wireless world

OCTOBER 1981 70p

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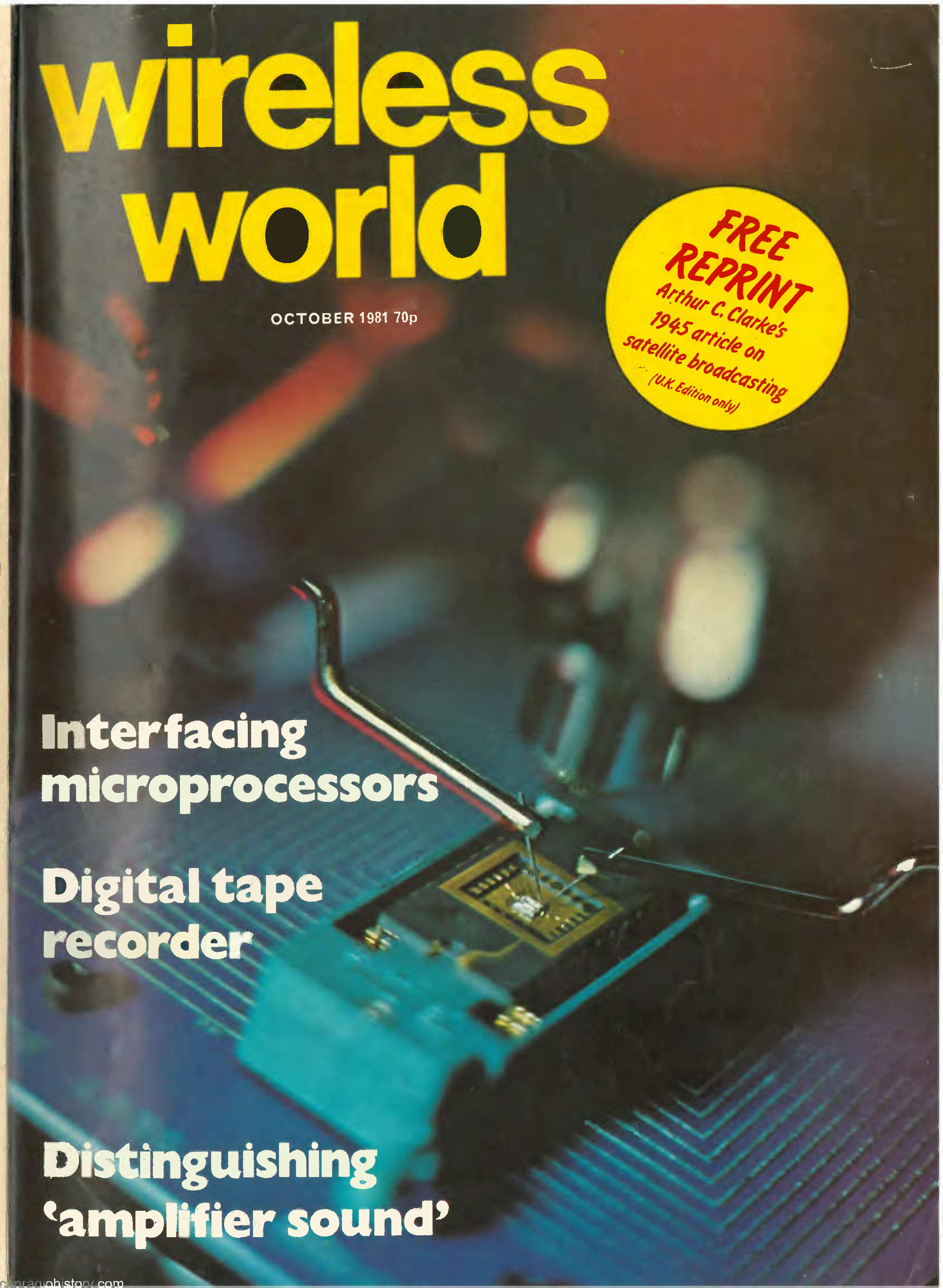
Arthur C. Clarke's 1945 article on satellite broadcasting
(U.K. Edition only)

Interfacing microprocessors

Digital tape recorder

Distinguishing 'amplifier sound'

WIRELESS WORLD OCTOBER 1981 VOL 87 NO 1549



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Front cover shows integrated-circuit probing equipment by Wentworth Laboratories. Photograph by Paul Brierley.

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ELECTRONICS / TELEVISION / RADIO / AUDIO

OCTOBER 1981 Vol 87 No 1549

IN OUR NEXT ISSUE

High-resolution satellite images. **The design of a receiving station for use with a computer to display weather pictures from spacecraft.**

C.b. frequency synthesizers. **Two kinds of synthesizer suitable for 27MHz are described, with a practical mixer synthesizer circuit to cover 40 channels.**

Cartridge alignment gauge. **Following on from this month's article, R. J. Gilson presents a simple device which makes alignment of pickup cartridges easier.**

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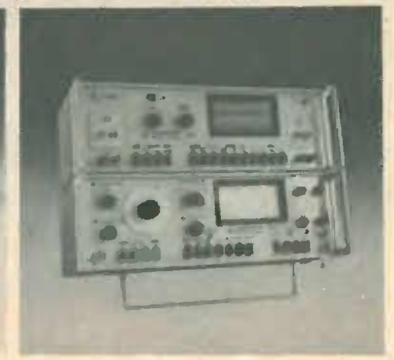

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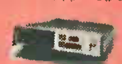

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7055 plus processor control and RS232 interface £900.00
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Super 50 Selectest. 20KΩ/volt. £77.00

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PM3243 Storage Scope 50MHz Dual Trace and Timebase. Variable persistence £2,500.00

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7704A Main Frame C/W 7A26, 7B80, 7B85 Dual Trace DC — 200MHz £3550.00

Telequipment.
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Tektronix
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P6015 40KV 1000 Probe ..... P.O.A.

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MG5-10 5V @ 10A switching £95.00
MG24-12 24V @ 12A switching £130.00

Weir.
762 Power Supply Unit 0-30V at 2A. Metered output. £90.00

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S.E.Labs.
6012 50 Channel UV Recorder. Servo paper drive upto 5 metrs/sec. 12" paper £1100.00

Watenabe.
MC641 6 Channel Chart Recorder 1mV-100V. 250mm scan width £1495.00

Yokagawa.
3047 2 Channel Chart Recorder. 0.5mV-100V. 2cm/hr-60cm/min £435.00

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4204A Decade LF Oscillator 10Hz-1MHz. 1mV-10V into 600Ω £695.00
606A AM Signal Generator. 50KHz-65MHz £475.00
606B AM Signal Generator 50KHz-65MHz AM 0-95% £850.00
616B UHF Signal Generator 1.8 to 4.2GHz Int pulse Mod. £1000.00
651B Test Oscillator 10Hz-10MHz. 0.1mV-3.16V £415.00
3320A Frequency Synthesizer 0.01Hz-13MHz. O/P range 0— +13dBm into 50Ω. Long term frequency accuracy ± 10 parts in 10⁸ of setting per year £995.00
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TF1370A Wide Range RC Oscillator. 10Hz-10MHz. Sine wave, square wave upto 100KHz £275.00
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TF2005R 2 Tone Signal Source. 20Hz-20KHz. 0-111dB in 0.1dB steps £295.00
TF2008 AM/FM Signal Generator 10KHz to 510MHz Stability 5P P M. built-in sweeper. RF output .2μV to 200MV £2725.00
TF2015 AM/FM Signal Generator 10 to 520MHz £875.00

Philips.
PM6456 Stereo Generator. Separate L and R Signals. Carrier frequency 100MHz ± 1%. RF O/P 3mV pk-pk £250.00

Radiometer
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1481C PAL Waveform monitor in as new condition £2275.00

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Wandel and Golterman.
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MISCELLANEOUS

Bruel and Kjaer.
2409 Electronic Voltmeter True RMS. Average and Peak 2HZ-200KHz £350.00

Ferrograph.
RTS2 Recorder Test Set Measures Wow & Flutter, Distortion, Gain £345.00
RTS2 + ATU-1 Recorder Test Set and auxiliary test set £375.00

Hewlett Packard.
332A Distortion Meter 5Hz-600KHz £595.00
334A Auto Nulling Distortion meter. 5Hz-600KHz. 0.1%-100% FS. AM detector I/P £800.00
4329A Insulation Resistance Meter. Range 500KΩ to 2 × 10¹⁶Ω £500.00
8745A S Parameter Test Set. Fitted with 11604A Universal Arms 0.1-2GHz £2750.00

Marconi.
TF791D Deviation Meter. 4-1024MHz £195.00
TF2163S UHF Attenuator 0-142dB 50Ω impedance DC-1GHz £250.00
TK2214 X-Y Memory £250.00
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TF2500 AF Power Meter. 7 ranges 100μ watts to 25 watts Impedance 2.5Ω to 20KΩ in 40 steps £275.00
TM8339 AC/DC mixer for TF2702 ..... £295.00

Miles Hivolt
IT30 Insulation Tester ministry version CT587/3 (mint) £850.00

Rohde and Schwarz.
MSC Stereo Coder. 30Hz-15KHz £500.00

Tektronix.
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KL8E Asynchronous Interface	£175.00
KL8JA Asynchronous Interface	£275.00
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TC08 DECtape control (Old type), 'as is condition'	£50.00
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PC11A Reader Punch with unibus control module **£1250.00**
TU60 Twin Cassette drive, new condition **£500.00**
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Calcomp
Drum Plotter model 565 **£1250.00**
Data General
Nova 1210 CPU type 8133-2 with 4KW "AS IS" **£395.00**
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Flat bed Plotter type DP3 — 1/5 **P.O.A.**
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Diablo Hyterm 1620
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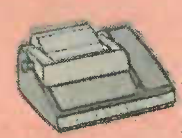

Low Cost Printer Offer.
Teletype 33 printer mechanism including case but no keyboard or electronics, 64 upper case ASCII, 10 cps, pinfeed platen, ideal for the electronic hobbyist ONLY **£85.00**

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P135/20 paper tape punch. Solenoid-actuated unit capable of punching 5 to 8 channel tapes at speeds up to 35 cps.
Pulse amplitude 27 VDC. Compact table-top unit **£95.00**

Facit 4070.
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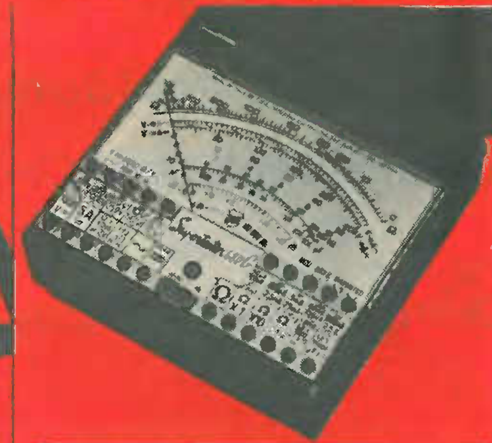

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Dual Trace. DC to 20MHz. 8 × 10cm display. Risetime 17.5nS. Sensitivity 5mV/cm-20V/cm. Timebase 0.5µs-0.2s. X5 magnifier. X-Y operation. Auto or variable trigger. Channel 1, Channel 2, line and external. Coupling AC, or TV low pass filter. Weighs only 6Kg. Size (m.m.) H. 145, W. 285, D. 380.
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I.C.E.

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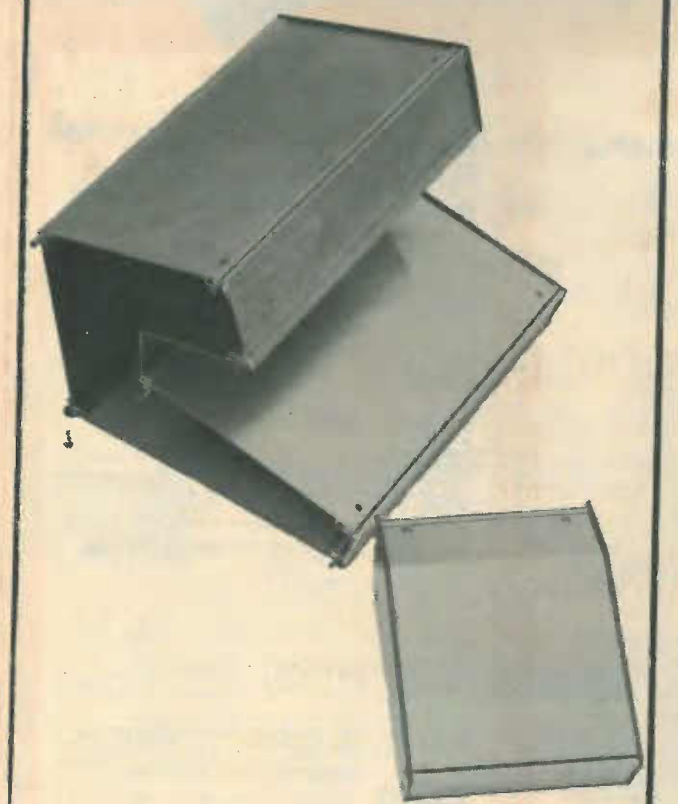

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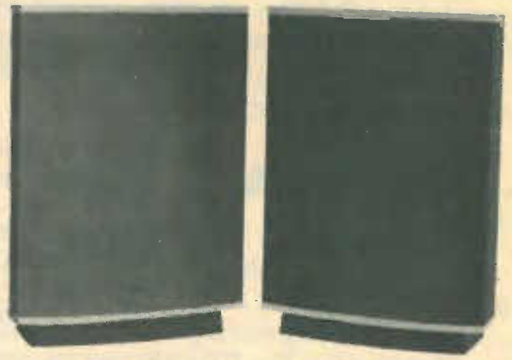
You will have to wait to acquire a pair of Quad ESL-63s, but in these gloomy days it is comforting to be associated with another Quad, hence British, success and your patience will be rewarded by many years of listening pleasure.

...or an apology for success.



The Quad ESL-63 is well worth waiting for.

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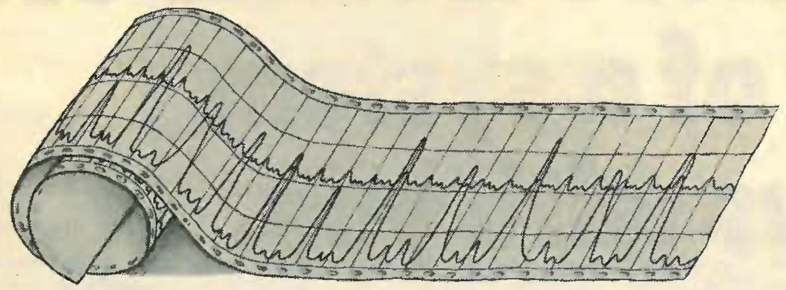
meet the time shrinker!



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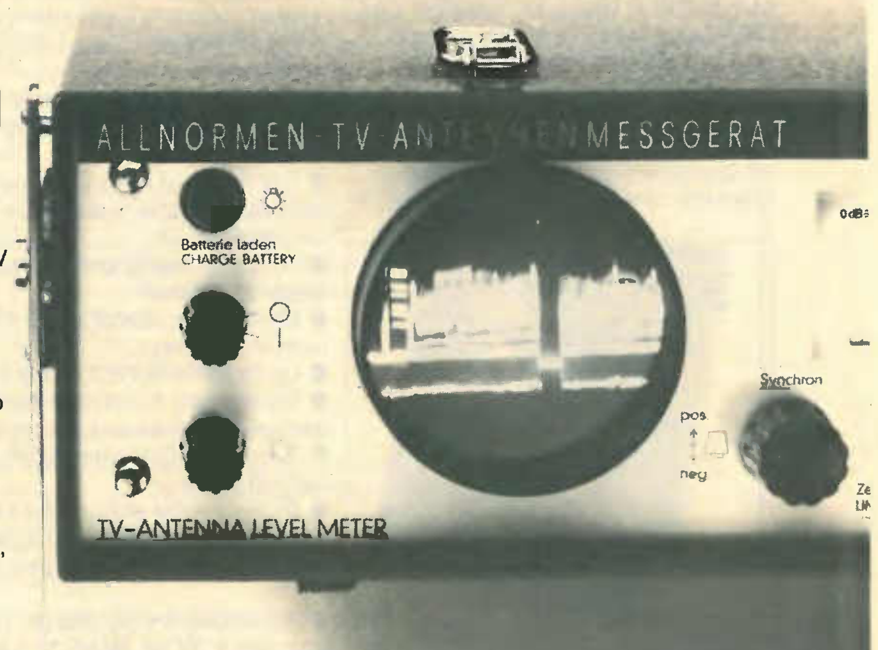

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- 2203 Sound Level Meter — 450
- 2305 Level Recorder — 650
- 4230 Sound Level Calibrator — 95
- 4424 Noise Dosemeter — 375

BRIDGES & V and I STANDARDS
BONTOON
- 71AR LC Bridge — 150

GENERAL RESISTANCE
- DAS56 DC V and I Calib 1µV-10V 30mA — 600

HEWLETT PACKARD
- 4261A Digital Automatic LCR Bridge — 975
- 4342 QLC Meter 22 KHz-70 MHz — 1100

MARCONI
- TF868A Universal LCR Bridge — 250
- TF1313 Universal LCR Bridge — 350

WAYNE KERR
- B224 LCR Bridge 0.1% — 500
- B521 LCR Bridge — 115

COMMS & CABLE TEST EQUIPMENT
CHASE
- 35A Field Strength Meter 20-850 MHz — 600

HEWLETT PACKARD
- 3556A psophometer 20 Hz-20 KHz — 250

STC
- 74216A Noise Generator CCIT — 275
- 74261A Psophometer CCIT — 375

TEKTRONIX
- 1502 TDR Cable Tester CRT + Recorder — 2950

COMPUTER EQUIPMENT
CENTRONICS
- 702 matrix printer — 500

TEKTRONIX
- 4610-1 Hard copy printer for 4010 series computer display terminals — 1800

COUNTERS & TIMERS
FLUKE
- 1910A-1 125 MHz 7 digit Cntr. AC/Batt — 300
- 1911 250 MHz 7 Digit Counter — 298
- 1912 520 MHz 7 Digit Counter — 375
- 1912A01 As 1912A but inc. re-charging batteries — 430
- 1920A 520 MHz 9 Digit Counter inc. Brst. mode — 575
- 1920A14 1250 MHz otherwise as 1920A — 750

HEWLETT PACKARD
- 5300A 6 Digit Display Unit — P/in reqd. — 99
- 5305B 1300 MHz Counter for 5300 — 325
- 5308A 75 MHz Counter Timer for 5300B — 112

RACAL
- 9024 600 MHz 7½ digit Counter — 220
- 9025 1 GHz 8 digit Counter — 450
- 9905 200 MHz 8 digit Counter Timer — 360

SYSTRON DONNER
- 6053 3 GHz 9 digit Counter BCD O/P — 790
- 5103B Strip Printer for 6053/6054 — 375

DIGITAL TESTING EQUIPMENT
HEWLETT PACKARD
- 1600A Logic Analyser 16ch 20 MHz — 2300
- 1607 Logic Analyser 16ch 20 MHz — 1500
- 5011T Logic troubleshooting kit — 125

TEKTRONIX
- 7D01F Logic Analyser 16ch 50 MHz P/in — 2650
- 832 Datacomm Test V24/RS232/I loop — 1150

MAINS TEST EQUIPMENT
COLE
- T1007 Volt/Freq/Spike Monitor Rec D/P — 110

DATALAB
- DL019 Mains Interface for DL905 — 300

GAY
- LDM AC/DC/Spike/Time inc Printer — 1250

MISCELLANEOUS
AVO
- RM215 AC/DC Breakdown/Leakage Tester — 475

COMARK
- 1601BLS Thermom 10ch 87 + 1000°C type K — 75

N.B. Thermocouples not included

DATALAB
- DL901 Digital Transient Recorder — 500

HEWLETT PACKARD
- X382A Rotary Vane Attenuator WG16 — 175

Prices from £

MULTIMETRICS
- AF120 Dual H/Pass L/Pass active filter 20 Hz – 2 MHz — 600

TEKTRONIX
- 521PAL Vectorscope — 2200
- 528 TV Waveform Monitor — 750
- 575 Semiconductor Curve Tracer — 425
- 1485C TV Waveform Monitor PAL/NTSC — 2300

NETWORK ANALYSERS/ PHASEMETERS
DRANETZ
- 305B/3001 Phasemeter 2Hz-700KHz — 990

HEWLETT PACKARD
- 8405A Vector Voltmeter 1–1000 MHz — 2000
- 8414A Polar Display for 8410 N.W.A. — 700

THIS MONTH'S SPECIAL OFFERS

PHILIPS PM 8251 CHART RECORDER
single pen writing on 10 inch chart paper.
12 chart speeds + 12 voltage input ranges from 3.5mV to 50V F.S.D.
£325

HEWLETT PACKARD 427A MULTI-FUNCTION METER
with linear and dB scale.
DC Volts 100mV-1000V F.S.D.
AC Volts 10mV-300V (10Hz-1MHz)
OHMS 10Ω-10MΩ F.S.D.
Battery operated
£135

PHILIPS PM 5129 FUNCTION GENERATOR
1 mHz to 1 MHz sine/square/triangle/pulse/ramp 30V p.p. output max.
Variable over 60dB.
Internal FM and sweep
Single shot and burst modes
±10V variable D.C. offset.
£595

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OSCILLOSCOPES & ACCESSORIES
GOULD ADVANCE
- OS3300B 50 MHz 1mV 2 Trace 2T base — 675

HEWLETT PACKARD
- 1804A 50 MHz 20mV 4 Trace Plug-in — 625
- 1825A Dual Timebase Plug-in — 500
- 1805A 100 MHz 5mV 2 Trace Plug-in — 625

PHILIPS
- PM3211 15 MHz 2mV 2 Trace TV trig — 390
- PM3212 25 MHz 2mV 2 Trace TV trig — 550
- PM3244 50 MHz 5mV 4 Trace 2T base — 1450
- PM3260 120 MHz 5mV 2 Trace 2T base — 1450
- PM3262 100 MHz 5mV 2 Trace 2T base Tr View — 1150

Prices from £

TEKTRONIX
- 465 100 MHz 5mV 2 Trace 2T base — 1150
- 465B 100 MHz 5mV 2 Trace 2TB, inc Probes — 1550
- 475 200 MHz 2mV 2 Trace 2T base — 1750
- 485 350 MHz 5mV 2 Trace 2T base — 2300
- 661/4S3/5T1A 1 GHz Sampling scope — 775
- 7A12 105 MHz 5mV 2 Trace Plug-in — 300
- 7A18 75 MHz 5mV 2 Trace Plug-in — 400
- 7A19 500 MHz 10mV 1 Trace Plug-in — 950
- 7A22 1MHz 10µV Differential Plug-in — 575
- 7A24 350 MHz 5mV 2 Trace Plug-in — 950
- 7A26 200 MHz 5mV 2 Trace Plug-in — 625
- 7B53A 2 Timebase Plug-in 100 MHz Trig — 500
- 7B80 Single Timebase 400 MHz Trig — 550
- 7B85 Timebase with delay 400 MHz Trig — 650
- 7603 100 MHz CRT r/out 3 slot M/Frame — 1250
- 7704A 200 MHz CRT r/out 4 slot M/Frame — 1350
- 7904 500 MHz CRT r/out 4 slot M/Frame — 3900
- P6013A X1000 12KV Probe — 95
- P6201 FET Probe DC-900 MHz — 300

TELEQUIPMENT
- D63/V1/V1 15 MHz 2 Trace 1mV — 525
- D83/V4/S2A 50 MHz 1mV 2 Trace 2T Big CRT — 750
- D1015 15 MHz 5mV 2 Trace TV trig — 295
- S1A Single T/base Plug-in 50 MHz trig — 135

VUDATA
- PS935/975 35 MHz 5mV 2 Trace – unit has built-in 3½ digit DMM + 3½ dig. cnter — 675

Note: we hold a range of cameras — P.O.A.

OSCILLOSCOPES (STORAGE)
HEWLETT PACKARD
- 1703A 35 MHz 10mV 2 Tr 2TB 1000 Div/ms — 1400

TEKTRONIX
- 466 100 MHz 5mV 2 Tr 2TB 1350cm/µs — 2950
- 603 Bistable Storage Monitor XYZ amps — 750
- T912 10 MHz 2mV 2 Tr 1TB 250cm/ms — 650
- 7834 400 MHz 4 Slot M/Frame 2500cm/µs — 4990

POWER MEASUREMENT
FLUKE
- 8921A 10 Hz-20 MHz 4½ Digit & Analogue TRMS dBm/V meter 50Ω-1200Ω & 10MΩ — 885

HEWLETT PACKARD
- 8481A Type N Coax sensor for 435A — 200
- 478A Type N Coax sensor for 432A — 90

MARCONI
- TF2512 DC-500 MHz Powermeter — 145
- TF893A 10 Hz-20 KHz Powermeter — 135

POWER SUPPLIES etc
ADVANCE
- 1V5S Inverter 24V DC to 240V AC 500W — 300

FARNELL
- FFSL 5 V – 20 A PSU module — 100
- L30B 0-30V variable 1A Metered — 50

FLUKE
- 415B 0-3.1 KV variable 30mA Metered — 550

HEWLETT PACKARD
- 6966A 0-36 V variable 10 A metered — 250

MARCONI
- TF2154/1 0-30V variable Metered — 75

PHILIPS
- PE1646 0-75V variable 6A Metered V + I — 495

PULSE GENERATORS
ADVANCE
- PG57 10 Hz-50 MHz 10V 50Ω Vari RT 6ns — 190

EH RESEARCH
- 132 10 Hz-3.5 MHz 50V 50Ω RT 10ns 2 pulse — 120

MARCONI
- TF2025 0.2 Hz-25 MHz 10V 50Ω RT 7ns 2 pulse — 350

RECORDERS & ACCESSORIES
BRUNO WOELKE
- ME102B Wow and Flutter meter — 75

BRYANS SOUTHERN
- BS314 Chart 10" 4 Pen 16 speed — 1950
- BS316 Chart 10" 6 Pen 16 speed — 2500

HEWLETT PACKARD
- 7015A XY 1 pen A4 size — 700
- 7046A XY 2 pen A3 size — 995

MEDELEC
- M/scope Combined 4 ch scope + UV rec'dr — 1900

SE LABS
- 994 6 ch galvo preamp + DC bridge supply — 450
- 6008 UV chart 8" 25 ch 16 speed — 950
- 6150/51 UV recorder 12 ch-inc 6 ch amps — 1000

Prices from £

SMITHS
- RE541 Chart 8" 1 pen 8 speed — 250
- RE501/4701 Cht 4" + XY 1ch 10 spd AC Batt — 200

SOLARTRON
- 3240 Modular Data Logger system — P.O.A.

Note: UV recorders are priced less galvos

SIGNAL ANALYSIS EQUIPMENT
MARCONI
- TF2300A Mod Meter 1 MHz-1 GHz AM/FM — 450
- TF2330 Wave Analyser 20 Hz-50 KHz — 425

Note: see also "Spectrum Analysers"

SIGNAL/FUNCTION/ + SWEEP GENERATORS
ADVANCE
- SG63D Generator 4-230 MHz AM/FM — 200

GENERAL RADIO
- 1362 Generator 220-920 MHz — 375

HEWLETT PACKARD
- 8640B Generator 500 KHz-512 MHz AM/FM Phase Lock — 3800
- 618B Generator 3.8-7.5 GHz — 975
- 612 Generator 450-1230 MHz — 750
- 614 Generator 0.8-2.1 GHz — 825

MARCONI
- TF144H/4S Generator 10 KHz-72 MHz AM — 550
- TF801D Generator 10 MHz-470 MHz AM — 180
- TF2012 Generator 400-520 MHz FM — 550

PHILIPS
- PM5127 Function 0.1 Hz-1 MHz Sin Sq Tri Rmp — 450

TEXSCAN
- 9900 Sweeper 10-300 MHz 6/in CRT disp — 525
- VS60 Sweeper 5-1000 MHz — 890

SPECTRUM ANALYSERS
HEWLETT PACKARD
- 141T/8552B/8555A Complete .01-18 GHz — 8500
- 3580A 5 Hz-50 KHz with digi store disp — 2450
- 8445A Pre-selector 0.01-18 GHz — 2000
- 8558B 0.1-1500 MHz Plug-in for 180 series — 3750

VOLT/MULTI-METER (ANALOGUE)
AVO
- 8 Mk4 AC/DC/—VI + Ω — 70

BOONTON
- 92C AC/RF 10 KHz-1.2 GHZ ½mV-3V — 350

HEWLETT PACKARD
- 400E 10 Hz-10 MHz 1mV-300V DC D/P — 285
- 400H 10 Hz-4 MHz 1mV-300V — 75
- 411A 0.5-500 MHz 10mV-10V DC D/P — 175
- 3400A TRMS 10 Hz-10 MHz 1mV-300V DC O/P — 390

MARCONI
- TF2603 50 KHz-1.5 GHz 300µV-3V — 300
- TF2604 20 Hz-1.5 GHz 300mV-300V — 425

PHILIPS
- PM2454B 10 Hz-12 MHz 1mV-300V DC D/P — 250

RACAL
- 9301 RMS 10 KHz-1.5 GHz 100µV-300V — 550

VIBRON/E.I.L.
- 33B-2 1mV-1V Electrometer — 200

VOLT/MULTI-METER (DIGITAL)
ADVANCE
- DMM7A 1999FSD AC/DC/V/I/Ω — 100

BOONTON
- 92AD 1999FSD 10 KHz-1.2 GHz 10µV res — 525

FLUKE
- 8010A 2000 FSD TRMS AC/DC/VIΩ — 140
- 8010A01 As 8010A + re-charging batteries — 159
- 8020A 2000 FSD Handheld AC/DC/VIΩ + cond. — 99
- 8022A 2000 FSD Handheld AC/DC/VIΩ — 75
- 8030A-1 2000 FSD AC/DC/VIΩ Batt + AC — 165
- 8050A 20000 FSD AC/DC/VIΩ dB TRMS — 215
- 8800A 200000 FSD AC/DC/VΩ — 550

HEWLETT PACKARD
- 3490A 100000FSD AC/DC/V/Ω — 625

SOLARTRON
- A200 19999FSD DC only 1µV-1 KV — 85
- A203 19999FSD AC/DC/V/Ω — 175
- A205 19999FSD TRMS AC/DC/V/Ω — 195

Carston Electronics Ltd
01-267 5311
Shirley House, 27 Camden Road, London NW1 9NR. Telex: 23920.

Full details and specification of equipment listed, available. Because of long copy dates this list is not comprehensive – ring for inventory update or tell us your SPECIFIC NEEDS. Hours Monday to Friday 9.00 am-5.30 pm (4.30 pm Fridays). **Prices exclude delivery and VAT.** We take Access or Barclaycard.

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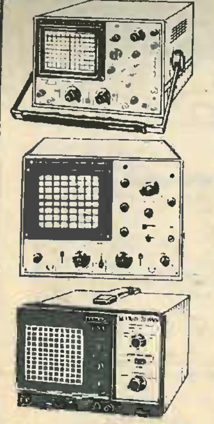


TRIO OSCILLOSCOPES

Range of mains operated Scopes with 5" displays, triggered sweep (UK c/p £3.50)

DUAL TRACE

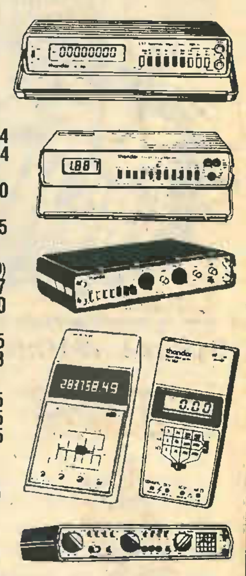
CS1562A 10 MHZ: 10 MV: 1 micro sec.	£267.95
CS1560A II 15 MHZ: 10mV: 0.5 micro sec.	£341.55
CS1566A 20 MHZ: 5mV: 0.5 micro sec.	£363.40
CS1577A 35 MHZ: 2mV: 0.1 micro sec.	£523.25
CS1830 Mk II 30MHZ, 2mV, 0.2 micro sec. (fitted delay line)	£626.75
CS1575 5 MHZ, 1mV, 0.5 micro sec. Multi display Audio scope.	£312.80

SINGLE TRACE

CS1559A 10 MHZ, 10mV, 1 micro sec. 5" display Triggered sweep (UK c/p £3.50)	£251.85
CO1303D 5 MHZ, 10mV, low sweep for observation below 1 HZ and up to 450 MHZ, 75mm display (UK c/p £2.00)	£124.20

Optional probes – all models.
X1 £8.50: X10 £10.50: X1-X10 £12.95 X100 £16.95

THANDAR – SINCLAIR



Reliable low cost portable instruments, bench models all 25.5 x 15 x 5cm. Generators mains operated rest battery (supplied). UK c/p Hand models 65p, bench £1.15)

DIGITAL MULTIMETERS (3½ digit LCD)

TM354 Hand held, DC 2A, 2m ohm, 1mV – 1000V DC, 500v AC.	£45.94
TM352 Hand held, DC 10A, Hfe test. Continuity test	£57.44
TM353 Bench, 2A AC/DC, 1000V AC/DC. 20M ohm. Typical 0.25%	£96.60
TM351 Bench, 10A AC/DC, 1000V AC/DC. 20M ohm Typical 0.1%	£143.85

FREQUENCY COUNTERS (8 Digit)

PFM200 Hand held LED. 200 MHZ, 10mV (600 MHZ with TP600)	£57.27
TF040 Bench LCD. 40 MHZ, 40mV (400 MHZ with TP600)	£126.50
TF200 Bench LCD. 200 MHZ, 10-30mV (600 MHZ with (TP600)	£166.75
TP600 600 MHZ + 10 Prescaler 10 mV	£43.13

GENERATORS (All bench models) mains operated

TG100 Function. 1 HZ-100 KHZ. Sine/SQ/Triangle/TTL	£90.85
TG102 Function. 0.2 HZ-2 MHZ. Sine/SQ/Triangle/TTL	£166.75
TG105 Pulse. 5 MHZ-5HZ (200nS-200mS) various outputs	£97.75

OSCILLOSCOPE (Bench model low power portable) 10 MHZ 2" trace. 10mV. 0.1 micro sec. All facilities.
Model SC110 £159.85
(Rechargable battery pack £8.63, AC adaptor/charger £5.69

OPTIONAL ITEMS
Carry case (bench only) £6.84 AC Adaptors (state model) £5.69

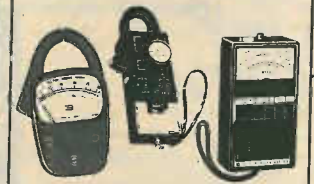
MULTIMETERS

(UK c/p 65p or £1.00 for two)

CHOOSE FROM UK's LARGEST RANGE

KRT101 10 range pocket 1K/Volt	£4.50
KRT100 12 range pocket 1K/Volt	£5.50
NH55 10 range pocket 2K/Volt	£6.50
AT1 12 range pocket Deluxe 2K/Volt	£7.75
ST5 11 range pocket 4K/Volt	£8.50
NH56 22 range pocket 20K/Volt	£11.95
YN360TR 19 range plus Hfe Test 20K/Volt	£13.50
ST303TR 21 range plus Hfe Test 20K/Volt	£15.95
KRT5001 16 range – range double 50K/Volt	£16.50
AT1020 19 range Deluxe plus Hfe Test 20K/Volt	£16.95
ETC5000 As KRT5001 plus colour scales 50K/Volt	£17.95
7081 18 range – range double 10A DC 50K/Volt	£20.85
TMK500 23 range. Plus 12A DC Plus Cont. Buzzer 30K/Volt	£22.75
AT205 21 range Deluxe 10A DC 50K/Volt	£25.75
C7080 26 range large scale 10A DC. 5KV AC/DC 20K/Volt	£27.50
AT210 23 range Deluxe 12A AC/DC 100K/Volt	£31.00
360TR 23 range Large scale 10A AC/DC Hfe Test 50Meg ohm. 1KV AC/DC 100K/Volt	£34.95

CLAMP-ON-METERS INSULATION TESTERS



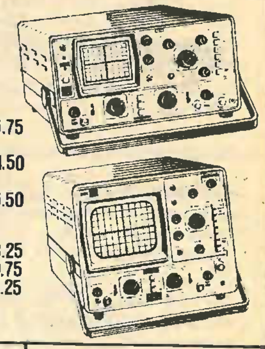
Multi-range clamps all with resistance range, carry case & leads. Also digital and OC clamp in stock (UK c/p 75p)

ST300 300A 600V 9 ranges	£25.95
ST310 300A 600V 9 ranges	£29.95
K2602 150A, 600V, AC 7 ranges	£35.95
*K2606 300A, 600V, AC 8 ranges	£44.00
K2803 300A, 600V, AC 9 ranges	£53.95
K2903 900A, 750V, AC 9 ranges	£77.50
K2103 1000A, 750V, AC 9 ranges	£95.00
*Optional temperature probe	£13.80

ELECTRONIC INSULATION TESTERS
Battery operated complete with carry case (UK c/p £1.00)

YF500L 500V/100Meg. Plus 0-100 ohm	£55.00
K3103 600V/100Meg. Plus 0-2.6K ohm	£95.00
K3106 500V & 1000V. 1000 & 2000Meg.	£104.00
K4101 Earth resistance tester	£136.50
M500 Hand cranked insulation tester 500V/100Meg.	£67.50

CROTECH OSCILLOSCOPES

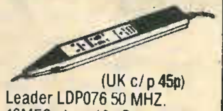


Range of Portable Scopes mains and battery operated. Plus special features (UK c/p £3.00)

3030 Single trace 15 MHZ, 5 mV. 0.5 micro secs. Plus built in component tester, 95mm tube	£166.75
3131 Dual trace 15 MHZ, trig to 35 MHZ, 5mV, 0.5 micro sec. 130mm tube, plus component tester.	£264.50
3034 Battery-mains dual trace 15 MHZ, trig to 20 MHZ, built in Nicads, 5mV, 0.5 micro secs. (Eliminator charger optional £28.75)	£356.50
Also Available 3033, single trace 3034	£293.25
3035, 130mm 3030	£189.75
3337, dual 30 MHZ, 130mm	£408.25

(Optional Probes all models – see Trio above)

LOGIC PROBE



(UK c/p 45p)
Leader LDP076 50 MHZ. 10MEG ohm. 10n Sec. with carry case £56.90

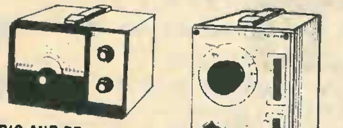
DIRECT READ HV PROBE

(UK c/p 65p)
0/40KV: 20K Volt £18.40

OSCILLOSCOPE PROBE KITS

(UK c/p 50p per 1 to 3)
Available BNC plug or Banana X1 £8.50: X10 £10.50: X1-X10 £12.95
Also X100 (BNC only) £16.95

TRIO GENERATORS

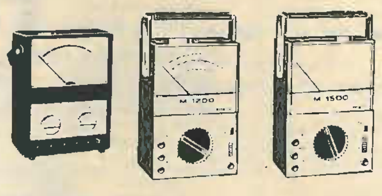


AUDIO AND RF
MAINS OPERATED PORTABLE EQUIPMENT (UK c/p £1.10)

SG402 6 range RF 100 KHZ – 30 MHZ ± 1.5, INT/EXT MOD. 100 mV o/p Variable	£68.00
AG202A Matching 4 range Audio, 20 HZ – 200 KHZ Sine Square. Distortion 0.5-1% Variable o/p to 10V	£78.20
AG203 5 band 10 HZ – 1 MHZ Sine/Square: 0.5dB deviation. Distortion Max. 0.1% Switched Alternator & variable.	£126.50

ALSO IN STOCK Newtronics 200P & 200SPC 0.002 HZ 5.5 MHZ Generators (200 SPC with 100 MHZ counter)

PROFESSIONAL MULTIMETERS

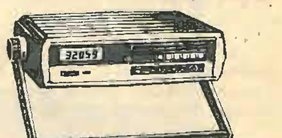


(UK c/p £1.50) All featuring AC/DC Volts/Current & Ohms ranges.

M1500 43 range 20K/Volt: AC/DC 10A	£53.50
M1200 30 range 100K/Volt: AC/DC 15A 200 Meg ohm.	£67.50
K1400 26 range large scale 20K/Volt: 10A AC/DC: 20 Meg ohm: 5kV AC/DC	£79.95
K200 39 range 10 Meg ohm input. 25 HZ-1 MHZ	£95.00

OPTIONS Cases: M 1500 & 1200 £16.50: K1400 £19.00
Temperature Probe for K1400 £13.80

THURLBY DIGITAL MULTIMETER



4¾ Digit. 0.05% 7 Function LCD
Made to exacting standards in UK
30 ranges: 1200V DC, 750V AC, 10A AC/DC, 32 Meg ohm. Also includes frequency measurement to 4 MHZ and 4 KHZ output.
Price is with batteries, test leads and mains adaptor.
(optional carry case £20.45)

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UK c/p £1.00

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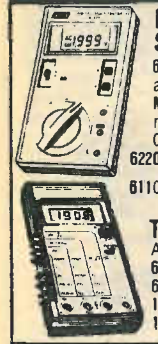


LCD DIGITAL MULTIMETRES SPECIAL PURCHASE – LIMITED PERIOD ONLY

6220 Reliable 22 range hand held 3½ digit LCD with volt/ohms auto range, unit and range signs, 10 amp AC/DC, battery warning, lower power ohms range; Model 6110 Also has range hold, continuity buzzer and improved accuracy. All models high quality rotary operation. Resolution 0.1 milli volt: 10-Micro amp: 0.1 ohm.

6220 1000v DC: 0.2/10A AC/DC: 600v AC: 2meg ohm. Was £55.95 **NOW £42.95**

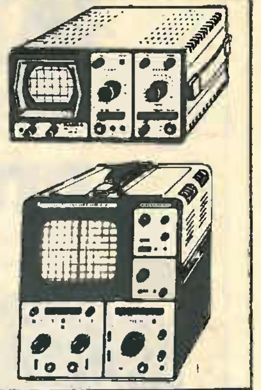
6110 As above plus 20mA AC/DC and improved accuracy. Was £85.95 **NOW £59.95**

THIS SPECIAL OFFER IS QUALITY WITH VALUE

Also in stock.

6200 20/200mA version of 6220 (i.e. no 10 amp)	£37.95
6100 0.2A version of 6110 (i.e. no 10 amp)	£49.95
188m 16 range with Hfe checker	£43.50
189m 30 range with Hfe checker	£69.95

HAMEG OSCILLOSCOPES



Range of top quality scopes for Amateur and Professional (UK c/p '307' £3.00, other £4.00)

307 Single trace 10 MHZ: 5mV: 0.5 micro sec. Plus built in component tester. 6 x 7cm display. (Optional carry case £18.40)	£158.70
203 Dual 20 MHZ: Trig to 30 MHZ: 5mV: 0.5 micro secs: 8 x 10cm display (replace model 312)	£253.00
412-5 Dual 20 MHZ delayed sweep: trig to 40 MHZ: 5mV: 0.1 micro sec. 8 x 10cm display.	£402.50
512 Dual 50 MHZ. Delayed sweep: Single sweep: Delay line: Trig to 70 MHZ: 5mV: 0.1 micro sec. 8 x 10cm display	£667.00
Options 203/412/512 Viewing hood	£6.90
Component tester 203, 412, 512	£46.00
Carry case (state model)	£21.85

Optional Probes (All models) X1 £8.50: X10 £10.50
X1-X10 £12.95: X100 £16.95

GLOBAL SPECIALISTS EQUIPMENT

Generators (UK c/p £1.00)

2001 1 HZ-100 KHZ sweepable function generator. TTL/Sine/Square/Triangle. All facilities: Mains	£89.00
4001 Ultra Variable Pulse Generator: 0.5 HZ-5 MHZ Every facility: Mains.	£109.00

Frequency Counters (UK c/p 80p)

MAX100 100 MHZ Counter. Battery operated. 8 Digit 30mV typical. Bench portable LED	£89.00
MAX50 6 Digit 50 MHZ Hand held LED	£56.00
MAX550 2 range 6 Digit 550 MHZ hand held LED	£97.75
PS500 500 MHZ prescaler for MAX50/100	£34.50

Circuit Powered Logic Probes Pulser and Monitors (UK c/p 65p)

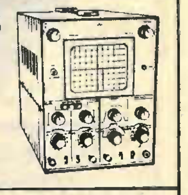
LP1 DTL/TTL/CMOS: 10 MHZ. 100K ohm: Pulse: Memory	£35.50
LP2 DTL/TTL/CMOS: 1.5 MHZ. 300K ohm: Pulse	£19.95
LP3 DTL/TTL/CMOS: 50 MHZ. 500K ohm: Pulse: Memory	£55.95
LM1 All, 8, 10, 12, 14 & 16 Pin IC'S. Indicates Pins on/off DTL/TTL/CMOS/HTL. Logic Monitor	£33.00
OP1 Digital Pulse. Single shot or 100pps.	£58.50

KEITHLEY PROFESSIONAL DIGITAL MULTIMETER

Model 130, 25 range. Easy to hold and use LCD DMM. Size 7 x 3.1 x 1.5

Ranges
DC Volts 200mV-1000V 0.5% 100 micro volt
AC Volts 200mV-750V 1% 100 micro volt
DC current 2mA-10AMP 1-2% 1 micro amp
AC current 2mA-10AMP 2% 1 micro amp
Resistance 200 ohm-20 Meg 0.5% 0.1 ohm £102.35

SAFGAN PORTABLE OSCILLOSCOPES

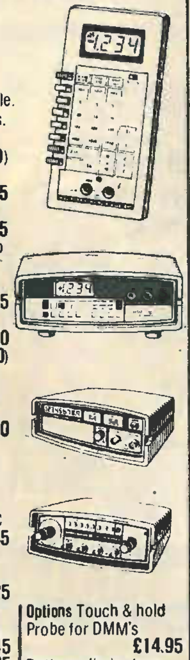


Range of low cost Dual Trace Scopes mains operated. Made in UK to exacting standards. Available as 10 MHZ: 15 MHZ or 20 MHZ. All feature 5mV sensitivity: 0.5 micro sec: 6.4 x 8cm display. (UK c/p £2.50)

DT410	Dual 10 MHZ	£194.35
DT415	Dual 14 MHZ	£201.25
DT420	Dual 20 MHZ	£216.20

Optional probes available (see Hameg above)

SABTRONICS EQUIPMENT



New reliable range of DMM's and frequency computers with those extra facilities and competitive prices. All battery operated (supplied). Except 5020A mains. Optional mains eliminators available.

8 DIGIT COUNTERS 0.1 HZ to 10 HZ Res. 10mV sensitivity to 100 MHZ (UK c/p £1.00)

8110A 20 HZ-100 MHZ in 2 ranges	£95.45
8610A 20 HZ-600 MHZ in 3 ranges	£113.85

9 DIGIT COUNTERS 30mV sensitivty to 1GHZ. Resolution 0.1 HZ-10 HZ

8610B 10 HZ-600 MHZ in 3 ranges.	£125.35
8000B 10 HZ-1GHZ in 3 ranges	£184.00

FUNCTION GENERATOR (UK c/p £1.00) with mains adaptor
5020A 1 HZ-200 KHZ Sine/Square/Triangle/TTC Freq. sweep. Low distortion £98.90

DIGITAL MULTIMETERS Two LCD hand held – one with temperature range. Also LCD and LED Bench models.

2035A 3½ digit LCD hand. 2A AC/DC 20Meg ohm ETC	£95.45
2037A As 2035A with -50°C to + 150°C Temp. range 0.1° resolution	£109.25
2010A 3½ Digit LED. Auto decimal & minus. 10A AC/DC: 20Meg ohm etc.	£95.45
2015A LCD version of above.	£109.25

(c/p 2035/37A 65p: All others £1.00)

Options Touch & hold Probe for DMM's £14.95
Battery eliminators (state model) £5.69

VARIABLE REGULATED POWER SUPPLIES

Mains operated regulated single metre (UK c/p £1.50)

241 0/12-0/24V 1 amp	£35.00
154 5-15 volt 3 amp	£44.00
423 0/12-0/24V 3 amp	£54.00

AMATEUR/CB

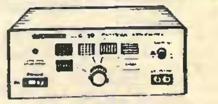
KDM6 1.5 to 250 MHZ Dip meter 6 ranges	£38.50
SWR9 3-150 MHZ SWR + F/S	£9.50
SWR50 Twin meter SWR + Power 1KW max.	£13.95
310 Single meter SWR + Power 10W	£8.95
110 SWR/Power/FS-10/100W	£11.95
171 As '110' Twin meter	£14.50
175 SWR/FS/AE Match (40 MHZ)	£13.80
176 As '175' + 0/5/50 Watt power	£16.95
178 As '175' + 0/10/100W + MOD Scale	£19.50
HM20 SWR meter Plus 20K/Volt 19 range Multimeter	£28.95

(Note: SWR-Power ETC to 144/150 MHZ)
Just a selection of a huge range in stock – send for latest lists including professional ranges.

LASCAR BENCH MULTIMETER

3½ Digit LCD Made in UK: 25 ranges with basic 0.1% accuracy: 2A AC/DC with 0.1, Microamp resolution: 1KV AC/DC 20Meg ohm ETC.
LM100 £89.00 (UK c/p £1.00)
Also model LM2001 hand held model 2Meg ohm. 2A AC/DC ETC. £57.70 (UK c/p 65p)

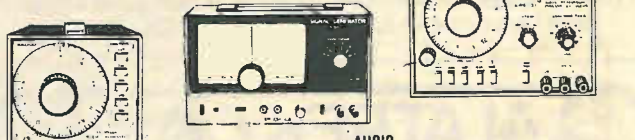
TV COLOUR BAR PATTERN GENERATORS



PAL UHF and VHF Models

LGC393 VHF 6 pattern	£143.75
LCG392u UHF 15 pattern	£228.85
LCG392v VHF 15 pattern	£231.15
LCG399 VHF/UHF 13 pattern	£572.70

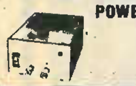
LEADER AUDIO RF FM TV GENERATORS



High quality mains operated equipment (UK c/p £1.50)
LSG16 RF Generator (matches LAG26) 5 ranges 100 KHZ-100 MHZ (300 MHZ harmonics). Int/Ext MOD: 100mV o/p: ± 1.5% accuracy £63.25

AUDIO
LAG26 4 range 20 HZ-20 KHZ. Sine/Square: 5V o/p: Distortion 0.5-1% £73.60
LAG120A 5 range 10 HZ-1 MHZ. Sine/Square: 3V o/p: 0.05-0.8% Max Distortion Switched attenuator. £146.00
LAG125 Low distortion (0.02%) version of LAG120A Plus 'Burst' type signals for Speaker test £273.00

POWER SUPPLIES



13.8 volt. Output regulated mains operated (c/p £1.00)

3 to 5 amp	£15.50
5 to 8 amp	£17.95

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AUDAX WFR15S	£23.95
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COLES 3100	£7.95
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CELESTION HF2000	£13.60
DALESFORD D10 ferrofluid	£8.95
DALESFORD D30/110	£11.95
DALESFORD D50/153	£12.95
DALESFORD D50/153 polymer	£15.25
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DALESFORD D100/250	£29.95
DALESFORD D300	£35.75
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JORDAN WATTS Module Mk III	£28.50
JORDAN WATTS HF kit	£10.50
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KEF B110	£12.25
KEF B200	£13.50
KEF B139	£27.75
KEF DN12	£9.95
KEF DN13 SP1106	£7.75
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LOWTHER PM6 Mk I	£67.75
LOWTHER PM7	£99.00
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RICHARD ALLAN DT20	£10.75
RICHARD ALLAN DT30	£11.50
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SHACKMAN ES units with cabs.	£170 pair
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TANNOY DC316	£159.00
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VOLT B250	£58.00
VOLT 10" ABR	£32.00
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WILMSLOW AUDIO

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Hi Fi News Tabor with H4 bass units	£73.50
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All prices include VAT at 15% and are correct at 1.6.81.

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4" to 6½"	90p each
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12", 13" x 8", 14" x 9"	£2.00 each
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18"	£5.00 each
Speaker Kits	£3.95 pair
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WILMSLOW AUDIO WE50. Total kit – Wharfdale E50 kit plus WE50 flat-pack cabinet kit including reflex port trims, acoustic wadding grille fabric, nuts, bolts, terminals etc **£182 pair. Carriage £8.**

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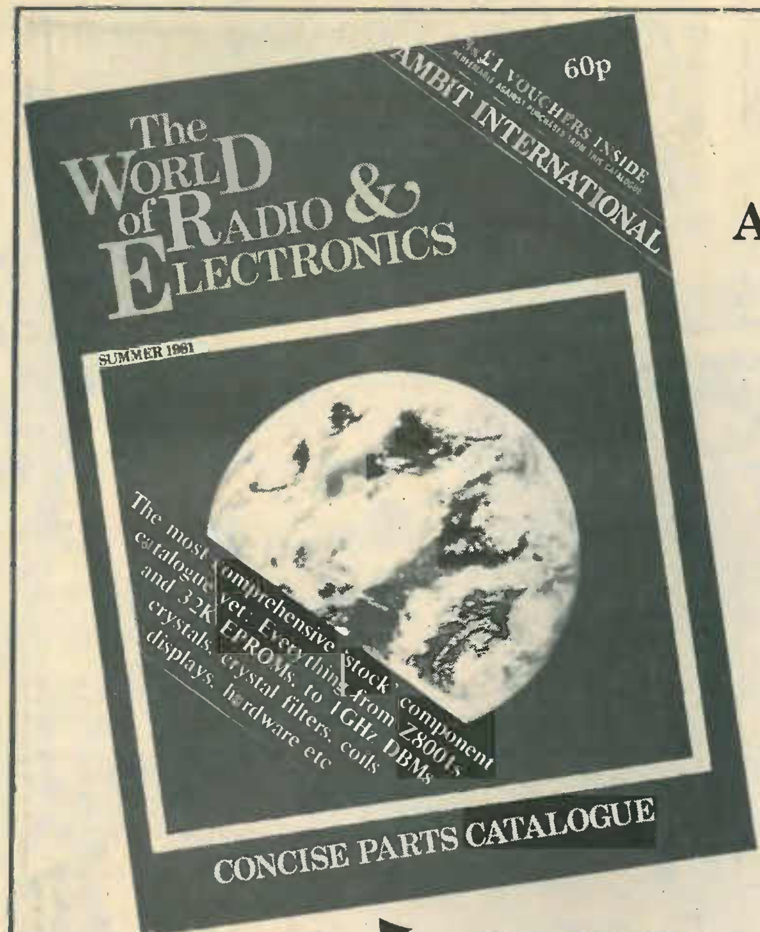

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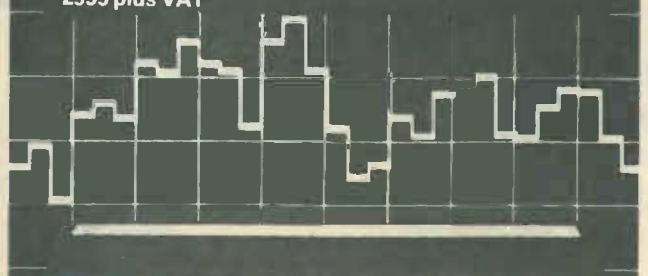

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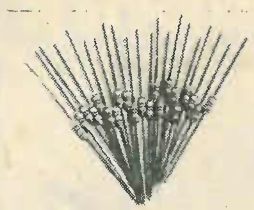

100R	1k	10k	100k
110R	1k1	11k	110k
120R	1k2	12k	120k
130R	1k3	13k	130k
150R	1k5	15k	150k
160R	1k6	16k	160k
180R	1k8	18k	180k
200R	2k	20k	200k
220R	2k2	22k	220k
240R	2k4	24k	240k
270R	2k7	27k	270k
300R	3k	30k	—
330R	3k3	33k	330k
360R	3k6	36k	—
390R	3k9	39k	—
430R	4k3	43k	—
470R	4k7	47k	470k
510R	5k1	51k	—
560R	5k6	56k	560k
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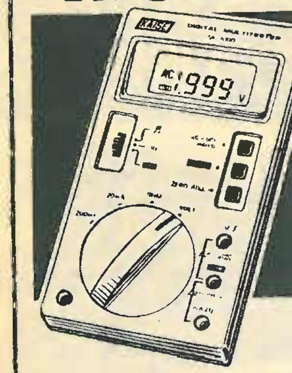
Practical Wireless

LCD DIGITAL MULTIMETERS

ALL PRICES INCLUDE VAT



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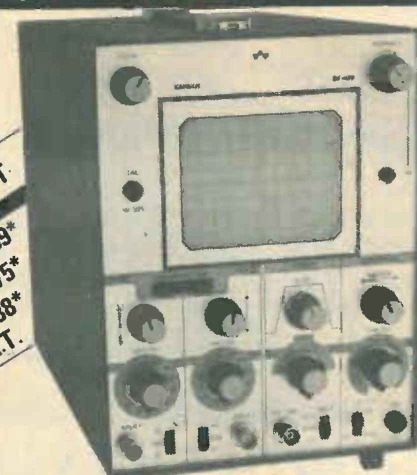

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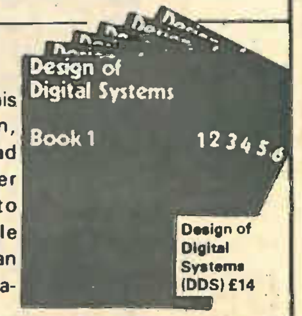

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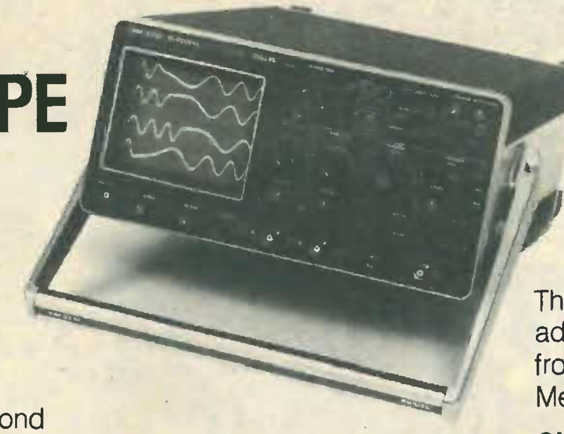

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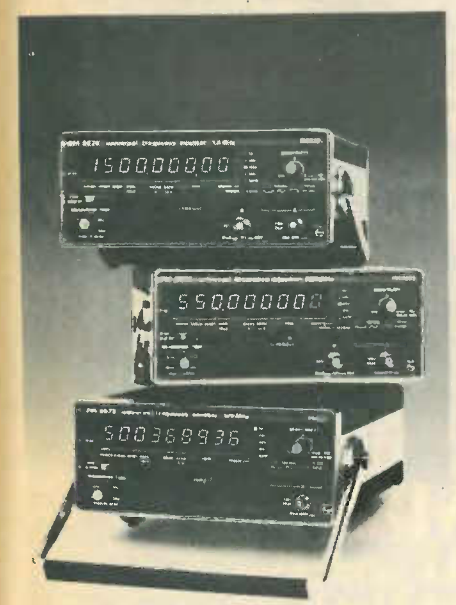

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wireless world

Invention – the orphan

In the last century, life for anyone but the intentionally static, self-sufficient and unaware was restricted, although it is probable that few were seriously worried by the restrictions: a delay of several days in hearing the news that Wellington had won, or a journey from York to London that occupied three whole days would cause little trouble. Nevertheless, any improvement in mobility, the spread of information and education and an easing of the severity of life in general was greatly to be desired. The Industrial Revolution had its origin in these conditions.

Since human nature does not change rapidly, if at all, it is unlikely that many of the engineers and scientists who brought about the fundamental change in life style of the western world did so with any sense of altruism. Then, as now, a man had an idea and was unable to rest until he had a piece of hardware that worked; and if it did, there was a chance of making some money out of it. There is nothing whatever wrong with that – it is a dream cherished by most engineers – and the inventions changed the world for the better, in most cases. The process was apparently logical: an engineer saw a need and proceeded to satisfy it. It may have been that the idea was attractive technically and the inventor would have gone ahead without any other stimulus, but needs were numerous and almost any advance in engineering was useful in some field or other.

Long before the middle of the present century, the majority of man's pressing material and cultural needs had been attended to, in the 'developed' countries, at least. But the drive to be inventive persisted: provision began to precede requirement and eventually to create it – not once, but annually. In the field of technology concerned with domestic, as opposed to industrial engineering, it is now commonplace for companies mesmerized by their own expertise not to perceive a need, but to satisfy a non-existent one. Devices are designed and produced in vast numbers before the public has shown any indication of wanting them or even knowing what they are, simply because they are technically possible. Not only that, but before the first round of production and the subsequent 'creation' of a market for it is finished, the next version is hurled at us, in slightly modified form and possibly incompatible with the first.

In recent years, this inversion has occurred at least four times. In the early 1970s, perfectly ordinary citizens suddenly discovered an inescapable need to possess pocket calculators. These devices were made solely because it was possible to make them, but that having been done, the market creation had to begin. Before long we were seeing housewives using calculators to add up their supermarket bills: they do not do that now – the 'need', created by advertising, evaporated as quickly as it was formed.

The same process brought into being the digital watch. It was not easy to make the digital output drive hands, so the inferior numerical display was adopted as a substitute. Advertising created a demand for the watches – no public outcry had forced their development – and we will no doubt find hands in fashion again quite soon. Passing over the sorry business of quadraphonic sound, in which the participant companies were too clever for their own good and did not manage to persuade the public to think otherwise, we have now reached the video disc, a development which appears to have little to offer over video tape, and which could conceivably prove to be the sticking point for a baffled and possibly resentful public. In this case, not only does the need not exist; it didn't exist when it was satisfied the first time, with tape machines, in the domestic sphere at any rate.

To pursue technology for its own sake and to pay for it by exploiting the public's total and uncomprehending belief in technology is, at the very least, open to question. A professional soldier may become a mercenary if he runs out of official wars, but an engineer has no need of that – the world is full of ready-made problems to solve without inventing them.

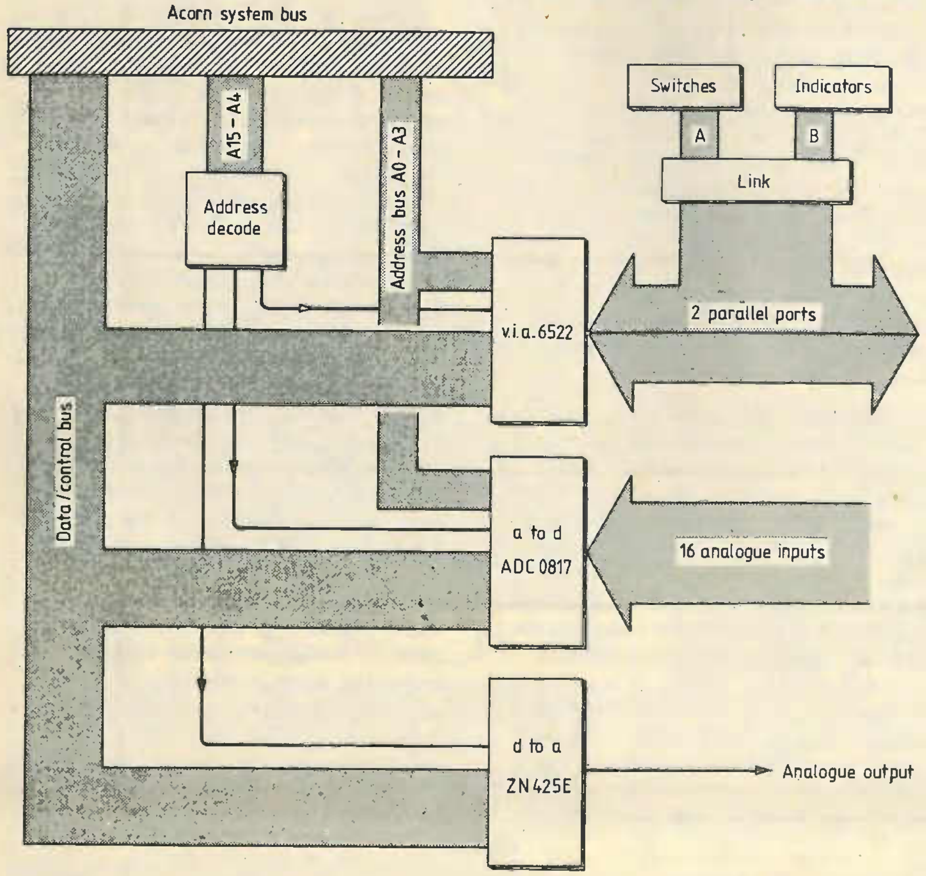
Interfacing microprocessors

Design, operation and application of a "universal" interface board

by **J. D. Ferguson,** B.Sc., M.Sc., M.Inst.P., **J. Stewart, and P. Williams,** B.Sc., Ph.D., M.Inst.P.
Microelectronics Educational Development Centre, Paisley College of Technology.

By using a range of practical circuits, this series of articles explains how to interface microprocessors to other electronic and electromechanical systems. The emphasis throughout is on providing economical solutions for educational and industrial applications, rather than achieving the ultimate performance. Part one describes a "universal" interface board which is directly compatible with the 6502, and later articles describe hardware and software modifications for other microprocessors. A range of extra funtions which do not need to be connected to the address or data bus will also be described later in the series.

Fig. 1. Basic interface system which uses three main i.cs to provide a range of functions. Further circuits which do not need to be connected to the address or data bus can be easily added on daughter boards.



Most people interested in or involved with microprocessors now realise that there is a gap between a microprocessor and its application. This is an inevitable result of the different aims and skills of the participants. The electronics engineer concentrates on the circuits, the architecture and the bus structure, but to the user these are almost irrelevant. A mechanical engineer may know the functions that the microprocessor should perform, but cannot connect the device to the hardware it must control. A strain-gauge speaks in millivolts, but a microprocessor listens in bytes and sends out binary data. For each problem there are solutions, although they are sometimes difficult and costly or involve additional work by the user.

When designing an interface the most important and difficult decisions to make are which microprocessors/microcomputers should the interface be directly compatible with, how many others could be adapted, and the bus structure and board format to meet these requirements. The board and bus structure are closely linked and may constrain the choice of microprocessor. For example, the S100 bus supports the 8080/Z80 family but boards have been designed for the 6800 family. However, the large board size and mixed power supplies make the S100 an unsuitable format. The high cost of Multibus and other industrial standards makes them inappropriate for educational use.

An economical and standard board is the Eurocard which is widely available and can be mounted in standard racks. The Acorn bus structure, which was chosen, enables the interface to be directly used with a low cost unit, a rack-mounted system, or the new BBC computer. However, with suitable interconnecting cables, the interface is equally compatible with the Aim 65, Apple and Pet.

The choice of functions is a compromise between the desirable and the economically feasible. The final design includes digital-to-analogue conversion, analogue-to-digital conversion with 16 input channels, 16 line i/o ports, 8 output drivers, 2 counter-timers, serial i/o and handshaking lines. As many microcomputers and microprocessor boards already offer a few functions, the corresponding i.cs can be omitted on the interface without affecting the remaining functions.

The parallel i/o ports, handshaking, counter-timer and serial i/o are all achieved with a 6522 v.i.a. (versatile interface adapter) whose programming can be as complex as the c.p.u. if all possible functions are considered. However, by starting with the parallel ports and gradually including the other options, programming can be kept manageable. The d-to-a converter is based on a ZN425 which, with extra gating, can be used as an a-to-d converter. To increase the d-to-a flexibility, the parallel ports can gate the d-to-a output to a series of sample-and-hold circuits. Although this technique slows down the response, it is satisfactory where, for example, multiple analogue outputs are required to drive electromechanical loads.

An important part of the interface is the ADC8017 16-channel successive approximation a-to-d converter. This device contains internal switching and gating which allows it to be connected directly to the address and data lines of most 8-bit microprocessors. Although the 8017 is not particularly fast, the ability to scan 16 analogue channels and load the data into memory with simple hardware and software makes it an ideal device for data-

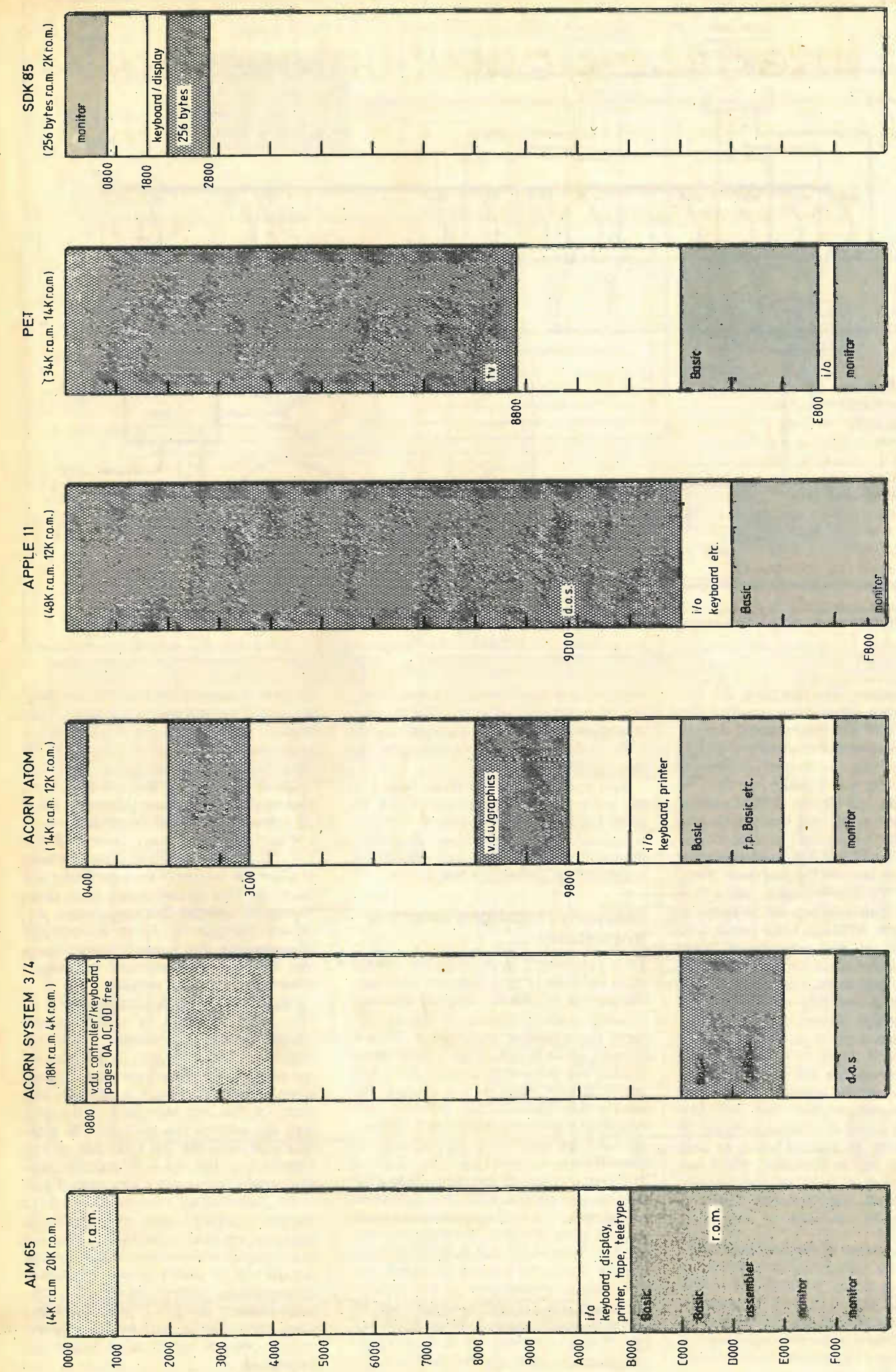


Fig. 2. Memory maps of the popular 6502 based microcomputers and an 8085 system.

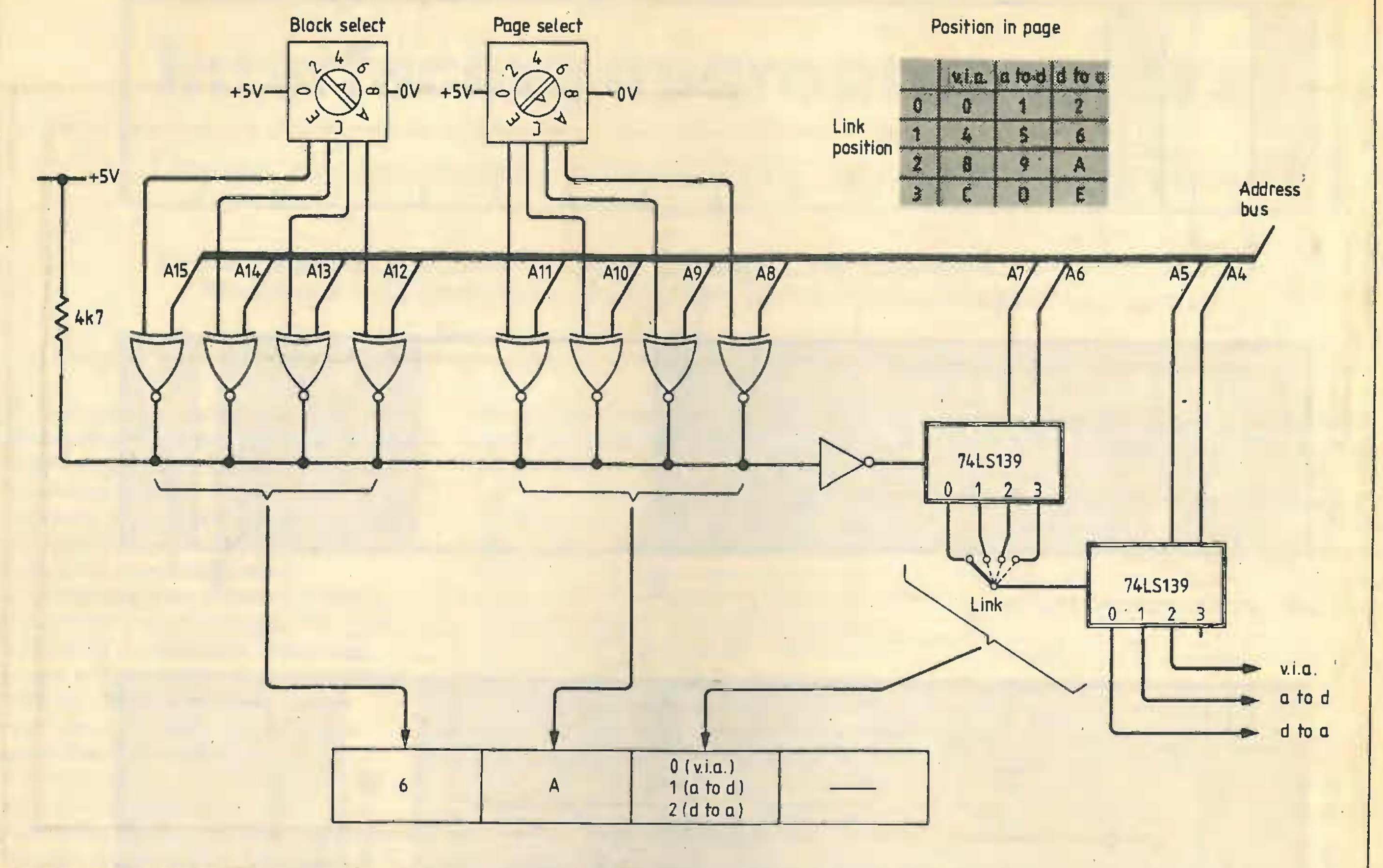


Fig. 3. Address decode circuit. Two binary-encoded switches allow independent selection of block and page.

logging systems. To complete the interface, output drivers are provided for switching l.e.ds and opto-coupled devices such as transistors and thyristors. C.m.o.s. logic is used for the output drivers to prevent loading port B which can then be used with external signals. If the l.e.ds on the board are used, and the interface is plugged into a suitable microprocessor system, no additional connections are required to run and test i/o programs. Also, by linking the d-to-a output and a-to-d input lines, both functions can be tested at the same time. Although these points may seem trivial to the experienced user, simple demonstrations have proved to be valuable for beginners. Extensions to this board could include opto-coupled switches, power control devices, signal conditioning, e.p.r.o.m. programming and microprocessor communication. These options, which will be covered later in the series, are not necessarily connected to the address or data bus and can therefore be added on a daughter board or via a socket on the original board. In each case a circuit will be described, which can plug into one or more of the popular microprocessors, together with details of how it can be modified to suit other systems.

The arrangement of the three main components in the basic interface is shown in Fig. 1. Each device is connected to the control and data bus, and is memory mapped by the address decode circuit. The v.i.a. and a-to-d converter each require sixteen memory locations for their internal registers and input channels respectively, and each location is selected using the four least-significant address lines A0 to A3. The single channel d-to-a converter requires only one location.

Sets of switches and l.e.ds are linked to the v.i.a. so that external loads can be driven and sensors monitored to allow i/o simulation while developing programs. The 6522 has several other capabilities which will be covered in a later article.

Memory maps and address allocation

Each component in a computer system must have one or more memory addresses assigned to it, and the designer allocates memory space according to the components importance or convenience. With a memory space of 64K bytes it might seem simple, but as a system expands more and more memory space is pre-empted. To ensure that this interface or other new board can be used with different systems, the memory maps must be compared to identify their unused areas. Fig. 2 shows the memory maps of several standard 6502 systems and a typical 8085 arrangement for comparison. All 6502 systems have r.o.m. at high order memory, although only the top few bytes which include locations used for the automatic start-up procedure are essential. It is convenient to use adjacent areas for r.o.m., though some gaps may be left for other functions as in the Apple and Pet. There is normally r.a.m. in zero page (the first 256 bytes) and memory access to this area requires fewer bytes and less time using the zero-page addressing mode of the 6502. It is sensible to use this facility for rapid access to data which is required repeatedly by various programs. Page one (location 256_{10} - 511_{10}) also contains r.a.m. because the 6502 uses this page as the stack for subroutine jumps and interrupts.

The Aim 65 and Acorn system 3/4 follow a similar format which allows the user to adapt the computer for a particular application. This flexibility has made them popular in colleges and universities. Although the Apple and Pet are also popular in engineering and scientific applications, they are ideal programmers' computers, filled to the brim with memory.

The address decode circuit shown in Fig. 3 is the result of several design changes to give the interface board the flexibility needed for operation with several systems. Two binary encoded switches allow independent selection of block and page, i.e. the first two digits of the four digit hex address, via exclusive-OR gates which monitor the top eight bits of the address bus. The 74LS139 decoders provide chip select to the d-to-a, a-to-d and v.i.a., allocating sixteen sequential memory locations to each, and modify the addresses via links by adding 40, 80 or C0 to the chosen base address. Therefore, the circuits can be added to any compatible microprocessor system that has an unallocated memory space of at least 64 consecutive bytes. Where memory space is not critical, parts of the decoding circuit can be omitted.

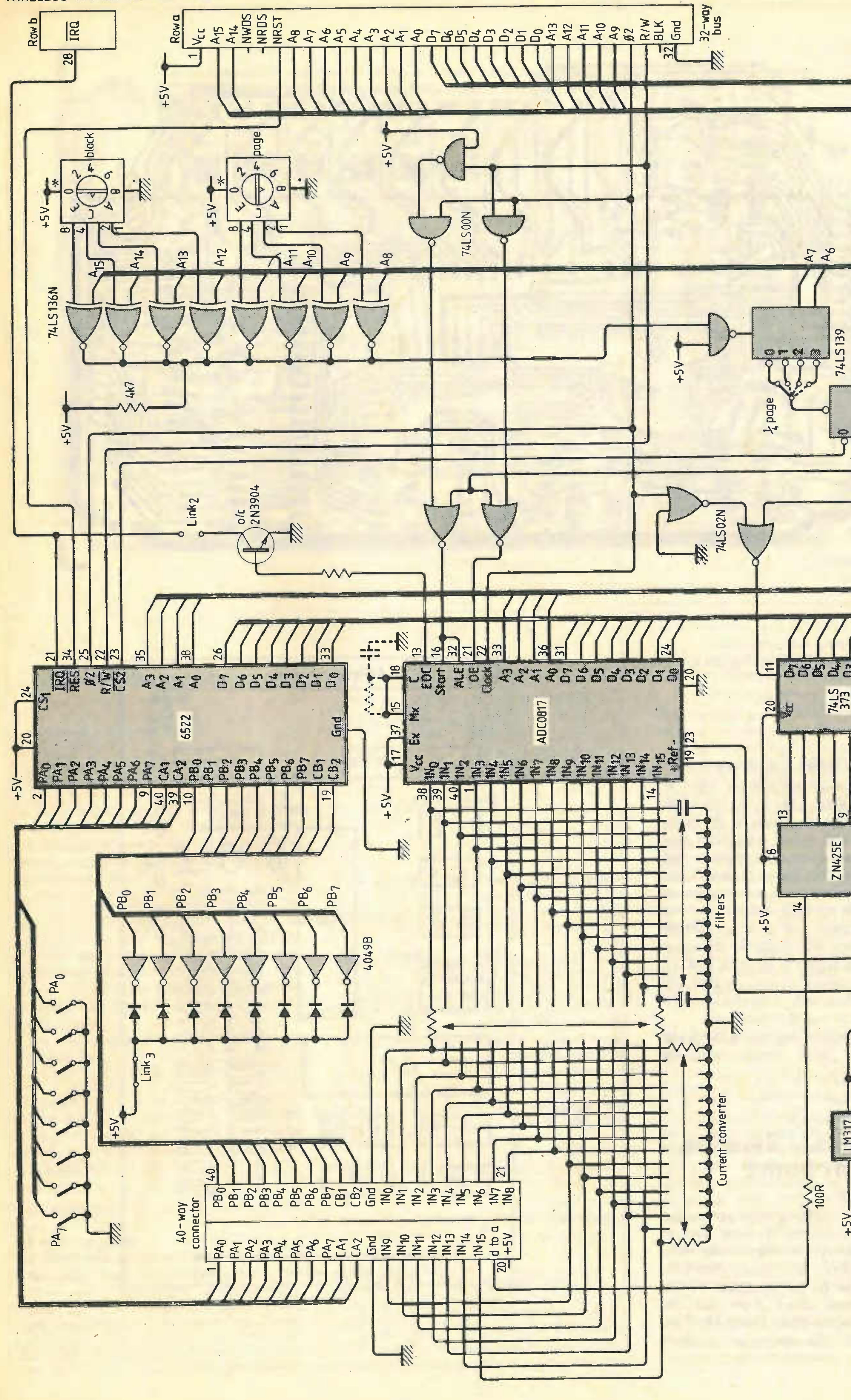


Fig. 4. Complete circuit for the main interface. I.c. sockets are recommended so that the board can be upgraded with little inconvenience.

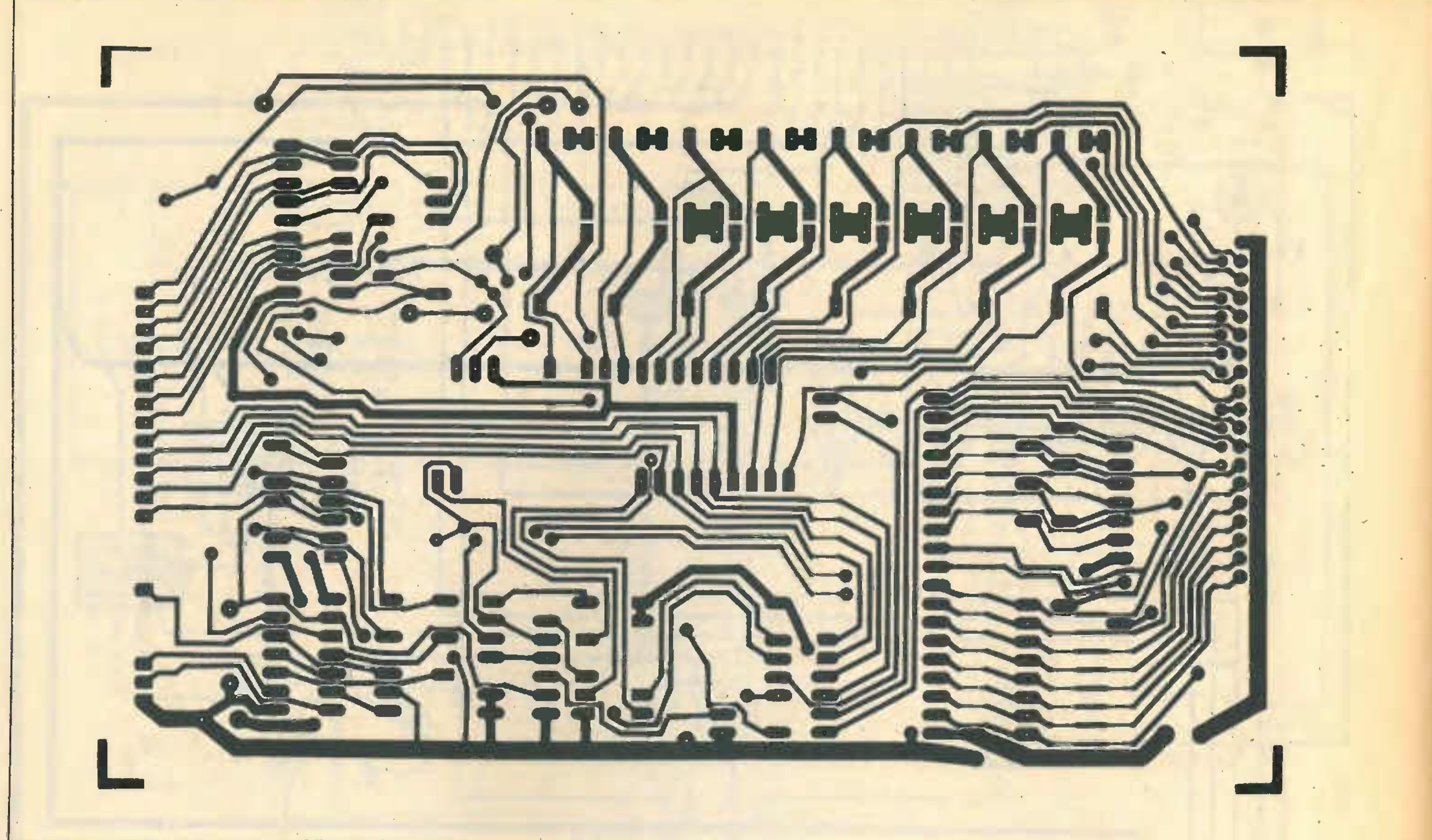
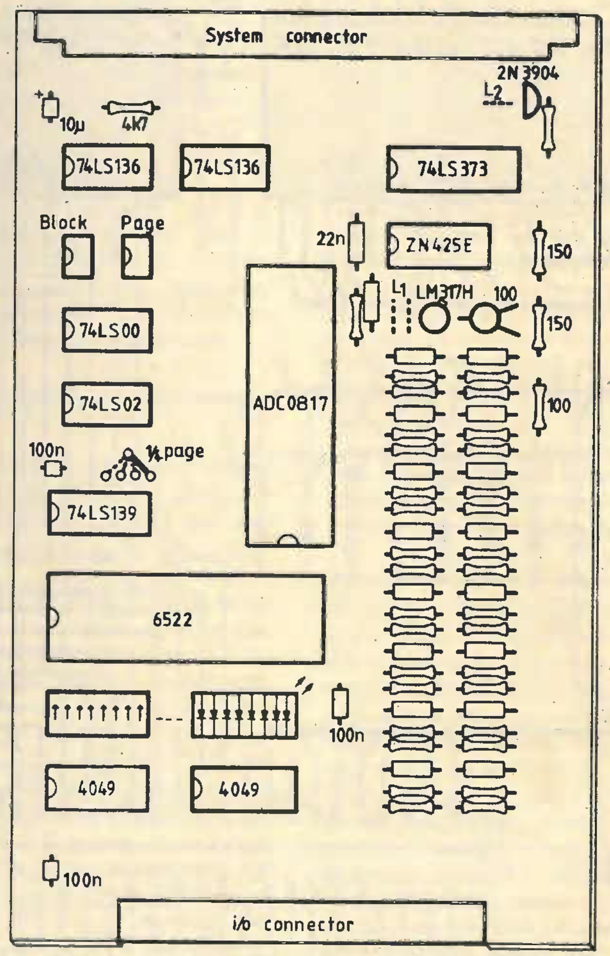


Fig. 5. Board layouts and component location diagram. Assembled boards and kits will be available from Control Universal, 11 Bush House, Bush Fair, Harlow, Essex. CM186NS.

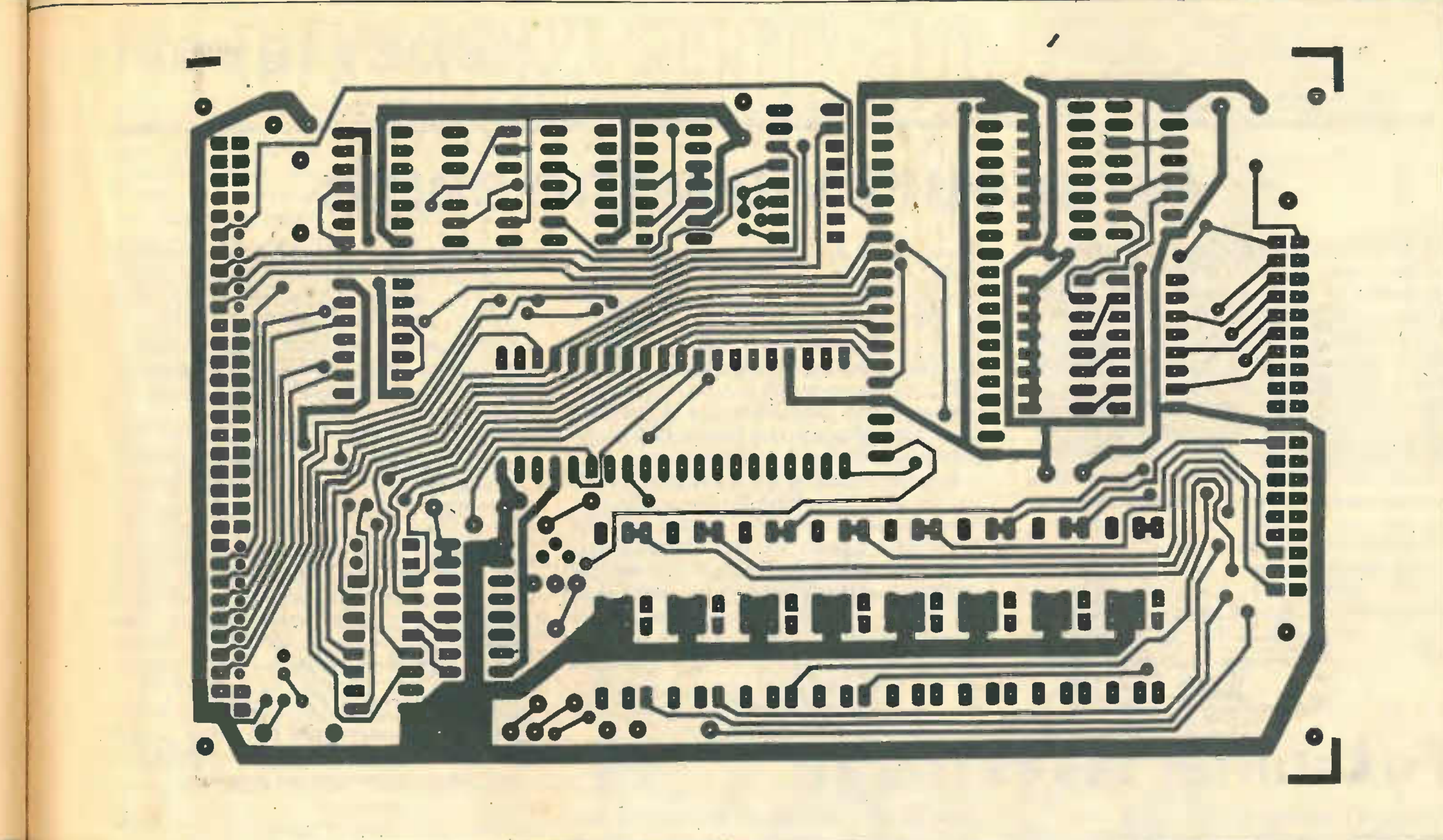
The complete interface is shown in Fig. 4. The decoding circuit on the right is wired to the address bus on the Eurocard connector. Additional gating is provided for the d-to-a and a-to-d converters and both devices have protective resistors and simple capacitive filtering to reduce noise. The a-to-d converter also has the option of a shunt resistor on each channel for use with current inputs. The reference for this device can be provided from the reference of the ZN425 or from the separate LM317 regulator. This can be trimmed to give, for example, 2.55V which is convenient for 10mV steps in the a-to-d conversion.

Part 2 describes the functional blocks in detail together with simple program examples.



Corrections – Satellite tracking by home computer

One or two errors in the second part of this article in the September '81 issue need to be pointed out: At the top of the third column on page 67, DC should read DE. In line 16 of the BURP program, N=CS SIN should read N=S SIN and the beginning of line 36 should read IF W>0 A=180 A + !. We apologise for these errors.



BOOKS

Atom Busiuess, by John Phipps.
110pp., paperback.
Phipps Associates, £6.95.
Nuclear physicists are not necessarily the prime audience for this little book, since it contains a set of programs for use with the Acorn microcomputer in business. Peripherals needed are a domestic television receiver and a cassette recorder: a printer is useful, but not essential for most of these programs.

Complexity varies from a simple listing to perform running addition to a program for determining the effects of various parameters on the length of a queue – probably this could be varied for use in other circumstances, such as the piling up of work in a machine shop. The reasons for each program are set out and operating instructions presented in simple terms prior to each listing.

The book is completely practical in that there is nothing on programming as an art – simply the programs, including one which will help to make a decision on whether to lease or buy equipment, a sales record, a nominal ledger and one for working out expenses. A cassette containing the programs is obtainable from Phipps Associates, 3 Down Avenue, Epsom, Surrey KT18 5HQ at £7.50, plus v.a.t.

Video/computers, by Charles J. Sippl and Fred Dahl.
246 pp., paperback.
Prentice-Hall International, £5.55.
Messrs Sippl and Dahl base this book on the premise that much of the domestic video, viewdata, audio and computing machinery currently considered as separate entities should, and inevitably will, be brought together to form what they call an integrated video computer. Such a collection of electronics would, as we are constantly told, bring about a social revolution in working habits, communication, banking, shopping . . . etc, but, although much of the technology already exists, there are problems still to solve.

Adopting an extremely methodical approach, the authors consider all the ingredients of such an integrated system, one by one, and try to form an opinion on the way they will develop if the i.v.c. is to come along. They view the concept from several angles – video, computing, data conversion and communications, explaining what is now in existence and what will have to be done in each sector to reach the goal of an integrated system. The conclusion is eventually drawn that the i.v.c. will be with us towards the end of the decade, given that legalities and vested interests can be surmounted, and that we will then have a "worldwide network of total communications".

Latest Developments in Sound Broadcasting, Ed: John Lovell.
77 pp., paperback.
IBA, London.
Local radio is the subject of the fourteenth in the IBA Technical Review series, and contains nine articles by IBA staff and consultants.

Studio design, installation and testing are covered in the first three sections, followed by a description of the contribution network, which enables local radio companies to send news items to IRN in London for national use. Technical features of the first i.l.r. transmitters are reviewed and there is a description of the solid-state Phase 2 stations, including a look at pulse-width modulated (Class D) power amplifiers.

A section on the Borehamwood m.f. aerial for the London area describes the design of an extremely complex system, which must not only cover its service area and avoid other areas serviced by other transmitters, but must do this at two frequencies simultaneously.

Surround sound is discussed, in theory and practice, and in the final section the head of Long-range Studies looks at the future of radio broadcasting. Technical Review 14 is available from IBA, Brompton Road, London SW3 IE7. No charge for single copies.

The ZX81 Pocket Book, by Trevor Toms.
136pp., paperback.
Phipps Associates, £4.95 (Cassette £5.00).
Owners of the Sinclair ZX81 microcomputer who have worked through the handbook supplied may feel the need of additional tuition in the techniques of programming. If so, this book should be of assistance.

To illustrate the sections of instruction (on efficient programming, using machine code, using data files, etc.) a number of programs, mainly games, are presented and explained. The book is not simply a rewritten version of the ZX80 Pocket Book, but is said to be almost completely new, since the ZX81 is quite different to the ZX80. Much of the content can be used by ZX81 owners who do not possess the 16K r.a.m. A cassette, containing the programs described in the book, is also available.

Phipps Associates are at 3 Downs Avenue, Epsom, Surrey KT18 5HQ.

NEWS OF THE MONTH

Amateurs view the earth

By the time you read this, Britain's first amateur satellite should be in orbit 330 miles up with an inclination of 97.5° and a 98 minute period. Organized by the University of Surrey department of electronic engineering, the launch was to accompany a NASA solar Mesosphere Explorer spacecraft, set for September 12 at the time of going to press, and is especially interesting in Vosat's ability to provide earth pictures for display on a domestic tv receiver. The satellite will not be fully operational immediately, though the v.h.f. and v.h.f. beacons should be in use for telemetry; picture transmission should start in a few weeks time.

the pictures are taken by a charge-coupled imaging array made by GEC's Hirst Research Centre and provides land and sea image data for digital transmission over the v.h.f. beacon using a.f.s.k. at 1200 bit/s, line synchronous. Image format is 256 by 256 element, with 16 grey levels, stored in an on-board 0.25 megabit memory. The camera is organized to cover 500 by 500km areas of the earth's surface, with a resolution of about 2km, optimized to enhance land features and land/sea boundaries. All being well, this experiment will also display processed telementry and experiment data in graphical form. The University team promise circuitry for interfacing with a tv set later in the year.

Beacons transmit on 145.825MHz with 450W and 435.025MHz at 800mW. Further beacons, phase-referenced on 7.050, 14.002, 21.002 and 28.002MHz support ionospheric experiments whilst two microwave beacons on 2.401 and 10.470GHz are for studying s.h.f. propagation – intended to help develop inexpensive microwave satellite ground stations. The two data beacons, v.h.f. and u.h.f., should provide a reliable satellite-to-ground link for reception by simple amateur stations: an unmodified narrow-band f.m. receiver together with a fixed pair of crossed dipoles should do for most passes. Besides normal telemetry data, computer outputs, synthesized speech for school demonstrations, and earth imaging data are also information sources for the beacons.

The computer is based around the RCA CDP1802 microprocessor, enables telemetry surveillance, command and status management, experiment data storage and processing, dissemination of orbital data and operating schedules, and closed-loop attitude control. It has direct high-speed data links with the magnetometer and radiation experiments, and access to the earth-imaging memory area for image processing. Program is resident in dynamic r.a.m. loaded from the ground *via* the telecommand link and which can be modified, or replaced during flight, from ground. Commands from ground stations take precedence in cases where commands emanate simultaneously from both the microcomputer and ground.

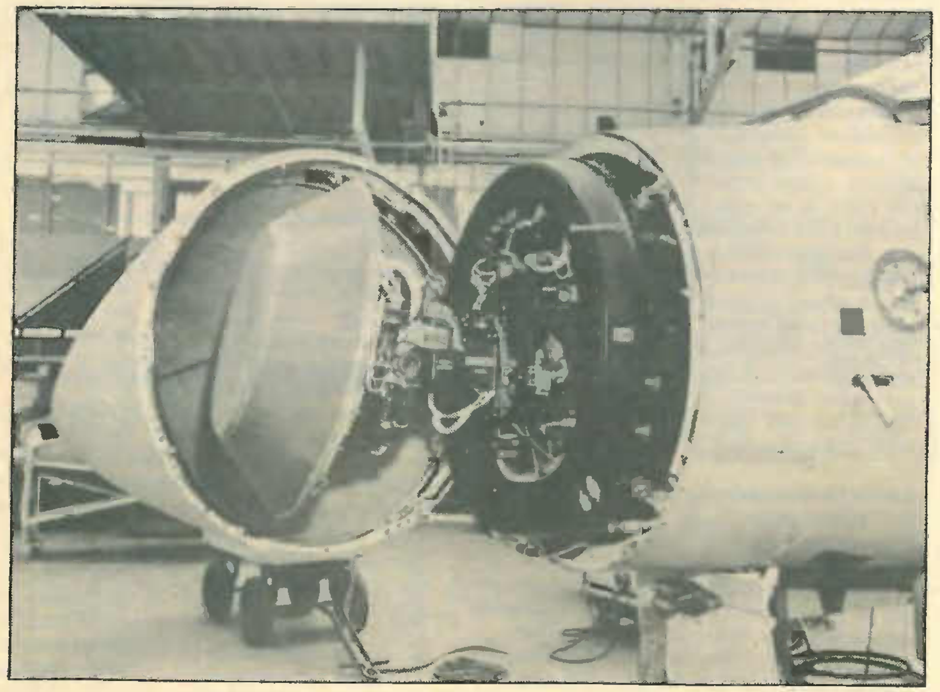
Foxhunter takes the air

Marconi's Foxhunter, the multi-mode radar for the Panavia Tornadó 2, made its first flight in the Tornado air-defence variant at Warton on June 17. Since that date, the radar has operated in all its major modes.

Trials have been progressing at Bedford in a modified Buccaneer since shortly after the initial MoD production contract was placed in late 1979. First deliveries of the equipment are due to start in 1982 for RAF service two years later: a run of 200-250 radars is the likely figure, assuming that defence cuts do not affect the number of Tornado ADVs to be built. Further, modified radars may be exported later.

Foxhunter operates in several ways. It uses the pulse-Doppler technique to avoid static clutter and will search an area, track several aircraft simultaneously while continuing to search, can display a map of the ground and 'locks on' to a target – the Skyflash missiles carried by the Tornado use reflected radiation from the target for guidance. In the photograph, the truncated cone is the two-reflector aerial. The signal is launched through the rear dish, is reflected from the forward element back to the rear one and collimated into the forward beam, the polarized radiation firing through the slatted forward reflector. A glass-fibre cover forms a skirt between the two.

Foxhunter is extremely compact, compared with existing radars, but even so occupies nearly all the cross-sectional area of the Tornado nose forward of the cockpit. The black object protruding from the nose is the 'brim' of a 'top-hat', around the crown of which are arranged all the signal-processing modules.



Cascade noise reduction

Integrated circuits for the new Dolby noise reduction system are being sent to licensees this month. Made by Hitachi, but designed by Pioneer, the i.cs are off the mark much sooner than expected. Dolby originally said that dedicated i.cs would not be available until some time next year, but it is undoubtedly market pressure that has accelerated development. Dolby say that better sound sources, higher listening levels and a "genuine market desire" for more noise reduction prompted the new system, which reduces noise by up to 20dB. But as competing n.r. systems abound now and threaten Dolby's growth in this area, Dolby clearly needed to come up with something new, having found that the B circuit couldn't be pushed far enough without adverse effects, both errors in frequency response and overshoot.

So they adopted the approach of cascading two sliding-band circuits, each working at different levels (rather than different frequencies). But although a good amount of noise reduction is obtained in this way, as many enthusiasts have discovered, it is not altogether satisfactory by itself – what is described as a mid-frequency "mud" still remains. Dolby found that to rebalance the amounts of high, mid and low-frequency reduction the turnover frequency, 1.5kHz in the B-system, was best set two octaves lower at 375Hz for the cascade-circuit. A slightly higher compression ratio of 2.2 instead of 1.9 is used and further circuitry is needed to keep side effects down: h.f. de-emphasis at the encoder input to reduce mistracking, and a network in the low-level stage to prevent tape saturation at high signal levels. With this additional circuitry the C system, as it is called, is claimed to be "at least as free of side effects as the B system". Altogether Dolby say the circuitry costs 2½ times as much as the B system but that falling i.c. prices partly offsets this.

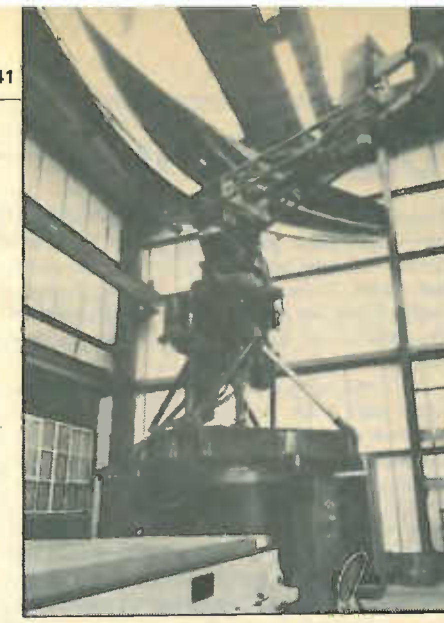
Naval radar

In an effort to help warships survive when attacked by an extremely varied, agile and 'intelligent' array of nastiness, Plessey has developed a new S-band radar, the AWS-5.

Many constraints are imposed on such a radar. High-flying aircraft must be detected at the same time as sea-skimming missiles; small, fast attack craft, which carry sufficient weaponry to embarrass a cruiser, may press their attentions at precisely the same time as a vertically diving missile aims for the funnel.

AWS-5 comes in several forms, for different kinds of warship, but the most comprehensive variety offers a dual-beam aerial on a stabilized platform. The differing sizes and direction of approach of attacking objects are catered for by aerial design and the use of pulse compression. Two aerials are used – a main parabolic reflector for low-angle detection and a smaller type mounted over the main one for high angles. No height-finding facility is incorporated, both beams being narrow horizontally, but some discrimination between high and low objects is obtained by comparing the returns from the two search patterns. The two are multiplexed and can be viewed separately. Coded-pulse compression confers long range, high resolution and a low peak power requirement: the technique is one variety of the 'chirp' process, in which the radiation is frequency-modulated during the pulse; here, phase changes at each transition of a pseudo-random code. Peak power can be kept low, which means that the radar is less easy for an attacker to detect, and resolution can be changed by simply modifying the code. The coding is difficult to analyse and confuse.

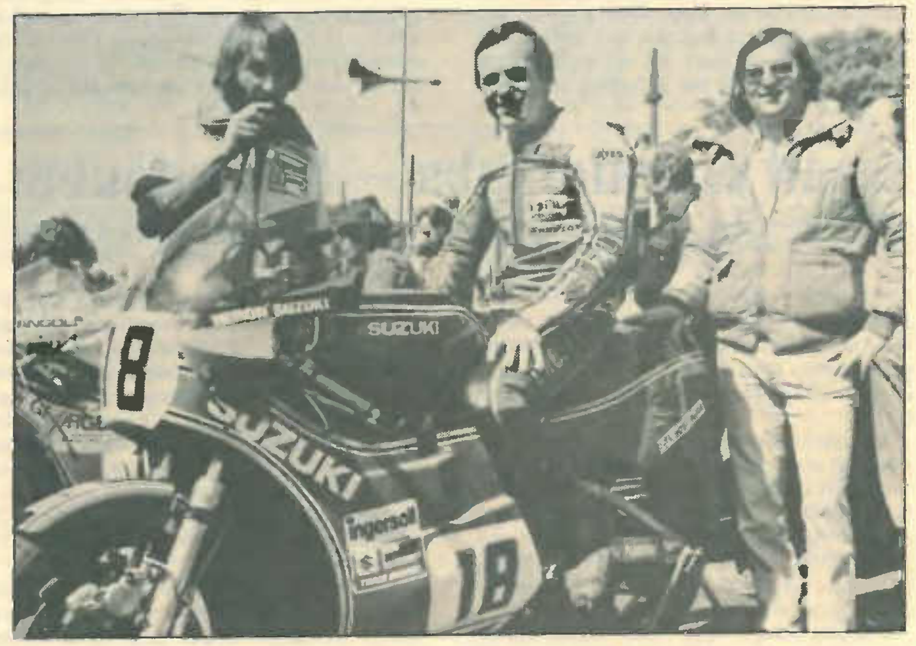
To avoid jamming, the transmitter uses a travelling-wave tube, which can be made frequency-agile, while sea clutter is reduced by the use of a developed version of Plessey's adaptive moving target indication technique. The search area is considered as a great number of 'cells', which are defined by the transmitted pulse length in the radial (distance from transmitter) dimension and bearing gates in the tangential direction, the average signal level in each cell being digitized and stored. Variations in this level on succeeding scans are assumed to mean that an object has entered or left the cell and the return is displayed: if nothing has changed, it is blanked. The process is complicated by the motion of the ship, but the principle remains. Two foreign navies have ordered the radar, and the Royal Navy is said to "interested".



Plessey's AWS-5 radar being vibration tested.

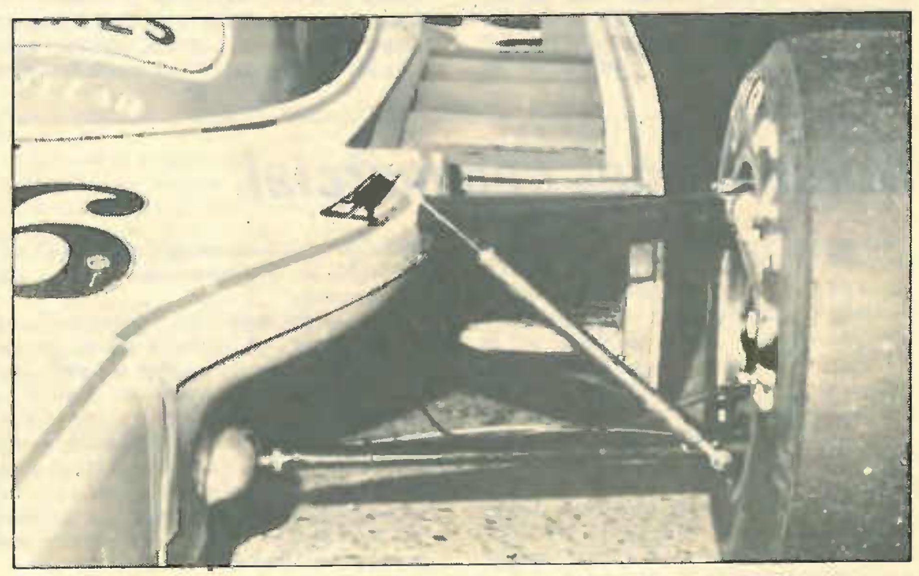
High-speed car radio

Richard Noble will make his attempt on the world land-speed record in October with Thrust 2, which is powered by a Rolls-Royce Avon jet engine and which, Noble hopes, will move at around 700 m.p.h. For the attempt, a number of communication channels are neede to link support crews at the ends of the run, a high-speed fire tender, timekeepers, an observation aircraft and the driver himself. Racal-Messenger have planned a multiway u.h.f. link which will allow several conversations to be carried on simultaneously, being overriden by transmissions from the car, which will be received automatically by all units. The aerial is of the 'blade' type normally used on high-speed aircraft.



Although of not such a high speed as the world land-speed record, racing motorcyclists have similar communications problems which have been solved by Pye Telecom whose mobile radios and 'pocketrones' were used at the Formula 1 and Classic T.T. Races on the Isle of Man. The Driver of the winning Suzuki machine, shown her, was Graeme Crosby and he and his team used the radios for a two-way flow of race information and tactical decisions. A further Pye Telecom radio system was used by the race marshalls, who had mobile radio telephones mounted on their motorcycles. This enables them to react quickly in the event of accidents, summoning help and warning the organisers.

Yet another application of electronics at high speed is in the Grand Prix racing car which uses the 'ground effect'. The effect is achieved by designing cavities in the underside of the car which act as venturis. When air passes under the venturis, the car is physically sucked down onto the track, offering greater stability and consequently higher speeds. unfortunately the gap between the venturi and the surface of the road is critical to the level of suction obtained, and it is necessary to establish which type of suspension system will give the best performance to take advantage of the effect in a car which is subjected to a constantly changing pitch through rapid acceleration and hard braking. The French Talbot-Ligier Grand Prix team use a Sangamo UDC 100 longstroke displacement transducer, linked to on-board monitoring equipment, to record the changes in suspension geometry as the car is driven at racing speeds.

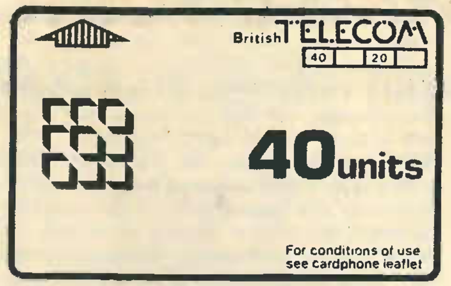


Cardphones – towards a cashless society?

Major London railway and tube stations are to have Cardphones installed in a trial for what British Telecom call "a step towards the cashless society." But before you get too excited, cashless does not mean broke – Cardphones eat Phonecards and that means purchasing either forty or two-hundred 5p units in advance.

A Phonecard is a piece of plastic similar to a credit card with 'holographically memorized' call units printed on it. As a unit is used up, so that unit is erased, but a warning is given twenty seconds before the last unit runs out to give you time to fumble for a new card, say goodbye or utter an expletive (Cardphones don't accept coins). Throughout a call, though, a readout tells the caller how many units are left without making a call by simply inserting the card in the slot – completely free of charge!

Long, consecutive and overseas calls to countries now on the direct dialling system are not hampered by the insertion (and availability) of coins, but if you make one call directly after another you still lose any remaining parts of a



unit, as is the case with present domestic and public telephones.

Each card has one or two tracks, depending on its price, on which the call units and other information are printed. The extra information tells one of the two microprocessors used in the system whether or not the card is acceptable, i.e., whether British Telecom intended the card for making public calls and whether the card is indeed issued by them and not by any other authorities with similar systems. Belgium and Austria already have such systems and the French are making trials.

The call units are read by a configuration of infrared detectors that pick up reflected light from the coded patterns on the card. As a unit is used up it is erased thermally.

A second microprocessor looks after the normal routine of the dialling system and allows normally free calls such as '999' and directory enquiries to be made without the use of a card. The system was developed in Switzerland by Sodeco, a subsidiary of the Landis and Gyr Group who are now supplying the apparatus to British Telecom.

Initially around 120 Cardphones are being installed in London and another 80 or so will appear in Birmingham, Glasgow and Manchester. Phonecards will be available from post offices and some retail outlets including railway station bookstalls and fare counters.

Small wavelengths – large doubts

Over the years there has been much mild controversy in most western countries over the maximum safe level of microwave radiation. Recent news that an American body for workers' compensation defined the cause of death of a radio technician as chronic exposure to microwave radiation will hopefully invoke a closer look into the long term effects of these rays.

As in most Western countries, the maximum safe level in Britain for continuous exposure to microwave radiation is defined as $10mW/cm^2$, a figure 1,000 times higher than that adopted by the Russians at $0.01mW/cm^2$. Taking into account that the conditions associated with these figures are not exactly the same, the difference is still enormous.

Our maximum level is based on that determined by the Americans nearly 20 years ago. According to an International Electrotechnical Commission report† from 1979, the very large discrepancies between standards are due to differences in approach, namely that the USSR standards are based on the possibility of any noticeable biological effect, in contrast to thermal injury, and most western countries take the view that minor reversible effects are not necessarily hazardous to man. Also, say the IEC, the Russians have used very much larger safety factors than most other countries in defining their limit. As there is, even now, much doubt as to the long-term effects of microwave radiation, large safety factors seem a sensible precaution.

There are stumbling blocks in researching the effects of microwaves on humans, though, that could account for the uncertainty as to the 'safe level' of radiation. One problem appears to be the lack of a suitable guinea-pig; the position, size, density and material of every part of the body can be critical. Also, measurements made in the 'near field', i.e., the complex field close to the antenna which contains electric and magnetic components additional to those of the main propagation field, are difficult to interpret in terms of potential hazards.

According to the National Radiological Protection Board, frequency dependent limits have been proposed in the USA to bring down the maximum exposure figure to $1mW/cm^2$ at frequencies where the radiation has the greatest effect on the human body. But only time will tell whether these new limits, if accepted, are safe, or indeed whether the old limits were on the safe side anyway.

● Research into electromagnetic coupling between a 'thin-wire' antenna and a biological body was reported in the IEEE's publication dealing with microwaves* last November. The research was carried out to assess the potential hazard of portable transmitters, especially those for c.b., and some interesting conclusions were drawn. According to the article, the power absorbed by a human adult standing 20cm away from an antenna with a 30W input at 90MHz or 140W at 27MHz is the same as would be absorbed from exposure to $10mW/cm^2$ plane-wave irradiation. An average human adult standing 20cm away from a quarter-wave antenna operating at 20MHz will absorb about 8.5% of the antenna input power, but at 90MHz, over 50% of the input power will be absorbed as the average height of an adult is about a resonant length at this frequency. Electromagnetic coupling is increased considerably when the body is in contact with the ground, as opposed to being isolated, and the body may act as a director element when placed close to the antenna.

† IEC report number 657 'Non-ionizing radiation hazards in the frequency range from 10MHz to 300,000MHz.'

* IEEE transactions on Microwave Theory and Techniques, Volume MTT-28, Number 11 (part 1), 'Electromagnetic coupling between a thin-wire antenna and a neighbouring biological body', by K. Karimullah, K.-M. Chen and D. P. Nyquist.

Satellite on a string

Sounding rockets remain in the air for only a few minutes; low-altitude, non-propulsive satellites can gather data for a few hours before their orbits decay. A possible solution could be a low cost satellite tethered by a long (very long – up to 60 miles) super-strong cord to the NASA space shuttle.

Engineers from the Marshall Centre have been carrying out feasibility studies with space scientists from Italy for what could be the first US/Italian cooperative space project. The Italians could build the satellite and the Americans would supply the equipment necessary to handle it. The satellite, attached to the shuttle by the tether line, would be trolled through the Earth's upper atmosphere in a very low orbit, perhaps only 80 miles above the Earth, for an extended period. It would be used to gather data on the atmosphere, the magnetosphere and gravity. The system is likely to become operational by the mid 1980s.

Teletext goes commercial

You can now buy a page of text on Oracle, the six-year old ITCA teletext service, for a weekly rate of £400. That page would then be available to the 180,000 teletext-equipped sets throughout the UK, though the service will be available on a regional basis. First Scottish Television start their own input unit this autumn followed by Channel TV some time later, culminating in a fully regional service by 1984/5 that can offer local news as well as advertising – by which time the number of teletext sets is predicted to be 3,000,000. Oracle Television Ltd, formed last year by the ITV companies, will promote itself mainly through "filler" tv advertisements to the tune of £2 million-worth of tv "commercials" over a six-month period.

At about the same time – probably during the Department of Industry's teletext promotion in October – the number of Oracle lines will be increased from two to four. This allows the access time to be reduced typically from 45 to 20 seconds, which otherwise would become unacceptably long as the volume of text pages is increased. One feature of this new service is that certain advertising not permitted on television will be allowed on Oracle – from football pool promoters and bookmakers in particular.

Integrated circuit design

Understanding the nature of black boxes may make a significant contribution to circuit performance

by J. L. Linsley Hood

The starting point for this series of articles is the i.c. that perhaps has done most to encourage the application of op-amps as a simple, cost-effective solution to circuit problems.

Historically, the 741 device was introduced by Fairchild at the end of the 1960s, along with several other second-generation i.cs from rival manufacturers, as an internally-compensated improvement upon Bob Widlar's classic 709. In the Fairchild µA741, most of the minor operational problems of the 709 were reduced to an extent that they were no longer inconvenient in use, and the 741 then became a nearly ideal building block for low frequency applications.

Understandably, many of the internal circuit facilities such as output short-circuit protection were similar to, and inspired by the same requirements as, those being introduced in the discrete component audio amplifier designs current at the time. However, the standardization on the use of separate + and – supply lines, together with nearly identical inverting and non-inverting inputs and the use of circuitry which allowed a high degree of supply line isolation, presaged developments which the discrete component amplifier designs were not to adopt at all widely for many years.

I have shown the circuit, in very simplified form, in Fig.1, with the necessary apology that a simplification of this type inevitably takes liberties with the actual design, simply because a more accurate representation of its form would hardly be a simplification at all. For example, although I have shown the input transistors as a p-n-p long-tailed pair, because this is effectively how they operate, they are in reality a rather more complex arrangement to allow the use of a pair of n-p-n devices in the input stage – in a modified cascode connection – of a form which is identical to, and perhaps inspired by, a circuit proposed a couple of years earlier*.

The reason for this rearrangement, shown in Fig.2, is that it is very difficult in conventional bipolar technology to fabricate p-n-p transistors which have any respectable current gain h_{FE}, except in the case where the collector is electrically connected to the p-type substrate such as in the output device. Other p-n-p devices must be of the lateral type, as shown in Fig.3. These are robust, but generally have h_{FE} figures only in the range 5 to 25, depending on the skill of the manufacturer in defining small gaps in his diffusion masking.

In addition, great use is made in i.c. manufacture of current mirrors of one kind or another. These are circuit arrangements in which the output limb looks and behaves like a conventional high-impedance constant-current source, but with an output current controlled by an input current fed into its other limb from some external source. The output current then mimics or mirrors the input current. I have shown three of the more commonly used types in Fig.4†. The popularity of this type of circuit element in i.c. manufacture arises from the fact that resistors and capacitors are inconvenient to construct in any large values, whereas transistors and diodes are easy. Moreover, if a current mirror is used as a load, an improvement in gain can be won.

This allows, for example, better operation of an input long-tailed pair wherein the loss of gain due to the normal halving of the g_m of the input devices is recovered together with an improved equivalence of gain between the two inputs of the long-tailed pair.

The operation of this type of circuit, taking for example the simplest arrangement of Fig.4a, hinges on the fact that if a transistor is forward biased so that it passes a certain collector current, the voltage across its base-emitter junction will then be precisely that which is required to cause

*Linsley Hood, J. L., *Electronic Engineering*, March 1967. (Letters).

†Davidse., J., Integration of Analogue Electronic Circuits. Academic Press.

Why look inside?

There are three ways in which a better understanding of the internal design of linear and quasi-linear integrated circuits can help the engineer: more satisfactory performance of circuits following from a greater appreciation of their strengths and limitations; possible use of accessible internal circuitry in unusual applications (a rich hunting ground in some of the more advanced units); and as an encyclopaedia of ingenious circuit design techniques, worked out by some of the most competent and resourceful design engineers.

Choice of the 741 as the starting point for this series stems mainly from a feeling that it was this i.c. more than any other which was responsible for the reconciliation of linear circuit engineers to the idea that most of the circuit functions they had implemented using discrete components could be done by integrated circuits, with improvements in simplicity and cost effectiveness.

In spite of all its limitations, in gain and phase-shift, slew rate and input bias current demand, there are many applications in which the 741 gives excellent service. This applies in audio and medium frequency applications so long as the associated circuitry is designed with an eye to its strengths and limitations. In addition there are a vast number of other circuit usages in which the very high d.c. and l.f. gain of this i.c., coupled with its good rejection of supply line voltage fluctuations and its ability to operate with input d.c. levels almost anywhere between the limits imposed by its supply voltage lines, make the life of the linear circuit engineer much simpler than it was some 15 years ago.

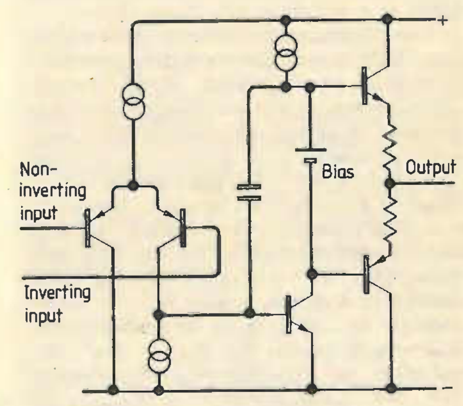
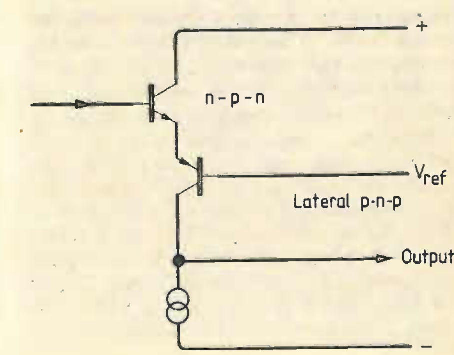


Fig. 1. Simplified input circuit operates as p-n-p long-tailed pair but in reality input devices are n.p.n.

Fig.2. Because of difficulty in fabricating high-gain p-n-p transistors input arrangement uses n-p-n types in modified cascode circuit.



an identical transistor (such as one diffused at the same time on the same chip and having the same junction area) to pass the same current. This is not strictly true in practice because the input current will be greater by two lots of base current. However, if this was important, the mask used in the diffusion process could cause Tr_2 to be slightly larger than Tr_1. The circuits of Fig.4b and 4c minimize this error. My own shorthand symbol for the current mirror configuration is shown beneath, and I have used this in subsequent drawings.

In the full circuit of the input stage, shown in slightly simplified form in Fig.5, a three-transistor current mirror of the type shown in Fig.4b is used as the load for the input long-tailed pair, and an ingenious combination of two simpler (4a type) current mirrors, transistors 8 and 9 and 10 and 11, is used to stabilize the operating currents of the input devices. The way this works is by means of a d.c. negative feedback loop. If the total current of Tr_1 and Tr_2, which should not contain any signal components, tends to increase, then the current output of the mirror Tr_8, Tr_9 will also try to increase. However, this is effectively fed from a constant-current source (the output of the current mirror formed by Tr_{10} and Tr_{11}) so the only thing which can happen is for the base voltage on the p-n-p transistors Tr_3 and Tr_4 to become more positive, which reduces the throughput current of the input stage because it effectively reduces the forward bias on the input transistors at the same time.

The interaction of these current mirrors also operates to minimize the magnitude of any unbalance currents in the input stage, which improves its symmetry, while simultaneously acting to lessen the extent of any breakthrough of signal components from the supply lines.

The second class-A amplifier stage and output stage, as shown in Fig.1, is of conventional form – the traditional high-gain small-signal amplifier followed by unity-gain power output stage, as spelled out so many years ago in this journal by Tobey and Dinsdale. High-frequency loop stabilization is achieved by the simple and effective expedient so common in early "hi-fi" amplifier circuits of a capacitor between collector and base, as shown. This leads to a few avoidable limitations which are discussed later.

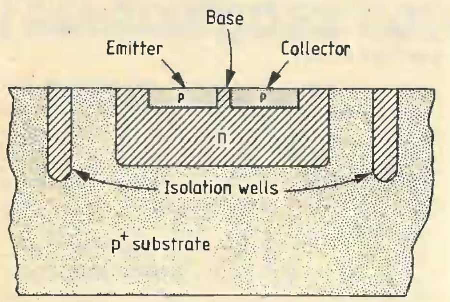


Fig.3. Lateral type of p-n-p transistor, though robust, has low value of h_{FE}, generally from 5 to 25.

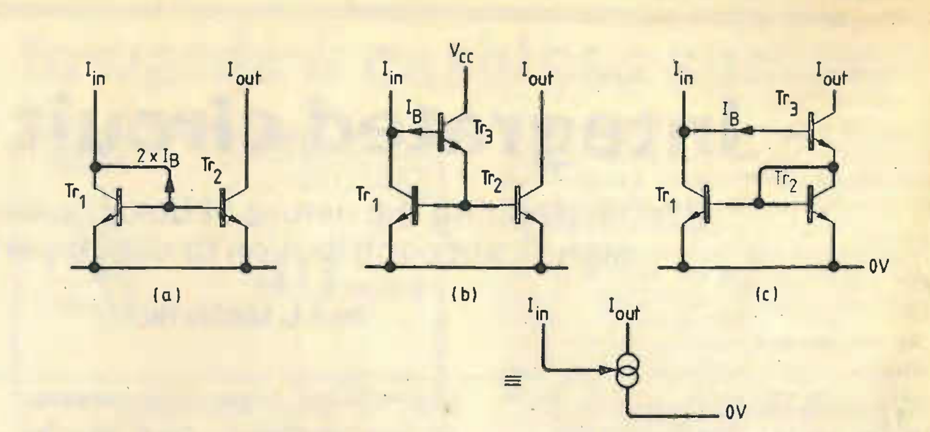


Fig. 4. Output current mimics input current in these current mirror variations, all much more easily integrated than resistors and capacitors.

As shown, the output stage would have no protection against damage due to output short circuits. This is accomplished by the use of a pair of n-p-n transistors (the preferred type), as shown in the more complete diagram of Fig.6, one of which is connected across the emitter resistor of Tr_{14}, and will take the current from the input to this if the voltage drop across this resistor exceeds its own base turn-on voltage, and the other (Tr_{22}) which acts in the same way in respect of the Darlington-pair class A amplifier stage Tr_{16}, Tr_{17}. The output stage forward bias is provided conventionally by an "amplified diode" Tr_{21} to give a quiescent current in class AB operation of about 1.5mA.

The final circuit of the complete i.c. is shown in Fig.7. I have actually shown that used by National Semiconductor, but all of the commercial 741s use a closely similar structure. In this, the only item not covered so far is the provision for offsetting inadvertent d.c. error at the output. This is done by putting a pair of resistors in the emitter leads of the current mirror used as the load for the input stage. If one or other of these is reduced relative to that in the other limb the current in that limb for balance will need to be greater; which calls for a change in input potential for that input device to maintain *status quo*. As this will not happen normally, the result of the adjustment will be to provide an output voltage shift equivalent to the stage gain multiplied by the required input offset. This provides a convenient means for obtaining a small shift in the output d.c. level, with minimal interference in the performance of the i.c. as a whole.

Performance

The d.c. and low-frequency voltage gain given by this circuit is very high – in excess of 50,000, with typical values of the order of 200,000. However, the presence of the h.f. stabilizing capacitor has a massive effect on the a.c. performance at frequencies higher than a few hertz, with the gain decreasing with frequency beyond some 5 to 10Hz at a rate of 6dB/octave.

A typical gain and phase-shift graph is shown in Fig.8.

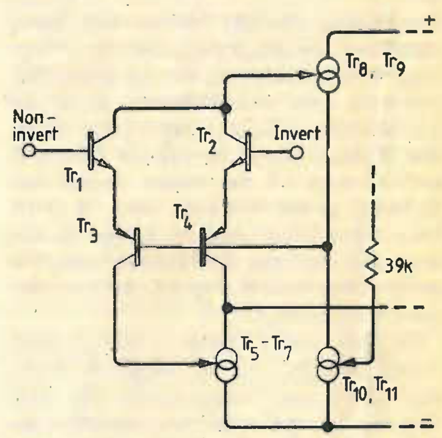


Fig.5. As well as a current mirror for the tail, type (b) in Fig.4, two (a)-types stabilize operating currents of the input transistors by d.c. negative feedback onto the base of Tr3, Tr4.

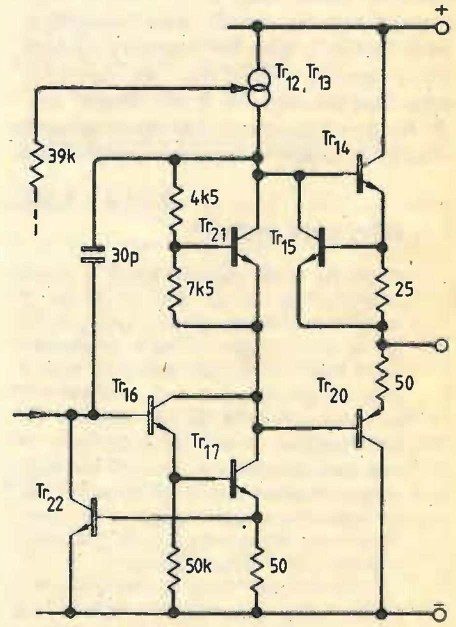


Fig.6. To provide short-circuit protection, Tr15 passes current from the input if the emitter resistor drop exceeds base turn-on voltage. Tr22 acts in a similar way for the Darlington pair.

An examination of this shows two important features. There is a significant additional phase error beyond 300kHz, which implies the presence of one or more

phase-lag inducing components within the i.c. having a pole at some few MHz. This is the reason why the unity-gain point for adequate unity-gain stability in a feedback configuration cannot be made much higher than 1MHz. And following upon this, the available open-loop gain at the upper end of the audio band, say 20kHz, is only of the order of 50.

Unless, therefore, the gain requirements of an audio amplifier stage using a 741 are kept deliberately low, neither the amplitude response and phase linearity, nor the harmonic distortion characteristics of the amplifier stage, are likely to be satisfactory in the context of contemporary expectations for "hi-fi" equipment. Fortunately there are now third-generation operational amplifiers, such as the Texas Instruments TL071 series, which offer substantial improvements over the performance of the 741-type i.c. in those regions which are of importance to the audio engineer, and I propose to examine this i.c. later in this series.

The other features inherent in the design of the 741 which must be borne in mind in its use if results are to be satisfactory are those which concern the input long-tailed pair of bipolar transistors and the effect of the h.f. compensation capacitor on the transient performance. Taking the first of these, the design of the input stage leads to a combined collector current for the long-tailed pair of around 25 microamperes. Assuming a current gain of 100 for the input devices, the necessary forward bias current for 25°C operation of the circuit will be 0.1 μA for each input, and this current must be supplied through any resistive circuit components in the input paths. While an output d.c. offset can be minimized by making sure that the total resistance value in each input circuit through which these bias currents must flow is the same (those components through which currents do not flow are unimportant in this calculation), it must be remembered that these currents increase significantly with temperature, and that the internal matching of the input devices may not hold over this range. For this reason, the total d.c. gain of the circuit and the amount of output d.c. offset which is tolerable must be considered when its circuit environment is being formulated, along with the temperature range over which it is to operate.

The second limitation of this i.c., that due to the nature of the internal h.f. compensation, is a rigid upper limit on the voltage slew rate which can be achieved at the output, around 0.5V/μs. If a composite signal is applied to the input which contains components calling for a greater rate of change in putput voltage than this, the total composite signal will be lost while the output moves from one instantaneous d.c. level to another, at the maximum rate possible. This self-evident fact applies to all amplifiers which are slew-rate limited, including some in the "hi-fi" field. It is, I think, a sad commentary on the state of our art that a fact which is so simple to comprehend and can be stated so simply, can be used as the basis for a whole series

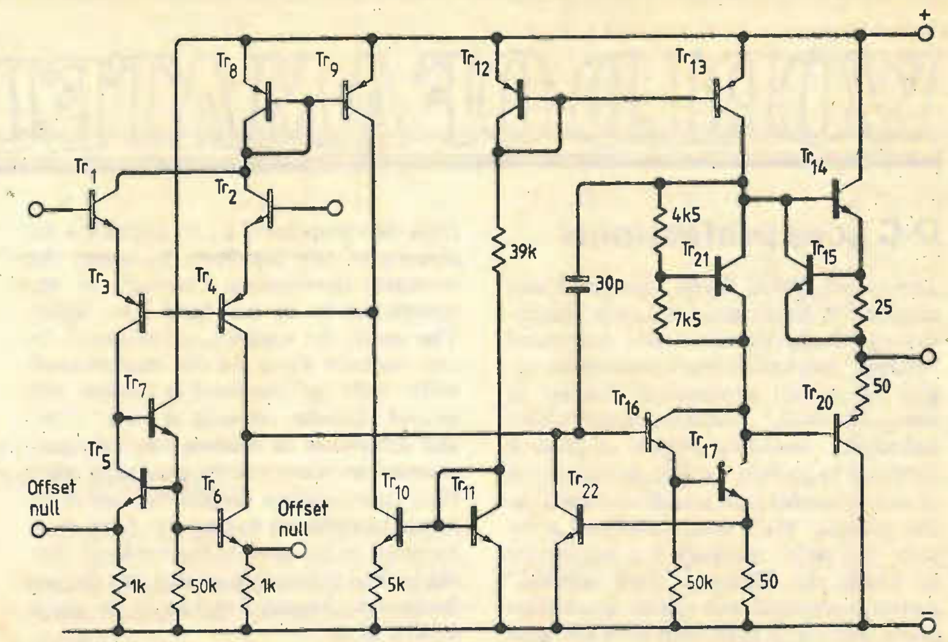


Fig.7. Complete circuit shows arrangement for offsetting d.c. error in the output. Reducing value of appropriate emmitter resistor in the input stage produces an output voltage shift equivalent to gain × input offset.

of technical papers aimed at proving the superiority of one or other commercial product.

To live within this limitation, it is necessary to ensure in all cases where slew-rate limited output saturation is important – and not all applications would be influenced by this – that the maximum rate of change of voltage in any input signal does not approach the output slew-rate divided by the effective a.c. gain.

There are now a large number of more recently designed and more expensive third generation operational amplifiers, in which both the small-signal bandwidth and slewing rate are much greater (by a factor of ten or more) than is the case for the 741. In some of these, such as the Ti TL071 and the RCA CA3140 types, the input bias requirements have been reduced to a level which is so low that the choice of input resistance values can be determined solely by other circuit requirements.

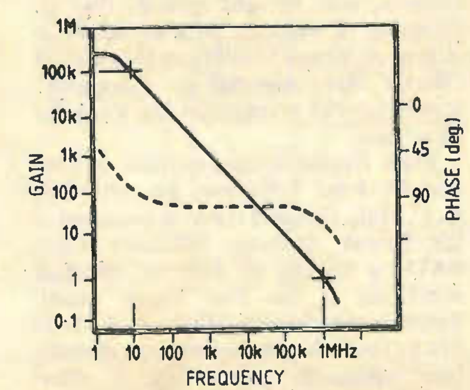


Fig.8. Frequency response shows phase error beyond 300kHz which limits unity-gain point to around 1MHz. Low open-loop gain at 20kHz limits usefulness in hi-fi applications.

FREE WITH THIS ISSUE

Extra-terrestrial relays

In October 1945 an article by a new author, a man named Arthur C. Clarke, appeared in *Wireless World*. At first glance, the subject of the article seemed to belong more to a science fiction periodical than to a technical journal like *Wireless World*; indeed, Mr Clarke subsequently became one of the best-known authors of science fiction. The second and succeeding readings, however, showed that what Mr Clarke had to say was sound sense. Here was nothing less than a scheme to use artificial geostationary earth satellites as broadcasting and communications platforms.

As everyone knows, space is now thick with satellites of all descriptions – there are 110 in geosynchronous orbit – but in 1945 it needed a great deal of thought to be sure that, by publishing such an article, *WW* would not be made to look foolish.

There is currently a crescendo of activity and speculation on the use of satellites for television and data communication, and readers might like to see how it all started. This month, therefore, we have included a reprint of the original article as an insert in those issues distributed in the UK. It was not possible to do this for overseas readers, but if anyone abroad would like a copy, they need only send a stamped, addressed envelope to Wireless World, Quadrant House, The Quadrant, Sutton, Surrey SM2 5AS, whereupon it will be sent off immediately.

WORLD OF AMATEUR RADIO

D-C goes professional

The recent IERE Clerk Maxwell Commemorative Conference at Leeds University on "Radio Receivers and Associated Systems" emphasised the considerable degree of current professional interest in direct-conversion (homodyne/synchrodyne) techniques, including the use of phasing networks to provide flexible demodulation of various modes such as narrow-band f.m. For example MEL have developed a 20-watt "Callpack" manpack h.f. transceiver in which the Weaver "third method" system, combined with digital quadrature phase shifting, is used both for s.s.b. generation and demodulation. Similarly, I.A.W. Vance, G3WMS and a team at STL have further developed their integrated-circuit n.b.f.m. d-c receiver to minimize the effects of oscillator phase-noise by means of what they term an "amplitude-normalized sine-cosine demodulator" and expect to see widespread use of this class of (mobile) receiver in the near future since it permits almost the complete receiver to be put on a chip. Philips Research Laboratories have been investigating the use of surface acoustic wave (s.a.w.) resonators to provide fixed-tuned d-c v.h.f. paging receivers suitable for the British Telecom National Paging System, the s.a.w. resonators being used to provide both a selective input filter and for oscillator frequency control at a fraction of the cost of using quartz crystals.

Although homodyne techniques date back to the 1920s, it seems fair to claim that much of the current professional interest stems from the work of J. P. Costas, W2CRR of General Electrics (US) in the 1950s followed by J. R. White, W2WBI, K. Spaargaren, PAoKSB and Wes Hayward, W7Z01 plus a whole decade of active amateur experimentation with simple direct-conversion techniques.

Vintage c.w. equipment?

Bob Locher, W9KNI writing in *Ham Radio* suggests that amateur h.f. equipment reached its operating zenith in the mid-1950s in the form of such equipment as the Collins 75A-4 receiver and the same firm's KWS-1 transmitter. He pinpoints what he regards as a subsequent decline in overall performance as starting with the general introduction of h.f. transceivers, following the success of the Collins KWM-1 in the late 1950s, due to the economic advantages and size reduction that was possible by combining both transmitter and receiver into a single compact unit.

His main complaint with the current generation of factory-built equipment from the viewpoint of a c.w. operator is the absence of any capability to ensure the accurate zero-beating (netting) of the transmitter to an incoming c.w. signal. The result, he suggests, continues to be c.w. contacts where the two stations gradually "walk up" the band in tandem, due to each operator retuning at each "over" and difficulties in competitive "pile-ups" because operators cannot accurately place their transmissions despite the use of receiver incremental tuning etc. Even more harmful, in his eyes, is the tendency during normal contacts to occupy two distinct frequency channels, separated by up to 0.5kHz or so.

There seems to be no easy solution to this problem, which today exists to some degree even where separate receivers and transmitters are used. The recent high-cost Collins h.f. transceiver type KWM380 does include a fairly simple means of minimizing the problem and ensuring that the offset between the transmitted and received frequency is exactly equal to the frequency of the audio c.w. monitor. Even better, Bob Locher adds, would be to make variable, with perfect tracking, both the frequency of the c.w. monitor and the offset differential, so that the audio monitor only could be keyed and then adjusted to zero beat precisely the incoming signals.

Many c.w. operators would agree that current h.f. equipment is not ideal; my own pet annoyance is the frequent absence of the ability to switch off the a.g.c. system.

Here and there

Very high levels of solar flare activity were recorded around the end of July, resulting in disturbed conditions on the h.f. bands and one of the most pronounced periods of auroral activity ever recorded in Europe, with stations as far away as Moscow being heard in the UK on 144MHz. It is, however, now thought unlikely that the reception of southern African signals in Athens on 50 and 144MHz on February 16 (WoAR May) was due to "long-path" transequatorial propagation but along the direct path.

Roger Appleton, chief engineer of London Weekend Television, has succeeded R. C. Hills, G3HRH (IBA) as president of the British Amateur Television Club. BATC is holding an amateur television exhibition at the Post House Hotel, Leicester on Sunday, October 4 (from 11a.m.) including demonstrations of members' equipment, narrow-band tv, slow-scan tv and F. J. ("Dud") Charman, G6CJ's "aerial circus" (on video tape).

The FCC is tightening up on its "waivers" procedure for home computers, which are providing a severe source of radio-frequency interference (r.f.i.). Test procedures and maximum permitted levels have been set for both radiated and mains-borne field strengths. These regulations are more severe than those for large professional computers since domestic models are considered more likely to be located close to television and radio receivers. Until recently a number of "waivers" to the regulations have been authorised to give manufacturers time to modify designs.

The Amsat Phase 3B amateur radio satellite is now expected to be carried on the same Ariane launch vehicle as the European Communications Satellite (ECS) now scheduled for launch on an unspecified date between June and October 1982.

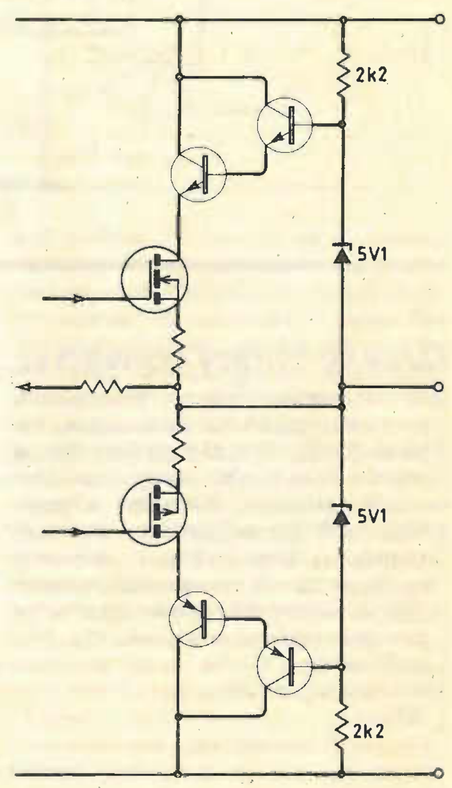
ARRL are investigating a number of cases where bogus or "altered" QSL cards have been submitted by amateurs for the DXCC listing. The suspect cards include a number for contacts made (or claimed to have been made) with "expeditions" in the 1960s and 1970s. One amateur connected with past expeditions is alleged to have admitted that 20,000 bogus cards were printed and issued by his group alone.

In brief

A new 70.12MHz solid-state beacon transmitter at ZB2VHF, Gibraltar has been well received in the UK and on July 19 several British amateurs made contact on 70MHz with ZB2BL Gibraltar, one of the very few European countries, apart from the UK, where 70MHz operation is permitted . . . RSGB's Senior Rose Bowl trophy for the 1981 Commonwealth Contest has gone to the Canadian amateur John Sluymer, VE60U with 480 contacts, 158 of them carrying bonus points . . . In a number of American states it is illegal to drive a car wearing headphones regardless of whether these block out road noises . . . Three Canadian amateurs have been experimenting with the use of the 10GHz band for mobile operation and have found it possible to make contacts over distances of up to about a mile despite screening from buildings and other vehicles . . . FCC has opened an enquiry into issuing permits to "advanced class" licence holders to experiment with spread-spectrum techniques in the 50, 144 and 220 MHz bands . . . In a policy statement, the RSGB has confirmed that it welcomes c.b. provided that it is suitably regulated; will continue to emphasize the differences between c.b. and amateur radio; is prepared to support low-power f.m. on 27MHz with officially-approved equipment; welcomes the 934MHz allocation; and "will do whatever is within its power to prevent c.b. operation within any amateur bands."

PAT HAWKER, G3VA

CIRCUIT IDEAS

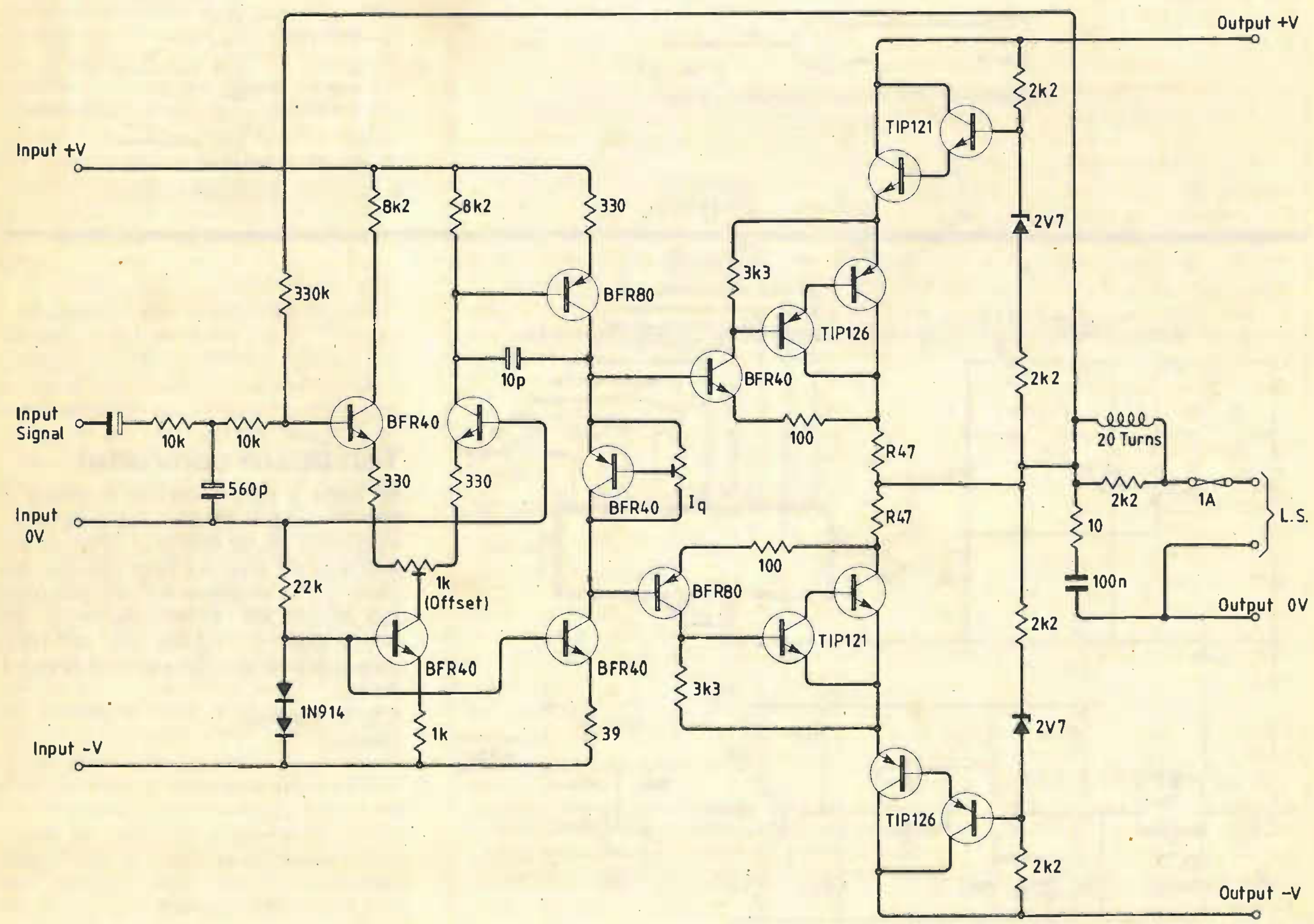


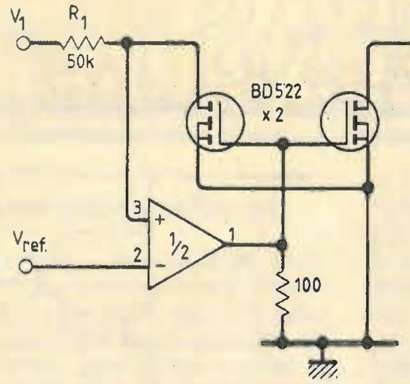
Linear power-amp offers high stability

Although many amplifiers claim good damping factors, speaker resonance often affects the input stages, even with low-impedance cables. This amplifier uses a novel output stage with control transistors connected as quasi-emitter-followers for high linearity. Slave devices provide a voltage to the control transistors about halfway between the output and the supplies, which helps to share the dissipation and reduce the possibility of secondary breakdown. The amplifier provides a 25V/μs slew rate and is unconditionally stable into any load down to 2Ω. Input and output supply rails should be connected at the power supply to eliminate the need for local decoupling. A 3A power supply at up to ±50V is recommended for a maximum output of 60W.

Fig. 2 shows an alternative output configuration which uses slave transistors to dump current into a low power output stage such as a class A or v.f.e.t. as shown. In this circuit most of the dissipation is in the slave transistors.

Q. Rice
New Malden
Surrey





Analogue multiplier

A simple high-impedance analogue multiplier can be constructed using two v.m.o.s. transistors and an op-amp to simulate a resistor which is proportional to $R_1V_{ref}/(V_1-V_{ref})$. This circuit represents a non-inverting amplifier whose output voltage is V_1V_2/V_{ref}.

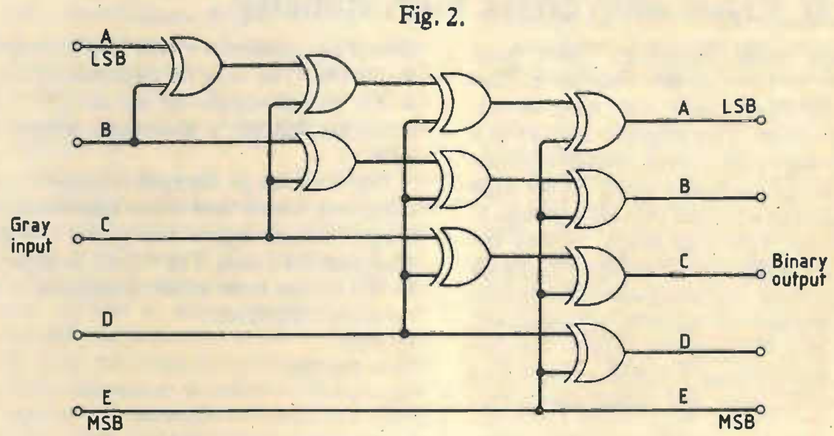
K. Kraus
Rokycany
Czechoslovakia

Gray to binary converter

When converting Gray to binary, each time a more significant bit is added, the relationship between the previous bits is inverted but the new bit has the same value in Gray and binary. Therefore, a single exclusive-OR gate will convert a Gray code to binary as shown in Fig. 1. For more bits, the circuit can be expanded as shown in Fig. 2. Converting from binary to Gray is also easily achieved as shown in Fig. 3.

J. J. Mouton
East London
S. Africa

Fig. 2.



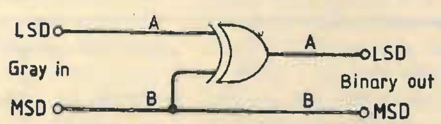
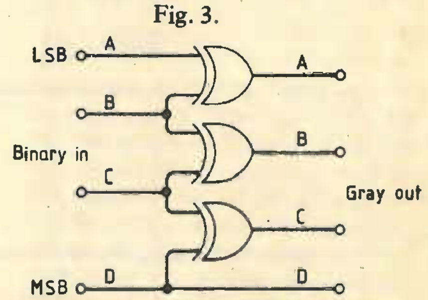
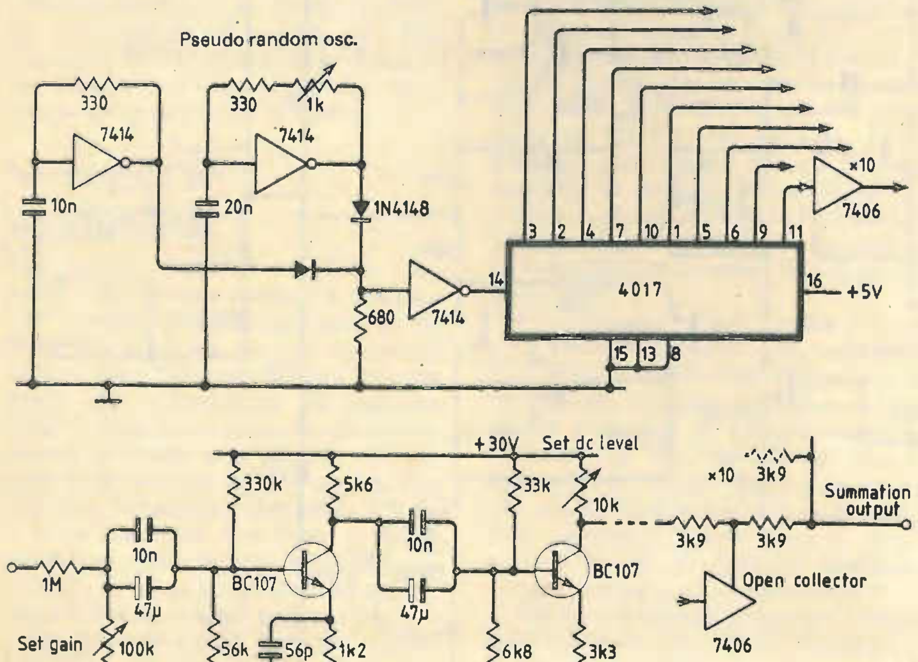


Fig. 1.

Fig. 3.





Ten beam converter

By using a pseudo-random oscillator to provide unequal periods, ten oscilloscope waveforms can be displayed which do not lock with the timebase from the chop frequency. The amplifiers are low-gain types and require 30V. Open collectors in the 7406 switch the signals and the 10kΩ potentiometers position the waveforms on the c.r.t.

J. R. V. Hawkins
London

Quantifying amplifier sound

Why can we distinguish amplifier sound in listening tests?

by Yoshimutsu Hirata, Waseda University, Tokyo

An important problem still unsolved in audio is the correlation between subjective and objective quantities. But it is more important to answer the question of why we can distinguish "amplifier sound" from the sound of a loudspeaker.

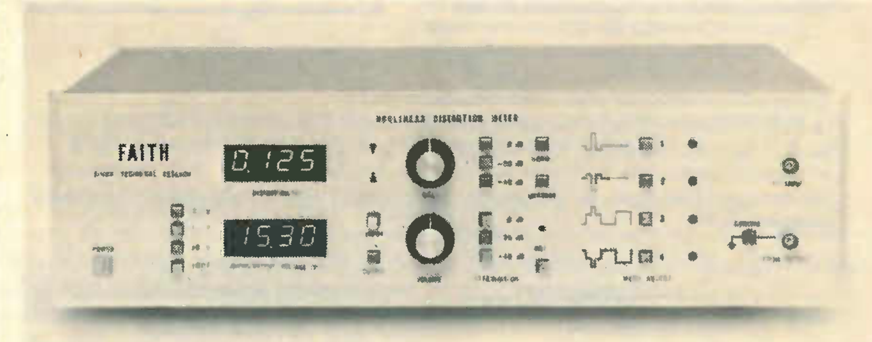
Audible differences of amplifiers are the inherent distortions, whatever they may be. The harmonic distortion of a high quality amplifier is usually less than 0.1%, while the distortion of a loudspeaker is more than 1%. In spite of this we can distinguish "amplifier sound" from the sound of a loudspeaker and point out differences in the quality of amplifiers. This implies that the distortions in amplifiers and loudspeakers differ in properties which cannot be expressed by the total harmonic distortion measurement.

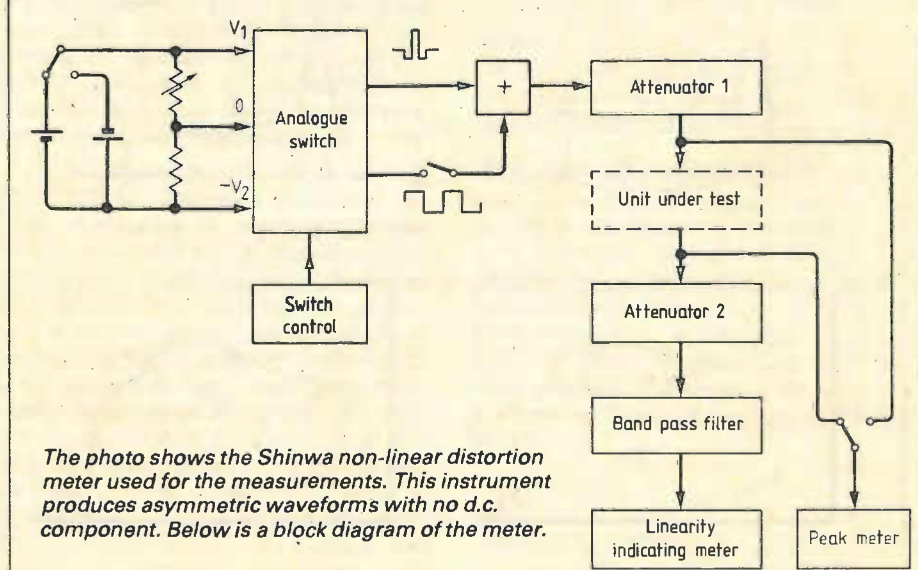
It has been recognized that harmonic distortion generally does not give good correlation with subjective assessments of sound quality. To give an improved subjective agreement, several methods of measuring non-linear distortion have been proposed [1]. The gap between the subjective quantity and the total harmonic distortion measurement is explained to start with by the difference of signals used for tests, *viz.* a musical sound and a sine-wave signal.

A musical sound involves many transient sounds whose waveforms are generally very complicated; see Fig. 1. What is a common feature in those waveforms? If anything is common in those waveforms, it is expressed by an asymmetric waveform having no d.c. component. Many waveforms have such properties: Fig. 2 (top) shows the waveform of a model transient sound, $h(t)$, which consists of half-sine waves a and b. Amplitudes of the positive and negative waves are different from each other and the area of the waveform above the zero axis is equal to the area below the zero axis. Thus, the asymmetric waveform $h(t)$ has no d.c. component. Fig. 2 also shows a plot of the frequency spectrum of the waveform where $S_1(f)$ and $S_2(f)$ are spectra of impulses a and b respectively. At low frequencies the spectrum shows a 6 dB/octave slope. When an amplifier under test alters such a waveform, the area of the altered waveform above and below the zero axis will be different, in accordance with non-linearity in gain. This difference gives rise to a d.c. component, coupled with an increase of low-frequency components of the waveform. The spectrum function of the altered waveform can be obtained mathematically by expressing the non-linearity in the form of an appropriate equation[2].

Effects of several different non-linearities on the waveform and spectrum of the model signal are shown in Figs 3-7, where $h_a(t)$ is the altered waveform, $\Delta(t)$ the deviation from the waveform $h(t)$, $D_a(f)$ the spectrum of $\Delta(t)$, $S_a(f)$ the spectrum of $h_a(t)$, and so on. An s-type non-linearity of an amplifier gives rise to an increase in low-frequency components of the signal, caught as soft or 'glossy' in listening tests. A soft distortion as represented by an s-type non-linearity is sometimes preferred when the distortion is not too great. Clipping is not an operational non-linearity in the proper sense of the word, but the saturation of a system being overloaded. The effect of clipping on the spectrum is the increasing of both low and high-frequency components, which is audibly irritating and disliked. In the case of a crossover distortion, the low-frequency component increases with decreasing amplitude of the input signal. This distortion is remarkable for low input signal levels. As an example of a dynamic distortion, take the distortion occasioned by a level compressor. For simplification, assume that the gain of a circuit is attenuated to reduce the amplitude of a positive pulse as in Fig. 6(a). In this case, the spectrum of the signal suffers heavy change, *viz.* the increasing and decreasing of low and high-frequency components. Unless the functioning of a level compressor and expander is ideal, a noise reduction system produces a similar distortion in an impulsive signal, which may be caught subjectively as somewhat dull or heavy by a listener.

As discussed, gain non-linearities in amplifiers give rise to a d.c. component coupled with an increase of low-frequency components of an impulsive signal. On the





The photo shows the Shinwa non-linear distortion meter used for the measurements. This instrument produces asymmetric waveforms with no d.c. component. Below is a block diagram of the meter.

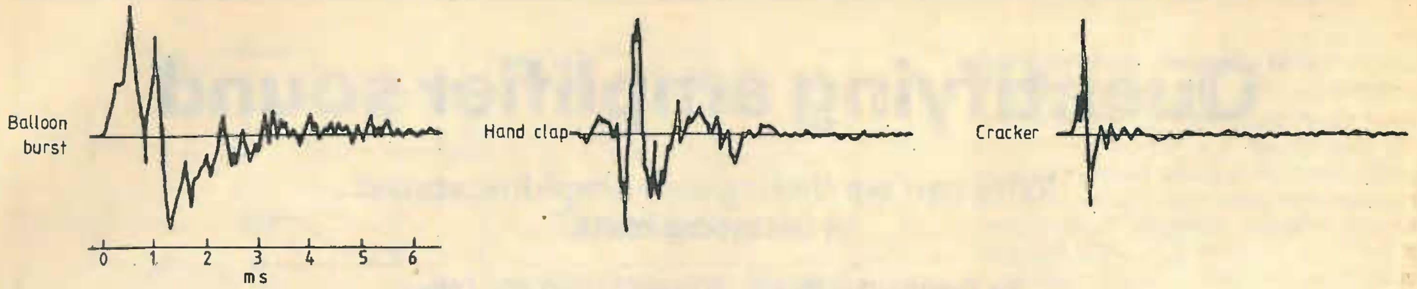
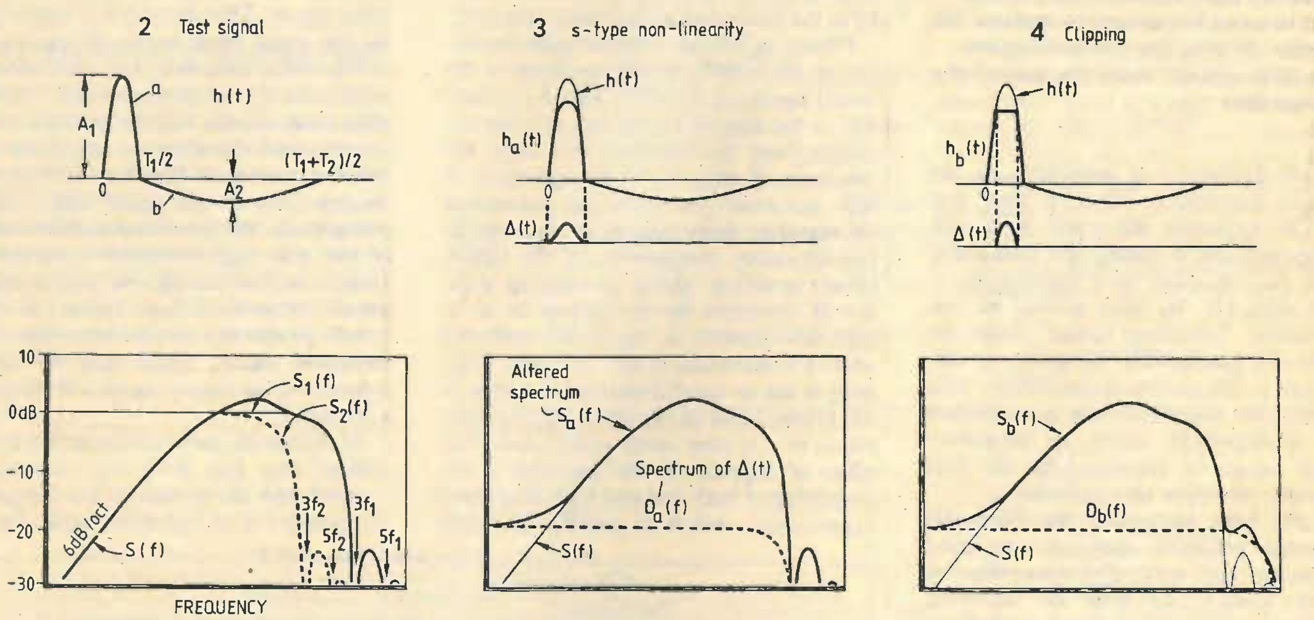


Fig. 1. Typical waveforms of transient sounds are asymmetric, with no d.c. component.



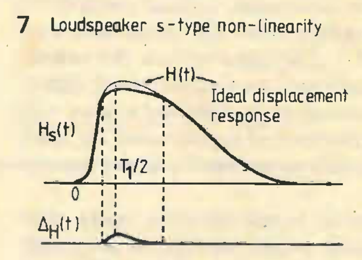
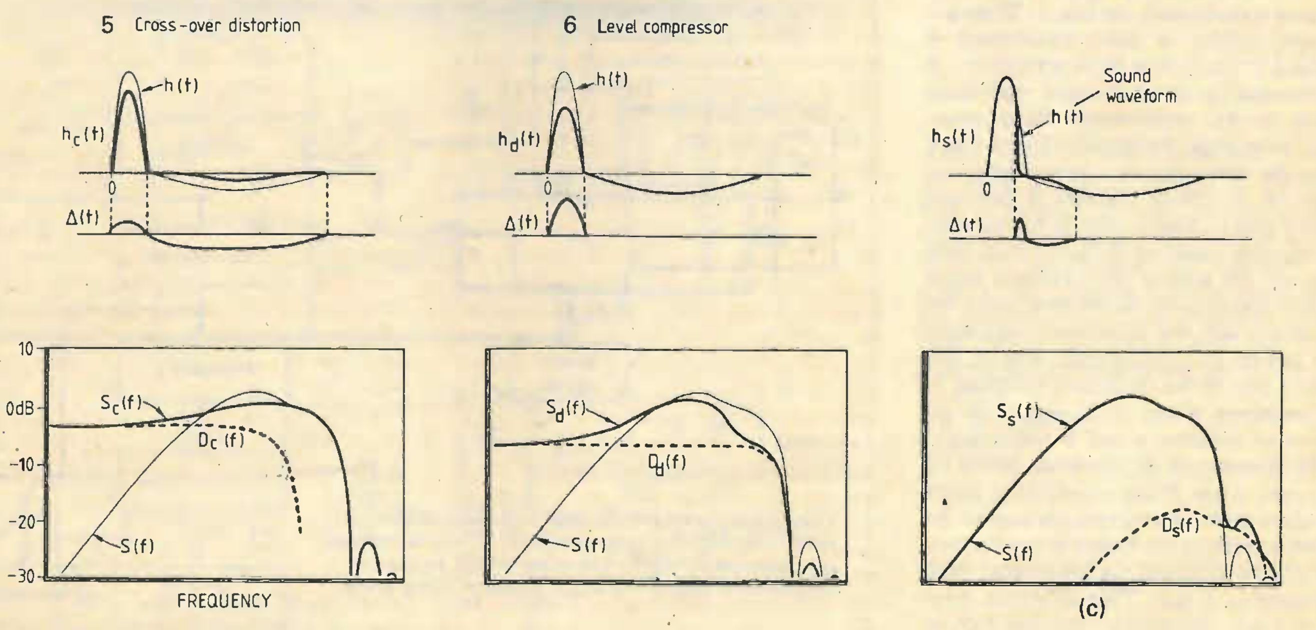


Fig. 2. Waveform of the pulse h(t) and its spectrum S(f), where $f_1 = 1/T_1$ and $f_2 = 1/T_2$. Fig. 3. S-type non-linearity. Fig. 4. Clipping. Fig. 5. Cross-over distortion. Fig. 6. Level compressor. Fig. 7. S-type non-linearity of a loudspeaker. (a) is the altered displacement response, (b) the altered waveform of sound $h_s(t)$ and (c) the altered spectrum $S_s(f)$.

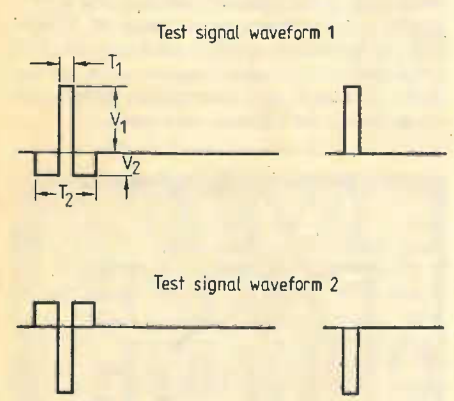


other hand the distortion in a loudspeaker does not, as a matter of course, give a d.c. component. As an example, consider the s-type non-linearity in the displacement response of a vibrating plate. To simplify an analysis, assume a sound pressure is in proportion to the velocity of a plate, *viz.* a sound is assumed to be radiated from an infinite plate. For a finite plate, the sound pressure of low-frequency components is proportional to the acceleration of the plate. The difference between the displacement response $Hs(t)$ which is altered by the s-type non-linearity and the ideal response $H(t)$ is expressed by $\Delta_H(t)$. As sound pressure is proportional to the velocity of an infinite plate, the sound distortion $\Delta(t)$ is given by the derivative of $\Delta_H(t)$ with respect to time. As shown in Fig. 7(c), the spectrum of Δt involves no d.c. component and few low-frequency components. Thus the distortion in a loudspeaker does not change the low-frequency component of an impulsive sound, which is in contrast with the case of the distortion in an amplifier. And this is the reason we can distinguish amplifier sound from the sound of a loudspeaker. This has been verified by experiments using a novel method for measuring non-linear distortion[3].

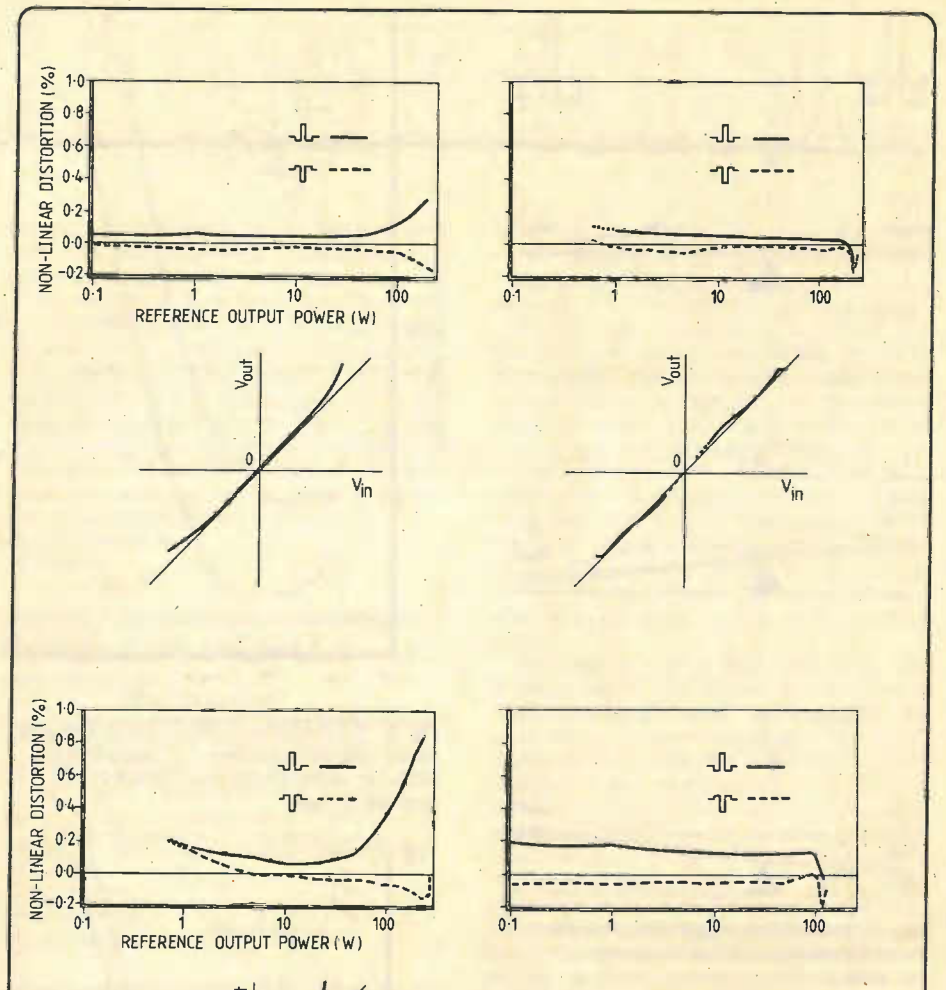
Experimental method

The model of a transient sound used for the theoretical study consisted of two half-sine waves of different amplitudes and polarities. The model signal was such that it could distinguish distortion due to non-linearities as a change of spectrum, which is related with a subjective quantity. It is however not convenient to use the model signal for the measurement of an objective quantity, *viz.* the rate of distortion.

A new method of making non-linear distortion measurements uses composite rectangular pulses, as shown below. The area of the part of each test signal waveform above the zero axis is equal to the area below the zero axis, so these asymmetric test signals have no d.c. component.



Asymmetric test signals have equal areas above and below axis and so have no d.c. component. Area of a positive pulse of the first test signal is V_1T_1 and the area of a negative pulse is $V_2(T_2-T_1)$.



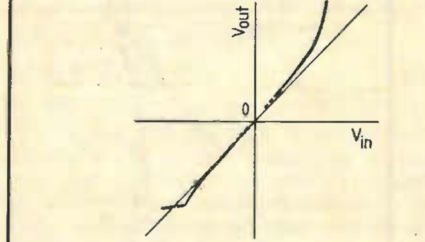


Fig. 8. These diagrams show non-linear distortion curves of three power amplifiers together with transfer function shapes. Reference power output is defined as $V_1^2/2R$ where R is 8 ohms. Top left is for the Marantz SM7, top right for the Threshold 4000, bottom left the Amcron SA2 and bottom right the Meridian.

When an amplifier under test alters the applied test signal the areas of the altered waveform above and below the zero axis will be different in accordance with the gain non-linearity. This difference gives rise to a d.c. component coupled with an increase or decrease of certain low-frequency components of the test signal, either of which can indicate the degree of linearity of the amplifier under test.

The repetition frequency of the test signals is 220 Hz, chosen to lie between the higher harmonics of power-supply frequencies 50 and 60 Hz. At low frequencies the envelope of the test-signal spectrum has a 12 dB/octave slope and the component at 220 Hz, normalized to that of the reference signal, is 0.002 (−54 dB) which is the theoretical value in the absence of non-linear distortion. Thus the difference between the normalized component at 220 Hz of an altered test signal and the theoretical value of 0.002, indicates the value of non-linearity of an amplifier under test. When the gain non-linearity is static, i.e. when the i/o property is expressed by a single curve, the non-linear distortion D corresponds to the linearity deviation Δ. The relationship is

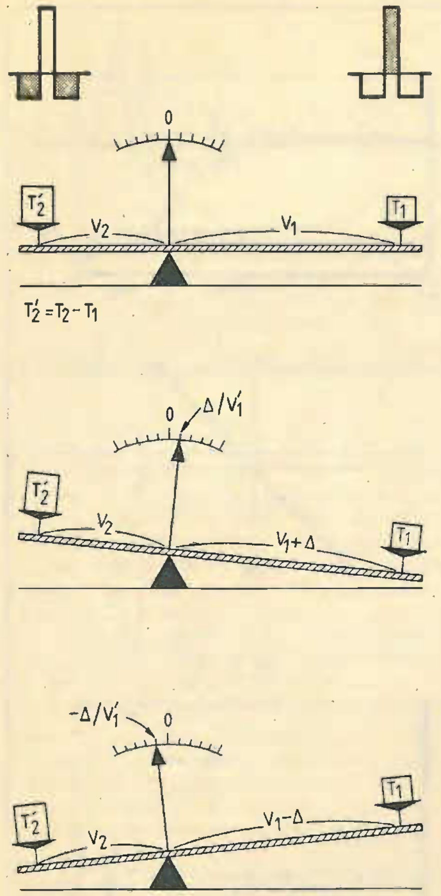
$$D = |\Delta/V'_1 + 0.002| - 0.002$$

where $V'_1 = V_1 + \Delta$,
which is nearly equal to V_1 for $\Delta \ll V_1$.
This gives $D = \Delta/V'_1$ for $\Delta/V'_1 \geq -0.002$
and $D = -\Delta/V'_1 - 0.004$ for $\Delta/V'_1 \leq -0.002$.

The figure of D involving V_1 indicates directly the form of gain non-linearity, i.e. an s-type non-linearity, cross-over distortion, and so on. (Notice that the figure of D given by the last equation corresponds with the figure given by $D = \Delta/V^1$ which turns up at $D = -0.002$. Thus the lower limit of the D (%) axis is −0.2%). The block diagram of the measuring apparatus is shown under the photograph on the first page.

Results

One of the advantages of this method is that calibration of the test signal can be



Beam in balance analogy tilts according to the increase (Δ> 0) or decrease (Δ<0) of V_1.

done with a circuit connected at the output of the apparatus, which enables distortion to be measured in a large system such as a console. Fig. 8 shows measured distortion figures of commercial audio amplifiers withn an 8Ω resistive load and the form of non-linearity given by the respective distortion figures. Rise and fall times of the test signal are about 10µs/V. When there is a difference in slew-rate between build-up and fall of voltage amplification it gives rise to a d.c. component coupled with low-frequency components, and distortion figures for the first and second test signal will be shifted upward in the case of a larger build-up slew-rate, and *vice versa*. Such an example is shown in Fig. 7. Figures 9 and 10 show the distortion figures of noise reduction systems used in cassette tape recorders, where the input test signal was superimposed on a rectangular waveform which has no d.c. component. Although large distortions are due to the inherent non-linearity of recording media, it is shown that noise reduction increases the low frequency components.

To verify the fact that amplifier sound can be distinguished from the sound of a loudspeaker, non-linear distortion measurements were made using a loudspeaker. Fig. 11 shows the distortion figure of a high-quality audio amplifier, measured by using the sound reproduced from a loudspeaker connected to the amplifier output. The measurement was made in a low-noise

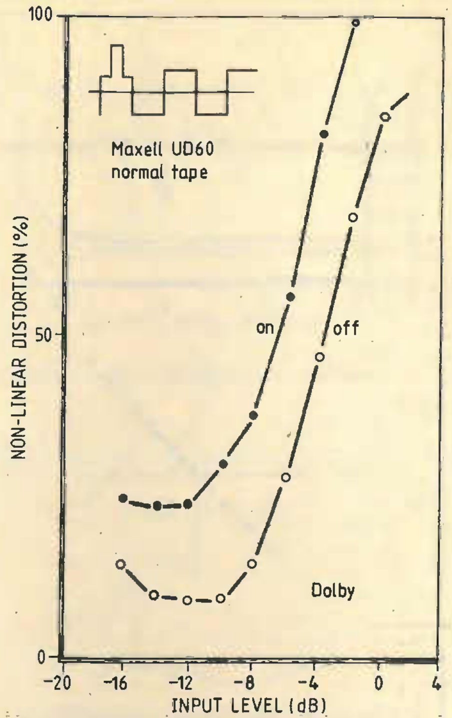


Fig. 9. The non-linear distortion graph of a noise reduction system in a cassette tape recorder using Dolby and Maxell UD60 (standard) tape.

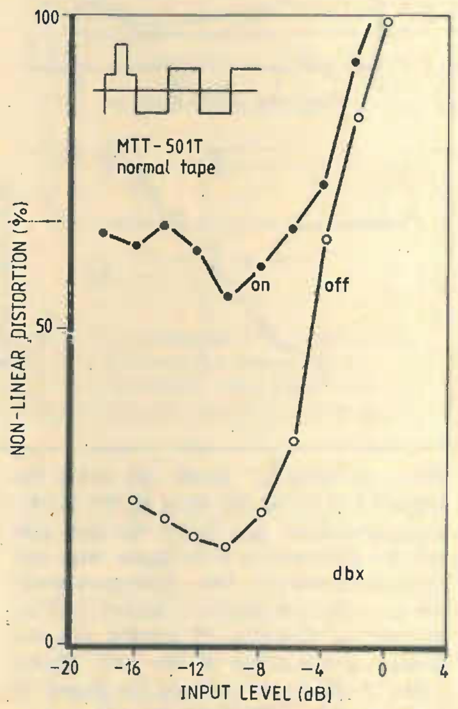


Fig. 10. As Fig. 9, but using dbx and MTT-501T cassette tape.

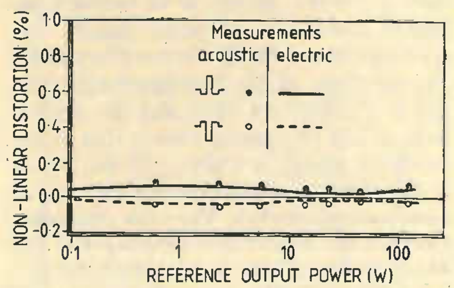


Fig. 11. Non-linear distortion of a Kenwood L-07 MII power amplifier. Broken lines represent electric measurement and solid lines acoustic measurement.

room so that the component of 220Hz could be correctly measured. Care was taken to avoid the effect of non-linearity in the microphone and measuring amplifier. Fig. 11 also shows the distortion figures obtained by measuring the voltage across the voice coil of the loudspeaker. Evidently, the non-linear distortion of an amplifier can be distinguished from the distortion of a loudspeaker. So we can really distinguish amplifier sounds from the sound of loudspeakers objectively as well as subjectively.

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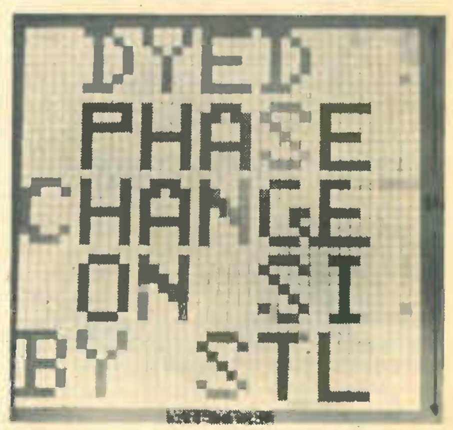
69×69mm flat-panel display

The first step towards production of a 69×69mm, 57,600 pixel flat-panel display has been made by Standard Telecommunication Laboratories in the form of a 36×36mm, 1,600 pixel prototype. The end product will be a liquid-crystal display only a few millimetres thick with built-in drive electronics and able to display, for instance, a full page of Prestel. STL, a subsidiary of STC, disclosed details of what they call the sub-miniature v.d.u. in New York recently at the International Symposium of the Society for Information Display.



LETTERS TO THE EDITOR

CITIZENS BAND

Mr Frost says in July letters that operators of illegal c.b. gear "have every reason to grumble at the expense of changing to the new system". I can think of a few other groups having "reason to grumble":

—housebreakers, forced by new lock designs to obtain a new set of lock picks;

—radio control aeromodellers, faced with a move to the 35MHz band;

—car and boat modellers, who have not been given a new band, but are left with one polluted by c.b.;

—hospital administrators, forced to buy new paging systems to avoid 27MHz c.b. interference – see the letter on page 2 of *Electronics Weekly*, May 13th, 1981.

After the publication of a letter in a local paper in which I contradicted inaccurate statements by a c.b. enthusiast, objects were thrown from a passing car at my house. For this reason, I would be grateful if you did not publish my name and address.

(Name and address supplied)

DECLINE OF THE PHILOSOPHICAL SPIRIT

The editorial of the July 1981 edition of *Wireless World* rightly draws attention to the impoverishment of science and philosophy by their artificial separation. It is therefore of great importance to consider what practical steps are available for us to remedy this unhappy situation. Fortunately, the words of the great scientists are full of practical advice on this. For example, Michael Faraday said "Here it is we are born, bred, and live, and yet we view things with an almost entire absence of wonder to ourselves respecting the way in which all this happens. So small, indeed, is our wonder that we are never taken by surprise". On a similar theme, Albert Einstein said "The fairest thing we can experience is the mysterious. It is the fundamental emotion which stands at the cradle of true art and true science. He who knows it not and can no longer wonder, no longer feels amazement, is as good as dead, a snuffed-out candle."

This wonder, which leads to 'natural philosophy', does not have to be acquired, it is there so obviously in the young child. It might well get covered over by the ideas described so well in the editorial and then a greyness appears to cover the physical world. It does seem, however, that it can be readily uncovered by the application of the scientific principles to our everyday experiences. To take a small example, on tuning a radio over its waveband, a large number of stations are heard. It does not take much reflection to appreciate that the radio is separating and manifesting what is all together in the room. The space in the room is pervaded not only by countless man-made electromagnetic signals, but also by the electromagnetic signals we call light from every body in the room, the magnetic and electric fields of the earth and beyond, the gravitational influence of every atom in the universe and so on. Such a reflection very easily turns into wonder.

As demonstrated so clearly by the great scientists, this emotional response is not a waste of time but "the cradle of true art and true science". Their works show the ability to simplify and unify previously complicated and separate areas of scientific endeavour.

It is the duty of those involved in the education of the young in science and engineering subjects to uncover this 'fundamental emotion' in the students by what is said, more so by what is done in the laboratory, but most of all by the teacher's love of his subject. Perhaps in this practical way it is quite possible by our scientific activities to, as Plato put it, "create the spirit of philosophy, and raise up that which is unhappily allowed to fall down".

M. J. Cunningham,
Manchester

WIRELESS WORLD 1911-1981

Congratulations on seventy years great service by *Wireless World* and *The Marconigraph*. As far as I can ascertain, the Technical Library operated by this company for its employees is the only library in Australia with a complete run from issue number 1 of *The Marconigraph* to date.

B. J. Simpson,
Chief of Patents,
Amalgamated Wireless (Australasia) Ltd

JAMES CLERK MAXWELL

Mr Wellard, in his article on Maxwell (May 1981) in my opinion spoiled an otherwise interesting article by some outrageously contentious statements which were unsubstantiated in the article and run completely counter to the present day knowledge of atomic and nuclear physics.

The *mass* ratio of electrons to protons is about 1:1820; but the *charge* ratio is exactly 1:1, but of opposite polarity. The body of experimental evidence for these ratios is extremely strong since the design of mass spectrometers, mass separators, synchro cyclotrons, electron beam systems (including c.r.ts) and many others depend on an accurate knowledge of the properties of electrons and protons.

His article implies that hydrogen atoms and molecules are not electrically neutral. The implications of this are very far reaching, but as far as I am aware, having had considerable experience of ion sources, electron beams and hydrogen gas handling, there is not a shred of evidence to support the implication of non neutrality.

Furthermore the statement about the neutron . . . 'this non-existent particle' . . ., is also highly contentious. How do nuclear reactors work if neutrons are non existent? Why do nuclear physicists refer to thermal neutrons and fast neutrons, and the resulting reactor types, thermal and fast-breeder reactors when, as Mr Wellard asserts, neutrons do not exist? Again, neutron radiography is an extremely useful alternative to X-raying for non-destructive testing of various engineering parts. How does this work if there are no neutrons?

It is difficult to follow Mr Wellard's logic in the article when such unsupported outrageous statements are made. If he is keeping some new important experimental data from the world then he should publish it properly in some form where it can be scrutinised and his experiments repeated, and so be tested for their universality.

If, on the other hand, he has no basis for his remarks, except that he does not like the idea of electrons and protons having an equal magnitude of charge, or does not like the idea of neutrons, then it is just too bad: they are an observational fact. Many people did not like the idea of a spherical earth or a helio-centric solar system, but both are observational facts, around which valid models can be constructed. We may not know what a neutron is, since we cannot handle it in the way we can handle familiar everyday things, but its properties can be quantified (e.g. mass) and used for prediction purposes.

Referring now to a letter concerning Mr Wellard's article by H. Aspden, he states that in 1980 experimental 'proof' that the aether can assert a force was reported in *Nature*. Not many of your readers would have access to the article named in *Nature*, which reports some interesting experimental results, as yet uncorroborated, which may be evidence for an aether (or something) but is not proof as yet. Mr Aspden is a very impressionable person if he considers one experiment to be proof. The experimental technique will have to be very carefully looked at to ensure that no other phenomenon was responsible for the effects measured, and the test repeated elsewhere with another apparatus to confirm the results. Only then can it be asserted that there is enough evidence to provisionally confirm the implication of an aether.

B. J. C. Burrows,
Benson,
Oxford

The author replies:

The question of the equivalence of electrons, protons and the hydrogen atom was answered in my reply to T. de Limelette (August letters). The equation does not satisfy Maxwell's test and is therefore absurd. The neutron is one half cycle of an electromagnetic wave; the anti-neutron is the other half cycle. They are mathematical myths. The infinite acceleration of an action-at-a-distance particle generates an infinite amount of energy. I assume from the third paragraph of Mr Burrows' letter that he is an experimental physicist. If he finds his working model of a particle more useful in his work than the working model of a wave, he should carry on as usual; he has no sensible alternative at present. The passing of an examination requires the unconditional acceptance of a working model and its associated equations. I am suggesting that the working model he was forced to accept was based on the wrong analogy. I am not suggesting that he or any other member of the scientific community is in any way responsible for the constrictions imposed by the mathematical extremism of Lorenz or the entrepreneurial skill of Einstein. The flaw in any idea or belief is its dogma; identify the dogma and you identify the flaw. As far as I know, a physicist is no different from any other human being infused with an extreme idea. Mr Burrows writes in plain English and must be capable of plain and open thinking.

Mr Burrows is unfair and unwise to include an innocent bystander in his deprecations. He asks whether I am withholding new important experimental data from the world. The answer is no. I would have thought that every experiment with an electromagnetic wave proved the presence of a medium. Does he know the whereabouts of another Einstein withholding experimental data that proves otherwise? Two books by Dr Aspden received a favourable mention in Dr Essen's attack on Relativity (October 1978 issue). Both gentlemen are Flat Earthers. Now that the memory of the late Professor Dingle has joined them, Mr Burrows is wasting his time seconding me to such an exclusive club.

To prove just how exclusive this club is, I have taken a look at Einstein's famous equation of energy, $E=mc^2$. The dimensions of work or energy are ML^2/T^2, the work done by a unit of force (ML/T^2), accelerating through unit of length measured in its own direction. This is mathematically equivalent to the product of a mass and the square of a velocity (L/T), and even if the velocity to be squared equalled one metre per year, work would still be performed by an accelerating mass. Einstein's famous equation has the dimensions of work or energy, but implies, in fact insists, that work is only performed by an accelerating mass when the velocity to be squared is equal to the 'constant' speed of light. His equation is meaningless, misleading, and very very slick. Einstein believed the Earth was round. Does this prove that Flat Earthers do not subscribe to the theory of the great philosopher 'Fats' Waller, " 'Tain't what you do, it's the way that you do it – that's what gets results."?

Mr Burrows asks some awkward questions in his fourth paragraph. Why not leave the microscopic dictatorship of nuclear physics and try the macroscopic democracy of astronomy, looking for analogies that do not allow energy to disappear? It has been suggested that this universe is the inside of a huge spherical atom. Taking the analogy further, a radioactive atom would be filled to bursting point with colliding quasars and the resulting big bangs and young galaxies. Quasars are continuously losing a vast amount of energy. Are they transformed from electromagnetic waves when the waves are absorbed, and transformed back again into waves when they are emitted at a lower energy level? Are they by analogy massive radioactive atoms which on devolution are emitted as spent neutron stars or helium atoms? Are there intermediate-size groups of atoms between the Earthly and the Universely? How many universes are there? As many as there are atoms in our universe? Would over 200 'particles' be emitted if a radioactive universe disintegrated? *Wireless World* regrets the decline of the philosophical spirit, and so do I.

"TRUTH TABLE" LOGIC SYMBOLS

There are many disadvantages in the system of intentional logic symbols (November 1980 issue, pp 61-62), the main ones being that once away from input stimulii no "true or assertive state" exists unless the circuit is further complicated by adding some form of flag to indicate an artificial "true or assertive state" at that point. In any case it is doubtful whether doubling the number of different symbols used can make diagnosis simpler, especially as identical gates may have different symbols.

There is a much simpler solution to the problem, that of what I would describe as Truth Table logic symbols. As far as I am aware the idea is original but I have not researched the matter. Fig. 1 shows the derivation of the symbol from what I have described as the singular logic state of the simplest form of the truth table. The gate outline shape is of no great importance provided its input and output are clearly distinguishable (this excludes the rectangular box as per BS3939). Each input has a logic state associated with it (TT input state) and each output a logic state associated with it (TT output state). In order to produce the TT output state it is necessary to make all inputs equal to the TT input state. It does not tax the brain too much to deduce that, to get the opposite state out, any state other than all inputs equal to the TT input state will do. If a gate is in an application where, for example, a signal is input to a gate and the other inputs are used to enable that gate then the gate will be enabled for that signal then all other inputs are equal to the TT input state.

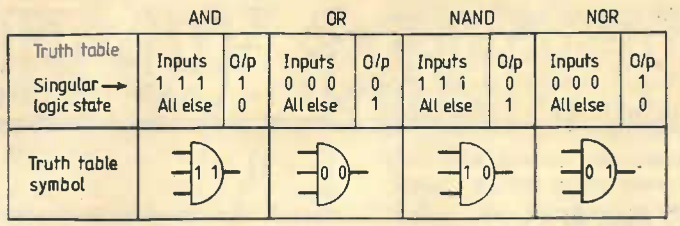


Fig. 1

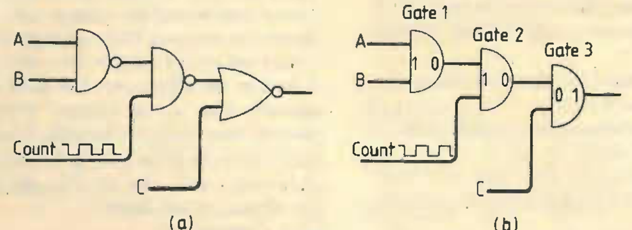


Fig. 2

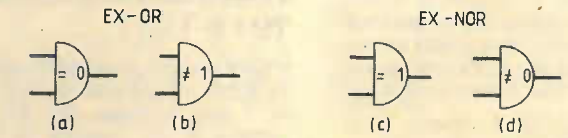


Fig. 3

Fig. 2(a) shows the problem as stated by Mr Cassera in his November article on intentional logic symbols. Fig. 2(b) shows how it would be drawn using TT logic symbols.

In order to enable gate 3 input C must be 0.

In order to enable gate 2 the output of gate 1 must be 1. Both are immediately obvious from the fact that to enable a gate enabling inputs must be equal to the TT input state.

To get 1 out of gate 1 any combination of A and B will do other than all 1s. This follows from the fact that the wanted state is not the TT output state.

The absence of the inverting symbols may worry some engineers used to conventional symbols but a NAND gate is only inverting because of the way the AND is defined. Few engineers would be entirely happy with this explanation but they can take comfort from the fact that is is extremely easy to tell conventionally inverting from conventionally non-inverting gates in that for inverting gates the TT input and TT output states are different implying an inversion.

The EX-OR and EX-NOR gates present a choice of symbols as shown in Fig. 3. If starting from scratch one would choose Fig. 3(a) and Fig. 3(c). Unfortunately Fig. 3(c), the symbol for an EX-NOR, is too similar to the conventional EX-OR symbol, therefore Fig. 3(d) should be used for an EX-NOR. This would be interpreted that in order to get 0 out the inputs must be not equal. Either 3(a) or 3(b) could be used for EX-OR, 3(a) being preferred on the grounds of simplicity. To summarise, the system is very simple, is very largely self-explanatory, and requires little mental effort to change from existing practices. All identical gates have identical circuit symbols, the symbols themselves being uncomplicated.

J. E. Kennaugh
Callington
Cornwall

TELEVISION AND THE DEAF

I have been interested to read the correspondence about amplifying tv sound for the deaf. Some years ago I was asked to help a deaf friend with this problem. A great deal of amplification was required but it was obviously going to be an advantage if normal hearing people could follow the programme without being blasted out of the room. Amplified headphones are not sufficient as the sound loses quality compared with that from a properly made earpiece with a mould made specifically for that person's ear. In addition many hearing aids have some form of equalization to try to compensate for hearing response curves changing with frequency.

It seemed to me that the answer was clear: use the amplification, equalization and earpiece already in the possession of the deaf person. Many hearing aids have some form of telephone pick-up coil and maybe some switching to allow "mic", telephone or both. Fortunately the Sony television to be used was equipped with both "listen" and "break" miniature jacks; giving the option of "silent" or joint listening. Providing a signal to drive the telephone pick-up was then very easy: a small coil fed from the earphone jack gave more than enough level. I used a 1kΩ Post Office 3000 relay coil to prove the system, but as this was, to say the least, somewhat clunky, I replaced it with a small plastic covered coil marketed as a telephone pick-up but, of course, used in the other direction to that intended! All that needed to be done was to extend the lead.

This method has the advantage of being cheap, safe, and very good quality. It is easiest to use with body worn hearing aids but even deaf people using ear-level aids often have a second, body worn aid as a spare.

Incidentally, I have also used a telephone pick-up coil to detect the ringing current in the coil of a telephone bell: suitably amplified and rectified this can be used to control flashing light repeaters without interfering with the Post Office installation. A few resistors, a 741, a diode and a BC108 to drive a relay are all that is required.

Roger Derry
Leytonstone
London E11

RADIO AMATEURS' LICENCE

As a citizens' band service is now proposed we, the undersigned, would like to suggest slight modifications to the radio amateur licence, as follows:

1. The use of c.w. by class "B" radio amateurs receiving and sending as part of the self-training in communication by c.w. on v.h.f. bands.
2. Limited use of station under supervision, e.g. jamboree-on-air, radio conventions, radio clubs, short wave league, XYLs, YLs etc.
3. The 27 and 930 MHz c.b. bands to be used by radio amateurs on the existing licence at no extra fee. Not type approved rigs.
4. The 10 and 4 metre amateur bands to be extended to class "B" radio amateurs; e.g. the 10 metre band to be used by licensed radio amateurs, not taken over by citizens' band.

M. Jackson (G8EOP)
Dewsbury
West Yorkshire
Also *G8WWE, G4LED, G3LHQ, G8PSE, G8EAH and 460 others.*

MICROCHIPS AND MEGADEATHS

I have no intention of cancelling my order for *Wireless World* as a result of your recent trend in editorials, but when I read some of the criticism against this, I sometimes wish I could cancel my subscription to the human race.

The electronics engineers referred to by Dr D. J. Dewhurst in June letters died, along with millions of others, in the hope that humanity would never again have to devote its time to finding ways of destroying itself.

Instead what is happening? The USA alone is planning to spend $1,500,000,000,000 on using *our* technology to produce still more weapons of death. *That is about 350 dollars for every man, woman and child alive on the earth.* That money alone could save most of the world from starvation, and I haven't even mentioned the Soviet Union and dozens of smaller countries yet!

It *is* up to us, as engineers and as ordinary people, to stand up against this. We must all throw down our arms and say that we will *not* fight their wars for them. And if I, for one, get shot as a result, it will just prove to me that this world was not the one I wanted to go on living in.

Who will stand beside me?

Tim Bierman
London NW11

BETTER RFI PROTECTION NEEDED

As a radio amateur, I wholeheartedly agree with Mr McLeod's observations (August Letters) that better r.f.i. (or e.m.c.) protection is needed for domestic electronic equipment. The problem of r.f. breakthrough is nothing new: it has existed for many years, but has become more prominent recently due to the number of illegal 27MHz a.m. c.b. transceivers now in this country.

Unfortunately, the manufacturers of domestic electronic equipment are unlikely to respond to Mr McLeod's plea for better r.f.i. protection. The design effort and extra components required would not be expensive, but the added cost would, no doubt, reduce their profit margin and/or competitive pricing. I can see two possible answers to the problem: (a) legislation to ensure that all domestic electronic equipment complies with a suitable e.m.c. standard; or (b) commercial pressure, i.e. bad publicity – if you have suffered with r.f. breakthrough, you are unlikely to buy the same make of equipment again. You would probably look for equipment which is better protected against r.f. breakthrough.

If there are any manufacturers whose products have a good e.m.c. performance, then they should say so. I am sure there are lots of customers in many countries waiting to buy their products.

P. J. Forshaw,
Runcorn,
Cheshire.

FILTER TRANSIENT RESPONSE

Thank you and Mr Hamill very much for the much-needed article on the "Transient response of audio filters" in the August issue of *Wireless World*.

In December 1977, I wrote an article entitled "A transient phase?" for *Hi Fi Answers* in which I introduced a new term t.p.d. (for transient phase distortion), and followed this up with a more descriptive article in *Hi Fi Answers*, August 1978, entitled "Transient phase distortion".

Although for the sake of clarity in these particular articles I confined the description of effects and equipment to fairly simple things, understandable to the average man in the street, I had in fact investigated the effects of impulses and their responses, and their relationship to real music signals using a "Fast Fourier" Transform analyser (as well as the storage 'scope and v.l.f. signal generator mentioned in the articles) and I very much agree with Mr Hamill that frequencies are produced which bear absolutely no relationship to any Fourier component present in the original signal – in fact the pitch is completely out of tune when a filter is used to limit the bandwidth, and noticeably so if the filter is at all "steep cut". Unlike Mr Hamill, however, I had been interested in the effects of bandwidth limitation from a truly musical viewpoint. This approach led me to conclude that *both* high-and low-frequency band limiting filters are detrimental, but that gentle h.f. filtering is *not* seriously degrading because air itself can act as a h.f. filter and often does so, thereby causing the ear to be used to the effect of mild filtering, which just sounds "natural" when introduced artificially.

Again, unlike Mr Hamill, I have found conclusively that l.f. filtering is very detrimental to realism in reproduced sound. It too has its natural counterpart, which takes the form of large areas of carpet suspended vertically close to the listener in the concert hall (to either side and rear I should point out!), or replacing the concert hall by a room 13ft square, say. Unfortunately, almost all reproducing equipment cuts off steeply below (at best) 45Hz in an average domestic living room, so the effects of filtering below this frequency are normally minimal as they are swamped by the inherent cut-off.

However, my system – demonstrated as "The" loudspeaker at "Hi-Fi 80" and recorded/photographed in July 1980 *Hi-Fi News*, page 52 – is truly flat down to *four hertz* in a room 16½ × 12ft (although this is not at all a function of room size), being +2½ dB at 6Hz and −2½dB at 4Hz in relation to 40Hz/400Hz/4kHz. For a good signal (which is not all that hard to come by on records!) it is very noticeable if the frequency response is out at 20Hz or left to go down to 4Hz flat, and the rate of cut is also critical between these figures. The effect of the filter is to remove spaciousness and precision of transient data. With the full range allowed to pass through to the ear, one can clearly hear the building boundaries, both their position and composition (the difference between a stone church or cathedral and a concert hall is very clear and real). Often air recordings sound as if they are in the open air, not in the listening room. As soon as the 20Hz filter is inserted all this disappears. The subjective effect is of an inferior performance, with both precision of tempo and accuracy of tuning or pitch affected a noticeable degree, and a removal of all that is considered "good" in the concert hall acoustic (as if the Colston Hall, for example, had been replaced by a large garden shed and the London Philharmonic Orchestra were replaced by the local youth orchestra). Of course, many recordings do not have the necessary range anyway, but many do, and it is such a pleasant surprise when this happens and realism comes through!

Graham Holliman,
Watford,
Herts.

'SPREADING'

The amateur fraternity here in Australia, and I suspect that it is much the same in other parts of the world, cling tenaciously to what I regard as a ridiculous superstition known as "spreading".

I should here explain for the general reader that the amateur fraternity these days employs almost exclusively the mode of transmission known as "single-sideband, suppressed-carrier", where the signal, before transmission, is passed through a band-pass filter restricting the bandwidth of the transmitted sideband to about 3 kHz.

The superstition to which I refer surfaces when a very strong signal is received and the receiving operator notices that he obtains an "S-9" indication on his signal-strength meter over perhaps 8 or 10 kHz on the dial of his receiver. The operator jumps to the conclusion that the transmitting station is actually radiating energy over a bandwidth of 8 or 10 kHz. Vain to tell them that this is an effect occurring in the receiver itself due to a combination of the effects of selectivity and a.g.c. Invariably, the transmitting operator is abused, and accused of negligence and incompetence.

I am wondering whether other readers of *Wireless World* have encountered this superstition and if so what they make of it.

R. C. Yates,
Charlestown,
N.S.W., Australia

Digital, multi-track tape recorder

Uses modified audio cassette deck for very low-cost, 12 channel recording

by A. J. Ewins, B.Tech., Research Department, London Transport

In applications where multi-track recording of experimental data is needed, but where several tape speeds and a wide bandwidth are not essential, a conventional f.m. instrumentation is an expensive luxury. This design uses a slightly modified Linsley Hood audio cassette recorder as the heart of a multi-track digital tape recorder. It can handle 2, 4, 6 or 12 channels with bandwidths of up to 42Hz and with zero wow and flutter.

In the field of electronic data collection and storage, that is the recording of signals from various electrical transducers with bandwidths of from zero to several kilohertz, the f.m. instrumentation tape recorder has played an almost unequalled role for many years. Perhaps because the market for such machines is small, they have become very complex, possibly in an attempt to answer every user's needs in just one design. The result of this is that they are very expensive. Typically, a 14-track machine using one-inch magnetic tape and operating with a range of six tape speeds can cost over £20,000. Reels of one-inch tape are also very expensive! Less-complex f.m. tape recorders are available that use quarter-inch tape, operate with a reduced range of tape speeds and with a reduced number of tracks, but which may still cost several thousands of pounds.

One of the main reasons for the f.m. i.t.r.'s expense is the very advanced tape deck used. In f.m. tape recorder designs it is necessary to reduce the wow and flutter content of the tape deck to a minimum to obtain a reasonable signal-to-noise ratio, since any wow and flutter of the recorded signals looks like frequency modulation and hence, when demodulated, represents an unwanted signal, or noise. Another reason for the high cost of f.m. recorders is that, to achieve multi-track recording, very expensive multi-track recording heads must be used. Nevertheless, in spite of these comments, when used to its fullest extent, the multi-track, multi-speed f.m. i.t.r. has yet to be bettered. There are instances, however, when multi-track (or multi-channel) recording is needed, but with reduced bandwidth requirements and without the need for a multi-speed option. To use an f.m. i.t.r. for this purpose, simply because it is the only type of machine available to offer multi-channel recording, is a very expensive solution. It is in an attempt to meet this need that the author has designed the multi-channel, digital tape recorder that is the subject of this article.

Essentially, a multi-channel machine with a bandwidth for each channel in excess of 50Hz was needed. A single-speed machine could be tolerated, provided it was possible to obtain wider bandwidths for each channel by reducing the number of channels available: digital techniques make it simple to do this. Another requirement was that the signal-to-noise ratio for each channel should be as good, or better, than that possible by f.m. recording. Again, digital techniques make it possible to obtain any desired level of signal-to-noise ratio by simply digitizing the analogue signal to the required number of bits. It is also possible, using digital techniques, to completely remove wow and flutter from the recorded data. This is very easily done if the digitized data is played back into a digital computer for analysis and not reconstructed into analogue form. However, the author's requirement was to remove it completely from the reconstructed analogue data outputs.

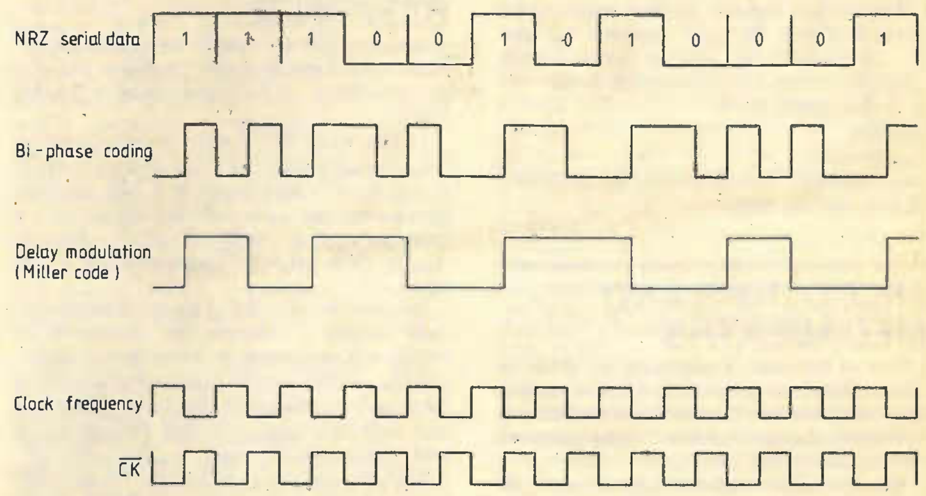


Fig. 1. Data is in non-return-to-zero (NRZ) form at input of recorder. Two possible encoding systems are bi-phase and Miller code. Miller has lower frequency content, allowing twice as much recorded data, but needs more extensive decoding circuitry. Higher data capacity considered to outweigh disadvantage of circuit complexity.

Removing wow and flutter from the recorded data eases the need for a tape deck with superior mechanical qualities, and the possibilities of using a cassette tape recorder were therefore considered: such recorders are cheap, compared with reel-to-reel machines, and tape cassettes offer the cheapest recording medium possible. To remove wow and flutter from the recorded data, the long-term speed stability of the tape must be accurately controlled. Commercially available cassette tape recorders for the hi-fi market are not easy to modify and it was thought that the best solution would be to obtain a recorder in kit form. There appeared to be only one such instrument available—the Hart version of J. Linsley Hood's excellent design. This did indeed prove to be the solution, for it was simple to modify the motor and speed control system of the VFL 910 deck used in the Hart kit. Since the kit costs around £110, a relatively cheap instrumentation tape recorder is therefore feasible.

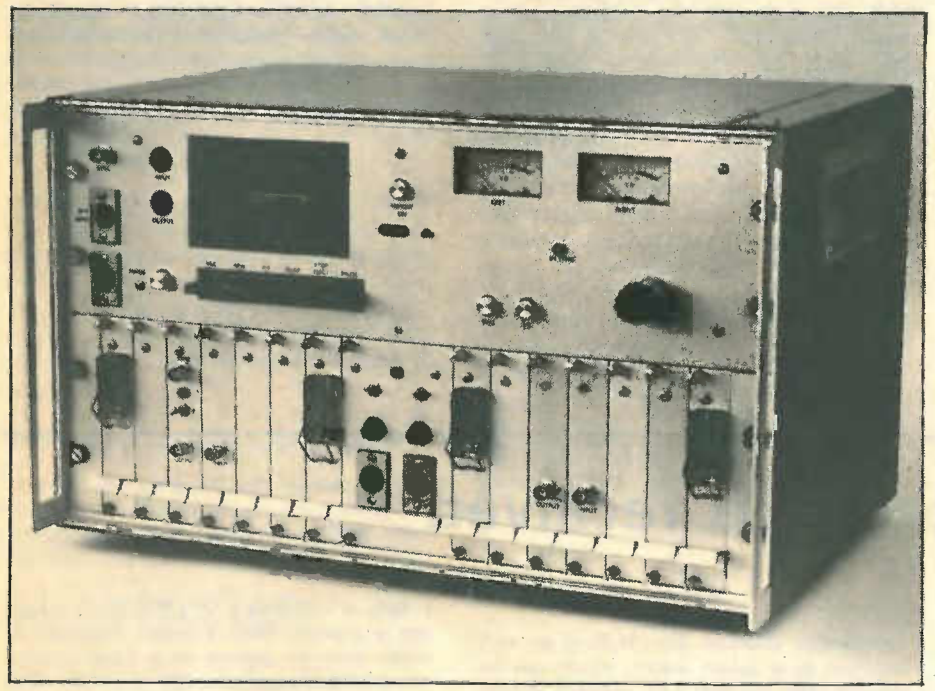
Replacing the front panel of the Hart recorder with one of 19in wide and 3U height (5¼in) made it possible to fit the recorder into a standard 19in instrument case. In the photograph of the complete instrument, the Hart recorder is mounted in a 19in case of 6U height (10½in) with the digital electronics mounted in a rack-ing-frame beneath it.

Specification

Before going on to a detailed description, the specification achieved by the prototype design is presented here so that readers may appreciate its qualities and limi-

tations. Twelve channels of analogue data may be recorded simultaneously, with a bandwidth of 70Hz (allowing for anti-aliasing filters); six channels are recorded on each track of the 'stereo' cassette recorder. Six, four or two channels can be used, with consequent increased bandwidths of 140Hz, 210Hz and 420Hz, respectively. Recorded data is reconstructed into analogue form on playback, with a signal-to-noise ratio of the order of 60dB, achieved by digitizing the analogue data into 10-bits. (Data words of 12 bits length are used, 2 bits being allowed for parity checks.) Wow and flutter content of the replayed analogue signals is zero.

An important parameter of any instrumentation tape recorder is its response to imperfections in the quality of the tape and also to vibration. Many such i.t.rs are used in the transport industries and are therefore subject to considerable vibration.



Complete multi-track tape recorder. Top half is modified Linsley Hood cassette deck.

The effect of both poor-quality tape and excessive vibration is to produce a momentary signal drop-out, resulting in a 'glitch' in the recorded data, the importance of which depends very much on the type of analysis that is subsequently carried out on the data, and which in some cases can be very embarrassing. Digital recording techniques are very much more sensitive to both these faults and it was therefore with some trepidation that the author embarked on such a design using a relatively cheap cassette-recorder and cassette tapes.

It is possible to eliminate the problem of poor-quality tape, provided the imperfections are not extensive, by distributing the serial data stream across several tracks of the tape and by using advanced error detection and concealment techniques.

It was not practical to attempt such sophistication in a relatively cheap recorder. The first attempt at eliminating the generation of glitches was thus a simple one and consisted of adding two parity bits to the 10 bit data word. In the event of a parity error being detected in the played-back data, the output signal of the particular channel is simply held at the level given by the last correct data word. The author's expectations of this simple error-detection system have been more than satisfactorily realised! using good-quality cassette tapes, such as the Maxell UDXL II, typically less than five glitches have been detected on one channel in 30 minutes of recorded data, with the recorder in its two-channel mode. Nearly all the glitches occurred at the beginning and end of a cassette and if half a minute's recording time were eliminated from each end of the cassette it would appear that most would be removed.

In an attempt to assess the recorder's response to vibration it was picked-up and shaken by hand – to and fro, side to side and back and forth. No glitch in the replayed data was observed. The cassette deck is, however, sensitive to rotation about the capstan's axis. Whilst the author appreciates that this vibration test meets no British Standard, the instrument's response (or lack of it) is better than some i.t.rs within his experience.

Design philogophy

To record the outputs from a number of analogue channels, digitally, on to one track of a tape-recorder they must be multiplexed, converted to digital words, formed into a continuous stream of serial data, and suitably encoded. The first decision to be made was the method by which the serial data stream should be encoded. The need for encoding arises from the fact that it is not possible to simply record the NRZ digital data directly onto tape, because long strings of zeroes or ones would contain a strong low-frequency content. Also, with no changes in the signal level taking place, there is no information being generated from which to recover the clock frequency. Because of the inability of conventional recording techniques to record signals down to zero frequency (20Hz is about the best lower limit of a good direct-recording tape-recorder) and the need to be able to extract the clock frequency from the recorded data, it is essential that the serial data be encoded in such a way that frequent changes occur in the output voltage. However, to maximize the recording density of the tape, these changes should be as infrequent as possible.

Two encoding systems were considered – *bi-phase* encoding and *delay modulation* (also called Miller code). Figure 1 shows a serial stream of digital, non-return to zero (NRZ) data and the resulting outputs from the two encoding systems. Bi-phase encoding results in a positive signal transition at the centre of 1 cells and a negative transition at the centre of 0 cells. Miller code is simply bi-phase encoding divided by 2 and results in a signal transition at the centre of a 1 cell and between adjacent 0 cells: the direction of a transition in Miller code is unimportant. In bi-phase encoding the highest fundamental frequency present in the encoded data (ignoring harmonics) is that of the clock oscillator. In Miller code, it is half that of the encoding clock oscillator. The lower frequency content of the Miller-coded data, compared with the bi-phase coded data, is the main advantage of Miller code. It means simply that twice as much data may be recorded on tape, within a given bandwidth, using Miller code than by using bi-phase encoding.

Miller code does, however, have disadvantages. It is relatively simple to extract the clock frequency from encoded bi-phase data, and also to decode it. The circuitry required to decode Miller code and to extract the clock frequency is very much more complicated. It is also desirable that the sequence 1, 0, 1 be included in the NRZ serial data stream since, in the absence of 1s or 0s, a string of encoded 0s looks exactly the same as a string of encoded 1s. However, there is a phase difference between encoded 0s and 1s which is only detectable when both are present in the data. The sequence 1,0,1 in the encoded data produces a unique time gap between signal transitions and thus correctly sets the Miller decoder for decoding 1s and 0s. To determine which encoding system should be used, the author had to decide between circuit complexity and high data capacity or relatively simple circuitry with reduced data capacity: the decision in favour of a higher data capacity brought with it the need to design a reliable clock-recovery circuit and Miller decoder.

One of the advantages of digital recording is that the recording process does not need to be linearized by using a bias oscillator, which can considerably simplify the recording circuitry. However, as the Hart cassette recorder used in the design was already complete with its linearizing circuitry, there was no reason to modify it.

Since the frequency response of the recorder extends to 15kHz, it was expected that it would cope with encoded data whose highest fundamental frequency was of the order of 12kHz. Using Miller code this meant that a bit rate of 24kbits/s could be handled, using a clock oscillator of 24kHz. To determine the number of channels that could be multiplexed and recorded on one track of the cassette recorder, a number of requirements needed to be considered, with an ultimate bit rate of 24kbit/s as the objective:

– the minimum required bandwidth of each channel,
– the desired signal-to-noise ratio,
– the number of parity bits per data word,
– the inclusion and length of a synchronization word.

First, the desired bandwidth was a minimum of 50Hz. To allow for anti-aliasing filters, this meant a sampling rate of 4 to 5 times the bandwidth, say 250Hz. Second, a signal-to-noise ratio of 60dB was considered desirable, which could be achieved by digitizing to 10 bits (in f.m. i.t.rs., the signal-to-noise ratio is the ratio of the peak signal level to r.m.s. noise level. If the ration of r.m.s. signal level to the r.m.s. noise level is taken, 10 bits produces a signal-to-noise ratio of only 57dB). With 10 bits for the data an additional 2 bits for parity was thought sufficient, making a total of 12 bits per data word. Eight channels of 12 bit data words, sampled at a frequency of 250 Hz, produces a bit rate of precisely 24kbits/s. However, as mentioned earlier, the sequence 1,0,1 is needed in the data stream so that the Miller-coded data can be decoded: a synchronization word in the data stream at regular intervals allows for this. It also allows correct synchronization of the data on replay and de-multiplexing. To insert a sync, word into the data stream, without interrupting its steady flow, temporary storage buffers are needed for the data. Two clocks also become necessary – one to clock the data into the buffers and a second, faster one – to clock the data and sync. word out. The ratio of these two clocks will be $(x\times12):(x\times12)+y$ where x equals the number of channels and y equals the number of bits in the sync. word. With 6 data words of 12 bits and a sync. word of 8 bits this ratio could be very conveniently made 9:10, i.e. $(6\times12)=72$: $(6\times12)+8=80$. A common crystal-controlled oscillator of 3,2768MHz, divided down by 16×9 and 16×10, gives frequencies for the two clocks of 22,755.5 Hz and 20,480 Hz respectively.

The faster one is referred to as the tape clock, since it runs at the rate at which the data, plus sync. word, is encoded on the tape-recorder. The slower one, running at the rate at which data alone is handled, is the data clock. The tape clock, at 22,755.5 Hz, is very close to the aimed-for bit rate of 24kbits/s and is the closest that can be achieved using standard crystals. For this reason, and because of the convenience of a 9:10 ratio for the data and tape clocks, it was finally decided to record six channels per track of the cassette recorder. With a data clock of 20,480 Hz, 12 bit data words and 6 channels, the sampling rate per channel works out at 284.4 samples/s, which makes possible a bandwidth per channel of around 60 Hz to 70Hz.

To be continued.

Literature received

Twenty four application notes from Datalab present information on the use of their **transient recorders** with a number of computers, calculators, tape punches and graphics terminals. Copies are obtainable from Data Laboratories Ltd, 28 Wates Way, Mitcham, Surrey CR4 4HR.
WW401

1981 Samtec catalogue contains 44 pages of specification and ordering information on a range of **plugs, sockets,** jumpers and terminal strips. Write to Symec Electronics Ltd, Lexden Lodge, Crowborough Hill, Jarvis Brook, Crowborough, Sussex.
WW402

Single, dual and triple-rail **power supplies,** mounted on Eurocards and covering all standard voltages from ± 5V to ± 30V are made by Vero, who describe them in a new brochure, available from Vero Systems, 362A Spring Road, Southampton SO9 5QD.
WW403

Ambit International have changed the name of their **components catalogue** to 'The World Of Radio And Electronics' and intend to produce it quarterly. Items stocked will, they say, complement their magazine. Price is 60p, but the catalogue contains three £1 vouchers. Ambit are at 200, North Service Road, Brentwood, Essex CM14 4SG.

Metal-film resistors from Mullard are well described in a colour leaflet, which can be obtained from Mullard Ltd, Mullard House, Torrington Place, London WC1E 7HC.
WW404

Catalogue covering a range of ceramic, chip and mica **capacitors** is available from RBS Capacitors Ltd, Orchard Works, Vencourt Place, Hammersmith, London W6 9LZ.
WW405

Publication from ICI discusses the **cleaning and drying** of metal, glass and plastics components and equipment in the electronics industry, using Arklone W solvent in special plant. Copies from ICI Solvents Marketing Department, ICI Mond Division, PO Box 19, Runcorn, Cheshire WA7 4LW.
WW406

Data conversion equipment is the subject of a short catalogue from Micro Networks. It includes brief descriptions of digital-analogue-digital converters, track/hold amplifiers, instrumentation amplifiers and complete systems. The company is represented in the UK by Pascall Electronics Ltd, Hawke House, Green Street, Sunbury-on-Thames, Middlesex TW16 6RA.
WW407

Catalogue of **software** for CP/M-based computers is available from Transam. Programs include those for general office work, business and accounting and scientific operations. Languages include several varieties of Basic and Pascal. TCL Software, 59/61 Theobald's Road, London WC1.
WW408

Catalogue of general **electronic components** from Vako contains descriptions of a wide range of discrete and integrated semiconductors, displays, passive components and hardware, including a large section of loudspeaker drive units. Write on company letterheads to Vako Electronics Ltd, Pass Street, Werneth, Oldham, Greater Manchester OL9 6HZ.

Short catalogue by Burr Brown on a-d-a converters, amplifiers, analogue circuit functions, power supplies and fibre-optic data links. Burr Brown International Ltd, Cassiobury House, 11-19 Station Road, Watford, Herts WD1 1EA.
WW409

Selection of **switches** from Lorlin is described in a new catalogue. Types shown are rotary, lever rotary, lock switches, p.c.b. types, sliders and mains switches. Catalogue from Lorlin Electronic Co. Ltd, Daux Road, Billingshurst, Sussex RH14 9SW.
WW410

The cartridge alignment problem

A new approach

by R. J. Gilson, M.I.Mech.E.

Of the two basic sources of distortion and wear involved in the use of a pivoted tone arm, tracking angle errors and lateral loading on the stylus, the customary approach assumes that the first is all-important. Hence use of overhang and offset figures that give the lowest tracking angle distortion over the playing area of the record. This approach may be fallacious, and a better overall performance may be attained by paying attention to the influence of overhang and offset on lateral loading conditions.

It must be clear to readers of the hi-fi press that there has long been a marked difference of opinion between manufacturers and hi-fi writers regarding the correct approach to arm geometry and alignment problems. The following quotations selected from recent publications are typical of many more:

"Most arms are sub-optimally designed."

"There is about a 50-50 split between those designed for minimum tracking error and those not really designed for anything at all other than high distortion."

"Pickup arms vary wildly in their geometry and few are properly designed."

"Current techniques for cartridge alignment are based on completely false assumptions and achieve . . . not alignment but misalignment."

"At present the importance of accurate arm alignment is highly under-estimated."

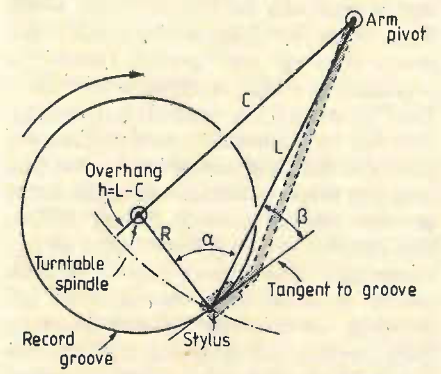
"If the arm geometry is wrong (sic) it can only be due either to cussedness or plain ignorance on the part of the designer."

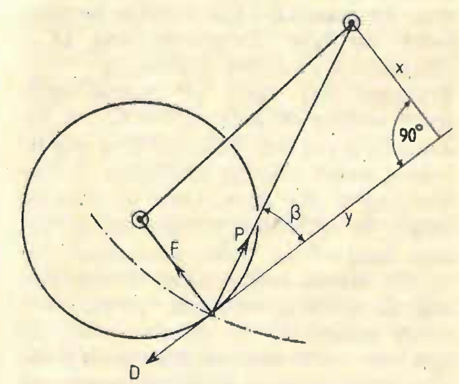
Strong words indeed! The most common ground of condemnation is that the amount of stylus overhang and head offset is insufficient to achieve the lowest possible tracking error distortion, over the playing area of a 12" record. The mathematical basis for achieving the lowest possible tracking error distortion has been examined by a number of people prominent amongst whom were Bauer and Baerwald more than three decades ago. These approaches have been well publicised and it seems to be the essence of the pundits' criticism that manufacturers are too ignorant, obtuse or disinterested to take notice of these well known methods of design. But there are disadvantages in aiming at the lowest possible tracking errors which do not seem to be considered by those who censure the manufacturers so strongly. This article examines both aspects and suggests a way of resolving this longstanding controversy.

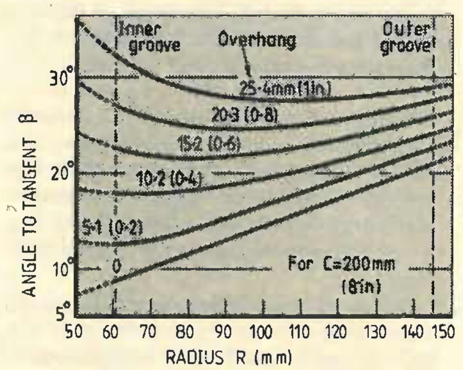
Tracking geometry

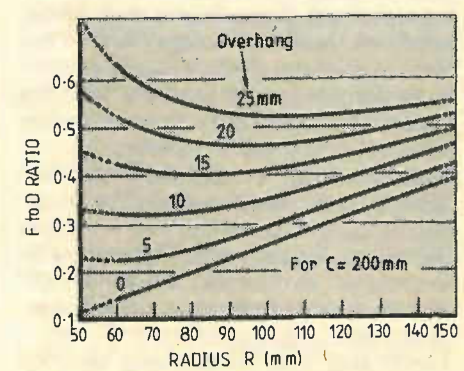
Tracking angle errors are minimized by offsetting the pick-up cartridge at an angle to the arm, as indicated by broken lines in Fig. 1. It is the amount of *variation* in β across the playing area of the record that determines the magnitude of the tracking error, i.e. the discrepancy between β and offset angle. Graph shows the relationship of β to R over the playing area for various values of overhang, and to obtain the minimum possible tracking errors an overhang figure in the region of 15 to 20mm is needed. (This is based on a value of C of 203mm (8in), which seems typical. For other values overhang required will vary inversely as C. This is dealt with more fully later). The formula proposed by Bauer/Baerwald for calculating "optimum" overhang, in the sense of achieving lowest possible tracking errors (see appendix) gives an h_0 of 17.9mm, for $R=146$, $r=60$, $L=221$mm, where R and r are outer and inner groove radii, which agrees with expectations from the graph of Fig. 1.A more recent rule (Stevenson, May, June 1966) gives the same figure of 17.9mm (see appendix). Another widely publicised rule is to set zero tracking angle error at radii of 121 and 66mm. The overhang figure necessary to meet the requirement of zero angular error at any two radii can be calculated from equation 3, and for $C=203$, $R=121$ and $r=66$mm gives $h=18.8$mm, which is in close agreement. Randhawa in *WW* March 1978, proposed overhang and offset figures comparable to those given by Bauer, although if anything slightly higher. The actual figure proposed for an 216mm value of L, is 16.5mm, which is somewhat smaller than the above figures because a smaller value of r has been assumed, i.e. 54mm in place of 60mm.

The next step is to evaluate the corresponding offset angle which will average out the angular errors to best advantage. (This is, of course, provided automatically









(by adopting the two-point zero error method as instanced by formula 3, but in the more general case it is necessary to find the optimum offset angle for any selected value of overhang.) Looking at Fig. 2 again, there are three potential points of maximum angle error, i.e. outer radius, inner radius, and intermediate radius at which β is a minimum. This last, the radius for minimum β, can be calculated from $R_{min}=\sqrt{L^2-C^2}$. For the curve in Fig. 2 for h=20mm, if the offset angle were to be set at 25°, tracking angle errors would be ×2, −0.4 and ×2.5° at inner, R_{min}, and outer grooves respectively. To put this into perspective, notice that distortion due to tracking error is proportional to angular error and inversely to groove radius, so we need to convert the angle errors to degrees per unit of radius i.e. +0.33, −0.06 and +0.17° per cm of radius. Obviously this is not the best that can be done, and the figure of 25° for offset angle needs increasing a little. The best value could be found by trial and error, or calculated from formula 4a, see appendix.

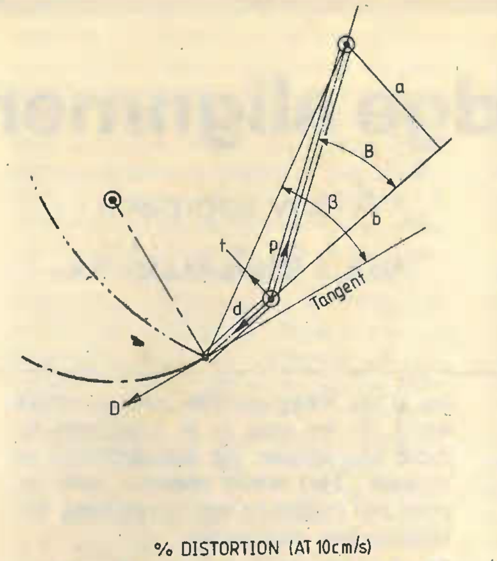
Lateral bias forces

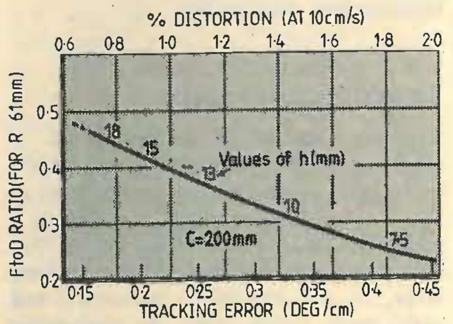
It is an unfortunate fact of life that with a pivoted arm moving in an arc, there must be a side force acting on the stylus tip which becomes greater with increasing overhang. The basic conditions are set out in Fig. 3, where F is the side thrust resulting from the directions of stylus drag D and resisting pull P. Taking moments about the arm pivot, force F can be evaluated in terms of drag D by $F=D\tan\beta$. Values of F/D are plotted in Fig. 4, which shows F can reach 50% of the drag D with 18mm overhang. The normal method of dealing with this side thrust is to apply an opposing outward torque or bias to the arm, but it seems not to be generally appreciated that such compensation is very much of an approximation. To understand this, examine the drag factor carefully. Tangential drag D is composed of a number of elements.

Frictional drag. With 45° groove walls, stylus loading on each wall is 0.7 of the down force, so that frictional drag will be 1.4μw, where μ is the coefficient of friction and w the down force or tracking weight. In addition to straight sliding friction, there will be "deformation drag" due to the elastic deformation of the disc material at the stylus contact point, and it seems reasonable to estimate that the effective coefficient of friction will be somewhere between say 0.1 minimum and 0.3 maximum, depending on stylus shape and finish, and disc surface finish. Thus the total frictional element of drag D could be between about 0.15 and 0.4 of the down force. In principle, this frictional element is independent of groove velocity.

Modulation drag. In addition to the frictional element which applied to an unmodulated groove, there will be further drag due to modulation of the groove. This modulation element can be sub-divided into three related elements, inertial drag, compliance drag and transducer drag.

Inertial drag is due to the energy absorbed





in accelerating the stylus/armature system as it responds to the groove modulation. Acceleration can be extremely high, up to 1000g or more, and inertial effects must be correspondingly great. The energy required to violently waggle the stylus/cantilever/armature system can only be supplied by the turntable motor, and on the assumption that for a given music content the energy requirement is constant, it follows that a constant torque is imposed on the turntable motor, which means that the drag at the stylus point varies inversely with groove radius. (In principle, no energy is required to waggle a mass, as energy absorbed during acceleration will be balanced by an equal amount given out during deceleration. But we are a long way from a perfect mechanism, and in practice the deceleration forces will be dissipated in the form of frictional losses.)

Compliance drag covers the energy absorbed in overcoming the stiffness and damping or hysterisis of the cantilever hinge system, and is presumably greatest at low frequencies when lateral movement of the stylus is at a maximum. It tends to have a constant energy characteristic, giving an inverse relationship to groove velocity.

Transducer drag covers the energy absorbed in converting mechanical energy input into electrical output from the armature/field system. Presumably small compared to inertial and compliance drag, it will also have an inverse relationship to groove velocity.

In the absence of measured figures for stylus drag one can try to make some sensible guesses based on a background of mechanical engineering principles. It seems obvious that a heavily modulated groove will impose more drag than a lightly modulated one, and bearing in mind the high acceleration figures involved, it seems reasonable to assume that modulation drag might reach a peak value of say 30% of the down force. (Modulation drag is not in fact directly influenced by down force, but in practice tracking weight or down force is affected by stylus mass and mechanical impedance, and in this sense modulation drag can be related to minimum tracking weight.)

Adding frictional drag to the assumed modulation drag, we get a total stylus drag varying from a minimum of perhaps 15% of down force up to a peak maximum of perhaps 60% or more. With 18mm overhang giving an F/D ratio of approximately 0.5, side thrust F could be anything between *say 8 and 30% of tracking weight*, Part of this thrust varies inversely with groove radius, increasing from rim to centre, and more importantly it can fluctuate violently with modulation characteristic. It is unrealistic to expect to cancel out the ill-effects of fluctuating side thrust by a fixed arm bias, although it may mitigate things a little. About the best one could achieve, assuming drag D could be accurately assessed, would be to reduce the maximum F by about ⅔, at the cost of increasing the minimum F in roughly the same proportion.

As well as force F increasing the stylus loading on the inner groove wall and reducing the loading on the outer wall, there is a separate force acting to displace the stylus from its free dead-centre position. This is due to the tendency of the tangential drag D to pull the stylus cantilever into line with the arm pivot, and might be termed the reverse toggle effect. The conditions are set out in Fig. 3, which shows that by taking moments about the arm pivot, effective stylus displacement force is $t=d\tan B$. Angle B will be nearly the same as the tracking angle β, and d will be almost the same as D, so t is substantially the same as F, Fig. 3. It follows that any arm bias applied to compensate F will also compensate t to almost the same extent. If we apply an outward acting arm bias of say ⅔ of maximum F, then in lightly modulated grooves the displacement force t will be over-compensated, and there could be a net force t of roughly 15% of the tracking weight acting to rotate the cantilever in a clockwise direction. Conversely, in a peak-modulated groove there will be a partially compensated force t acting to rotate the cantilever in an anticlockwise direction. The amount by which the cantilever/armature system is displaced will depend on the static compliance of the cartridge, and any ill-effects on sound quality will depend on the sensitivity of the transducer system to non-linearity due to displacement from the true dead-centre position.

In addition to any audible effects, the existence of force F must result in added wear on stylus and disc. It may not be realised that the effective increase in stylus loading against the inner groove wall is double the net force F. If a given cartridge

requires $w$gm tracking weight with zero F, as could be the case with a true radial or straight-line arm movement, then if F becomes say 0.2gm the tracking weight will need to be increased to w+0.2gm to prevent mistracking on the groove outer wall, and the lateral loading on the inner wall will be increased by 0.4gm.

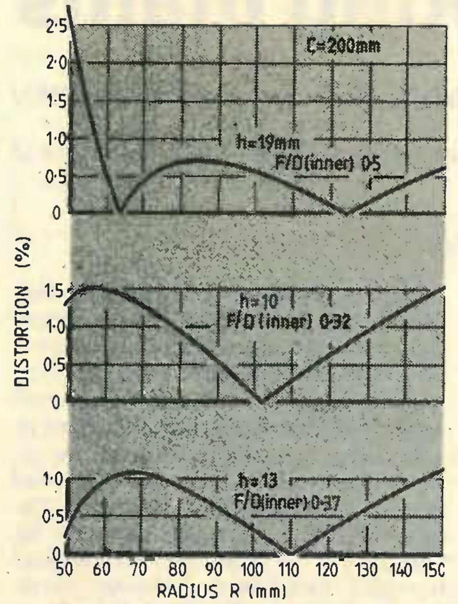
The existence of so many factors in the lateral bias problem, and the difficulty of assessing the optimum balance between conflicting requirements, is doubtless the reason for the widely differing approaches adopted. The Hi-Fi press seem to regard an arm bias of about 10% of tracking weight as "correct", whilst manufacturers adopt anything between 5 and 30%. And at least one major record company recommend setting arm bias on a plain ungrooved section of their test record! There are also differences of opinion on the question of whether bias should increase or decrease over the arm travel. The press seem to regard increasing bias from rim to centre as desirable, whilst some manufacturers adopt reducing bias, presumably on the reasonable assumption that the tendency for modulation drag to increase with reducing groove radius is more than balanced by the tendency for F/D to fall towards the inner grooves when overhang is small.

Optimization

To date, the emphasis in the press seems to be confined to the problem of achieving minimum possible angular errors, without regard to the possible penalty in increasing the lateral forces F and t.

Distortion due to angular error is proportional to angular error per unit of groove radius and is quantified in a formula attributed to Baerwald, $d=4.44e/r$, where d is % tracking error distortion for modulation velocity 10 cm/s, e is tracking error in degrees, and r is groove radius in cm. Using this formula in conjunction with Fig. 2 for values of F/D and formulae 4 for optimum offset angle, one can plot F/D against distortion, as shown in Fig. 4. *The whole controversy is summed up in this curve.* It shows simply that the lowest possible tracking angle errors can be achieved only at the cost of increasing the values of F and t; and conversely forces F and t can only be reduced by accepting increased angular errors. In the absence of published information on the audible effects of the opposing factors, the optimum balance is anybody's guess, but it is hard to see justification for the assumption that the lowest possible angular error must necessarily be the best condition. Tracking distortion is said to be predominantly second harmonic, and the question arises of what level becomes audible. According to one source* 5 to 10% second harmonic distortion is normally undetectable, so it seems highly improbable that a mere 1% distortion would be audible, bearing in mind the overtone content and highly complex waveform of musical modulation. Would the 1½% imposed by the usual overhang of only 10mm adopted by manufacturers become audible? Or the

*"*Pickups, the key to hi-fi*", by *J. Walton* (*Pitman*).



2% necessary to halve the force F at inner grooves? Without a definite answer, it is difficult to formulate an optimum balance between the conflicting factors.

There are two essential factors to investigate, the audible effects of angular error and of lateral force, and it should be a fairly simple matter to undertake this with the aid of a straight-line arm for a reference. The cartridge could be twisted round in say 1° steps up to a maximum of perhaps 7°, and side loading could be applied (perhaps by tilting the deck bodily) in steps of say 5% of tracking weight up to perhaps 50% maximum. If such tests were assessed by listening panels, using a number of top-grade cartridges of differing characteristics, this would surely provide a firm basis for arriving at a generally acceptable balance. The listening tests could be supplemented by wear tests on the stylus, and by measurements of stylus drag.

In thinking about these problems, it is necessary to keep a sense of proportion; tracking error is only one source of distortion and possibly a minor one. Probably the worst source is tracing error, which can easily run into double figures percentage at the inner grooves, particularly with slight stylus wear. Then there is vertical tracking angle error, which is difficult to avoid. Another source is that due to any longitudinal compliance in the stylus/armature system; it is usual to mount the cantilever in an elastomeric grommet or block, and this is not adapted to providing much rigidity in the longitudinal direction.

Bearing all these factors in mind, it seems not unlikely that the manufacturers are doing the right thing in using lower overhang and offset figures than those favoured so strongly by the hi-fi pundits.

It may be helpful to examine the distortion figures across the playing area of the record, for the extreme conditions favoured by one side or the other. Fig. 5 shows the distortion profiles at 10cm/s for three different overhang conditions: 19mm

Appendix

The "two sides and included angle" trig. formula $a^2=b^2+c^2-2bc\cos A$ applied to Fig. 1 gives

$$\sin\beta=\frac{L^2}{}+\frac{R^2-C^2}{2LR}$$

$$\text{or } R=L\sin\beta\pm\sqrt{(L\sin\beta)^2+C^2-L^2} \quad (1)$$

Bauer/Baerwald formula:

$$h_o=\frac{r^2}{L\left[\frac{r}{R}+\left(\frac{R+r}{2R}\right)^2\right]} \quad (2)$$

where R and r are outer and inner groove radii.

Stevenson:

$$h_o=L-\sqrt{L^2-7600}$$

$$\text{or } \sqrt{C^2+7600}-C$$

Overhang for zero angular error at any two radii:

$$h=\sqrt{C^2+Rr}-C \quad (3)$$

where R and r are points of zero angle error.

Offset angle:

$$\frac{\beta_i R_{min}\times\beta_{min}R_i}{R_{min}\times R_i} \quad (4a)$$

where β_i is the angle at inner groove and R_{min} the radius at which β_{min} occurs. For the present example, this works out to β_{opt}=26.1°, which gives errors of +0.15, −0.16 and +0.09° per cm.

This is the best we can do when rounding to the nearest 0.1°, the points of maximum error being at inner grooves and R_{min} of 93mm. With smaller overhang figures, as often used by manufacturers, the points of maximum error will usually be at outer grooves and R_{min}, and the new formula for β_{opt} becomes

$$\frac{\beta_o R_{min}\times\beta_{min}R_o}{R_{min}\times R_o} \quad (4b)$$

where β_0 is angle at radius R_0 (normally 146mm). If the overhang is small enough to place R_{mm} less than the inner groove radius, usually below 10mm. then the formula for β_{opt} becomes

$$\frac{\beta_o R_i\times\beta_{in}R_o}{R_o\times R_i} \quad (4c)$$

as required to give lowest possible distortion, 10mm as favoured by many manufacturers, and my proposal of 13mm the low distortion achieved by the 19mm condition is only maintained if the inner groove radius does not fall below 60mm, and in practice figures down to 58 or even 56mm can occur with 33 rev/min discs, while 45s can go down to about 50mm. At the other extreme, the 10mm overhang gives 2½ times greater distortion at the nominal 60mm inner groove radius, in return for 35% reduction in lateral forces F and t. The proposed 13mm condition seems to make sense; it holds distortion down to a maximum of about 1%, and provides 25% reduction in lateral forces as

Continued on page 64

Tracking mains filter

High-Q active network rejects low frequency interfering signals

by K. Radhakrishna Rao and R. S. Moni, Indian Institute of Technology, Madras

The circuit described is a high-Q, self-tuned band-rejection filter for suppressing low-frequency interfering signals, particularly 50Hz power-line interference. It makes use of four op-amps and a phase-correction scheme and needs no precision components. Because the notch frequency of the filter tracks the frequency of the interference signal, tolerances and temperature coefficients of the frequency-determining passive components do not affect the performance.

Active band-elimination filters have become important in instrumentation used in biomedical and other fields, to eliminate low-frequency interference signals, particularly those due to the 50Hz mains frequency and its harmonics. High-Q band-stop filters are required, but without affecting the physiological data, which carries a wide range of frequency components (normally from zero to 5000Hz). Such a high-Q filter requires excellent performance characteristics. The zero-frequency of the filter has to be accurately determined by the passive components, and it must exactly coincide with the pole-frequency. Such stringent requirements need precision passive components with zero temperature coefficients. But even if the filter satisfies all these conditions there is no guarantee of the frequency stability of the interfering signal. This frequency might fluctuate from its nominal value and result in the feed-through of a significant portion of the interference signal at the output of the filter. This problem can be tackled only by using a self-tuned high-Q band-elimination filter whose pole-frequency is determined by the same components as the zero frequency.

Many of the known active RC band-stop filters[1-3] require precision passive components with zero temperature-coefficients to achieve satisfactory performance. Further, in a few self-tuned notch filters reported earlier[4,5] the notch response is obtained by subtracting from the input signal the interference-frequency components, derived from a switched RC network. The switching frequency is synchronised to the frequency of the actual interference signal through a clock generator, thereby providing a tracking capability. With this scheme the self-tuning range attained is limited and, furthermore, all the stringent conditions with regard to passive components must be fulfilled. Moreover it is quite complicated, because additional circuitry has to be incorporated to suppress the switching-noise generated and to keep both inputs to the subtracting circuit equal in magnitude and phase at all the tracking frequencies. In this article a relatively simple scheme, which does not require zero temperature-coefficient, precision, passive components, is proposed. It uses the four-amplifier circuit shown in Fig. 1, which is a modified Kerwin Huelsman Newcomb biquad[6]. Self-tuning in such an arrangement involves making the filter voltage-tunable[7] by replacing the frequency determining resistor, R in Fig. 1, by a voltage-dependent resistor, R′, shown in Fig. 2, and then locking it to the interference signal V_{i1}, by applying phase corrections[8] as shown in Fig. 3.

An analysis of the circuit is given in the Appendix.

Experimental results

The filter shown in Figs. 1, 2 and 3 was built with dual op-amps and a pair of matched j.f.e.t.s. The phase-correction system was made up of a LM711C dual voltage-comparator, a CA3028A differential amplifier for temperature compensation of the output levels of the comparator, and a low-pass filter for smoothing the output. The control voltage from the phase correction scheme was used to vary the resistance offered by the f.e.t. The filter was tested for self-tuning and frequency response characteristics. The input signal-levels, V_{i1}, and V_{i2}, shown in Fig. 1 were kept low enough (100mV) to facilitate linear operation of the f.e.t.

Fig. 4 shows the filter attenuation for different Q_0s (Q_0=54.67,100), as the frequency of the interfering signal, V_{i1} (self-tuning frequency) is varied. It can be seen that the attenuation decreases slightly as the frequency and Q_0 are increased (see equation (6) in Appendix). Fig 5 shows the frequency response characteristic with the filter self-tuned to the 50Hz mains interference signal, V_{i1}, and the incoming physiological data with interference signal present added as V_{i2}.

Attenuation of the interfering signal and the self-tuning range of the filter are found to be more than adequate in many applications, as, in practice, the drift in power line frequency is much less than the self-tuning capability of the filter. The filter can be used to suppress any undesired frequency component, in any range, by properly choosing capacitor C in Fig. 1 within the limits permitted by the phase correction circuit. This scheme can be extended to suppress the harmonic component of the interference signal also, by adding one more filter of the same kind but self-tuned to the harmonic component to be eliminated. A single phase correction scheme is sufficient to drive both the filters, using a "follow-the-master" principle[9]. In this case, the first filter will have two inputs, V_{i1} and V_{i2}, as discussed earlier and the second filter will have its input taken from the notch output, V_{04} of the first filter. The desired output, devoid of the interference signal and its harmonic component, is obtained from the notch output of the second filter.

In the filter constructed a small amount of jitter is observed at the notch output at 50Hz. This is due to the presence of the 50Hz ripple in V_c, used to control the f.e.t. To get rid of this, the phase correction scheme in Fig. 3 has to be slightly modified, by replacing the 711 dual comparator by another comparator circuit, shown in Fig. 6. It makes use of two single comparators and an Exclusive-OR gate. The output of the gate has a frequency twice that of the input and therefore can be easily smoothed by the succeeding filter stage in the phase correction scheme. However, the price to be paid is a slight decrease in the attenuation at 50Hz, due to the increase in error introduced by the phase correction scheme, if the comparators used are not perfectly matched. With this modification, using a pair of randomly chosen comparators, the attenuation at 50Hz is found to be about 36.5 dB.

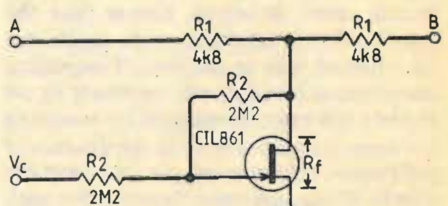


Fig. 2. Voltage-dependent resistor using a f.e.t., to be used for R in Fig. 1. Its value $R'=(2R_1+R_1^2/R_f)$=effective resistance between the terminals A and B. R_f=f.e.t. resistance, determined by the control voltage, V_c. R_2=resistors for equalising the f.e.t. characteristics.

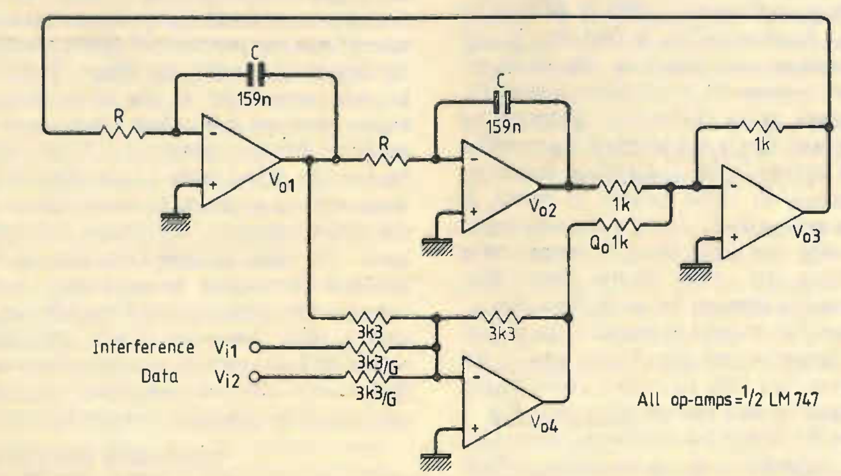


Fig. 1. Modified biquad circuit, providing band-pass (V_{01}), low-pass (V_{02}), high-pass (V_{03}), and band-elimination (V_{04}) functions. V_{i1} is the interference signal to be eliminated and V_{i2} is the physiological data signal containing the interference-component. (ω_0(=1/RC)=ideal pole frequency; Q_0=ideal pole Q; and G=gain of filter.)

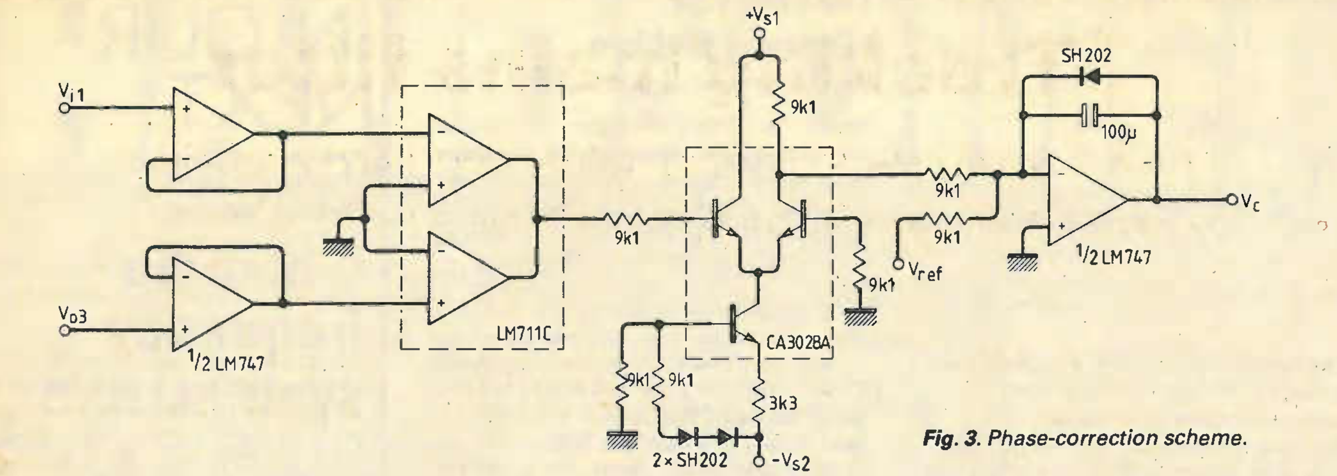


Fig. 3. Phase-correction scheme.

Appendix: circuit analysis

Considering the finite-gain of the op-amps used as $A=1(1/A_0+s/GB)$, where A_0 is the finite d.c. gain and GB is the finite gain-bandwidth product of the op-amp, and assuming all the op-amps to be identical, the transfer-function of the notch filter with G=1 can be derived as:

$$\frac{V_{04}}{V_{i1}}=\frac{\left[\frac{s^2}{\omega_z^2}+\frac{s}{\omega_z Q_z}+1\right]}{\left[\frac{s^2}{\omega_p^2}+\frac{s}{\omega_p Q_p}+1\right]} \qquad (1)$$

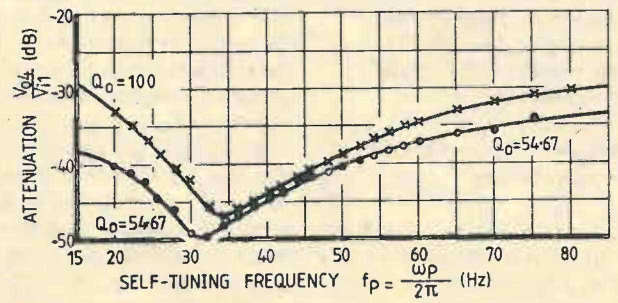


Fig. 4. Attenuation of the band-elimination filter with self-tuning frequency. V_{04}=output of filter; V_{i1}=interference signal.

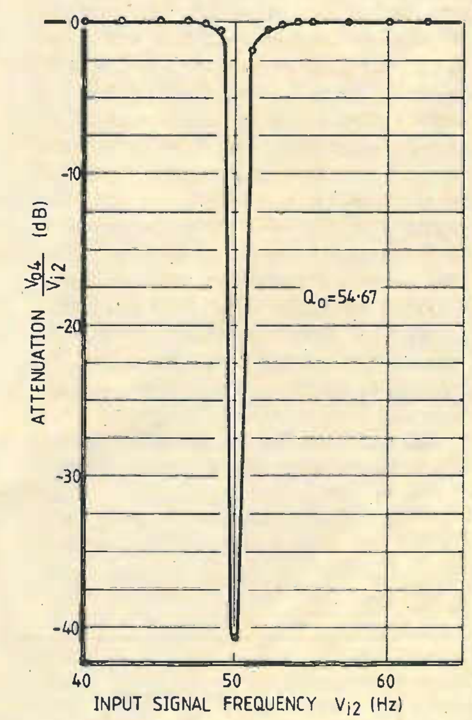
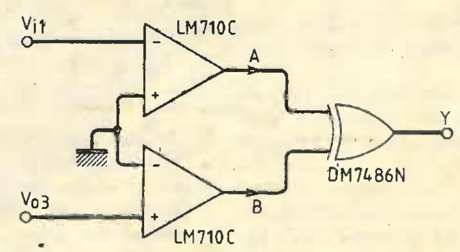


Fig. 5. Attenuation of the filter with frequency of input signal, V_{i2}, the filter being self-tuned to the 50Hz power-line interference signal, V_{i1}. V_{04}=output of filter; V_{i2}=physiological data signal with the interference signal present.

Fig. 6. The comparator circuit with Exclusive OR-ed output Y=A⊕B.



where $$\omega_p = \frac{\omega_o}{\left[1+\frac{\left(4+\frac{1}{Q_o}\right)}{A_o}+\frac{\omega_o\left(2-\frac{1}{Q_o}\right)}{GB}\right]^{1/2}} \quad (2)$$

$$\omega_z = \frac{\omega_o}{\left[1+\frac{\left(4+\frac{1}{Q_o}\right)}{A_o}+\frac{2\omega_o}{GB}\right]^{1/2}} \quad (3)$$

$$Q_p = \frac{Q_o}{\left[1+Q_o\frac{\left(2-\frac{2}{Q_o}\right)}{A_o}-\frac{4\omega_o}{GB}\right]} \quad (4)$$

$$Q_z = \frac{1}{\left[\frac{2}{A_o}-\frac{\left(4\omega_o+\frac{1}{Q_o}\right)}{GB}\right]} \quad (5)$$

Taking into account the tuning error, $\epsilon/2Q_p$, where ϵ is the error due to the phase-detector used in the phase-correction scheme[8], the equation (1) can be simplified for the self-tuned filter as:

$$\frac{V_{04}}{V_{i1}} = Q_p\left[\frac{\epsilon^2}{Q_p{}^2}+\frac{1}{Q_z{}^2}\right]^{1/2} \simeq \left[\epsilon^2+Q_0{}^2\left(\frac{2}{A_0}-\frac{4\omega_0}{GB}\right)^2\right]^{1/2} \quad (6)$$

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The cartridge alignment problem

Continued from page 61

against the 19mm condition: and distortion drops away nicely in the 50-60mm inner groove region. These profiles are based on a figure for C of 200mm, which seems typical. For other values, overhang should vary in inverse proportion, $h = k/C$.

In the case of arms having L as the fixed dimension, this can be transposed as $h=\frac{1}{2}(L-\sqrt{L^2-4k})$. For the high-overhang condition as represented by Bauer/Baerwald/Stephenson k is 3600, assuming an inner radius of 60mm; for the Randhawa proposal (54mm) k is 3,300; for 10mm overhang k is 2000; and for the 13mm condition proposed it is 2600.

It remains to formulate a method of evaluating the optimum setting radius, i.e. the radius at which offset angle is the same as tracking angle β. Calculate β for various values of C and h at the three controlling raddii, i.e. inner grooves, outer grooves and R_{min} as $\sqrt{L^2-C^2}$. Then calculate the optimum offset angle from formula 4. Finally, calculate the radii for zero angle error, from formula 1. Plot these radii against h for various values of C, and against C for various values of h. The resulting curves are practically straight lines over the usable range of C and h, which means that the setting radii have the form of a $y=a+bx$ relationship. The figures obtained are $R_0=79+(hC/84)$ and $r_0=12+(hC/71)$, where R_0 and r_0 are radii for zero angle error. (Strictly speaking, it is undesirable, from the point of view of accuracy, to use two empirical formulae when the product of the two quantities are precisely related (refer to formula 3), and it would be better to evaluate r_0 from the formula $(L^2-C^2/R_0$. For the proposed rule $h=2600/C$, $R_0=110$mm and $r_0=49$mm, for any value of C within the normal range of say 170 to 230mm. The maximum tracking error distortion can be calculated from the empirical expression $d(\%)=210/C$. Offset angle can be calculated from formula 1, or closely approximated by the empirical expression $4380/C$. Using high quality equipment I have been unable to detect any audible difference between the points of maximum tracking error distortion and zero error.

*"Pickups, the key to hi-fi", by J. Walton (Pitman).

Displacement current
Will Mr Lawrence A. Jones, who submitted two articles on displacement current, please write to Martin Eccles, *Wireless World*, Quadrant House, The Quadrant, Sutton, Surrey or ring 01-3500, extension 3589.

IN OUR NEXT ISSUE

C.b. radio frequency synthesizers

Direct and mixer-type frequency synthesizers are described by Dr E. F. da Silva of the Open University. Points for and against each type are mentioned and there is a practical circuit design for a mixing synthesizer to cover the 40 c.b. channels.

Display aid for microprocessors

This is a device designed by Prof. K. Padmanabhan of Madras University which enables a simple oscilloscope to display the values of digitized signals in alphanumeric form, complete with software-generated annotation.

Cartridge alignment gauge

R. J. Gilson presents a simple device which plugs into the stylus position of a pickup cartridge to enable the correct position to be set up more easily than with a protractor.

On Sale October 21

Long-distance television reception

2 – Why tv signals sometimes cover long distances

by Keith Hamer and Garry Smith

This month the authors discuss the theory behind various conditions under which the range of tv transmissions are extended and – what is more important to the prospective DX-tv enthusiast – how to look for, identify and make the best use of these conditions.

Temporary effects caused by certain weather conditions, meteor showers and even lightning can affect the distance over which tv signals can be received. In this article we will take a closer look at some of the conditions briefly discussed in the last article and some others not previously mentioned. We hope that experienced DX-tv enthusiasts will bear with us for the benefit of newcomers to the hobby as we intend to cover news and development in the field in subsequent articles. For readers who missed the first article, DX-tv is an abbreviation for long-distance television reception.

Tropospheric propagation

This is probably the easiest propagation mode for the newcomer to DX-tv to experiment with as, provided one is not interested in receiving sound channels, a standard u.h.f. tv set can be used to pick up signals from the Continent if the aerial is pointed in the right direction.

The troposphere extends from the surface of the earth to around 7600 metres above and within it atmospheric pressures vary in different areas. From time to time, slow-moving areas of above normal pressure can occur (anticyclones). Clear blue skies by day and clear but cold nights are often associated with high-pressure areas, but sometimes a high-pressure area together with a low-pressure zone can exist that leads to conditions normally associated with winter.

Assuming that the weather condition is purely anticyclonic, a noticeable improvement in the strengths of usually weak signals will be experienced. Long-distance signals will be at their best on the u.h.f. bands (Bands IV and V) in the early morning and late evening. If you pick up an unfamiliar programme during this atmospheric condition, the first sign that it may come from overseas is a picture without sound. Table 1 reveals that other European television services have different sound channel spacings to the one we use, system *I*.

A picture without sound does not necessarily mean that the signal you are receiving comes from the Continent but by looking through programme guides or briefly tuning into the British transmitters you can usually make certain by a process of elimination. Many European stations transmit the test card – the easiest means of identification – for a few minutes after close down, in the early morning and sometimes even all through the night.

A cold or occluded front at the boundary of the high-pressure region can increase the range of tv signals even further. In October of 1975, an exceptionally good 'opening' in the UK allowed signals transmitted some 850 miles away to be received. Figure 1, kindly supplied by the Meteorological Office, is a weather chart for that period showing the high-pressure region over central Europe and the associated front, line AA.

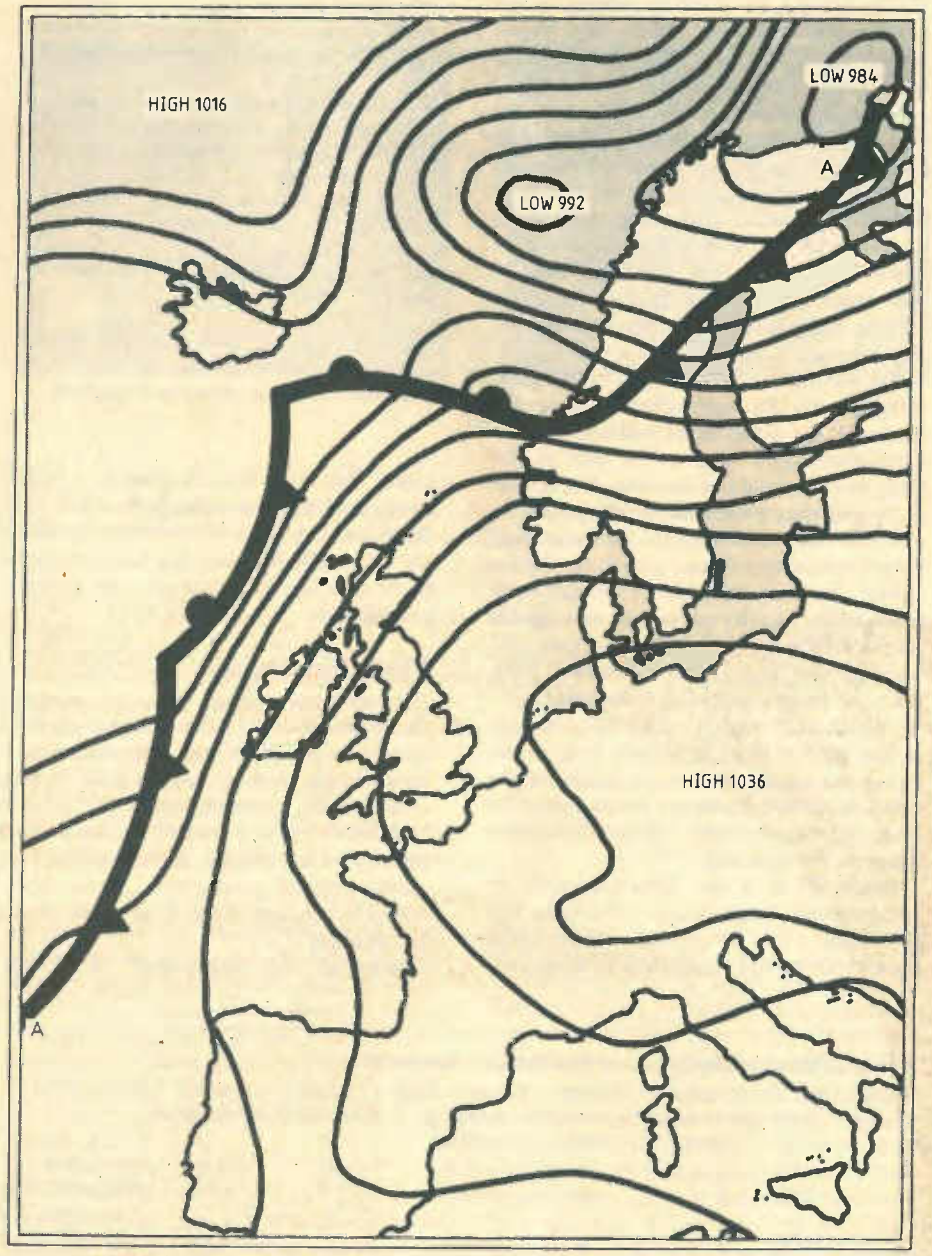


Fig. 1. In October of 1975 the weather conditions shown here produced very good tropospheric reception conditions and signals from some 850 miles away were received in the UK.

Daily weather forecasts on BBC-1 are a means of keeping watch for tropospheric propagation conditions, as the Atlantic chart is always shown for a few moments and approaching high-pressure areas can be monitored. More detailed information can be obtained by taking out a subscription for weather charts from the Meteorological Office in Bracknell.

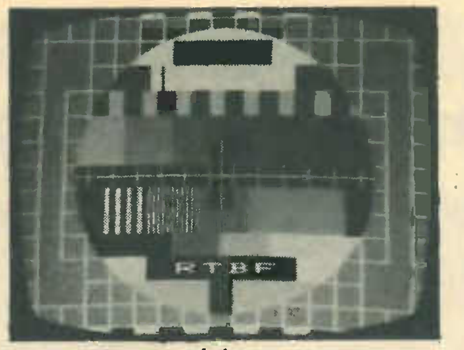
During anticyclonic weather conditions, the earth warms up in the daytime because of lack of clouds and for the same reason the heat built up escapes quickly in the evening. Tropospheric propagation is often greatly enhanced by a frequent result of this heating and cooling process called temperature inversion, where the troposphere forms a waveguide for directing signals above around 70MHz.

Reception under tropospheric-propagation mode conditions tends to be best in a path parallel to the isobars (lines showing where atmospheric pressures between low and high-pressure areas are equal) on weather charts. As a high-pressure area moves away from you reception will be best from transmitters in line with the trailing edge of the area by means of tropospheric ducting.

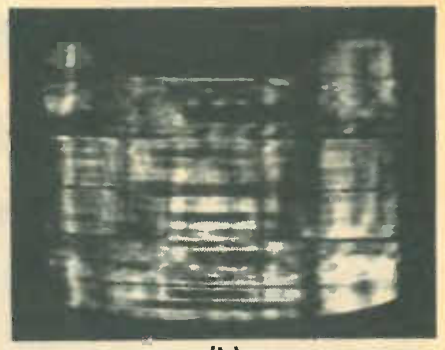
Yet another indication that Continental reception via the troposphere may be enhanced is the presence of widespread fog. Conditions again tend to be best in the early morning and late evening, but fall off as the sun warms the lower troposphere.

Under the conditions so far described, long-distance signals can sometimes be received for several days. Tropospheric propagation has the advantage that signals received by it are not subject to rapid fading and that little phase-distortion takes place, so programmes can sometimes be of 'entertainment quality'. The disadvantage is that irregularities in terrain tend to obstruct the signal path: enthusiasts on the east coast of Britain have a better chance of receiving signals from Europe than those living on the west side.

Bands III to V are most enhanced by advantageous tropospheric conditions but even Band I can be affected. Programmes most received will come from France, Belgium, East and West Germany, Luxembourg and the Netherlands but it should be noted that good tropospheric openings are relatively few and far between compared with the frequency of other forms of propagation.



(a)



(b)

(a) Radiodiffusion-Télévision Belge's (Belgium's French-language service) PM5544 electronic test-pattern received in the UK by tropospheric propagation.

(b) Ionized F2 layer conditions caused reception of this picture in the UK from the USSR's TSS service close to the Chinese border on channel R1 (49.75MHz vision). The image shown features distorsions typical of pictures received by F2 propagation.
(c) The origin of the image shown above, the TSS "0249" test-card.

(d) A PM5544 test pattern on channel E3 (55.25MHz vision) received by F2 propagation. This picture is thought to have come from an Arab Emirate country some 4,500 miles away.

(e) A Voice of Kenya (VOK) news caption received via trans-equatorial skip in September '80.
(f) An official VOK station identification

(g) Ghosting associated with sporadic E reception is shown in this photo of an image from Sveriges-Radio (SR-Sweden). Signals received during sporadic E can, however, be very clear and last for several hours.

(h) A further example of reception possible through sporadic E showing TVE (Television Española) on channel E2 (48.25MHz vision).

Meteor shower

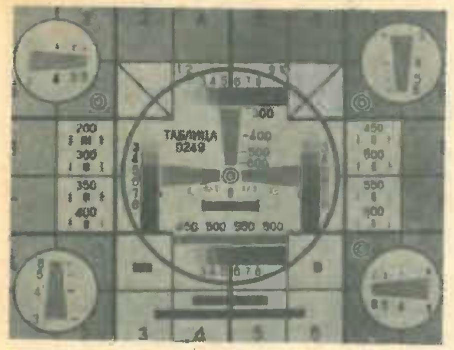
Long-distance signals can be received for short periods when meteor showers cause ionization of the atmosphere's E layer. These meteors, which may be very small indeed, move through the E layer at high velocities and friction causes ionized trails to be left behind. Meteor-shower (or meteor-scatter) propagation, often abbreviated to ms, can occur at any time of the day or night.

Although the occurrence of meteor showers in any 24-hour period is random there are certain times of the year when meteor showers appear more frequently. Details of these annual periods can be found in certain astronomical handbooks. Table 2 gives a rough guide of the best times of the year to look for long-distance signals as far as meteor-shower propagation is concerned.

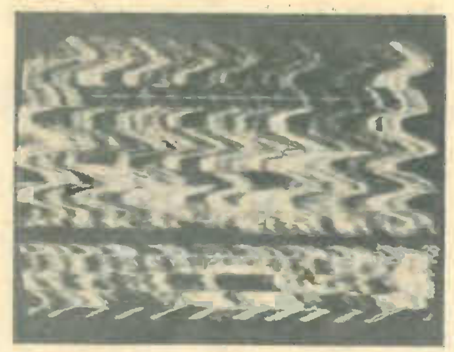
It is not possible to predict the direction from which signals enhanced during meteor showers might come from and, as the effects of a shower usually last for only a few seconds, identification of the transmitter is difficult. Band I signals are most likely to be improved under this mode of propagation but sometimes intense ionization in the E layer can improve reception on Band III channels.

Table 1: World television transmission standards

System code	No. of lines	Channel bandwidth (MHz)	Vision bandwidth (MHz)	Sound/vision spacing (MHz)	Vision modulation	Sound modulation	Regions
A	405	5	3	−3.5	+	a.m.	UK, Eire (v.h.f., to be phased out within the decade)
B	625	7	5	+5.5	−	f.m.	Western Europe, Albania, parts of Africa, Middle East, Australasia (v.h.f.)
C	625	7	5	+5.5	+	a.m.	Luxembourg (v.h.f.)
D	625	8	6	+6.5	−	f.m.	Eastern Europe, Albania, USSR, China (v.h.f.)
E	819	14	10	±11.15	+	a.m.	France (v.h.f., possibly changing to system L on v.h.f.), Monaco (v.h.f., 625-line scanning)
G/H	625	8	5	+5.5	−	f.m.	Western Europe (u.h.f., system G), Belgium, Cyprus, Greece, Israel, Malta, Yugoslavia (u.h.f., system H with 1.25MHz vestigal side-band), Monaco (u.h.f., system G)
I	625	8	5.5	+6	−	f.m.	UK (u.h.f.), Eire (v.h.f./u.h.f.), Rep. of S. Africa (v.h.f./u.h.f.), some Central African countries (v.h.f./u.h.f.), Hong Kong (u.h.f.)
K	625	8	6	+6.5	−	f.m.	Gabon (v.h.f.), Eastern Europe (u.h.f.), French Territories (system K)
L	625	8	6	+6.5	+	a.m.	France, Luxembourg, Monaco (u.h.f.)
M	525	6	4.2	+4.5	−	f.m.	N. and S. America, Caribbean, parts of Pacific, Far East, US Forces (AFRTS), Japan
N	625	6	4.2	+4.5	−	f.m.	Argentina, Bolivia, Paraguay, Uruguay



(c)



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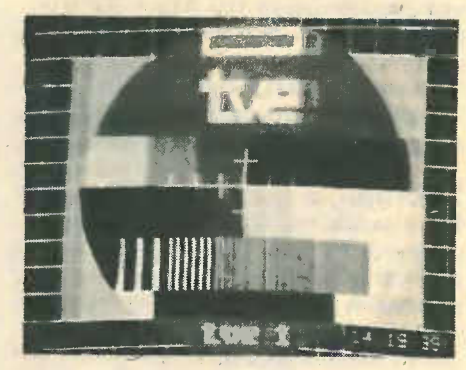


(h)

Table 2: Approximate annual meteor-shower periods. Between the dates given here long-distance reception aided by meteor-shower ionization in the atmosphere is most likely.

Meteor shower name	Beginning	End	Chances of long-distance reception
Quadrantids	Jan. 3	Jan. 4	average
Lyrids	Apr. 19	Apr. 22	moderate
Aquarids	May 1	May 13	good
Perseids	July 27	Aug. 17	best
Orionids	Oct. 15	Oct. 25	moderate
Taurids	Oct. 26	Nov. 16	average
Leonids	Nov. 15	Nov. 17	unpredictable
Geminids	Dec. 9	Dec. 13	good

Lightning flash

During severe thunderstorms lightning causes the atmosphere to become highly charged, thus causing incident-signal reflection. With this form of propagation, both v.h.f. and u.h.f. transmissions may be enhanced. For optimum results the lightning should occur mid-way between the transmitter and receiving site. Conditions may initially be monitored by listening to the radio, since lightning causes interference, especially in the long-wave band.

Auroral reflection

From time to time, particularly around the equinoxes, there are periods of intense solar activity. Solar flares erupt and cause vertical r.f. reflecting sheets within the earth's atmosphere due to magnetic disturbance and ionization of the D, E and F layers. Visual evidence of such disturbances is the Aurora Borealis or "Northern Lights" ("Southern Lights" in the southern hemisphere). In the northern hemisphere, the charged particles emitted by the solar flares spiral towards the earth and are concentrated at the auroral zone. Hence, television signals are received from a northerly direction irrespective of the location of the transmitter. It follows that aerials should be directed northwards.

A rumbling or 'sleigh-bell' effect on sound and horizontal bars on vision are associated with signals propagated by auroral reflections. It is possible to receive trans-Atlantic transmissions during exceptionally high solar-flare activity. Signals received tend to be of poor quality but nevertheless auroral reflection (ar) is an interesting form of propagation. Due to the rotation of the sun, there is a tendency for recurrence of auroral reflection after approximately 27 days. Normally only Bands I and II are affected so far as video reception is concerned but Band III channels may well suffer from severe noise-distortion of the type mentioned above. Usually auroral reflection manifests itself during the mid-afternoon and recurs later in the evening.

F2 propagation

During intense solar activity, the F2 layer becomes ionized and reception from tv transmitters over 2000 miles away is possible. The F layer divides into two belts in the daytime; the F2 layer forms the outer belt at about 200 miles above the earth's surface. During recent solar activity Australian television signals have been received several times in the UK.

F2 layer reception occurs when solar activity is at a maximum in cycles of approximately eleven years. An observation of the sun's surface will indicate whether F2 (and also auroral reflection) reception is likely as magnetic storms in the sun's photosphere, visible as sun spots, are responsible for the ionization in our atmosphere that causes radio waves to be reflected. To avoid damage to the eyes look at the sun through a piece of smoked or filter glass, or project its image onto a piece of white card: never use a telescope or binocular.

Theoretically, F2 reception is best when it is noon at a point mid-way between the transmitter and receiver. In our experience during the present sun-spot-cycle peak, signals from the Far East are noted soon after sunrise. Reception from Australian stations on channel 0 (vision frequency 46.25MHz) has also been reported at around this time. Signals from central Russia will be received towards mid-morning. During December 1979, trans-Atlantic signals were received on many days from shortly after mid-day until late afternoon. African signals, thought to have originated from central countries and Zimbabwe, have also been received, mainly during the equinoxes and after mid-day. Reception from the south was noticeably weaker than that from east and west. An interesting point which several enthusiasts have noted about F2 reception, especially in the early morning, is that signals tend to increase from zero to maximum strength within the space of a few minutes.

Trans-equatorial skip

As sunset approaches, the F1 and F2 layers break up and merge to form a single layer at an altitude of approximately 250 miles. As the F2 layer disintegrates another effect can occur known as trans-equatorial skip (normally abbreviated to te). Reception usually occurs within a limit of 40° north or south of the equator. Signal quality is similar to that experienced with F2 layer propagation, that is, distorted with multiple images. It is often difficult to decipher signals received by these two modes but where there is a possibility of 'double-hop' paths, reception of transmitters at vast distances can be achieved. Normally, only Band I is affected.

Sporadic *E*

Every year between May and September (in the Northern Hemisphere), many part-time DX-tv enthusiasts come out of winter hibernation for the "sporadic E" season (sporadic E is abbreviated to sp.E). As many readers will know, short-wave radio communication is possible due to reflection in the various layers, including the E layer. This particular layer lies approximately 75 miles above the earth's surface and, although it is capable of reflecting short-wave signals, television signals normally pass straight through it. However, during the summer months the E layer becomes highly ionized. If the electron density is sufficiently high, Bands I and II signals will be reflected.

Patches of ionized gases within the E layer move about at great speeds, sometimes approaching 300 mile/h. Several transmissions can be received simultaneously on the same channel, the stronger and more stable signal being accompanied by one or more 'floaters'. But signal bandwidth can be severely restricted and sometimes strong video will be present without sound and chroma signals. We have noticed a tendency for the lower Band I channels to suffer more from this peculiarity than channels above around 60MHz.

As the name suggests, sp.E reception is very sporadic and can occur at any time of the year either day or night, although conditions are less favourable outside the main season. Sp.E cannot be relied upon for entertainment-quality signals and the countries likely to be received cannot be predicted. Reception via sp.E in Band II tends to be more stable and resembles that enhanced by tropospheric propagation. Signals are normally received within 1,000 miles of the transmitter although double-hop or even multi-hop sp.E is possible. At times during the sp.E season, signals from Zimbabwe (ZTV) have been received, usually in the later afternoon. These were assumed to have been propagated by a combination of trans-equatorial skip and sporadic E as Italian television transmissions were normally present simultaneously.

Depending on the state of the E layer, reception can last from a few minutes to several hours. Extremely low-power transmitters can be received via sp.E and it is possible to receive virtually every national television service operating in Europe. Some Middle East countries can also be received within the UK, notably Jordon (JTV). A survey conducted by us (published in the EBU Technical Review, October 1979) revealed that the USSR television service, TSS, was the most commonly received station for this location. Signals from the USSR could easily be received with good picture quality using nothing more than a length of standard wire for an aerial. So for sp.E signals, the minimum of extra equipment will suffice. For serious DX work, however, an external aerial mast is recommended with facilities for rotating the aerial(s).

Under very favourable sp.E conditions transmissions on Band III may also be received so when reception on Bands I and II is good, make a check on the lower-frequency channels of Band III. For newcomers to DX-tv who are mystified by references to Bands and channels, all will be revealed in the next article when we will be covering channel allocations.

Acknowledgements
The authors would like to thank Mr Fish of the Met. Office for supplying the weather chart and Mr Sturgess for the meteor shower periods shown in Table 2.

A slightly more detailed version of Table 1 will be published in the 1982 WW diary with up-dates provided by the European Broadcasting Union.

Another engineer persecuted in USSR

Following our report on the detention of two electronics workers in the USSR (News, July issue) we have been told of a further case by Dr Yosef Ahs, a hospital anaesthetist who was born in the USSR but now lives in Israel. This is Boris Chernobilsky, aged 37, a Jewish radio and electronics engineer from Moscow. He used to work in a high-security institute, possibly on radar. Like Fridman and Brailovsky (July issue) he applied for a visa to emigrate to Israel but was refused on the grounds of "secrecy". That was in 1975. Since then Chernobilsky, his wife Elena (also a radio engineer) and their two daughters have been constantly harassed by the KGB. In October 1976 he went with a number of other Refuseniks to the offices of the Praesidium of the Supreme Soviet where they hoped to find out why they were being refused visas and for how long they would be refused. Instead of being received, the men were rounded up and taken to a site outside Moscow where they were beaten. Two of them, Dr Ahs and Chernobilsky, were detained while the others were set free. They were held in prison for 22 days for "malicious hooliganism". Dr Ahs was allowed to leave for Israel in 1978. Chernobilsky has not been able to work in his profession, in spite of efforts to obtain employment in the general field of radio, and so has been working as a plumber in order to support his family. The Chernobilskys' flat has been searched and they were threatened with arrest more than once.

On 10th May 1981 a number of Jews set out to an area near Moscow called Opalikha to have a picnic to celebrate Israel's Independence Day. Towards the end of the picnic, militiamen who had been standing nearby told the Jews to move. There was an acrimonious argument involving Chernobilsky. Everyone went home without incident, but several days later Chernobilsky received in the post a summons to report to the police station. As there was no mention of why he was being summoned, he did not report. In early June he disappeared for two days – he had been picked up by the police and held overnight. At the end of the first week in June he returned home after having signed an undertaking that he would not leave Moscow.

A criminal file against Chernobilsky has been compiled under which he is alleged to have violated Article 191-1, "resisting the police". The indictment claims that he was asked to give his name and produce his internal passport in Opalikha but refused to do so. The file was due to be completed by the end of June 1981 and then Chernobilsky was expected to be brought to trial.

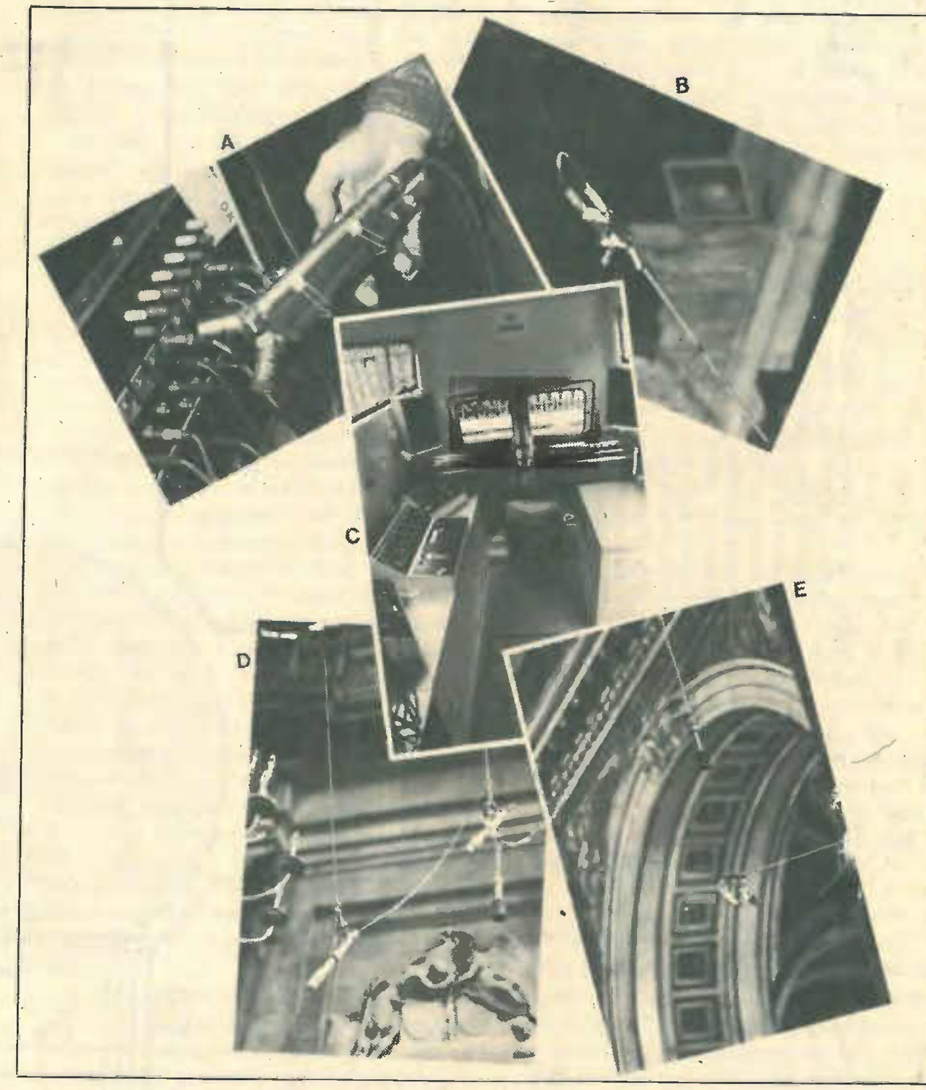
Boris Chernobilsky

Royal Wedding – a sound spectacular

BBC sound broadcasting and recording at St. Paul's.

by John Flewitt B.Eng., MIEE, BBC Engineering Information Department

An estimated 1000 million people were provided with sound from St. Paul's Cathedral during the Royal Wedding. BBC engineers had not only to arrange a variety of mono and stereo sound feeds for broadcasting on radio and television, but also to cope with both stereo and surround-sound for BBC recordings. This article explains how it was done.



A – Portico microphone; coincident pair before insertion into windshield. B – Coincident pair mounted in chancel for cathedral choir. C – Interior of BBC's digital recording van: the mixer and monitoring equipment is on the left and beyond are the two video recorders adapted to record digital sound by the 16-bit p.c.m. unit underneath. D – Main and spare stereo mics for the ochestra are left and top right suspended in north transept; below on right is a sound field microphone. E – One of the microphone positions in the cathedral: at the top a sound field mic and below a stereo pair.

Engineers from BBC Radio Outside Broadcasts rigged 57 microphones to bring the sound of the wedding service to the worldwide audience, including listeners to ILR and viewers of ITV. The sound was fed to the BBC sound control room in St. Paul's Crypt, where a 64-channel mixer produced a 'clean' feed of stereo sound (i.e. sound without the commentaries) and a second 'mixed feed' mixer added the commentaries to produce feed for BBC Radio 4. BBC Television carried out their own sound mixing and other broadcasting organizations either took direct microphone feeds or outputs from one of the mixers.

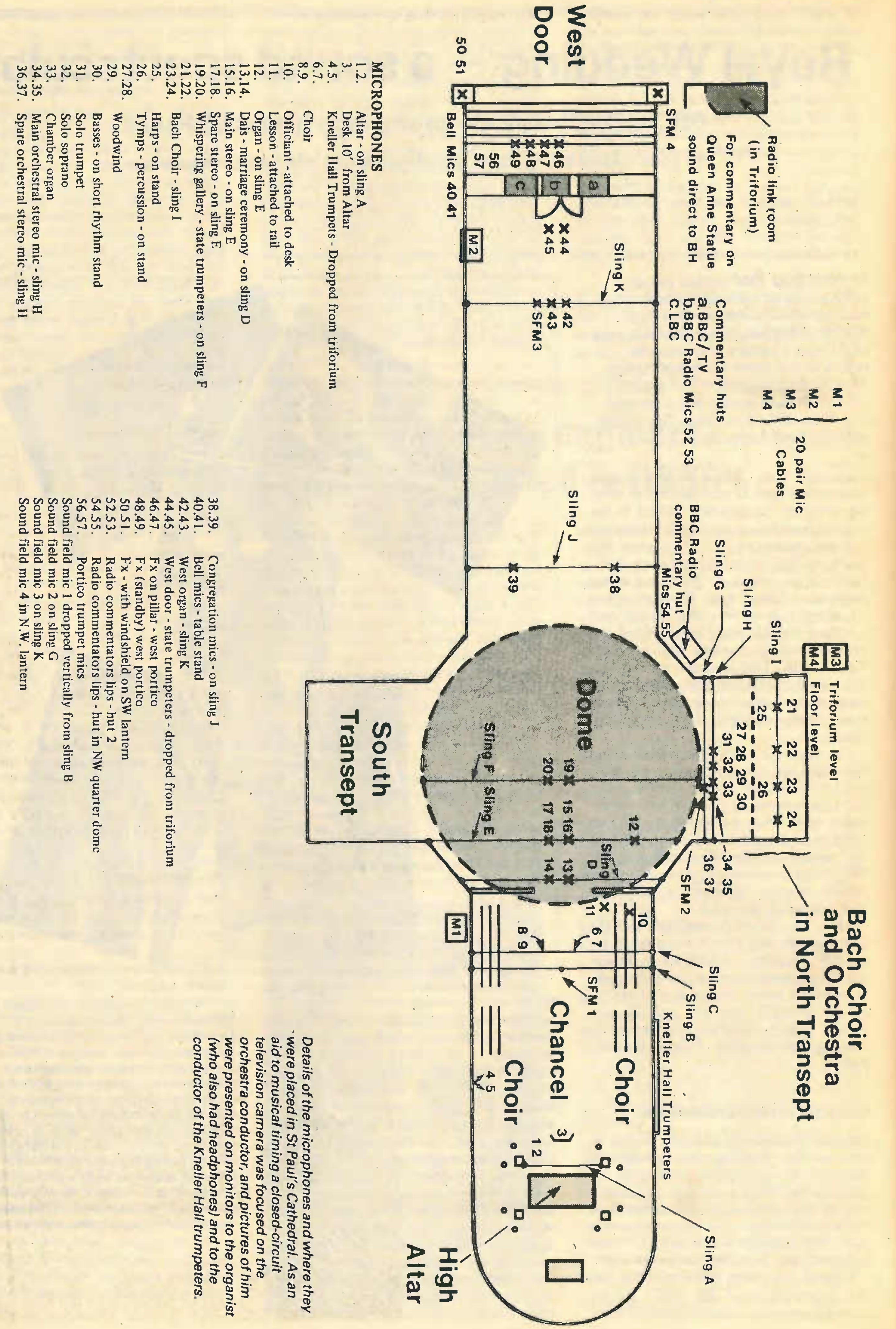
The needs for producing various sound recordings had also to be considered: BBC Enterprises needed a clean feed of sound for their commercial disc and cassette released soon after the wedding and two digital recordings were made, one of clean feed sound and the other including the BBC Radio commentary. And, as a completely separate exercise, a surround-sound recording was made.

All in all, the whole operation had the largest number of stereo o.b. routeings for any BBC broadcast: in addition to the eleven radio commentary positions along the processional route, roving radio links on the day provided interviews with the public and sounds of street celebrations to create a wide spectrum of sound for BBC Radio.

Microphone installation

Detailed engineering planning began as soon as the wedding was announced. Much of the microphone placing in the cathedral was based on past experience but, on this occasion, the use of the Bach Choir and the large orchestra positioned in the north transept was something more ambitious than anything done previously.

Planning the sound in the cathedral was the responsibility of the BBC's Senior Sound Supervisor, Harold Kutscherauer. He arranged coverage around twenty stereo capacitor microphones (mostly coincident pairs), eleven of which were mounted on slings and others suspended on strengthened cables from the 70 ft high triforium gallery of the cathedral. The main internal 'sound stages' to be covered were the dais and the altar for the marriage ceremony itself, the Cathedral Choir and Kneller Hall trumpeters in the chancel, the State Trumpeters in the Whispering Gallery and by the west door, the orchestra and Bach Choir in the north transept, the organ speaking in the north-east quarter-dome and above the west door, and the cathedral bells. An external stereo pair was suspended from the west portico to catch the west door trumpeters immediately below, sounding their fanfare on the arrival



Details of the microphones and where they were placed in St Paul's Cathedral. As an aid to musical timing a closed-circuit television camera was focused on the orchestra conductor, and pictures of him were presented on monitors to the organist (who also had headphones) and to the conductor of the Kneller Hall trumpeters.

of Lady Diana. The remaining complement consisted of spot microphones for soloists and sections of the orchestra and choir, and lectern positions for the ceremonial.

When it comes to siting microphones in St. Paul's, the problems are more physical than acoustical. The cathedral's good-natured acoustics and the use of 'close-mic' techniques ensure that sound levels rarely rise high enough to excite any troublesome echoes. The three requirements borne in mind when siting for this particular event were:

- primarily, to provide complete coverage, bearing in mind the sound radio presentation. Radio listeners, lacking any visual component, easily become confused when any of the action inadvertently wanders beyond microphone coverage;
- to make the microphones unobtrusive to a television audience without sacrificing sound quality. An example of this was the siting of the Cathedral Choir microphones on either side of the chancel instead of choosing the classical midway positions (the black finish of some of the microphones helped make them less conspicuous);
- to provide tighter control of balance by the use of spot microphones. This gave the television sound mixers the useful option of favouring the sounds of small groups of orchestral performers, for instance, when they were being shown in close-up by the cameras.

Virtually all microphones were capacitor types, used in cardioid configuration, and most routed signals down to the crypt control room on 20-pair cables. Certain key microphones were individually cabled as an extra precaution against a multipair failure.

Control and mixing

In the control room, each microphone's signal was fed firstly to a splitter, one output of which was taken to the clean-feed 64-channel mixer, a second to a 'ceremonial' bay* and, in the case of speech microphones, a further feed was taken to the cathedral's public address system. The outputs of the 'ceremonial' bays provided both direct microphone signals, for BBC Television and Thames Television, for example, and a mixed feed to BBC Broadcasting House for the radio network and other purposes.

For large ceremonies, it is normal practice in BBC Radio for two mixers to be installed where possible. The mixers are used in adjacent but acoustically isolated rooms, as they were on this occasion in the crypt. This isolation enabled the mixer at the 'clean feed' desk to concentrate more fully on balancing the ceremonial. The 'clean feed' desk output was then fed to the 'mixed feed' position where the operator mixed the commentators' microphones using cues from talkback.

* 'Ceremonial' bay is BBC parlance for a type of microphone distribution amplifier used chiefly on ceremonial occasions – hence its name. Each bay will handle nine microphone inputs and each input has two buffered outputs.

Recording the wedding

This sonically grand occasion also gave the impetus to make two extra forms of sound recording, over and above the standard analogue ones.

In the first instance, two experimental digital recordings were made, one of clean-feed sound carried out in the BBC's digital recording van parked in the Cathedral Churchyard, the other of mixed-feed sound, undertaken at Broadcasting House.

The digital van was equipped with twin video recorders with a 16-bit pulse-code-modulation unit plus the normal sound monitoring and mixing facilities. Previous problems with tape drop-outs, more noticeable in digital recording, are now largely overcome by ensuring a dust-free recording area and using only highest-quality, pen-tested recording tape.

Finally, the surround-sound recording project was undertaken by BBC engineers as a technical experiment to aid British industry. Four sound-field microphones of an improved design were specially loaned for the event, three being used internally in the cathedral to cover the chancel, north transept and the nave towards the west door. The fourth was mounted near the cathedral steps in the north-west Lantern. The four component outputs from each microphone were separately assigned to individual tracks of a 24-track recorder without any form of surround-sound coding. Special noise-reduction devices were ruled out by interference from nearby thyristor lighting dimmers and, instead, a higher tape speed of 30 i.p.s. was used to improve signal/noise ratio. A problem then arose with sound linking on tape changeovers, since at this high speed each reel of tape ran for only 30 minutes. This was overcome by arranging changeovers to occur during pauses in the wedding service or, at least, on non-musical items, and further arranging for a standby two-channel recorder to make a linking recording in HJ-coded stereo. These stereo recordings would then suffice in any subsequent System HJ matrixing to bridge the gaps in the multichannel surround-sound recording.

Setting the sound-field microphones was relatively simple: each unit's four encapsulated microphones, combined with a microphone processor, gave the system a unique versatility enabling an extremely wide range of operating modes to be electronically selected in a subsequent re-mixing session. In the cathedral, each sound-field microphone's physical height was set by listening to the output of a unit in omnidirectional mode and fixing the height when the most satisfactory balance was heard. A height in the range 30-50 ft proved about right.

"Mixed-feed" mixer in BBC control room in cathedral crypt. "Clean-feed" input was faded on the operator's left; commentator's left; commentator's microphones were controlled on mixer's right.

Royal success

It was a complex exercise and, with 1000 million people listening for the marriage vows, had to be reliable. How did it perform?

Well, very successfully – it could hardly have been otherwise; but, bearing in mind that much of the ceremony could not be properly rehearsed, the quiet sighs of relief from the engineers at the successful conclusion can be well understood.

The introduction of television and its accompanying lighting into a large, completely 'wired-for-sound' cathedral certainly presented numerous hum problems; for instance. But after the below-par cable screening was tracked down and some cable re-routeing undertaken, the seven miles of microphone cable and the 10-mile long lighting network co-existed successfully, each in its own way making a vital contribution to Britain's and the world's biggest outside broadcast.

Acknowledgement

The author would like to thank the engineers of Radio O.Bs for their assistance, especially Harold Kutscherauer for his diagram of the microphone placings, and the BBC's Director of Engineering for permission to publish.

John Flewitt joined the BBC in 1967 after obtaining a degree in electronics at Sheffield University. He worked initially in television studio maintenance before joining Studio Capital Projects Department. He is now a publicity engineer in Engineering Information Department with special responsibility for technical photography.

Digital storage and analysis of speech

3–Spectral analysis

by Ian H. Witten, M.A., M.Sc., Ph.D., M.I.E.E., University of Calgary

Digital recordings of speech provide a jumping-off point for further processing which can alleviate the difficulty of synthesizing natural sounds by concatenating individually-spoken words. Perhaps the most significant contextual effect which must be taken into account when forming connected speech out of isolated words is pitch. The intonation of an utterance, which is a continually changing pitch, is holistic, in that the utterance contains more information than the sum of its components determined by the individual words alone. Happily, and quite coincidentally, communications engineers in their quest for reduced-bandwidth telephony have invented methods of coding speech that separate the pitch information from that carried by the articulation.

Most speech analysis views speech according to the source-filter model* which aims to separate the effects of the sound source – the vocal cords – from those of the vocal tract filter. The frequency spectrum of the vocal tract filter is of great interest, and the technique of discrete Fourier transformation will be discussed. For many purposes it is better to extract the formant frequencies from the spectrum and use these alone (or in conjunction with their bandwidths) to characterize it. As far as the signal source in the source-filter model is concerned, its most interesting features are pitch and amplitude – the latter being easy to estimate. Hence we go on to look at pitch extraction. Related to this is the problem of deciding whether a segment of speech has voiced or unvoiced excitation, or both.

The channel vocoder

A direct representation of the frequency spectrum of a signal can be obtained by a bank of bandpass filters. This is the basis of the channel vocoder, which was the first device that attempted to take advantage of the source-filter model for speech coding (the word "vocoder" is a contraction of voice coder). The energy in each filter band is estimated by rectification and smoothing, and the resulting approximation to the frequency spectrum is transmitted or stored. The source properties are represented by the type of excitation (voiced or unvoiced), and if voiced, the pitch. It is not necessary to include the overall amplitude of the speech explicitly, because this is conveyed by the energy levels from the separate bandpass filters.

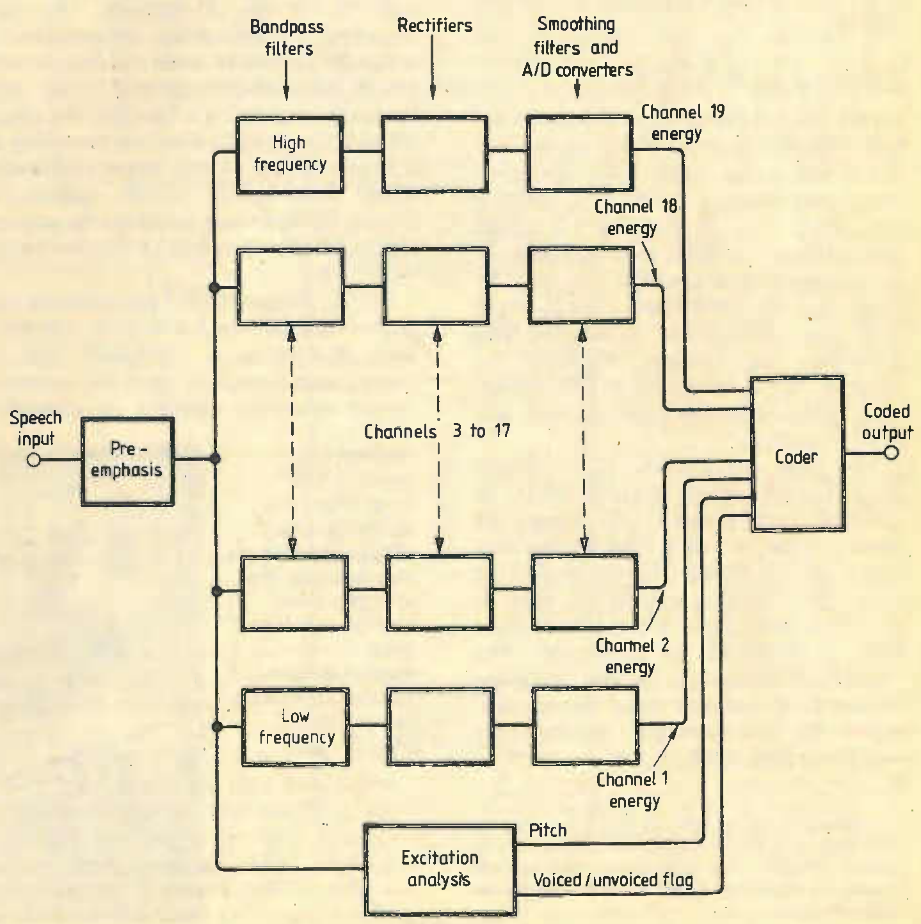
Figure 11 shows the encoding part of a channel vocoder which has been used successfully for many years. We will discuss the block labelled "pre-emphasis" shortly. The shape of the spectrum is estimated by 19 bandpass filters, whose spacing and bandwidth decrease slightly with decreasing frequency to obtain the rather greater resolution that is needed in the lower frequency region, as shown in Table 3. The 3 dB points of adjacent filters are halfway between their centre frequencies, so that there is some overlap between bands. The filter characteristics do not need to have very sharp edges, because the energy in neighbouring bands is fairly highly correlated. Indeed, there is a disadvantage in making them too sharp, because the phase delays associated with sharp cutoff filters induce "smearing" of the spectrum in the time domain. This particular channel vocoder uses second-order Butterworth bandpass filters.

*See the Chatterbox article, cited earlier.

Table 3: Filter specifications for a vocoder analyser

channel number	centre frequency (Hz)	analysis bandwidth (Hz)
1	240	120
2	360	120
3	480	120
4	600	120
5	720	120
6	840	120
7	1000	150
8	1150	150
9	1300	150
10	1450	150
11	1600	150
12	1800	200
13	2000	200
14	2200	200
15	2400	200
16	2700	200
17	3000	300
18	3300	300
19	3750	500

Fig. 11. Block diagram of the encoding side of a channel recoder, which determines and encodes the energy in each of nineteen channels.



For regenerating speech stored in this way, an excitation of unit impulses at the specified pitch period (for voiced sounds) or white noise (for unvoiced sounds) is produced and passed through a bank of bandpass filters similar to the analysis ones. The excitation has a flat spectrum, for regular impulses have harmonics at multiples of the repetition frequency which are all of the same size, and so the spectrum of the output signal is completely determined by the filter bank. The gain of each filter is controlled by the stored magnitude of the spectrum at that frequency.

The frequency spectrum and voicing pitch of speech change at much slower rates than the time waveform. The changes are due to movements of the articulatory organs (tongue, lips, etc.) in the speaker, and so are limited in their speed by physical constraints. A typical rate of production of phonemes is 15 per second, but in fact the spectrum can change quite a lot within a single phoneme (especially a stop sound). Between 10 and 25 msec (100 Hz and 40 Hz) is generally thought to be a satisfactory interval for transmitting or storing the spectrum, to preserve a reasonably faithful representation of the speech. Of course, the entire spectrum, as well as the source characteristics, must be stored at this rate. One channel vocoder uses 48 bits to encode the information. Repeated every 20 msec, this gives a data rate of 2400 bits/s – very considerably less than any of the time-domain encoding techniques.

It needs some care to encode the output of 19 filters, the excitation type, and the pitch into 48 bits of information. Six bits are needed for pitch, logarithmically encoded, and one bit for excitation type. This leaves 41 bits to encode the output of the 19 filters, and a differential technique can be used which transmits just the difference between adjacent channels – for the spectrum does not change abruptly in the frequency domain. Three bits are enough for the absolute level in channel 1, and two bits for each channel-to-channel difference, giving a total of 39 bits for the whole spectrum. The remaining two bits per frame can be reserved for signalling or monitoring purposes.

A 2400 bit/s channel vocoder degrades the speech in a telephone channel quite perceptibly. It is sufficient for interactive communication, where if you do not understand something you can always ask for it to be repeated. It is probably not good enough for most voice response applications. However, the vocoder principle can be used with larger filter banks and much higher bit rates, and still reduce the data rate substantially below that required by log. p.c.m.

Pre-emphasis

It has often been noticed that there is an overall -6 dB/octave trend in speech radiated from the lips, as frequency increases. For vocoders, and indeed for other methods of spectral analysis of speech, it is usually desirable to equalize this by a +6dB/octave lift prior to processing, so that the channel outputs occupy a similar range of levels. On regeneration, the output speech is passed through an inverse filter which provides 6 dB/octave of attenuation.

For a digital system, such pre-emphasis can either be implemented as an analogue circuit which precedes the presampling filter and digitizer, or as a digital operation on the sampled and quantized signal. In the former case, the characteristic is usually flat up to a certain breakpoint, which occurs somewhere between 100 Hz and 1 kHz – the exact position does not seem to be critical – at which point the +6 dB/octave lift begins. Although de-emphasis on output ought to have an exactly inverse characteristic, it is sometimes modified or even eliminated altogether in an attempt approximately to counteract the $\sin(\pi f/f_s)/(\pi f/f_s)$ distortion introduced by the desampling operation, which was discussed in an earlier section. Above half the sampling frequency, the characteristic of the pre-emphasis is irrelevant because any effect will be suppressed by the presampling filter.

The effect of a 6 dB/octave lift can also be achieved digitally, by differencing the input. The operation

$$y(n)=x(n)-ax(n-1)$$

is suitable, where the constant parameter a is usually chosen between 0.9 and 1. The latter value gives straightforward differencing, and this amounts to creating a d.p.c.m. signal as input to the spectral analysis. Figure 12 plots the frequency response of this operation, with a sample frequency of 8 kHz, for two values of the parameter, together with that of a 6 dB/octave lift above 100 Hz. The vertical positions of the plots have been adjusted to give the same gain, 20 dB, at 1 kHz. The difference at 3.4 kHz, the upper end of the telephone spectrum, is just over 2 dB. At frequencies below the breakpoint, in this case 100 Hz, the difference between analogue and digital pre-emphasis can be very great. For a = 0.9 the attenuation at zero frequency is 18 dB below that at 1 kHz, which happens to be close to that of the analogue filter for frequencies below the breakpoint. However, if the break point had been at 1kHz there would have been 20dB difference between the analogue and a=0.9 plots at z.f. And of course, the a=1 characteristic has infinite attenuation at z.f. In practice, however, the exact form of the pre-emphasis does not seem to be at all critical.

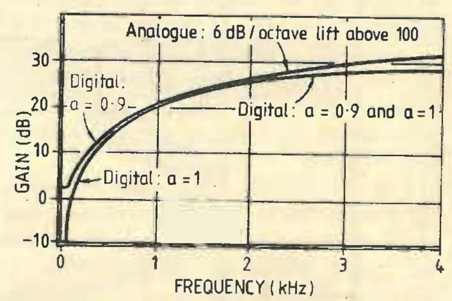


Fig. 12. Frequency response of digital pre-emphasis block shown in Fig. 11. Analogue and digital responses shown.

The above remarks apply to voiced speech. For unvoiced speech there appears to be no real need for pre-emphasis; indeed, it may do harm by reinforcing the already large high-frequency components. There is a case for altering the parameter a according to the excitation mode of the speech: a=1 for voiced excitation and a=0 for unvoiced gives pre-emphasis just when it is needed. This can be achieved by expressing the parameter in terms of the autocorrelation of the incoming signal, as

$$a=\frac{R(1)}{R(0)},$$

where $R(1)$ is the correlation of the signal with itself delayed by one sample, and $R(0)$ is the correlation without delay (that is, the signal variance). This is reasonable intuitively because high sample-to-sample correlation is to be expected in voiced speech, so that $R(1)$ is very nearly as great as $R(0)$ and the ratio becomes 1; whereas little or no sample-to-sample correlation will be present in unvoiced speech, making the ratio close to 0. Such a scheme is reminiscent of a.d.p.c.m. with adaptive prediction.

However, this sophisticated pre-emphasis method does not seem to be worthwhile in practice. Usually the breakpoint in an analogue pre-emphasis filter is chosen to be rather greater than 100Hz to limit the amplification of fricative energy. In fact, one channel vocoder has the breakpoint at 1kHz, limiting the gain to 12dB at 4kHz, two octaves above.

Digital signal analysis

You may be wondering how the frequency response for the digital pre-emphasis filters, displayed in Fig. 12, can be calculated. Suppose a digitized sinusoid is applied as input to the filer.

$$y(n)=x(n)-ax(n-1).$$

A sine wave of frequency f has equation $x(t)=\sin 2\pi ft$, and when sampled at $t=0, T, 2T, \ldots$ (where T is the sampling interval, 125ms for an 8kHz sample rate), this becomes $x(n)=\sin 2\pi fnt$. It is much more convenient to consider a complex exponential input, $e^{j2\pi fnT}$ – the response to a sinusoid can then be derived by taking imaginary parts, if necessary. The output for this input is

$$y(n)=e^{j2\pi fnT}-ae^{j2\pi f(n-1)T}$$
$$=(1-ae^{j2\pi fT})e^{j2\pi fnT},$$

a sinusoid at the same frequency as the input. The factor $1-ae^{-j2\pi fT}$ is complex, with both amplitude and phase components. Thus the output will be a phase-shifted and amplified version of the input. The amplitude response at frequency f is therefore

$$|1-ae^{-j2\pi fT}|=[1+a^2-2a\cos 2\pi fT]^{1/2},$$

or

$$10\log_{10}(1+a^2-2a\cos 2\pi fT)\mathrm{dB}.$$

Normalizing to 20dB at 1kHz, and assuming 8kHz sampling, yields

$$20+10\log_{10}(1+a^2-2a\cos\frac{\pi f}{4000})$$

$$-10\log_{10}(1+a^2-2a-2a\cos\frac{\pi}{4}).$$

With a=0.9 and 1 this gives the graphs of Fig. 12.

Frequency responses for analogue filters are often plotted with a logarithmic frequency scale, as well as a logarithmic amplitude one, to bring out the asymptotes in dB/octave as straight lines. For digital filters, the response is usually drawn on a *linear* frequency axis extending to half the sampling frequency. The response is symmetric about this point.

Analyses like the above are usually expressed in terms of the *z*-transform. Denote the unit delay operation by z^{-1}. The choice of the inverse rather than z itself is of course an arbitrary matter, but the convention has stuck. Then the filter can be characterized by Fig. 13, which signi-

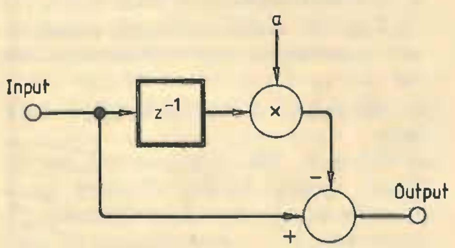


Fig. 13. Digital pre-emphasis filter. Block labelled Z^{-1} is delay operator.

fies that the output is the input minus a delayed and scaled version of itself. The transfer function of the filter is

$$H(z)=1-az^{-1},$$

and we have seen that the effect of the system on a (complex) exponential of frequency f is to multiply it by

$$1-ae^{-j2\pi fT}.$$

To get the frequency response from the transfer function, replace z^{-1} by $e^{-j2\pi fT}$. Amplitude and phase responses can then be found by taking the modulus and angle of the complex frequency response.

If z^{-1} is treated as an *operator*, it is quite in order to summarize the action of the filter by

$$y(n)=x(n)-az^{-1}x(n)=(1-az^{-1})x(n).$$

However, it is usual to derive from the sequence $x(m)$ a *transform* $X(z)$ upon which z^{-1} acts as a *multiplier*. If the transform of $x(n)$ is defined as

$$X(z)=\sum_{n=-\infty}^{\infty} x(n)z^{-n},$$

then on multiplication by z^1 we get a new transform, say $V(z)$:

$$V(z)=z^{-1}X(z)=z^{-1}\sum_{n=-\infty}^{\infty} x(n)z^{-n}$$

$$=\Sigma x(n)z^{-n-1}=\Sigma x(n-1)z^{-n}.$$

$V(z)$ can also be expressed as the transform of a new sequence, say $v(n)$, by

$$V(z)=\sum_{n=-\infty}^{\infty} v(n)z^{-n},$$

from which it becomes apparent that

$$v(n)=x(n-1).$$

Thus $v(n)$ is a delayed version of $x(n)$, and we have accomplished what we set out to do, namely to show that the delay *operator* z^{-1} can be treated as an ordinary *multiplier* in the z-transform domain, where z-transforms are defined as the infinite sums given above.

In terms of z-transforms, the filter can be written

$$Y(z)=(1-az^{-1})X(z),$$

where z^{-1} is now treated as a multiplier. The transfer function of the filter is

$$H(z)=\frac{Y(z)}{X(z)}=1-az^{-1},$$

the ratio of the output to the input transform.

It may seem that little has been gained by inventing this rather abstract notion of transform, simply to change an operator to a multiplier. After all, the equation of the filter is no simpler in the transform domain than it was in the time domain using z^{-1} as an operator. However, we will need to go on to examine more complex filters. Consider, for example, the transfer function

$$H(z)=\frac{1+az^{-1}+bz^{-2}}{1+cz^{-1}+dz^{-2}}.$$

If z^{-1} is treated as an operator, it is not immediately obvious how this transfer function can be realized by a time-domain recurrence relation, However, with z^{-1} as an ordinary multiplier in the transform domain, we can make purely mechanical manipulations with infinite sums to see what the tranfer function means as a recurrence relation.

It is worth noting the similarity between the z-transform in the discrete domain and the Fourier and Laplace transforms in the continuous domains. In fact, the z-transform plays an analogous role in digital signal processing to the Laplace transform in continuous theory, for the delay operator z^{-1} performs a similar service to the differentiation operator s. Recall first the continuous Fourier transform,

$$G(f)=\int_{-\infty}^{\infty} g(t)e^{-j2\pi ft}dt,$$

where f is real, and the Laplace transform,

$$F(s)=\int_{0}^{\infty} f(t)e^{-st}dt,$$

where s is complex. The main difference between these two transforms is that the

range of integration begins at $-\infty$ for the Fourier transform and at 0 for the Laplace. Advocates of the Fourier transform, which typically include people involved with telecommunications, enjoy the freedom from initial conditions which is bestowed by an origin way back in the mists of time. Advocates of Laplace, including most analogue filter theorists, invariably consider systems where all is quiet before $t=0$ – altering the origin of measurement of time to achieve this if necessary – and welcome the opportunity to include initial conditions explicity *without* having to worry about what happens in the mists of time. Although there is a two-sided Laplace transform where the integration begins at $-\infty$, it is not generally used because it causes some convergence complications. Ignoring this difference between the transforms (by considering signals which are zero when $t<0$, the Fourier spectrum can be found from the Laplace transform by writing $s=j2\pi f$; that is, by considering values of s which lie on the imaginary axis.

The z-transform is

$$H(z)=\sum_{n=0}^{\infty} h(n)z^{-n}, \text{ or}$$

$$H(z)=\sum_{n=-\infty}^{\infty} h(n)z^{-n},$$

depending on whether a one-sided or two-sided transform is used. The advantages and disadvantages of one- and two-sided transforms are the same as in the analogue case. Z plays the role of e^{sT}, and so it is not surprising that the response to a (sampled) sinusoid input can be found by setting

$$z=e^{j2\pi fT}$$

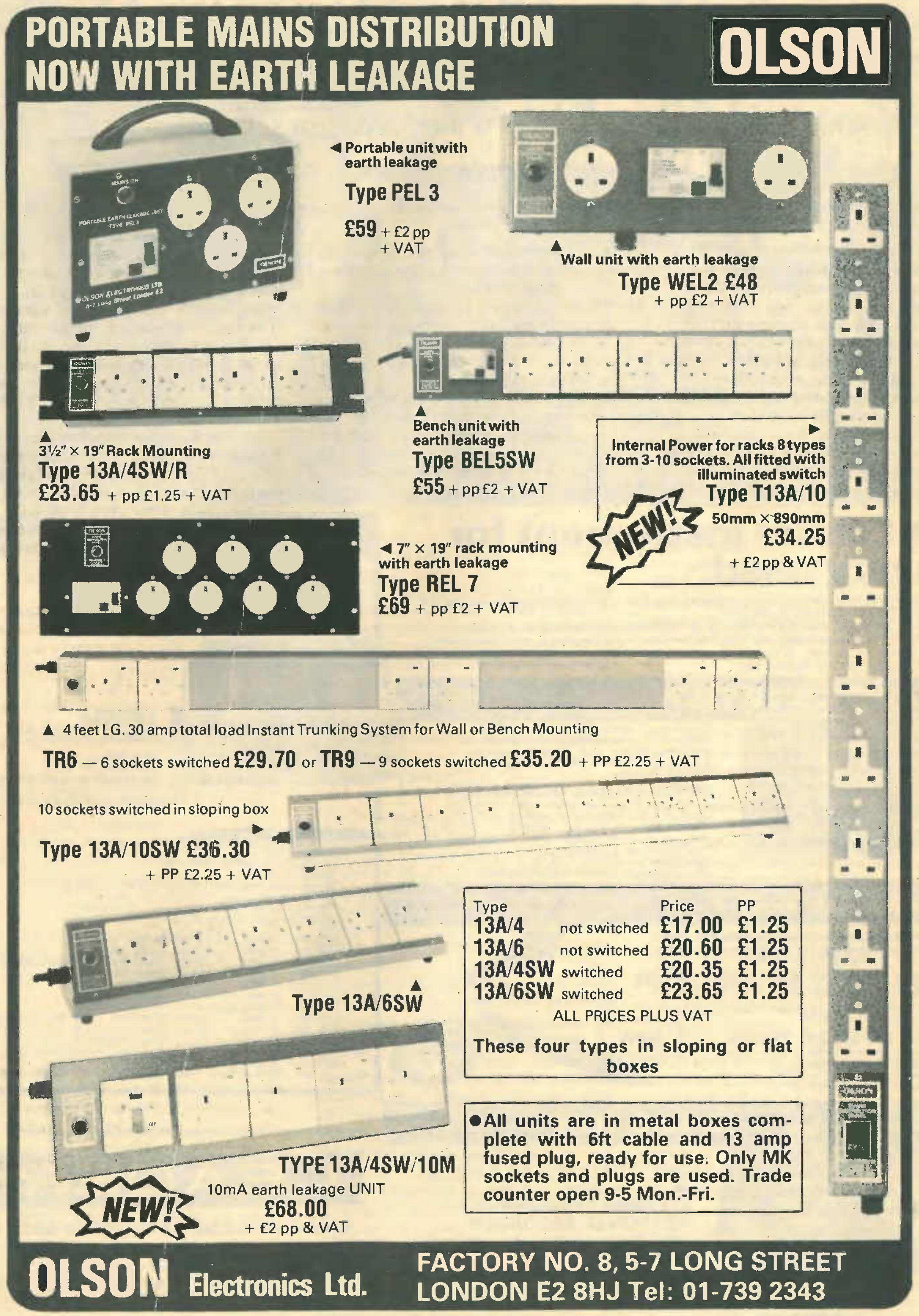
in $H(z)$, as we proved explicitly above for the pre-emphasis filter.

The above relation between z and f means that real-valued frequencies correspond to points where $|z|=1$, that is, the unit circle in the complex z-plane. As you travel anticlockwise around this unit circle, starting from the point $z=1$, the corresponding frequency increases from 0, to $1/2T$ half-way round ($z=-1$), to $1/T$ when you get back to the beginning ($z=1$) again. Frequencies greater than the sampling frequency are aliased back into the sampling band, corresponding to further circuits of $|z|=1$ with frequency going from $1/T$ to $2/T$, $2/T$ to $3/T$, and so on. In fact, this is the circle of Fig. 3 which was used earlier to explain how sampling affects the frequency spectrum!

To be continued

Corrections – Frequency synthesizer for c.b.

Figure 1 of the above article in the September issue contained the following errors for which we apologize: the anode of the variable-capacitance diode connected to the frequency up/-down rail should have been connected to ground, the unmarked capacitor of the v.c.o. circuit is 1nF and the 1μF capacitor at the bottom of the diagram should be 10μF.



A.m. receivers without interference

A method of interference cancellation for double sideband signals

by **Lewis Illingworth**, B. Eng.

Two systems which work with d.s.b. amplitude modulated carriers are described. As the signals are those propagated in long, medium and short wave bands throughout the world, there is a universal application. The systems make use of the fact that a.m. signals have symmetrical sidebands spreading out each side of the carrier frequency so that a modulated carrier has a constant phase, that of the carrier. Interference is not symmetrical even when it spreads through the same frequencies and this asymmetry can be used against it.

Double sideband a.m. transmission has especially desirable characteristics. In the *IRE Proceedings* for December 1956, John P. Costas wrote an article "Synchronous communications — the optimum a.m. system" which explains that double sideband, supressed carrier a.m. signals, similar to broadcast signals but with the carrier removed, are easier to generate than single sideband and permit straightforward synchronous reception with superior performance in the presence of jamming and other interference. What was not mentioned was the additional possibility of cancelling out some of the received interference when synchronous reception is employed. One article which did cover this was an excellent paper in *Wireless World* by P. L. Taylor (July 1977) showing how one overlapping signal can be completely separated from another. Another fascinating paper on co-channel interference was given by J. S. Lothian at the *International Broadcasting Convention*, September 1974.

Our approach to the interference problem does not look for particular types of interfering signal and home in on them, as in P. L. Taylor's system, but to apply a general correction to a received band and accept whatever improvement one piece of circuitry will give. The systems described here completely eliminate interfering signals that are restricted to either side of the carrier. For the more difficult case of a fully modulated signal with carrier, at a slightly different carrier frequency from the wanted signal, the improvement in signal-to-noise ratio drops to about 10dB. As interference becomes progressively complex and finally degenerates to noise, the improvement drops to zero. This performance could be improved but at the expense of some intermodulation between signal and interference.

While such performance is less than ideal, it must be remembered that the figures show improvement over the generally accepted theoretical limit for reception and represent considerable improvement over the performance of an equivalent s.s.b. receiver. This is achieved by conventional circuits and although the quantity of circuitry is not trivial, it is straightforward and works automatically.

Synchronous reception

An unexpected and welcome benefit from the addition of these systems is apparent when used with a synchronous receiver. Such sets, for example the General Electric AN/FRR-48 (XW-1) while operating well on fixed frequencies within the range of carrier phase lock are not at all nice to use when searching for signals. Off frequency unsuppressed carriers are demodulated as ear splitting whistles. For experiments on signals from a receiver, rather than instruments, a synchronous adapter was tied into the 455kHz i.f. of a conventional receiver, leading to adequate and painful listening experience. The new circuits however see such off-tune carriers as interference and eliminate them accordingly. A synchronous receiver now becomes quite nice to use with off-tune stations sounding like single sideband "wasp in the matchbox" accompanied by low background whistle with the mess disappearing as the carrier is tuned within frequency limits.

Development

My work on interference began some four years ago with a system that measured interference amplitude at carrier zero crossings. An initial guess at interference phase was taken to be that of the incoming signal, containing both signal and interference components, and the interference amplitude was estimated using that assumption. Signal amplitude was then deduced and subtracted from the incoming signal-interference composite to provide a better guess at incoming interference phase. Such a recursive system has to operate within tight, almost impossible envelope delay restrictions. Practical tests suggested however that such a system could be developed for general use and so a patent application (Canadian) was made. Further development showed fair operation with simpler forms of interference and the system would even reduce noise levels at very low signal-to-noise rations. However it was abominably unstable and the nonlinear recursive operation nigh on impossible to analyze. Response to it in official circles was negative — (See my letter to *Wireless World*, 15 September, 1977).

The simple system described here is an outcome of a search for a nonrecursive solution. While it is intended as a basis for forays into the realm of reducing levels of complex interference and even noise it stands in its own right as an extremely useful tool in cleaning up radio reception.

Theory

A double sideband modulated carrier with carrier frequency f_c and modulation frequency f_m can be written down as:

$$m\cos2\pi(f_c-f_m)+c\cos2\pi f_c+m\cos2\pi(f_c+f_m)$$

where m and c are amplitudes. This is for a simple sinusoidal modulation and ignores modulation phase, but suffices for this basic analysis.

Demodulating the signal by multiplying by the carrier frequency $\cos2\pi f_c$ gives us:

$$m\cos2\pi f_c\cos2\pi(f_c-f_m)+c\cos^22\pi f_c+m\cos2\pi f_c\cos2\pi f_c(f_c+f_m)$$

$$=\frac{m}{2}[\cos2\pi f_m+\cos2\pi(2f_c-f_m)]+\frac{c}{2}(1+\cos4\pi f_c)+\frac{m}{2}[\cos2\pi f_m+\cos2\pi(f_c+f_m)]$$

$$=m\cos2\pi f_m+\frac{m}{2}[\cos2\pi(2f_c-f_m)+\cos2\pi(2f_c+f_m)]$$

and when filtered leaves a lower side-band at modulation frequencies: $(c/2+m\cos2\pi f_m)$. The carrier product $c/2$ is constant and removed by a.c. coupling to leave the modulation $m\cos2\pi f_m$.

Demodulating the signal by multiplying with the carrier frequency shifted through 90°, $\sin2\pi f_c$, gives us:

$$m\sin2\pi f_c\cos2\pi(f_c-f_m)+c\sin2\pi f_c\cos2\pi f_c+m\sin2\pi f_c\cos2\pi f_c\cos2\pi(f_c+f_m)$$

$$=\frac{m}{2}[\sin2\pi f_m+\sin2\pi(2f_c-f_m)]+\frac{c}{2}\sin4\pi f_c+\frac{m}{2}[-\sin2\pi f_m+\sin\pi(2f_c+f_m)]$$

When this is filtered the $\sin 2\pi f_m$ terms cancel to leave absolutely nothing.

The addition of interference leads to low frequency products when demodulated by both sin and $\cos 2\pi f_c$. Let us add two interfering tones; $U\cos 2\pi(f_c+f_u)$ above the carrier and $L\cos 2\pi(f_c-f_l)$ below the carrier. Demodulating with $\cos 2\pi f_c$ produces:

$$U\cos 2\pi f_c \cos 2\pi(f_c+f_u)+ L\cos 2\pi f_c \cos 2\pi(f_c+f_l)$$

which has low frequency products:

$$\frac{U}{2}\cos 2\pi f_u+\frac{L}{2}\cos 2\pi f_l$$

Demodulating with $\sin 2\pi f_c$ produces:

$$U\sin 2\pi f_c \cos 2\pi(f_c+f_u)+ L\sin 2\pi f_c \cos 2\pi(f_c-f_l)$$

which has low frequency products:

$$\frac{U}{2}\sin 2\pi f_u-\frac{L}{2}\cos 2\pi f_l$$

It is convenient to shift the phase of these by 90° to give the signals:

$$\frac{U}{2}\cos 2\pi f_u-\frac{L}{2}\cos 2\pi f_l$$

To summarize this part, demodulation of the modulated carrier and interfering signals by the carrier, in its natural phase, produce the sum of all modulating and interfering signals; demodulation by a carrier in quadrature phase produces the difference between the interfering signals above and below the carrier, each shifted in phase by 90°.

Take the case of a single interfering tone, say $U\cos 2\pi(f_c+f_u)$, demodulated by a quadrature carrier to $U/2\sin 2\pi f_u$ and shifted 90° to $U/2\cos 2\pi f_u$. This can be easily doubled and subtracted from the in phase demodulated output to leave only signal components. Unfortunately this can only be done if you know that the interference is sitting above the carrier. If it were below, then the subtraction of $-L\cos 2\pi f_l$ will double the interference level in the output!

The key to the new system is to determine the polarity of the interference in the signal demodulated by the 'phase' carrier. To do this audio from the 'phase' demodulator is again modulated using the frequency of the 'quadrature' demodulated audio, shifted through 90°.

For audio derived from a signal with interference in the upper sideband: $m\cos 2\pi f_m+U\cos 2\pi f_u$, modulation by $\cos 2\pi f_u$ produces:

$$mU/2\cos 2\pi(f_m+f_u)-\cos 2\pi(f_m-f_u)+ U(1+\cos 4\pi f_u)$$

This signal is a mess, and most of it is generally unusable. However by filtering at frequencies below the audio range the d.c. value U is left in the clear, the lower intermodulation product $mU/2\cos 2\pi(f_m-f_u)$ may pass the filter but being smaller does affect the polarity of the filter output.

This d.c. value U can be used in two ways:

1. By providing the polarity of the interference it is simple to devise a circuit to either add or subtract the quadrature demodulated interference from the phase demodulated signal-with-interference composite.
2. In real life interference, many frequencies are present and there is no guarantee that the amplitude of the quadrature demodulated interference reflects the required 'phase' demodulated interference amplitude. The envelope of the interference at maximum, when interference is in quadrature with the carrier, may well be a minimum when it is in phase. Here the d.c. value filtered from the second modulation can provide a more accurate amplitude reference. It is re-modulated by the quadrature derived interference frequency to form an interference estimate and is then subtracted from the phase demodulated signal-interference composite.

While it is easy to see how the system works in rejecting a single interfering tone, the many frequencies present in real interference lead to complex analyses that are out of place here. Difficulties arise in estimating the phase of multitone interference, for example what is the instantaneous phase of a mixture of frequencies ranging between 300 and 3,000Hz? This difficulty is overcome by artificially raising signal and interference frequencies before processing so that they appear to be sinusoidal and the phase demodulated signal-with-interference component can be readily modulated by a signal having the instantaneous frequency and phase of the quadrature interference component. Not only does this make modulation possible, it has the added advantage of removing many modulation and intermodulation products to a high frequency where they can be eliminated by a low pass filter. The phase of even a complex difference signal remains a good estimate for the phase of the interference appearing in the 'phase' demodulation and permits excellent operation with complex signals. Following interference cancellation, the correct frequency range has, of course, to be restored.

Systems employing both approaches are described. The performance of each can be modified by changing the bandwidth of the low pass filter in the interference amplitude/polarity circuit. As the frequency response here is increased, the system follows increasingly rapid changes in interference amplitude and frequency, accompanied by increasing intermodulation between signal and interference. The limit occurs when the low pass filter bandwidth equals that of the signal modulation. At this point interfering white noise is attenuated by some 6dB, but there is an associated loss in signal level of about 3dB due to the rapidly changing interference phase continually passing through the carrier and collecting bits of signal as it goes.

Circuit description

The amount of circuitry involved is quite extensive, so rather than getting involved in endless details, this description is limited to the functions of the various parts and only a couple of circuits are shown for clarification. There are two parts to the system: a synchronous receiver adapter that puts received radio signals into a suitable form for processing, and the interference cancelling system itself.

Synchronous receiver adapter

This system, shown in Figure 1, operates from a 455kHz signal taken from a conventional receiver i.f. amplifier. It is modulated by 555kHz for a 100kHz second i.f. frequency chosen to be high enough so that signals are well clear of the audio range and yet not too high for c-m.o.s switches to operate effectively as modulators and inexpensive operational amplifiers such as the LM218 operate without significant delay. The carrier is extracted by a 200Hz bandwidth bandpass filter, and frequency lock achieved by a frequency discriminator which generates a control voltage for the 555kHz heterodyne oscillator.

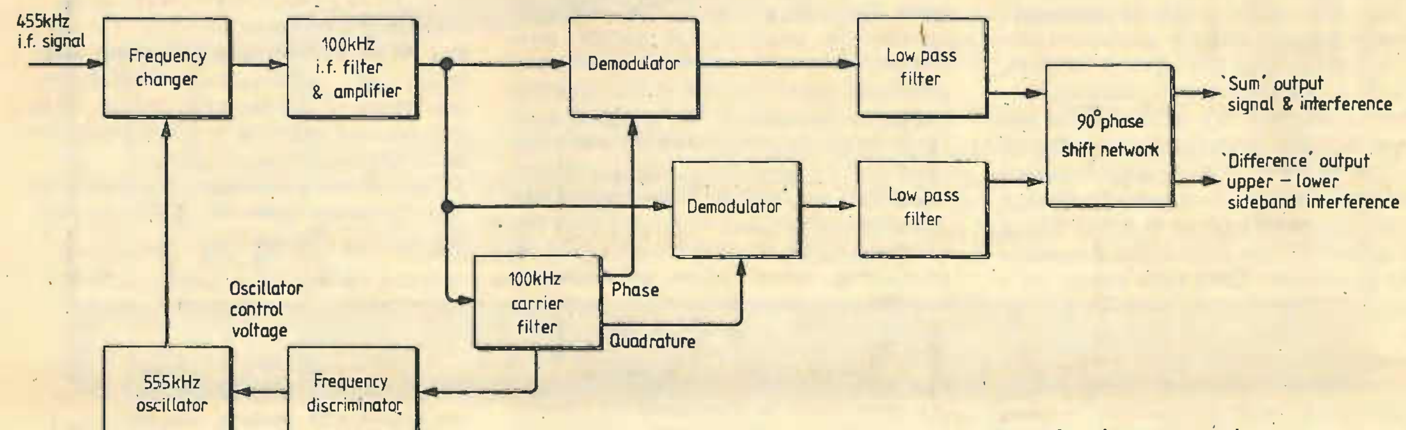


Fig.1. Synchronous receiver

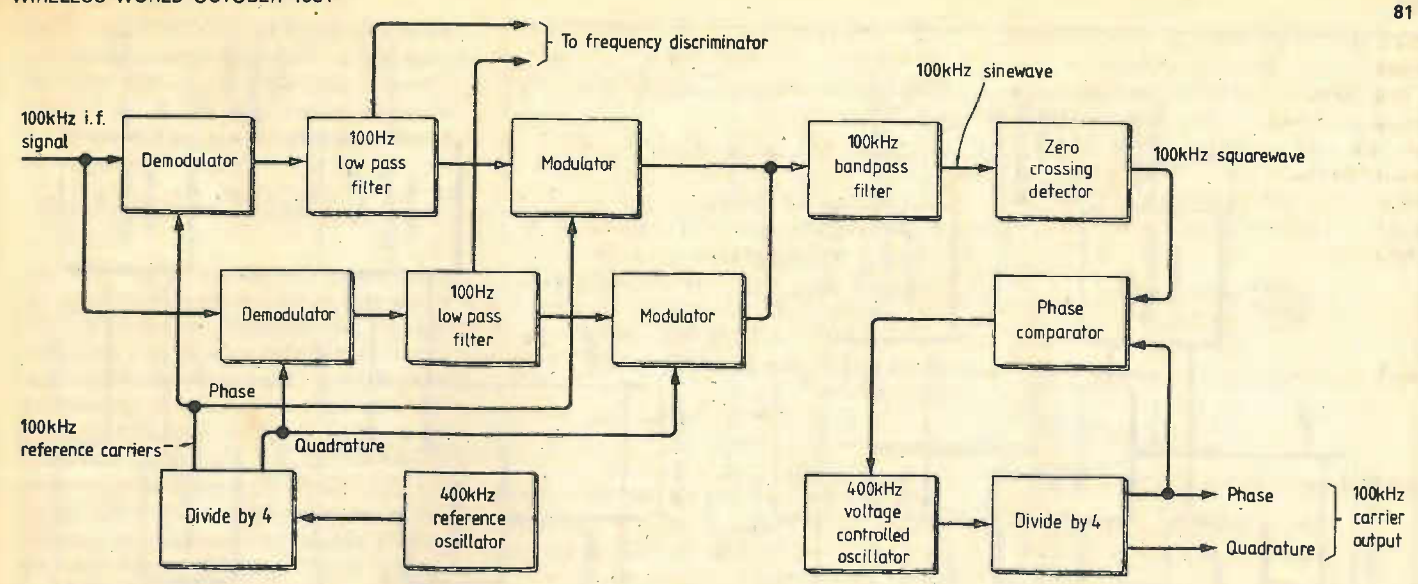


Fig.2. Carrier filter for synchronous receiver

At first sight this system must appear cumbersome in an age of phase locked loops. The reason for it is quite simple: phase locked loops do not operate well under high interference conditions; the loop frequency response must be high enough to permit a lock to be regained after a disturbance, a necessary response that allows interference to get into the loop and so leads to phase jumps in the oscillator output. Narrowing down the response to prevent this happening also prevents phase locking. In the early synchronous receivers phase jitter would not cause much of a problem because a 10° phase error would only reduce the detected amplitude to cos 10°, 0.985, a drop of only 1.5%. With these interference cancelling systems, however, the quadrature value of the signal under the same conditions, sin 10°, 0.174 is considerable, 17.4%, and is seen by the circuitry as interference. A stable narrow band filter minimizes these phase perturbations in the detected carrier.

Two carrier phases are required, one in phase with the incoming signal and the other in quadrature. Each is used to demodulate the 100kHz i.f. signal to produce audio. The 'in phase' demodulation contains the sum of modulation and interfering signals, the 'quadrature' demodulation has the difference between interfering frequencies lying above and below the carrier. A final step adjusts the relative phases of these two outputs by 90° so that signal and interfering components are either in phase or 180° out of phase. It is convenient to refer to the phase corrected phase and quadrature demodulated outputs as 'sum' and 'difference' signals because they are similar to equivalent signals derived from other radio and non-radio sources.

Figure 6 shows a typical circuit of a 90° phase shift network. Two RC networks both attenuate and phase shift the 'sum' and 'difference' audio signals in such a way that attenuation is uniform over the frequency range but there is a constant 90° phase difference produced in signals traversing them.

Figure 2 shows the carrier filter in detail. This has phase and quadrature demodulators and modulators together with los pass active filters to produce extremely stable amplitude and phase characteristics. It is the same type of filter that is used with remarkable success in navigation equipment and is not difficult to make.

The 100kHz i.f. signal is demodulated using 100kHz phase and quadrature reference carriers to produce audio low frequency outputs. Figure 7 shows a typical divider to generate these carriers from a 400kHz stable source. A matched pair of 100Hz cut off low pass active filters eliminate modulation, interference and other higher frequencies. These signals are remodulated, again in phase and quadrature, back to 100kHz and added together. A simple 100kHz conventional filter removes higher modulation products to leave a clean reconstructed carrier. The result is an overall 200Hz bandwidth with stability equal to that of the reference oscillator. One point to watch is the complete cancellation of carrier leaks in the modulators, these will tend to produce jitter in the carrier output.

A 400kHz voltage controlled oscillator drives a divide by 4 circuit to generate 100kHz phase and quadrature carrier outputs. A conventional phase locked loop ties these carriers to the squared up 100kHz filter output. Such a phase lock loop in this position causes no problem because interference has been eliminated from the system.

Figure 3 shows the frequency discriminator used to derive the frequency control voltage for the 555kHz heterodyne oscillator. This is driven by low frequency signals from within the carrier filter. The 'quadrature' demodulated low pass filter output is differentiated to generate a signal that increases with amplitude as the frequency difference increases between the 100kHz received carrier and 100kHz reference. It is in phase with the 'phase' signal when the carrier frequency is higher than the reference and is 180° out of phase when the carrier frequency is lower. The differentiated quadrature signal is added to the phase signal and results in a low frequency a.c. voltage whose amplitude increases when the carrier is above the reference and decreases when it is below. A complimentary a.c. signal is produced by inverting the differentiated 'quadrature' low pass output before adding to the phase signal. Here the amplitude decreases when the carrier is above the reference and increases when below. Both signals are rectified and applied to a differential amplifier to generate a d.c. oscillator control voltage.

The frequency lock is to within a couple of hertz under most operating conditions,

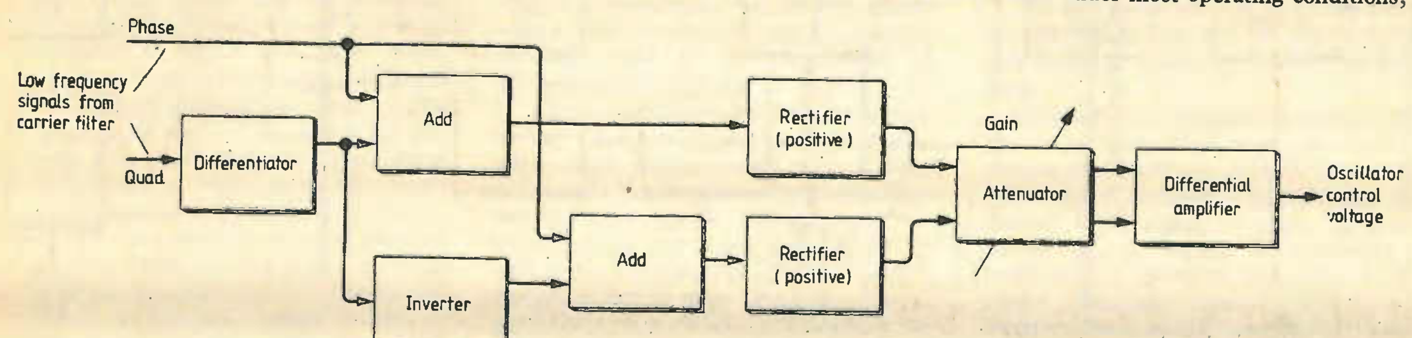


Fig.3. Frequency discriminator for synchronous receiver

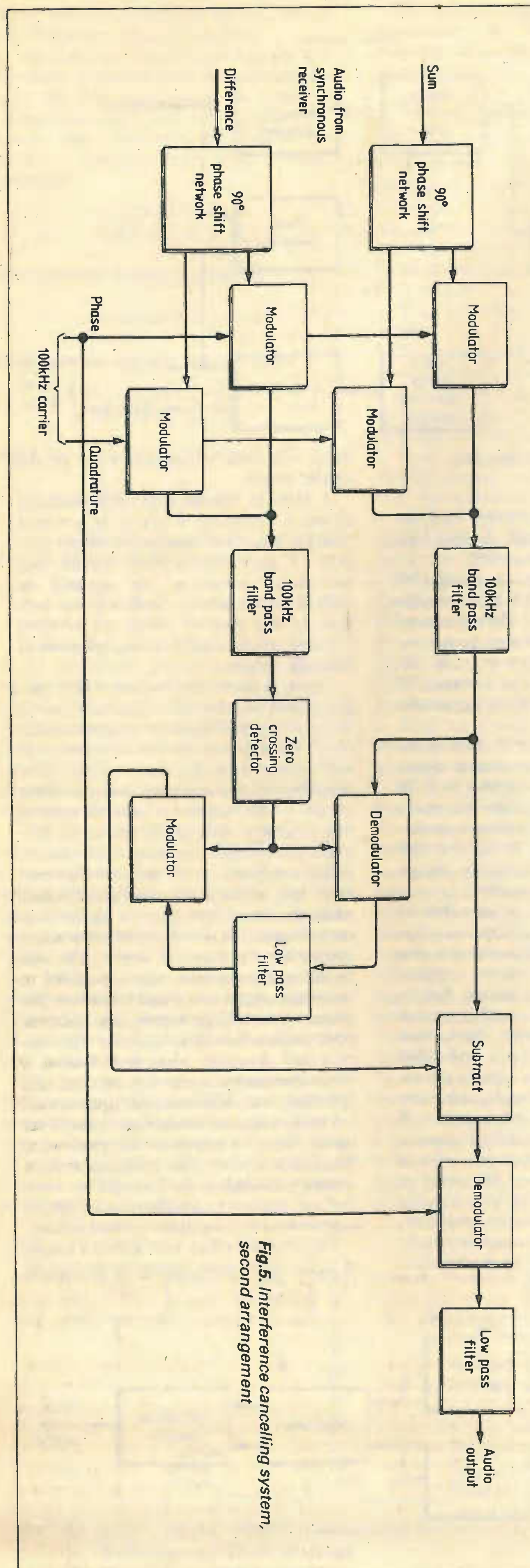


Fig.5. Interference cancelling system, second arrangement

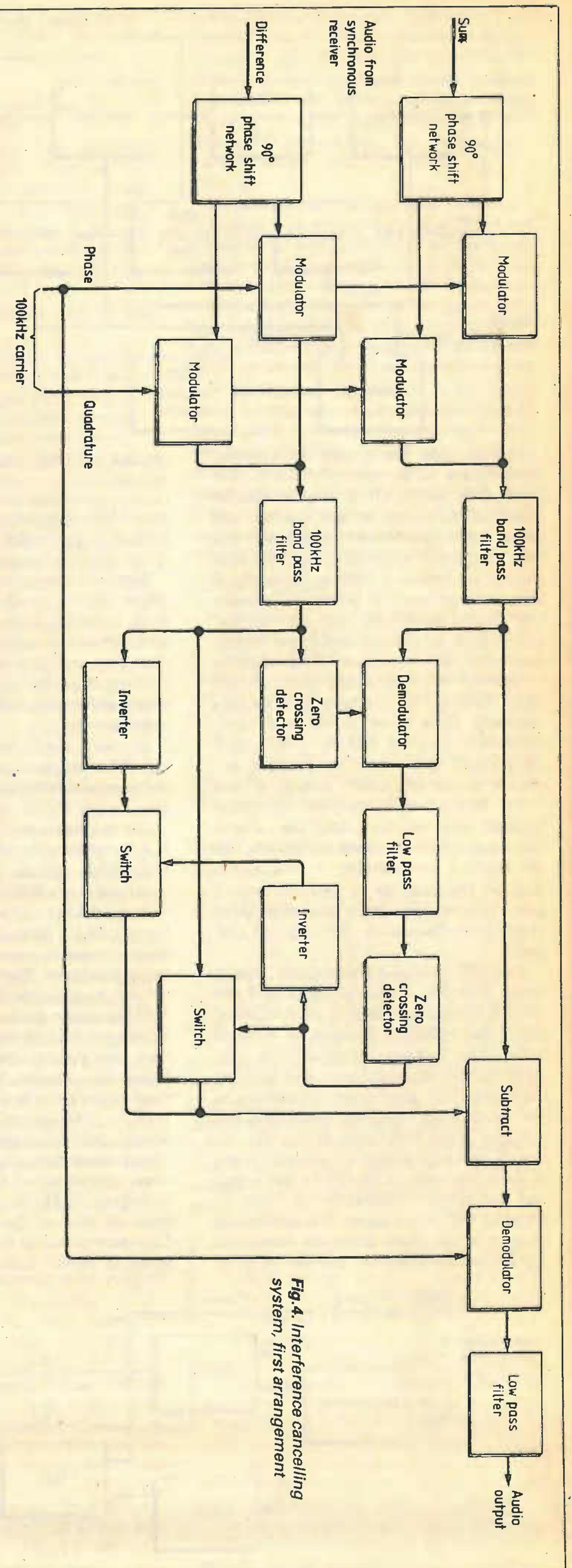


Fig.4. Interference cancelling system, first arrangement

and by virtue of control by the reference oscillator is not subject to drift. It is important that the carrier frequency be maintained close to the reference in order to minimize off tune phase errors introduced by the carrier filters.

Interference cancelling system

The first step in interference cancellation is to raise the frequency up out of the audio range so that everything appears sinusoidal. In the first arrangement, Figure 4, the 'sum' and 'difference' audio outputs from the synchronous receiver are applied to 90° phase shift networks, and modulated by phase and quadrature 100kHz carriers to form upper sidebands of 100kHz. This is a conventional phase-shift method single sideband generation. Each signal is then filtered to remove higher order components. The 'difference' signal is squared up by a zero crossing detector and the output used as a carrier to demodulate the 'sum'. A low pass filter passes only the d.c. component of the demodulation. From this the interference polarity is extracted by a zero crossing detector, converted into a logic level and used to control switches to select either the direct or the inverted 'difference' signal. This puts the interference into the correct polarity where it is subtracted from the 'sum' to produce an improved signal-to-noise ratio. The signal is still at 100kHz, so a final step is demodulation using the 'phase' 100kHz carrier to generate an audio output.

The second system, shown in Figure 5, starts off in a similar manner. Signals are shifted to 100kHz, the 'difference' signal is squared up and used to demodulate the 'sum', and the output low pass filtered. In this system the low pass filter output is taken to represent both the amplitude and polarity of the interference. It is remodulated by the same squared up 'difference' signal to generate reconstituted interference for subtraction from the 'sum' signal-with-interference composite to be followed by demodulation down to audio.

Comments

Many system improvements suggest themselves and the most straightforward are currently being developed and assessed. There is unfortunately a practical limit beyond which processing errors in these analogue systems outweigh advantages gained by increased complexity. We can look forward though to steadily improved signal-to-noise ratios as time and innovation permit.

This article does not pretend to describe an optimum or ultimate method for interference cancellation. The intent is to show that the performance we now accept from our radios is not as good as it could be and that significant improvements can be made by straightforward circuitry. The approach has value both in the improvement of radio reception and also as a basis for further experimentation, particularly into the nature of interference and band limited noise.

The circuits are particularly adaptable to attenuation of interference when something is known about the nature of the interference, for example, military jamming or c.w. interference on frequency shift radio telegraphy.

There is an equivalent system under development for f.m. receivers. This takes the constant f.m. signal amplitude as reference, as opposed to the a.m. signal's constant phase. While showing promise, the technique is not yet sufficiently developed to warrant publication.

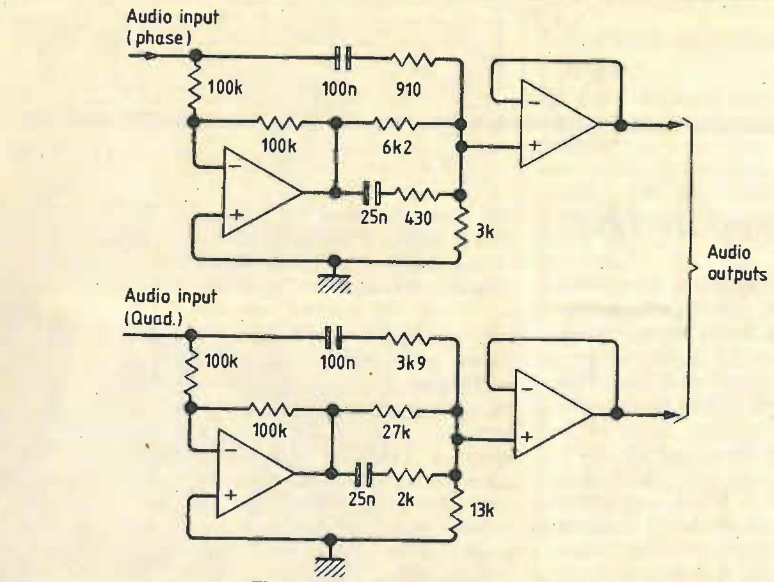


Fig.6. 90° phase shift network

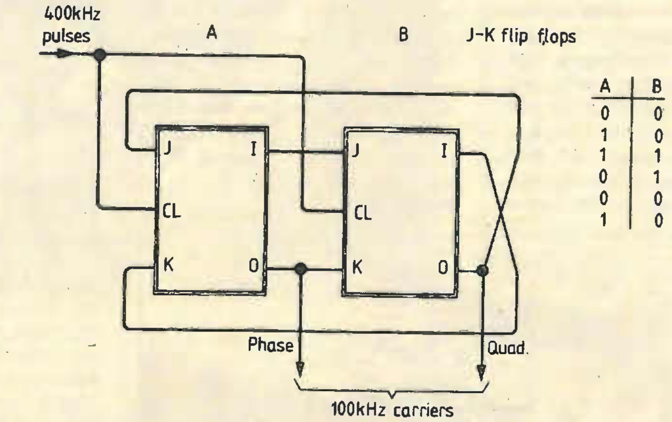


Fig.7. Phase and quadrature carrier generation

New UK group supports semiconductor manufacture

One of the first commercial ventures of the new British Technology Group formed by the merger of NRDC and NEB (News, September issue) is in an advanced field of semiconductor manufacture. This is a joint venture with the UK company Plasma-Therm Ltd, a subsidiary of Plasma-Therm Inc of Kresson, New Jersey, USA, who supply plasma process equipment (see below). The two partners will share the total cost of £170,000 of a two-year programme to develop new process control equipment for sale to European manufacturers of semiconductor devices. The equipment will be based on radio-frequency plasma chemistry techniques, which offer advantages over traditional wet chemical methods used in the fabrication of semiconductor products.

A microprocessor-based monitoring system will be developed to give manufacturers more precise control in dry etching procedures based on radio-frequency plasmas by using optical emission spectroscopy. Also, a new power unit will be developed to complement Plasma-Therm's present range of radio frequency generators and so offer the ability to improve the adhesive qualities of plasma deposited passivation layers which protect the i.cs.

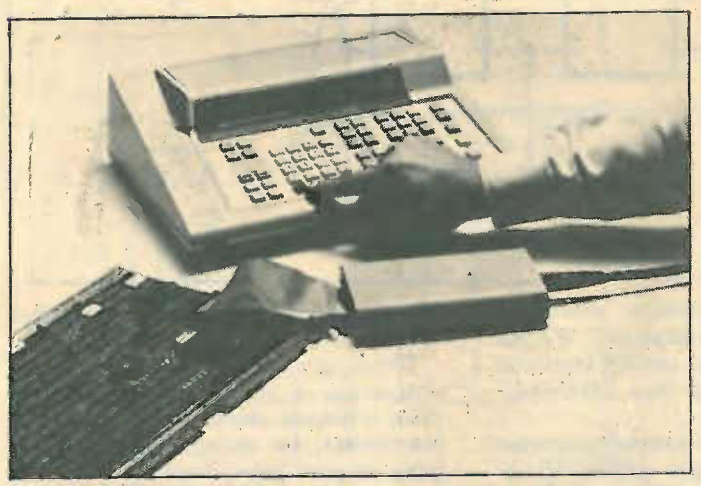
The agreement includes an arrangement for the NRDC part of BTG to recover its investment by a sales levy on relevant products.

Radio-frequency plasma chemistry techniques are being used more and more in making semiconductor devices, in place of wet chemical methods. One technical advantage is the ability to create the finer circuit patterns needed for producing a larger number of circuit elements per unit area. Certain gases, when ionized, form a reactive plasma which interacts with solid surfaces to selectively remove unwanted material without residual contamination.

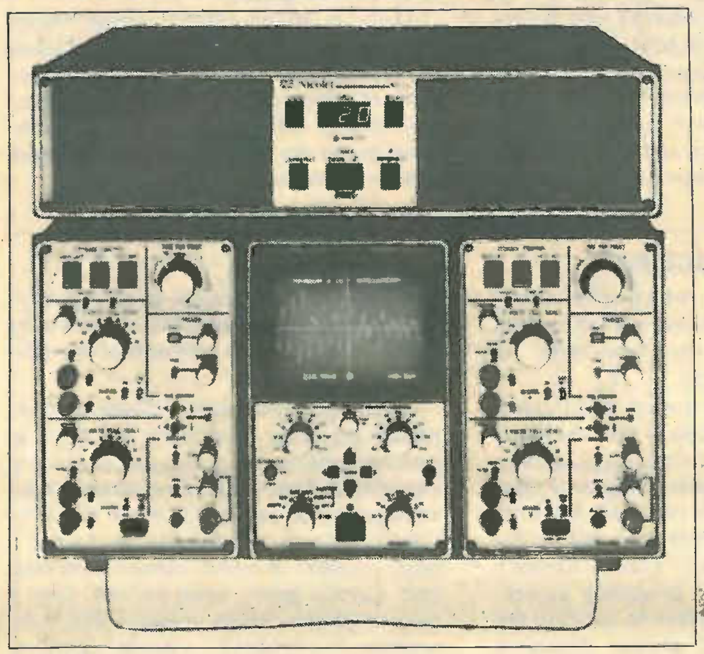
NEW PRODUCTS

Micro-system fault finder

A fault finding routine for r.o.m., r.a.m., bus, clock and power supply of a microprocessor-based system can usually be carried out in under five minutes using the 9010A Micro-System Troubleshooter, regardless of system complexity, claim Fluke. Test programs need not be written since the 9010A has a 'learn' mode in which it examines and defines all the digital locations and functions of a working system and stores the information in its memory. When a similar but suspect system is connected to the tester the information from the working system is compared with that from the suspect equipment and any faults indicated on a 32-character alphanumeric display. The tester's programs, whether 'learnt' or presented manually, can be loaded onto cassette for future use. Operators need not have a knowledge of programming language, and to use the tester, a plug from an eight or sixteen-bit processor module is inserted into the microprocessor socket of the system to be tested. If the system malfunction does not appear when the tester is connected the fault is narrowed down to the processor removed. The tester module contains a microprocessor which replaces the one removed from the board. Using other algorithms, the tester can be used to check peripherals such as character generators, keyboards, readouts, printheads and relays. Modules can be obtained for testing 8080, 8085, Z80, 6502, 6800 or 9900 based systems. A further seven modules should become available within the next ten months. Fluke International Corporation, Colonial Way, Watford, Herts WD2 4TT.
WW301



WW301



WW302

WW303

Storage oscilloscope

The main unit of Nicolet's latest digital-storage oscilloscope is the 4094 with 16K-word×16-bit memory capacity. Two dual-channel input-amplifier modules, the 4851 with 15-bit a-to-d conversion and 100kHz sampling rate and the 4562 with 12-bit conversion and 2MHz sampling, can be added to the main unit in any combination for either two or four-channel operation as the main unit's memory can be shared. Permanent waveform storage is possible using a single floppy-disc drive, the F-43, with a dual-channel oscilloscope or the XF-44 dual disc-drive with four-channel versions. Cursor positioning, display expansion (up to ×256), r.m.s. calculation and waveform addition, subtraction and inversion are standard on the 4094 and further programs for waveform multiplication, integration, etc, are available on disc. Both plug-in input amplifiers have pre- and post-trigger delay controls and where two amplifiers are used they can be operated independently thus forming two dual-channel oscilloscopes with a common display. An RS232 and IEEE488 i/o interface and digital plotter are available for use with the oscilloscope. The manufacturers have also introduced a small 4000-line FFT spectrum analyzer, the 100A, for 0 to 20kHz. Nicolet Instruments Ltd, Budbrooke Road, Warwick CV34 5XH.
WW302

Softy 2

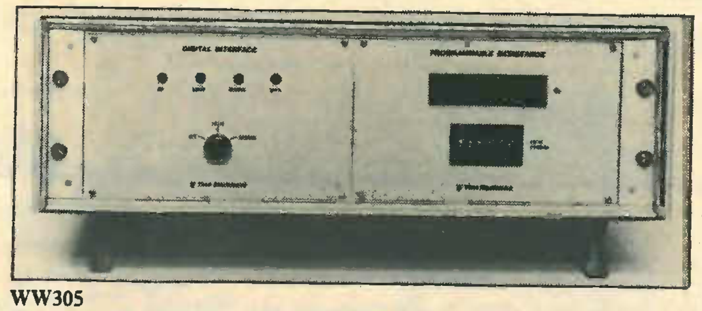
Following the success of Softy, a low-cost micro tool for e.p.r.o.m. programming, copying and r.o.m. emulation, the designer has recently introduced an enhanced version called Softy 2. This unit is similar to the original version which displays the contents of 512 contiguous addresses in hex form on a television screen via an internal modulator. Improvements include an expanded monitor and keypad (28-key, two-level) to provide extra functions such as serial (RS232) and parallel (Centronics) routines for connection to other computer systems or printers. Code can also be stored on cassette tape using a new system called Transwift which is claimed to be tolerant of speed and level changes. The buffer r.a.m. has also been increased to 2K and the unit will now program or copy the 2716, 2732, 2532 family of single-rail e.p.r.o.ms. To make r.o.m. emulation easier, the address and data lines have been buffered and the unit is supplied with a ribbon cable and 24-pin plug. Softy is only available built and tested in a black plastic case and is supplied with a separate power supply for around £169 + v.a.t. Dataman Designs, Lombard House, 24 Cornwall Rd, Dorchester, Dorset DT1 1RS.
WW303

Low-noise f.e.t. preamplifier

At 1MHz, the noise characteristic of the AH0013 linear wideband preamplifier is quoted as $800pV/\sqrt{Hz}$. An f.e.t. input is used in this hybrid device giving a typical input impedance of 100GΩ and maximum bias-current requirement of 50pA. Sonar, audio, infra-red detection and communication equipment applications are suggested for the device. Packaging is 8-pin d.i.l. and the operating temperature range is from −65°C to +125°C. DI-AN Data Systems Ltd, Mersey House, Battersea Rd, Heaton Mersey, Stockport, Cheshire SK4 3EA.
WW304

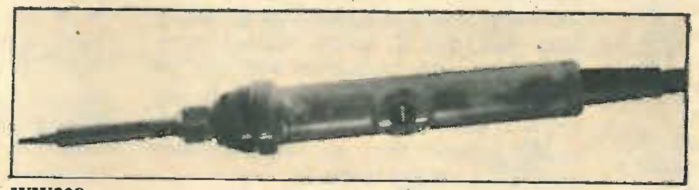
IEEE-programmable resistance box

Resistances from 1Ω to 1MΩ in 1Ω steps can be programmed through an IEEE-488 link from a microcomputer on the 9811 resistance/potential-divider box from Time Electronics. Manual resistance selection is also possible by means of thumbwheel switches on the front panel and in both modes the value is displayed on a six-digit l.c.d. Separate outputs for each other decade are provided so the unit can be used as a programmable potential divider. The resistance outputs are electrically isolated from the digital input. Maximum power dissipation, voltage and current rating of any resistance value are 1W, 100V and 1A respectively. Resistance-value error and temperature coefficient figures are ±0.1% (for values above 100Ω) and 50 p.p.m./°C. Both rack-mounting and free-standing versions can be supplied and operation through RS232 or parallel b.c.d. is possible using an IEEE converter. Time Electronics Ltd, Botany Industrial Estate, Tonbridge, Kent.
WW305

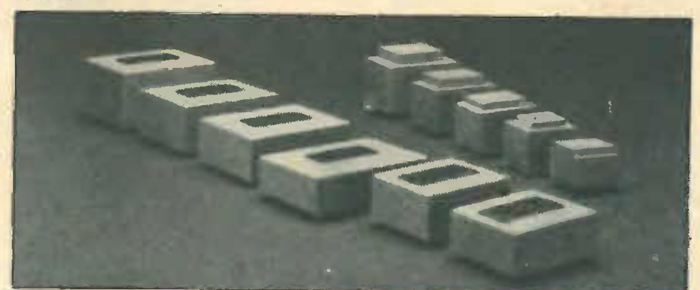


WW305

WW307



WW308



WW311

High-voltage meter

Measurement of voltages up to 150kV with less than ±5% error is claimed for this high-voltage meter. The Field Mill Voltmeter FM150, available through Hunting Hivolt, takes its reading from the field created by a voltage as opposed to measuring the potential difference directly, and is said to give negligible current drain. One switch is used to select 5kV, 25kV and 150kV ranges and battery check and on/off functions. The coupling cable between the measuring transducer and the control box can be any length (within reason, we suppose) without affecting the instrument's capabilities. Rechargeable batteries and charger are supplied as standard. Hunting Hivolt Ltd, Riverbank Works, Old Shoreham Rd, Shoreham by Sea, Sussex BN4 5FL.
WW306

Hard-disc for micros

The Seagate 5¼in Micro-Winchester disc-drive and intelligent controller have been added to the range of computer peripherals stocked by Euro Electronics Ltd. Fitting into the same space as a 'mini-floppy', the ST-506 hard-disc drive with two double-sided discs and four heads uses the same power supply as a 5¼in floppy-disc but offers 5Mbytes of storage, 170ms access time (access time can be reduced by a half using software), data transfer at 5M-bit/s and a projected mean-time-between-failure of 8,000h. The hard-disc and a 5¼in floppy-disc can be used together, the former providing large memory capacity and the latter system back-up and input/output. One or two drives can be controlled by one DTC-510 intelligent controller. There are standard adaptors available for microcomputers such as the Apple II, TRS-80, Motorola's Exorciser, the Dec LSI-11 mini and many others with S-100 or Intel Multibus systems. A second controller, the DTC-520, can control up to four drives in any combination of Micro-Winchesters and 5¼in floppy-disc drives. The ST-506 drive is available at a one-off price of £763, the DTC-510 is £595 and a typical interface, that for the Apple, costs £195. A range of software drivers is also available. Euro Electronics Ltd, Twyman House, 31 Camden Rd, London NW1 1YE.
WW307

Soldering irons

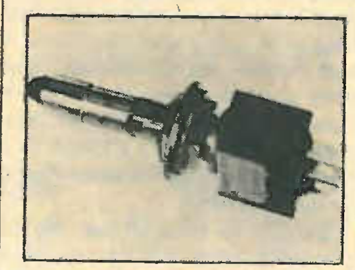
Power-control electronics and associated setting dial of Permax soldering irons are contained within the handle. An aluminium/ceramic base on which the heating element is printed is said to give good heat conductivity combined with high electrical insulation; 17½W versions have an average leakage current of 0.78μA (19pF) and 26W versions 1.1μA (22pF). To aid heat transfer, the flat heating element fits directly into a slit in the iron's alloy coated tip. Both versions can be obtained for 220V or 240V mains operation at around £20 and a wide range of easily interchangeable tips are manufactured for these irons. Special Products Distributors Ltd, 81 Piccadilly, London W1V 0HL.
WW308

Current regulator diodes

Two-lead constant-current sources in eight values from 0.24mA to 4.3mA are available through Semiconductor Specialists. The Siliconix CRR range diodes can be operated at temperatures from −55°C to +150°C and have a temperature coefficient rating of 0.15%/°C. For each type, the maximum operating voltage is 100V; limiting voltages for the 0.24mA and 4.3mA types are 0.5V and 1.45V respectively. These regulators are housed in TO-18 packages. Semiconductor Specialists (UK) Ltd, Carroll House, 159 High St, Yiewsley, West Drayton, Middx UB7 7XB.
WW309

Small lever-locking switches

Accidental switching of the M range miniature switches is prevented by inclusion of a lever-locking system. The lever ends are interchangeable to facilitate colour coding. Ratings of these switches are: 6A at 125V a.c. and 3A at 250V a.c. (resistive), 500V insulation breakdown and 1GΩ insulation resistance. Versions are available with 1, 2, 3 or 4 poles, either gold or silver-plated contacts, and with sealed casings. NSF Ltd, Switches and Controls, Keighley, Yorkshire BD21 5EF.
WW310



Small transformers

Two ranges of small p.c.b.-mounting encapsulated transformers have been introduced to the market by Clairtronic. Transformers in the 'flat' range have power ratings from 3 to 25VA in standard voltage ratings of 6, 12, 15 or 20V secondary, 240V primary. Four mounting screw holes are provided on each unit. The smallest version measures 44×53×17mm and the largest 69×58×35mm. In each case the third dimension, the height, is excluding 5mm long pins. A series of 'top-cap' types are also available in the same range of secondary voltages but with split primaries (2×120V) and power ratings from 0.8 to 25VA. Screw mounting holes are provided for versions above 8VA. Dimensions of the 0.8VA version are 27×27×20mm and of the 25VA type, 80×54×47mm. Prices range from around £1.20 for the smallest 'top-cap' type to around £4.50 for the largest 'flat' type. Clairtronic Ltd, 6 Wayside Gardens, Gerrards Cross, Bucks.
WW311

SIDEBANDS

By Mixer

Basic holidays

For as long as I can remember, I've been very much in favour of lots of healthy exercise. I can sit around and watch people playing games, digging gardens and jogging round the park all day long without feeling any the worse for it, and I'm sure it has all helped to shape my easy-going and tolerant outlook on life.

Those summer camps for kids in America have, likewise, long been a focus of admiration for me: a stroke of absolute genius, I've often thought. We British have developed hypocrisy to a pretty exceptional level, but whoever thought of those camps has nothing to learn from us. No sooner do the 2.4 brats come belting home from school for the summer holidays than they are given a hose down, provided with enough of everything to avoid the need for further communication in the foreseeable future and packed off to the hinterland to be coped with by others, while Ma and Pa take off for the sun. And it's all done in such a way that even the kids themselves, and probably the parents too, sometimes, think it's a Good Thing. The Outdoor Life, and all that. That's what I thought, anyway.

But it's all gone wrong. No longer do emergent Americans go on long walks, swim, climb grizzly bears and fish for waffles. What they do now is compute! In short pants and deadly seriousness, they are hooked on computing.

I haven't been able to find out any more details except that, at the camp I've heard of, the kids are turned loose on the computers at the crack of dawn and, apart from a few seconds to swallow a pecan pie washed down with a glass of clam chowder (I really must try to discover what all these things are) stick with it till sack-time, as I believe it is called. According to the man who runs the operation, it is quite difficult, short of resorting to the garrotte, to stop the little people computing away like crazy when it is time for the Sandman to come and put sleep in their eyes. What they compute, I have no idea. Perhaps it's the answer to imaginary questions such as "Why aren't I in Florida with Mom and Pop?".

Oasis

I've always been a bit envious of anyone who can eat snails, or mussels, or oysters, or sheep's eyes or any of those things with evident relish and without any apparent coercion. There must be a lot I'm missing by being so pernickety, but even just writing about eating oysters is making me feel all peculiar. It isn't just the dishes themselves, either, that make me curl up – you only have to mention sheep's eyes when there is a hard-boiled egg with the salad to put me right off. Association, you see, that's what it is.

I hope this flaw in my make-up isn't characteristic of the majority of c.b. enthusiasts, because the fraternity now has its own restaurant: its name – The Eyeball Bistro. It's in Princes Street, near Oxford Circus. Of course, it doesn't mean they're going to serve eyeballs (it doesn't, does it?) but the damage is done, so far as I am concerned. Why couldn't they call it 'The Breaker Bistro' or 'Eighty-eights' and accept the risk of being overwhelmed by demolition workers or bingo players?

Further grounds for misgivings arise from the declared intention to print the menu in c.b. jargon. Now, at that point, I do really think one has to demur. With something as serious as food and drink, there must be no room for misunderstanding, and this idea is a most dangerous precedent. I realise that menus are sometimes written in French for reasons of snobbery, but I've learned to get along with that, and it doesn't change every other fortnight. "Superslab with i.cs, in tears" is no way to talk about steak and chips in onion gravy and I do most earnestly suggest that, on compassionate grounds, the idea is dropped.

Postillions beware!

Translation by computer of foreign languages has always been good for a laugh. There was a story that a lucrative civil engineering contract was once turned down because the contract, translated by computer from the Russian, insisted that a flock of water-sheep would be needed. Since the company didn't have anywhere to keep the animals, and couldn't find out what they were anyway, it decided to forget the whole thing and work for the Arabs, instead. So a German company, whose sales manager knew how to translate the Russian for 'hydraulic ram', got the job.

But we're over all that sort of thing now. At least, I hope we are, because the EEC is wanting to use a new system, Eurotra, which they hope will be able to cope with the 72 language pairs in use in the community when Portugal and Spain get their tickets – that's each of nine languages translated into all the other eight.

So it had better be a good system, and it ought really to be able to handle the odd bit of idiom, slang and dialect. The thought of a British MEP, infuriated by yet another bland explanation of why butter mountains and wine lakes are not bad economics, rising to his feet and bellowing "What a load of old cobblers!" is a sombre one. Unless Eurotra can deal effectively with vituperative outbursts of this kind, there are problems ahead. One can imagine the blind panic which would be the natural result of the foregoing innocent remark: proceedings would be halted while the premises were searched for the elderly shoemakers and the whole fabric of civilized Community life would be imperilled.

Maybe Esperanto would be a better idea.

Man-powered flight

Pilots have a lot to cope with. It isn't all just a matter of sitting there, looking at the pretty clouds and sending the air hostess for cups of coffee every few minutes: not by any manner of means. It might be like that some of the time, I dare say, but every now and then their reactions are put to the test when things go wrong. You don't want to be too relaxed when, for example, an engine goes on strike just after take-off, or when the controller tells you that another aircraft is making a head-on approach at a closing speed of around 1200 knots.

To help pilots get used to little problems of this kind, airlines make use of flight simulators, which can be programmed with all kinds of 'failure'. Accelerations, visual effects and sounds are provided to make the simulation as realistic as it can be, the sounds being recorded in a real aircraft and played back under the control of the simulator.

Clearly, as many sounds as possible are needed, but a chap I know who used to record the aircraft noises tells me there was one he never could get in the normal way. Birds are unco-operative little beasts, and he found it quite difficult to persuade them to fly into windscreens to order so he could record the splat and allow birdstrikes to be simulated on the ground.

So, I am about to explain a bizarre little scene which someone may have seen and wondered about (though not for long, probably – the world is full of strange people). What they had to do, it seems, was to buy some frozen, oven-ready chickens from Sainsbury's, thaw them out and give them to several brawny characters with good right arms. On the word of command, the chaps hurled the fowls with considerable enthusiasm at the aircraft's nose cone, while the tape recorder inside collected the bangs. Played back at a higher speed, it gave just the right effect and everyone (except the chickens) was happy.

electronics today INTERNATIONAL

The November issue of Electronics Today International features no less than 100 circuits designed for ETI by Tim Orr — amplifiers, timers, alarms and filters.

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2N3632	6.03	2N5090	8.44	BLX66	4.91	BLY83	7.45
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2N4041	10.97	2N5642	8.11	BLX92A	13.06	BLY88A	8.66
2N4127	9.18	2N5643	12.44	BLX93A	19.19	BLY88C	9.10
2N4128	11.03	2N5913	2.34	BLX94A	35.79	BLY89A	12.65
2N4129	12.08	2N6080	5.94	BLX95	44.59	BLY89C	11.90
2N4427	1.15	2N6081	9.87	BLX98	84.95	BLY91A	6.95
2N4429	9.89	2N6082	10.17	BLY33	1.87	BLY91C	6.90
2N4430	11.30	2N6083	11.08	BLY34	1.07	BLY92A	9.25
2N4431	12.50	2N6084	12.27	BLY35	7.50	BLY92C	9.06
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74LS86	24p
74LS90	35p
74LS92	40p
74LS93	35p
74LS95	45p
74LS96	110p
74LS107	45p
74LS109	30p
74LS112	34p
74LS113	30p
74LS114	30p
74LS122	42p
74LS123	50p
74LS124	120p
74LS125	30p
74LS126	30p
74LS132	45p
74LS133	30p
74LS136	30p
74LS138	36p
74LS139	36p
74LS145	75p
74LS147	160p
74LS148	90p
74LS151	70p
74LS153	60p
74LS154	200p
74LS155	40p
74LS156	40p
74LS157	35p
74LS158	36p
74LS160	40p
74LS161	40p
74LS162	40p
74LS163	40p
74LS164	48p
74LS165	100p
74LS166	90p
74LS170	120p
74LS173	70p
74LS174	60p
74LS175	50p
74LS181	140p
74LS190	50p
74LS191	50p
74LS192	50p
74LS193	60p
74LS195	50p
74LS196	60p
74LS197	65p
74LS221	60p
74LS240	90p
74LS241	90p
74LS242	80p
74LS243	80p
74LS244	80p
74LS245	90p
74LS247	70p
74LS251	40p
74LS253	40p
74LS257	45p
74LS258	45p
74LS259	90p
74LS266	25p
74LS273	90p
74LS279	45p
74LS283	45p
74LS298	160p
74LS323	250p
74LS324	150p
74LS348	150p
74LS352	100p
74LS353	100p
74LS365	36p
74LS367	36p
74LS368	50p
74LS373	80p
74LS374	80p
74LS375	50p
74LS377	90p
74LS378	70p
74LS390	55p
74LS393	50p
74LS399	200p
74LS540	135p
74LS541	135p
74LS670	170p

4000 SERIES

Type	Price
4000	12p
4001	14p
4002	16p
4006	65p
4007	16p
4008	60p
4009	35p
4010	40p
4011	14p
4012	18p
4013	35p
4014	60p
4015	60p
4016	30p
4017	45p
4018	60p
4019	32p
4020	60p
4021	65p
4022	70p
4023	20p
4024	40p
4025	20p
4026	130p
4027	32p
4028	60p
4029	75p
4030	40p
4031	170p
4034	160p
4035	80p
4036	295p
4039	295p
4040	55p
4041	70p
4042	55p
4043	60p
4044	70p
4046	70p
4047	75p
4048	55p
4049	27p
4050	27p
4051	60p
4052	80p
4053	60p
4054	130p
4055	125p
4056	120p
4059	500p
4060	90p
4063	100p
4066	35p
4067	400p
4068	18p
4069	18p
4070	18p
4071	18p
4072	18p
4073	20p
4075	20p
4076	60p
4081	16p
4082	20p
4086	72p
4089	150p
4093	40p
4094	150p
4095	95p
4096	95p
4097	340p
4098	90p
4099	120p
40085	90p
40097	£120p
40098	120p
40102	180p
40103	180p
40106	50p
40109	100p
40163	100p
40173	120p
40174	90p
40175	100p
40193	120p
40257	160p
4502	75p
4503	50p
4507	40p
4508	200p
4510	65p
4511	50p
4512	65p
4514	150p
4515	150p
4516	75p
4518	45p
4520	70p
4521	150p
4526	75p
4527	90p
4528	75p
4532	90p
4534	500p
4536	300p
4538	120p
4539	110p
4543	100p
4553	290p
4555	50p
4556	60p
4560	180p
4569	180p
4572	30p
4583	100p
4584	45p
4585	100p
4724	150p
14411	700p
14412	900p
14599	290p

INTERFACE ICs

Type	Price
AD561J	1400p
AD7524	600p
DM8131	375p
DP8304	450p
DS8835	250p
DS8836	150p
DS8838	225p
MC1488	85p
MC1489	65p
MC3446	300p
MC3480	850p
25910	350p
75107	160p
58174	£12
75107	160p
75110	160p
75114	160p
75150	140p
75154	150p
75182	230p
75324	375p
75361	150p
75363	150p
75365	150p
75451/2	72p
75491/2	70p
8T26	120p
8T28	140p
8T95	140p
8T97	140p
81LS95	120p
81LS96	140p
81LS97	120p
81LS98	140p
9602	220p

93 SERIES

Type	Price
9301	160p
9302	175p
9308	316p
9310	275p
9311	275p
9312	160p
9314	165p
9316	225p
9321	225p
9322	150p
9334	360p
9368	250p
9370	300p
9374	200p

74S SERIES

Type	Price
74S00	60p
74S04	80p
74S05	75p
74S08	75p
74S10	60p
74S20	60p
74S30	60p
74S32	90p
74S37	90p
74S64	60p
74S74	90p
74S85	300p
74S86	180p
74S112	180p
74S114	120p
74S124	300p
74S132	160p
74S133	75p
74S138	225p
74S139	225p
74S157	250p
74S163	300p
74S174	250p
74S175	320p
74S194	350p
74S241	450p
74S260	70p
74S373	400p
74S374	400p

LINEAR ICs

Type	Price
AN103	200p
AY1-0212	650p
AY1-1313	668p
AY1-1320	320p
AY1-5050	90p
AY3-1270	840p
AY3-8910	700p
AY3-8912	825p
CA3046	70p
CA3048	225p
CA3080E	60p
CA3086	48p
CA3089E	225p
CA3090AQ	375p
CA3130E	90p
CA3140E	45p
CA3160E	100p
CA3161E	150p
CA3162E	450p
CA3189E	300p
CA3280	180p
DAC1408-8	200p
HA1388	260p
ICL8038	300p
ICM7555	80p
LC7120	600p
LC7130	475p
LF351	48p
LF356P	90p
LM10C	425p
LM301A	27p
LM311	70p
LM318	200p
LM319	225p
LM324	45p
LM339	65p
LM348	75p
LM358P	175p
LM377	175p
LM380	75p
LM381AN	180p
LM386	80p
LM393	100p
LM709	36p
LM710	50p
LM725	350p
LM733	75p
LM741	16p
LM747	70p
LM748	35p
LM2917	200p
LM3302	100p
LM3900	55p
LM3909	70p
LM3911	100p
LM3914	225p
LM3915	225p
LM3916	225p
LM13600	125p
MB3712	200p
MC1310P	150p
MC1458	40p
MC1495L	350p
MC1496	70p
MC3340P	120p
MC3403	75p
MM57160	620p
NE531	150p
NE555	18p
NE556	55p
NE564	420p
NE565	120p
NE566	155p
NE567	140p
NE571	425p
NE5534A	250p
PLL02A	500p
RC4136	60p
S566B	240p
SAD1024A	1250p
SFF96364	800p
SN76013N	170p
SN76477	175p
SP8515	750p
TA7120	200p
TA7204	250p
TA7205	200p
TA7222	200p
TA7310	200p
TBA231	150p
TBA641B11	300p
TBA651	200p
TBA800	90p
TBA810	100p
TBA820	90p
TBA950	300p
TCA220	350p
TCA940	175p
TDA1004	300p
TDA1008	320p
TDA1010	225p
TDA1022	570p
TDA1024	100p
TDA1034B	250p
TDA1170	300p
TDA2002V	325p
TDA2020	320p
TL071/81	45p
TL072/82	75p
TL074	130p
TL084	110p
UAA170	175p
ULN2003	100p
UPC1156H	300p
XR2206	300p
ZN414	90p
ZN419C	225p
ZN424E	135p
ZN425E	350p
ZN427E	650p
ZN1034	200p

VOLTAGE REGULATORS
Fixed Plastic TO-220

1A	+ve		-ve	
5V	7805	50p	7905	55p
12V	7812	50p	7912	55p
15V	7815	55p	7915	55p
18V	7818	55p	7918	70p
24V	7824	55p	7924	70p
100mA	TO-92			
5V	78L05	30p	79L05	60p
12V	78L12	30p	79L12	60p
15V	78L15	30p	79L15	60p

OTHER REGULATORS

Type	Price	Type	Price
LM309K	135p	78HGKC	600p
LM317T	250p	78H05KC	550p
LM323K	500p	78MGT2C	135p
LM723	37p	79HGKC	600p
		TL497	300p

OPTO-ELECTRONICS

Type	Price	Type	Price
2N5777	45p	ORP60	120p
OCP71	120p	ORP61	120p
ORP12	120p	TIL78	55p

OPTO-ISOLATORS

Type	Price	Type	Price
ILD74	130p	TIL111	90p
MCT26	100p	TIL112	90p
MCS2400	190p	TIL116	90p

LEDS

Type	Price	Type	Price
0.125''		0.2''	
TIL32	55p	TIL220 Red	16p
TIL209 Red	13p	TIL222 Gr	18p
TIL211 Gr	20p	TIL228 Red	22p
TIL212 Ye	25p	Rectangular LEDs (R, G, Y)	30p
TIL216 Red	18p		

DISPLAYS

Type	Price	Type	Price
3015F	200p	NSB5881	570p
DL704	140p	TIL311	800p
DL707 Red	140p	TIL312/3	110p
FND357	120p	TIL321/2	130p
FND500	110p	TIL330	140p
FND507	110p	7750/60	200p
MAN3640	175p		
MAN4640	200p		

DRIVERS

Type	Price
9368	250p
9370	300p
UDN6118	320p
UDN6184	320p

TRANSISTORS

Type	Price
AC126	25p
AC127/8	20p
AC176	25p
AC187/8	25p
AF116	50p
AD149	70p
AD161/2	45p
AU107	200p
BC107/8	13p
BC109	14p
BC117	20p
BC147/8	9p
BC149	10p
BC157/8	10p
BC159	11p
BC169C	12p
BC172	12p
BC177/8	17p
BC179	18p
BC182/3	10p
BC184	11p
BC187	30p
BC212/3	11p
BC214	12p
BC237	15p
BC327	16p
BC337	16p
BC338	16p
BC461	36p
BC477/8	30p
BC516/7	40p
BC547B	16p
BC548C	9p
BC549C	18p
BC557B	16p
BC559C	18p
BCY70	18p
BCY71/2	22p
BD131/2	50p
BD135/6	54p
BD139	56p
BD140	60p
BD189	80p
BD232	95p
BD233	75p
BD235	85p
BD241	70p
BD242	70p
BDY56	200p
BF200	32p
BF244B	35p
BF256B	70p
BF257/8	32p
BF259	36p
BFR39	25p
BFR40	25p
BFR41	25p
BFR79	25p
BFR80	25p
BFR81	25p
BFX29	40p
BFX30	34p
BFX84/5	40p
BFX86/7	30p
BFX88	30p
BFW10	90p
BFY50	30p
BFY51/2	30p
BFY56	33p
BFY90	90p
BRY39	45p
BSX19/20	24p
BU104	225p
BU105	190p
BU108	250p
BU109	225p
BU126	150p
BU180A	120p
BU205	200p
BU208	200p
BU406	145p
E300	50p
E308	50p
E310	50p
MJ2501	225p
MJ2955	90p
MJ3001	225p
MJE340	60p
MJE2955	100p
MJE3055	70p
MPF102	45p
MPF103/4	40p
MPF105/6	40p
MPS6531	50p
MPS6534	50p
MPSA06	30p
MPSA12	50p
MPSA13	50p
MPSA20	50p
MPSA42	50p
MPSA43	50p
MPSA56	32p
MPSA70	50p
MPSU06	63p
MPSU07	60p
MPSU45	90p
MPSU65	78p
OC28	130p
OC35	130p
TIP29A	40p
TIP29C	55p
TIP30A	48p
TIP30C	60p
TIP31A	58p
TIP31C	62p
TIP32A	68p
TIP32C	82p
TIP33A	90p
TIP33C	114p
TIP34A	115p
TIP34C	160p
TIP35A	225p
TIP35C	290p
TIP36A	270p
TIP36C	340p
TIP41A	65p
TIP41C	78p
TIP42A	70p
TIP42C	82p
TIP54	160p
TIP120	120p
TIP122	130p
TIP142	130p
TIP147	130p
TIP2955	78p
TIP4055	70p
TIS93	30p
ZTX108	12p
ZTX300	13p
ZTX500	15p
ZTX502	18p
ZTX504	30p
2N457A	250p
2N696	35p
2N697	25p
2N698	45p
2N706A	30p
2N708	30p
2N918	45p
2N930	18p
2N1131/2	36p
2N1613	25p
2N1711	25p
2N2102	70p
2N2160	350p
2N2219A	30p
2N2222A	30p
2N2369A	25p
2N2484	30p
2N2646	45p
2N2904/5	30p
2N2906A	30p
2N2907A	30p
2N2926	9p
2N3053	30p
2N3054	65p
2N3055	54p
2N3442	140p
2N3553	240p
ZN3584	250p
2N3643/4	48p
2N3702/3	12p
2N3704/5	12p
2N3706/7	14p
2N3708/9	12p
2N3773	300p
2N3819	25p
2N3820	50p
2N3823	70p
2N3866	90p
2N3902	700p
2N3903/4	18p
2N3905/6	20p
2N4037	65p
2N4061/2	18p
2N4123/4	27p
2N4125/6	27p
2N4401/3	27p
2N4427	90p
2N4871	60p
2N5087	27p
2N5089	27p
2N5172	27p
2N5191	90p
2N5194	90p
2N5245	40p
2N5296	55p
2N5401	50p
2N5457/8	40p
2N5459	40p
2N5460	60p
2N5485	44p
2N5875	250p
2N6027	48p
2N6041	160p
2N6044	160p
2N6052	300p
2N6059	325p
2N6107	65p
2N6247	190p
2N6254	130p
2N6290	65p
2N6292	65p
2SC1172	150p
2SC1306	150p
2SC1307	250p
2SC1957	90p
2SC1969	200p
2SC2028	120p
2SC2029	250p
2SC2078	200p
3N128	120p
3N140	120p
3N141	110p
3N201	110p
3N204	120p
40290	250p
40361/2	75p
40408	90p
40409	100p
40410	100p
40411	300p
40594	120p
40595	120p
40673	75p
40871/2	100p

DIODES

Type	Price
BY127	12p
BYX36-300	20p
OA47	9p
OA81	15p
OA85	15p
OA90	9p
OA91	9p
OA95	9p
OA200	9p
OA202	10p
1N914	4p
1N916	7p
1N4148	4p
1N4001/2	5p
1N4003/4	6p
1N4005	6p
1N4006/7	7p
1N5401/3	14p
1N5404/7	19p
IS920	9p

HEATSINKS
For TO220 Voltage Regs. and transistors 22p
For TO5 12p

BRIDGE RECTIFIERS

		Price
1A	50V	19p
1A	100V	20p
1A	400V	25p
1A	600V	30p
2A	50V	30p
2A	100V	35p
2A	400V	45p
3A	200V	60p
3A	600V	72p
4A	100V	95p
4A	400V	100p
6A	50V	80p
6A	100V	100p
6A	400V	120p
10A	400V	200p
25A	400V	400p

ZENERS
2.7V-33V
400mW 9p
1W 15p

TRIACS PLASTIC

	Price
3A 400V	80p
6A 400V	70p
6A 500V	88p
8A 400V	75p
8A 500V	95p
12A 400V	85p
12A 500V	105p
16A 400V	110p
16A 500V	130p
T2800D	130p

THYRISTORS

	Price
1A 50V	70p
1A 400V	90p
3A 400V	100p
8A 600V	140p
12A 400V	160p
16A 100V	160p
16A 400V	180p
BT106	110p
C106D	40p
MCR101	36p
TIC44	27p
2N3525	130p
2N4444	140p
2N5060	34p
2N5064	40p

LOUD-SPEAKERS

Size		
2½''	64R	80p
2½''	8R	80p
2''	8R	90p
1½''	8R	100p

MODULATORS
6MHz UHF 375p
8MHz UHF 450p

MEMORIES

Type	Price
2101-4L	400p
2102-2L	120p
2107B	500p
2111-4	300p
2112-4	300p
2114L	130p
2114-2L	160p
4027-3	300p
4044	700p
4116	200p
4118	500p
5101	300p
6116 (150ns)	£11
6514-45	£4
6810	200p

ROM/PROMs

Type	Price
71301	700p
74S188	225p
74S287	350p
74S471	650p
74S571	650p

CPUs

Type	Price
1802CE	750p
2650A	£16
6502	400p
6800	370p
6802	500p
6809	1260p
8080A	450p
8085A	550p
1NS8060	1000p
TM59980A	£20
Z80	370p
Z80A	550p

EPROMs

Type	Price
1702A	500p
2708	300p
2716 (+5V)	400p
2532	600p
2732	600p

SUPPORT DEVICES

Type	Price
3242	800p
3245	450p
6522	500p
6532	775p
6821	160p
6845	£10
6850	180p
6852	370p
6875	600p
8154	950p
8155	800p
8205	320p
8212	180p
8216	180p
8224	250p
8226	250p
8228	250p
8251	400p
8253	800p
8255	400p
8257	800p
8259	800p
8279	950p
TMS9918	POA
Z80P10	400p
Z80AP10	450p
Z80CTC	400p
Z80ACTC	500p
Z80ADART	£8
Z80ADMA	£12
Z80S10-1	£20

UART

Type	Price
AY-3-1015P	400p
AY-5-1013P	350p
IM6402	450p

CHARACTER GENERATORS

Type	Price
RO-3-2513 U.C.	750p
RO-3-2513 L.C.	700p
SN74S262AN	£10

KEYBOARD ENCODER

Type	Price
AY-5-2376	700p
74C922	500p

CRYSTALS

Frequency	Price
32.768KHz	250p
100KHz	300p
200KHz	370p
1.0MHz	320p
1.008MHz	350p
1.8432MHz	250p
2.00MHz	250p
2.45760MHz	250p
3.276MHz	250p
3.5795MHz	175p
4.00MHz	250p
4.194MHz	250p
4.43MHz	125p
5.0MHz	250p
6.0MHz	250p
6.144MHz	250p
7.0MHz	250p
7.168MHz	250p
8.00MHz	250p
8.867MHz	250p
10.00MHz	250p
10.7MHz	250p
1p.0MHz	350p
16.00MHz	250p
18.00MHz	250p
18.432	250p
19.968MHz	390p
26.690MHz	300p
27.145MHz	250p
38.6667MHz	350p
48.0MHz	300p
55.5MHz	400p
116MHz	350p

LOW PROFILE DIL SOCKETS BY TEXAS

8 pin	9p	18 pin	16p	24 pin	24p
14 pin	10p	20 pin	18p	28 pin	26p
16 pin	11p	22 pin	22p	40 pin	30p

ZERO INSERTION FORCE SKT
24 pin £6

WIRE WRAP SOCKETS BY TEXAS

8 pin	25p	18 pin	50p	24 pin	70p
14 pin	35p	20 pin	60p	28 pin	80p
16 pin	40p	22 pin	65p	40 pin	100p

HEADER PLUGS

	Price
14 pin	40p
16 pin	50p
24 pin	90p
40 pin	250p

COUNTERS

Type	Price
74C925	550p
74C928	600p
ICM7216B	£20
ZN1040E	£7

TTL & ECL

Type	Price
MC4024	325p
MC4044	325p
10116	70p
10231	350p

DISC IC

Type	Price
FD1771	£20
FD1791	£30
WD1691	£15
WD2143	550p

★ SPECIAL OFFERS ★

	1-24	25-99	100
2114L (450ns)	1.10	1.05	1.00
2716 (+5V)	2.75	2.50	2.25
4116 (200ns)	0.90	0.85	0.80

ACORN ATOM. Supplied with full size QWERTY Keyboard, 8K ROM, 2K RAM and can be expanded to 12K ROM, 12K RAM. Included are UHF Modulator for connection to domestic TV and Free manual on Programming in BASIC or Machine Code.
Kit **£129**. Built **£150**. P&P £2.50.

UK 101/OHIO II TWO BOARD HARDWARE INTERFACE

DECODING BOARD plugs into UK 101/OHIO II socket to provide 16 bit user port plus a wide variety of decoded lines for interfaces (inc. full decoding for AY3-8910/12 PSG) and a 40-pin socket for further expansion. Kit **£27.50.**
Analogue Board plugs into decoding board to provide D/A converter, 8-channel multiplexed A/D converter, AY3-8910 PSG plus 6522 timer IC allowing complex timing and counting functions, plus 16-bit port. Kit **£39.95.**
Reprint of 'P.E.' articles **£1** + SAE.

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Ideal software/hardware tool. Develop your programmes, debug/verify and then commit them to EPROM. Ready-built for +5V EPROMS, **£169**.

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A microprocessor trainer based on powerful state-of-the-art 6809. Facilities for further expansion (to be published in WW). An excellent system for training centres and especially useful to control application engineers. Complete Kit (inc. PCBs, case, prog. 2716, etc), **£80**. (As described in "W.W." July '81.) 6809 Prog. Ref. Manual, **£5.**

ADD SOUND TO ZX80/81 PLUS RELAY CONTROL AND LIGHT DETECTION

★ ZX80/81 USER PORT ★

(as described in "P.C.W." Oct. '81)

Port module plugs directly into ZX80 or ZX81 to provide 8 input and 8 output lines. These allow input of data from switches, photocells, joy-sticks, etc., and control of up to 8 relays. Also 7-segment LED displays or LED lamps may be used and solid-state buzzers may be directly connected to the port. Variable tone audio output may be produced. The port is accessed by simple PEEK and POKE commands. The port kit includes plugs, sockets, double-sided PCB and all components. Price **£11.50**. Software and application sheet 40p + SAE.

SEE US AT "P.C.W." EXHIBITION IN EARLY SEPTEMBER FOR ABOVE AND OTHER PROJECTS

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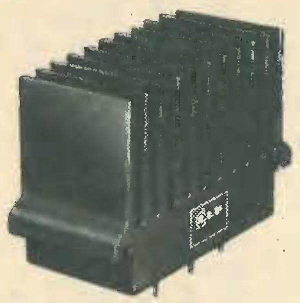
20 POWER AMPS   19 FUNCTIONAL MODULES   DAWN OF A NEW ERA   ILP

POWER UP TO 480 WATTS RMS SINGLE CHANNEL

Which amplifier?

I.L.P. Amplifiers now come in three basic types, each of which is available with or without heatsink. Having decided the system you want – home hi-fi (models HY30, 60 or 120 for example), super quality hi-fi with extra versatility (MOS120, MOS200) or Disco/PA/Guitar (HD120, HD200 or HD400) you will then decide whether amplifiers housed within their own heatsinks or plate amplifiers for bolting to a metal chassis will suit. With choice such as this and a brilliant new range of I.L.P. functional modules to choose from you now have the chance to build the finest audio system ever offered to the constructor.



AMPLIFIER WITH HEAT SINK

BIPOLAR Standard, with heatsinks

MODEL NUMBER	OUTPUT POWER Watts rms	DISTORTION T.H.D. Typ at 1kHz	DISTORTION I.M.D. 60HZ/7kHz 4:1	SUPPLY VOLTAGE TYP/MAX	SIZE mm	WT gms	PRICE	VAT	Without heatsinks: MODEL NUMBER	SIZE in mm	WT gms	PRICE	VAT
HY30	15w 4-8Ω	0.015%	<0.006%	±18±20	76x68x40	240	£7.29	£1.09					
HY60	30w 4-8Ω	0.015%	<0.006%	±25±30	76x68x40	240	£8.33	£1.25					
HY120	60w 4-8Ω	0.01%	<0.006%	±35±40	120x78x40	410	£17.48	£2.62	HY120P	120x26x40	215	£15.50	£2.33
HY200	120w 4-8Ω	0.01%	<0.006%	±45±50	120x78x50	515	£21.21	£3.18	HY200P	120x26x40	215	£18.46	£2.77
HY400	240w 4Ω	0.01%	<0.006%	±45±50	120x78x100	1025	£31.83	£4.77	HY400P	120x26x70	375	£28.33	£4.25

Protection: Load line, momentary short circuit (typically 10 sec) *Slew rate:* 15V/μs *Rise time:* 5μs
S/N ratio: 100db *Frequency response (–3dB):* 15Hz – 50kHz
Input sensitivity: 500mV rms *Input impedance:* 100kΩ *Damping factor:* (8Ω/100Hz)>400

HEAVY DUTY with heatsinks

MODEL NUMBER	OUTPUT POWER	T.H.D.	I.M.D.	SUPPLY VOLTAGE	SIZE mm	WT gms	PRICE	VAT	Without heatsinks: MODEL NUMBER	SIZE in mm	WT gms	PRICE	VAT
HD120	60w/4-8Ω	0.01%	<0.006%	±35±40	120x78x50	515	£22.48	£3.37	HD120P	120x26x50	265	£19.84	£2.98
HD200	120w/4-8Ω	0.01%	<0.006%	±45±50	120x78x60	620	£27.38	£4.11	HD200P	120x26x50	265	£23.63	£3.54
HD400	240w/4Ω	0.01%	<0.006%	±45±50	120x78x100	1025	£38.63	£5.79	HD400P	120x26x70	375	£34.28	£5.14

Protection: load line, PERMANENT SHORT CIRCUIT (ideal for disco/group use should evidence of short circuit not be immediately apparent).
The Heavy Duty range can claim additional output power devices and complementary protection circuitry with performance specs. as for standard types.

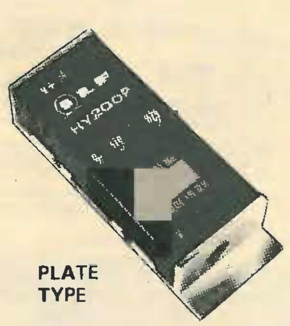


PLATE TYPE

MOSFET Ultra-Fi, with heatsinks

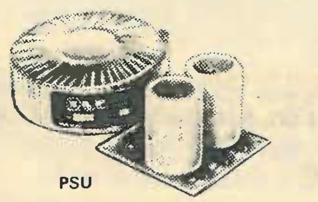
MODEL NUMBER	OUTPUT POWER	T.H.D.	I.M.D.	SUPPLY VOLTAGE	SIZE mm	WT gms	PRICE	VAT	Without heatsinks: MODEL NUMBER	SIZE in mm	WT gms	PRICE	VAT
MOS120	60w/4-8Ω	<0.005%	<0.006%	±45±50	120x78x40	420	£25.88	£3.88	MOS120P	120x26x40	215	£23.32	£3.50
MOS200	120w/4-8Ω	<0.005%	<0.006%	±55±60	120x78x80	850	£33.46	£5.02	MOS200P	120x26x80	420	£28.53	£4.28
MOS400	240w/4Ω	<0.005%	<0.006%	±55±60	120x78x100	1025	£45.39	£6.81	MOS400P	120x26x100	525	£38.91	£5.84

Protection: Able to cope with complex loads, without the need for very special protection circuitry (fuses will suffice).
Ultra-fi specifications:
Slew rate: 20V/μs *Rise time:* 3μs *S/N ratio:* 100db *Frequency response (–3dB):* 15Hz – 100kHz
Input sensitivity: 500mV rms *Input impedance:* 100kΩ *Damping factor:* (8Ω/100Hz)>400

POWER SUPPLY UNITS

MODEL NO.	FOR USE WITH	PRICE	VAT
PSU30	± 15V combinations of HY6/66 series to a maximum of 100mA or *one* HY67	£4.50	£0.68
	The following will also drive the HY6/66 series except HY67 which requires the PSU30.		
PSU36	1 or 2 HY30	£8.10	£1.22
PSU50	1 or 2 HY60	£10.94	£1.64
PSU60	1 x HY120/HY120P/HD120/HD120P	£13.04	£1.96
PSU65	1 x MOS120/1 x MOS120P	£13.32	£2.00
PSU70	1 or 2 HY120/HY120P/HD120/HD120P	£15.92	£2.39
PSU75	1 or 2 MOS120/MOS120P	£16.20	£2.43
PSU90	1 x HY200/HY200P/HD200/HD200P	£16.20	£2.43
PSU95	1 x MOS200/MOS200P	£16.32	£2.45
PSU180	2 x HY200/HY200P/HD200/HD200P or 1 x HY400/1 x HY400P/HD400/HD400P	£21.34	£3.20
PSU185	1 or 2 MOS200/MOS200P/1 x MOS400/ 1 x MOS400P	£21.46	£3.22

All models except PSU30 and PSU36 incorporate our own toroidal transformers.



PSU

FP480
BRIDGING UNIT FOR DOUBLING POWER

Designed specially by I.L.P. for use with any two power amplifiers of the same type to double the power output obtained and will function with any I.L.P. power supply. In totally sealed case, size 45 x 50 x 20mm, with edge connector. It thus becomes possible to obtain 480 watts rms (single channel) into 8Ω. Contributory distortion less than 0.005%.
Price: £4.79 + 72p. V.A.T.

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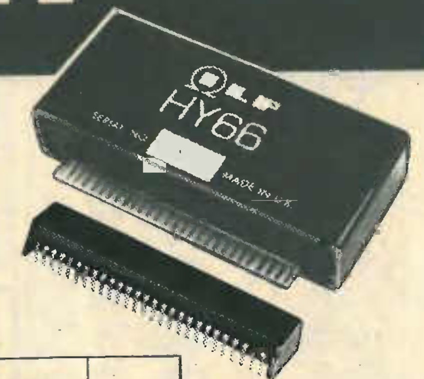
Available also from MARSHALLS, TECHNOMATIC, WATFORD ELECTRONICS and certain other selected retailers.



Which modules?

In launching eighteen different units all within amazingly compact cases to help make complete audio systems using I.L.P. power amplifiers, we bring the most exciting, the most versatile modular assembly scheme ever for constructors of all ages and experience. Study the list – see how these modules will combine to almost any audio project you fancy – and remember *all I.L.P. modules are compatible with each other*, they connect easily. Modules HY6 to HY13 measure 45 x 20 x 40mm. HY66 to HY77 measure 90 x 20 x 40mm. They are so reliable that all I.L.P. modules carry a 5 year no quibble guarantee.

MODEL NO.	MODULE	DESCRIPTION/FACILITIES	CURRENT REQUIRED	PRICE	VAT
HY6	MONO PRE AMP	Mic/Mag. Cartridge/Tuner/Tape/Aux + Volume/Bass/Treble	10mA	£6.44	£0.97
HY7	MONO MIXER	To mix eight signals into one	10mA	£5.15	£0.77
HY8	STEREO MIXER	Two channels, each mixing five signals into one	10mA	£6.25	£0.94
HY9	STEREO PRE AMP	Two channels mag. Cartridge/Mic + Volume	10mA	£6.70	£1.01
HY11	MONO MIXER	To mix five signals into one + Bass/Treble controls	10mA	£7.05	£1.06
*HY12	MONO PRE AMP	To mix four signals into one + Bass/Mid-range/Treble	10mA	£6.70	£1.01
*HY13	MONO VU METER	Programmable gain/LED overload driver	10mA	£5.95	£0.89
HY66	STEREO PRE AMP	Mic/Mag. Cartridge/Tape/Tuner/Aux + Volume/Bass/Treble/Balance	20mA	£12.19	£1.83
HY67	STEREO HEADPHONE	Will drive headphones in the range of 4Ω – 2KΩ	80mA	£12.35	£1.85
HY68	STEREO MIXER	Two channels, each mixing ten signals into one	20mA	£7.95	£1.19
HY69	MONO PRE AMP	Two input channels of mag. Cartridge/Mic + Mixing/Volume/Treble/Bass	20mA	£10.45	£1.57
HY71	DUAL STEREO PRE AMP	Four channels of mag. Cartridge/Mic + Volume	20mA	£10.75	£1.61
*HY72	VOICE OPERATED STEREO FADER	Depth/Delay	20mA	£13.10	£1.97
*HY73	GUITAR PRE AMP	Two Guitar (Bass/Lead) and Mic + separate Volume/Bass/Treble + Mix	20mA	£12.25	£1.84
†HY74	STEREO MIXER	Two channels, each mixing five signals into one + Treble/Bass	20mA	£11.45	£1.72
†HY75	STEREO PRE AMP	Two channels, each mixing four signals into one + Bass/Mid-range/Treble	20mA	£10.75	£1.61
†HY76	STEREO SWITCH MATRIX	Two channels, each switching one of four signals into one	20mA	*To be announced*	
†HY77	STEREO VU METER DRIVER	Programmable gain/LED overload driver	20mA	£9.25	£1.39

* Ready August – may be ordered now
† Ready September – may be ordered now

All the above modules operate from ±15V minimum to ±30V maximum higher voltages being accommodated by use of dropper resistors. HY67 can only be used with the PSU 30 power supply unit.

The modules are encapsulated and include latest design high quality clip-on edge connectors.

For easy mounting we recommend
B6 Mounting board for modules HY6 – HY13 **78p** + 12p. V.A.T.
B66 Mounting board for HY66 – HY77 **99p** + 13p. V.A.T.

All I.L.P. modules include full connection data.

I.L.P. Products are of British Design and Manufacture.

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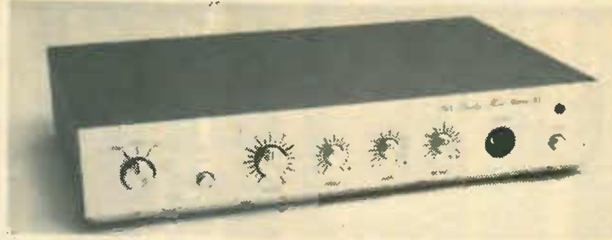
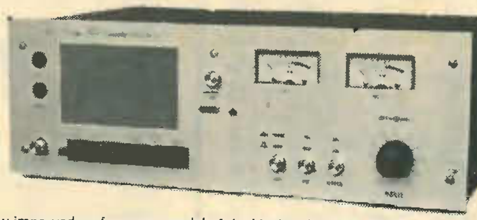
HART

LINSLEY-HOOD 300 SERIES AMPLIFIERS



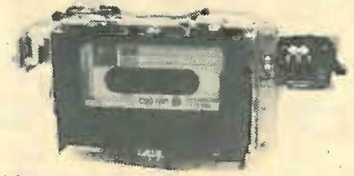
These latest designs from the drawing board of John Linsley-Hood, engineered to the very highest standard, represent the very best that is available on the kit market today. The delicacy and transparency of the tone quality enable these amplifiers to outperform, on a side-by-side comparison, the bulk of amplifiers in the commercial market-place and even exceed the high standard set by his earlier 75-watt design.
Three versions are offered, a 30-watt with Darlington output transistors, and a 35- and 45-watt, both with Mosfet output devices. All are of identical outside appearance which is designed to match and stack with our Linsley-Hood cassette recorder 2.
As with all Hart kits the constructors interests have been looked after in a unique way by reducing the conventional (and boring) wiring almost to the point of extinction.
Any of these kits represents a most cost-effective route to the very highest sound quality with the extra bonus of the enjoyment of building a sophisticated piece of equipment.
30-watt Darlington amplifier, fully integrated with tone controls and magnetic pick-up facility. Total cost of all parts is **£81.12.** Special offer price for complete kits **£72.**
35-watt Mosfet amplifier. Total cost of parts **£98.41.** Special offer for complete kits, **£87.40.**
45-watt Mosfet amplifier. Total cost of parts **£104.95.** Special offer price for complete kits **£94.80.**
Reprints of original Articles from Hi-Fi News 50p. Post free. No VAT.
Reprints of MOSFET article 25p. No V.A.T. Post free.

LINSLEY HOOD CASSETTE RECORDER 2



Our new improved performance model of the Linsley Hood Cassette Recorder incorporates our VFL 910 vertical front mechanism and circuit modifications to increase dynamic range. Board layouts have been altered and improved but retain the outstandingly successful mother-and-daughter arrangement used on our Linsley-Hood Cassette Recorder 1.
This latest version has the following extra features: Ultra low wow-and-flutter of .09% – easily meets DIN Hi-Fi spec. Deck controls latch in rewind modes and do not have to be held. Full Auto-stop on all modes. Tape counter with memory rewind. Oil damped cassette door. Latching record button for level setting. Dual concentric input level controls. Phone output. Microphone input facility if required. Record interlock prevents rerecording on valued cassettes. Frequency generating feedback servo drive motor with built-in speed control for thermal stability. All these desirable and useful features added to the excellent design of the Linsley-Hood circuits and the quality of the components used makes this new kit comparable with built-up units of much higher cost than the modest, **£94.90** + V.A.T. we ask for the complete kit.

FEED YOUR MICRO BYTES WITH OUR SOLENOID CONTROLLED CASSETTE DECK



Front loading deck with full solenoid control of all functions including optional search in fast wind modes. 12 volt operation. Fitted 3-digit memory counter and Hall IC Motion Sensor. Standard erase and stereo R/P Heads. Cheapest price ever for all these features. Only **£38.90** plus VAT. Full technical specification included.

LINSLEY-HOOD CASSETTE RECORDER 1

We are the Designer Approved suppliers of kits for this excellent design. The Author's reputation tells all you need to know about the circuitry and Hart expertise and experience guarantees the engineering design of the kit. Advanced features include: High-quality separate VU meters with excellent ballistics. Controls, switches and sockets mounted on PCB to eliminate difficult wiring. Proper moulded escutcheon for cassette aperture improves appearance and removes the need for the cassette transport to be set back behind a narrow finger trapping slot. Easy to use, robust Lenco mechanism. Switched bias and equalisation for different tape formulations. All wiring is terminated with plugs and sockets for easy assembly and test. Sophisticated modular PCB system gives a spacious, easily-built and tested layout. All these features added to the high-quality metalwork make this a most satisfying kit to build. Also included at no extra cost is our latest HS 16 Sendus Alloy super head, available separately at **£8.20** but included free with the complete kit at **£75** plus VAT.
Reprints of the 3 original articles describing this design **45p.** No VAT.
Reprint of the subsequent postscript article **30p.** No VAT.

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One inexpensive test cassette enables you to set up VU level, head azimuth and tape speed. Invaluable when fitting new heads. Only **£2.70** plus V.A.T. and 50p postage.

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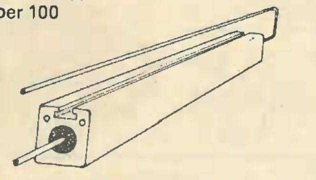

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Complete PROFESSIONAL Case beautifully constructed with cut out for one 'CHERRY' keyboard, plus ample room to house a COMPLETE SYSTEM and our power supply complete with fittings. (Case top detachable). Unit is silver-grey in colour. Robust construction. Sloping front with side ventilation
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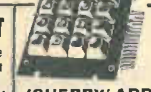
Size 18" x 15½" x 4½" (Front slopes)
A snip at **£39.95** incl. VAT post free.

'CHERRY' ADD-ON KEYPAD



'CHERRY' ADD-ON KEYPAD
A compact 12 button keypad suitable for use with Cherry keyboard to extend its functions plus four extra keys.
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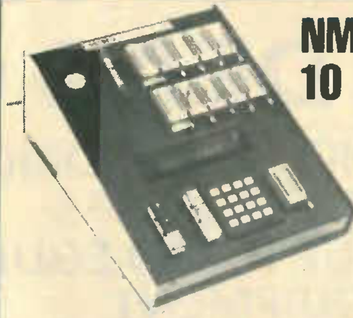
EPROM PROGRAMMER



NMOS, CMOS 10 COPY SOCKETS

MODELS:
S20 for copying
S30 for copying and Editing

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TYPE	SERIES No	SECONDARY RMS Volts	Current	PRICE
30VA 70 × 30mm 0.45Kg Regulation 18%	1X010	6+6	2.50	**£4.48** + 0.87p P/P + 0.80p VAT
	1X011	9+9	1.66	
	1X012	12+12	1.25	
	1X013	15+15	1.00	
	1X014	18+18	0.83	
	1X015	22+22	0.68	
	1X016	25+25	0.60	
	1X017	30+30	0.50	
50VA 80 × 35mm 0.9Kg Regulation 13%	2X010	6+6	4.16	**£4.93** + £1.10 P/P + 0.90p VAT
	2X011	9+9	2.77	
	2X012	12+12	2.08	
	2X013	15+15	1.66	
	2X014	18+18	1.38	
	2X015	22+22	1.13	
	2X016	25+25	1.00	
	2X017	30+30	0.83	
	2X028	110	0.45	
	2X029	220	0.22	
	2X030	240	0.20	
80VA 90 × 30mm 1Kg Regulation 12%	3X010	6+6	6.64	**£5.47** + £1.43 P/P + £1.04 VAT
	3X011	9+9	4.44	
	3X012	12+12	3.33	
	3X013	15+15	2.66	
	3X014	18+18	2.22	
	3X015	22+22	1.81	
	3X016	25+25	1.60	
	3X017	30+30	1.33	
	3X028	110	0.72	
	3X029	220	0.36	
	3X030	240	0.33	
120VA 90 × 40mm 1.2Kg Regulation 11%	4X010	6+6	10.00	**£6.38** + £1.43 P/P + £1.17 VAT
	4X011	9+9	6.66	
	4X012	12+12	5.00	
	4X013	15+15	4.00	
	4X014	18+18	3.33	
	4X015	22+22	2.72	
	4X016	25+25	2.40	
	4X017	30+30	2.00	
	4X018	35+35	1.71	
	4X028	110	1.09	
	4X029	220	0.54	
	4X030	240	0.50	
160VA 110 × 40mm 1.8Kg Regulation 8%	5X011	9+9	8.89	**£8.44** + £1.43 P/P + £1.48 VAT
	5X012	12+12	6.66	
	5X013	15+15	5.33	
	5X014	18+18	4.44	
	5X015	22+22	3.63	
	5X016	25+25	3.20	
	5X017	30+30	2.66	
	5X018	35+35	2.28	
	5X026	40+40	2.00	
	5X028	110	1.45	
	5X029	220	0.72	
	5X030	240	0.66	

Also available at Electrovalue, Maplins, Marshalls, Technomatic and Watford Electronics.

TYPE	SERIES No.	SECONDARY RMS Volts	Current	PRICE
225VA 110 × 45mm 2.2Kg Regulation 7%	6X012	12+12	9.38	**£10.06** + £1.73 P/P + £1.77 VAT
	6X013	15+15	7.50	
	6X014	18+18	6.25	
	6X015	22+22	5.11	
	6X016	25+25	4.50	
	6X017	30+30	3.75	
	6X018	35+35	3.21	
	6X026	40+40	2.81	
	6X025	45+45	2.50	
	6X028	110	2.04	
	6X029	220	1.02	
	6X030	240	0.93	
300VA 110 × 50mm 2.6Kg Regulation 6%	7X014	18+18	8.33	**£11.66** + £1.73 P/P + £2.01 VAT
	7X015	22+22	6.82	
	7X016	25+25	6.00	
	7X017	30+30	5.00	
	7X018	35+35	4.28	
	7X026	40+40	3.75	
	7X025	45+45	3.33	
	7X033	50+50	3.00	
	7X028	110	2.72	
	7X029	220	1.36	
	7X030	240	1.25	
500VA 140 × 60mm 4Kg Regulation 4%	8X017	30+30	8.33	**£15.53** + £2.05 P/P + £2.64 VAT
	8X018	35+35	7.14	
	8X026	40+40	6.25	
	8X025	45+45	5.55	
	8X033	50+50	5.00	
	8X042	55+55	4.54	
	8X028	110	4.54	
	8X029	220	2.27	
	8X030	240	2.08	
625VA 140 × 75mm 5.0Kg Regulation 4%	9X017	30+30	10.41	**£21.54** + £2.20 P/P + £3.56 VAT
	9X018	35+35	8.92	
	9X026	40+40	7.81	
	9X025	45+45	6.94	
	9X033	50+50	6.25	
	9X042	55+55	5.68	
	9X028	110	5.68	
	9X029	220	2.84	
	9X030	240	2.60	

IMPORTANT: Regulator references – All voltages quoted are **FULL LOAD.** Please add regulation figure to secondary voltage to obtain off load voltage.

See also our ads. pages 90 and 91.

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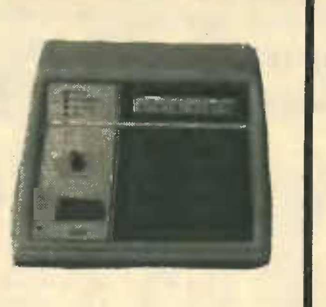

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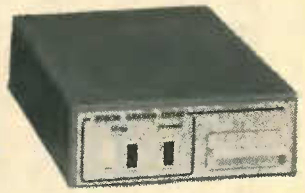

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CHOPPERS RULE O.K!

Switch mode power supplies are now the rule in colour sets, having finally ousted the thyristor regulated power supply. Servicing them is another matter however ... Derek Snelling provides some useful guidelines.

TELEVISION

October issue
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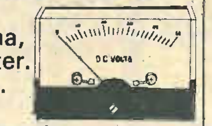

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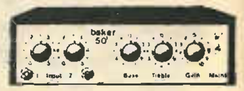

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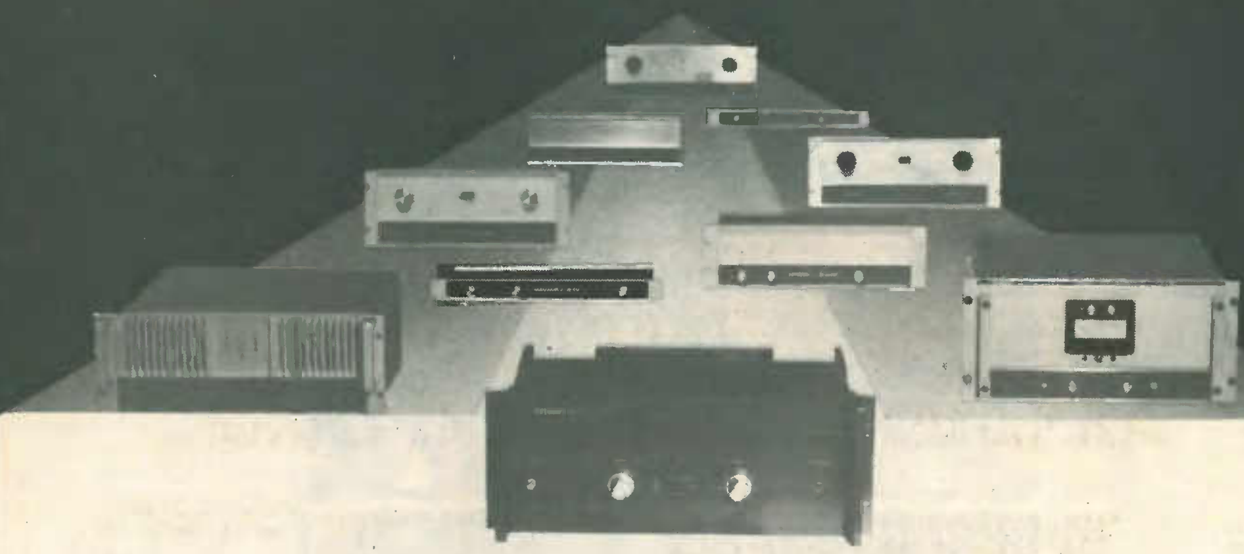

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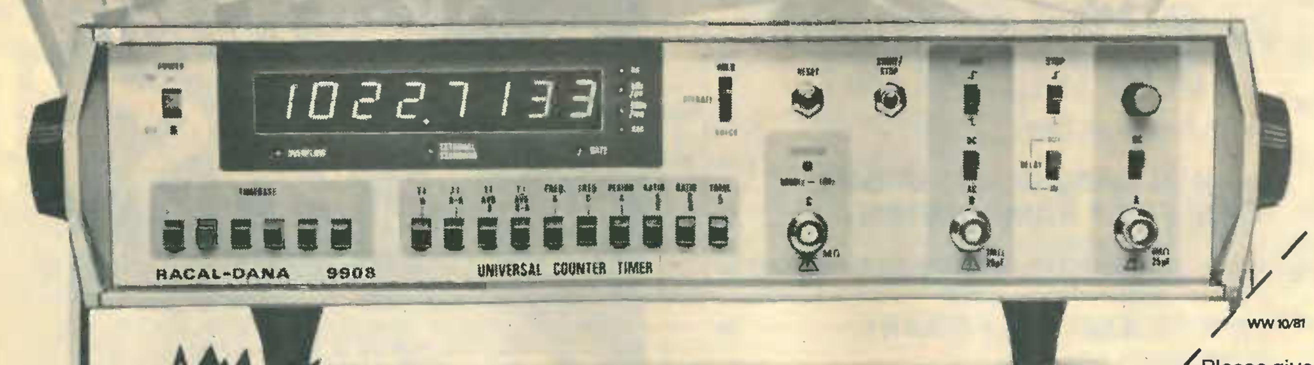

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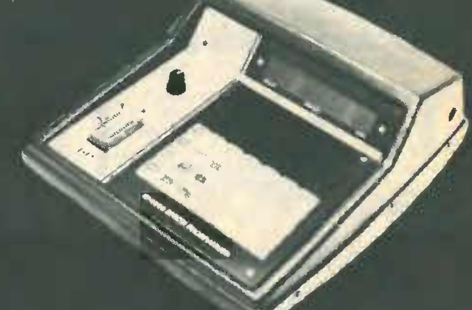

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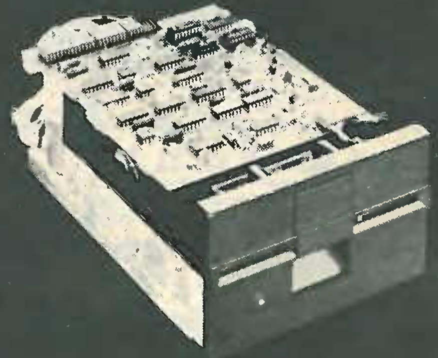

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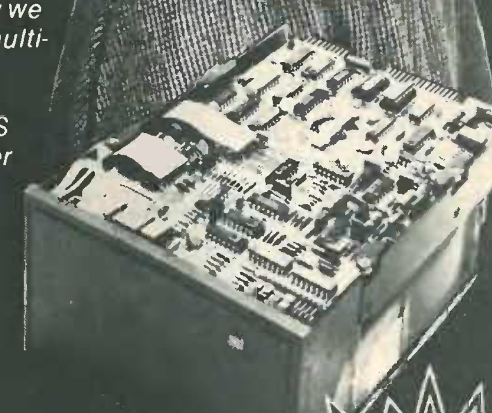

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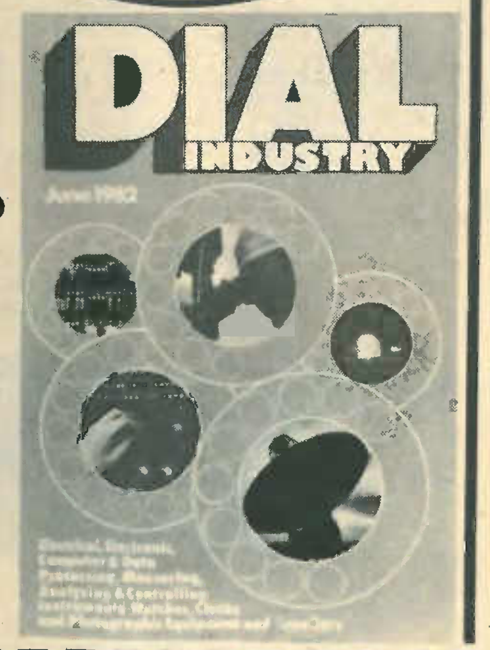

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SIEMENS

Siemens-Printer terminals are our business

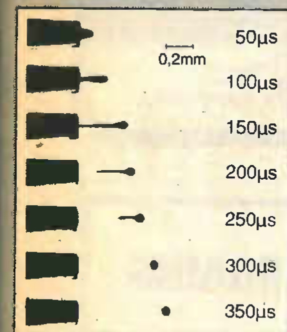
A quiet word...

The characteristics of an ideal printer, whether it be for use as a teleprinter or a printer terminal might be stated as an efficient, quiet machine that can produce high quality text on plain paper in clear, indelible script.

Siemens' Model 1000 teleprinter and the versatile PT80 series printer terminals possess these attributes and many more. Indeed they are designed to satisfy a variety of today's text and communication requirements.

Teleprinters and Data printers for the office

The conventional teleprinter, has been greatly refined over the years but some machines still need frequent maintenance checks and are too noisy for use in an open office. However, Siemens' development of printing mechanisms has allowed the noise producing parts to be reduced. This, together with the introduction of advanced electronic techniques have been built into the Model 1000 teleprinter.



This teleprinter offers many advantages over more conventionally styled teleprinters, the most significant features being quietness of operation and the extensive use of electronics to replace mechanical parts. The Model 1000 is so quiet it can be used in an open office, eliminating the necessity for a separate telex room housing a 'chattering' teleprinter.

The use of electronics also greatly reduces the need for preventive maintenance and regular periodic checks and, furthermore, makes for a smaller, more economical, easier-to-use machine. A prominent feature of the Model 1000 is the easily replaceable daisy wheel, and typewriter style red and black ribbon.

Variants for different purposes

Siemens Model 1000 teleprinters are made in a number of variants. For example, receive-only versions, with paper tape attachments, and magnetic tape memories for bulk message storage and editing are available. For use in special environments there is a Model 1000 V designed for military type applications.

Model 1000 S

The teleprinter Model 1000 S is an exciting new development using the latest technological advances. The versatility of the Model 1000 S is illustrated by the fact that it will produce both latin and non-latin script and switch between them when required e.g. Arabic, Greek, Cyrillic, Hangul or Farsi. The Model 1000 S is available with either a daisy-wheel, a needle-printer head, or an ink-jet. Optional items include a visual display unit and floppy disk message store.

Security – a growing problem

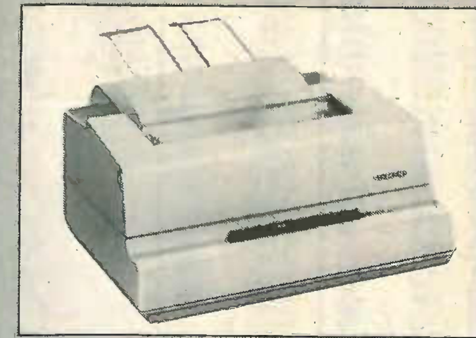
Industry, commerce, Government departments, and large international concerns and Institutions frequently have a communications security requirement. The Model 1000 CA (Cryptographical Application) gives the message originator and recipient protection from any electronic 'eaves-dropping'. This is done by encyphering and decyphering the message through a built-in cryptographic device. This machine has been designed to be compatible with all standard telegraph circuit options.

PT80 – a concept for today

The PT80 printer terminals are a result of many years' operational experience in both text and data communications. In essence, these machines are electronic terminals suited to a wide range of communications and data networks as well as process control.

The PT80 printer terminal uses either a 12 needle printing head for refined print quality, or alternatively the Siemens revolutionary ink-jet mechanism to achieve the ideal particularly in respect of minimal noise. The PT80 uses the ink-jet principle to attain a printing speed of up to 270 characters per second. The principle is featured very simply in the illustration on this page, with the droplet being ejected by means of a shockwave which causes a momentary increase in pressure in the nozzle.

What happens immediately afterwards in front of the nozzle orifice is shown in our illustration. The shockwave in the nozzle is generated by a piezo-electric transducer to which a voltage is momentarily applied. Siemens has ensured that ink is ejected only as and when needed.



Versatility

As with Siemens' Model 1000 teleprinters a number of PT80 variants are available to suit specific requirements. For example, there are receive-only machines with needle or ink-jet printing, a teleprinter (PT80-5) and a variant with paper tape attachments for automatic send/receiver. There is also a wide variety of character sets and an extensive range of interface modules to suit most telecommunications and data peripheral requirements.

PT80-H

Also available is the PT80-H, designed to print airline-style tickets, multi-part forms and continuous pre-printed stationery. This machine has the ability to recognise the validity of each ticket by series and type and adjust the print format accordingly. It can also be fitted with an integral guillotine, so that forms can be cut to size as they are used.

Easy servicing

Again, as with the Model 1000 teleprinter, these printer terminals are based on the modular design concept. For example, plug-in modules of the PT80i enable a fast and therefore economical service support.

Operational flexibility

PT80 machines generally operate with seven-bit codes or alternatively the PT80-5 teleprinter variant uses the five-bit code. The standard terminals are suitable for operating at speeds of up to 600 baud, the teleprinter variant at up to 200 baud, and the PT80i up to 4800 baud. All the PT80 terminals satisfy the requirements for a flexible character set.

Notwithstanding their advanced specification, the PT80 range of printer terminals is compact and simple to use and along with the Model 1000 teleprinters they are perfect examples of 'quiet words from Siemens'.

For full information, cut out the coupon and send to Marketing Services Dept., Siemens Ltd., Siemens House, Windmill Rd. Sunbury-on-Thames, Middlesex TW6 7HS. Or telephone Sunbury-on-Thames (09327) 85691. Tx. 8951091.

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Salary offered will be very competitive. We are situated in an attractive part of north Hampshire within easy reach of several other major towns.

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(1328)

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The Job

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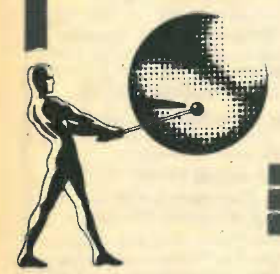

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PHILIP DRAKE ELECTRONICS
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PHONE: 01-221 1476

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ELECTRONICS ENGINEER

with experience in design and construction of digital frequency synthesis circuitry involving phase-lock techniques

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France
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Tel: 01-253 7611.

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with experience in design and construction of microwave phase-lock loops and phase drift monitoring systems of high accuracy

Candidates with relevant experience should request an application form and return it before October 23, 1981, to:

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38017 GRENOBLE CEDEX
France
Reference VC/17/LW

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