputs i.e. high quality ceramic pick-up, stereo tuner, stereo tape deck etc. Simple to install, capable of producing really first class results, this unit is supplied with full instructions, black front panel knobs, main switch, fuse and fuse holder and universal mounting brackets enabling it to be installed in a record plinth, cabinets of your own construction or the cabinet available. Ideal for the beginner or the advanced constructor who requires Hi-Fi performance with a minimum of installation difficulty (can be installed in 30 mins).

TRANSFORMER £2.45 plus 62p p &p
TEAK CASE £3.65 plus 62p p & p.

AL 60 25 Watts (RMS)



★ Max Heat Sink temp 90C. ★ Frequency response 20Hz to 100KHz ★ Distortion better than 0.1 at 1KHz ★ Supply voltage 15-50v ★ Thermal Feedback ★ Latest Design Improvements ★ Load — 3,4,5, or 16 ohms ★ Signal to noise ratio 80db ★ Overall size 63mm. 105mm. 13mm.

Especially designed to a strict specification. Only the finest components have been used and the latest solid-state circuitry incorporated in this powerful little amplifier which should satisfy the most critical A.F. enthusiast.

£3.95

NEW

PA12

NEW PA12 Stereo Pre-Amplifier completely redesigned for use with AL10/20/30 Amplifier Modules. Features include on/off volume, Balance, Bass and Treble controls. Complete with tape output.

Frequency Response 20Hz-20KHz (—3dB). Bass and Treble range 12dB. Input Impedence 1 meg ohm. Input Sensitivity 300mV. Supply requirements 24V .5mA. Size 152mm x 84mm x 33mm.

£6.50

PS12

Power supply for **AL10/20/30, PA12, SA450** etc.

Input voltage 15-20v A.C. Output voltage 22-30v D.C.
Output current 800 mA Max. Size 60mm x 43mm x 26mm.

OUR PRICE £1.20
Transformer T538 £2.30

Stabilised Power Supply Type SPM80

SPM80 is especially designed to power 2 of the AL60 Amplifiers, up to 15 watts (R.M.S.) per channel simultaneously. With the addition of the Mains Transformer **BMT80,** the unit will provide outputs of up to 1.5A at 35V. Size: 63mm. 105mm. 30mm. Incorporating short circuit protection.
Transformer BMT80
£2.60 + 62p postage

£3.00

BI-PAK

P.O. BOX 6, WARE, HERTS.

news digest

MAGNETIC CARD PROGRAMMABLE CALCULATOR

Slough, December 10, 1975. . . . . A handheld magnetic card programmable calculator with a recommended retail price of £250.00 (inc.VAT) has been introduced by Texas Instruments Limited, European Calculator Division.

This new programmable calculator from TI offers 20 memory registers, 10 user-definable keys and has the capability to accept 224 program steps. The algebraic entry combined with 9 levels of parentheses allows problems to be entered exactly as they are written.

The SR-52 is able to store up to 224 program steps and numbers on a single magnetic card. 20 independent addressable memory registers permit addition, subtraction, multiplication and division of any displayed quantity with any memory register without affecting the keyboard calculation in progress.

With the calculator's 23 preprogrammed key functions, trig and log, powers and roots, factorials, reciprocals, three conversions and pi can be directly executed from the keyboard.

Ten different decision instructions and five user-set flags allow the user to program the SR-52 to make repetitive decisions and branch to appropriate program segments automatically without interruption.

Other features include 10 user-defined keys and 72 user-defined labels. While any portion of a program may be called by an absolute address number, these keys and labels permit prompt and unique identification and call-out of any pertinent program segment. Indirect addressing, decrement-and-skip on zero, and two levels of subroutines provide additional programming flexibility.

In addition to an operating guide and owner's manual, the SR-52 package includes a Basic Library manual, a card case, a library of pre-recorded programs on magnetic cards, diagnostic programs for testing the SR-52 and head cleaner.

Pre-recorded and diagnostic programs will be available through retail outlets and from Texas Instruments Limited together with other pre-recorded libraries for various professional disciplines.

The Texas Instruments SR-52 carries a one-year limited warranty, from the original purchase date -- under normal use and service -- against defective materials or workmanship. Any implied warranties are also limited in duration to the one-year period from the original purchase date.

IMPORTANT WARNING

Readers are asked to note the considerable danger involved in the production of homemade LDRs by cutting open silicon power transistors, as described on page 47 of Tech-Tips Special in the January issue. The danger lies in the fact that many power transistors contain Beryllium Oxide which is EXTREMELY TOXIC. The inhalation of Beryllia dust or fumes on a single exposure lasting mintutes or seconds can cause injury to the skin or mucous membrane severe enough to endanger life or cause serious injury. If particles of Beryllia enter the skin through cuts or grazes chronic ulceration is liable to result. Our thanks go to M. P. Hearne of Campbeltown for calling our attention to this.

SIGNETICS DROP CMOS

Our US correspondent advises us that Signetics have telexed all their US sales staff and distributors advising them that the company will be discontinuing their CMOS logic series.

Signetics' sales of CMOS products fell short of their projected target by a considerable margin. We understand that whilst the 1975 target was US$200 million actual sales are not expected to exceed US$120 million.

SOLAR CELL IS 20% EFFICIENT

A 20% efficient solar cell has been developed by Varian Associates.

According to the firm, its 8mm diameter cell produces 10 watts of electricity directly from a focused sunlight beam. Varian makes the cell from a gallium arsenide material developed by IBM. Although the cells are not yet in commercial production, Varian says it plans eventually to build a system of cells that can generate 1kW.

ELECTRONIC WATCH HAS CONTINUOUS DISPLAY

A challenge to LED and LCD watches has been launch by America's Optel Corporation. Optel have just released details of a prototype unit incorporating an 'electro-chromic' display claimed to be capable of showing a continuous readout without constant battery power.

The prototype unit is a three function device which shows hours and minutes continuously and seconds on demand. The corporation say that there are still 'some problems' with the design but claim a two to five year life for the prototype readout which they say has a 200 millisecond response time.

NEW HEATKITS

The January 1976 Heathkit catelogue is now available (send a 10p stamp for return postage) free from their HQ at Bristol Road, Gloucester, GL2 6EE.

The most interesting new product is a seven-function programmable stop-watch. For £74.50 you get the following–

(i) Ability to time two events simultaneously.

(ii) Timing of two minor events (two lap times, for instance) and displaying these, whilst timing the overall event.

(iii) Accumulation of a series of timings and display of the total; simultaneously the time elapsed from the first to the last timing is measured. Total journey time and total driving time, for example, can be measured.

(iv) Split facility – the display will show the time-so-far at the touch of a button at any instant in the course of timing an event.

(v) Separate timings can be taken for each "leg" of an event and then these can be totalled.

(vi) Alarm or "upcount": counts up to a programmed time and then gives an output pulse. The count can be interrupted at any point.

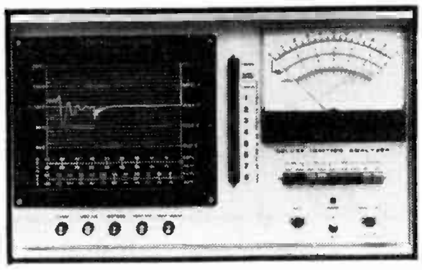
(vii) "Downcount" – the timer will count down from a programmed number and gives out a pulse at zero. Ideal for launching rockets, etc.

By combining the functions a time can be "learned" by transferring a displayed number to upcount or downcount.

Start/Stop jacks allow external triggering from photoelectric sensors, etc. Measuring range is up to 100hrs. and programming range up to 10hrs, with a resolution of 1/100th second. Although there is nothing this device can do which can't be done using the ETI Timing Modules (Dec 75 and Jan 76 issues) the Heathkit GB-1201E is packaged nicely in a hand-hold case (5 1/3" x 2 1/6" x 2").

Other new products in the catalogue are:

A car intrusion Alarm (£18.80).
An Electronic Doorbell (£31.50) which allows you to compose your own tune (within 16 beats and 1 octave). The tune can be changed (for a birthday party or for Christmas, etc) behind the front panel door.
A Digital Rev Counter (£31.80) gives a two digit readout.



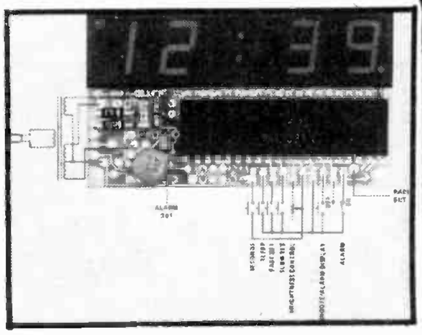
An Ignition Analyser (£280.00) with a 12 inch screen and professional facilities.
An SSB Transceiver (£490) for 80 thru 10m.
A CW Transceiver (£108.00) for 80 thru 15m.
A Synthesised 2 Meter FM Transceiver (£255.00).
A Hand-Held 2 Meter Transceiver (£144.00).
A 30MHz Frequency Counter/Timer (£78.00).

All prices include VAT and delivery in the UK and all refer to products in kit form.

DIGITAL CLOCK COSTS CUTS

A new clock module subsystem MA1001 requires only a transformer and switches to become a pre-tested electronic digital clock for use in radios, alarms, domestic and timing instrumentation, at approximately 50% of the cost of conventional similar designs.

The MA1001 is supplied with timing circuitry and a 4-digit LED display ready mounted on a 1.5" x 3" pcb. The LED's are ½" characters, and the ready-mounted MM5385N integrated MOS clock circuit (packaged 40-pin DIL) eliminates the need for separate bipolar segment-driver and digit-drive circuits. It also eliminates more than 30 resistors, as well as the RF filter capacitors sometimes required to attenuate RF interference



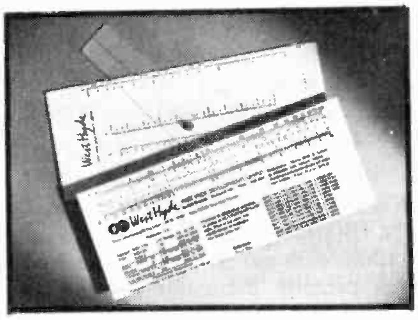
generated by multiplexed LED displays. The clock IC drives LED segments and digits directly, without multiplexing, which suppresses most RFI. The residual is eliminated by using a slow transition time in the output stages – 100μsecs as against the usual 1μsec.

The time-keeping frequency source can be either 50 or 60Hz, and 12-hour or 24-hour display options are available. Time is set through "Fast" and "Slow" scanning controls.

Features include alarm ON and PM indication, blinking colon, SLEEP and DOZE timers and variable brightness control capability. Alarm clock options include a transistor oscillator circuit for use with a low-cost earphone transducer. National Semiconductor UK Ltd., 19 Goldington Road, Bedford MK40 3LF.

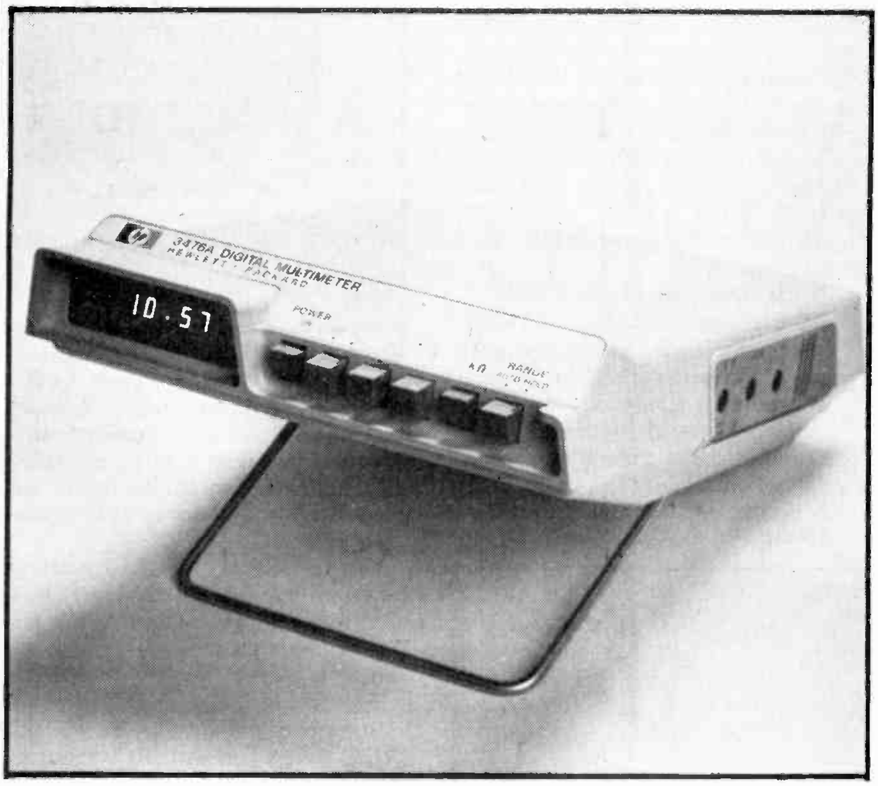
A SWINGING CALCULATOR (CURSOR?)

Now that West Hyde Developments are supplying a big range of carbon film, wirewound, vitreous wirewound resistors, and polystyrene capacitors, they have designed a 5" plastic Resistor Calculator with a built-in slide rule with which they are introducing these products to their customers. It has a



calculator on one side with a swinging cursor, which efficiently calculates power, current, voltage and resistance – any two known and two unknown, and the 5" slide rule is on the other side.

These are available from West Hyde Developments Ltd., Ryefield Crescent, Northwood, Middlesex, price 61p including P and P and VAT.



LOW-COST HP DMM

This new 3½ digit, five function, fully autoranging digital multimeter from Hewlett-Packard sells for only £144, and measures voltages from ±0.0001V DC and from 0.0003V to 700V rms AC. Resistance is measured from 0.001 kΩ to 1,000 kΩ. Current can be measured from 0.0001A to 1.1A DC and 0.0003A to 1.1A AC. Autozero, autopolarity and autoranging are built in.

Typical accuracy for DC voltage measurements is 0.5% DC current accuracy is 1.0%. One AC voltage ranges, frequency is specified to 10kHz, while AC current is to 5kHz. Accuracy of resistance measurements is within 0.6% on the three highest ranges and 0.4% on the two lower ranges. Open circuit voltage is less than 4V.

Input resistance on all voltages is 10MΩ with input capacitance of less than 30pF. The 3476 is protected to 1100V peak on all ranges. The fuse that protects the ohms function is rated 250V rms. The current function is fuse protected to 1.5A. No special fuses are required and they can be quickly replaced without dismantling the instrument.

A range hold feature is included that allows the instrument to be locked to any desired range. This feature is necessary, for example, when measuring diode resistance. It also makes repetitive measurement faster. The LED readout gives all voltage readings in volts, all resistance readings in kilohms and all current readings in amperes.

Model 3476A is AC line powered only; Model 3476B is AC line powered and also includes rechargeable nickel cadmium batteries. Model 3476A weighs 0.71kg and Model 3476B weighs 0.9kg. UK price is £175.00 for the 3476B. Hewlett-Packard Ltd., King Street Lane, Winnersh, Wokingham, Berks RG11 5AR.

HEATHKIT 'SCOPE COMPETITION

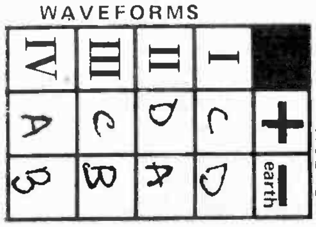
We now have the results of the competition in our December issue.

FIRST PRIZE of a 10-4540 scope goes to Mr. J. Humberstone,
57 Woodhead Road,
Sheffield 52 4TB.

SECOND PRIZE of a 10-4560 scope goes to Mr. A Eaves.
'Salway', Crowbrook Road,
Askett, Aylesbury,
Bucks, HP17 9LS.

THIRD PRIZE of another 10-4560 scope goes to Mr. M.L. Shirtcliffe,
50 Lupton Crescent,
Sheffield S87NA.

WAVEFORMS



'VERTICAL' PROBES

STILL TIME FOR HELPING HAND

The closing date for this award is March 31st, 1976, so if you haven't sent your ideas in yet there is still time. Details of the problems to be tackled are given in previous issues of ETI – choose any one of three specific problems facing deaf people. By the application of your knowledge of electronics you can help spare deaf people some of the hardships they face; and ETI, in conjunction with the RNID, will help develop your ideas.

If you need more information send a large SAE to ETI HELPING HAND, ETI Magazine, 36 Ebury Street, London SW1W 0LW.

ASTRONOMERS DETECT INDIVIDUAL PHOTONS

A digital television system for astronomy, developed by University of Arizona astronomers, is sensitive enough to detect individual photons of light coming into a telescope and record them for immediate playback.

As a result it is possible to see objects 100 times fainter than previously possible. The system is based on a special television tube – a silicon intensified target tube – that records 64 000 points of light simultaneously. The information then is sent in digital form to a computer, which removes the image's noise. Photographs or a spectrogram can be produced.

BIPOLAR PROMs AND ROMs BOOKLET

Intel have published a 36-page booklet which provides technical information on 13 different ROMs and 24 different PROM types they manufacture. All the devices in the booklet are of Schottky bipolar construction and erasable PROMs are not included.

The booklet incorporates a data sheet for each device, an equivalents chart and gives details of PROM programming equipment.

For every Intel PROM there is a pin and performance compatible PROM, which means that no circuit changes have to be made when a project is finalized and information in PROM is committed to mask-programmed ROM.

All Intel PROMs employ polysilicon fuses which coat the surrounding area in a aprotective oxide layer when they are 'blown'. With polysilicon fuses there is no danger of a fuse 'growing back' as has happened before with conventional metal fuses.

AMPLIFIER DISTORTION

—it may sometimes be advantageous!

In this article written specially for Electronics Today International, Gordon King explains transient intermodulation distortion and how it affects amplifier/speaker combinations.

OF RECENT MONTHS I have been endeavouring to establish meaningful correlation between the measured parameters of amplifiers and the listening room sound. This is a very difficult area of research because there are so many inter-related and variable factors involved; also because the net result is obviously a subjective impression rather than a meter reading! Nevertheless, a number of points of interest have emerged which merit discussion.

For starters, it would appear that certain parameters are advanced out of sheer 'specmanship' rather than on a basis of electro-acoustic requirements; in fact, it is sometimes possible to enhance the sound by deliberately *diminishing* a parameter. There would also appear to be a fairly important link between the subjective and the objective at the electrical interface between the amplifier and the loudspeaker. The sound field is certainly affected by the acoustical interface between the loudspeaker and the listening room.

Although this article is concerned primarily with electrical parameters, it is necessary to look at some of the acoustical aspects, too, for after all, it is the resulting sound that we listen to, not electrical signals.

The loudspeaker 'loads' both electrically into the amplifier and acoustically into the listening room; in other words, the amplifier is the electrical source for the loudspeaker and the loudspeaker the acoustical source for the listening room. It is well known of course that the output into or across a load, is influenced by the nature of the source and the load, this being applicable to acoustics as well as electrics. Sound *pressure,* in fact, is the acoustical analogue of voltage.

It is reasonable to conclude, therefore, that just as some

The acoustical load presented by a particular listening room may be more acceptable to some loudspeakers than others . .

electrical sources are more critical of loading than others, so are some loudspeakers. The acoustical load presented by a particular listening room may be more acceptable to some loudspeakers than others, which is one reason why a pair of loudspeakers which yield acceptable results in one room may audition less favourably in a different room. There is a case, therefore, for the loudspeaker and listening room to be measured in partnership. Although free-field (anechoic) pressure *versus* frequency plots are commonly adopted for optimising the design of loudspeakers, they are far from revealing how different loudspeakers will audition in different rooms.

It is neither difficult nor expensive for a hi-fi dealer or audiophile to measure loudspeaker/room combinations, and an inexpensive, though surprisingly accurate, method is based on the reproduction of a third-octave bands of pink noise. A 'linear' sound level meter at the normal listening position is then used to measure the sound pressure level at each band in turn over the range of 20 Hz-20 kHz, leading to the construction of a graph. This simple technique reveals eigentones and absorption effects quite dramatically, thereby indicating the adjustments required for improving the results.

Noise signal has the effect of automatically averaging the

It is neither difficult nor expensive to measure loudspeaker/room combinations . . .

sound in the listening room. Steady-state sinewave signal cannot be used. Pink noise, which is white noise with −3 dB/octave (or −10 db/decade) weighting, is used because it correlates more closely with the spectral distribution of music than unweighted noise, which is white noise. It is noteworthy that the voltage of white noise is proportional to the square root of the bandwidth, and contains frequency components of constant energy per unit bandwidth. The weighting thus endows pink noise with components of constant energy per octave bandwidth.

Bruel and Kjaer have produced a calibrated record of third-octave pink noise bands (Type QR2011) which, along with a B&K sound level meter, such as Type 2206, makes it possible to 'sweep' the loudspeaker and room. The resulting overall response needs to be interpreted with care, however,

Detailed measurements may place two amplifiers well into the accepted hi-fi category yet, in a common signal source, loudspeaker and listening room situation, one may produce a very fine sound and the other a distinctly fatiguing sound. Fact or fiction?
FACT!

Given an amplifier of top-flight-measured parameters and two pairs of similar style but different make, measurement-acclaimed loudspeakers, one pair to a critical ear can be far more acceptable than the other pair, yet if the amplifier is changed the other pair may then be preferred.
Fact or fiction?
FACT!

since at the higher frequencies the response fails to correlate to what we hear. This is because we judge a sound more on its starting transient rather than on its overall integrated energy. Nevertheless, low-frequency standing waves are brought to light, and modifications to loudspeaker positions, furniture positions and amplifier tone controls can often improve matters. It is hoped to publish an article in these pages later describing listening room optimisation.

The prime discrepancies between what the meter reads and what the ear discerns are related to non-linear effects both in the amplifier and loudspeaker; also, sometimes, to how the non-linearities interact electrically and acoustically. One problem in obtaining subjective correlation from a meter reading lies in the nature of the signals we are obliged to use for the measurements. Sine and square wave signals are useful, being component parts of music signal, but real music signal is much more complicated than both of these.

If it were possible to feed a loudspeaker with a perfect electrical representation of the originating sound, it is likely that the reproduction would be less palatable than that obtained by first passing the source signal through a distorting amplifier! A non-distorting signal would tend to emphasise the loudspeaker non-linearities in terms of crossmodulation of spatial, spectral and temporal co-ordinates. The reproduction would thus be modified by all the practical inadequacies of even the best of loudspeakers. Further modification would result from distortion on the electrical signal, and there is reason to believe that distortion on the signal prior to its application to the loudspeaker can, in certain circumstances, lead to more acceptable reproduction[1]. (So much for the straight piece of wire with gain theory — Ed)

The loudspeaker distortion co-ordinates would then themselves be crossmodulated by similar distortion

. . . a perfect representation of the original sound may be less palatable than by first passing it through a distorting amplifier!

co-ordinates on the signal before the loudspeaker, leading to an acoustical result more closely related to the originating sound as humanly judged, than if the loudspeaker distortion alone were present.

The nature of the distortion produced by both the amplifier and loudspeaker is thus critical, so that different types of distortion would give different subjective impressions, which is not uncommon in a system of units of different distortion types. For example, the distortion from a radio tuner can interact with the distortion of an amplifier to which it is connected in such a way that the instrument-indicated *change* in the net distortion from the partnership, as the FM tuning is adjusted within the passband of a tuned signal, lacks subjective correlation. A test condition can be established where a fall in meter distortion is accompanied by an obvious rise in subjective distortion!

It thus seems to be perfectly feasible that after establishing the most acceptable reproduction by selection

. . . a fall in measured distortion may be accompanied by a rise in subjective distortion.

of the amplifier and loudspeaker partnership, a detraction in subjective acceptability could well result by changing either the amplifier or loudspeaker. I believe that this is one reason why a hi-fi system whose amplifier parameters measure in advance of those of another may not necessarily audition any better. Indeed, it could be judged subjectively inferior!

It should be understood that we are now considering hi-fi at the top equipment level, where the amplifier distortion figure is, at least, *one* place to the right of the decimal point. Distortion from this class of amplifier seems to be falling more swiftly than the distortion from comparable class loudspeakers, which is not making it particularly easy to select suitable loudspeakers for the parameters of the amplifiers.

One amplifier from the Pioneer range comes to mind. In the lab this was found to have a very low level of distortion — one of the lowest ever measured — with harmonic components down to the distortion threshold of the measuring oscillator (0.002%) over the whole dynamic range, as measured with a wave analyser to read below the wideband noise power of the simple distortion factor meter. The intermodulation distortion was also correspondingly low and there was no crossover discontinuity; yet, in partnership with acclaimed well known loudspeakers, the amplifier was judged to be less subjectively acceptable than a counterpart of similar power, bandwidth, etc. but of much higher measured distortion.

Clearly, it is becoming more important to audition loudspeakers in partnership with the amplifier with which it is going to be used. The ultimate performance of the Pioneer, just exampled, was eventually realised only after careful loudspeaker selection.

Most of the important parameters of amplifiers are measured into resistive loads, which does not make much sense because no loudspeaker presents a purely resistive load to an amplifier. The load analogue of a loudspeaker is an impedance composed of resistance, capacitance and inductance, but the impedance is not very easily defined since it is affected to some extent by the electrical drive signal and, of course, by the impedance of the separate units and nature of the frequency dividers. Different designs of loudspeakers present different loads to amplifiers, and it is not feasible to construct load analogues corresponding to all loudspeakers for testing amplifiers! Neither is it good for the loudspeakers (nor the neighbours!) to use real loudspeakers at test loads. Thus for testing we are back to R with, perhaps, a dash of C and/or L.

There has been a tendency for designers to optimise in terms of the smallest rise-time into resistive loads, and rise-times as small as 2 µsec. can be seen in the specs. However, there can be a dramatic change in scene when the load is made reactive by the addition of C.

For example, in Fig. 1(a) the step function applied to the input of an amplifier was around 100 nsec. rise-time. The oscilloscope was set to 5 µsec./div., giving the display a rise-time, via the amplifier and *resistive* load, of about 2 to 3 µsec., corresponding to about 140 kHz −3 dB upper-frequency response.

The trace at (b) shows the same signal from the same amplifier, but this time with the load consisting of 8 ohms in parallel with 1 µF. Rings as bad as this can certainly affect the tonal quality of an amplifier, depending on their amplitude and period. I have suggested[2] a definition of settling-time as an important parameter of amplifiers when measured into a reactive load arranged either to evoke the worst condition (i.e., by selecting shunt C for the most prolonged ring, when R corresponds to the rated load) or to the load analogue of the loudspeaker which will be used with the amplifier. The definition of settling-time under ref. 2, is the time elapsed from the application of the step-function to the time that the amplifier enters and remains within a ± 5% error band, corresponding to $E_0 \pm \Delta E$, where E_0 is the final settling voltage. With the

AMPLIFIER DISTORTION

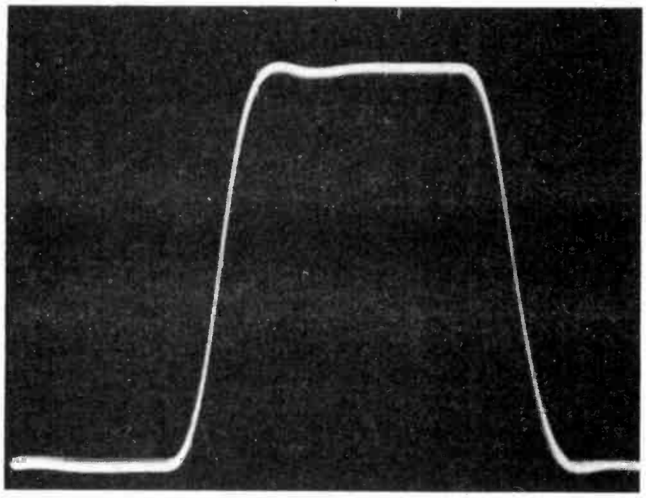


Fig. 1. Amplifier of small rise-time. (a) resistive rise-time about 4µsec.

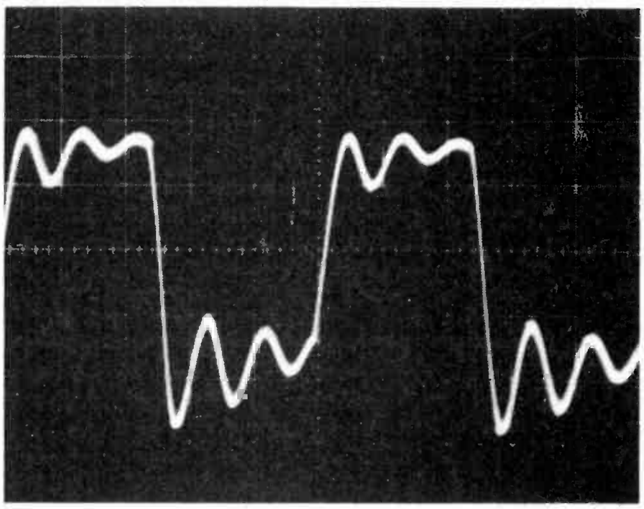


Fig. 1(b). Severe rings into reactance resulting in protracted 50µsec. settling-time.

settling-time referred to $\Delta E/E_0 \times 100$, it is in advance of 50 µsec. in Fig. 1(b), which is unacceptable for hi-fi.

Too long a settling-time, therefore, appears to be another valid reason why some amplifiers fail to audition as well as might be expected from distortion measurements alone; also another reason why a change in loudspeaker may modify the subjective result (i.e., by changing the electrical transient performance of the amplifier).

The oscillograms in Fig. 2 reveal a more acceptable state of affairs. That at (a) is based on a 3 µsec./div. sweep and corresponds to the resistive rise-time of a different amplifier of around 7 µsec. (50 kHz – 3 dB response), (b) shows what happens when the resistive load is shunted by capacitance. There are no rings in this case, just a mild kink at the leading and trailing corners of the squarewave, with the settling-time corresponding to about 10 µsec. (waveform on 20 µ sec./div. sweep). This shows the worst condition obtained with a shunt capacitance of 0.68 µF.

There appears to be a definite tendency for amplifiers designed for dramatically small resistive rise-times (some as small as 1.8 µsec. have been measured) to suffer prolonged settling-time as the result of severe rings into reactive loads, and hence tonal impairment when used with loudspeakers constituting critical load conditions.

The value of designing for very small rise-times and hence for extended small-signal high-frequency response is thus obvious. A rise-time of 1.8 µsec. implies that the amplifier is responsive well up to 200 kHz (the LW radio band!). Rise-time is related to the upper-frequency response by K/f, where K is a constant defined by the response characteristic of the amplifier or network, and f the upper-frequency where the response is 3 dB below the mid-spectrum response. When the upper-frequency roll-off approaches the so-called gaussian characteristic (i.e., when the –3 dB upper-frequency is approximately half the –12 dB frequency), K is close to 0.35; but it can range between 0.3 and 0.5, depending on the nature of the roll-off.

Of course the upper-frequency response needs to extend beyond audibility to accommodate the transient components of the music signal and thus to preserve the musical attack. However, it is difficult to argue in favour of a response much above 30 kHz, corresponding to around 12 µsec. rise-time. We have seen that an extended response might encourage rings and increased settling-time. There is also the possibility that it might encourage 'blocking' following fast transient signals. This is called transient intermodulation distortion.

Transient components of music *signals* rarely exceed about 16 µsec. owing to the limitations associated with the response and transfer characteristic of their sources. For example, a high quality FM transmission has an

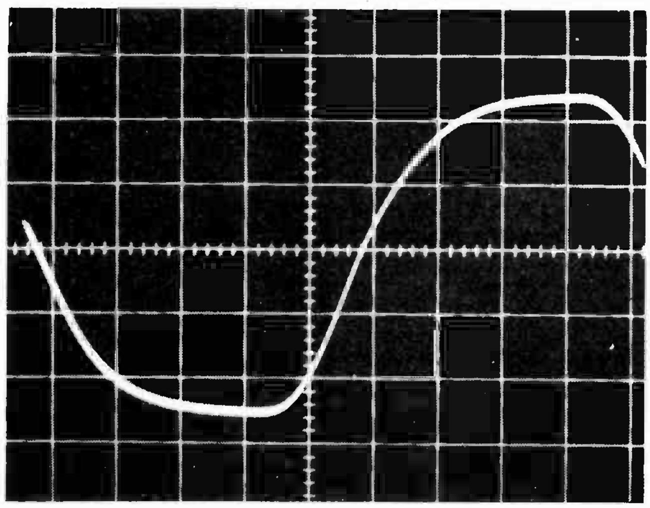


Fig. 2(a). Resistive rise-time about 7µsec.

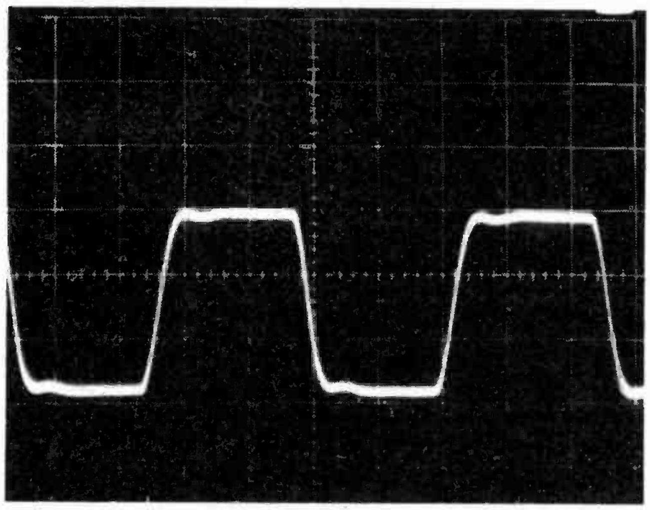


Fig. 2(b). Well controlled overshoot into reactance resulting in settling-time of 10µsec.

upper-frequency response limit of 15 kHz, with a swift fall into the 19 kHz pilot tone notch, thereby limiting the maximum equivalent rise-time to about 20 µsec. Few gramophone records carry high energy information much above 18 kHz, and the same applies to tape recordings, so even from sources of this kind the music transients are not likely to be much faster than 17 µsec. It is thus difficult to commend small-signal responses down to 2 µsec. or less and up to 200 kHz or more!

Transient components of music signals rarely exceed 16 µsec.

If one regards the source as a network of a given response, then, clearly, further limiting by a relatively slow amplifier response is undesirable. However, it must be remembered that the total rise-time (T_0) of two cascaded networks of rise-times T_1 and T_2 is equal to the vector sum (not to the simple sum), such that $T_o = \sqrt{T_1^2 + T_2^2}$. Thus the degree of response limiting of the source signal by the amplifier is relatively small — certainly not calling for a rise-time as small as 2 µsec.

Contemporary hi-fi amplifiers rely on negative feedback for extending and flattening the frequency response and for reducing non-linear distortion, particularly of the power amplifier section. The open-loop bandwidth of a power amplifier is dictated by the transistors which are available to drive the required audio power into reactive loudspeaker loads without veering too close to the secondary breakdown characteristic. This generally means that quite a lot of negative feedback must be applied to yield a viable closed-loop power response, and that lead and/or lag networks are necessary to maintain a reasonable stability margin. Unless the ratio of the response of the power amplifier in the open-loop mode to the response of the preamplifier is of unity or greater value, the amplifier is likely to exhibit transient intermodulation distortion — tid, for short.

In other words, the overall frequency response of the hi-fi amplifier should be dictated by the roll-off of the preamplifier section and not by the power amplifier section[3]. This, then, clearly places a limit on the small-signal response or rise-time of the preamplifier, beyond which it is subjectively imprudent to engineer.

The mechanics of tid can be described in the following way. The total input to a feedback amplifier consists of the sum of two signals, the source signal proper and the error signal fed back antiphase. If the source signal is a very fast transient and the error signal slightly delayed owing to a relatively slow power amplifier response, the input stage of the power amplifier will momentarily receive a signal of greater amplitude than it is designed to accommodate, and severe overloading may ensue. The transient may thus be distorted, and the sudden 'shock' to the input stage may result in this closing down for a brief period, followed by a relatively slow recovery due to the action of circuit time-constants, so that information immediately following the transient is lost.

. . . overall frequency response should be determined by the roll off of the pre-amplifier . . .

A method for the display of tid has been promulgated[4] and attempts have been made to measure it[5], but so far there is no accepted standard for the measurement.

AMPLIFIER DISTORTION

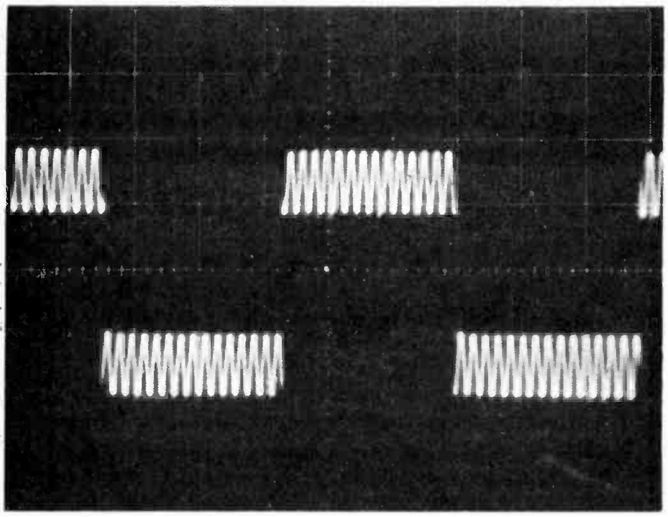


Fig. 3. *Transient intermodulation distortion. (a) Test signal of squarewave plus sinewave.*

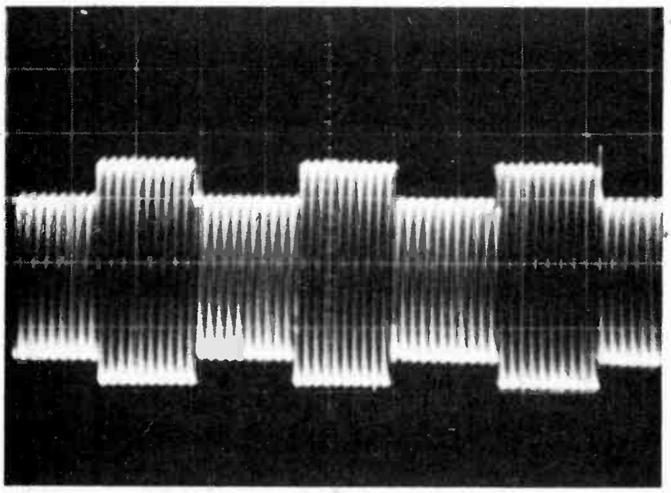


Fig. 3(b). Severe form of tid.

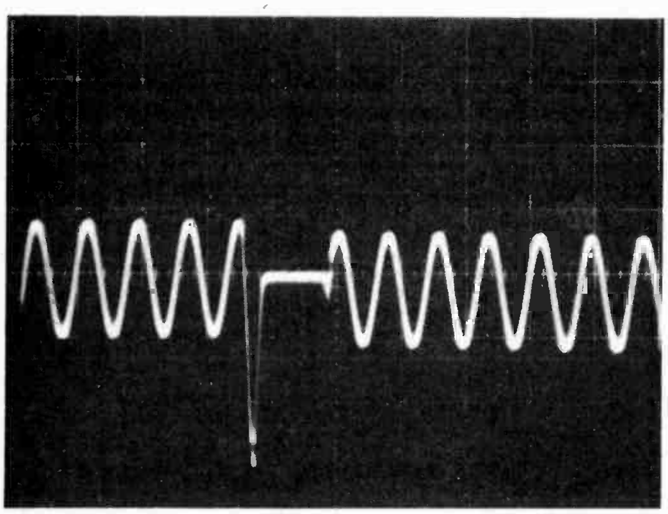


Fig. 3(c). Showing 'blocking' effect.

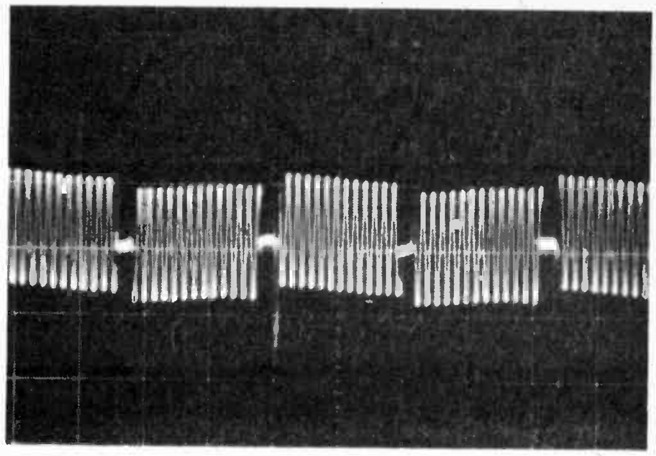


Fig. 3(d). 'Blocking' following each squarewave transient.

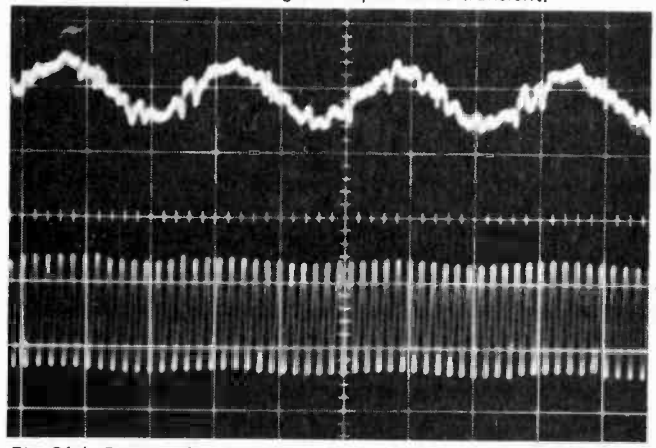


Fig. 3(e). Relative freedom from tid; bottom trace reconstituted sinewave; top trace distortion content of waveform.

The oscillograms in Fig. 3 may be of interest. Display (a) shows a test signal consisting of the addition of sinewave and squarewave signals. This composite signal is applied to the input of the amplifier under test, and at the output the squarewave component is cancelled out so that the sinewave component only is left for oscilloscope analysis. The squarewave is cancelled by applying to a bridge circuit an inverted *replica* of the squarewave component of the composite signal.

Display (b) shows a severe form of tid, giving asymmetrical sinewave components on the positive and negative going squarewave cycles. Display (c) shows the 'blocking' effect following transients. (d) is a similar display but with less expansion. Display (e) shows the sinewave components fairly well fitted together on the bottom trace, thereby indicating minimal tid, and the distortion signal on the top trace, after passing through a distortion factor meter.

Transient intermodulation distortion tends to affect the quality of the reproduction more towards the full power drive of the amplifier, and is emphasised by treble lift. It manifests as stridence and harshness on signal peaks. While there is a real possibility of tid being responsible for lack of objective/subjective correlation (for it does not appear as a parameter in specifications or reviews), it can only occur when the rate of rise of a signal transient at the *power amplifier* input is in advance of the response speed of the power amplifier in open-loop mode. It is thus encouraged by a very small rise-time which is not matched by the response speed of the power amplifier, indicated by a poor or mediocre slew-rate.

It is becoming apparent that an amplifier of very low distortion factor may not necessarily produce better sound than a counterpart which fares objectively less well. In fact, the latter may audition better! Here, then, is still another reason why an amplifier of very low measured distortion may fail to perform subjectively as one might expect.

A clue to this paradox is contained in the oscillograms shown in Fig. 4. A distortion factor meter responds to the average energy in the distortion signal, but the ear is more critical of signal peaks composed of high-order harmonics than lower-order harmonics of higher energy. Display (a) indicates relatively high energy third-harmonic distortion, which would produce a fairly substantial reading on a distortion factor meter compared with display (b), where the energy is small but the amplitude of the peaks large at the crossover points. The distortion factor of (a) was around 0.25% and of (b) a mere 0.05%, yet the amplifier responsible for (a) was more acceptable in the listening room than that responsible for (b), in spite of (a) being the much greater readout!

When comparing amplifiers in terms of distortion factor, it is essential to take account of the *nature* of the distortion, since the figures alone rarely provide adequate comparative information. Alternatively, attention should be directed to the intermodulation distortion, for with suitably high measuring frequencies, such as f_1 = 5 kHz and f_2 = 9 kHz (1:1 ratio), a relatively high $2f_1-f_2$ readout is a sure indication that the crossover distortion is not very well tamed, particularly when this order increases as the power of the amplifier is reduced. Crossover distortion is generally more troublesome at low amplifier power, than at high power, the converse of tid.

Another form of bad crossover distortion is shown by display (c), where the energy is also high. This corresponds to about 0.4%, which is barely hi-fi. A commendable result is shown by display (d), the distortion being virtually down to noise threshold with no crossover artifacts; this corresponds to 0.02%.

. . . it is essential to take account of the nature of distortion – figures alone are insufficient.

Other factors responsible for the auditioning differences of amplifiers include asymmetrical overload allied with abnormally long recovery time-constant and changing quiescent current under dynamic conditions. The damping factor, too, has a bearing on the amplifier/loudspeaker partnership, and it is desirable for the amplifier's source impedance to remain at a low value right down to infrabass.

In conclusion, it is hoped that this article has given a few interesting points over which to ponder. We are learning all the time, which is half the fun of hi-fi. ●

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5. Gordon J. King, *The Audio Handbook,* published by Newnes-Butterworths.

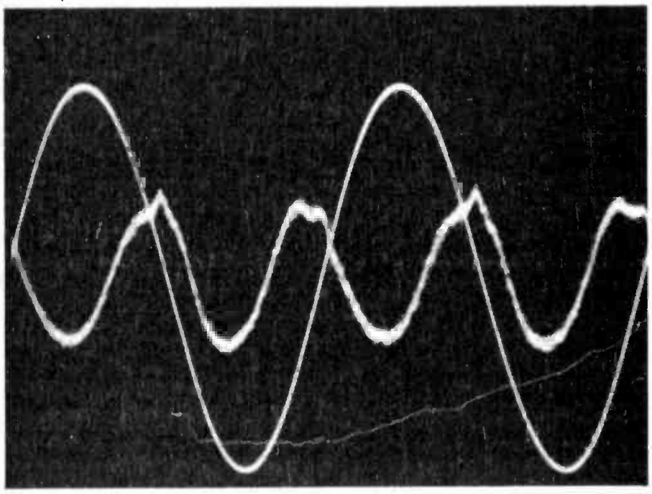


Fig. 4. Distortion factor oscillograms. (a) relatively high energy third-harmonic distortion.

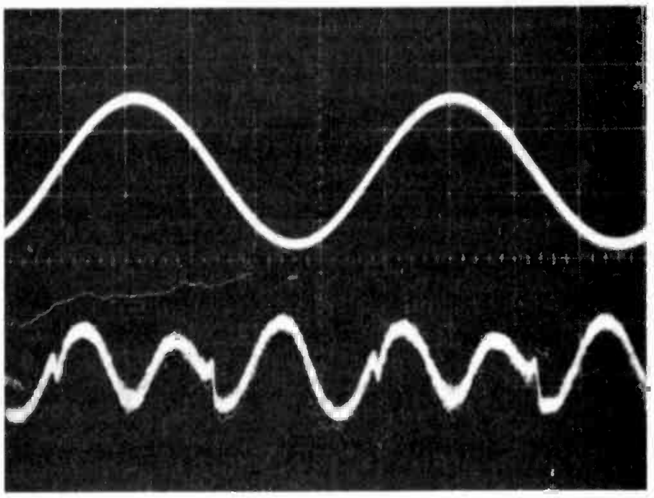


Fig. 4(c). High total energy, including crossover distortion.

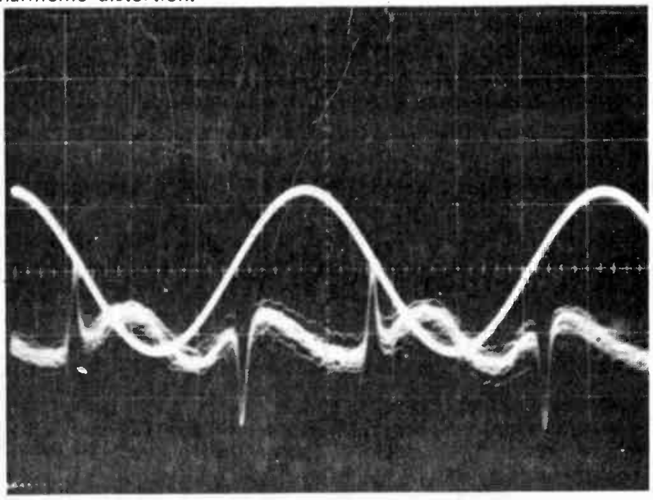


Fig. 4 (b). Low total energy but 'peaky' crossover.

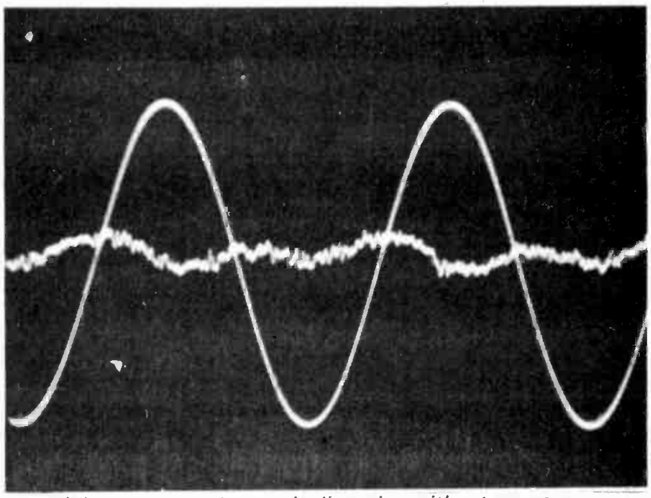
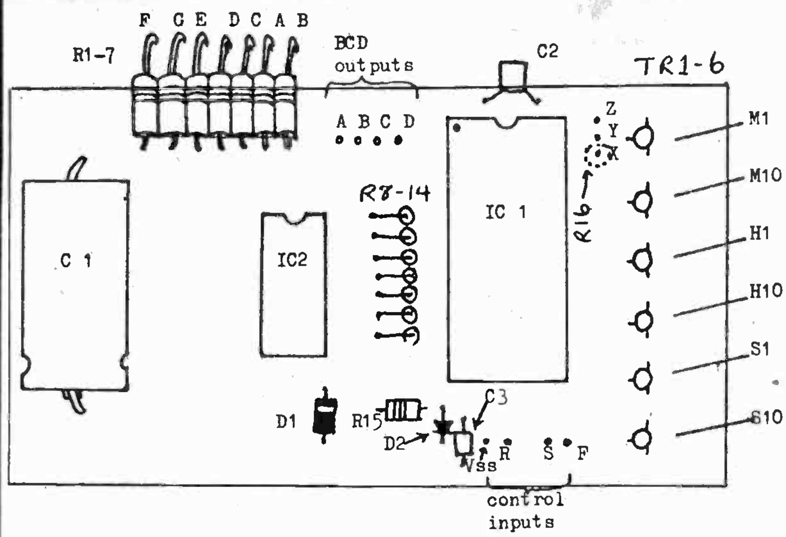


Fig. 4(d). Low energy harmonic distortion without crossover artifacts.

MHI-5309, MHI-5311, MHI-5318.

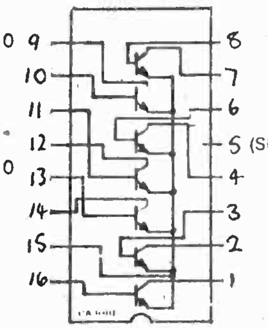


COMPONENTS LIST.

IC 1	MM5309/11/18	R1-R7	390 ohm ½ W
IC 2	CA3081	R8-R14	4K7 ¼ W
TR1-6	BC187 or BC213	R15-R16	100K 2 W
D1	IN4001	C1	500μF,25V
D2	IN914 or IN4148	C2	0.02μF
		C3	0.01μF

SW1-SW3 Push-ON, release-OFF

TFMR Transformer must be able to deliver at least 150mA (100mA for 4 digits) and may be 6-0-6, 9-0-9, 12-0-12.



CA3081

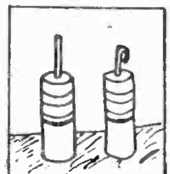


Fig 2. How to bend the leads of the seven segment output resistors.

ASSEMBLY INSTRUCTIONS.

Take care when soldering not to bridge any of the fine gaps between adjacent tracks on the PCB. Do not solder pins of ICs or transistors for too long or damage from overheating may occur. Never solder when power is applied.
Insert and solder into place the 4 or 6 digit drive transistors. The board has been drilled to accept BC187 or BC213, if a unit with a different pinout is used then refer to makers data.
Insert large smoothing capacitor (C1), being sure to check that it is the right way round. Solder into place.
Insert and solder IN4001 (D1), again being careful about polarity. White band on diode indicates positive end.
The power supply may be checked at this point by attaching the live transformer outputs to the points on the underside of PC board. The voltage across C1 should be within the range of 12-18v DC.
Disconnect transformer. Insert and solder diode D2. Insert and solder the sixteen resistors and capacitors C2 & C3.
Insert socket and solder. Insert the two ICs. DO NOT SOLDER — check the polarity, then solder.
Solder wires on to the tags on the back of the board and lead these to the three push switches and to the transformer outputs (POWER OFF).
The main clock board is now complete and ready for wiring to the display board. Only one end of resistors R1-R7 are soldered to the PCB, with the other ends sticking up to form connecting pins.
Cut the resistor leads to about ¼ inch long and then bend them back as shown in fig. 2. The wires to the display can now be hooked into the resistors and soldered into place. Cut the wires to be at least six inches long.

MHI-5309

A,B,C,D, are inverted BCD outputs
F,S,R, = Fast, Slow and Reset, Vss is the common for these switch inputs.
HOLD By connecting 50Hz to Vdd.
100K between X & Y (R16).
Z connected to Vdd to disable display.

MHI-5311

as 5309 except:
F,S,R = Fast, Slow, Hold.

MHI-5318.

as 5311 except:
X,Y,Z are ABC (in BCD format) of input digit address.
Break their track next to legend "MHI-5309" on rear of PCB which connects Y to Vdd.

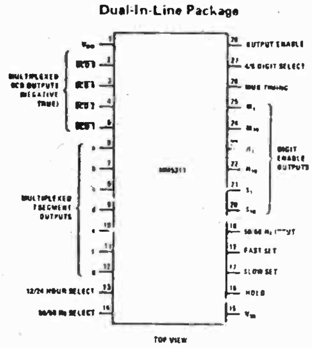
MM5318.

As MM5311 except pins 26,27 and 28 which are digit select lines.
(X-26, Y-27, Z-28.)

PIN CONNECTIONS.

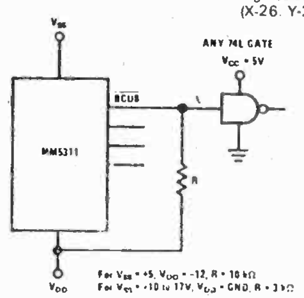
MM5309.
As MM5311 except pin 16 is reset.

Dual-In-Line Package

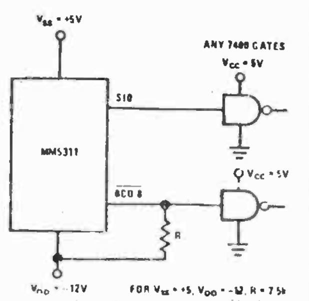


USES

MM5309— Stop watch or applications requiring a reset to zero function on both BCD & 7 seg outputs.
MM5311— General purpose clock chip interfacing to TTL or MOS.
MM5318— For applications requiring external digit selections such as micro-processors, printers. Interfaces with MM5841 TV time display chip.



MOS to Low Power TTL Interface



MOS to TTL Interface

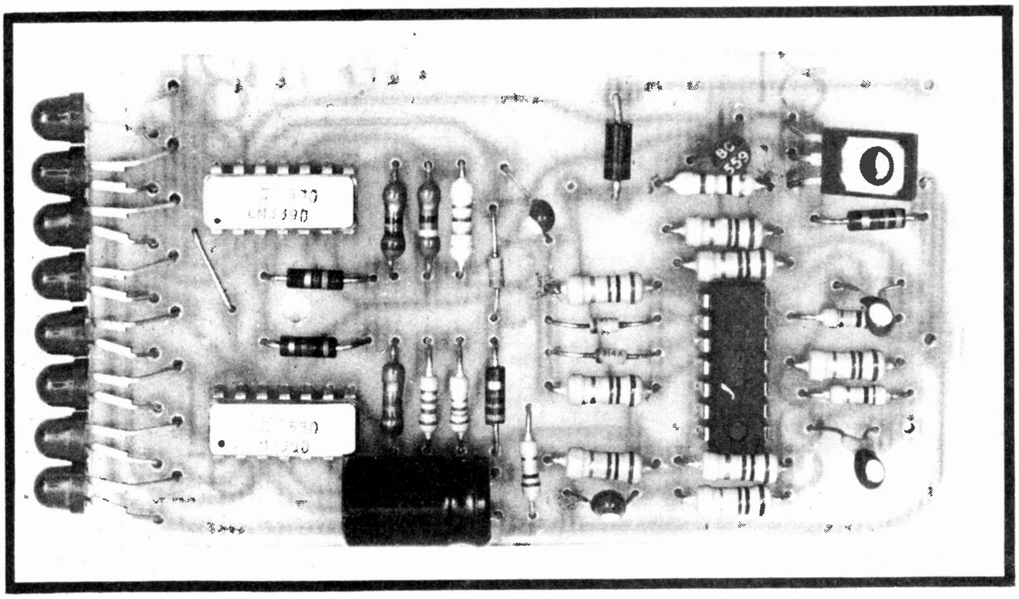
electrical characteristics

T_A within operating range, V_{SS} = +14V, V_{DD} = 0V, unless otherwise specified.

PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
Power Supply Voltage	V_{SS} (V_{DD} = 0V)	11	14	19	V
Power Supply Current	V_{SS} = +14V (No Output Loads)		8.0	15	mA
50/60 Hz Input Frequency		dc	50 or 60	60k	Hz
50/60 Hz Input Voltage					
Logical High Level		V_{SS} − 2.0	V_{SS} − 1.0	V_{SS}	V
Logical Low Level		−2.0	0	4.0	V
Multiplex Frequency	Determined by External R & C	dc	1.0	60	kHz
All Logic Inputs					
Logical High Level	Internal 20kΩ Resistor to V_{SS}		V_{SS}		V
Logical Low Level		−2.0	0	4.0	V
BCD and 7-Segment Outputs					
Logical High Level	Loaded 7kΩ to V_{DD}	2.0	5.0	20	mA source
Logical Low Level				0.01	mA source
Digital Enable Outputs					
Logical High Level				0.3	mA source
Logical Low Level	Loaded 100Ω to V_{SS}	5.0	10.0	25	mA sink

AUDIO LEVEL METER

eti project 438



Peak and average audio levels are indicated by a bar of light.

HIGH-POWER amplifiers usually incorporate meters to indicate the output-power levels in each channel. These meters are often called VU meters but in most cases they resemble proper VU meters only in the way they are scaled.

A professional VU meter is the industry standard for measuring the levels of complex music waveforms. It has a scale marked from −20 to +3 VU (on a steady state signal VU correspond to dB) where '0' VU corresponds to a level of one milliwatt into 600 ohms. The meter has a carefully controlled time constant such that if a reference tone level is applied the pointer of the meter will take 0.3 seconds to reach 99% of the reference level, and will then overshoot by not more than 1.5% and not less than 1.0%.

The professional VU meter is thus an instrument that has been designed to give a reasonable compromise between indicating the fast peaks and the average levels of a complex music waveform.

In contrast the meters fitted to some amplifiers have scales calibrated in VU but usually relying on the inertia of the meter movement to provide meter averaging. Apart from this the 0 VU point corresponds to the rated power output of the amplifier – not to 1 mW into 600 ohms (equivalent to 75 mW in 8 ohms). Strictly speaking therefore such meters should be called level or power meters, not VU meters.

Even the best of such meters are not fast enough to indicate accurately the peak levels which occur in music and hence are useless for detecting the onset of amplifier clipping. This is vital as at clipping amplifier distortion rises rapidly.

One alternative is to use in addition to the level meter a clipping indicator that detects fast peaks which exceed a preset level. The ETI 417 OVER-LED project (Nov. 73 issue) was such an instrument – it flashed an LED when a music transient exceeded clipping level.

The circuit described in this project is best described as a 'level meter'. It uses an array of LED diodes set to illuminate at successively higher increments in music level. With this type of display an estimate can quite easily be made of channel balance, and all transients, no matter how fast, are detected and indicated.

SPECIFICATION

Supply voltage	20 to 32 volts dc 15 to 20 volts dc
Supply current	16 mA dc approx,
Input sensitivity (VU meter)	500 k/v
Indication	8 LEDs 3 dB apart
Attack time	1 ms
Release time	0.5 sec.

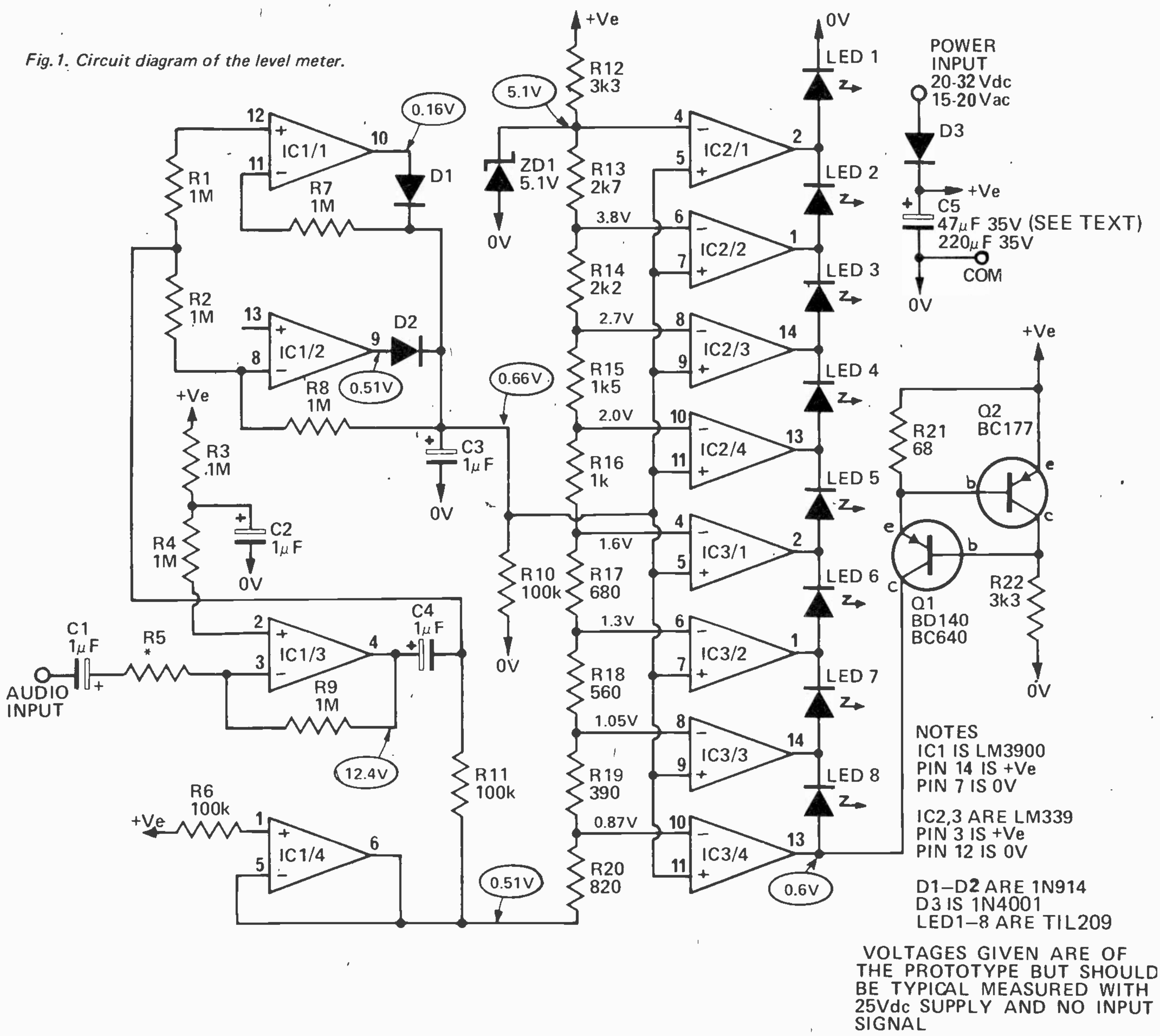
DESIGN FEATURES

The ETI 438 Level Meter can be arranged to indicate levels either in 'VU meter' format or in output power format. In the 'VU-meter' format the eight diodes light at 3 dB intervals from −18 to +3 VU where 0 VU corresponds to the nominal voltage required. Alternately as a power meter (remember that an amplifier cannot be driven beyond the clipping point) the top LED indicates maximum power and each lower LED indicates half the power of the one above it. The LEDs of the meter could thus be labelled, for example (for a 100 watt amplifier) 100, 50, 25, 12.5 watts etc.

The fast attack time of the meter

AUDIO LEVEL METER

Fig. 1. Circuit diagram of the level meter.



HOW IT WORKS – ETI 438

Although the circuitry of the level meter looks complicated the complete instrument only uses three ICs. These are an LM3900 which is a quad amplifier and two LM339s which are quad voltage comparators.

The input signal is amplified and buffered by IC1/3 to provide about 2.5 volts out at 0 VU input. The value of R5 is selected to give the sensitivity required for amplifiers of different power outputs. The gain of this amplifier is equal to the ratio of R9/R5.

A positive peak detector, IC1/1, and an inverting negative peak detector, IC1/2, give an output which represents the absolute peak level. Capacitor C3 and resistor R10 provide the peak hold and decay time. IC1/4 provides compensation for the 0.6 volt offsets of the LM3900 inputs.

The eight comparators are connected to a resistor divider chain the top of which is fed from a 5.1 volt supply which is stabilized by a zener. The resistor values are calculated to provide reference voltage steps at 3 dB intervals. The output of the detector is applied to all the non-inverting inputs of the comparators.

The LEDs are all connected in series and supplied with a constant current of 10 mA by the source consisting of Q1 and Q2. The outputs of the comparators are via open collector transistors which are "ON" if the input is lower than the reference voltage at the particular comparator input. With no input signal at all the comparators are all on thus shorting out all the LEDs so that none is on. As the input voltage rises the comparators turn off in sequence allowing the 10 mA to flow through the LEDs. Thus as the voltage increases a bar of light of increasing height is formed by the LEDs.

The current drawn from the power supply is about 16 mA and is independent of the number of LEDs which are on. Supply voltage is not critical and may be anywhere between 20 and 32 volts. Providing the supply is between these limits the unit will also be insensitive to supply ripple. When working from a dc supply a 47 microfarad filter capacitor is required but if an ac supply is used then the capacitor should be increased to 220 microfarad to minimize ripple. A single diode is used to both rectify the ac input and to prevent damage due to accidental reversed polarity if a dc supply is used.

PARTS LIST – ETI 438

R21 Resistor 68 ohm ½W 5%
R19 " 390 ohm ½W 5%
R18 " 560 ohm ½W 5%
R17 " 680 ohm ½W 5%
R20 " 820 ohm ½W 5%

R16 " 1k ½W 5%
R15 " 1k5 ½W 5%
R14 " 2k2 ½W 5%
R13 " 2k7 ½W 5%
R12,22 " 3k3 ½W 5%

R6,10,11 Resistor 100k ½W 5%
R1,2,7,8 " 1M ½W 5%
R3,4,9 " See Table 1 ½W 5%
R5 " See Table 1 ½W 5%

C1,2,3,4 Capacitor 1 μF 35V
*C5A " 47 μF 35V
*C5B " 220μF 35V

* use 47 μF for dc operation 220 μF for ac operation

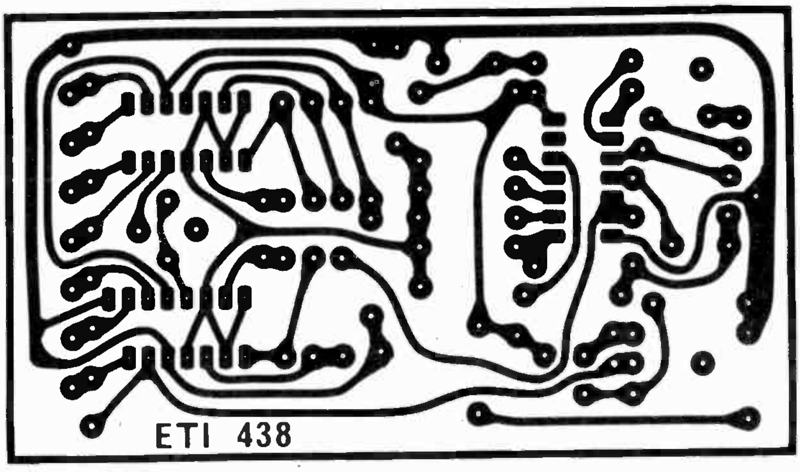
IC1 Integrated Circuit LM 3900
IC2,3 Integrated Circuit LM 339

D1,2 Diode IN914, BA318 or similar
D3 " 1N4001 or similar
ZD1 Zener diode 5.1 V 400 mW

Q1 Transistor BD 140,
Q2 " BC177,

LED 1-8 L.E.D. TIL209 or similar

PC board ETI 438



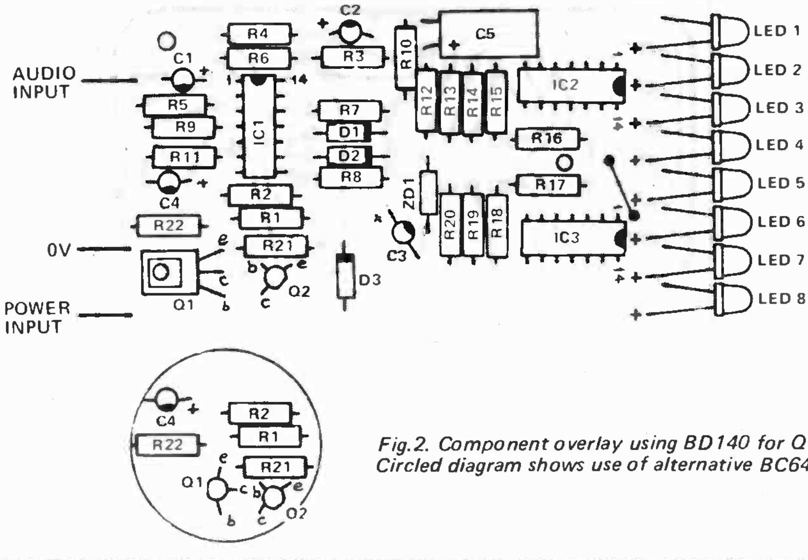


Fig.2. Component overlay using BD140 for Q1. Circled diagram shows use of alternative BC640

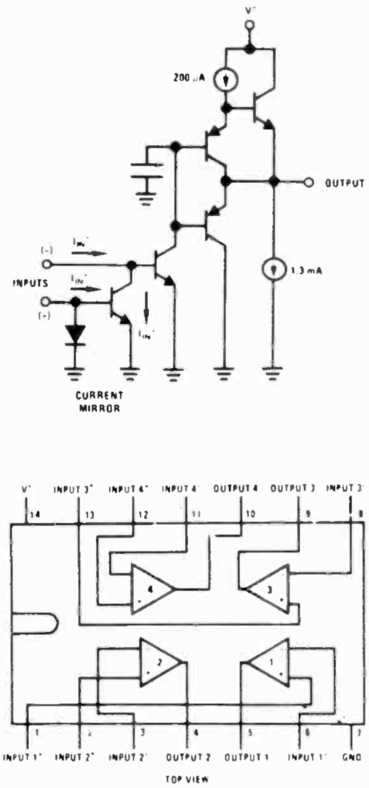


Fig.4. Internal circuitry and pin connections of the LM3900 IC.

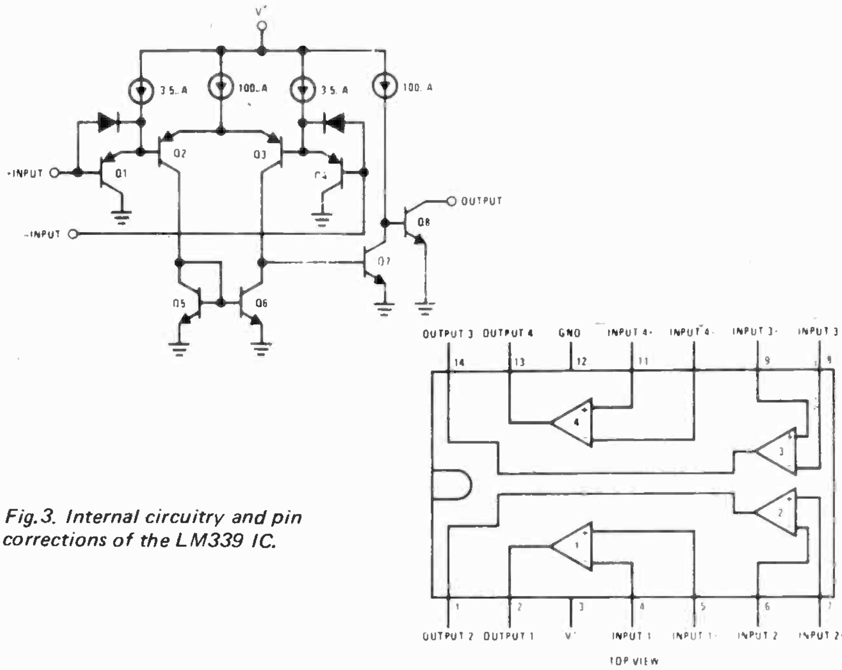


Fig.3. Internal circuitry and pin corrections of the LM339 IC.

AUDIO LEVEL METER

(less than one millisecond) ensures that even very short transients are detected, whilst the relatively slow release time (0.5 seconds) provides a reasonably-accurate, average — level indication.

In most previous designs for such meters, discrete transistors were used to build level detectors. Temperature effects and variations in gain led to inaccuracies and to calibration difficulties. These problems have largely been overcome in the ETI 438 meter by using the LM339 IC which contains four accurate level detectors in one package. Additionally the LM339 also has an open-collector output stage which enables a constant current supply for the LEDs to be used. Thus the current and LED brightness are the same no matter how many LEDs are alight.

If required the interval between LEDs may be altered by changing the values of R13 to R20. Thus for example, a 6 dB interval could be used. Additionally the display could be extended to 12 or even 16 diodes by adding comparators and LEDs and by substituting another divider chain for R20 (values would have to be calculated for the levels required). The positive inputs of the comparators would also be fed from C3 and R10.

A separate current source would be required as there is insufficient supply voltage available to light 16 LEDs in series. If the bottom LED in such a system indicates a level more than 30 dB down it may also be necessary to use a trimpot as the bottom resistor of the second divider chain to adjust for offsets etc.

The LM3900 is a quad differential amplifier which uses a current balancing technique at the input rather than the voltage balancing that is used with conventional operational amplifiers. Both the inputs "look" like the base-emitter junctions of normal transistors and both are at 0.6 volts with respect to ground. The currents into the two inputs must be equal if the output of the amplifier is to be in the linear region. In the case of IC1/3 the current into the positive input is set at about 12 microamps by R3 and R4. Current into the negative input is provided from the output by R9. If the current into the negative input is too low the output voltage will rise thus increasing the current into the negative input until balance is achieved. This self balancing ensures correct static biasing.

Gain is obtained by feeding a signal into R5 which adds or subtracts current into the negative input. For the amplifier to remain balanced there must be a corresponding shift in output voltage. The voltage gain is the ratio of R9 to R5.

TABLE 1B — POWER METER
FSD = 0 dB
R3, 4 and 9 are 100 k

POWER OUTPUT IN WATTS	VALUE OF R5 4 Ohms	8 Ohms	16 Ohms
5	150 k	200 k	270 k
10	200 k	270 k	390 k
15	240 k	330 k	470 k
20	270 k	390 k	560 k
25	330 k	430 k	620 k
30	360 k	470 k	680 k
40	390 k	560 k	820 k
50	430 k	620 k	910 k
75	560 k	750 k	1.1 M
100	620 k	910 k	1.2 M
150	750 k	1.1 M	1.5 M
200	910 k	1.2 M	1.8 M
250	1 M	1.5 M	2 M

$R5 = 32\sqrt{PR}$ Where P = power in watts
R = speaker impedance in Ohms.

SPECIFICATION LM3900

Maximum supply voltage	32 V
Supply current	6 mA typical
Voltage gain	2800 V/V typical
Input current range	1 µA — 1 mA
Current balance	0.9 — 1.1 at 200 µA
Bias current	30 nA typical
Output current capability	18 mA source typical. 1.3 mA sink typical

The LM339 is a quad voltage comparator where the output of each is an NPN transistor which has an unterminated collector and its emitter connected to ground.

SPECIFICATION LM339

Maximum supply voltage	36 V
Supply current	0.8 mA typical
Voltage gain	200 000 V/V typical
Offsett voltage	2 mV typical
Bias current	25 nA typical
Response time	1.3 µS typical
Output sink current	16 mA typical
Input common-mode voltage range	0 to (V^+ — 2 volts)

CONSTRUCTION

The meter will most likely be mounted in an existing amplifier or piece of equipment and for this reason the board construction only is given.

Layout of components is non-critical but, as with any multiple IC device, construction is greatly simplified by using the printed-circuit board specified. The usual precautions with polarities of components, such as capacitors, diodes, ICs and transistors should be observed. Some care must be taken when mounting the LEDs in order to obtain even spacing and good alignment. The long lead of the LED should be inserted in the hole furthest from the edge of the board. Put a slight curvature in the leads so that the LEDs can be aligned against the edge of the board (see photo). Take care not to bend the leads too often or too close to the body of the LED as the leads break very easily.

TABLE 1A — VU METER
FSD = +3 dB
R3, 4 and 9 are 1 megohm

SENSITIVITY	VALUE OF R5*
50 mV	22 k
100 mV	47 k
250 mV	120 k
500 mV	220 k
1 V	470 k

**Sensitivity equals R5 x 500 000 ohms.*

CALIBRATION

Resistor R5 is selected from Table 1 and this will ensure a result within 10 percent of that required. Greater accuracy may be obtained by using a variable potentiometer, in series with R5. To adjust this potentiometer inject a signal (around 1 kHz) equal to 0 VU (VU meter) or maximum power ($E = \sqrt{RP}$, e.g. 4 ohms and 100 watts, E = 20 volts) and adjust such that the second top LED (VU meter) or the top LED (power meter) just lights. ●

The only
6 Function L.E.D. Watch Kit
in the world . . .

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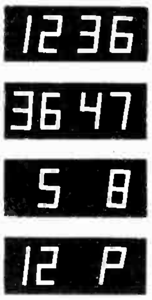
SABCHRON DIGITAL

Probably the only six-function LED watch kit available today, the Sabchron Digital offers more features than many other LED WATCHES. At the touch of a single button you see the exact hour, minute, second, month and day-of-month . . . PLUS an automatic date change memory which remembers whether the month contains 28, 30 or 31 days. Also, an AM/PM indicator which lights up while setting hours.



HOURS/MINUTES

MINUTES/SECONDS

DATE

AM/PM INDICATOR

AUTOMATIC LED BRIGHTNESS

The Sabchron Digital features a special light sensor circuit which automatically adjusts LED display brightness according to varying light levels, thus conserving battery power. In a bright room the display brightens; in a dim room or darkness the display dims — a feature not found in many other LED watches.

COMPLETE KIT

Contains all the components needed to build this space-age watch module, including solder, batteries and easy to follow assembly manual. If you can use a fine-point soldering iron and follow step-by-step assembly instructions, you can build the Sabchron Digital. If you have any problems, remember we're only as far as your nearest mailbox or telephone.

PLUS THESE OTHER FEATURES

- ★ Long-life silver-oxide power cells last up to one full year with normal use.
- ★ Shock-protected, anti-magnetic.
- ★ Low component count — IC, LED package, Quartz crystal trimmer, photocell and two chip capacitors.
- ★ Universal module size — fits many LED watch cases (optional gold plated case available).
- ★ Easy time and date resetting — you can change the time or date without affecting the accuracy of other functions. A very useful feature when crossing time or date lines.

NEW CMOS IC CHIP

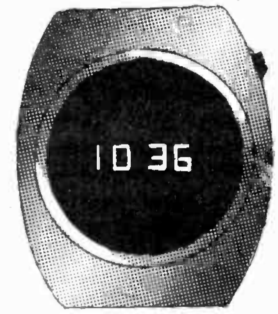
Containing the equivalent of over 1,500 transistors, this new IC chip features STATIC-PROTECTED TERMINALS — this means the IC can be handled safely with bare hands. It contains all the logic, decoding, multiplexing, display drive and counting circuits on a single silicon slice 1/50 square inch.

QUARTZ CRYSTAL ACCURACY

The heart of the Sabchron Digital is a quartz crystal which provides a basic frequency of 32,768 Hz. This frequency is tuned to ± 2 parts per million via the trimmer capacitor and is divided in half 15 successive times to provide one incredibly accurate pulse every second. This means a timekeeping accuracy to ± 5 seconds per month. Careful adjustment of the trimmer can bring this accuracy to within ± 1 second per month or ± 12 seconds per year!

OPTIONAL GOLD-PLATED CASE

The above price does not include a case or band. A gold-plated case as pictured (without bracelet) is available for £5.95. Note: The case is gold-plated; not "gold-tone", "gold-colour" or plastic. Elegantly styled and only slightly larger than a conventional watch, it is shown here actual size. Wear a time computer that looks like a watch, not vice versa. Cut out this case outline and try the size on your wrist.



★ ORDERING INFORMATION

All prices shown in British £'s are approximate equivalents of the following U.S. Dollar prices. Complete watch kit (less case and bracelet) $38.00; Gold-plated case only $12.00. Total $50.00 postpaid via registered airmail to customers in U.K., Eire and Canada. Other countries please add $3.00 additional for postage/packing. REMITTANCE BY BANK DRAFT or INTERNATIONAL MONEY ORDER IN U.S. DOLLARS (available from all banks and national Giro offices).

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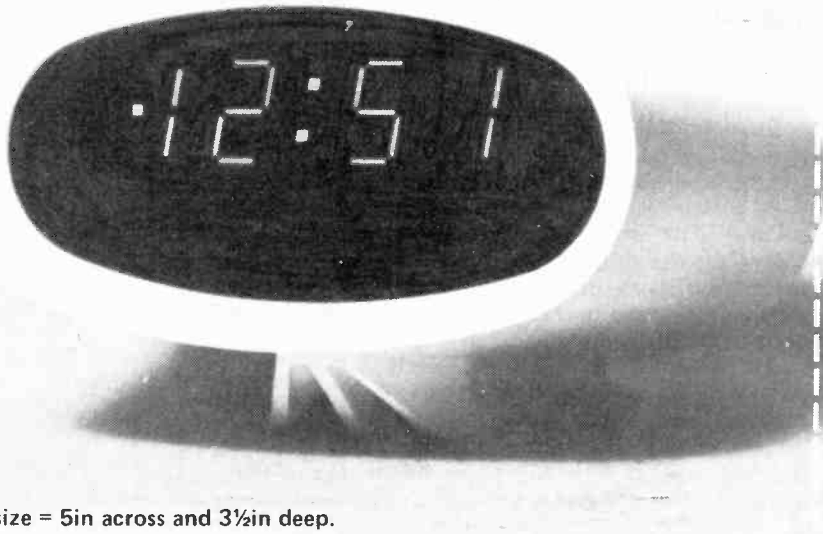

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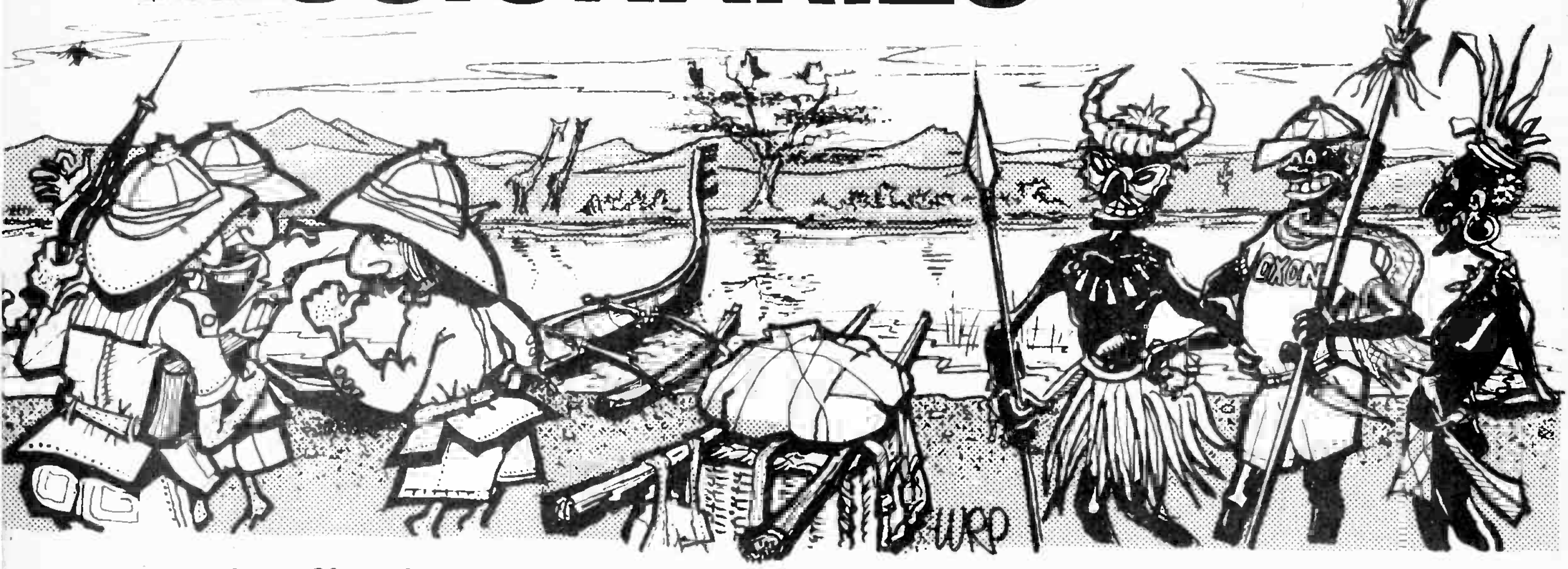
CANNIBALS and MISSIONARIES



— the final river crossing problem

Designed by A.J. LOWE

HERE'S a particularly perplexing problem provided for people with painstaking propensities. It's an electrical model of the puzzle which goes like this:

Three missionaries and three cannibals come to a river they want to cross. A little boat at the bank will carry only two people. All the missionaries can row, but only one of the cannibals can row — he'd been to Oxford. He also wears a red shirt! If at any time, on either side of the river, cannibals outnumber missionaries then the cannibals will eat the missionaries, which, understandably, the missionaries don't want. Problem: how do they cross safely?

In the model shown in Fig.1 the missionaries are represented by three switches M1, M2 and M3, and the cannibals by three switches C1, C2 and C3. The missionary switches have white levers. Two of the cannibal switches have black levers, but the switch representing the cannibal who wears a shirt and learned to row at Oxford – C2, has a red lever.

By operating the switches to represent crossings of the people involved — never more than two at a time as that's the limit of the boat, you try to solve the problem. If at any time a situation arises where, on either bank, cannibals outnumber missionaries then an alarm sounds and you've failed.

The circuitry detects situations where cannibals can satisfy their taste for eating missionaries, but it does not detect cheating — such as putting three people in the boat, or allowing a cannibal who can't row to be in the boat on his own.

CONSTRUCTION

The prototype was assembled in a plastic box 140 mm x 100 mm x 75 mm high with an aluminium front panel. Modern telephone-type key switches were used each having four changeover switches on each side of the switch.

Figure 3 shows the bottom view of one of the switches and how its terminals are laid out. It also shows, by means of the arrow-headed lines, which terminals connect with the moving parts of the switches. The eight changeover switches which comprise one key switch have been lettered a to h for convenience and to tie in with the lettering in Fig.2.

Note carefully that the switches in Fig. 3 are shown making the circuits which they make when the switch lever is in its *central* position. When the switch lever is moved from the central position to the start side of the river it changes over only the switches on the opposite half of the switch — i.e. switched a, b, c, d. When the switch lever is moved from the centre position to the far side of the river it changes over only switches e, f, g, h.

These key switches can be bought with push-on handle covers of various colours. The prototype used white covers for the missionaries, black for two of the cannibals, and red for the cannibal C2 who wore a shirt (and went to Oxford).

Although key switches each containing a total of eight changeover switches were used in the original, actually it is only the missionaries who need all eight switches. The cannibals need only five changeovers, but it was

ETI project 803

HOW IT WORKS

The circuit is a switching logic circuit. See Fig. 2. The cell and buzzer are between the outer vertical rails, and if ever a way between these two rails is set up by the switches then the alarm sounds. The circuit shows all the switches in the *starting* position i.e. all the missionaries and cannibals are on the near bank. Note that when any person goes over the river *all* switches changeover. The customary dotted lines showing the connections between coupled switches have been omitted for clarity. Thus, if M1 crosses the river, switches M1a, M1b, M1c, M1d, M1e, M1f, M1g and M1h all changeover.

You can work out the circuits for the alarm to sound. Here are three examples. Suppose all three cannibals stay on the start side and M1 goes over. Then the cannibals outnumber the missionaries on the start side and so the alarm sounds – through C2b, C3b, C1b, M2d, M2c and M1d which has changed over.

Similarly, if M2 went over alone then the alarm would sound through C2b, C3b, C1b, M3d, M1d, M3b and M2a which has changed over.

And if M3 went over alone the alarm sounds through C2b, C3b, C1b, M1c, M2c, M1d and M3b which has changed over.

You can check all the 'alarm should sound' configurations on each bank of the river by visualising an alarming situation – cannibals outnumbering missionaries, and then tracing through the switches to find a circuit. Similarly the 'alarm should no sound' circuits, or rather 'no' circuits can be checked in the same way.

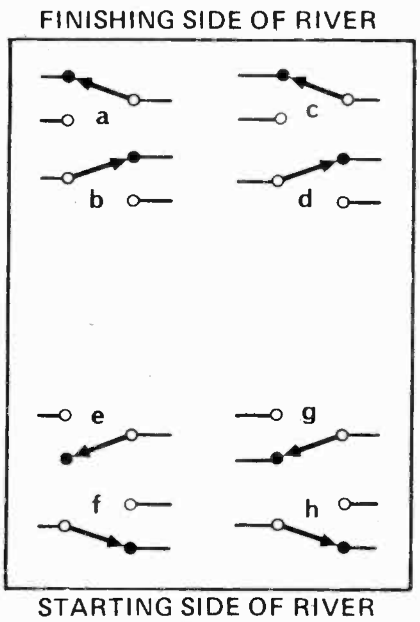


Fig. 3 Terminal layout on standard key switch with four changeover on each half. Contacts shown as being made in this diagram are with the lever in the central position.

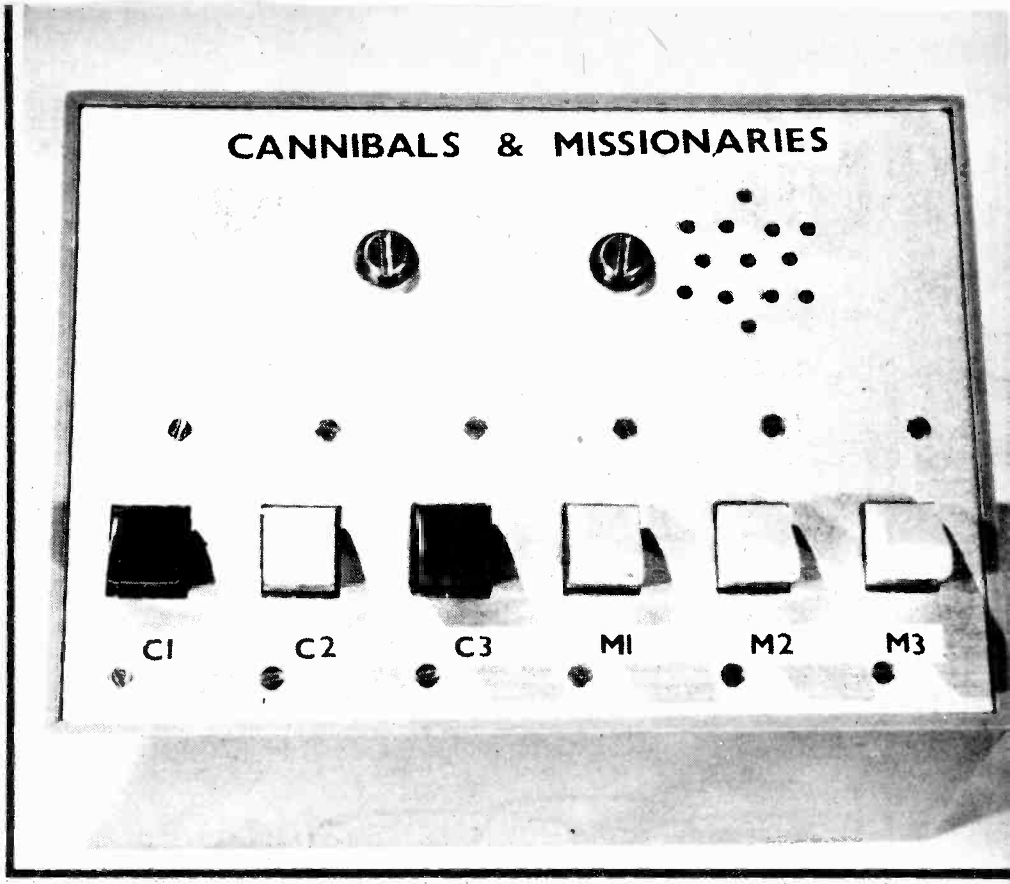


Fig. 1. The finished model. Lettering done with press-on letters on white Contact.

thought simpler to buy six identical switches.

Those with access to disposal stores could probably buy enough old style key switches comparatively cheaply, to build up the necessary number of switching functions needed.

The panel aperture dimensions for a standard key switch are shown in Fig.4.

The buzzer alarm and battery holder are all one piece — taken from a bicycle horn.

Wiring must be done carefully — very carefully! Bare wire was used in the prototype. Figure 5 shows the wiring diagram for the start side half of the switches and Fig. 6 shows the wiring of the other half. They are shown separately to minimise confusion. As can be seen from Figure 7, the switch wiring needs considerable care. On each of Fig. 5 and Fig. 6 one lead is marked 'To Buzzer' and another 'To battery —ve'. These leads, i.e. both buzzer leads are joined together and to the buzzer; and both battery —ve leads are joined together and run to battery —ve.

When wiring up — work logically. Start with switch M3 which is on the

PARTS LIST

Six standard phone type key switches type 4CL/4CL which means 4 changeover locking switches on each half of the switch.
e.g. Doram lever key switch type 338-018
One buzzer and battery holder — ex bicycle horn
One suitable box and panel
Hook up wire etc.

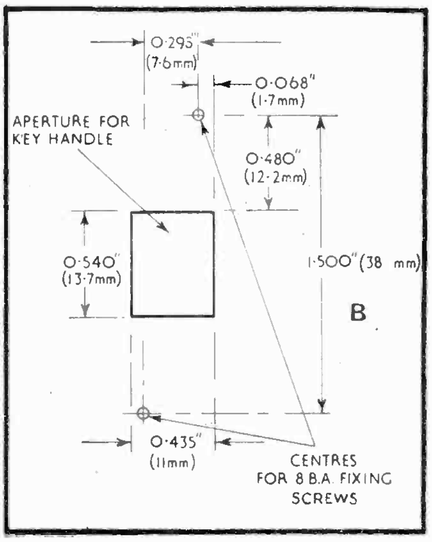


Fig. 4 Dimensions of aperture needed and hole positions for standard key switch.

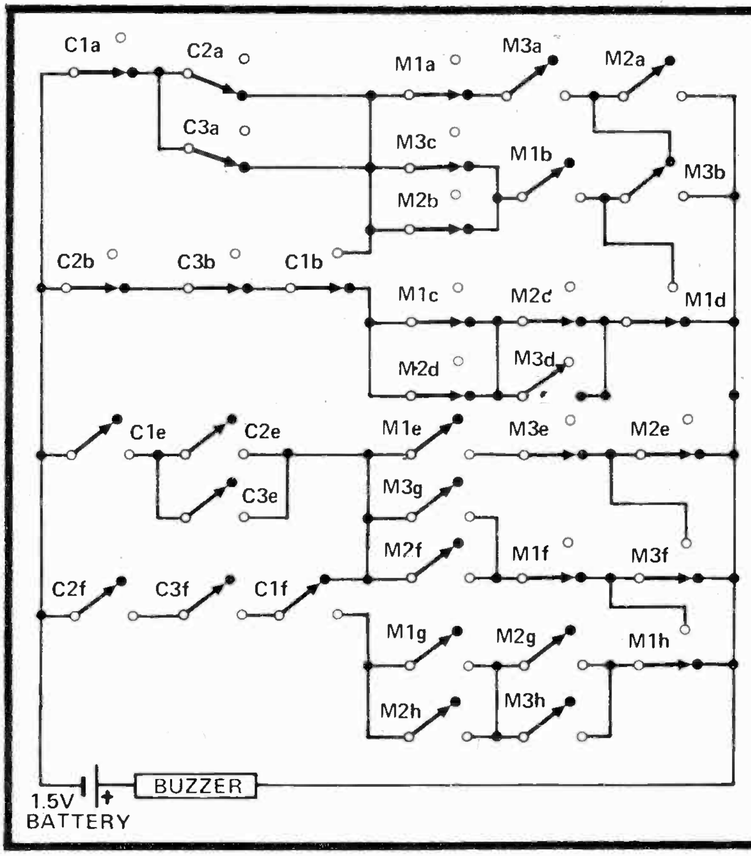


Fig. 2 The circuit diagram with all switches shown with levers on the start side of the river

CANNIBALS and MISSIONARIES

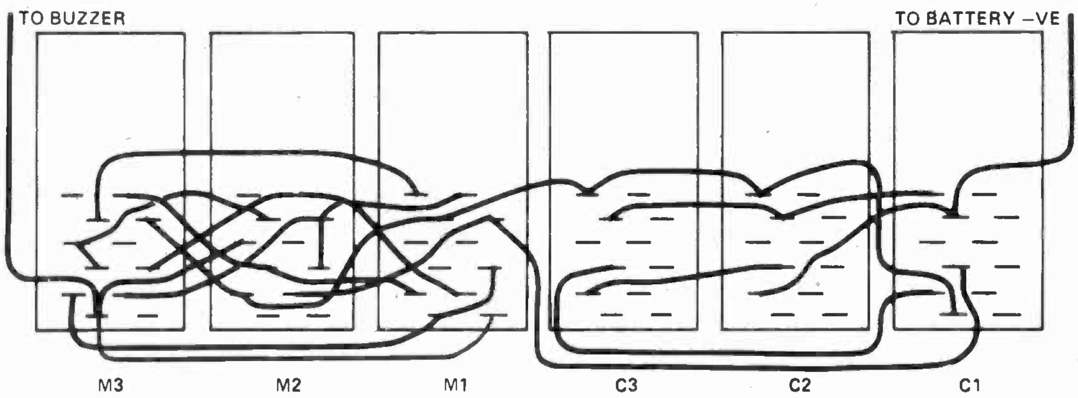


Fig. 5 Wiring on switches on the start side i.e. on switches 'e' to 'h' of each key switch.

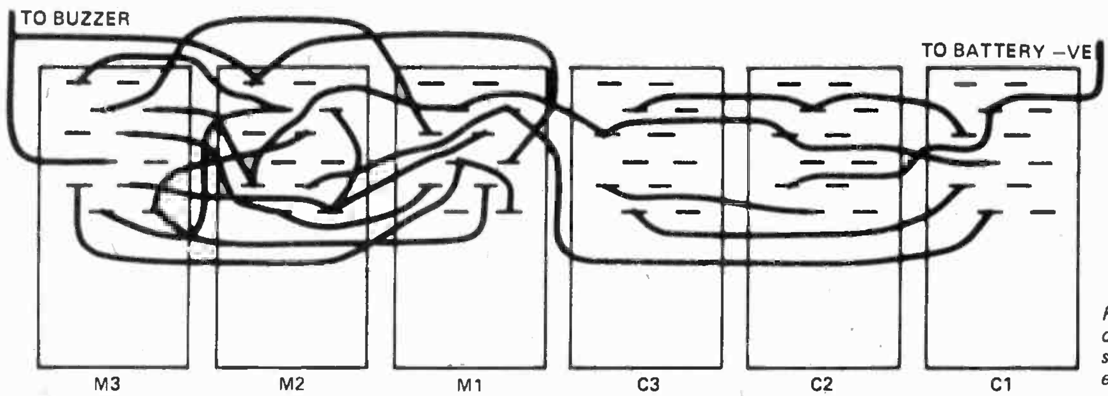


Fig. 6 Wiring on switches on finish side i.e. on switches 'a' to 'd' of each key switch

left when the panel is upside down. Start with the top left hand terminal and make all connections to it. Then move down each terminal in turn down the left hand row of terminals. Proceed row by row to the right making and checking connections to each terminal. It's a good idea to cross off with a pencil, each connection shown on the wiring diagrams, as soon as that connection has been made on the switches. Be sure not to miss the short connection between switches M3e and M3f.

If you want to check through the wiring diagrams, the circuit, and the switch diagram — bear in mind that the switch diagram shows connections with the switch lever in the central position, and the circuit diagram shows the connections with the levers in the start side position.

On completion of the wiring and after insertion of the cell, the puzzle should work. Check out all the alarm situations on both banks and see that the alarm sounds when it should. Also check the no-alarm situations — i.e. when cannibals do not outnumber missionaries on either bank of the river.

Fault finding is not as daunting as it may appear at first. A logical working through the circuit diagram should help to pin point any problem.

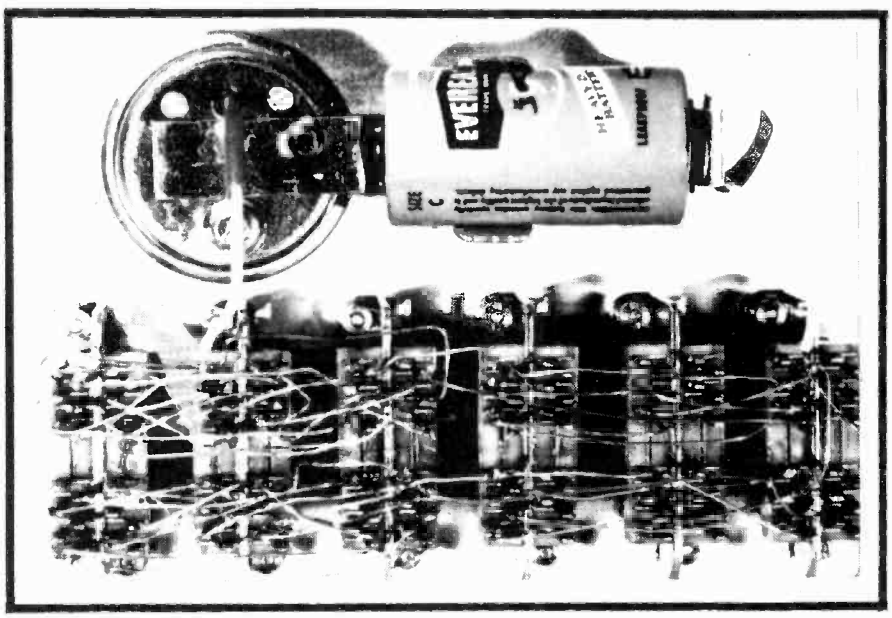


Fig. 7 Underside view of the panel.

HAVE A GO

Having built the puzzle — try to solve it. It's far from easy. If all else fails — check the correct answer which is hidden somewhere in this issue.

ETI DATA SHEET SPECIAL

SN72741-741 OP AMP — TEXAS

The SN72741 is Texas Instruments' version of the ubiquitous 741 op-amp. The 741 was originally introduced as the μA741 in May 1968, and was the first monolithic op-amp to be internally frequency compensated by an on-chip capacitor. The SN52741 is a military temperature range version of the SN72741. Both are high-performance operational amplifiers featuring offset voltage null capability. The high common-mode input voltage range and the absence of latch-up makes the 741 ideal for voltage follower applications. The devices are short-circuit protected and the internal frequency compensation ensures stability without external components. A low-value potentiometer may be connected between the offset null inputs to null out the offset null voltage as shown in Fig. 1.

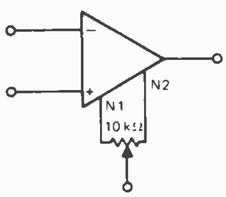
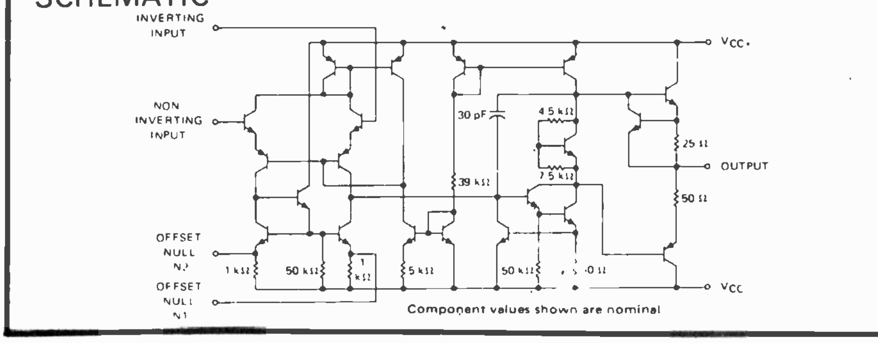


Fig. 1. Input offset voltage null circuit.

SCHEMATIC



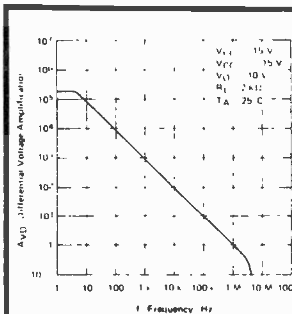


Fig. 2. Open-loop large signal differential voltage amplification vs. frequency.

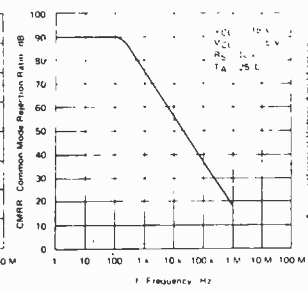


Fig 3. Common-mode rejection ratio vs. frequency.

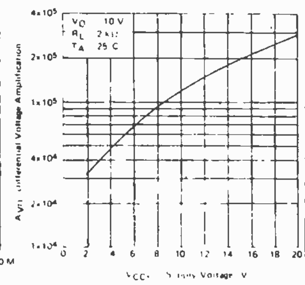


Fig 4. Open-loop large signal differential voltage amplification vs. supply voltage.

ELECTRICAL CHARACTERISTICS

Input Offset Voltage	7.5mV max.
Offset Voltage Adjust Range	15mV
Input Offset Current	300mA max
Input Bias Current	800mA max
Input Voltage Range	12V min
Max Pk-Pk Output Voltage Swing	
R_L 10k	24V min
R_L 2k	20V min
Large Signal Differential Voltage Amplification	15000 min
Input Resistance	2Mohm
Output Resistance	75ohm
Input Capacitance	1.4pF
Common-mode Rejection Ratio	70dB min
Power Supply Sensitivity	150μV/V max
Short-circuit Output Current	25mA
Supply Current	3.3mA max
Total Power Dissipation	100mW max

ABSOLUTE MAXIMUM RATINGS

	SN52741	SN72741
Supply Voltage Vcc+	22V	18V
Supply Voltage Vcc−	−22V	−18V
Differential Input Voltage	30V	30V
Input Voltage	15V	15V
Voltage between Offset Null Terminals (N1/N2) and Vcc−	0.5V	0.5V
Duration of Output Short-circuit		unlimited
Continuous Total Power Dissipation		500mW

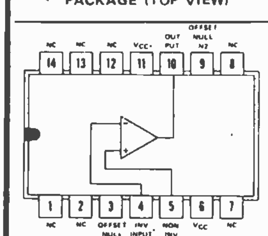
OPERATING CHARACTERISTICS

V_{CC+}=15V, V_{CC-}=−15V

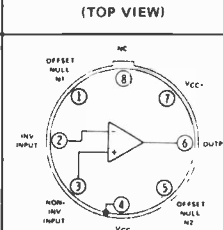
Rise Time (tr)	0.3μS
Overshoot	5%
Slew Rate at Unity Gain	0.5V/μS

TERMINAL ASSIGNMENTS

J OR N DUAL-IN-LINE PACKAGE (TOP VIEW)

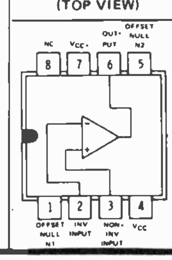


L-PLUG-IN PACKAGE (TOP VIEW)

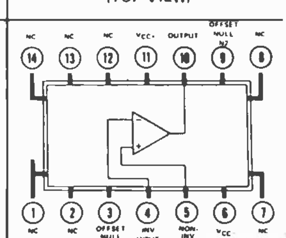


PIN 4 IS IN ELECTRICAL CONTACT WITH THE CASE

P DUAL-IN-LINE PACKAGE (TOP VIEW)



Z FLAT PACKAGE (TOP VIEW)



OP-AMPS

SN72301A-301 OP-AMP — TEXAS

The SN52101A is the Texas equivalent of the LM101A. The LM101 first appeared in 1967 and the LM101A is similar to the 101 but features better dc input characteristics. The SN52101A is a full military temperature range device and so the average constructor is likely to use the SN72301A which is a commercial temperature range device for operation from 0°C to 70°C.

Both are high-performance operational amplifiers, featuring very low input bias current and input offset voltage and current to improve the accuracy of high-impedance circuits using these devices. The high common-mode input voltage range and the absence of latch-up make these op-amps ideal for voltage-follower applications. The devices are protected to withstand short-circuits at the output. The external compensation of the 101A and 301A allows the changing of the frequency response (when the closed-loop gain is greater than unity) for applications requiring wider bandwidth or higher slew rate.

A potentiometer may be connected between the off-set null inputs (N1 & N2) as shown in Fig. 5, to null out the offset voltage.

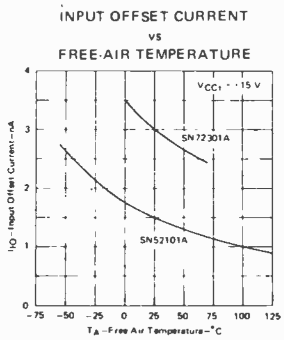


Fig. 1. Input offset current vs. free-air temp.

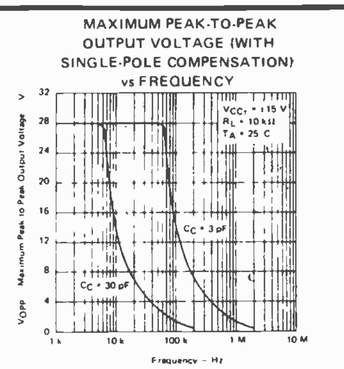


Fig. 2. Maximum p-p output voltage vs. frequency (for single-pole compensation).

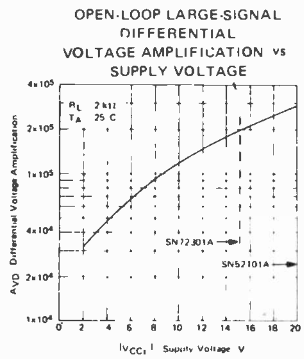


Fig 3. Open-loop large-signal differential voltage amplification vs. supply voltage.

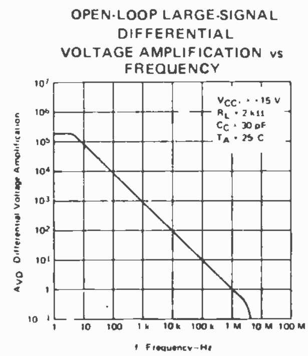
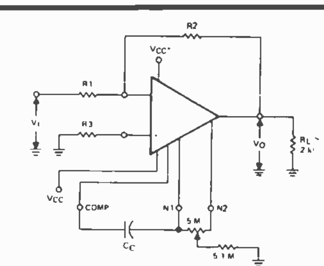


Fig. 4. Open-loop large-signal differential voltage amplification vs. frequency.

ELECTRICAL CHARACTERISTICS

	SN72301A
Input offset voltage	2mV
Average temperature coefficient of input offset voltage	6μV/°C
Input offset current	3nA
Average temperature coefficient of input offset current	0.02nA/°C
Input bias current	70nA
Input voltage range	±12V min
Maximum pk-pk output voltage saving	
(VCC = ±15V, RL = 10K)	28V
(VCC = ±15V, RL = 2K)	26V
Large signal differential voltage amplification	200,000
Input resistance	2MΩ
Common-mode rejection ratio	90dB
Power supply rejection ratio	96dB
Supply current	1.8mA



$$\frac{V_O}{V_I} = -\frac{R2}{R1} \qquad C_C \geq \frac{R1 \cdot 30\,pF}{R1+R2}$$

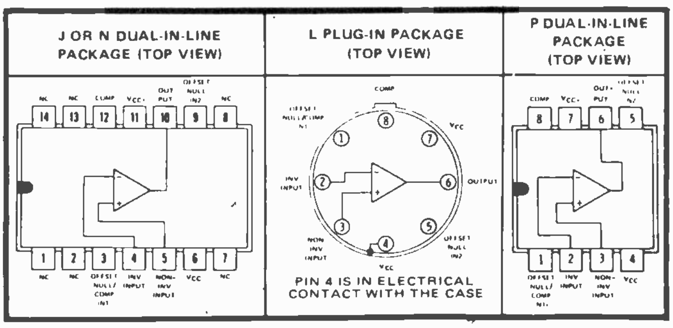
$$R3 = \frac{R1 \cdot R2}{R1+R2}$$

Fig. 5. Inverting circuit with adjustable gain single-pole compensation and offset adjustment.

ABSOLUTE MAXIMUM RATINGS

	SN52101A	SN72301A
Supply voltage VCC+	22V	18V
Supply voltage VCC—	—22V	—18V
Different input voltage	±30V	±30V
Input voltage	±15V	±15V
Voltage between N1/N2 (either offset null terminal) and VCC—	0.5V to 2V	—0.5V to 2V
Duration of output short circuit	unlimited	unlimited
Continuous total output power	500mW	500mW

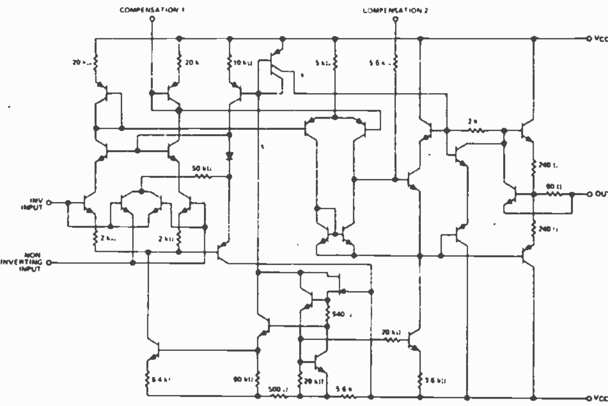
TERMINAL ASSIGNMENTS



NC—No internal connection

The 108A and 308A are designed for applications requiring extremely low input bias and offset currents and offset voltages. The SN52108A has a typical input offset voltage of 300μV and typical input offset current of 50pA at25°C. Input bias current is 3nA maximum over the full temperature range. The output is protected against damage from shorting to ground or either supply and the input stage is diode-protected against excessive differential input signals. External compensation permits optimization of the frequency response for each application. The compensation circuit shown in Fig.3 can be used to make the amplifier particularly insensitive to supply noise.

SCHEMATIC



Component values shown are nominal.

APPLICATION DATA

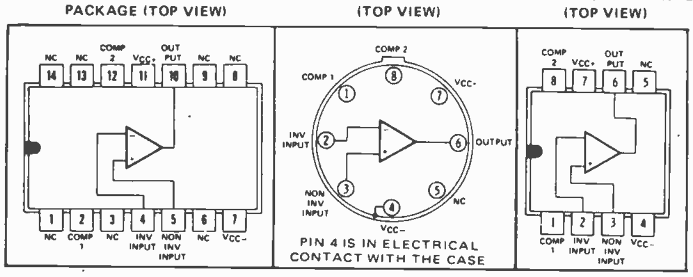
Extra care must be taken in the assembly of printed circuit boards to take full advantage of the low input currents of these amplifiers.

Even with properly cleaned and coated boards, leakage currents may cause trouble at 125°C, particularly in the L plug-in package (TO-99) where the input pins are adjacent to pins that are at supply potentials. This leakage can be reduced by surrounding the input terminals with a conductive guard ring. The guard ring is connected to a low-impedance point that is at approximately the same voltage as the inputs. As shown in Figure 4, input guarding of the 8-lead L package may be accomplished by using a 10-lead pin circle, with leads of the device formed so that the holes adjacent to the inputs are empty when it is inserted in the board. The conductive guard ring should be used on both sides of the board.

The pin configurations of both the dual-in-line and flat packages are designed to facilitate guarding, since the pins adjacent to the inputs are not internally connected to the chip. Pin connections in these packages are different from the standard pin connections used in other operational amplifiers such as the SN52741/SN72741.

TERMINAL ASSIGNMENTS.

ALL TYPES J OR N DUAL-IN-LINE PACKAGE (TOP VIEW) — ALL TYPES L PLUG-IN PACKAGE (TOP VIEW) — SN72308AP, SN72308P P DUAL-IN-LINE PACKAGE (TOP VIEW)



ELECTRICAL CHARACTERISTICS

(values typical except where specified)	SN 72308A	SN 72308
Input Offset Voltage	0.3mV	2mV
Av. Temp. coefficient of Input Offset Voltage	1μV/°C	6μV/°C
Input Offset Current	0.2nA	0.2nA
Av. Temp. coefficient of Input Offset Current	2pA/°C	2pA/°C
Input Bias Current	1.5nA	1.5nA
Input Voltage Range	±14V min	±14V min
Max. Pk-Pk Output Voltage Swing	28V	28V
Large-signal Differential Voltage Amplification	300,000	300,000
Input Resistance	40M ohm	40M ohm
Common-mode Rejection Ratio	110dB	100dB
Supply Voltage Rejection Ratio	110dB	96dB
Supply Current	0.3mA	0.3mA

ABSOLUTE MAXIMUM RATINGS

	SN 52108A SN 52108	SN 72308A SN 72308
Supply Voltage Vcc+	20V	18V
Supply Voltage Vcc—	—20V	—18V
Input Voltage	±15V	±15V
Differential Input Current	±10mA	±10mA
Duration of Output Short-circuit	unlimited	unlimited
Continuous Total Power Dissipation (below 35°C)	500mW	500mW

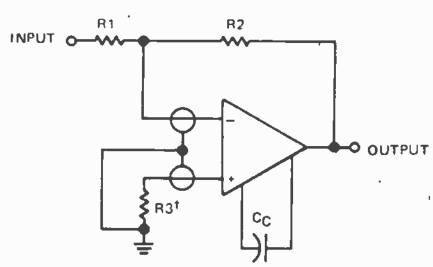


Fig. 1: Inverting amplifier.

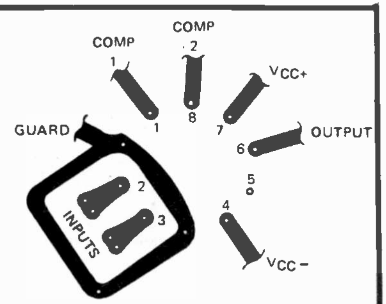


Fig. 4. Top view of board layout for L plug-in package.

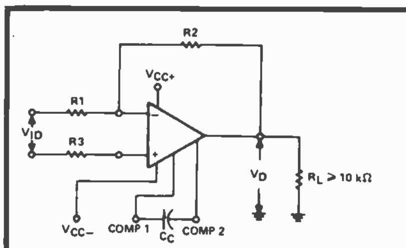


Fig. 2. Standard compensation circuit.

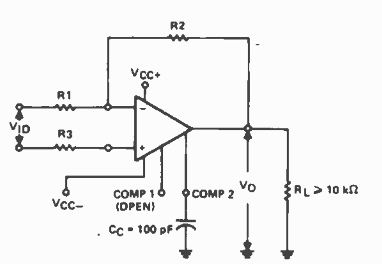
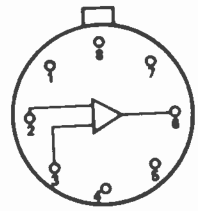


Fig. 3. Alternate compensation circuit.

536 FET-INPUT OP-AMP

SIGNETICS

The 536 is a special-purpose high-performance operational amplifier utilizing an FET input stage for extremely high input impedance and low input current. The device features internal compensation, standard pinout, wide differential and common mode input voltage range, high slew rate and high output drive capability. Two versions are available: the SU536T has a temperature range of –55°C to 85°C, while the NE536T covers 0°C to 70°C.



1. Offset Null
2. Inverting Input
3. Non-inverting Input
4. V⁻
5. Offset Null
6. Output
7. V⁺
8. NC

PIN-OUT DIAGRAM

ABSOLUTE MAXIMUM RATINGS

Supply Voltage	±22V
Differential Input Voltage Range	±30V
Comon Mode Input Voltage Range	±VS
Power Dissipation (Note 1)	500mW
Operating Temperature Range	0°C to ±70°C
Storage Temperature Range	–65°C to +150°C
Lead Temperature (Solder, 60 sec)	300°C
Output Short Circuit Duration (Note 2)	Indefinite

NOTES:

1. Rating applies for case temperatures to +25°C; derate linearly at 6.5mW/°C for ambient temperatures above 75°C.
2. Short circuit may be to ground or either supply. Rating applies to +125°C case temperature or +75°C ambient temperature.

ELECTRICAL CHARACTERISTICS

INPUT CHARACTERISTICS	
Large Signal Voltage Gain @ + 25°C	100V/mV
Input Offset Voltage @+25°C	30mV
vs Temperature (drift)	30μV/°C
vs Common Mode Voltage (C.M.R.R.)	80dB
vs Power Supply (P.S.R.R.)	100μV/V
Input Current @ +25°C	30pA
Input Offset Current @+25°C	5pA
Input Impedance	
Differential Resistance	10^{14}ohm
Differential Capacitance	6pF
Input Noise (0.1Hz — 100kHz)	
Voltage Noise	20μV r.m.s
Common Mode Voltage Range	±11V
OUTPUT CHARACTERISTICS	
Output Current	5mA min
Open Loop Output Impedance	100 ohm
Output Voltage Swing	
RL=2k ohm	±10V
RL=10k ohm	±13V
Short Circuit Current	17mA
FREQUENCY AND TRANSIENT RESPONSE	
Gain Bandwidth Product	1MHz
Unity Gain Frequency	1MHz
Full Power Bandwidth	100kHz
Slew Rate	
Inverter	6V/μS
Follower	6V/S
POWER SUPPLY REQUIREMENT	
Power Supply Range	±6V to ±18V
Quiescent Supply Current at ±15V Vs	6mA
Quiescent Power Dissipation	180mW

All values typical unless specified

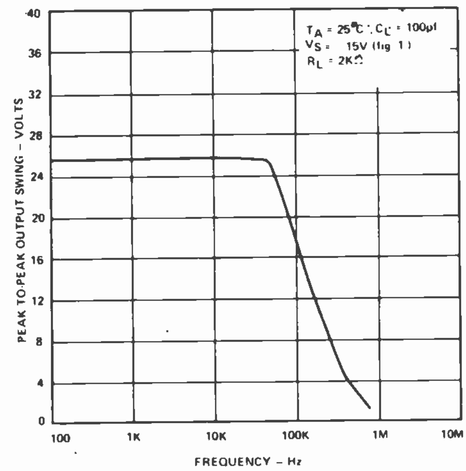


Fig 1. Output voltage swing vs frequency

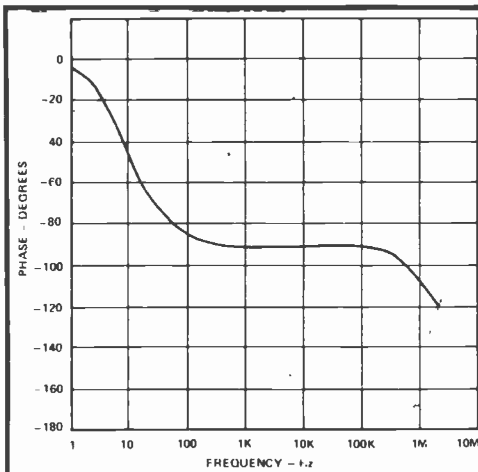


Fig 2. Open loop phase vs frequency

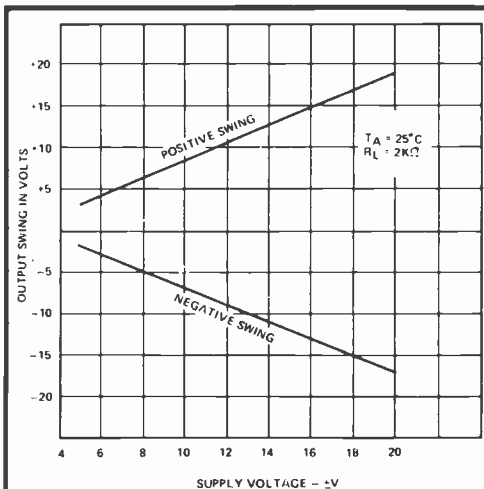


Fig 3. Output voltage swing vs supply voltage

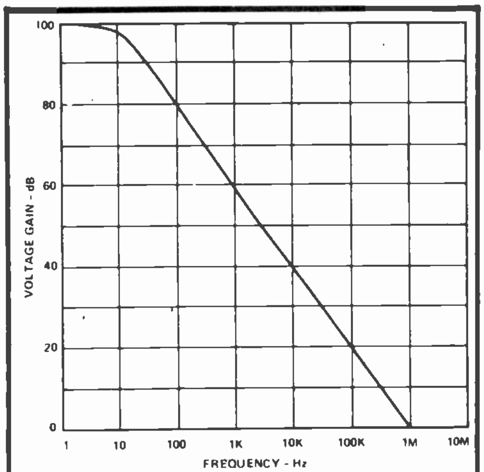
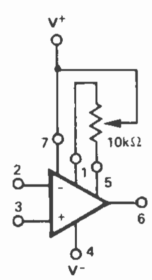


Fig 4. Open loop voltage gain vs frequency



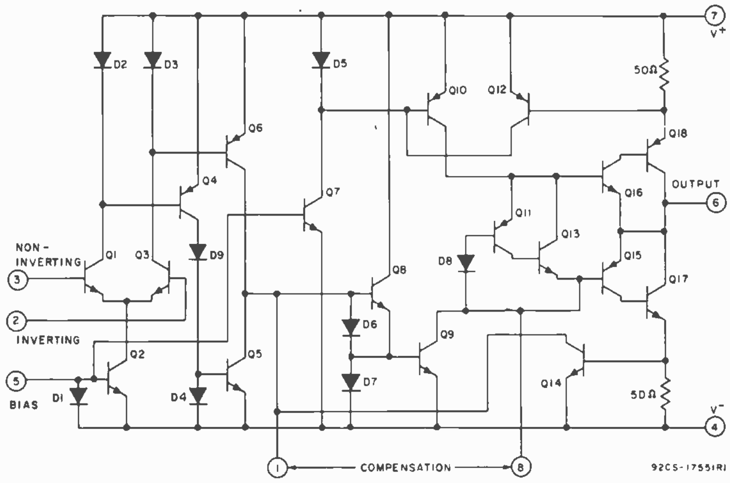
OFFSET NULL CIRCUIT

Fig 5. Offset null circuit

CA3078T, CA3078AT MICROPOWER OP-AMPS

The RCA CA3078T and CA3078AT are high-gain monolithic op-amps which can deliver milliamps of current yet only consume microwatts of standby power. Their operating points are externally adjustable and frequency compensation may be accomplished with one external capacitor. They provide the designer with the opportunity to tailor the frequency response and improve the slew rate without sacrificing power. Operation with a single 1.5V battery is a practical reality with these devices.

The CA3078T has a minimum supply voltage of ±0.75V and a maximum supply voltage of ±15V, with an operating temperature range of −55°C to 125°C. The CA3078T has the same lower supply voltage limit but the upper limit is ±6V and the temperature range is 0°C to 70°C.



SCHEMATIC

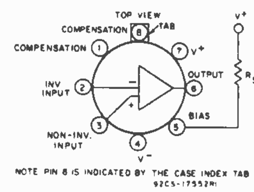
ELECTRICAL CHARACTERISTICS

All values typical unless specified.

	CA3078AT	CA3078T
Input Offset Voltage	0.70mV	1.3mV
Input Offset Current	0.50nA	6nA
Input Bias Current	7nA	60nA
Open-Loop Diff Voltage Gain	100dB	92dB
Total Quiescent Current	20μA	100μA
Device Dissipation	240μW	1200μW
Maximum Output Voltage	±5.3V	±5.3V
Common-Mode Input Voltage Range	−5.5 to ±5.8V	−5.5 to ±5.8V
Common-Mode Rejection Ratio	115dB	110dB
Maximum Output Current	12mA	12mA
Input Offset Voltage Sensitivity	105μV/V	93μV/V

Absolute values at $T_A = 25°C$.

FUNCTIONAL DIAGRAM



ABSOLUTE MAXIMUM RATINGS

	CA3078AT	CA3078T
DC Supply Voltage (between V+ and V− terminal)	36V	14V
Differential Input Voltage	±6V	±6V
DC Input Voltage	V+ to V−	V+ to V−
Input Signal Current	0.1mA	0.1mA
Output Short Circuit Duration*	No Limitation	No Limitation
Device Dissipation	250mW (up to 125°C)	500mW (up to 70°C)

*Short circuit may be applied to ground or to either supply

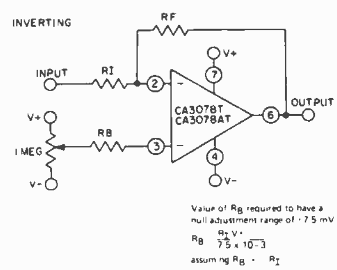


Fig 1. Offset voltage null circuit

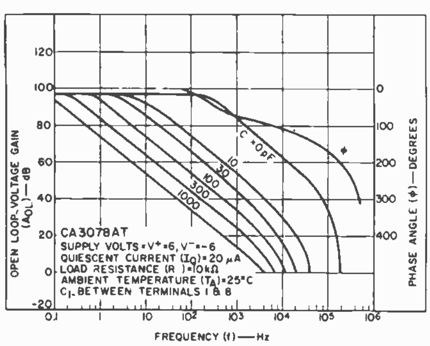


Fig 3. Open-loop voltage gain vs frequency for I_Q = 20μA — CA3078T

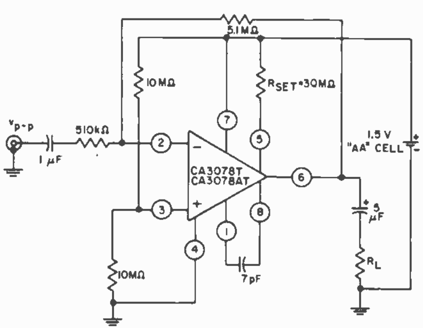


Fig 4. Inverting 20-dB amplifier circuit

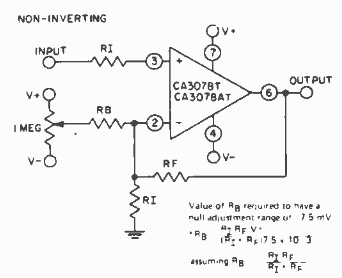


Fig 2. Offset voltage null circuit

DEFINITION OF OP-AMP TERMS

Input Offset Voltage (V_{IO}) The d-c voltage which must be applied between the input terminals to force the quiescent d-c output voltage to zero. The input offset voltage may also be defined for the case where two equal resistances (R_S) are inserted in series with the input leads.

Average Temperature Coefficient of Input Offset Voltage (α_{VIO}) The ratio of the change in input offset voltage to the change in free-air temperature. This is an average value for the specified temperature range.

$$\alpha_{VIO} = \left| \frac{(V_{IO}@T_{A(1)}) - (V_{IO}@T_{A(2)})}{T_{A(1)} - T_{A(2)}} \right|$$

where $T_{A(1)}$ and $T_{A(2)}$ are the specified temperature extremes.

Input Offset Current (I_{IO}) The difference between the currents into the two input terminals with the output at zero volts.

Average Temperature Coefficient Of Input Offset Current (α_{IIO}) The ratio of the change in input offset current to the change in free-air temperature. This is an average value for the specified temperature range.

$$\alpha_{IIO} = \left| \frac{(I_{IO}@T_{A(1)}) - (I_{IO}@T_{A(2)})}{T_{A(1)} - T_{A(2)}} \right|$$

where $T_{A(1)}$ and $T_{A(2)}$ are the specified temperature extremes.

Input Bias Current (I_{IB}) The average of the currents into the two input terminals with the output at zero volts.

Input Voltage Range (V_I) The range of voltage which if exceeded at either input terminal will cause the amplifier to cease functioning properly.

Maximum Peak-to-Peak Output Voltage Swing (V_{OPP}) The maximum peak-to-peak output voltage which can be obtained without waveform clipping when the quiescent d-c output voltage is zero.

Large-Signal Differential Voltage Amplification (A_{VD}) The ratio of the peak-to-peak output voltage swing to the change in differential input voltage required to drive the output.

Input Resistance (r_I) The resistance between the input terminals with either input grounded.

Output Resistance (r_O) The resistance between the output terminal and ground.

Common-Mode Rejection Ratio (CMRR) The ratio of differential voltage amplification to common-mode voltage amplification. This is measured by determining the ratio of a change in input common-mode voltage to the resulting change in input offset voltage.

Power Supply Sensitivity ($\Delta V_{IO}/\Delta V_{CC}$) The ratio of the change in input offset voltage to the change in supply voltages producing it. For these devices, both supply voltages are varied symmetrically.

Total Power Dissipation (P_D) The total d-c power supplied to the device less any power delivered from the device to a load. At no load: $P_D = V_{CC+} \cdot I_{CC+} + V_{CC-} \cdot I_{CC-}$

Rise Time (t') The time required for an output voltage step to change from 10% to 90% of its final value.

Overshoot The quoteint of: (1) the largest deviation of the output signal value from its steady-state value after a step-function change of the input signal, and (2) the difference between the output signal values in the steady state before and after the step-function change of the input signal.

Slew Rate (SR) The average time of change of the closed-loop amplifier output voltage for a step-signal input. Slew rate is measured between specified output levels (0 and 10 volts for this device) with feedback adjusted for unity gain.

ETI HELPING HAND COMPETITION

STILL TIME TO ENTER

CLOSING DATE 31 MARCH 1976

THE PROBLEMS

1 A sick person is being looked after by a deaf person. The deaf person has no useful hearing and requires to know whether the sick person is all right and above all needs to know if the sick person is in a state of distress anywhere in the sick room.

2 A hard of hearing person is attending a College of Further Education and has considerable difficulty in understanding what the lecturer says due to his distance from the lecturer and to the background noise in the room. A device is required to enable him to make the best possible use of his hearing.

3 Many deaf people have great difficulty in using the telephone and in fact many of them cannot use the telephone at all. The development of a writing tablet which would allow them to write a message on a small pad and for this to be communicated over the telephone line to a pad at the other end would have many great advantages. In addition the communication should be two way so that the person can receive a message or an indication that the message has been received.

This is our open competition to find solutions for problems facing the deaf.

This closing date is March 31st 1976. ETI and the Royal National Institute for the Deaf (RNID) are co-operating fully in the organisation of this competition.

Three problems are shown above. We invite individual readers, clubs, schools, universities, companies, in fact anybody, to develop a practical solution. The rules are as basic as possible and impose virtually no restriction apart from insisting that any Patent Royalties are waived if the idea is produced.

The prizes, three in all, will each be a silver trophy specially designed for ETI. At the close of the competition the magazine will hand over £250 to the RNID to help with development costs. There is a £1.00 entry fee (payable to RNID) and this will be added to the £250.

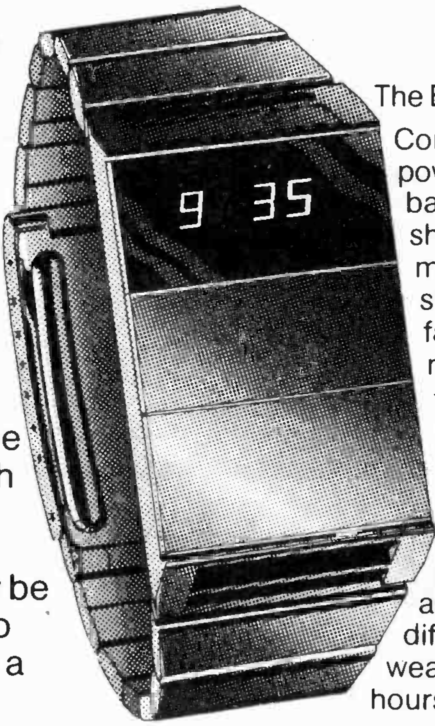
Background information has been prepared to help readers and say what is alreayd known. This is available from ETI on receipt of a large self-addressed envelope. Enquiries should be sent to:

Helping Hand,
ETI Magazine,
36 Ebury Street,
London, SW1W 0LW.

Now...the most exciting Sinclair kit ever

The Black Watch kit

At £14.95, it's



★**practical**–easily built by anyone in an evening's straightforward assembly.

★**complete** –right down to strap and batteries.

★**guaranteed.** A correctly-assembled watch is guaranteed for a year. It works as soon as you put the batteries in. On a built watch we guarantee an accuracy within a second a day–but building it yourself you may be able to adjust the trimmer to achieve an accuracy within a second a week.

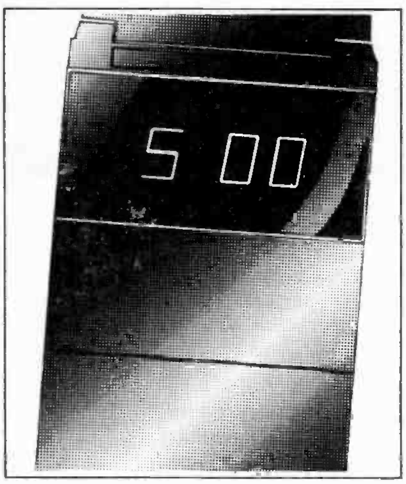
The Black Watch by Sinclair is unique. Controlled by a quartz crystal... powered by two hearing aid batteries...using bright red LEDs to show hours and minutes and minutes and seconds...it's also styled in the cool prestige Sinclair fashion: no knobs, no buttons, no flash.

The Black Watch kit is unique, too. It's rational–Sinclair have reduced the separate components to just four.

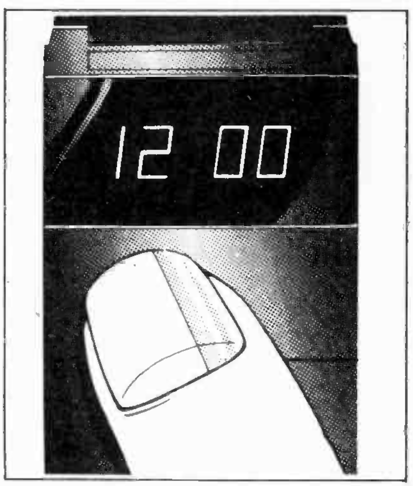
It's simple–anybody who can use a soldering iron can assemble a Black Watch without difficulty. From opening the kit to wearing the watch is a couple of hours' work.

The special features of The Black Watch

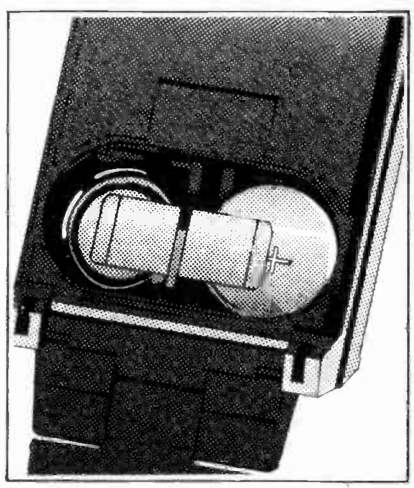
Smooth, chunky, matt-black case, with black strap. (Black stainless-steel bracelet available as extra–see order form.)



Large, bright, red display–easily read at night.
Touch-and-see case–no unprofessional buttons.



Runs on two hearing-aid batteries (supplied). Change your batteries yourself–no expensive jeweller's service.



The Black Watch–using the unique Sinclair-designed state-of-the-art IC.

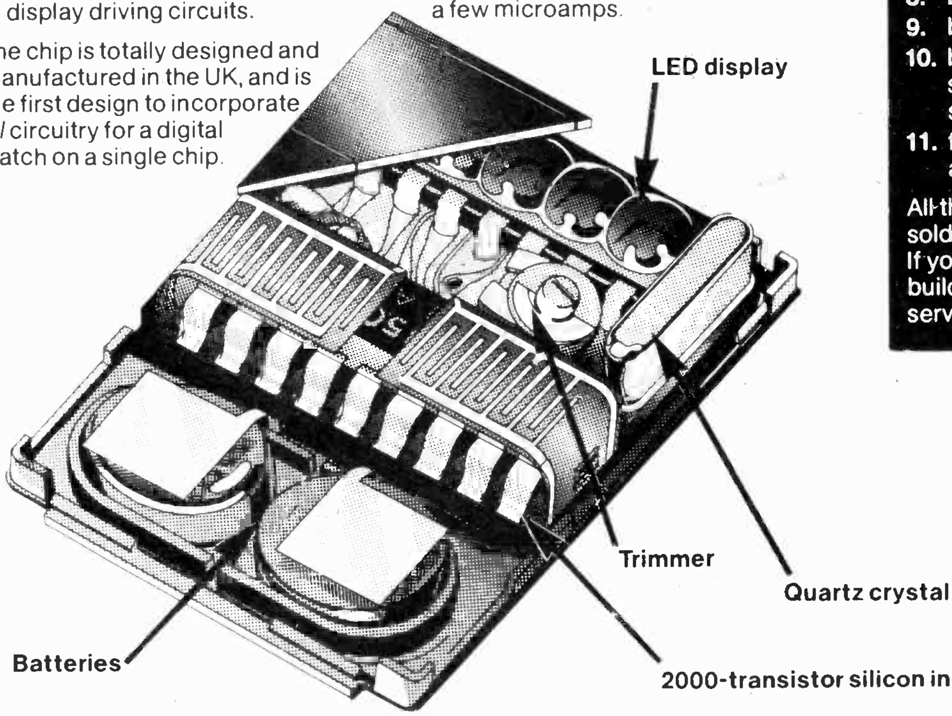
The chip...
The heart of the Black Watch is a unique IC designed by Sinclair and custom-built for them using state-of-the-art technology–integrated injection logic.

This chip of silicon measures only 3 mm x 3 mm and contains over 2000 transistors. The circuit includes

a) reference oscillator
b) divider chain
c) decoder circuits
d) display inhibit circuits
e) display driving circuits.

The chip is totally designed and manufactured in the UK, and is the first design to incorporate *all* circuitry for a digital watch on a single chip.

...and how it works
A crystal-controlled reference is used to drive a chain of 15 binary dividers which reduce the frequency from 32,768 Hz to 1 Hz. This accurate signal is then counted into units of seconds, minutes, and hours, and on request the stored information is processed by the decoders and display drivers to feed the four 7-segment LED displays. When the display is not in operation, special power-saving circuits on the chip reduce current consumption to only a few microamps.



Complete kit £14.95!

The kit contains

1. printed circuit board
2. unique Sinclair-designed IC
3. encapsulated quartz crystal
4. trimmer
5. capacitor
6. LED display
7. 2-part case with window in position
8. batteries
9. battery-clip
10. black strap (black stainless-steel bracelet optional extra–see order form)
11. full instructions for building and use.

All the tools you need are a fine soldering iron and a pair of cutters. If you've any queries or problems in building, ring or write to Sinclair service department for help.

Take advantage of this no-risks, money-back offer today!
The Sinclair Black Watch is fully guaranteed. Return your kit in original condition within 10 days and we'll refund your money without question. All parts are tested and checked before despatch–and correctly-assembled watches are guaranteed for one year. Simply fill in the FREEPOST order form and post it–today!

Price in kit form: £14.95 (inc. black strap, VAT, p&p).

sinclair

**Sinclair Radionics Ltd,
London Road, St Ives,
Huntingdon, Cambs., PE17 4HJ.
Tel: St Ives (0480) 64646.**
Reg. no: 699483 England. VAT Reg. no: 213 8170 88.

To: Sinclair Radionics Ltd, FREEPOST, St Ives, Huntingdon, Cambs., PE17 4BR.

Please send me Total £

.... (qty) Sinclair Black Watch kit(s) at £14.95 (inc. black strap, VAT, p&p). ....

.... (qty) black stainless-steel bracelet(s) at £2.00 (inc. VAT, p&p). ....

*I enclose cheque for £................................ made out to Sinclair Radionics Ltd and crossed.

*Please debit my *Barclaycard/Access/American Express account number

....

Name ____________________

Address ____________________

____

ETI/3

Please print. FREEPOST–no stamp required. *Delete as required

"I used to read 'comics'....until I discovered ETI."

"Now I get it regularly!"

There are only two ways to ensure a regular copy of ETI — place a regular order with your newsagent ot take out a subscription. If you don't believe us — take a look at the recent issues no longer available listed with the Index!

Help us to help you: please write your name and address on the back of your cheques.

To: SUBSCRIPTIONS DEPARTMENT,
ETI MAGAZINE,
36 Ebury Street,
London SW1W 0LW

I enclose £5.00 (£5.50 overseas subscriptions except Canada: $10) for the next twelve issues of ETI.

Name . . . .

Address . . . .

. . . .

. . . . March 1976

UP TO £500 TO BE WON

TEN PRIZES OF £12.50

worth of goods selected from Henry's Self-Service Centre or Components Catalogue (inc. VAT).

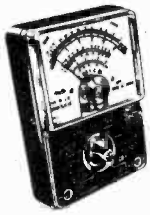
AND FIRST PRIZE OF £12.50

PLUS

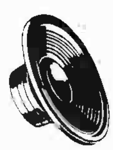
'SALE-OF-THE-CENTURY'

The first prize is £12.50 cash but the winner is entitled to a 90% discount if he spends the money on goods selected from Henry's Self-Service Centre or Components Catalogue.

EXAMPLES



Normally £8.10. 81p!



Normally £7.78 77p!

Normally £93.98 £9.40!

DOUBLE YOUR MONEY

ETI-HENRY'S COMPETITION

To celebrate the recent opening of Henry's Self-Service Electronics Centre at 404 Edgware Road, London W2, ETI and Henry's have arranged an easy-to-enter competition with prizes of at least £250 and up to £500 worth of goods.

First prize is £12.50 plus a 90% discount on goods purchased with this sum — or equivalent to £250 worth of goods if you affix to your entry coupon the price circle from the current Henry's catalogue.

Additionally there are 10 other prizes of £12.50 worth of goods, again worth double with proof of purchase of the catalogue.

Henry's have perhaps the best selection of components in the country and their new Self-Service Centre — the first of its kind in Britain — makes browsing and selection more enjoyable than it ever has been.

RULES

The competition is open to readers in the United Kingdom and Northern Ireland except employees of Electronics Today International and the Henry's Group of companies and their agents. There is no entry fee but all entries must be made on the coupon cut from the magazine. No other correspondence should accompany the entry. If you wish to show proof of purchase of the latest Henry's catalogue, entitling you to double the value in prizes follow the instructions given below.

The prizes will be awarded to the first eleven correct entries drawn after the closing date of March 31st, 1976. The first prize, entitling the winner to a 90% discount will be awarded to the first correct entry selected. A method of draw will be used which gives no preference one way or the other to those entrants opting for 'Double-your-Money'.

The prize value includes VAT. Readers may submit as many entries as they wish but each must be accompanied by an original coupon. It is a condition of entry that the judge's decision is regarded as final.

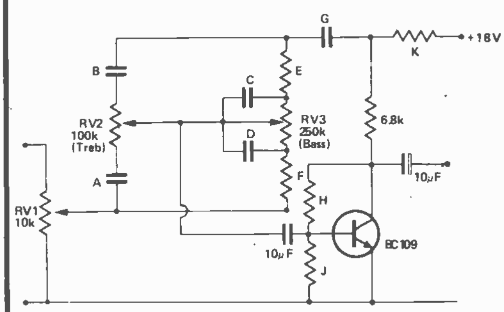
DOUBLE YOUR PRIZE VALUE!

If you stick or staple the '50p' price circle from the top right hand corner of the current Henry's Radio Components Catalogue to your entry coupon, the value of your prize will be doubled! The ten basic prizes will then be £25 worth of components and the first prize the same but still qualifying for 90% discount – making the first prize worth £250!

How to enter

The circuit shown below is of a fairly standard tone control. Many of the component values are missed out and replaced by the letters A–K. A list of the correct component values is listed.

Use your skill and judgement to allocate the components next to the letters on the entry coupon. The electrolytic in the circuit is deliberately shown as a non-polarised capacitor. Components should be allocated for the best conventional performance.



Capacitors	Resistors
2,7nF	3.3kΩ
2.7nF	15kΩ
15nF	22kΩ
15nF	22kΩ
10µF electrolytic	100kΩ

A .... F ....
B .... G ....
C .... H ....
D .... J ....
E .... K ....

To: ETI/HENRY'S COMPETITION
ETI MAGAZINE,
36 EBURY STREET,
LONDON SW1W 0LW.

Paste cat. price circle here if you try for Double-Your-Money

NAME..............................

ADDRESS..............................

....

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APRIL ETI

MORE DETAILS ON THE

CA3130

THIS CMOS OP-AMP WAS FEATURED IN DATASHEET LAST NOVEMBER. NEXT MONTH WE LOOK AGAIN AT THIS REMARKABLE DEVICE AND GIVE MORE APPLICATIONS CIRCUITS.

CONVERT YOUR CALCULATOR INTO A STOPWATCH

OR IF YOU WANT TO BUILD A CHEAP STOPWATCH BUY A CHEAP CALCULATOR AND A COUPLE OF CMOS ICS THEN FOLLOW OUR DESIGN. THE INSTRUMENT WE BUILT GAVE ±0.2% ACCURACY AND MEASURED UP TO 9999.99 SECONDS (OVER 2½ HRS).

SWITCHING REGULATOR TTL POWER SUPPLY

FIVE VOLTS AT TEN AMPS FROM THIS SIMPLE PROJECT. SWITCHING REGULATION GIVES MUCH MORE EFFICIENCY THAN SERIES REGULATION SO THE DEMANDS ON THE TRANSFORMER ARE LOWER.

ETI AND Marshall's SPECIAL OFFER

THIS WILL BE RUN ON LINES SIMILAR TO THE OFFER WITH MARSHALL'S LAST YEAR: BIG REDUCTIONS ON TTL, CMOS, 7-SEGMENT DISPLAYS, ETC.

ETI MICROFILE

KEEPS YOU IN TOUCH WITH THE MOST EXCITING NEWS IN ELECTRONICS. NEXT MONTH FIND OUT HOW ONE SMALL CHIP HANDLES PROBLEMS WHICH WERE PREVIOUSLY TACKLED ONLY BY ENORMOUS COMPUTERS!

TECH-TIPS & DATA SHEET INDEX

TECH-TIPS has now been running for four years. With our index you will be able to find creative circuitry, in almost any application, without having to search for hours.

SPACE ENVIRONMENT

The space capsule has to provide the astronaut with a comfortable environment despite the hostile temperatures, radiation, pressure, meteors, etc

APRIL ISSUE ON SALE MARCH 5TH

electronics today international

ANOTHER FABULOUS ETI OFFER

THE EXELAR Digital Watch £14.95!

INCLUDING VAT AND POSTAGE

The Exelar. Never heard of it? You certainly haven't heard of any digital electronic watch at this price.

Exelar is a new name for the Digital Watch products of National Semiconductor; in fact N.S. make all the electronics themselves for this new range and Exelar is a name that you'll probably get to know pretty well.

This offer is a prelude to the launch of this watch — and ETI once again has been able to arrange an enormous discount for readers. Response to offers in the past has shown that ETI readers are very quick to appreciate new technologies and the resulting products and this enables us to come up with the really good deals for which we are known.

The NW1-WS is a three-function watch: Hours, Minutes and Seconds with a typical accuracy of 5 seconds per month. The bright, easy-to-read LED display is one of the few in the industry which takes advantage of the crisp clarity of monolithic digits. The bezel is chrome plated, the back is stainless steel and the strap is leather.

The recommended retail price of the Exelar is over £20, itself setting a new record low for a built watch but using the coupon you pay just £14.95 including VAT and postage. Don't delay — an enormous response is expected.

For operation of the Exelar, press the button and the hours and minutes are displayed. Keeping the button depressed will display flashing seconds. The display stays on for 1½ seconds after touching the button. Time setting is accomplished by pressing a recessed push switch in conjunction with the display button.

AVAILABILITY

ETI Offers in the past have been so successful that initial estimates of quantity have sometimes proved to be inadequate. Several thousand units are being stocked but if orders exceed this, delays may result.

If you have *not* received your watch before publication of the April issue, see that month's News Digest for the situation.

Although orders will be despatched as soon as possible after receipt, please allow 35 days for delivery.

For security reasons, no stocks of the Exelar are being held at ETI offices for callers.

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ELECTRONICS IN GEOPHYSICAL SURVEYS PART TWO

by Martin Blanchard

THE THREE CHARACTERISTICS studied in modern airborne surveys are (a) Magnetic, (b) Radioactive and (c) Thermal. Last month we looked at the techniques and equipment used in magnetic surveys; this second part of the article covers the other two methods.

RADIATION DETECTORS

Gamma rays are detected by the ionisation effect that they have on some materials. Early radiometric work was done using Geiger tubes but these have now been superseded by scintillation type detectors (in airborne geophysical work). The basic detector consists of a crystal (usually sodium iodide) which gives a flash of light (usually in the blue to UV region) caused by a gamma ray being absorbed and giving up its energy.

The scintillation crystal is superior to Geiger tube for two reasons: (a) it is a denser medium so detection efficiency is improved, and (b) its 'recovery time' is short. This second factor is important because several gamma rays may arrive at the detector within a short interval.

The scintillations are amplified in a photo-multiplier (PM) tube which is in direct contact with the crystal. This assembly is then mounted in a light proof container. The process of amplification begins when photons emitted during a scintillation strike the photo-cathode and displace a quantity of electrons. Due to the voltage arrangement of the PM tube plates these electrons are drawn to the next cathode in the tube where they displace a greater number of electrons. In turn these strike the next cathode and so on up the cathode chain until the output appears as a voltage pulse at the anode. The combination of crystal and PM tube is termed a detector head and is shown in Fig. 1.

SCINTILLOMETERS

A scintillometer consists of a detector head, a power supply and a pulse counting circuit. PM tubes require an EHT supply of about 1.5kV but do not draw much current. Fig. 2 shows an early scintillometer which uses three detector heads. The heads are connected to a mixer circuit where the incoming pulses are chanelled on to one signal line. The incoming pulses trigger a blocking oscillator which gives constant amplitude and width pulses to feed an integrating circuit. The integrator output is displayed on an analogue recorder. Other scintillometers use a ratemeter (for counting the pulse rate) and a digital read-out. The scintillometer counts one pulse regardless of the energy of a detected gamma ray and is therefore often called a total count detector.

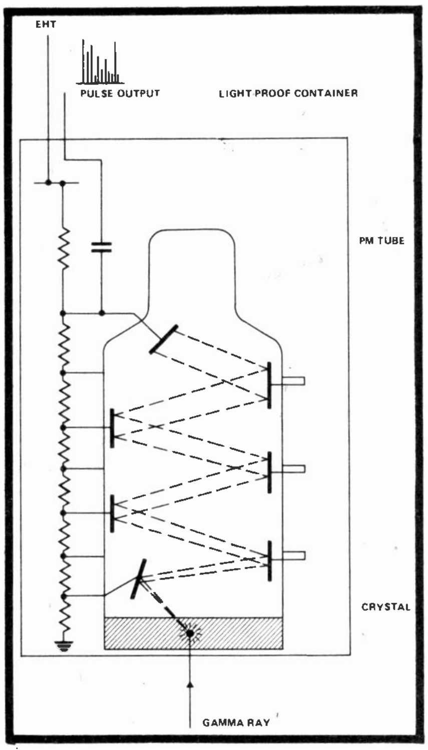


Fig. 1. Scintillation Detector Head.

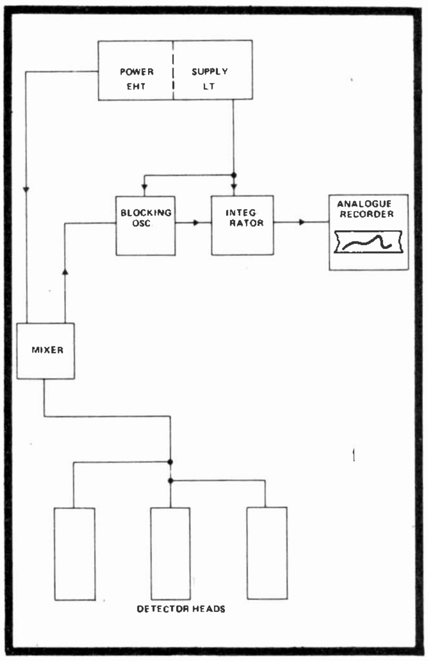


Fig. 2. Scintillometer.

SPECTROMETERS

A spectrometer is a scintillation counter with a facility for discriminating between the various energy levels of the gamma rays being absorbed by the crystal. The intensity of each scintillation is directly proportional to the energy of the gamma ray causing it.

Each part of the radiation spectrum which is counted is termed a 'channel'. Four channels are commonly used with the channel limits set as follows:

Channel 1: 0.9 to 2.9MeV
Channel 2: 1.36 to 1.56MeV
Channel 3: 1.66 to 1.86MeV
Channel 4: 2.3 to 2.9MeV

Channel 1 is usually called the total channel and the other channels are often named after the radioactive element that has a major energy peak (or peaks) between the various limits: Channel 2, Potassium; Channel 3, Uranium; Channel 4, Thorium.

A complete spectrometer is shown in Fig. 4. This has a common detection system comprising three units each of which contains three separate detector heads. The total number of detector heads is therefore nine, each with a 6 by 4 inch crystal. This arrangement provides a good sensitivity combined with a relatively small space requirement.

OPERATIONAL TECHNIQUES WITH SPECTROMETERS

Because of the attenuation of gamma rays in the atmosphere and in surface material the altitude at which spectrometers may usefully be operated is limited to about 200 metres and in general the best results are obtained at the lowest possible altitude. All the spectrometer equipment is carried inside the aircraft, including the detector heads. These must be carefully sited so that there is a minimum of scattering and absorbtion of gamma rays by the aircraft structure. Background radiation from the aircraft (for example from luminous instruments) must also be kept to a minimum. Survey areas to be covered using a spectrometer are flown in parallel lines like magnetometer surveys. The low altitude means that a narrow line spacing must be used.

INFRA-RED SCANNERS

IR line scanners are electro-optical devices which reflect radiation from the ground on to an IR detector. Much of the IR radiation spectrum is absorbed in the atmosphere but there are two windows at wavelengths of between 0.0003 and 0.0005cm and 0.0008 to 0.0014cm. The detectors used are manufactured from various semiconductor materials which have different responses in parts of the IR spectrum. The materials most frequently used are, for the 0.0003 to 0.0005cm region, Indium Antimonide (InSb) and for the 0.0008 to 0.0014cm region, doped Germanium. In order to reduce their internal thermal noise these detectors are cooled to low temperatures — Indium Antimonide to 77°K (−196°C) and doped Germanium to less than 10°K (−263°C). This cooling is achieved by the use of gases such as nitrogen and helium in their liquid state.

The detector output is usually recorded on a video tape recorder and on ordinary black and white film. The film is exposed to a light source that is modulated by the detector signal so that the density variations in the film are proportional to the IR radiation. The video tape recorder however, produces the detector output much more exactly.

Fig. 5 illustrates a practical line scanner. The actual scanning mechanism is rotated by a motor at about 100rps and has three main parts: the main mirror, the synchronizing device and the film scan motor. The main mirror is angled at 45° and reflects radiation from the ground into a focusing device which concentrates the radiation on to the detector. The radiation can only enter the scanner through a slot, representing a scan angle of about 120°, so the detector only receives a signal for one-third of a rotation cycle. The detector is cooled from an insulated flask containing liquid gas and the detector output is taken to a pre-amplifier stage. The pre-amplifier output goes to a gain control and the main amplifier and is also sampled by a gate controlled with a synchronizing pulse from a lamp and photo-electric device mounted on the scanner shaft. This device is triggered when the scanning mirror is 'looking' at the inside of the scanner housing so that the gate passes a sample of the pre-amplifier output when it is theoretically zero and any undesired signal which does occur, due to dc drift or changes in detector response, is integrated and fed back to the pre-amplifier input to adjust the no signal output to zero. After the gain control and main amplifier the signal is fed to another gate which is also controlled from a photo-electric device on the scanner shaft. This gate is 'opened' only during the 120° of scan rotation when a signal is present.

The final amplifier stage provides outputs to the monitoring oscilloscope, the video tape recorder and a voltage-to-current converter which modulates the film light source. The modulated beam from the light source is reflected on to the film by a mirror mounted on the scan shaft so that each scan builds up an image on the film. The oscilloscope display is

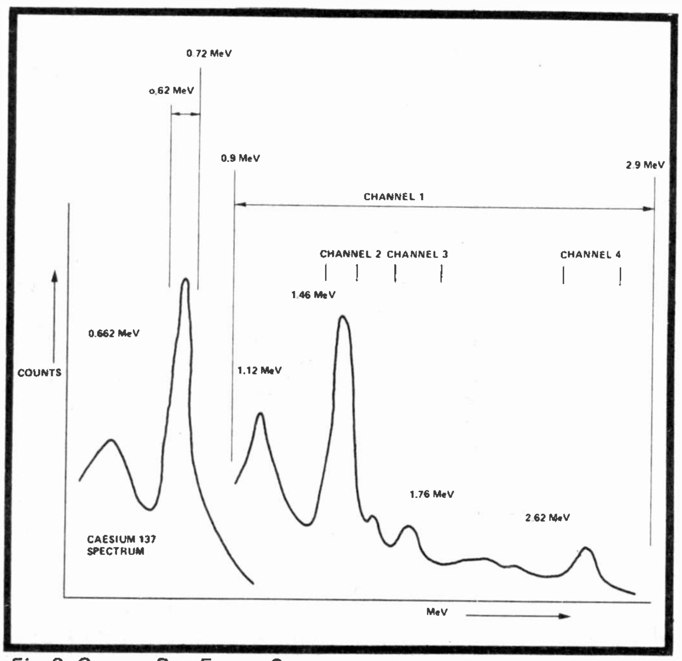


Fig. 3. Gamma Ray Energy Spectra.

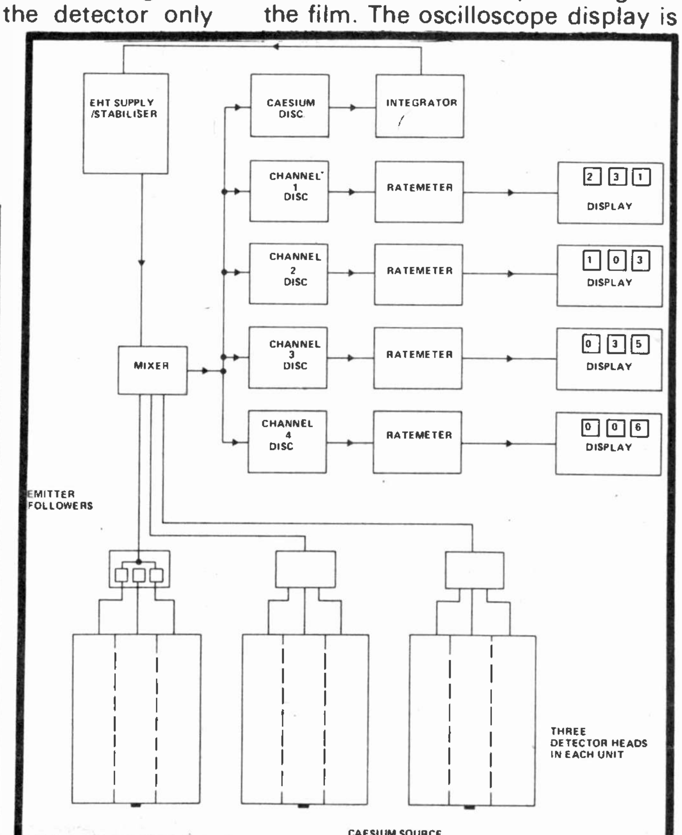
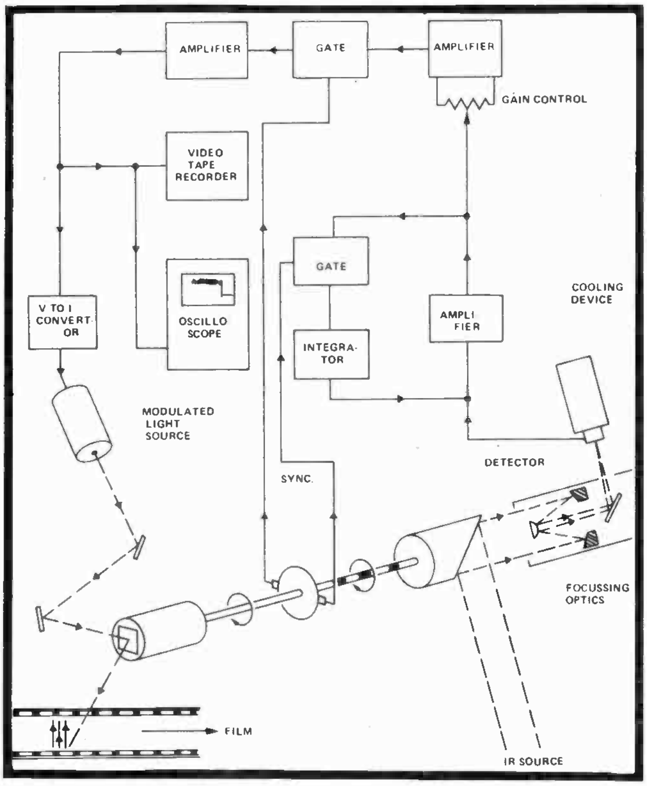


Fig. 4. Spectrometer.



viewed by the scanner operator to monitor the equipment performance and any necessary adjustment of the system gain.

Fig. 5. above. IR Line Scanner.

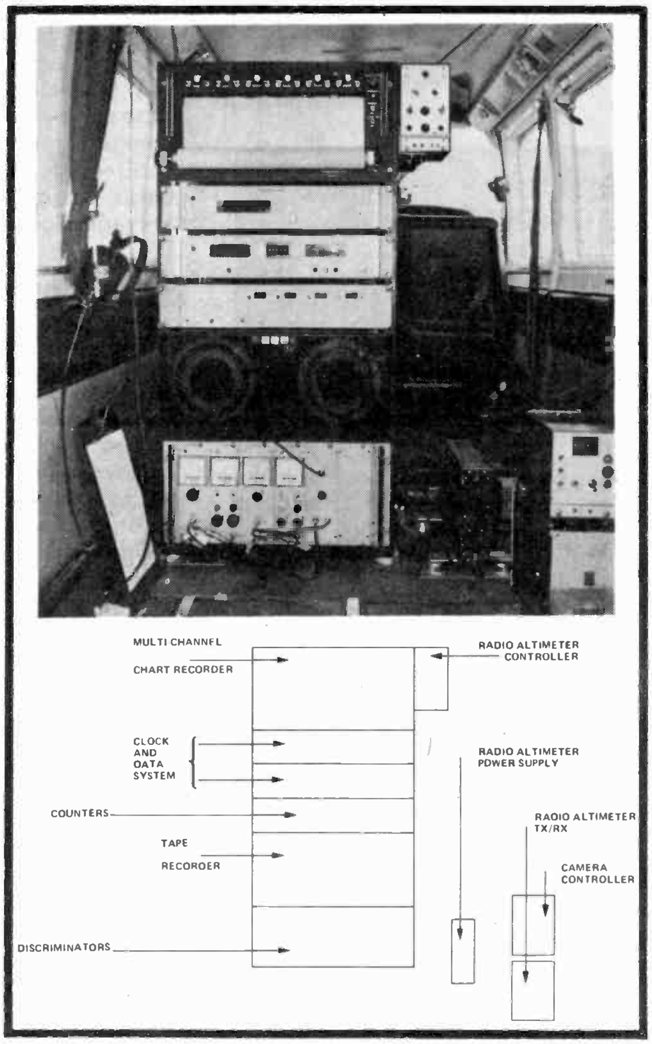


Fig. 6. Equipment used in a Spectrometer Survey.

OPERATIONAL TECHNIQUES WITH IR LINE SCANNERS

The height at which IR line scanners are operated is determined by the resolution required. The lower the altitude the better the resolution, although, of course, at low altitude a smaller area is covered in each scan. The minimum operating altitude is limited for any particular aircraft speed by the scan speed if complete coverage is to be maintained. Normally temperatures differences of about 0.5° C can be detected from a height of 500 metres.

The time of day when IR line scan surveys are flown is important, due to the different temperature effects occuring through a day. During the night temperature differences are often accentuated but any mist and fog tends to absorb radiation. Often a survey area is flown at two or more times of the day to provide more information for interpretation. Operation is usually stopped during high winds as these tend to 'smear' the temperature profiles.

NAVIGATION TECHNIQUES

The problem of navigating in survey areas is a critical one. Obviously geophysical results are not very useful if the location in which they were obtained is not known exactly. Often there are no accurate maps of a survey area and navigation must be carried out on mosaics of survey photographs (the accuracy of which is sometimes doubtful). A frequently used navigation aid is the Doppler navigator, which has the important characteristic that it requires no beacons or other external equipment. The Doppler navigator works by measuring the Doppler 'shift' of radio-frequency transmissions from the aircraft which are received back as reflections from the ground. Information derived from the frequency shifts and from the compass is used to guide the aircraft along a straight course.

The main navigation aid for checking where an aircraft has flown is the tracking camera which is run continuously while the aircraft is on a flight line and after a flight the processed film allows the flight path to be plotted on maps or photo mosaics.

All the geophysical instruments described here are flown so that a constant ground clearance is maintained. However, over rough terrain this is not possible using barometric altimeters so radio altimeters are used. A radio altimeter transmits a swept range of RF towards the ground and then receives the reflections. The transmitted and received frequencies are mixed to produce an audio frequency signal which is the difference between the frequency transmitted at a particular instant and the frequency being transmitted when that signal returns. This frequency difference is proportional to time (i.e. to the sweep speed) and therefore to the altitude of the aircraft. The audio frequency is converted to dc to drive a pilot s indicator and provide outputs for the recording devices. Radio altimeters have accuracies of better than ± 1 meter in 150 metres.

DATA SYSTEMS

In recent years it has become the practice to record all information from the navigational and geophysical equipment on magnetic tape. The instrument outputs are sampled at a fixed interval, the timing being controlled from a master clock which also registers on the tracking camera film to ensure correct orietation of

Rack of geophysical electronic equipment—Top shelf—Spectrometer console and mixer box. Centre shelf—Magnetometer console and various flight recorders. Background—Two spectrometer detector heads.

the survey data with the flight path plot. The taped data can only be read out if special facilities are available so the survey data is also recorded in analogue form on multi-channel chart recorders. This procedure also guards against un-noticed faults in the magnetic recording.

GENERAL OPERATIONAL TECHNIQUES

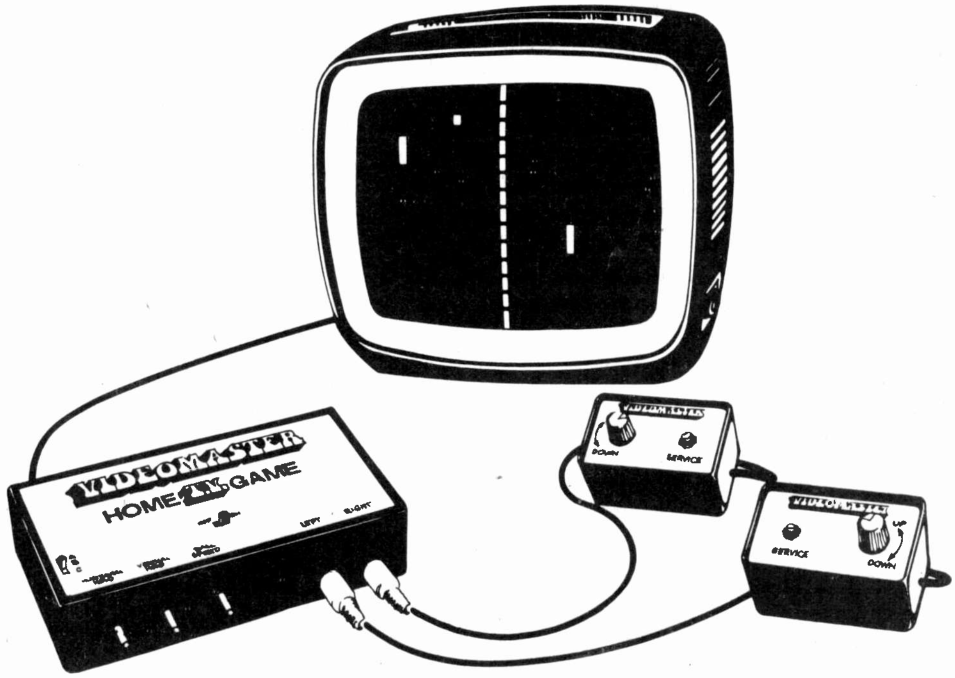
Generally the need for an airborne survey is decided upon by the possibility of a specific mineral appearing in a particular area. The geophysical instrument to be used is selected for its efficiency in detecting that mineral.

Once the basic instrument has been selected a second instrument is often chosen to provide additional information to aid in interpretation. Fig. 6 illustrates some of the equipment required for a spectrometer survey. The detector heads are not shown. The type of geophysical equipment selected governs the survey operating parameters. If a spectrometer is in use the maximum operating altitude is determined by the spectrometer detection sensitivity and the minimum operating altitude by safety considerations.

Frequently, spectrometer surveys are flown at an altitude of 150 metres but this means that to maintain adequate ground coverage the flight lines must be fairly close together, say at about 400 metre spacing. The flight lines are marked on to navigational maps or photo mosaics and are flown progressively with the aid of navigational equipment. Lines that are flown in the wrong place or are subject to an equipment failure are usually reflown at the end of the survey.

When all the survey area is satisfactorily covered the geophysical information is plotted on to the grid of flight lines and contour maps are drawn. Computerised data may be printed out in the form of contour maps. Depending on the type of survey the contour lines are of equal magnetic field or equal radiometric count rate and the maps are interpreted by geophysicists and geologists to predict the location of possible mineral deposits.

Field investigation of selected sites can begin immediately and so it is that airborne geophysical surveys enable the few commercially viable sites in a very large unexplored area to be located in a relatively short time. ●


VIDEOMASTER
HOME T.V. GAME
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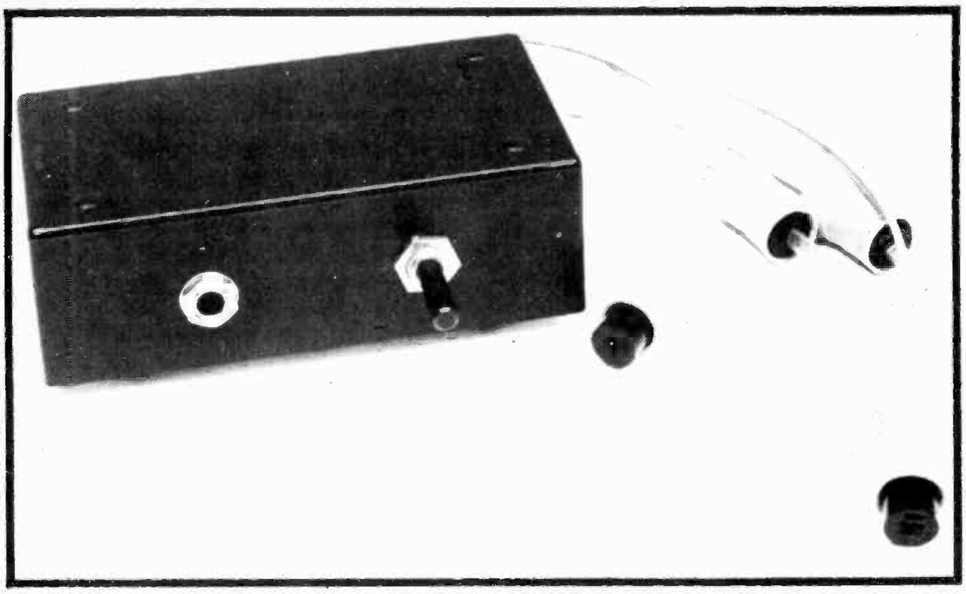


eti project 238

PARTS LIST — ETI 238	
Two	100 ohm 1 watt resistor
Two	10 ohm ½ watt resistors
Two	4-way screw type speaker connectors
One	100 ohm linear potentiometer
One	stereo headphone socket with double-pole break contacts
One	box (as required)
Two	2-pin DIN plugs (if required)
Two	2-pin DIN line sockets (if required)

HEADPHONE ADAPTOR

Unit restores damping and natural stereo separation.

HEADPHONES have impedances which range from 8 ohms to 2 k ohms or more and handle a typical maximum power of 500 mW. To limit the power that may be delivered into the 8 ohm types, commercial amplifiers generally supply the headphones from the amplifier output via series resistors of around 220 ohms.

Although this technique allows the use of practically any type of headphones without fear of damage the series resistor drastically reduces the amount of damping the amplifier can apply to the phones.

A further problem with headphone listening is that the stereo separation is unnatural in that there is little right channel information fed to the left ear and vice versa.

This simple little adaptor is inserted between the amplifier and the leads to the speakers. It restores damping, by supplying the phones from a 10 ohm source, and has a blend control by which the separation between channels can be varied to obtain a more natural sound.

CONSTRUCTION

We mounted the two four-way terminal strips onto the lid of a small box and the headphone socket and potentiometer through one side. The stereo-headphone socket should be the type which incorporates double-pole break contacts. This type is generally of sealed construction and has 8 pins on the back. Such a socket is necessary so that normal speaker operation is obtained until the headphones are plugged in whereupon the speakers are automatically disconnected.

Wire the unit as shown in the interconnection diagram, and, if the amplifier has DIN connectors, attach short leads, with DIN plugs and sockets, to the unit as shown in the main photograph. ●

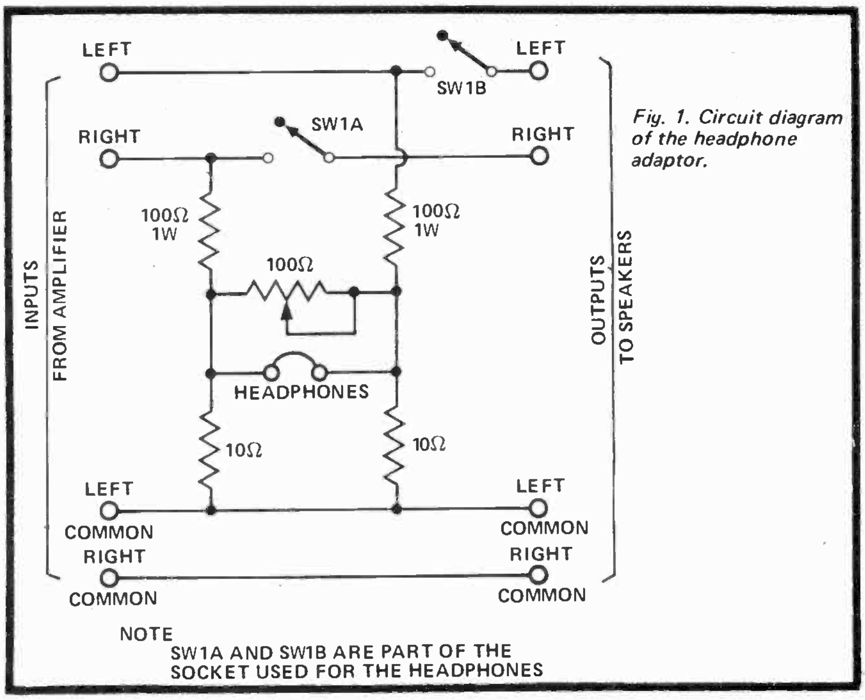


Fig. 1. Circuit diagram of the headphone adaptor.

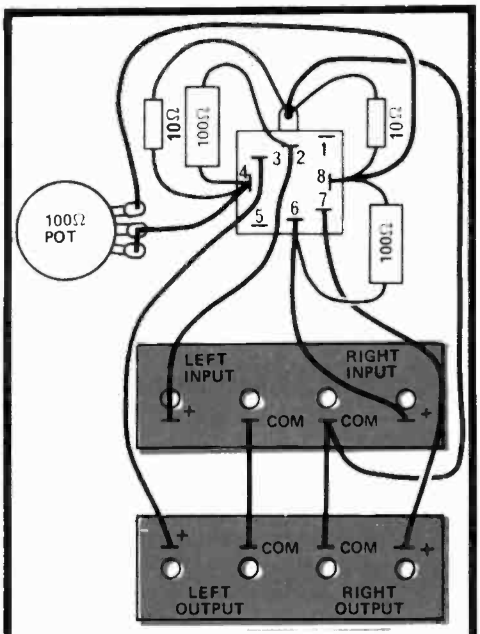
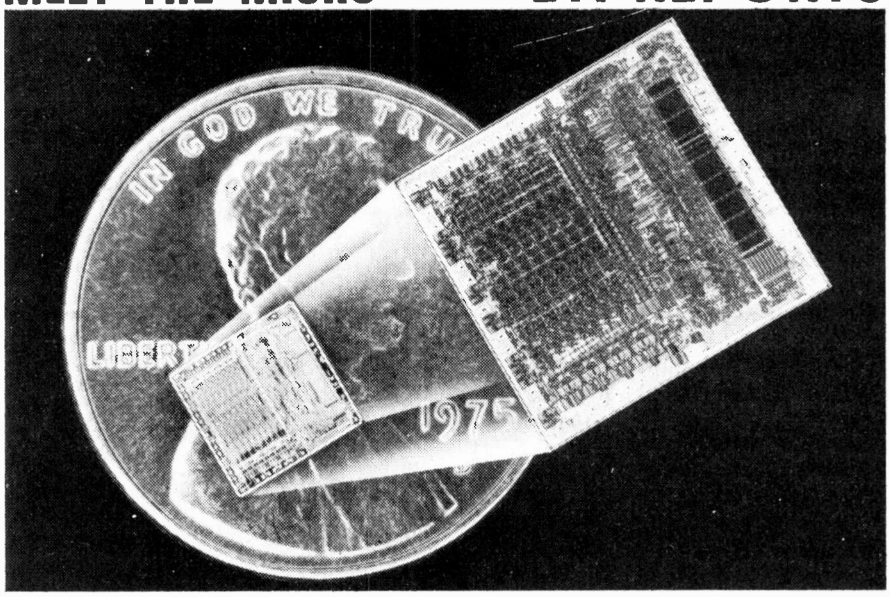


Fig. 2. Interconnection diagram for the adaptor.

eti microfile

MEET THE MICRO — ETI REPORTS

NO, YOU CAN'T USE IT FOR YOUR HOMEWORK . . . . I'M LOADING IN TOMORROW'S RUNNERS AT KEMPTON!



IN THE FIRST OF several articles under the title ETI Microfile, we'd like to introduce you to our star performer who is currently enjoying fantastic reviews in the States. We refer, of course, to the multi-talented microprocessor.

This device has been getting mentions in ETI for the past few months, (N.D., Desk Top Graphics), but we've never really said what it is, what it does, or how it does it. In fact, nobody really seems to have thought of passing on microprocessor information to the interested experimenter in a comprehensive way. Micros are rapidly coming down in price and up in applications — make no mistake, the micro is here to stay, very soon many electronics enthusiasts will be using one.

"But what will they be using it for?" we hear you ask. Here's a short list of potential applications:

- Programmable calculator
- Electronic games
- TV games
- Microcomputer
- Controlling heating systems
- Controlling measurement systems etc.

Many more applications will suggest themselves with a little thought. A fairly typical system might be built around a micro to control an electric cooker in a sophisticated way so a to permit programming of complete menus with the micro automatically controlling oven settings and timing — the applications are limited only by the imagination.

The main area which we will discuss is the microcomputer, as this is the ideal way to learn about microprocessors, and is a) interesting b) educational and c) useful. It is interesting to look at American activity in this area. Magazines such as *Popular Electronics* carry full page adverts from companies supplying complete computer kits, and programs to run on them. These computers, although small and cheap, are nonetheless very powerful — in fact on many of them it is possible to use BASIC language, which makes programming quick, easy and fun. Obviously, we'll have a lot more to say on the amateur microcomputer later.

FINE, BUT WHAT IS IT?

We still haven't said what a microprocessor actually is, so here goes. Please bear with us while we explain a bit about computer design. The micro has been described as an 'LSI computer on a chip'. This is a slight misnomer, and may lead you to expect more from a micro than it can reasonably give. It is perhaps more accurate to say that a microprocessor is the arithmetic, logic and control circuits of a computer implemented in LSI form. The definition given by one major microprocessor manufacturer is 'A Microprocessor is a very large scale integrated circuit which by the action of a sequence of instructions, externally programmable, can fulfil a wide variety of different electrical functions'. None of these definitions is particularly revealing, so let's take a closer look at what's actually in a computer and, by analogy, what's in a microprocessor unit (MPU).

The modern stored-program digital computer (Fig. 1) consists of three main units — the central processing unit (CPU), the memory and the input/output interfacing. The sequence of instructions (program) which the computer is to execute is passed through the input interfacing into the memory along with the data to be operated on, and step by step, the machine works through these instructions. Depending on the results of calculations, the machine can jump forward to the next step of the program; otherwise it may go back to the beginning with a new set of numbers.

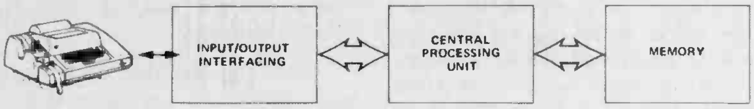


Fig. 1. The digital computer.

The instructions are fed, one by one, into the central processing unit which decodes them and 'decides' what they mean. The instructions are then executed by the arithmetic and logic unit of the CPU and the results may be tested to decide what to do next, or retained for use in the next calculation, or stored away in memory. (End of slight diversion into computers.)

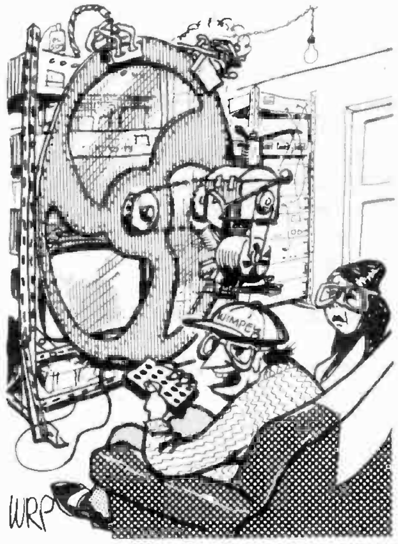
This, believe it or not, is precisely what a microprocessor system does — the microprocessor itself (MPU) corresponds to the CPU, and normally has associated memory containing a program for it to execute. The important point to note here, is that to change the application of an MPU, one does not usually have to change any components, but only the program; this usually costs little or nothing and programming is actually fun — we know our readers will take to it like ducks to water.

HOW DOES IT DO THIS?

From this point on, we have to assume that you, the reader, know a bit about digital logic, which you will, if you've been reading ETI for long. If you're not too sure, please, please find out now — it's definitely the coming thing — and we advertise some good books. We also have had to make a decision to work with one particular MPU, the reason being simply that if everyone works with different types, there will be a lot of duplicated and wasted effort, as really good and interesting programs worked out for one device won't work on a different type. After considerable spadework and deep thought we have opted for the Motorola M6800 family of devices, for a variety of reasons. Please note, however, that a lot of what we say applies to MPU's in general and not just the M6800.

Figure 2 shows a block diagram of a basic M6800 microprocessor system. The most noticeable thing on this diagram is the way all the devices are connected together by three buses — the control bus, the address bus and the data bus. The control bus has several lines carrying clock signals, read/write, halt, reset and other signals for the control of the system. The address bus has 16 lines in total $A_0 - A_{15}$, which means that the MPU can access up to 2^{16}, i.e. 65 536 bytes, (8 bit words) of memory directly. This amount of memory is equivalent, roughly, to 13 completely printed pages of ETI with all kinds of special characters, typefaces and colour specified in a none-too-efficient code. So you can see there is sufficient memory potentential for most applications.

The data bus has only 8 lines and will enable the MPU to input or out-



NOW I'VE GOT ALL THE INTERNAL CONTROL FUNCTIONS PRE-PROGRAMMED I CAN RUN IT UP FROM COLD IN WELL UNDER HALF AN HOUR!

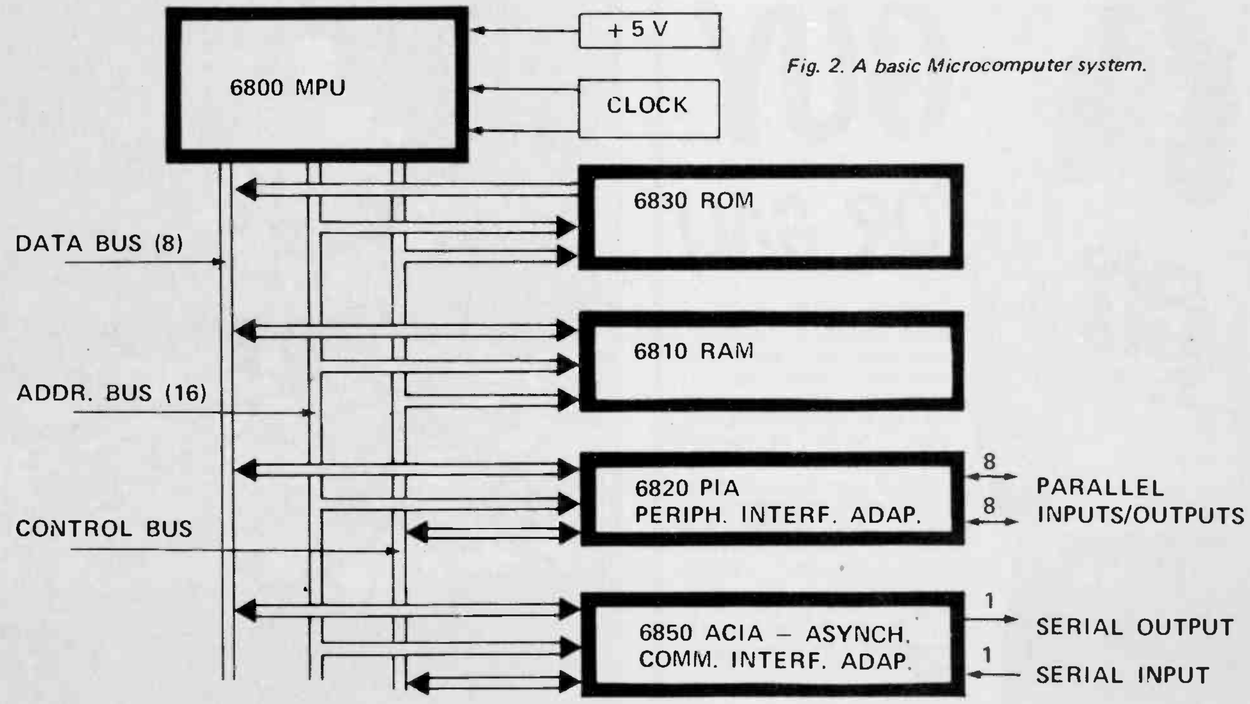


Fig. 2. A basic Microcomputer system.

Motorola's 'MPU in a suitcase'.

put any 1 of 256 different bit patterns. This will be covered in more detail when we discuss number systems. Both buses are bi-directional, so that the MPU can send or receive on either, with the added facility of disconnecting itself for special applications (Direct Memory Access, for those in the know).

THANKS FOR THE MEMORY

Let's now look at the two blocks below the MPU — the ones labelled ROM and RAM. We'll start with RAM, or Random Access Memory. This is the memory that holds both the program and the data it will operate upon, and the MPU can recall and store data in this memory as it works through a program. It functions in a manner similar to a rack of pigeon holes — upon selecting a particular pigeon-hole or *address*, the MPU can either read a word from it or stick in a word. The 6810 RAM shown in the diagram is a 128 byte semiconductor RAM chip made by Motorola to operate with the 6800 MPU.

RAM chips have one disadvantage — when the power is turned off, the contents are lost. This means that any programs which have to be permanently resident in the computer have to be in a different form of memory — this is usually Read Only Memory. This is programmed either during manufacture or before insertion into the circuit and cannot normally then be altered by the micro, which can only read information from it. Again, the 6830 is a Motorola part designed for the 6800 — many other parts are suitable, however.

INPUT, OUTPUT

Referring back to Fig. 2, the two, blocks at the bottom constitute the bulk of the input/output (I/O) interfacing mentioned earlier. The 6820 Peripheral Interface Adaptor has two eight-bit parallel buses each line of which can function as an input or an output, while the 6850 Asynchronous Communications Interface Adapter is a serial I/O device. Both these devices will be dealt with in greater detail later.

With the addition of a few devices such as a clock oscillator and a power supply, these devices can be made into a complete microcomputer. However you also need some kind of input/output device e.g. teletype or video display unit and also a monitor and debug program which will enable you to load programs into memory, run them and alter them. We'll have more to say on this when we discuss programming. The final (optional) piece of hardware is a low speed modem which converts digital signals into audio tones so that programs can be saved on a cassette tape via a cheap tape recorder. This means that programs do not have to be typed in each time the computer is switched on, and is probably the computer experimenter's best value-for-money labour saving device.

Well, that rounds off this general, and very simple, introduction to microprocessors. If you're already using one, and have any interesting thoughts, we'd like to hear from you. If you're just thinking about it — watch for the Microfile heading — that's where you'll find the facts.

SILENT A-B SWITCH

Speakers may be A-B tested using this simple modification to our tone-burst generator

WHEN evaluating speaker systems in A-B listening tests, the first few seconds of listening convey the truest impression of sound quality. Listening for longer than a few seconds not only fails to give further information, but may well give a false indication. For this reason it is usual to switch rapidly between the reference speaker and the speaker under test. This is generally done by using the amplifier's A/B speaker selector switch, or by wiring a change-over relay in the speaker wiring.

Whilst such switching methods are simple and reliable they have one major drawback. That is that switching may take place at any point in the waveform and as a consequence switching transients may be introduced which tend to mask the subtle differences for which one is listening. Hence a method of switching at zero-crossing points would be of great value.

When the ETI Tone-Burst Generator was constructed it was realised that it contained all the circuitry needed to performance this switching task and that it could be modified to do so very simply.

The switching must be done at low level and hence the unit is used at the input of a stereo power amplifier. The reference speaker and the speaker under test are each connected to one channel of the amplifier and the silent switch switches the input to the amplifiers as required. Thus the arrangement is mono only but this is all that is required to assess the transient response and performance of a speaker in comparison to a reference speaker.

HOW IT WORKS – ETI 124 AB

As this unit is based on the operation of the tone-burst generator ETI 124 described last month, that article should be thoroughly read first. Only the changes necessary to that unit are detailed in this article.

Whilst an A-B switch would be a little simpler if designed specifically for that purpose, the modifications required to the tone-burst generator are so simple that we thought it not worth while to design a special circuit.

To make the generator act as an A-B switch it is necessary to disable the existing mode switch. We do this by plugging in an external control switch, SW6, via a stereo phone socket. The phone socket has two change-over contacts fitted which are used to disconnect the plus and minus six volts supplies from SW3 when the jack is inserted. One of the phono contacts also disconnects the plus six volts from the common of the socket when the jack is removed. As the common of the socket is required to be at plus six volts the phono socket must be insulated from the front panel which is at 0 volts.

The control switch, SW6, effectively shorts either R4 or R5 thus stopping the pulses from C2 or C3 triggering the flip-flop. When the switch is actuated there is a delay until the number of cycles as set by the front panel switch have occurred and then, at the next zero crossing, the change-over occurs. The delay is necessary to ensure that any contact bounce of the SW6 contacts does not cause unwanted switching of the circuit.

CONSTRUCTION

The ETI 124 Tone-burst Generator should first be constructed, as detailed last month, except that the wiring to SW3 is changed as detailed in Fig. 1 and 2 of this article. The dual phono socket and the phone socket are then mounted on one side of the box. If a metal box is used make sure that the phono socket is insulated from the case of the box as it is at a potential of six volts. The switch, SW6, should be mounted in a small pill container or similar housing and fitted with a three-core cable that is terminated at the other end by a stereo phone jack. Note that the common of the switch should be connected to the common of the jack but that the other wires may be wired to either of the remaining contacts.

USING THE SWITCH.

The audio switch requires a reasonably high level of signal to ensure correct zero-crossing switching. There are two suitable points in a conventional amplifier. The first position is between the tape-in and tape-out sockets but the second and preferable position is between the pre and main amplifiers provided that the main amplifier has a volume control that is independent of the preamplifier.

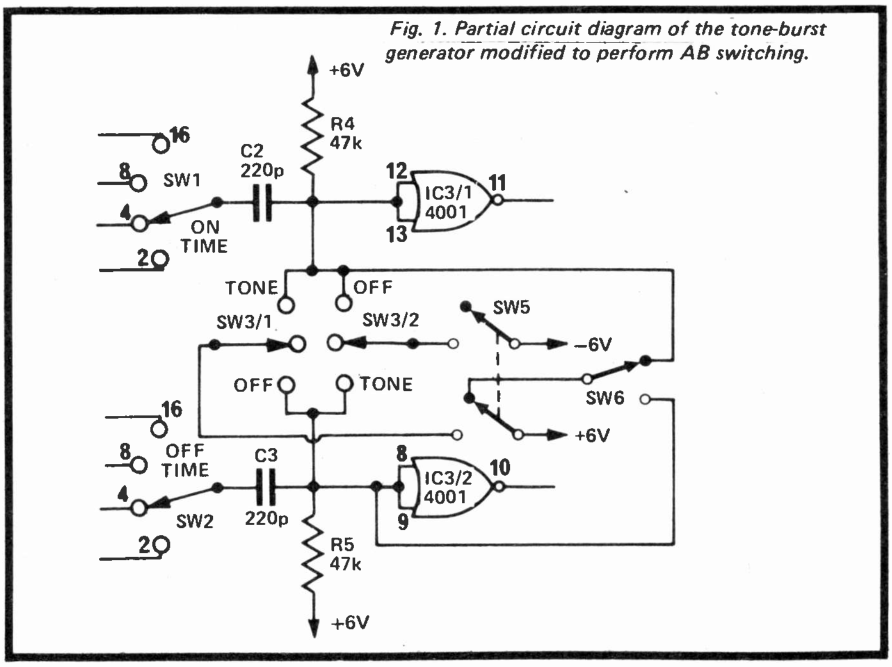


Fig. 1. Partial circuit diagram of the tone-burst generator modified to perform AB switching.

eti project 124 AB

SILENT A-B SWITCH

To connect the unit for AB testing apply a single input, from the preamplifier (switched to mono), to the normal input socket of the generator. The normal output socket of the generator is not used but the two phono output sockets are connected back to the left and right channel inputs of the main amplifier. When SW6 is operated the mono input will be silently switched between right and left channel speakers.

If using the tape sockets the monitor switch should be in the 'monitor' position and the balance control should be adjusted so that the levels from the two speakers are apparently the same. Make sure that the tone controls are in the flat position, as they can cause phase shifts which prevent the switching occuring at the zero-crossing point.

If the pre and main amplifier terminals are used the preamplifier volume should be adjusted to about half way and separate volume controls used to balance for the difference in efficiencies of the two speakers. If the main amplifier does not have separate volume controls then external ones must be added if balance is to be achieved. In this case the tone controls may be used if required without upsetting the crossover point.

Change over may be effected by using either a toggle switch or a push button. The tone-burst generator controls should be set for eight cycles on and off as this position will effectively remove any contact bounce. ●

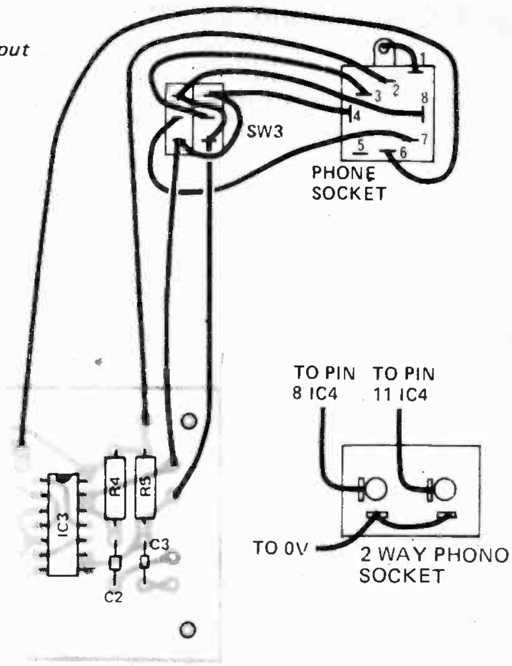


Fig. 2. Interconnection diagram to phone socket and phono output sockets of AB switch.

[CMOS]

THE 4016A

THE NEXT DEVICE we are going to look at has already made a considerable impact on amateur electronics. The 4016A quad bilateral switch consists of four transmission gates of the type discussed in the last part, each with its own control input. Each switch also has a signal input and output (although these are interchangeable). When the control input is held high the input to output path behaves like a pure resistance of about 300Ω but when it is low the equivalent value is of the order of 10^9Ω at low frequency, even with fairly low supply voltages. It is impossible to give all the data which might be necessary for diverse applications here and data sheets from a manufacturer may be required by the more adventurous experimenters. In any case the pinout diagram in fig. 1 should now be self explanatory.

It should be appreciated that the output impedance of the switch is fairly high and so for low signal distortion, a load greater than 10kΩ is necessary. Using a high supply voltage (10-15V) also helps to achieve this end. The gates will pass signals above the 10MHz mark but as the frequency becomes higher, crosstalk between the switches and distortion will inevitably increase. It should be fairly clear how complicated switching systems may be realised but fig. 2 has been included to guide constructors along the right lines.

ANALOGUE APPLICATIONS

Many uses of this device in audio equipment have already appeared in constructional articles in this magazine and so it is to two slightly less obvious applications that we shall turn now. Fig. 3 shows a sample and hold unit; when the control input is high the output tracks the input but when it goes low the output remains frozen at the value it was at the instant of transition. The operation of the circuit is generally self evident and it may be regarded as two voltage followers, one consisting of two op-amps with the output following the input, the other is just the second op-amp which "follows" the voltage stored on the capacitor. It is advisable to take care with the layout as with all op-amp circuits due to the huge open loop gain of these devices. The value chosen for C is a compromise between "slewing rate," that is the rate at which the circuit tracks a sudden change of input and "holding ability"

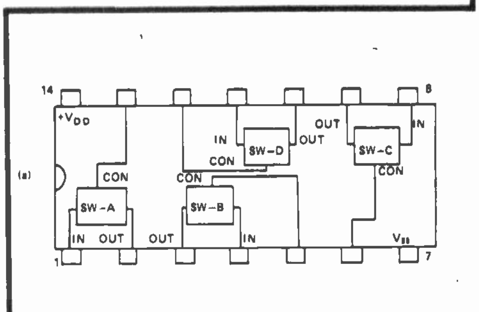
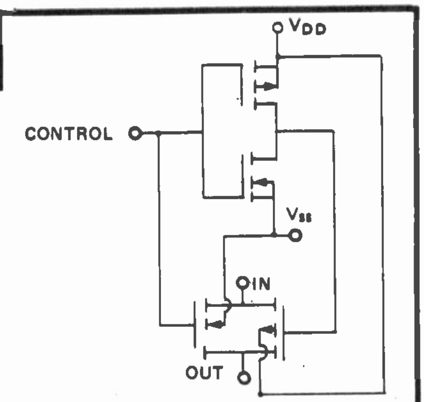


Fig.1 (a) Pin-out diagram of 4016A.



(b) Circuit diagram of a single switch.

which is the length of time the circuit will hold a signal without unreasonable decay. To give some sort of guide, for a 10kHz square wave to the control input, a 0.01µF capacitor seems to optimise the performance. The value of the resistors is also worth experimenting with.

An extension of the sample and hold concept is the analogue delay line which is shown in its basic form in fig. 4. The sequence of amplifiers and gates can be extended to any desired length to achieve a longer delay, the only limitation being that in extreme cases the control lines may need to be buffered. It should be observed that alternate stages of the circuit are driven by an identical clock waveform and so the circuit works by shifting the voltage on alternate capacitors during alternate clock phases. The value of the passive components and the clock frequency will have to be optimised for specific applications, low frequencies give long delays but high distortion.

DIGITAL COMPONENT SELECTION

There are a few fairly straight-forward uses of the 4016A in digital component selection which we will mention here because, in certain fields, they are very useful. Fig. 6 shows how to produce digitally controlled resistance and capacitance networks which will vary the magnitude of the quanity in question from its basic value up to 2^n-1 times that amount, where "n" is the number of gates and binary control bits. The resistor network can be used to produce a digitally gain controlled amplifier by placing it in the feedback loop of an op-amp and this can be used as a staircase generator as well as to produce more interesting waveforms. One application of the digital capacitor is to produce a digitally controlled sweep generator by using it as the capacitor in one of the multivibrators we discussed in the last part.

Clearly, any type of component may be switched in and out of circuit by the 4016A. One possibility that is useful in some circumstances is to use the information on filter design in "Electronics — it's easy" to produce digital filters of different descriptions. The main thing to remember when using all these ideas is that the impedance of the component that is being switched must, at the desired frequency of operation, be large compared to the 300Ω of the 4016A gates.

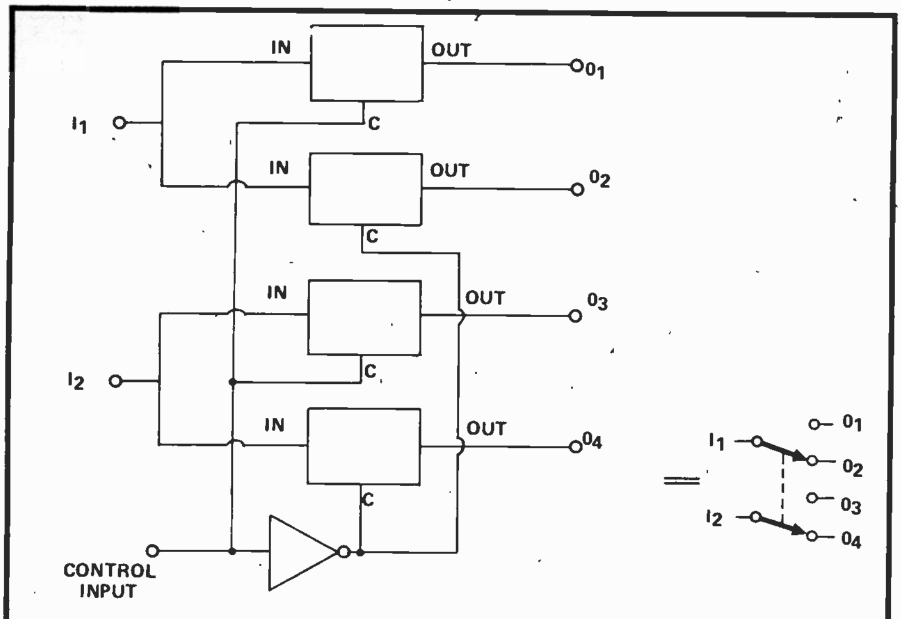


Fig. 2. DPDT electronic switch realised using a single 4016 and an inverter.

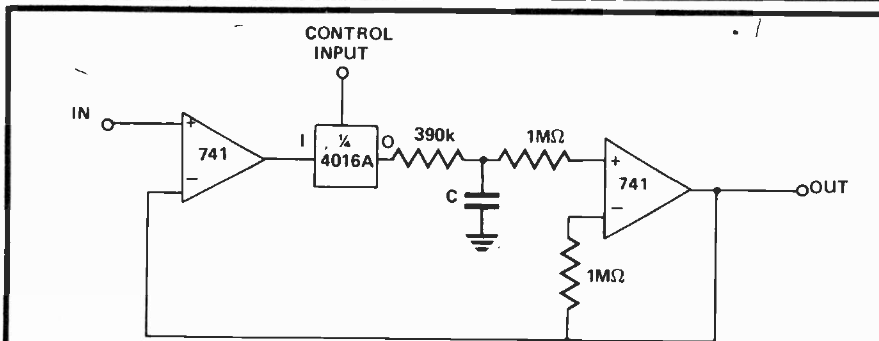


Fig. 3. A simple sample and hold circuit.

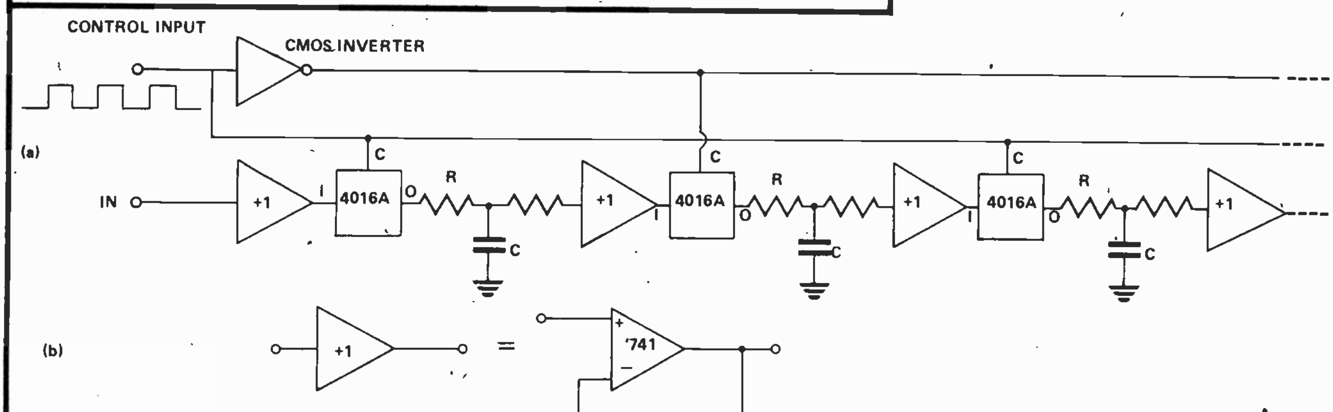


Fig. 4. An analogue delay line (a) Basic circuit and (b) composition of the voltage followers used.

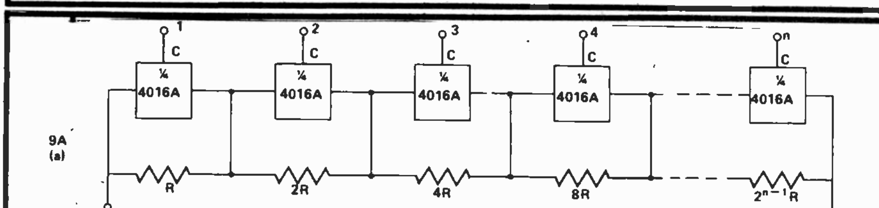
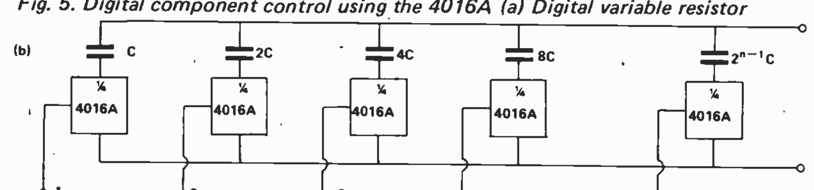


Fig. 5. Digital component control using the 4016A (a) Digital variable resistor



(b) Digital variable capacitor.

D-A AND A-D TECHNIQUES

The next use of the quad bilateral switch we are going to consider is in the field of digital to analogue conversion but first we are going to look at the subject of conversion on its own. Fig. 6 shows a D-A converter of the type known as a weighted resistor network. The working of this circuit is easy to see and the reason it is practical in so simple a form in CMOS is that the high and low output levels from a simple gate are within a few millivolts of the supply line thus providing accurate voltage levels to the summing resistors. This simple form has disadvantages, one of which will become immediately apparent if you consider the diversity of resistor values required for an eight or nine bit version. It is for this reason that the R — 2R network shown in fig. 7 is more popular for most applications. It is difficult to explain how this configuration works without becoming involved in mathematics but it becomes fairly clear if a three bit version is written down with the voltages in and analysed.

The basic idea for one sort of A-D converter or encoder as they are often called is shown in the block diagram of fig. 8. This variety is called a binary ramp encoder and is similar to the ramp integration method often encountered in modified form in digital measuring equipment. The principle of operation of our binary version is that a binary counter counts up from zero until it reaches the equivalent of the analogue input. Contrast this approach with the continuous counter method outlined in fig. 9. This provides a continous approximation to the input and is thus generally more useful for continuously monitoring a single channel

CMOS

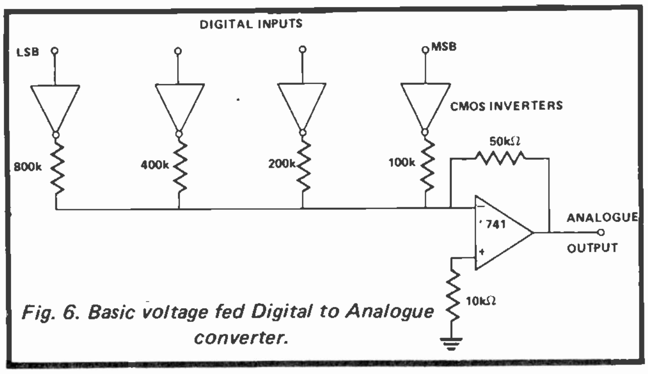


Fig. 6. Basic voltage fed Digital to Analogue converter.

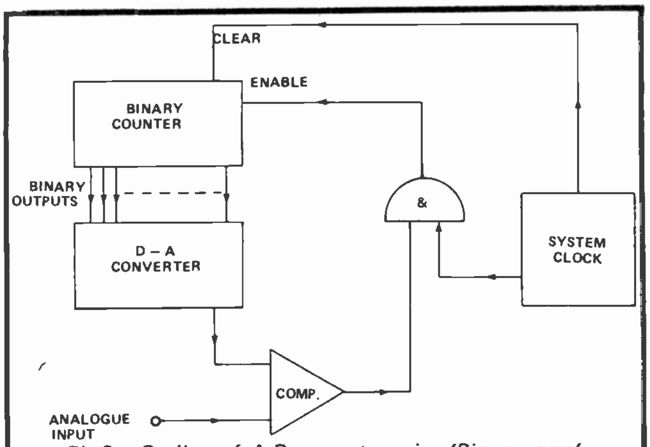


Fig.8. Outline of A-D converter using 'Binary ramp' method. The binary counter counts up to the required value so that the output is correct during only part of the cycle.

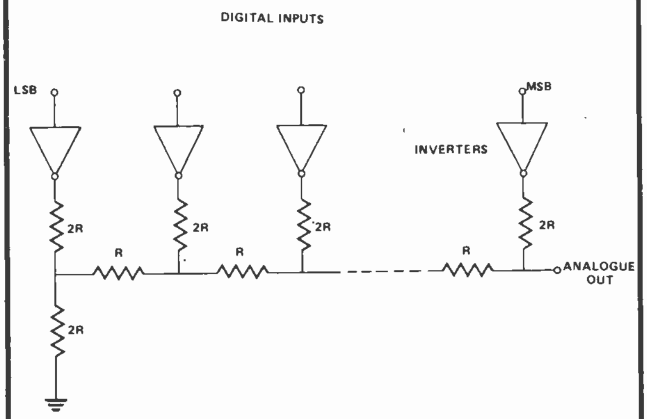


Fig. 7. The R–2R ladder network D–A converter.

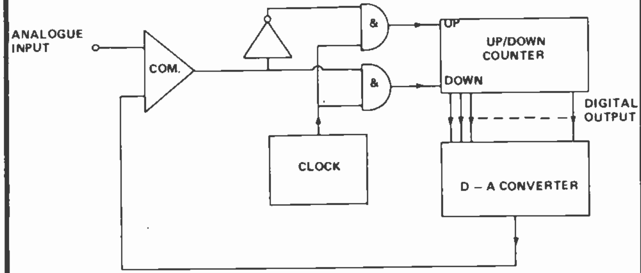


Fig. 9. A continuous counter type A–D converter. The clock just prevents the system oscillating at high frequency.

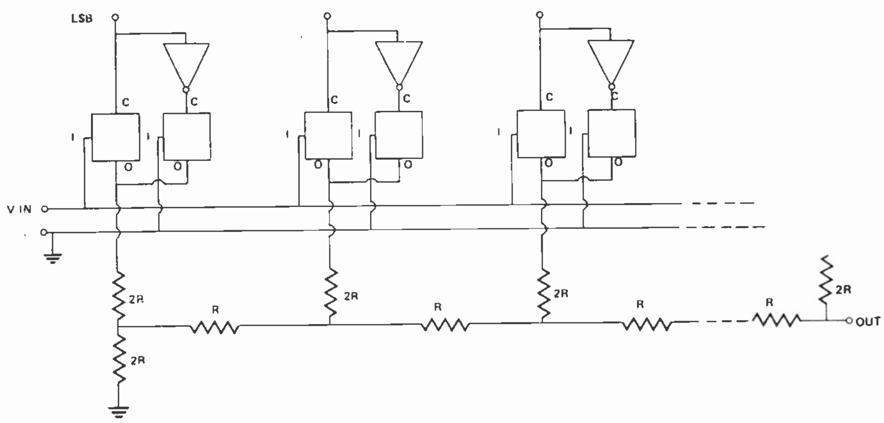


Fig.10. An analogue-digital multiplier – $V_{out} = V_{in}$ x Digital Input.

of information. It would be an advantage in many cases to have the counters working in BCD for ease of readout but this leads to complications which can not be gone into here. Finally on this subject it should be pointed out that these are circuits for experimentation and are unlikely to be directly applicable to any given situation. They have been included because of the ease with which they may be realised in CMOS compared, say, to TTL.

A-D MULTIPLIER

As far as digital to analogue conversion is concerned, using the 4016A we can take the idea a little further. What in fact we do (fig. 10) is to use an arbitary analogue voltage to feed the resistor ladder and so we multiply this input by the digital input and produce an analogue result. This "hybrid multiplier" is an interesting circuit, particularly because the analogue input voltage may be ac. thus producing several interesting waveforms and, on a more serious note, it may find application in hybrid computing experiments. We shall now leave the 4016A having, it is hoped, suggested some of the slightly less obvious uses of this versatile IC.

Continued next month . . . .

DYNAMIC NOISE FILTER

PART TWO

by M.G. Strange

THE FILTER output is coupled to the detector via a small capacitor to make the low-frequency rolloff even steeper below 1.6 kHz. The precision full-wave detector uses diodes in the feedback circuit of an op-amp to effectively produce ideal rectification characteristics down to the millivolt region. The output amplifier doubles as a post-detection filter. Resistor R determines the gain, and capacitor C makes this stage behave as an operational integrator with time constant RC. A switch is provided for increasing the time constant by paralleling capacitor C1; this is helpful with sources having sharp impulse noise. The output of the detector/filter circuit controls the bandwidth of the dynamic suppression filter according to the curve of Fig. 9.

Early experiments showed that it is undesirable to make the no-signal cutoff lower than absolutely necessary to substantially reduce noise with a particular signal source. When the cutoff is made lower than actually needed, weak signals are unnecessarily band-limited and the dynamic filter produces such a level-dependent bandwidth contrast that its action is much more likely to be audible. Hence a BASE CUTOFF (not "BASS CUTOFF") control was found to be desirable. This control is simply a pot which offsets the detector output at zero signal level by applying a variable reference voltage to the op amp non-inverting inputs. This voltage, variable from about −1 volt to −6 volts, establishes a "starting point" or base cutoff frequency which can be set just low enough to virtually eliminate no-signal noise.

VARIABLE-CUTOFF FILTER

The variable-cutoff filter, Fig. 8, is the very heart of the system. Since there is some part selection and adjustment necessary, it must be checked out separately. The basic configuration is similar to that of the pre-filter, except the latter's switch-selected resistors have been replaced by field-effect transistors (FETs). FET channel resistance R_{DS} changes as a function of gate voltage V_{GS} as shown in Fig. 11, thus varying cutoff frequency. A resistor across each FET establishes a solid lower cutoff limit and smooths the control characteristic as the FETs approach their "off" state. The gate circuit network, consisting of diode D1 and resistors R1 through R5, is used to empirically shape the control curve (Fig. 9) for best audible results. Diode D1 prevents excessive positive gate drive, maintaining isolation between the gate and signal circuits.

An input attenuator (R10 and R11) limits the signal amplitude presented to the FETs to about 0.1 volt p-p at 0 VU to ensure low distortion. Output amplifier A7 makes up exactly for this loss. An op amp having external frequency compensation was used here so that this relatively high-gain stage could be tailored for flat response to 15 kHz (a μa741 could be used, but would roll off slightly above 10 kHz). Resistors R16 and R17 attenuate the output signal by an amount equal to the gain, so that this amplifier doubles as the unity-gain buffer required for filter operation. The highest cutoff frequency is dictated by minimum FET resistance and capacitors C1 and C2. The latter should have values in a ratio of about 3:1 to produce the desired Butterworth response. Figure 10 shows the measured response of the complete filter for four values of control voltage.

Unfortunately, FETs vary widely in characteristics, even between units of the same type, so these devices must be selected. The two FETs must be reasonably well matched over at 15:1 R_{DS} range for a 15:1 range in cutoff frequency (15 kHz to 1 kHz). (Dual matched FETs are available, but are more expensive and not necessarily matched for the parameter of interest here.) A transistor curve tracer is most convenient for this purpose and permits selection for best linearity as well as matching. I used N-channel 2N4220s on hand (60p each) and selected the best matched pair out of a group of six units. Figure 11 shows the VI characteristics of one of these. There are many other inexpensive FETs which should work as well, such as the 2N5484, 2N5716, and 2N5717. In fact, any general-purpose, depletion-type FET with fairly low zero-bias current (I_{DSS}) and pinch-off voltage (Vp) should be usable. P-channel units would require reversing diodes D1 and D2 and the polarity of the control voltage.

If a curve tracer is not available, the

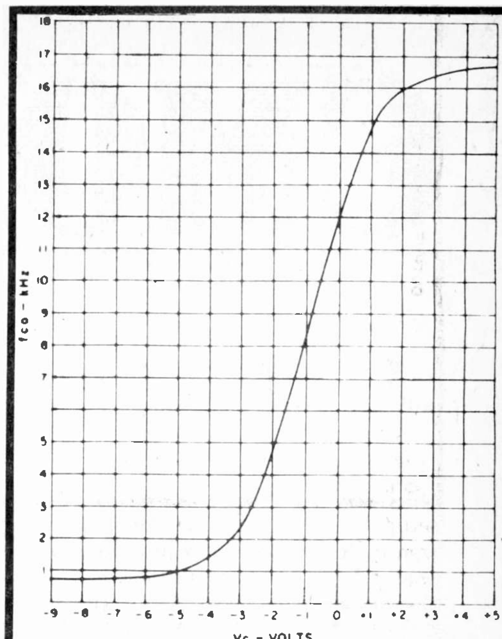


Fig.9. Variable-bandwidth filter cutoff frequency vs. control voltage.

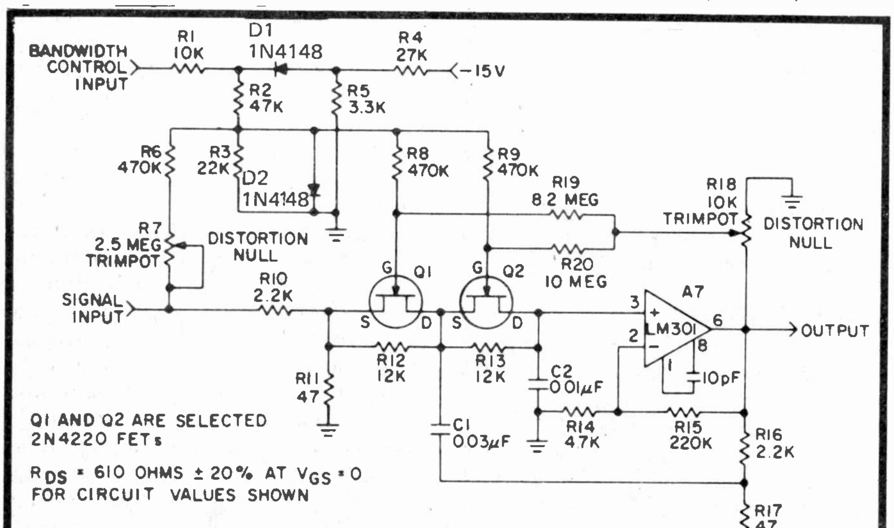


Fig.8. Voltage-controlled filter schematic. FETs Q1 and Q2 are critical and must be selected (see text).

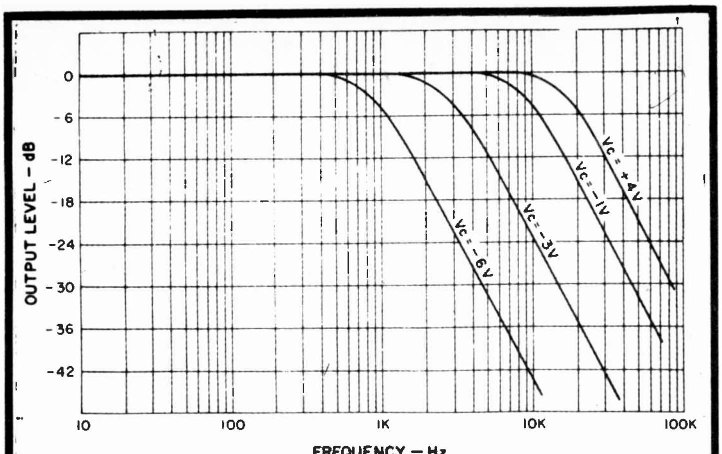


Fig. 10. Variable-bandwidth filter characteristics for several control voltage values.

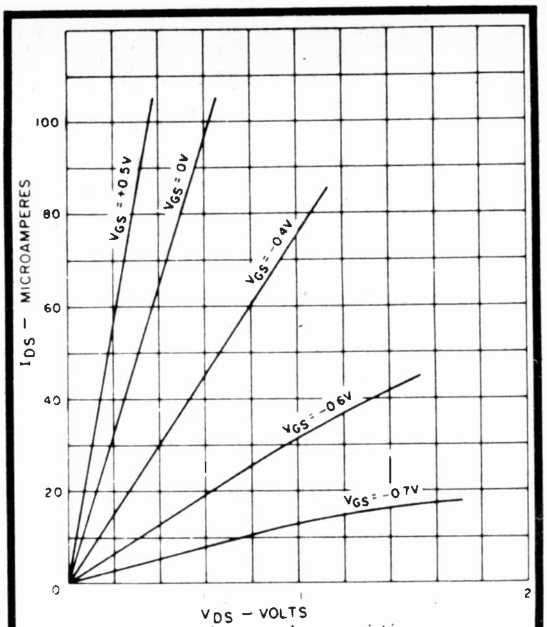


Fig. 11. Variable-resistance characteristics of a junction field-effect transistor with low values of drain-to-source voltage.

setup of Fig. 12 can be used. A transistor socket will facilitate changing FETs. A good procedure is to first measure R_{DS} at $V_{GS} = 0$. Then increase V_{GS} (negatively for N-channel FETs) until R_{DS} is about three times the zero-bias value; this corresponds to a mid-range cutoff frequency where matching is the most critical. With this V_{GS} setting try different FETs until a 10 percent or better match is found. If R_{DS} values seem to cluster higher or lower, try another unit as a reference and try matching to it. When matched units are found, check the match at minimum R_{DS} (V_{GS} = +0.5V) and at 10 times this value of R_{DS}. A 20 percent mismatch can be tolerated at these extremes. My 2N4220s measure 610 ohms at zero bias, 360 ohms at V_{GS} = + 0.5V., and about 8 kilohms at V_{GS} = –0.7 V. R11 and R12 are chosen for a cutoff of between 800 Hz and 1 kHz with the control voltage at its maximum negative value of about –6 volts. Circuit cutoff at zero FET bias should be roughly 12 kHz (see Fig. 9). A slight forward bias, limited to about +0.5 volt at the FET gates by diode D2, then boosts the cutoff to at least 15 kHz with maximum positive output from the precision detector.

Resistors R6, R7, R18, R19, and R20 reduce harmonic distortion significantly. R6 and R7 feed some signal to the FET gate circuit so that signal voltage does not appear between source and gate, which would make R_{DS} vary slightly with instantaneous low-frequency signal amplitude and polarity. R18, R19, and R20 feed back some output signal to the gates to further reduce distortion (this is a cancellation effect, not true negative feedback).

Distortion settings are best made in the vicinity of cutoff, where FET linearity is the most critical. Connect a variable-voltage d.c. source (the slider of a 5 k pot temporarily connected between – 15 V and ground will suffice) to the bandwidth control input and set it for a cutoff frequency of 2 kHz. Then, with a 2 kHz sinusoidal input at about 0 VU (2.2 V p-p), set trimpots R7 and R18 for lowest harmonic distortion at the output. It should be possible to sharply null the total harmonic content, which consists primarily of the 2nd and 3rd harmonics, to at least 60 dB below 0 VU. Then vary the cutoff frequency and make sure distortion is low for all settings. Of course, the filter itself will reduce harmonic distortion appreciably at its lower cutoff values. Lacking a distortion meter or wave analyzer, these adjustments can be made quite well by driving the input at 7 volts p-p (10 dB above 0 VU) to accentuate the distortion and setting very carefully for a symmetrical output waveform as monitored by a 'scope. Fixed resistors, determined by two decade boxes (the settings interact somewhat), could replace the pots. These adjustments, once made, are permanent unless the FETs are changed.

Figure 13 shows the distortion of the complete noise filter measured at two fixed values of bandwidth control voltage. At normal levels, distortion is so low that it is largely a measurement of the harmonic distortion of the test oscillator. The large margin above 0 VU passes the highest programme peaks ever likely to be encountered without clipping.

The simple power supply of Fig. 14 easily supplies the power requirement of ± 15 volts at about 10 mA.

CONSTRUCTION

The entire filter can be duplicated for about £40 with new parts. Very few components are critical and substitutes can be used in most cases. Quarter-watt, 5 percent composition resistors are suitable. Layout is not critical, since signal levels are high and impedances are relatively low. I strongly recommend that each of the functional blocks of Fig.2 be built and checked for reasonable conformance with the curves before integration into the system. This makes troubleshooting for errors and occasional bad components much easier, practically ensuring success. My unit (Fig. 1 and lead photo) is a "breadboard in a box." The circuit is still undergoing occasional changes, even though it is a third-generation model. Parts are mounted on terminal boards which were on hand. A neater approach would be to use the

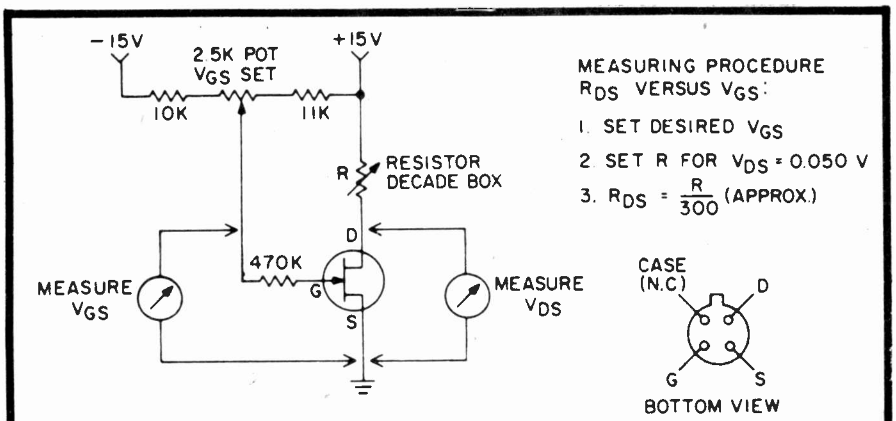


Fig. 12. Set-up for selecting FETs by static measurements (see text). Small 15 V batteries or the power supply of Fig. 13 may be used.

DYNAMIC NOISE FILTER

commercially-available matrix-board with snap-in terminals.

OPERATION

After checking the wiring, apply power to the unit and check for proper power supply voltages. Positive and negative supplies should both be between 14 and 16 volts with respect to ground. Much lower values would indicate a short circuit or bad op amp. Current drain should be on the order of 10 mA.

The noise filter can be conveniently connected to your audio system by means of the *Tape In* and *Tape Out* jacks included on most preamplifiers. An advantage to this connection is that the processed signal passes through the pre-amp tone controls, which can be set for the most pleasing final balance. For taping, the recorder input is paralleled with the output which drives the power amplifier.

For intial set-up experience, a record having a good frequency range and moderate, steady surface hiss is desirable. (A slightly noisy FM station can also be used, but results will not be quite as good because of the latter's flatter noise spectrum.) Initial control settings should be:

Pre-Filter: Off
Rumble Filter: Off
Time Const.: Off
Peak Rej. Freq.: 5 kHz
Base Cutoff: CCW
Suppr. Gain: CCW
Dyn. Suppr.: Off
Sig. Compare: Input

The signal should now pass through the unit unaffected, except the *Level Set* control will vary the gain from zero to 3.2 (10 dB). Set the level for 0 VU on signal peaks as you would set a recording level. Whenever the source is changed, the signal level should be reset as necessary.

Now switch the *Sig. Compare* switch to "output". The signal is now passing through the rumble filter (if used) and pre-filter, but bypassing the dynamic filter. Lowering the *Pre-Filter* cutoff setting should progressively cut off the highs. At the lower settings, which are primarily for acoustic records, the signal will sound severely band-limited. The best setting is the lowest cutoff which does not significantly affect the recorded bandwidth. I have found that with vocal music, the unfiltered sibilant sounds provide a means of judging bandwidth. If sibilants are quite strong and natural, a 7 kHz or higher cutoff is indicated. If they are weak or have a slight "whistling" sound, the upper limit is about 5 kHz. If sibilants are lacking, a 4 kHz or lower setting is best. Of course, the presence of high-frequency distortion may dictate a compromise setting a notch or two lower than indicated above. The filtered and unfiltered sounds may be compared at any time by means of the *Sig. Compare* switch.

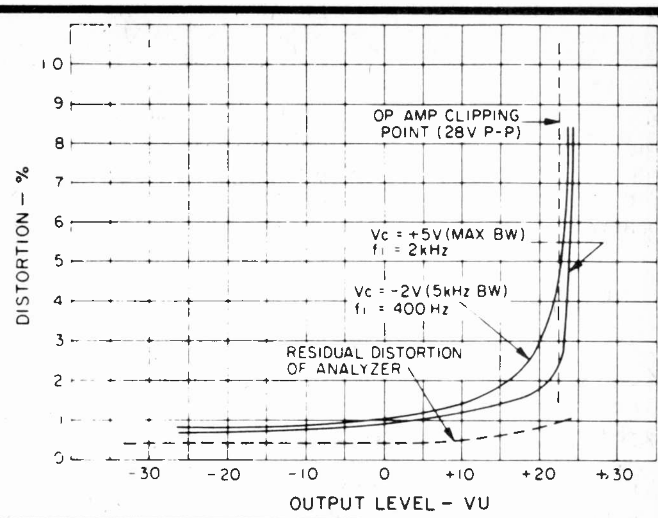


Fig. 13. Overall harmonic distortion of the noise filter for two constant values of bandwidth-control voltage.

The optional rumble filter is used for the occasional records which have warpage or bumps or low-frequency noise in the recording. For *acoustic* records it can be routinely left at 150 Hz, as nothing is recorded below about 200 Hz.

Next flip the *Dyn Suppr* switch to "on", putting the dynamic suppressor in the circuit. The sound should become very dull and lifeless, as the high-frequency cutoff is now 1 kHz or less. Increase the *Base Cutoff* setting until record noise just begins to be audible. The signal will probably still be quite lacking in high-frequency content (if it is not, only the pre-filter may be needed for this particular source). Now turn up the *Suppr Gain* slowly. This should "magically" restore the highs without increasing the noise level. The highest possible setting which does not noticeably increase the noise is normally best.

At this point it is edifying to monitor the bandwidth control input signal to the variable-bandwidth filter with a d.c.-coupled oscilloscope. The instantaneous voltage here is a measure of high-frequency programme amplitude and dynamic filter bandwidth (see Fig. 9). It should follow transients rapidly and may reach saturation (about + 14 volts) on musical passages having high harmonic content and on strong voice sibilants.

The *Peak Rej. Freq.* switch selects the frequency of peak rejection by choosing the appropriate filter curve (Fig. 7) for separating the bandwidth-control voltage from the input signal. The *5 kHz* position is used for most electrical 78 rpm records. For acoustic records or very noisy electrical 78s where the pre-filter is set for 4 kHz or less, the *3.5 kHz* position gives better results. Here the *Time Const.* switch can be set for *15 mS*. The longer time constant also helps to attenuate sharp clicks and pops occuring in quiet passages, as it prevents the bandwidth from increasing rapidly enough to follow their steep wavefronts. The *7.5 kHz* position is used for wideband recordings and tape.

With a little practice, you will be able

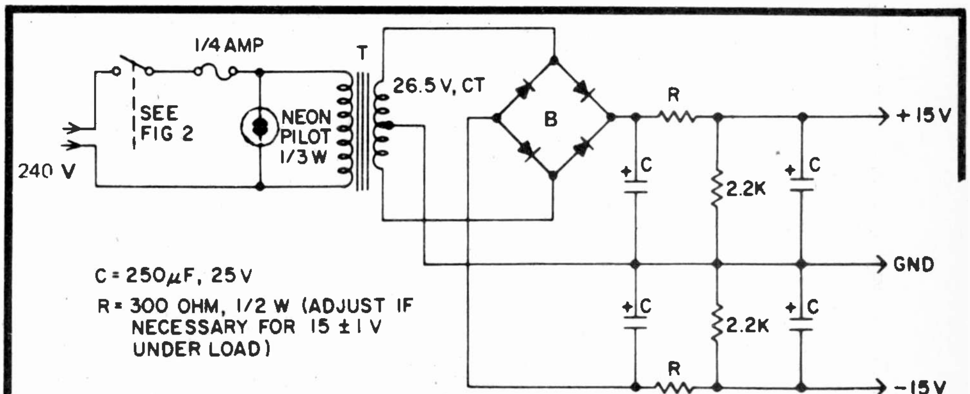


Fig. 14. Power supply. A two-channel suppressor may easily be powered by reducing R slightly.

to set the controls quickly for optimum performance. It is often best to set the *Base Cutoff* for a significant improvement, rather than to try to eliminate the noise completely. This will minimize low-level band limiting, and the suppressor will be less likely to betray its presence with obvious bandwidth changes.

PERFORMANCE

Figures 5 and 10 indicate the bandwidth ranges available. The pre-filter and dynamic filter (slope is 24 dB/octave above both cutoffs) can together provide well over 60 dB of noise attenuation at 10 kHz and over 40 dB at 5 kHz. The overall improvement in signal-to-noise ratio is strongly determined by the character and spectrum of the noise, which varies greatly with records. With the steady hiss typical of new electrical recordings on shellac, an average improvement of 8 dB (unweighted) is realized from the dynamic filter alone. Including the effects of the rumble filter and pre-filter on band-limited material, S/N improvement can be more than 12 dB. The apparent improvement is even greater, since the ear heavily weights the higher frequencies where record noise is concentrated.

The effect of the noise filter is surprisingly great on records which were originally thought to be quiet without filtering. It is a little weird at first to hear a familiar old record with realistic strings and brass and clear voice sibilants, but with the background suddenly rendered deadly quiet. I have spent many hours listening to the records and tapes in my collection and enjoying them anew.

The noise filter works very well on tape noise, providing at least 8 dB total S/N improvement. A stereo version built for tape only could be simplified considerably, as only the *Level Set, Base Cutoff, Suppr. Gain, and Sig. Compare* controls would be needed. The power supply as shown can easily handle two channels.

The noise level of the filter itself depends mostly on output amplifier A7. Of several units I tried, the noise level ranged from 62 to 68 dB below 0 VU.

A few tips on the mechanical aspects of copying records are in order here. The importance of good tracking cannot be overemphasized. More can be gained here than with any amount of electronic processing. Groove radius, depth, and angle were not standardized on early discs, and experimentation with tracking force and stylus size, if possible, may yield a considerable improvement in both noise and distortion. The playback stylus should, of course, ride on the sides of the groove. If it is too small it may ride the bottom of the groove and skate from side to side in a partially uncontrolled manner, creating severe distortion. If too large, it will ride high in the groove where it is more sensitive to surface blemishes. Also, larger styli cannot follow high-frequency modulation as well, especially on the inner record grooves. Elliptical styli are helpful on relatively wide-range 78s if the latter have not been damaged by previous playings.

Acoustic records (1925 and earlier) tend to have a larger groove, since with acoustic playback the mechanically-imparted stylus motion had to supply all the sound power. For these, a stylus of 4-mil (.004") radius may produce better results than the standard 3-mil size. Custom-made styli with a "truncated" tip (really a smooth transition from a 2- or 3-mil radius to about a 4-mil radius at the very tip) have been used to track the groove sides of 78s properly while avoiding contact with the bottom. (Truncated and other special styli are available from the International Observatory Instruments, 5401 Wakefield Drive, Nashville, Tenn. 37220). Although not a cure-all, these can give dramatic results on selected discs. A 2.5 mil stylus is best for most post-1946 transcriptions. Obviously, the pickup should have adequate lateral compliance and should produce no output for vertical motion. Incidentally, electrical recordings made before the mid-1940s are mostly recorded flat, that is, they have no high-frequency pre-emphasis, while later records have pre-emphasis of as much as 16 dB at 10 kHz.

Edison cylinders (160 rpm) and discs (80 rpm), some Pathe discs, and some early wax transcriptions are vertically modulated. Here the stylus does ride on the groove bottom, and the pickup should have only vertical response. This can be obtained (as can lateral-only response) from a suitably-phased stereo cartridge. Stylus radii of 4 to 10 mils are typical here; as always, experimentation is in order.

FUTURE DEVELOPMENT

The experimenter may want to try to improve the performance of the circuit described. Of course, additional types of processing can be added, such as more effective click suppression at the filter input or multi-channel equalization at the output. These would be electrically independent of the noise filter, and beyond the scope of this article. However, there are some possibilities for improving the noise filter itself. Many of these, unfortunately, would require an incongruous increase in complexity and cost.

Sharper filter cutoffs give a marginal improvement on very noisy material, but setup adjustments become more critical. Dynamic high-pass (low-cut) filtering using a simple 6 dB/octave slope might be a reasonable addition. Since the noise-rejection frequency band of the low-pass dynamic filter should complement the noise spectrum of the signal, a statistical study of record and tape noise spectra might lead to a better shape for the bandwidth-control-signal separation filter of Fig. 7. The separation filter selector could be ganged with the pre-filter cutoff switch to eliminate one control knob. Perhaps a noticeable improvement could be realized by experimenting with the shape of the bandwidth control characteristic, Fig. 9. The attack time constant could be shortened by using a more elaborate filter at the precision detector output; this would improve the response to occasionally encountered wide-band transients.

An obviously desirable change would be to replace the FET bandwidth-control filter with one of the voltage-controlled state-variable types. This would eliminate the need for FET selection, but would increase the cost severalfold. It therefore appears that the original goal of high performance per dollar has been achieved, yielding a practical design which is within reach of the hobbyist.

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Maplin Electronic Supplies have told us that they plan to market a kit of this project. Interested readers should contact them directly.

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ELECTRONICS —it's easy! PART 25

The algebra of logic.

MATHEMATICS is a kind of shorthand language which enables us to present a physical process, on paper, with symbols which may be manipulated in order to gain a better understanding of the process. It is thus a tool which aids understanding.

The familiar kind of algebra which relates two variables, x and y, in combinations such as x+y, x–y, x.y, x/y, x^y and others is a linear process because the two variables can hold any value. It is this kind of algebra that is performed by analogue operational amplifiers.

However, if x and y can only have two possible states, such as a voltage which is there or not there, we can ignore the actual value of the voltage (or whatever) and regard the variables as behaving according to a two-state or binary number system. Just what the two states are is of no importance whatsoever — they can be high or low, positive or negative, there or not there and even true or false.

A mathematical algebra has been developed to cope with such binary systems. It is known as Boolean algebra — the algebra of logic, and it's rules are somewhat different to those of linear algebra. Before delving into the operation of Boolean algebra, it is worth tracing it's historical development.

HISTORY OF SWITCHING MATHEMATICS

Philosophers, those people who apply special skills to resolving paradoxes by the use of logic, have existed since the earliest civilisations. The Ancient Greeks were so impressed with logic that they wrote plays around Aristotle's formally arranged rules of logical deduction. The rules for this process of reasoning were handed down, largely by word of mouth, through the Dark Ages, with little, if any, recognition of their value for logic in computation. It was not until the early 19th century that the use of logical rules in calculation was established. This work was very much the result of George Boole's 1854 work (see Fig. 1) entitled "An Investigation of the Laws of Thought on which are Founded the Mathematical Theories of Logic and Probabilities", Augustus de Morgan, a contemporary, also contributed to the first systematic arrangement of Aristotle's logic.

Boole took the concepts further than the Ancients by substituting mathematical symbols in place of the basic logical situations. This symbolic logic became known as Boolean algebra.

Little was achieved with Boole's work for the next few decades. The first machine to utilize his algebra to solve logic problems, faster than by hand, was William Jevons' logical piano of 1869. Boole's contribution, however, had to wait until the early 20th Century to find extensive application. One by one, logicians advanced the techniques of logical algebra: Pierce, Venn, Dodgson, Marquand, Pastore, Bollee. The "Principia Mathematica" of Whitehead and Russell (1910-1913) and the Hilbert and Ackermann work "Mathematical Logic" (1928) were further milestones in digital computer realisation.

Fig.1. In 1854 George Boole, an English logician, showed how ordinary algebra could be applied to logic situations.

Shannon's 1938 paper "A Symbolic Analysis of Relay and Switching Circuits" was a paper of very practical relevance for it described how to put Boole's rather abstract logical albegra to work in engineering and computer design. But this was not the first recorded use of electrical logic circuits. In a letter Charles S. Peirce wrote to his former student, Marquand, around 1890 he expressed, in the words and circuit diagrams shown in Fig. 2, that logical algebra could be performed with three switches in parallel or in series, also stating that he felt electricity to be one of the best ways to implement logical equipment.

Later theoretical studies concentrated on ways to ensure that switching networks contained no more switch contacts than were absolutely necessary. Unnecessary contacts can easily be unwittingly designed into complex switching networks — the "spares" are called redundant switches. Shannon, in his M.Sc. thesis (Fig. 3) prepared at the famous Massachusetts Institute of Technology, realised ways to systematically set about analysing a given switching networth in order to reduce the contact requirements to a minimum. Thus it was realized in the early 1940's that really powerful digital computers could be built using entirely electronic components.

Later in the course we will be dealing specifically with computer systems. They are, however, but a part of the total use of digital electronic methods — digital electronics finds use in an ever increasing number of instruments and devices.

BASIC LOGIC GATES

A quite satisfactory way to begin to comprehend basic switching algebra is to think in terms of mechanical switch contacts arranged in various different configurations. That we draw them and consider them as mechanical contacts that are either open or closed, does not imply that the contacts necessarily need to be mechanical — they are, today, more often than not the solid-state switches we discussed in the last part.

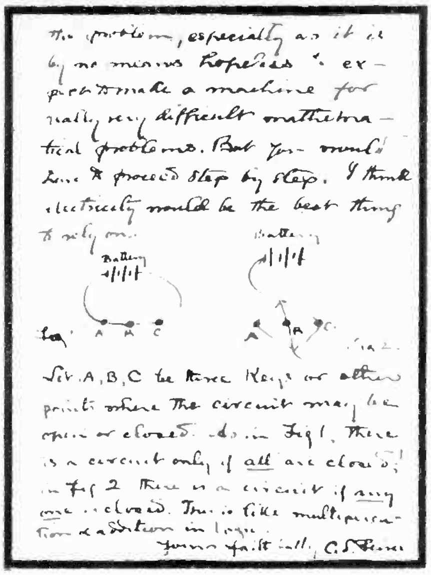
Groups of switches combining digital signal levels are known as gates. We begin by considering the simplest possibilities where there are just two contacts to build with.

They can be placed in series or in parallel, as shown in Fig.4. In each case different conditions exist between the transmission made through them for the two positions of each of the switches. We denote the switch inputs as A and as B (and C,D, etc., if more are involved) and the transmission as Z, thus using mathematical symbols to represent a physical situation. Imagine that the switches are wired in series with a lamp: when a circuit is made the lamp lights.

In the series case we need switch A and switch B to be made to obtain a transmission function Z. In the parallel case either switch A or switch B will provide transmission.

The AND and OR are basic logical functions. They need not necessarily be used only to describe electrical circuitry. They did, in fact, as we have seen, arise originally from philosophical study of truths and falsities.

Note that switch contacts are always shown in their non-actuated condition and this brings us to another basic gate function which can be realised using only one switch. If, as shown in Fig. 5, the switch A is actuated, Z is NOT enabled. If A is not actuated Z is enabled. A single switch,



The problem, especially as it is by no means hopeless to expect to make a machine for really very difficult mathematical problems. But you would have to proceed step by step. I think electricity would be the best thing to rely on.

Let A, B, C be three keys or other points where the circuit may be open or closed. As in Fig 1, there is a circuit only if all are closed; in Fig 2 there is a circuit if any one is closed. This is like multiplication & addition in logic.

Yours faithfully, C S Peirce

Fig.2. The first known description of electrical switching carrying out logic is in this letter of C.S. Peirce written around 1880.

Fig.3. Claude Shannon published details of "modern" digital computing design in 1937.

therefore, can provide a NOT function if its contacts are closed in the non-actuated state.

Attempts to explain switching circuit action in words, as above, only applies for the simplest of situations. The descriptive method becomes prohibitive when, say, we have two switches in series, in series with two switches in parallel, as shown in Fig. 6. Describing the action of all possible switch combinations on the lamp Z using words, is an inadequate way with which to communicate the idea. And few digital systems are that easy: many contain literally thousands of AND, OR and NOT gates.

We designate an OR function by means of the '+' symbol. This does *not* mean the same as our normal understanding of addition. When applied to decimal numbers it means addition as we normally understand it. With binary numbers, however, it has a different meaning and still another meaning when designating an OR function. For example:-

In decimal addition	1+1=2
binary addition	1+1=10
OR addition	1+1=1

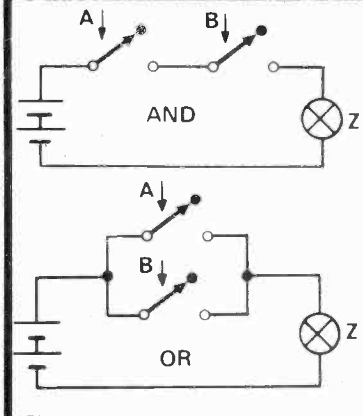


Fig.4. Two contacts wired in series or in parallel provide the basic logic functions of AND and OR.

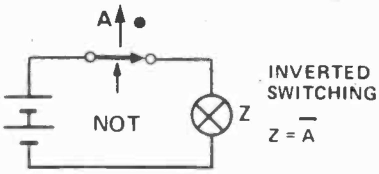


Fig.5. The NOT function is obtained by reversing the state of the switch operation.

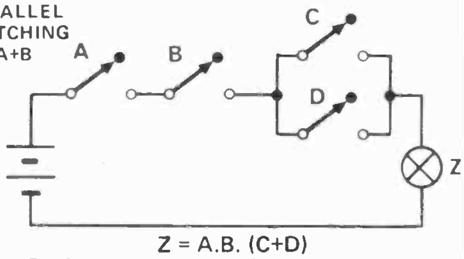


Fig.6. More complex switching functions are best described in terms of logic algebra than by words.

In Boolean algebra the OR meaning of addition is the one that applies. Thus A+B=Z means that A OR B switch closed will produce a transmission Z.

We designate an AND function with a dot. The dot means logical multiplication and is not to be confused with normal multiplication. However the truth tables for AND multiplication and normal multiplication are the same. Thus when we give the Boolean equation A.B=Z we mean that if switch A and switch B are both closed there will be a transmission Z.

The NOT function is designated as a line over the switches algebraic symbol giving $Z = \overline{A}$ to mean Z is NOT transmitted when A is actuated.

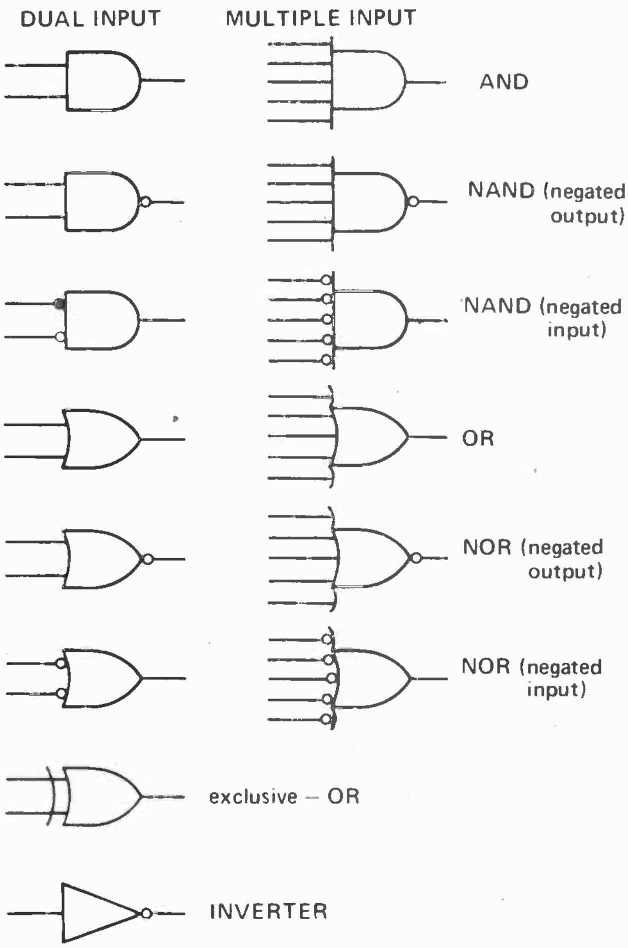


Fig. 7. Two-state logic gate symbols used in this course.

ELECTRONICS - it's easy!

Each of these functions have a symbolic representation as black-boxes with inputs that act in certain ways to give the output. The shape of the box (or the designation within a square box) tells the viewer the function of the box.

Unfortunately there still exists more than one conventional way to draw these symbols. For this course we will use those given in Fig. 7, which are also those used in projects in Electronics Today.

The NOT function bar can be applied to any function to signify that it is negated. For instance, an OR such as A+B=Z becomes $\overline{A+B}$ which is called a NOR function. Similarly so $\overline{A.B}$ is a NAND function.

The OR, AND, NOR and NAND functions can each have more than two inputs, for example, A+B+C+D=Z. When a function is negated its graphical symbol is also altered in some way to signify this. The convention used is the convention of the addition of a small round circle. If the circle is at the output the output is negated; if at the input the inputs are negated. The inverter (that provides negation) is basically an amplifier providing 180° phase shift so its symbol is that of an amplifier with the circle added.

TRUTH TABLES

Before we discuss more complex gate networks by studying their inter-connection, we need to understand the concept of a truth table. This is a simply drawn table that lists the output state for the various combinations of input states.

Rather than write on and off, or high and low, true or false, it is simpler to express the two states merely as '0' and '1'. The *positive* logic convention considers a high-voltage level as a '1' and the low level as a '0'. Fortunately, today, just about all logic circuits used are now in integrated circuit form and they nearly all work between just two levels - which are the same for any devices from a particular logic family. This provides a compatible arrangement whereby gates and other logic system boxes (that are yet to be introduced) can each be intercoupled without having to worry about matching voltage and impedance levels. However, when transferring logic signals between devices from different logic families translator circuits will be needed to make voltage levels compatible.

Occasionally, but not commonly, it is more convenient to reverse the levels calling a 1 the lower voltage and an 0 the higher. This is denoted *negative* logic. Such a system is however seldom used in modern integrated-circuit logic families.

Consider then the series contacts of Fig. 4. Assuming we use the positive logic convention where 0 represents an open contact and 1 a closed contact; it is easy to draw up columns as given in Fig. 8.

When A and B are both 0 then so also is Z, for no contacts are made. Similarly, if either A or B are open. When both A and B are closed, that is a 1 each, then Z is made. This is called a truth table.

Fig. 9 is the truth table for the parallel contacts of Fig. 4. In this case Z is 1 when A or B are 1.

An interesting property of the AND and OR functions is their dual nature when negated. For example, if we negate the inputs of the OR gate the truth table becomes that of Fig. 10, the output of which is the NAND function. Hence a negated input OR gate is a NAND gate. Also, by similar reasoning, a negated input AND is a NOR. Put another way, in negative logic an OR becomes an AND and vice versa.

INPUTS		OUTPUTS
A	B	Z
0	0	0
0	1	0
1	0	0
1	1	1

AND
Z = A.B

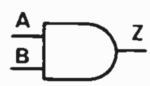


Fig.8. Truth table for A.B.=Z, the AND function.

A	B	Z
0	0	0
0	1	1
1	0	1
1	1	1

OR
Z = A + B

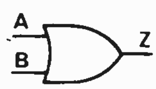


Fig.9. Truth table for A+B=Z, the OR function.

POSITIVE LOGIC		NEGATIVE LOGIC		NAND AND (IN POSITIVE LOGIC)	
A	B	$\overline{A}$	$\overline{B}$	Z	$\overline{Z}$
0	0	1	1	1	0
0	1	1	0	1	0
1	0	0	1	1	0
1	1	0	0	0	1

Fig.10. Truth table showing that negative logic (or negated positive logic) input to an OR gate provides NAND output.

UNIVERSAL GATES

Using the basic gates, AND, OR and NOT, we can build a logic circuit for any given Boolean expression. Where there is a plus sign (+) we use an OR gate, where there is a dot we use an AND gate and we use an NOT gate for those functions that are negated.

However it is interesting that the NAND gate can be used to obtain any desired function. It can be used to build AND, OR or NOT gates. In other words it is a universal building block, as is the NOR gate also.

Thus the majority of gates used in modern logic systems are NAND gates with the occasional use being made of NOR gates and inverters (NOT) to minimize complexity. The use of one major form of gate simplifies manufacture and reduces costs.

FAN OUT

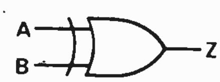
There exists a finite number of circuits that can be safely connected to the input, or the output, of logic elements. This number is called the fan-in and fan-out respectively, and gives the number of standard loads that can be accommodated. Fanouts of 10 and 30 are typical load factors.

EXCLUSIVE OR

One other important gate is a special class of the OR – the exclusive OR. The logic action of this gate is seen by studying its truth table which is given in Fig. 11. In this variation of the basic OR gate the output is 1 for either A or B but not when both are 1 simultaneously. Written in Boolean algebra symbols this gate performs $A.\overline{B}+\overline{A}.B=Z$. (Symbols written as AB imply that a dot exists between them; it is common practice to omit the AND dot).

A	B	Z
0	0	0
0	1	1
1	0	1
1	1	0

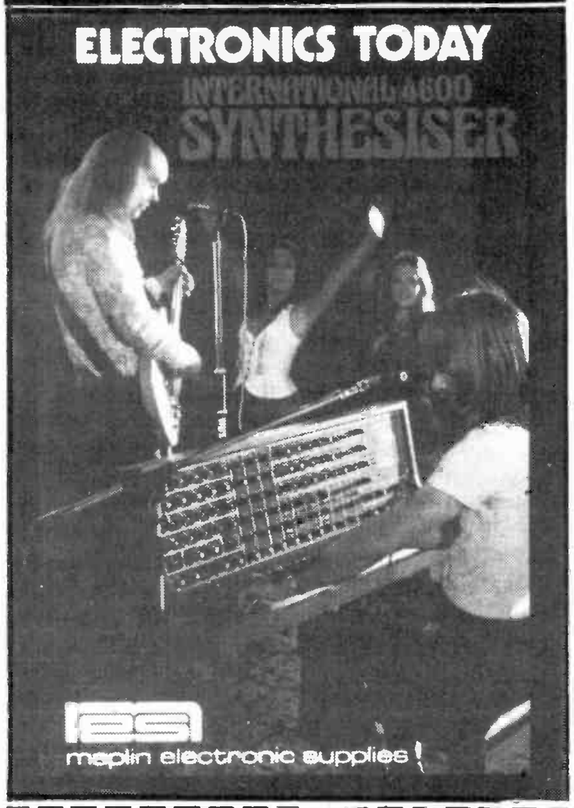
EXCLUSIVE–OR
$Z = A.\overline{B} + \overline{A}.B$



(Z = A⊕B)

Fig.11. Truth table for two input exclusive – OR gate.

SPECIALS



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ELECTRONICS TOMORROW

by John Miller-Kirkpatrick

I had found several interesting snippets about new products that I had hoped to relate to you in full this month but I have suffered from the great Post Office holdup. Information which was sent to me from Bedford had not arrived over a week later and so we will devote most of this month's column to a new American product; the data arrived from Ohio within three days of my enquiry. The new product is only a new approach to the age old problem of breadboarding circuits before setting them permanently onto Veroboard or PCBs.

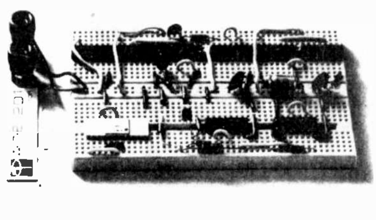
The version I have, is called an ACE board made by AP products Inc of Ohio, but a similar unit is called a PROTOBOARD and is advertised regularly in the American amateur magazines. The idea is similar to the S-Dec and T-Decs that have been on sale for several years except that the new boards are built on a 0.1 inch matrix with sockets that are good enough to take ICs of any size. The board that I received is called an ACE-201K and comes in a kit form containing three plastic mouldings, aluminium base with self-adhesive area, 195 terminal strips, rubber feet and two binding posts. The unit took me less than 30 minutes to assemble and it went to work almost immediately with a new clock circuit on it. Of the three mouldings one is 48 holes long by two holes wide and contains a continuous strip of sprung metal connecting all 48 holes in one channel with a similar strip connecting all the holes in the other channel, this acts as a distribution bus for ground and supply voltage. The aluminium base is connected to the ground supply to give a good backplane for circuits that require it. The other two mouldings sit one on either side of the distribution bus and each contains two lots of 5 by 48 holes with the metal strips connecting each set of 5 holes. Thus you end up with four lots of 48 five connector terminal strips with the distribution strip running down the centre.

This board has the capability of handling up to twelve 14 pin ICs where each pin of each IC can have up to four leads connected to it, it will handle a couple of 28 or 40 pin ICs plus transistors, capacitors, resistors, etc with plenty of tie-points left over. The two binding posts are intended to connect out to your power supply although I believe that one of the Protoboard range has a 5v 1A supply built into the unit to give a complete TTL or CMOS test bed.

PROTOTYPER

Something that I built for my own lab a few years ago could be of interest to readers who do a lot of TTL development work and could be built using one of the above boards. I used an S-Dec as the basis of my prototyping unit and built it into the top of a smallish instrument case. The idea if the unit is to use the S-Dec sockets as inputs and outputs to standard TTL units which are built on veroboard inside the case. To build a simple unit you might require a couple of 7400s, 7402, 7404, 7406, 7410, 7420, 7490s, 7493s, etc, in fact whichever gates you usually use. You have to open up the S-Dec and remove the copper connecting strips. Each strip is then cut up into individual socket units and a wire soldered onto the back of it before carefully gluing it back into the body of the S-Dec. This is a very tedious job but you end up with about 100 individual sockets each connected to a short length of wire, these wires can then be connected to the inputs and outputs of the TTL gates which are powered separately. If you are not too worried about propagation delays you can connect the gate outputs to invert gates so that a 7400 could behave like a 7400 or a 7408. When the unit is completed you can have a row of five holes connected to each two input gate such as — INPA, INPB, OUT, OUT, INV-OUT. By simply pushing 22swg wires into these holes you can lead off a complete 7400 gate function to the bradboard that you are presently working on without having to place a 7400 on that breadboard. The second stage of the prototyping unit is to include counters, seven-segment displays, LED lamps, BCD switches, etc and then to include a Protoboard or ACE kit on the same layout. This gives you the capability of building just about any circuit you like as a prototype without much additional cost, having proved the circuit you can then strip it down and build something completely different. One further advantage of this type of unit is that you can use up all of your 7400s with one gate blown or 7490s with no A input, etc as these can now be hidden away inside the unit. A word of warning — if you decide to build a prototyping unit make sure that all of the internal ICs are in sockets for easy replacement or to be extra sure buffer each input and output by connecting a 7407 between the wire socket and the IC in use. This means that if you do overload a gate it will be the cheap 7407 which goes not the 7490 or whatever, in this case make all of the 7407s easily accessible for replacement. Perhaps we might have an ETI Prototyper in a future issue.

If you would like to buy an ACE-201K breadboard kit you can try to contact a company called GDS in Slough but you will probably meet the unhelpful buck-passing that I did; they do sell AP products but have never heard of ACE boards. You can definitely get one from AP Products Inc, Box 110, 72 Corwin Drive, Painesville, Ohio 44077 if you send them about 27 dollars (24.95 plus postage) — this is a worthwhile investment in any lab whatever its size.



DISPLAYS TIME AND FREQUENCY

I have previously mentioned the Futaba fluorescent (Phosphor-diode) display type 5-LT-01 for use in digital clocks and the 5-LT-03 display for use with counters especially for use with the GI digital FM tuner chip. For those of you who want to build a digital readout for your tuner and want to use the same display for a digital clock wait for a new display from Futaba. This has the same four digits, colon and AM/PM as the 5-LT-01 but has also got an extra digit '1' on the left, MHZ/KHZ on the right and a complete set of decimal points. Thus it can display time with colon and AM or PM or can display tuned frequency from 1.9999 to 19999 with the choice of MHZ or KHZ. The type number and price have not yet been announced but I suspect that it will cost about £7 and be available about April. At the same time Futaba have also announced a display for a digital car clock to interface with the MM5378 (the higher voltage version MM5379 might be better), the display size is about 0.25in and contains four digits plus colon. The drive voltage for the display is about 30-35 volts at about 15mA, this could easily be derived from the car voltage through a simple inverter circuit.

CLOCK CHIPS

One clock chip that flared into light a few years ago and then almost disappeared was Nationals MM5316. This is a non-multiplexed chip with four digit output which will show hours and minutes, minutes and seconds, alarm time or 'Sleep' time, two outputs provide switching of radio or alarm with the 'Sleep' timer allowing you to go to sleep with the radio turned on. This chip was quite expensive but would drive Dynamic Liquid-Crystals or Fluorescent displays directly or LEDs via thirty transistors, the total cost of a clock using these was in the area of £50. National used the same logic on some other chips (MM5371, MM5370) but these also died in this country. Now several other manufacturers have second-sourced this chip (S1998, uPD5316, KM5139, etc) and also National have brought out some LED drive versions (I can't remember the numbers) and whichever chip you use the price is now a lot lower.

Futaba have now made the 5316 clock a little bit simpler and cheaper with the introduction of the 5-LT-02. This fluorescent display is externally very similar to the 5-LT-01 the only difference being that the display is non-multiplexed, 'Great', I hear you say, 'a non-multiplexed display for a non-multiplexed clock chip'. Futaba have been just a bit cleverer than that by making the 5-LT02 almost pin compatible with the 5316, wiring up on veroboard would be a cinch, hand drawing a PCB would be even simpler, and at a price of about £7 it is cheaper per digit than LCD or LED. As the additional components for the clock cost less than £1 (plus a small transformer) this could work out to a very cheap DIY alarm clock. The 5-LT-02 displays should be available within a month or may well be in the UK by the time this article appears.

In case you have never heard of fluorescent displays they are very bright, green, low current (1 or 2 mA per segment), the disadvantages are a heater (5V 50mA) and a drive voltage of about 35V.

—

References: Futaba Displays and National Clock chips — Bywood Electronics, 68 Ebberns Rd, Hemel Hempstead, HP3 9QRC.
NEC (uPD5316) and Futaba — Walter Scott, Imp House, Ashford Rd, Ashford, Middx.
Toko (KM5139) — Ambit International, 37a High St, Brentwood, Essex.
ACE-201K — GDS Sales, ??? Bath Rd, Slough. Tel Slough 30211.
PROTOBOARDS — Ancrona Corp, Box 2208, Culver City, California 90230.

tech-tips

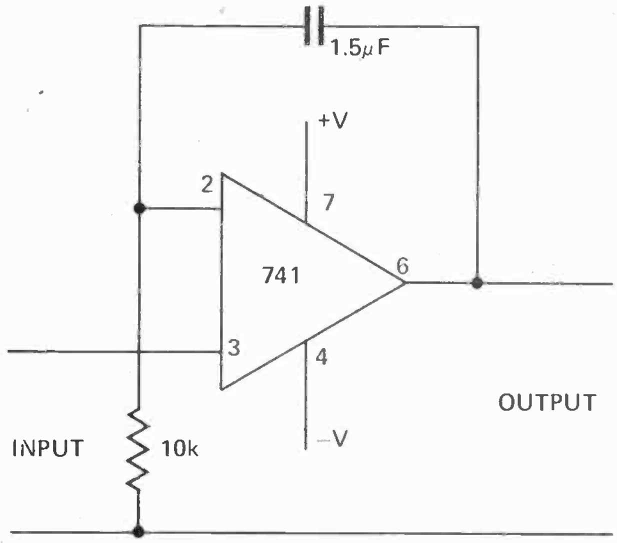
LOW FREQUENCY EXTENDER

In circuits which have a variable frequency input, e.g. optical tachometers, vibration measuring equipment etc., the low frequency response can leave a lot to be desired. The circuit shown brought the lower 3dB point of a measuring instrument down to 0.5Hz when placed in circuit between the transducer and the instrument.

Being of small size, the circuit may be fixed inside the case of the instrument it is to serve.

The gain of the circuit may be altered by means of the feedback capacitor to give a level response compatible with the instrument to which it is connected, i.e. a higher value will give a lower gain and vice-versa.

The 741 IC will operate at voltages between ±5 and ±15V.



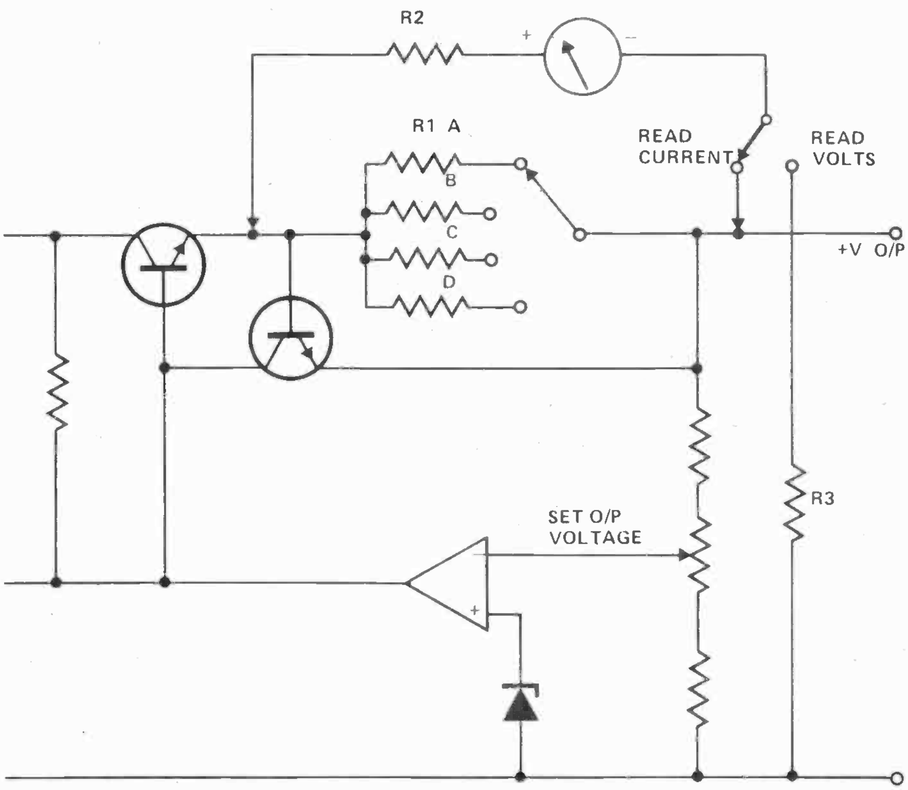
METERING A STABILISED POWER SUPPLY

It is not easy for the home constructor to make shunts of the correct value for a meter when wishing to alter its current range.

One way to monitor the current supplied by a current limited power supply is simply to measure the voltage drop across the current limiting resistor. This is usually of the order of .65V, and if the series meter (R2) is calculated to give fsd when .65V is applied, will indicate the limiting current at fsd no matter what the value of R1 may be. In effect, the limiting resistor becomes the meter shunt.

For a basic 1mA meter, the series resistor R2 will need to be about 560Ω, for a 50μA meter it will be about 12kΩ, for a 5mA meter about 120Ω and so on.

Unless individual adjustment of resistors and calibration of range is undertaken, this method cannot be absolutely accurate, but it will show whether a circuit is drawing something like its expected current. The method does have the advantage however, of the meter being within the feedback loop, and will therefore not add to the power supply output impedance, which can be important in some applications.



The addition of a single pole C/O switch as indicated will enable the meter to be used to set up the desired output voltage also, though it must be remembered that this will include the voltage drop across R1 if any current is being drawn, which could lead to a difference of .5V or so between indicated and actual output voltage. R3 should be chosen to give fsd at maximum output voltage and the meter scaled accordingly.

Tech-Tips is an ideas forum and is not aimed at the beginner. We regret we cannot answer queries on these items.

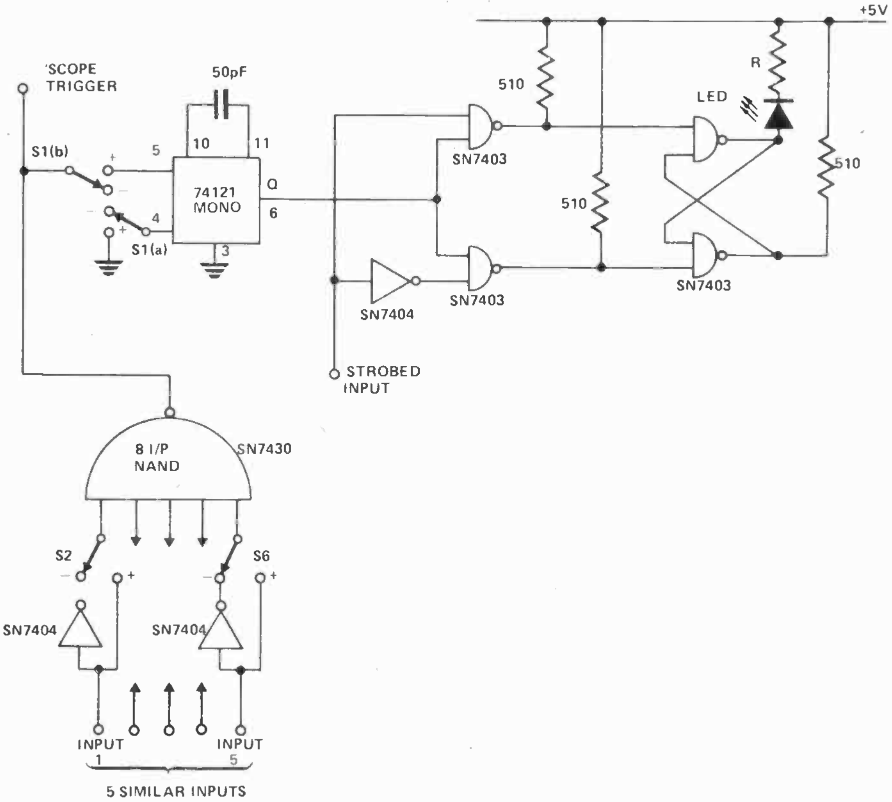
ETI is prepared to consider circuits or ideas submitted by readers for this page. All items used will be paid for. Drawings should be as clear as possible and the text should preferably be typed. Circuits must not be subject to copyright. Items for consideration should be sent to the Editor, Electronics Today International, 36 Ebury Street, London SW1W 0LW.

LOGIC ANALYSER

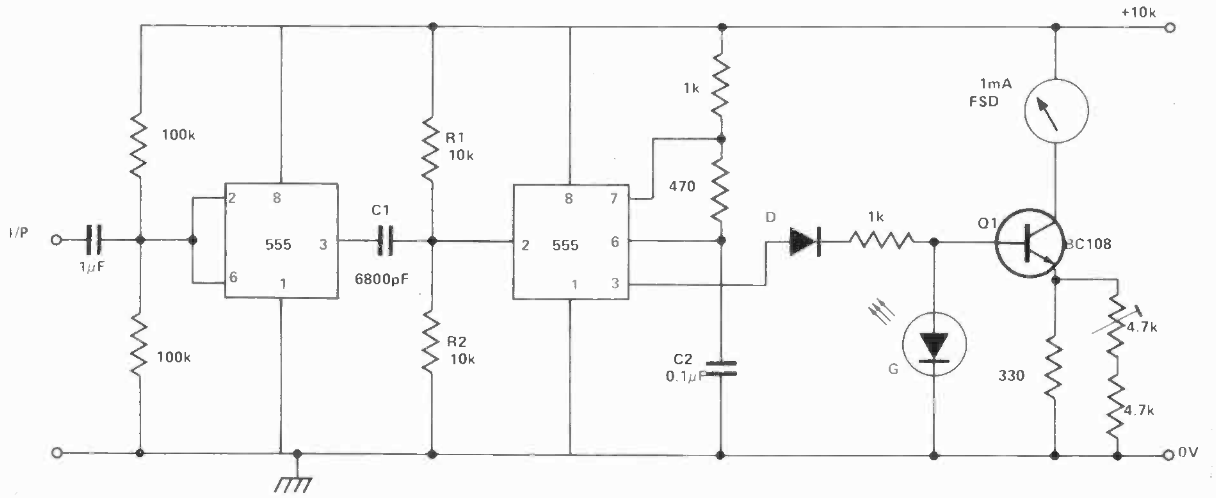
This circuit has been found useful for in-situ testing of TTL logic elements and general circuit development. An LED is used to indicate the state of the strobed input a short time (about 100nS) after the output of the 8 I/P NAND gate changes. Switch S1 is used to control whether the strobe occurs on the leading or trailing edge of the NAND gate.

The five control inputs and associated switches S2 to S6 determine when the strobe occurs relative to logic states in the circuit under test. In use, one or more of the inputs would be connected to appropriate points in the circuit, with switches S2 to S6 as described. For example, the state of a particular circuit point can be strobed each time a selected set of five other circuit points have the logic level 1,0,1,1,0 by setting the switches to +,–,+,+,–, respectively.

An output to trigger an oscilloscope is provided to overcome a common problem in logic circuit testing of finding suitable trigger points in the circuit.



SIMPLE ACCURATE FREQUENCY METER



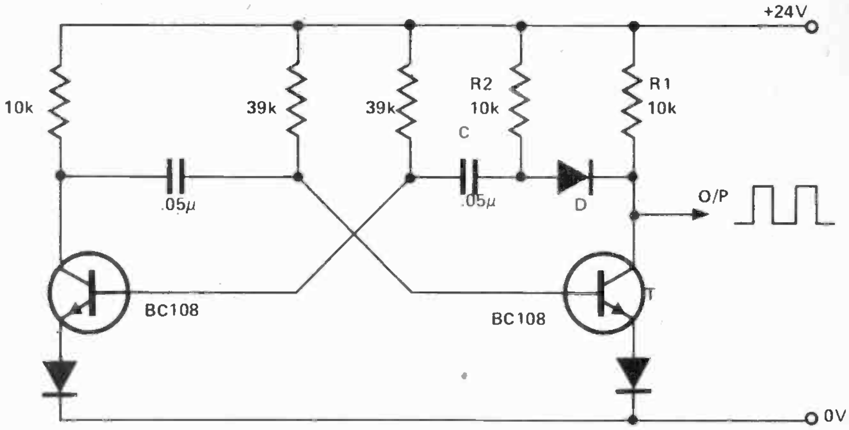
This circuit provides a meter deflection that is strictly proportional to the frequency of the input signal over the range 10Hz–300Hz. The first 555 timer IC is used as a Schmitt trigger, to convert the I/P signal to a fast-edge square wave. This is differentiated by the network C1, R1 and R2, and the resulting spikes used to trigger the second 555, which operates as a monostable, generating constant-width pulses. These are used to turn on the constant-current source Q1, so that the average current in the meter movement is proportional to the number of pulses arriving per second. A green LED is used to bias the current source as this gives near-perfect temperature compensation; the 4k7 preset pot gives a fine adjustment for calibration purposes. When the 1mA meter shown is used, fsd is given by 100Hz. To extend the range, reducing C2 to .01μF gives an fsd of 1kHz.

tech-tips

IMPROVED MULTIVIBRATOR

Conventional astable multivibrators suffer from the disadvantage that they do not produce a good square-wave output; the leading edge of the waveform has a very slow rise since the collector resistor R1 is tied to a slowly charging capacitor C when the transistor T turns off.

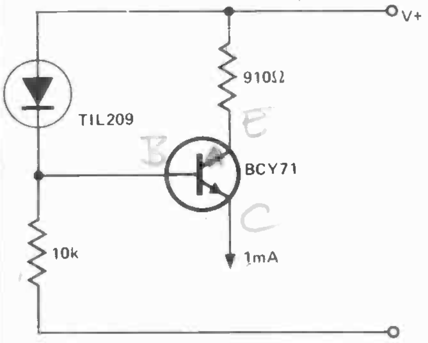
This circuit prevents this effect and thus generates a clean square-wave with 400nS rise-times and 100nS fall-times. This is because diode D turns off when the output begins to rise in voltage, and a fast rise is then possible. C is charged by a separate resistor R2, and apart from this multivibrator action is normal. The components shown give an operating frequency of about 700Hz.



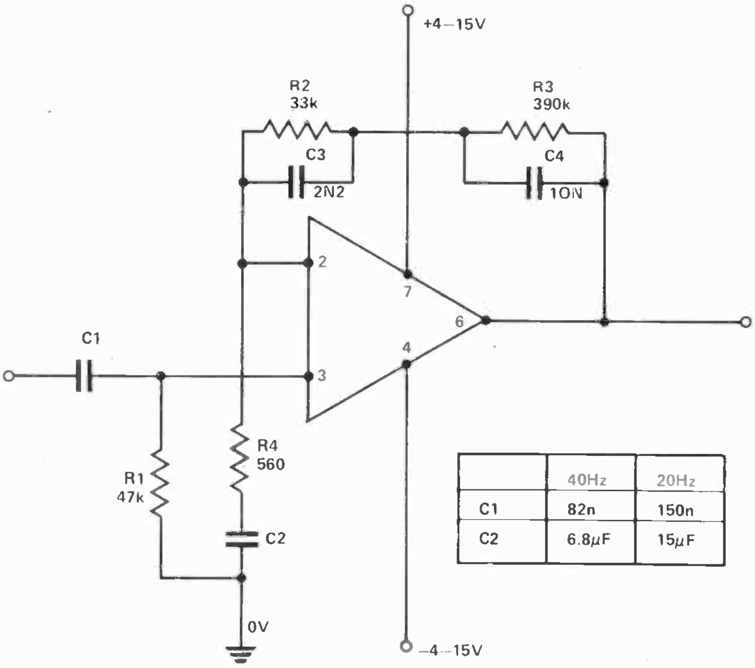
TEMPERATURE-STABLE CURRENT SOURCE

This current source is very temperature-stable; the output current varies by less then 1% over the temperature range −55°C to +100°C. This is possible because the transistor is biased by an LED, whose forward voltage drop has a temperature coefficient of −2mV/°C, the same as the base-emitter voltage of a silicon transistor. Hence near-perfect temperature compensation is possible, a great improvement over conventional methods of biasing with zener diodes.

The circuit values shown give an output current of about 1mA, though wide variation is possible by altering the value of emitter resistance. They are good for supply voltages in the range 25V to 5V.



CARTRIDGE EQ AND RUMBLE FILTER



	40Hz	20Hz
C1	82n	150n
C2	6.8μF	15μF

In this circuit a 741 op amp is used to provide standard RIAA equalisation for a magnetic pickup cartridge. The input signal is coupled via C1 into the non-inverting input of the IC. R1 damps the inherently high impedance at this point and provides the correct load for the cartridge. Feedback from the output, pin 6, is taken through the equalisation network R2, C3, R3 and C4 to the inverting input.

The ratio of R2 to R4 sets the midband gain at 65, 35dB. C1 and C2 together form a steep cut rumble filter whose cut off point can be set at 20 or 40Hz by selecting the appropriate component values in the table.

C2 also reduces the dc gain of the circuit to unity so that the output offset voltage will be ±5mV with reference to 0V.

One of the major disadvantages of discrete equalisers is overload distortion. Although the output of a magnetic cartridge may be only about 5mV normally a musical peak may well force the cartridge output to 100mV. Clearly unless a large signal swing is possible the sound emmanating from the speaker is not going to be Hi-Fi.

This circuit, operating from a ±15V supply, has an overload factor of +35dB refered to a nominal input of 5mV, equivalent to a maximum input of 325mV!

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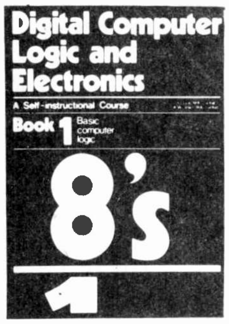

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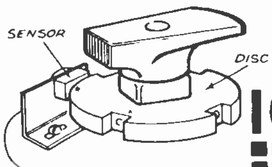

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