wireless world

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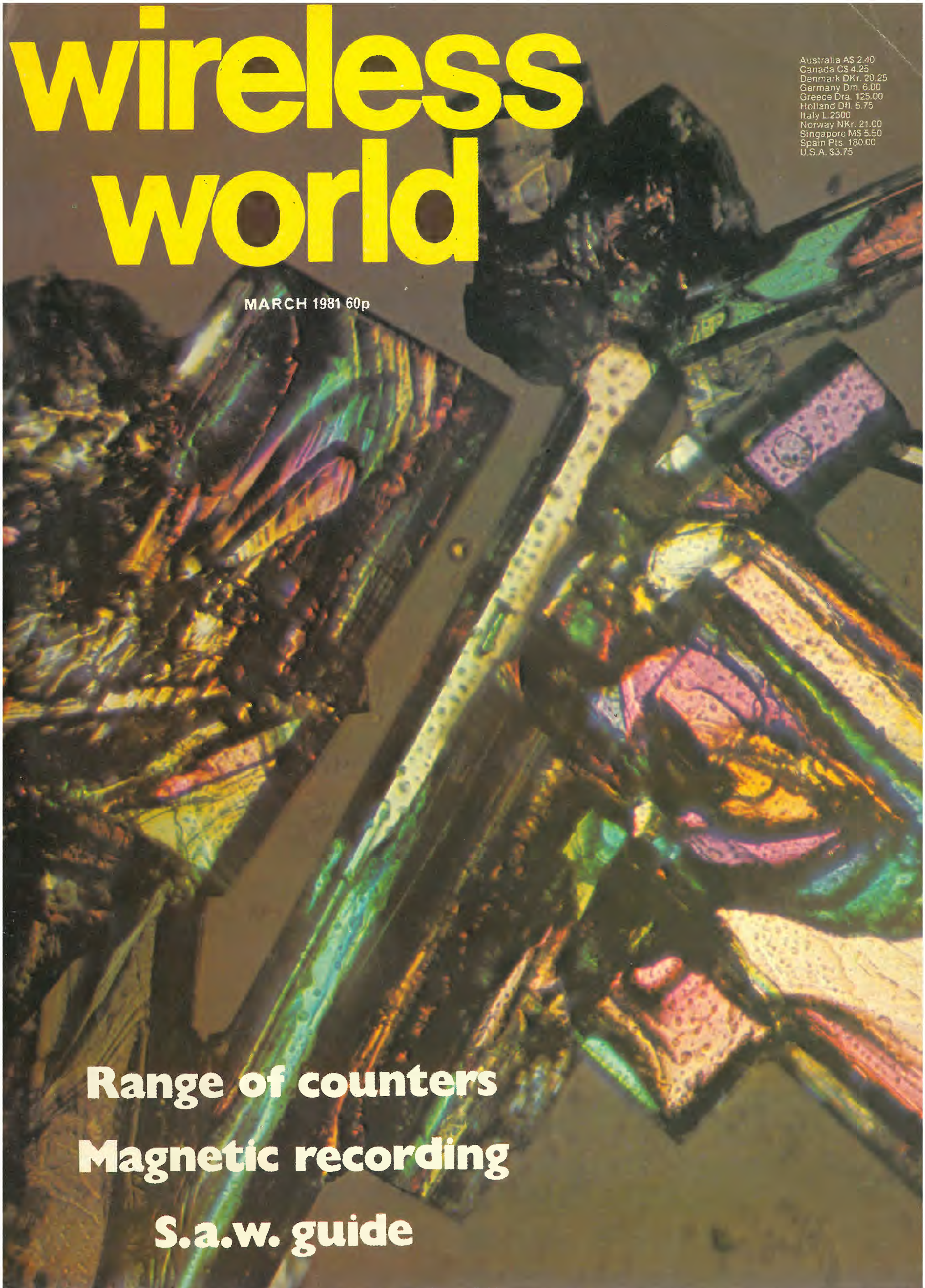
Canada C$ 4.25
Denmark DKr. 20.25
Germany Dm. 6.00
Greece Dra. 125.00
Holland Dfl. 5.75
Italy L.2300
Norway NKr. 21.00
Singapore M$ 5.50
Spain Pts. 180.00
U.S.A. $3.75

MARCH 1981 60p



Range of counters

Magnetic recording

S.a.w. guide

WIRELESS WORLD MARCH 1981 VOL 87 NO 1542

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Front cover shows piezo-electric Rochelle salt crystals, photographed by Paul Brierley in polarized light.

IN OUR NEXT ISSUE

Transient intermodulation distortion **is reviewed in a tutorial by Matti Otala that gives a comprehensive picture of distortions of this kind.**

David Read's modification to his ***active crossover networks*** **allows Quad electrostatic loudspeakers to be used with a common bass unit.**

Opto-electronic contact breaker **is a compact circuit which directly drives an electronic ignition unit and overcomes the heel wear and timing scatter inherent in mechanical contacts.**

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wireless world

ELECTRONICS/TELEVISION/RADIO/AUDIO

MARCH 1981 Vol 87 No 1542

Electronic Brokers~Europe's Premier Test Equipment Company offer

SUPERLATIVE USED EQUIPMENT CALIBRATED TO ORIGINAL SPECS

The Electronic Brokers GUARANTEE

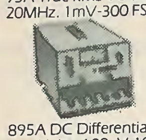
Unless otherwise stated, all test equipment sold by us carries a 12 month warranty; For VDUs and Teletypes we offer a 90 days warranty, and computers are offered with on site acceptance and diagnostic tests (which may qualify them for independent on-going maintenance). When you buy from Electronic Brokers you know the equipment is in 'top notch' condition. It is refurbished in our own service laboratories and checked to meet the original manufacturer's sales specifications. And it's serviced by our own highly qualified technicians.

A copy of our Trading Conditions is available on request.

....March 1981...... Latest Test Equipment.....

ANALOGUE VOLTMETERS AND MULTIMETERS

Boonton.
93A True RMS Voltmeter. Bandwidth 10Hz-20MHz. 1mV-300 FSD £375.00



Fluke.
883AB AC/DC Differential Voltmeter. 20Hz-100KHz. 1mV-1100V. Very high accuracy £975.00
895A DC Differential Voltmeter. 0-1100V. Null sensitivity. 100µV-100V. Very high accuracy £950.00

Hewlett Packard.
400E AC Millivoltmeter. 10Hz-10MHz. 1mV-300V DC o/p proportional to i/p £235.00
7563A DC Log Voltmeter / Amplifier. Single i/p range 360µV-100V O/P 0-1.1V @ 10mV/dB £400.00

Marconi.
TF2600 AC Millivoltmeter. 10Hz-5MHz. 1mV-300V £175.00
TF2603 RF Millivoltmeter. 50KHz-1.5GHz. 1mV-3V £525.00
TF2604 Electronic Voltmeter. AC 20Hz-1.5GHz. 300mV-1KV. DC 10mV-1KV. 0.2Ω-500MΩ £350.00
TF2606 DC Differential Voltmeter. 1mV-1100V on 5 decade dials £200.00

Philips.
PM2454B AC Millivoltmeter. 10Hz-12MHz. 1mV-300V £225.00

ANALYSERS

Biomation.
1650D 16 Channel Logic Analyser. Displays Timing, Map, hex, octal. Clock rate upto 50MHz. Supplied with display formatter and a Hameg HM312-8 oscilloscope £3600.00

General Radio.
1911A Sound and Vibration Analyser. C/W Graphic leve recorder. 4.5Hz-25KHz. 1/3 or 1/10 octave £1750.00

Hewlett Packard.
8407A/8412A Network Analyser. C/W 8600A/8601A sweeper. 0.1-110MHz £3500.00
141T-8552B-8554B Spectrum Analyser. 100KHz-1250MHz. + 10 to —122dBm. Variable percistence display £5400.00

Marconi.
TF2330 Wave Analyser. 20Hz-20KHz, 30µV-300V £395.00

Tektronix.
491 Spectrum Analyser. 10MHz-2GHz direct. Up to 40GHz with optional mixers. . £4150.00

BRIDGES & COMPONENT TESTERS

AVO/BPL.
CB154/4 Electrolytic Capacitance Tester. 0.1µF-1F. Accuracy 1% to 2% £515.00

Boonton.
63H Inductance Bridge. 0-110mH. Bridge frequency 5-500kHz £1950.00
280AP VHF Q Meter. Q 20 to 25,000. 210-610MHz £650.00

General Radio.
1607A Transfer Function and Immitance Bridge, 25-1500MHz. Direct reading . £750.00

Marconi.
TF1245-TF1246-TF1247 'Q' Meter and Oscillators. Q 5 to 500. 40KHz-300MHz £950.00
TF1313/2MI 0.25% LCR Bridge. 0.1µH-110H, 0.1pF-100µF. 3mΩ-110MΩ. 1 and 10KHz £325.00
TF2 700 1% LCR Bridge. 0.2µH-110H, 0.5pF-110µF, 0.01Ω-11MΩ. Battery operated £210.00

Rohde & Schwarz.
KRT (BN5100) Capacitance Meter. 1pF-100µF. 2.2-285KHz £395.00
LRT (BN6100) Inductance Meter. 1pH-100µH. 2.2-285KHz £395.00

Wayne Kerr.
B641 0.1% LCR Bridge. 0.002pF. 50,000µF, 200nH-5MH, 2mΩ-50,000MΩ, 20pυ-500υ £445.00
A60 Testmatic Dynamic Circuit Tester. Automatically checks upto 60 functions in 10 seconds on PCB's £895.00

CALIBRATION EQUIPMENT

Advance.
OFS2B Off Air Frequency Standard. O/P 1 and 10MHz £95.00

Fluke.
332A DC Voltage Calibrator. 0 to 1111.1110V. 0.1ppm resolution. 0.003% calibration accuracy £1750.00
760 Meter Calibrator. DC/AC Volts and current £2150.00

Hewlett Packard.
741B DC Voltage Source and AC/DC Differential Voltmeter £695.00

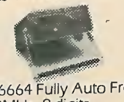
Tektronix.
191 Constant Amplitude Generator. 350KHz-100MHz £350.00

FREQUENCY COUNTERS

Advance.
TC15 & P1 Frequency Counter. DC — 500MHz, 9 digits £400.00

Fluke.
1900A-01 Frequency Counter. 5Hz-80MHz. 6 digits. Mains / battery operation £195.00
1912A Frequency Counter. 5Hz-520MHz. 7 digits £340.00
1925A Frequency Counter. 5Hz-125MHz, 9 digits, EMI proof case £375.00

Philips.
PM6604 Timer Counter. DC — 50MHz, 6 digits £150.00
PM6614 Frequency Counter. 10Hz-520MHz, 9 digits £450.00



PM6615 9 digit 1GHz Frequency Counter. 10mV sensitivity. Period, ratio and totalizing £695.00
PM6664 Fully Auto Frequency Counter. 10Hz-520MHz, 8 digits £250.00

DVM's AND DMM's

Datron.
1051 5½ Digit DMM. AC/DC Volts, Resistance. True RMS. 0.1µV resolution £900.00

Fluke.
8800A 5½ Digit DMM. AC/DC volts, resistance. 1µV resolution £495.00

Hewlett Packard.
34702A & 34740A 4½ digit DMM. AC/DC volts, resistance £225.00
3490A 5½ Digit DMM. AC/DC volts, resistance, 1µV resolution £495.00

Philips.
PM2514 3½ Digit Autoranging DMM. AC/DC volts and current, resistance £95.00
PM2517E 4 Digit Auto ranging DMM. AC/DC volts and current, resistance £115.00
PM2527 4½ Digit DMM. AC/DC volts, current and resistance. 10µV resolution. True RMS £400.00

Solartron.
A243 5½ Digit DMM. AC/DC volts, resistance. 1µV resolution £425.00

7055 Microprocessor DMM. Scale length 200,000. AC/DC volts, resistance 1µV resolution ... £850.00
7055 plus processor control and RS232 interface £1150.00
7065 Microprocessor DMM. Scale length 1,400.000. AC/DC volts, resistance .. £950.00
7065 plus processor control and RS232 interface £1250.00

Weston.
4440 3½ Digit DMM. AC/DC volts, DC current resistance £90.00

MULTIMETERS

Avo.
8 MK 5 Multimeter 20KΩ/volt £70.00
Full lead kit £5.25
Test Set Number 1. 20KΩ/volt, very robust £75.00
Full lead kit £5.25
EA113 Electronic Multimeter. High input impedance £115.00

S.E.I.
Super 50 Selectest. 20KΩ/volt £77.00

OSCILLOSCOPES

Hewlett Packard.
130C XY Oscilloscope. DC — 500KHz. 200µV-20V/div £250.00
1707A Dual Trace Portable Oscilloscope. DC — 75MHz Full delayed sweep. 30 day warranty £550.00

1707B Dual Trace Portable Oscilloscope. DC — 75MHz. 10mV-5V/div. Full delayed and intensified sweep. Trigger hold off £725.00

Philips.
PM3212 Dual Trade Portable Oscilloscope. DC — 25MHz. 2mV-10V/div £625.00
PM3262 Dual Trade Portable Oscilloscope. DC — 100MHz. Full delayed sweep.... £1250.00

SE Labs.
SM121 6 Channel Monitor. 12" crt. Internal sweep £395.00

Tektronix.

432 Dual Trace Portable Oscilloscope. DC — 25MHz. 1mV sensitivity. 20nS-5S. Auto setting mains input 100-250V £510.00
465 Dual Trace Portable Oscilloscope. DC — 100MHz. 5mV-5V/div. Full delayed sweep £1195.00
521A PAL Vectorscope. Measures Luminance Amplitude, Chrominance Phase, Chrominance Amplitude, Differential Phase and Gain. As new condition £2650.00
7704A Main Frame C/W 7A26, 7B80, 7B85. Dual Trace DC — 200MHz £3550.00

Telequipment.
DM64 Dual Trace B1-Stable Storage Oscilloscope. DC — 10MHz. Max writing speed 250div/mS £715.00
D67A Dual Trace Portable Oscilloscope. DC — 25MHz. Full delayed sweep £575.00
D75 Dual Trace Portable Oscilloscope. DC — 50MHz. Full delayed sweep £715.00

VU-Data.
PS940 Dual Trace Portable Battery Operated Oscilloscope. DC — 20MHz 10mV-20V/div. Weight 9.5lbs £395.00

OSCILLOSCOPE PROBES

EB90 X1 Probe. 1.2 mtr length DC — 20MHz £9.00
EB91 X10 Probe. 1-2 mtr length DC — 100MHz £11.00
EB95 X1, X10 Probe. 1.2 mtr length DC — 10MHz or DC — 100MHz £15.00

.......March 1981..... Latest Test Equipment......

POWER SUPPLIES

Advance.
PMA50 0-15V preset @ 5A £45.00
MG5-60 5V @ 60A switching £160.00
MG5-20 5V @ 20A switching £120.00
MG5-10 5V @ 10A switching £95.00
MG24-12 24V @ 12A switching £130.00

RECORDERS

Bryans Southern.
40000 12 Channel UV Recorder Plus 2 off 40501 Galvo Amps. 6" paper £950.00

Hewlett Packard.
2FA XYY A3 Plotter. 2 Pens on Y axis. Internal X sweep. Electric paper grip £725.00
7035B XY A4 Plotter. 4mV-4V/cm X and Y. Electric paper grip £650.00

Philips.
PM8110 Single Channel Chart Recorder. 10mV-10V £195.00

Racal.
Store 4 Tape Recorder. 4 Channels FM, DC — 20KHz £1650.00

S. E. Labs.
3006/12 12 Channel UV Recorder. Grid and timing lines. 6" paper £550.00
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 £1950.00
MC611-S4L. 4 Channel Chart Recorder; 100µV-2V. 400-800-1600mm/Hour/Minute £700.00

Yokagawa.
3047 2 Channel Chart Recorder. 0.5mV-100V. 2cm/hr-60cm/min £450.00

SIGNAL SOURCES

Avo.
HF135 AM Signal Generator. 100KHz-240MHz. 1µV-100mV into 75Ω. AM 0-30% @ 1KHz £190.00
HF136 AM/FM Signal Generator. 4.0-120MHz. 0-100mV in 20dB steps plus fine control. O/P Z = 75Ω £410.00

Hewlett Packard.
203A Variable Phase Oscillator. 0.005Hz-60KHz. 0-360° £450.00
3320B Frequency Synthesizer. 0.01Hz-13MHz. + 26.99 to —69.99dBm. O/P Z = 50Ω £1675.00
606B AM Signal Generator. 50KHz-65MHz. AM 0-95% £850.00
612A AM/PM UHF Generator. 450-1230MHz. 0.1µV-0.5V £775.00
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

4204A Digital Oscillator. 10Hz-1MHz. 1mV-10V into 600Ω. Frequency accuracy ±.2% or ±0.1Hz £695.00
8640B opt 002 Phase Locked Digital Readout AM/FM Generator. 500KHz-1024MHz £3650.00
8600A + 8601A RV Sweeper + Markers. 100KHz-110MHz £1500.00
8690B RF Sweeper Main Frame C/W the following RF Modules: 8698B 400KHz-110MHz. 8699B 0.1-4GHz. 8694A 8-12.4GHz £2500.00

Marconi.
TF144H/4 AM Signal Generator. 10KHz-72MHz. 2µV-2V £750.00
TF995B/2 AM/FM Signal Generator. 200KHz-220MHz. 1µ-200mV £675.00
TF995B/5 AM/FM Signal Generator. 200KHz-220MHz. 1µV-200mV. Narrow deriation for mobiles £695.00
TF1370A Wide Range RC Oscillator. 10Hz-10MHz. Sine wave, square wave upto 100KHz £275.00
TF2002B AM/FM Signal Generator. 10KHz-88MHz. 0.1µV-1V. 20Hz-20KHz Mod frequency £1200.00
TF2005R 2 Tone Signal Source. 20Hz-20KHz. 0-111dB in 0.1dB steps £295.00
TF2100 AF Oscillator. 20Hz-20KHz. 0.05% distortion £150.00

Philips.
PM5108 Function Generator. 1 Hz-1MHz. Sine, square and triangle. Triangle O/P £250.00
PM5127 Function Generator. 0.1HZ-1MHz. Sine, square and triangle £395.00
PM5167 Function Generator. 1MHz-10MHz. Sine, square, ± pulse, ramp and triangle £500.00
PM5509 CTV PAL Pattern Generator. Covers Bands I, III, IV and V. 10 Test Patterns. O/P 0-10mV into 75Ω. Video O/P £435.00
PM6456 Stereo Generator. Separate L and R Signals. Carrier frequency 100MHz ± 1%. RF O/P 3mV pk-pk £250.00

Radiometer.
SMGiC Stereo Generator. Separate L and R Signals. Carrier frequency 100MHz. RF O/P 10µV-100mV into 75Ω £375.00

Schlumberger.
4000C Frequency Synthesizer. 300Hz-520MHz. Max O/P 1V into 50Ω. 0-139.9dB attenuator £1250.00

Tektronix.
2101 Pulse Generator. 2.5Hz-25MHz. + and — O/P £375.00

Telonik.
2003 RF Sweeper System. 1-300MHz. 5 and 10MHz markers £950.00

TRANSMISSION MEASURING EQUIPMENT

Marconi.
OA2805A PCM Regenerator Test Set. Comprised of TF2801/2, TF2802/2 and TF2823. For use on systems at 1.536 and 2.048Mbits £2700.00
TF2807A PCM Multiplex Tester. Comprises Noise / Sinewave source and level meter £1500.00
TF2332 AF Transmission Test Set. 20Hz-20KHz £425.00
TF2333 MF Transmission Test Set. 30Hz-560KHz £600.00

Siemens.

D2040 Selective Level Analyser and Voltmeter. 10Hz-60KHz. —110dB to + 50dB. 3µV-300V Lin and log indication. 5 digit frequency readout £1200.00
D2072 + W2072 Level Meter and Oscillator. 50KHz-100MHz. —110dB-0dB. Receive bandwidth 3.1 and 10KHz £2200.00
D2074 + W2074 Level Oscillator and Level receiver. 50KHz-100MHz. —100 to + 20dBm. Receive bandwidth 80Hz, 3.1KHz and 10KHz. C/W G2006 to perform as a programmable system £2600.00
W2006 + D2006 Carrier Level Test Set. 10KHz-17MHz. —100 to + 10dB £1650.00
W2007 + D2007 Carrier Level Test Set. 6KHz-18.6MHz. —120 to + 20dB .. £1800.00
D2021 + W2021 + G2021 Carrier Level Test Set. 10KHz-25MHz. —100 to 10dBm £1700.0000

Wandel and Golterman.
PF-1 Digital Error Rate Measuring Set. Consisting of PFM-1 Digital Error Rate Meter and PFG-1 Pettern Generator £2490.00
PSO-5 and PMO-5 Level Measuring Set. 10KHz-36MHz. —110 to + 20dB. C/W AZD-1 Scale expander £2050.00
SPM-6 and PS-6 Level Measuring Set. 6KHz-18.6MHz. —110dB to + 20dB. Mains / battery operation £2150.00
Andimat (PSM-4 2MHz Automatic PCM Testing System) composed of the following PCM-1 PCM Test Set, PDA-64 PCM Signalling Analyser, PMD-1 Digital Level and Noise Measuring System, PDG-1 Digital Signal Generator, PDA-1 PCM Digital Signal Analyser, PSM-4 Level Measuring Set, MU-4 Scanner, R1-1 computer interface... £6500.00

SOUND LEVEL METERS

Bruel and Kjaer.
2203 Sound Level Meter. C/W 4145 1" Microphone. A, B, C, weighting..... £395.00

General Radio.
1933 Sound Level Meter C/W 1935 Cassette Data Recorder. 10-130dB. ABC weighting Octave band filters at 10 centre frequencies from 31.5Hz to 16KHz £995.00
1981 Sound Level Meter. 70-120dB. Digital and analogue reading. Peak hold. A weighting £300.00
1983 Sound Level Meter. 70-12db. A weighting £195.00

MISCELLANEOUS

Avantek.
UA101 RF Modular Amplifier. 5-500MHz, 10.5dB gain £39.00
UA103 RF Modular Amplifier. 10-500MHz, 10dB gain £39.00
UA301 RF Modular Amplifier. 1-400MHz, 7dB gain £39.00
UA305 RF Modular Amplifier. 2-500MHz, 13dB gain £39.00

Dymar.
1581 RF Power Meter. DC — 500MHz. 30mW-100W £350.00

Ferrograph.
RTS2 Recorder Test Set Measures Wow & Flutter, Distortion, Gain £345.00
RTS2 + ATU2 Recorder Test Set. C/W Auxiliary Test unit. Superb condition £650.00

Hewlett Packard.
335E Programmable Attenuator. 0-12dB. DC — 10GHz, 50Ω £75.00
432A Power Meter with 478A Thermistor Head. 10MHz-10GHz. 100µW-10mW £450.00
4329A Insulation Resistance Meter. Range 500KΩ to 2 × 10^{16}Ω £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
TF893A AF Power Meter. 1mW-10W£185.00

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

Radiometer.
BKF10 Automatic Distortion Analyser. 20Hz-20KHz. 0.02%-10% £1100.00

Rifa.
EVV 300 Microprocessor Development Kit. Based on Motorola 6800 TTY interface £495.00

Tektronix.
TM515 Main Frame C/W FG504. 0.001Hz-40MHz Function Generator and 2 off PS503A Power Supplies £1250.00
TM515 Main Frame C/W SC502. 15MHz oscilloscope, FG503 1.MHz-3MHz Function Generator, DM502 3½ digit DMM, DC503 100MHz counter £1495.00

WW-200 for further details

Electronic Brokers is Europe's largest specialist in quality, second user test equipment, computers and associated peripherals. Established 14 years ago, we have pioneered the second user concept in Britain, and many overseas territories.

To support our growth we have a skilled team. This includes trained sales staff, whose role is not only to sell, but provide a helpful information service to our many customers. Backing this team is our own service laboratory where technicians monitor each item of equipment we sell. Our maxim is service, and those who have dealt with us will know that we endeavour to always live up to our reputation.

Wherever your business is — UK, Europe, in fact anywhere in the world, the Electronic Brokers Organisation is just a telephone or telex call away!

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ADD 15% VAT TO ALL PRICES
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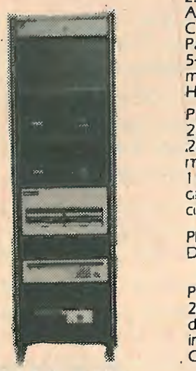
SUPERLATIVE USED COMPUTER EQUIPMENT~AS NEW CONDITION

March 1981 ....Latest Computer Equipment

PDP11 SYSTEMS



DEC

PDP11/34A System complete with 256KB MOS memory DL11W Asynchronous Interface / Line Clock, KY11-LB Programmers Panel, VT100 Terminal, 2 × RL01 5-meg Disk Drives, DZ11A 8-Line multiplexer, FP11A Floating Point, H960 Cabinet. **£17,500.00**

PDP11/34A System complete with 256KB MOS memory, 2 × RK07 28-meg Disk Drives, TMB11 magnetic tape station, DZ11 16-line multiplexer, Data system cabinet and expander box, LA36 console terminal. **£27,750.00**

PDP11/60 System with 2 RK07 Disk Drives, choice of console **P.O.A.**

PDP11/03 System (desk-mounted) 20KW MOS, RX11 dual floppy disk drive and control, 3 serial interfaces, LA180 Printer, VT100 Console terminal. .... **P.O.A.**

PDP11 C.P.U.s

DEC.

PDP11/03-SD Processor in 3½" chassis complete with 16KW MOS memory and E1S option.
BRAND NEW — ONLY **£1495.00**

PDP11/04 Processor in 10½" chassis complete with 16KW MOS memory and DL11W Interface.
BRAND NEW (can be enhanced to 28KW) **£4500.00**

PDP11/34 in 10½" chassis complete with 69KB MOS memory **£5500.00**

PDP11/34 Processor (non-A) in 10½" chassis complete with 128KB MOS memory, DL11W Interface and programmers front panel **£5900.00**

PDP11 MEMORY

DEC.

MF11L 8KW core memory c/w 9-slot system unit	**£975.00**
MM11L 8KW core memory	**£795.00**
MM11LP 8KW parity core memory	**£750.00**
MM11DP 16KW core (ex DEC-maintained 11/34 systems). BARGAIN OFFER — ONLY	**£395.00**
MM11UP 16KW Parity core memory complete with 9-slot system unit	**£1500.00**
MM11UP 16KW parity core memory (pre-requisite MF11UP)	**£1250.00**
MS11JP 16KW MOS memory	**£675.00**
MSV11C 16KW MOS memory (LSI11)	**£495.00**
MJ11-BE 64KW 11/70 memory	**£4500.00**
MK11-BE 64KW MOS memory for 11/70	**£3950.00**
MS11-MB 256KB ECC MOS memory for 11/44 (NEW)	**£3942.00**

DISKS

DEC.

RK05J add-on disk drive 2½ meg. Exchangeable cartridge type. From **£1850.00**

RK06 Add-on disk drive 14 meg. Free standing. From	**£3250.00**
RK07-ED Add-on Disk drive 28 meg.	**P.O.A.**
RL01-AK Add-on Disk drive 5 meg.	**£1975.00**
RP02 20 meg add-on disk drive	**£1250.00**
RP04 add-on 88 meg disk drive	**£8000.00**
RP06 add-on 176 meg disk drive	**£15,950.00**
RX11-BD Dual floppy disk drive and control	**£1800.00**

Controllers usually available for above drives.

MAGNETIC TAPE

DEC.

Available from time to time — TU10, TU45, TE16 TS03 etc.

PDP11 OPTIONS

DEC.

AA11D VT01 Controller with 4-slot System Unit	**£125.00**
DD11A 4-slot System Unit	**£125.00**
DD11B 4-slot System Unit	**£150.00**
DD11-DK 9-slot backplane	**£465.00**
DL11 Serial Interface	**£250.00**
DL11WA/B Serial Interface/Line Clock	**£395.00**
DR11C General Purpose Interface	**£250.00**
DRV11B 16-bit Parallel Interface	**£225.00**
DZ11A 8-line E1A MUX	**£1395.00**
FP11A Floating Point	**£1750.00**
KL11 TTY Interface	**£150.00**
KW11L Real Time Clock	**£150.00**
KW11P Programmable Clock	**£345.00**
M105 Device Selector	**£30.00**
M792 ROM Diode Matrix	**£195.00**
M920 Unibus Connector	**£12.00**
M9301-YB Bootstrap	**£325.00**

POWER SUPPLIES

DEC.

H720 Power supply for BA11 Expander Box, BRAND NEW SURPLUS **£175.00**

PDP8A C.P.U.

DEC.

PDP8A Processors, systems and add-on memory usually available.

PDP8A MEMORY

DEC.

MM8AA 8KW Core	**£750.00**
MM8AB 16KW Core	**£995.00**

PDP8E CPU, MEMORY, OPTIONS

DEC.

DP8EB Communications Adaptor	**£395.00**
KA8E Positive I/O Bus	**£95.00**
KD8E Databreak	**£145.00**
KL8E Asynchronous Interface	**£250.00**
KP8E Power fail/auto restart	**£95.00**
MM8E 4KW Core memory stack	**£350.00**
TA8E Dual Cassette Drive and Controller	**£525.00**
VT8E Set graphics Control Modules	**£250.00**

TERMINALS

DEC.

LA34 DECwriter 1V 30cps desk-top terminal, 132 columns, upper/lower case ASCII **£495.00**

LA36 DECwriter 11 Keyboard Printer Terminal. The Terminal that has become an industry standard, with 132 column upper/lower case printing and switch-selectable speeds of 10, 15 & 30cps. Available with either 20mA or RS232 interface **£750.00**

VT52 24 lines by 80 columns, 7 × 7 dot matrix, upper and lower case ASCII, includes numeric keypad, 20mA or RS232 interface. **£450.00**

VT100 — Also available from time to time, DEC's latest 132 column display terminal. Please enquire for latest price and delivery **£750.00**

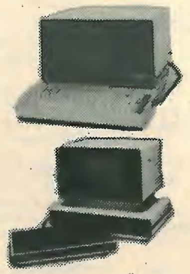
VDUs

Hazeltine.

H1000 The low, low priced teletypewriter-compatible video display terminal, offering your choice of transmission speeds up to 9600 baud as well as parity generation and checking. 12 × 80 display upper case ASCII, RS232 interface, choice of baud rates SUPER VALUE **£199.00**

VDUs

Hazeltine.



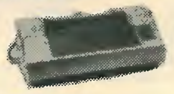
H2000 Superb spec. including full XY cursor addressing and edit facility, 27 × 74 display, upper case ASCII, RS232 interface, switch-selectable baud rates **£299.00**

H2000C NOW ALSO AVAILABLE with 25 × 80 line format and C-MOS logic **£375.00**

Modular One. Now with upper/lower case, XY cursor addressing, 24 × 80 line display, dual intensity detachable keyboard, choice of 8 transmission rates up to 9600 baud ... **£399.00**

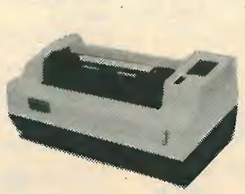
PRINTERS

Centronics 101.



Heavy Duty Matrix printer with 64 ASCII upper case character set. 165 cps operation. 132 print positions with adjustable tractor feed. Parallel input. ONLY..... **£495.00**

GP80 GRAPHICS.



BRAND NEW — LOW COST MATRIX PRINTER IDEAL FOR MICROPROCESSOR USERS SUCH AS HOBBYISTS & EDUCATIONALISTS OR ANY LOW-BUDGET APPLICATION

- ★ Full upper/lower case ASCII PLUS GRAPHICS Mode
- ★ 80-column printing with adjustable tractor feed
- ★ Standard and double-width characters (12 cpi and 6 cpi)
- ★ 30 cps print speed with 1-line buffer
- ★ Standard parallel (Centronics-type) interface
- ★ Optional interfaces available for RS232, IEEE488, Tandy, PET, Apple II
- ★ ONLY .... **£199.00** (Mail order total ..... £240.35)

GE Terminet 1200.
RO printer, 80 columns, tractor feed, upper/lower case, ASCII, 20mA Interface **£325.00**

Hazeltine.
Thermal printer, 80 columns, 30 cps silent RO with parallel TTL input **£395.00**

Tally 1602.
Matrix printer, parallel input, upper/lower case, tractor feed, as new **£995.00**

Teletype.
ASR 33 input/output terminal incorporating paper tape punch and reader. 64 ASCII upper case character set. 110 baud operation, even parity keyboard choice of RS232 or 20mA interface.
NOW ONLY **£595.00**

Options.

ICL-type keyboard	**£50.00**
8th level marking	**£25.00**
Remote reader control	**£50.00**
Reader step	**£20.00**
Auto reader	**£25.00**
Pedestal	**£30.00**

March 1981.........Latest Computer Equipment

PRINTERS

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**

Termiprinter 7075.
Impact printer, upper/lower case, pin feed, RS232 .... **£325.00**

Texas.
725 portable terminal with acoustic coupler **£625.00**
733 ASR terminal utilising high performance twin cassette drive for fast time-saving transmission and off-line storage.

- ★ Silent operation
- ★ Switch-selectable transmission speeds 10, 15, & 30 cps
- ★ Full tape editing capability
- ★ High-speed duplication and rewind
- ★ Standard RS232 interface **£1375.00**

745 LIGHTWEIGHT (13lbs) terminal with integral acoustic coupler, 10/30 cps, carrying case **£750.00**

MONITORS

EMI.

15in. diagonal tube, integral power supplies, accepts composite or separate video input. BRAND NEW SURPLUS......... **£100.00**

Ball Miratel TTL15.
15in. diagonal green phosphor tube, integral power supply. Requires separate horizontal and vertical video input. BRAND NEW SURPLUS **£75.00**

FLOPPY DISK DRIVES

NEW

Shugart SA400.
5¼" MINI-FLOPPY (NEW)

- ★ Capacity (unformatted): 110Kb (single density) 220Kb (double density)
- ★ Transfer rate: 125/250 kilobits/sec.
- ★ Average access time: 463ms.
- ★ Read/Write and Write Protect electronics
- ★ Power requirements +5VDC + 12VDC Dimensions 5¾" × 3¼" × 8", weight 3lbs

Supplied complete with technical manual.
Application notes are available on interfacing the SA400 to Intel 8080A, Motorola etc **£195.00**

Shugart SA450.
Double-sided, double density minifloppy providing 440KB unformatted storage capacity, yet the same compact size and weight as the SA400 .... **£299.00**

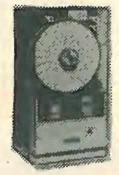
Shugart SA801.
8" FLOPPY (NEW)

- ★ Capacity (unformatted): 400Kb (single density) 800Kb (double density)
- ★ Transfer rate: 250/500 kilobits/sec.
- ★ Average access time: 211ms
- ★ Read/Write and Write Protect electronics
- ★ Power requirements +24VDC +5VDC —5VDC Dimensions 4⅝" × 9½" × 14¼"

Supplied complete with technical manual **£395.00**

PAPER TAPE PUNCHES

Digitronics.
P120 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**

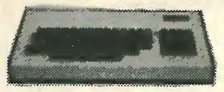


Facit 4070.
The top quality punch that has become an industry standard. Asynchronous 75cps operation. Adjustable for punching 5, 6, 7 or 8 level tape. Self-contained desk-top unit incorporating supply and take-up spools, chad box, and TTL-compatible control logic . **£650.00**

KEYBOARDS

George Risk.
New ASCII Keyboards
New Low Prices

NEW



KB771 Superb 71-station ASCII keyboard incorporating separate numeric/Cursor control pad and installed in custom-built steel enclosure with textured blue enamel finish. Ideal for the VDU builder. Case dimensions: 17¼" × 7½" × 3⅜".
Total weight 4kg. PRICE .. **£85.00**
Mail Order Total £103.50

KB756 56-station ASCII keyboard mounted on P.C.B.	**£39.50**	Mail Order Total £47.15
KB756MF as above, fitted with metal mounting frame for extra rigidity	**£45.00**	Mail Order Total £53.48
K8710 10-key numeric pad, supplied with connecting cable	**£8.00**	Mail Order Total £9.78
KB701 Plastic enclosure for KB756 or KB756MF	**£12.50**	Mail Order Total £15.24
KB702 Steel enclosure for KB756 or KB756MF	**£18.00**	Mail Order Total £23.00
KB2376 Spare ROM encoder	**£12.50**	Mail Order Total £15.24
KB15P Edge connector for KB756 or KB756MF	**£3.25**	Mail Order Total £4.31
DC-512 DC convertor to allow operation at 5V only (plugs in to P.C.B.)	**£7.50**	Mail Order Total £9.20
DB25S Mating connector for KB771	**£4.25**	Mail Order Total £5.46
PERK 56-station ASCII keyboard for PET complete with PET interface, built-in power supply and steel enclosure	**£95.00**	Mail Order Total £115.00

Discounts available for quantities.

Hours of Business: 9 a.m.-5 p.m. Mon-Fri.
Closed lunch 1-2p.m.
ADD 15%VAT TO ALL PRICES
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Model — M600

- ★ POWER RESPONSE DC — 20KHz ± 1dB.
- ★ OUTPUT POWER IN EXCESS OF 1.5kW INTO 2.75 Ohm LOAD (CONTINUOUS R.M.S.).
- ★ D.C. OUTPUT 20 AMPS AT 100 VOLTS OR 2KVa.
- ★ HARMONIC DISTORTION LESS THAN 0.05% DC-20KHz AT 1kW INTO 6 OHMS
- ★ PLUG-IN MODULES: CONSTANT VOLTAGE/CURRENT, PRECISION OSCILLATORS ★ UNIPOLAR AND BIPOLAR DIGITAL INTERFACES, FUNCTION GENERATORS, AND MANY OTHERS.
- ★ OUTPUT MATCHING TRANSFORMERS AVAILABLE TO MATCH VIRTUALLY ANY LOAD.
- ★ FULL OPEN AND SHORT CIRCUIT PROTECTION GUARANTEED STABLE INTO ANY LOAD.
- ★ TWO UNITS MAY BE CONNECTED TO PROVIDE UP TO 4kW.
- ★ INTERLOCK CAPABILITY FOR UP TO EIGHT UNITS.
- ★ 3-YEAR PARTS AND LABOUR WARRANTY.

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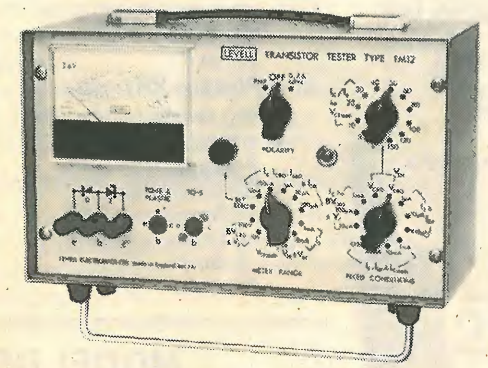

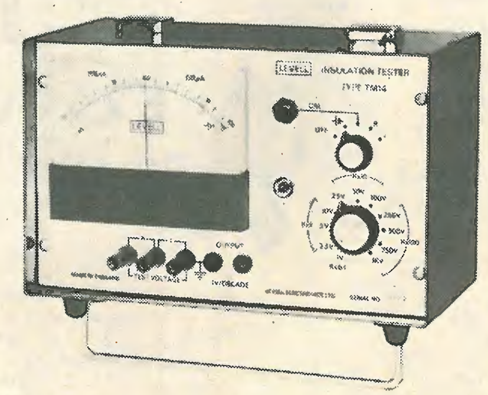

Tests bipolar transistors, diodes and zener diodes. Measures leakage down to 0.5 nA at 2V to 150V. Current gains are checked from 1µA to 100mA. Breakdown voltages up to 100V are measured at 10µA, 100µA and 1mA. Collector to emitter saturation voltage is measured at 1mA, 10mA, 30mA and 100mA for I_C/I_B ratios of 10, 20, 30. The instrument is powered by a 9V battery.

TRANSISTOR RANGES (PNP OR NPN)

I_{CBO} & I_{EBO}:	10nA, 100nA, 1µA, 10µA and 100µA f.s.d. acc. ±2% f.s.d. ±1% at voltages of 2V, 5V, 10V, 20V, 30V, 40V, 50V, 60V, 80V, 100V, 120V, and 150V acc. ±3% ±100mV up to 10µA with fall at 100µA <5%+250mV.
BV_{CBO}:	10V or 100V f.s.d. acc ±2% f.s.d. ±1% at currents of 10µA, 100µA and 1mA ±20%.
I_B:	10nA, 100nA, 1µA . . . 10mA f.s.d. acc. ±2% f.s.d. ±1% at fixed I_E of 1µA, 10µA, 100µA, 1mA, 10mA, 30mA, and 100mA acc. ±1%.
h_{FE}:	3 inverse scales of 2000 to 100, 400 to 30 and 100 to 10 convert I_B into h_{FE} readings.
V_{BE}:	1V f.s.d. acc. ±20mV measured at conditions on h_{FE} test.
$V_{CE(sat)}$:	1V f.s.d. acc. ±20mV at collector currents of 1mA, 10mA, 30mA and 100mA with I_C/I_B selected at 10, 20 or 30 acc. ±20%.

DIODE & ZENER DIODE RANGES

I_{DR}:	As I_{EBO} transistor ranges.
V_Z:	Breakdown ranges as BV_{CBO} for transistors.
V_{DF}:	1V f.s.d. acc. ±20mV at I_{DF} of 1µA, 10µA, 100µA, 1mA, 10mA, 30mA and 100mA.

type TM12 **£160**

A logarithmic scale covering 6 decades is used to display either insulation resistance or leakage current at a fixed stabilised test voltage. The current available is limited to a maximum value of 3mA for safety and capacitors are automatically discharged when the instrument is switched off or to the CAL condition. The instrument operates from a 9V internal battery.

RESISTANCE RANGES

10M Ω to 10T Ω (10^{13} Ω) at 250V, 500V, 750V and 1kV.
1 M Ω to 1 T Ω at 25V, 50V and 100V.
100k Ω to 100G Ω at 2.5V, 5V and 10V.
10k Ω to 10G Ω at 1V.
Accuracy ±15% +800 Ω on 6 decade logarithmic scale.
Accuracy of test voltages ±3% ±50mV at scale centre.
Fall of test voltages <2% at 10µA and <20% at 100µA.
Short circuit current between 500µA and 3mA.

CURRENT RANGE

100pA to 100µA on 6 decade logarithmic scale.
Accuracy of current measurement ±15% of indicated value.
Input voltage drop is approximately 20mV at 100pA, 200mV at 100nA and 400mV at 100µA.
Maximum safe continuous overload is 50mA.

MEASUREMENT TIME

<3s for resistance on all ranges relative to CAL position.
<10s for resistance of 10G Ω across 1µF on 50V to 500V.
Discharge time to 1% is 0.1s per µF on CAL position.

RECORDER OUTPUT

1V per decade ±2% with zero output at scale centre.
Maximum output ±3V. Output resistance 1k Ω.

type TM14 **£170**

Optional extras are leather cases and mains power units. Prices are ex works, V.A.T. extra in U.K.

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GR1657 Digibridge LCR, auto, LED display	850
CINTEL	
277 Measures iron core inductances 0,01H-1000H (with a Q value not less than 2)	130
HEWLETT PACKARD	
4342A 'Q' Meter QLC complete	1250
MARCONI	
TF868A Universal Bridge	250
TF1313A Universal LCR Bridge 0.1%	375
WAYNE KERR	
B224 Wide range LCR Bridge	475
B500 Log LCR Bridge	225
B601 RF LCR Bridge (Detector and Oscillator not incl).	125
B641. Measures L/C/R/G Accuracy of 0.1%	450
Q801. Y parameter test set. Plus transistor adaptor unit	230
Cable Test Equipment	
MARCONI	
TF2333 Transmission Test set	575
HEWLETT PACKARD	
3556A For psophometric measurements from 20 Hz-20kHz. 0.1mV-30V input level	475
NEC	
TTS-378. Noise, level and VU measurement. Sensitivity -80dBm up to +20dBm	275
STC	
74216A Noise Generator CCITT	240
74261A Psophometer CCITT	475
TEKTRONIX	
1502 Portable TDR Cable Tester	2,725
WANDEL u. GOLTERMANN	
DLM-1. Send/receive system	1500
LDS-2. 200Hz-600kHz sender for measuring group delay and attenuation variations	3250
LDEF-2. Filters for DLM unit	250
Counter Timers	
HEWLETT PACKARD	
5300A/5303B DC-520 MHz 6 digits	210
5300A Display Module. 6 Digits. 3 × 10⁷	90
5302A DC-50 MHz. 100mV sens. Time interval. Period. Ratio. Totalise.	75
5303B DC-520 MHz. (Plug-on) 125mV sens. 50Ω	120
5308A 0-75 MHz. Universal Module. 50mV sens. 1MΩ	100
5267A Time Interval Plug-in 10ns	120
5345 DC-500 MHz Time Int. Ave. Burst Total Ratio	1225
10590A Adaptor converts 5245 Plug-ins to 5345	225
RACAL	
9024 10 Hz-600 MHz 7 + 1 digits	250
9835 6 Digit DC-20 MHz 10mV	100
9837 DC-80 MHz 6 digits	130
SYSTRON DONNER	
6053 9 Digit 20 Hz-3 GHz BCD O/P	850

	Prices from £
6054A/04 11 Digit 20 kHz-18 GHz BCD O/P	2800
Function Generators	
INTER-STATE ELECTRONICS	
F51A Multi-Mode. + and − offset: 0.0005 Hz to 10 MHz. 10/15V/50Ω	250
F55A Multi-Model 0.0025 Hz-10 MHz. 10V/50Ω. Ext. VGC. Burst O/P up to 100k bursts/sec	350
PHILIPS	
PM5127. 0.1 Hz-1 MHz. Sine/Square/Triangular/Pulse outputs. External sweep facility 30Vp. p max output	325
Logic Analysers	
HEWLETT PACKARD	
1601L Logic state analyser 12 channel display	250
1600A 16 channel 20 MHz clock MAP A & B store	1850
1607 16 channel 20 MHz clock (Display scope required)	1500
TEKTRONIX	
7D01F 16 channel up to 50 MHz clock MAP	2650
Mains Monitors	
COLE	
T1007 200-260V. 35-65 Hz Thresholds 10V, 50V, 100V, 200V	75
DATALAB	
DL019 Power line interface for transient recording	350
DL905 Digital Storage Unit DC-3 MHz 10mV	1055
DRANETZ	
606-3 Disturbance Analyser Avg, Sag/Surge	2625
GAY	
LDM Records +ve/−ve transients of 50ns on AC or DC Lines	1250
Modulation Meters	
AIRMEC	
409 3-1500 MHz. AM/FM	295
MARCONI	
TF2300A 1-1000 MHz. AM/FM	450

Oscilloscopes	Prices from £
ADVANCE	
OS1000A DC-20 MHz. dual trace	200
3300B Dual Trace DC-50 MHz 5mV/div. Dual Timebase	600
GOULD ADVANCE	
OS1000B DC-20 MHz Dual Trace X-Y TV Sync	250
HEWLETT PACKARD	
1703A Storage 1000Div/ms. DC-35 MHz. Dual trace Mains/Ext DC	1200
181A Storage 1000Div/ms DC-100 MHz Main frame only	650
182C DC-100 MHz Mainframe, large screen	525
MEDELEC	
M-scope 4 channel DC-100 kHz U/V Chart	1650
PHILIPS	
PM 3211 DC-15 MHz Dual Trace 2mV	425
PM3233 Dual Beam DC-10 MHz 2mV/div.	400
TEKTRONIX	
475 Dual Trace DC-200 MHz 2mV	1125
485 Dual Trace DC-350 MHz 50Ω 1 MΩ 250 MHz	2100
545B/1A1. DC-30 MHz. dual trace. Delayed timebase	325
585A/82. DC-80 MHz. dual trace 10 mV sensitivity	525
547/1A1. DC-50 MHz. dual trace DTB	525
547/1A4. DC-50 MHz. four trace DTB	625
7403N DC-60 MHz 3 Plug-in Mainframe	450
7704A DC-200 MHz. CRT Readout. Mainframe for 4 Plug-in	1200
TELEQUIPMENT	
D34 Dual Trace DC-15 MHz 2mV Mains/Batt	525
D75 Dual Trace DC-50 MHz Dual Timebase	600
D83 DC-50 MHz. Dual trace. Large 6½" CRT. Dual Time Base	650
Oscilloscope Plug-ins	
HEWLETT PACKARD	
1804A DC-50 MHz Four channel 20 mV-10V/div.	575
1825A Dual Timebase 50ns-1s/div.	525
1805A Dual Trace DC-100 MHz 5mV. 1MΩ/50Ω	550
TEKTRONIX	
Type R. Transistor R.T. tester. Pulse rate 120 pulses/sec. R.T. Less than 5 mμs	100
Type G. Differential amplifier, 100:1 CMR DC-20 MHz. 50 mV sensitivity	50
Plug-ins for 500 series	
1A1 dual trace Plug-in DC-50 MHz	225
1A2 dual trace Plug-in DC-50 MHz	180
1A4 four trace Plug-in DC-50 MHz	375
1A5 Differential Plug-in	175
Z Differential Plug-in	140
81 Adaptor Plug-in 1A Series to 580 Series	75
7A12 Dual Trace DC-105 MHz 5mV/div.	410
7A22 High gain diff. amp. 0.1 Hz-1 MHz 10μV	450
7A26 Dual Trace DC-150 MHz 5mV-5V/div.	525
7B53A Dual Timebase 5ns-5s/div.	550
Oscilloscopes (storage)	
TEKTRONIX	
549/1A1. DC-30 MHz. 5mV sensitivity. Dual trace. Storage scope, Writing speed: 5cm/μs with enhancement. Includes trolley	675
564B/3A6/2B67. DC-10 MHz. Dual trace 10mV sensitivity, split screen storage oscilloscope	750
466 Storage 1350 cm/μs Variable Persist DC-100 MHz	2225
7313 Split screen 4.9 cm/μs. DC-25 MHz (M/F for 3 Plug-ins)	1650
Phase Meter	
DRANETZ	
301A 5 Hz-500 kHz. Z in 100kΩ. Accuracy ±1° to ±2°. Analogue O/P	400
Power Meters	
DYMAR	
2081/100 True RMS. DC-500 MHz. 30mW-100W	425
HEWLETT PACKARD	
478A Thermistor Mount for 432A	90
435A 0.3μW to 100mW 5 MHz-18GHz	475
8481A Power Sensor for 435A	200
MARCONI SANDERS	
6460 10 MHz-40 GHz (Depending on Head)	300
6420 10 MHz-12.4 GHz 10mw	110
MARCONI	
TF2512 DC-500 MHz 0.5-30w 50Ω	130
TF 893A 10 Hz-20 kHz. 20μW-10W.	120
Power Supplies	
BRANDENBURG	
475R 10-2100V 5mA DC Stab.	150
FARNELL	
L30B 0-30V 1A DC Stab.	55
FLUKE	
415B 0±3100V 30mA 0.005% reg. Protected	350
ITT	
Power Lab. up to 30V Dual Supply	90
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TF2154/1 0-30V 1A. 0±15V 2A 0±7.5V 4A	60
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4701 5-7V o/p Power Pack	32
SORENSEN	
DCR 300-2.5 0-300V 2.5A DC Stab.	375
Pulse Generators	
DB ELECTRONICS	
150. I.C. pulse generator	50
EH RESEARCH	
122. 1 KHz-200 MHz 5V/50Ω RT 12ns	220
139(L). 10Hz-50 MHz 10V/50Ω RT 5ns	175
1221. Timing Unit 6 Channel 0-10 MHz 5V/50Ω RT 8ns	50
HEWLETT PACKARD	
214A 100V/50Ω. Double pulse O/P. W50ns-10ms. 10 Hz-1 MHz. 15ns RT	350
MARCONI	
TF2025 0.2 Hz-25 MHz ±10V/50V RT 7ns	350
Recorders and Signal Conditioning Equipment	
AMPEX	
PR2200 Instrumentation Recorder up to 16 channels. FM/DR. Record replay all speeds. 1" tape FM/DR I.R.I.G. DC-40 kHz FM. 100 Hz-300 kHz DR	6500
BRUNO WOELKE	
ME102B. Wow and flutter meter	75
ME102C. Wow and flutter meter	90

As New Ex Stock delivery

OSCILLOSCOPES

TEKTRONIX 465 DC-100 MHz Dual Trace 5mV-5V/Div 0.05μs-0.5s/Div Delayed T/B XY DC 4 MHz £1250

TEKTRONIX 475A DC-250 MHz Dual Trace 5mV-5V Div 0.01μs-0.5s Div Delayed T B XY DC 3 MHz £1950

These instruments sold with
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Bigger stock investment greater equipment range means wider choice

	Prices from £
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BS314 4 channel 1mV-10V 16 speeds	1650
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7046A Two pen A3 0.25mV-5V/cm	995
KUDELSKI	
Nagra 4.2 LSP Professional Audio Recorder (Batt optd)	1215
PHILIPS	
PM 8251 Single pen 10in chart 10mV-50V FS	450
RACAL	
Store 4. Uses D/4 inch magnetic tape. Will record 4 F.M. channels. Operates at 7 different speeds.	1675
S E LABORATORIES	
6150/6151 12 channel UV 1250 mm/s-25 mm/min 6 in chart	1400
994 6 Channel Pre-Amp ± 1% ± 1V o/p	450
6008 25 Channel μV 8 in 4m/sec to 25mm/min	895
SMITHS INDUSTRIES	
RE541.20 Single Pen. 0.5mV-100V FSD. 3-60cm/min and hour	350
YOKOGAWA	
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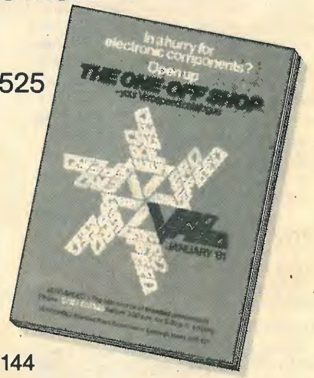

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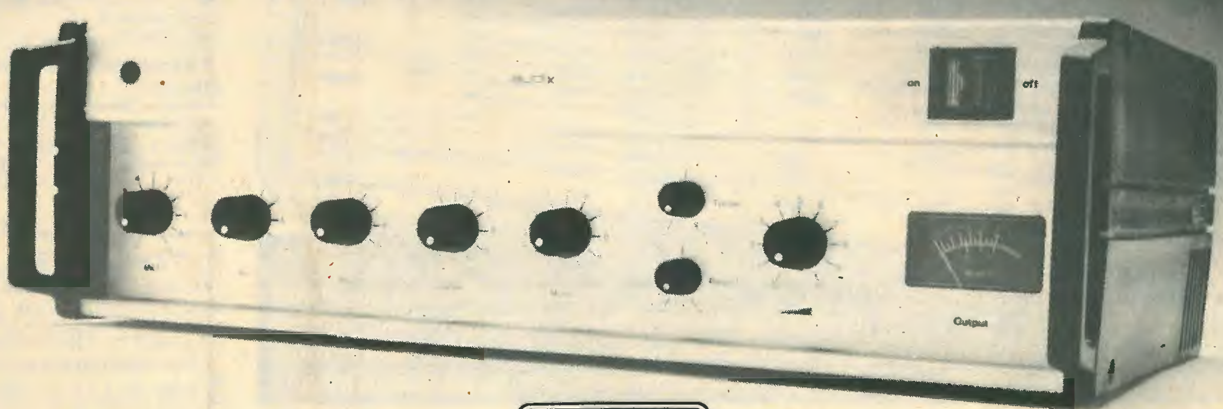

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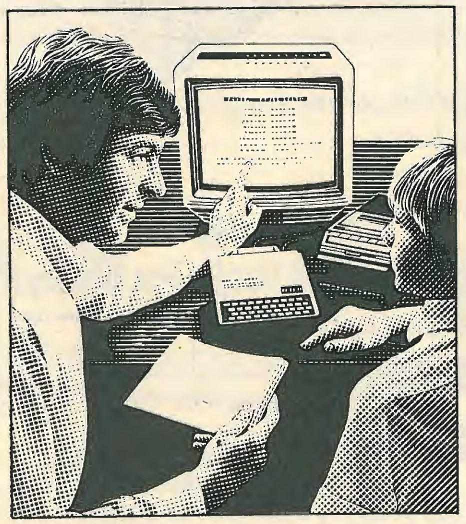

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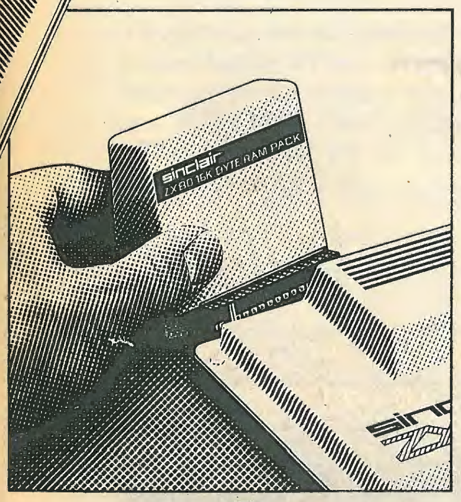
Kit or built – it's up to you

In kit form, the ZX80 is pleasantly easy to assemble, using a fine-tipped soldering iron. And you may already have a suitable mains adaptor – 600 mA at 9V DC nominal unregulated. If not, see the coupon.

Both kit and built versions come complete with all necessary leads to connect to your TV (colour or black and white) and cassette recorder. Plug in and you're ready to go. (Built versions come with mains adaptor.)

Now available for the ZX80... New 16K-BYTE RAM pack



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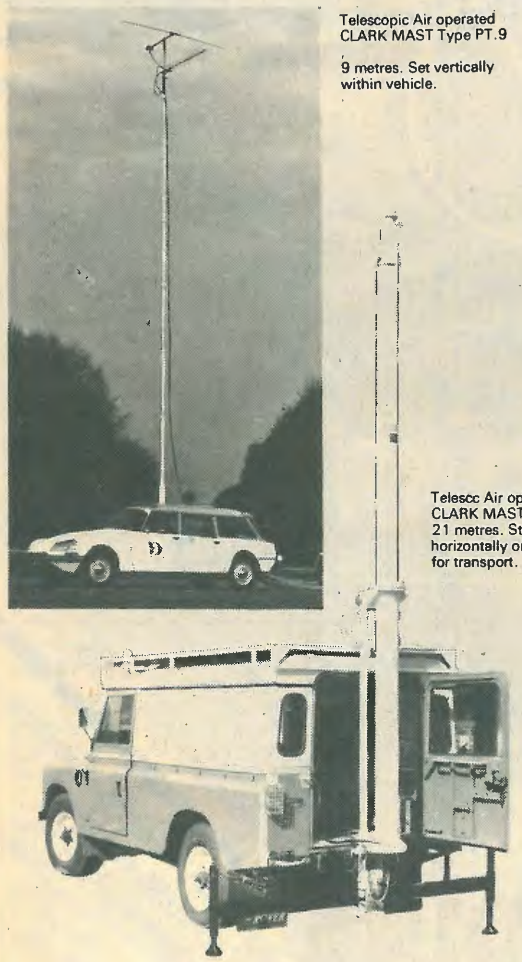

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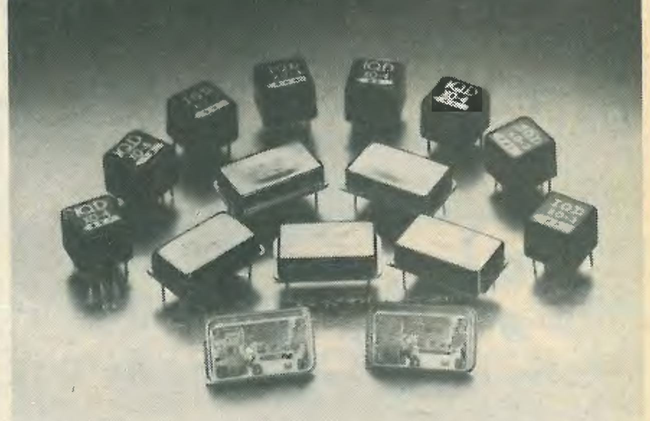

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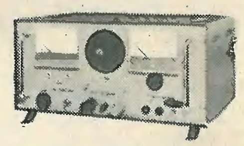

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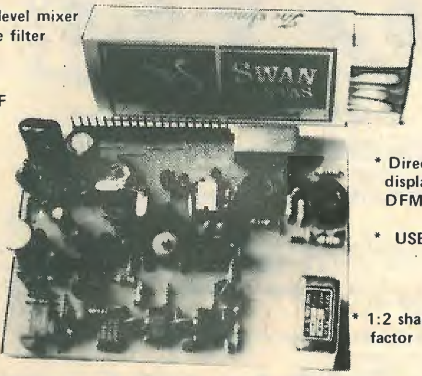

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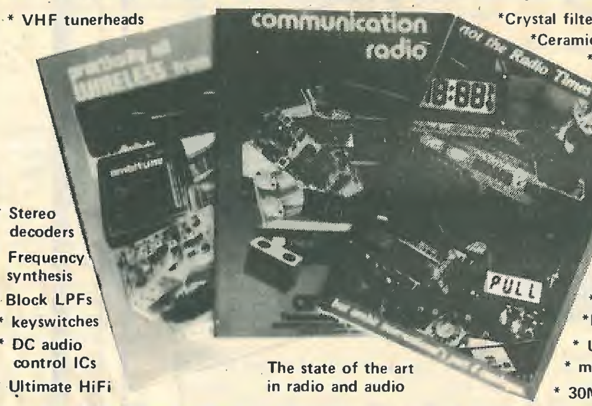

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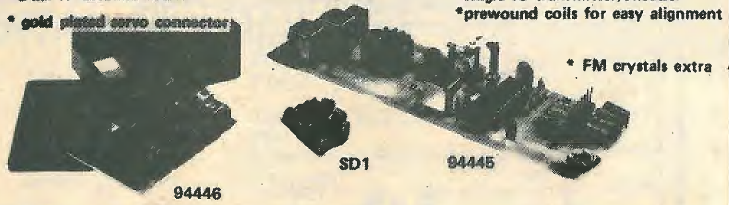

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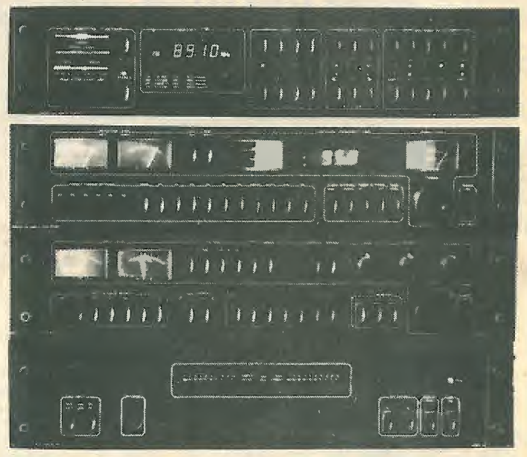

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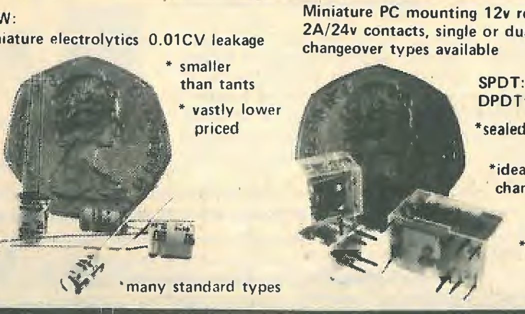

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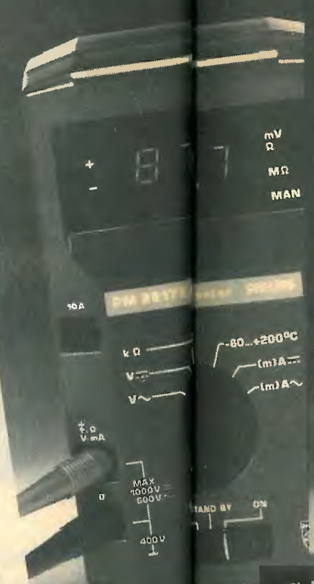

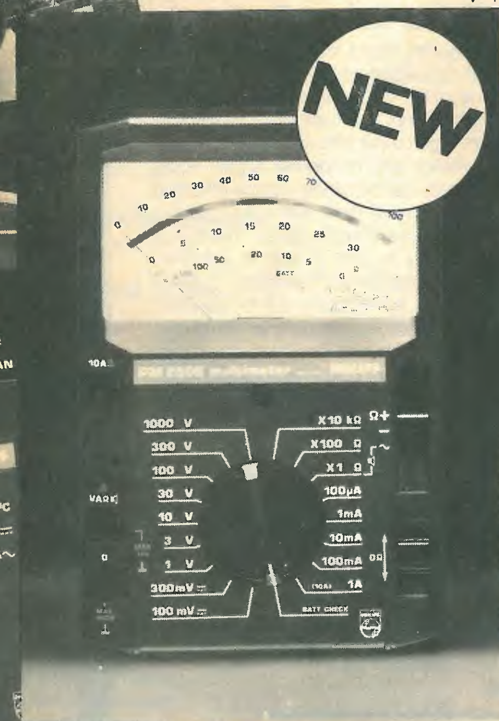

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The dream of objectivity

All good scientists and engineers are dedicated to the principle of objectivity. This presumes, however, the existence of something that can never be – a human observer who is both detached from the world he observes and also free of all value judgements, assumptions, prejudices, desires, unconscious motivation and psychological conditioning. Because some of the productions of the human mind, such as logic, are value free, we tend to think that the mind as observer should have this property too. Experience should remind us, however, that such purity is attained only occasionally and fleetingly, perhaps after much painful effort. The principle of objectivity is, as the French scientist Jacques Monod reminds us, an act of faith, a moral rule. In his book *Chance and Necessity* he says: "True knowledge is ignorant of values, but it has to be grounded on a value judgement, or rather an *axiomatic* value. It is obvious that the positing of the principle of objectivity as the condition of true knowledge constitutes an ethical choice and not a judgement reached from knowledge, since, according to the postulate's own terms, there cannot be any 'true' knowledge prior to this arbitral choice."

Our human limitations in reaching towards the ideal of objectivity have been demonstrated in this journal in the various articles and letters discussing the nature of a propagating signal. Must an electromagnetic signal be understood in terms of Maxwell's equations or can it be accepted as a primitive, a thing in itself that requires no further analysis, as suggested by Catt, Davidson and Walton? The problem here is that it is easy enough to propose mental models, but far from easy to validate these models with the uncertain and fragmentary information we get by observing the real world. The first question, though, is *what* exactly do we observe? Initially it seems to us that we mentally observe our own experiences. These experiences are happenings inside the body including those in the sense organs resulting from stimuli from the outside world. We have to admit, though, that these experiences are not really observed by some detached agent, some extra little creature inside us, because there is no such mythical observer apart from the process of observing. It is only possible to say with certainty that observing is taking place. The preceding discussion of what "we" do as active agents is no more than a convenient form of words, fostering an illusion.

This illusion of the separate, autonomous "I" or "we" doing things to the world (here, observing it) arises from the historical mind-body duality, which may have originated from the religious concept of an immortal soul. It was formalized by Descartes in his *Meditations* and is now supported by some modern psychology in the assumption of an ego or self. Much of Gilbert Ryle's book *The Concept of Mind* is devoted to attacking this Cartesian "dogma of the Ghost in the Machine". Ryle calls it a category-mistake, like the mistake of thinking a university to be some entity that is distinct from all the buildings and activities which make it up. Just as the university would be non-existent without the buildings and activities so our minds would be non-existent without the brain and sense organs continually receiving stimuli from the outside world. The observer would not exist if it were not for the phenomena of the world: equally the phenomena of the world would not exist if it were not for the observer (e.g. there is no sound without someone to hear it). So they both cannot exist *independently* at the same time. World and observer, object and subject, are indissolubly linked by phenomena. The observer cannot be truly detached and objectivity remains a dream.

Editor:
TOM IVALL, M.I.E.R.E.

Deputy Editor:
PHILIP DARRINGTON
01-661 3500 X3586

Technical Editor:
GEOFF SHORTER, B.Sc.
01-661 3500 X3590

Projects Editor:
MIKE SAGIN
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News Editor:
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01-661 3500 X3589

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01-661 3500 X3561

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Surface acoustic wave devices

A practical guide to their use for engineers

by R. J. Murray and P. D. White Philips Research Laboratories

This article, intended specifically for professional applications engineers, covers three common types of surface acoustic wave devices: bandpass filters, delay lines and oscillators. The main part of the article summarizes their performance limits, specification and application. Subsidiary sections give the basic principles of s.a.w. devices and (next month) fuller information on the specification, operation and performance trade-offs of the three types of component.

Among the many signal processing techniques available to today's engineer it is easy to lose sight of one of the more versatile and yet lesser known technologies – that of surface acoustic waves.

Surface acoustic wave (s.a.w.) devices are now being incorporated into advanced electronic systems in both the professional and consumer markets and can, in many cases, implement signal processing functions that are not easily achievable with alternative technologies. The following sections describe three types of device from the wide range of available s.a.w. components: bandpass filters, delay lines and oscillators.

Bandpass filters

The range of s.a.w. components includes both transversal filters, which are broadband, and resonant filters, which are narrowband. In a transversal filter, filtering is achieved by passing the signal through a number of delay paths and adding these delayed signals. In the passband the various signals add constructively while in the stopband they add destructively. Thus, s.a.w. transversal filters use travelling waves while, in contrast, s.a.w. resonators employ standing waves and have properties similar to LC and quartz crystal resonant filters. The range of realisable filter bandwidths is shown in Fig. 1.

The best known example of this type of s.a.w. device is undoubtedly the television i.f. filter. Work on the s.a.w. tv filter began more than ten years ago and has resulted in the development of devices which are now in large-scale production in England, France, West Germany, Japan and the USA.

This surface wave filter replaces an LC filter which uses five adjustable inductors and one adjustable resistor (all of which need setting at the factory) as well as several other components, all assembled on a printed circuit board and occupying a volume of 50cm³. The surface acoustic wave replacement needs no alighnment and is mounted in a TO-8 package occupying a volume of less than 2 cm³. The two types of filter are shown in Fig. 2.

Unlike LC filters, s.a.w.. transversal filters are not usually constrained to be minimum phase filters. This means that, to a large degree, the amplitude and phase responses may be designed and specified independently of one another. If required, a linear phase response may be achieved while simultaneously achieving a steep sided, flat topped, or equi-ripple amplitude response.

S.a.w. filters have many potential applications in communications and radar systems which can take advantage of their small size and weight. Table 1 lists the performance which might currently be achieved by filters of this type.

The ranges quoted in Table 1 are not intended to suggest that all of these extremes may be met simultaneously. For example, it would not be reasonable to expect a filter with a very narrow bandwidth at 10MHz to fit into a TO-8 package. Moreover, the values shown are typical and not necessarily firm limitations. Wide bandwidths (>50%) can only be achieved with high insertion loss.

Table 1: Surface wave broadband transversal filter capabilities

Centre frequency range	10MHz–1.5GHz
Bandwidth (–3dB)	Minimum: 100kHz or 0.2% of centre frequency (whichever greater) Maximum: 100%
Transition bandwidth (–50dB to –3dB)	Minimum: 100kHz or 0.2% of centre frequency (whichever greater)
Group delay	Typically 1-5 µs
Group delay ripple	Typically <2% pk-pk
Insertion loss	Typically 15-25dB
Passband amplitude ripple	Typically <0.5dB pk-pk
Stopband	Typically 50dB close to passband 70dB further out from passband
Package size	Small – usually TO-8 (1.5 cm diameter)

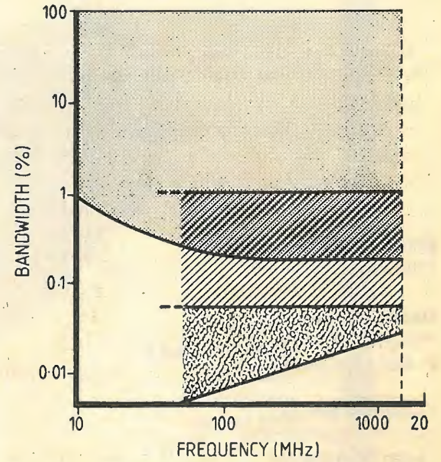


Fig. 1. Range of frequency and bandwidth achievable with surface acoustic wave filters.

Frequency range and bandwidth. There are two main restrictions here: physical size and fabrication considerations. The maximum acceptable substrate size determines the obtainable steepness of the filter skirts, while the available pattern definition determines the upper frequency limit.

● *Low frequency/bandwidth limitation.* The device length is determined by the transition bandwidth, which is the rate of cut off (expressed in db/Hz) of the filter amplitude response, and is *independent of centre frequency*. Shape factors (i.e. bandwidth at –50dB divided by the bandwidth at – 3dB) of better than 2:1 can be achieved.

● *High frequency limitation.* For routine device fabrication, current photo-lithographic techniques set an upper frequency limit for filters of approximately 500MHz (although some higher frequency devices have been made to suitably relaxed specifications). However, with the increasing use of electron beam lithography this limit is expected to rise in the near future to 1.5GHz.

Group delay and insertion loss. There is an absolute delay through the filter which is usually in the range of 1-5 µs, although it may be more for filters with very steep skirts. In a subsidiary section it is shown that there is a trade-off of insertion loss against amplitude ripple and group delay ripple.

A loss of 20dB is typical for most s.a.w. filter applications. This might give an amplitude ripple of less than 0.3dB and a group delay ripple of less than 2%.

Transversal filter example. Fig. 3 shows the amplitude response of a trasversal filter that has been developed as an i.f. filter. Fig. 4 shows the response of the same filter measured over a wider bandwidth. The specification that is achieved is:

Centre frequency	124MHz
Bandwidth (–3dB)	3.7MHz
Amplitude ripple	<±0.5dB
Group delay ripple	<±40ns
Insertion loss	20dB
Stopband (close in)	–51dB
Stopband (ultimate)	better than –70dB
Package	TO-8

Bandwidths less than 1% (resonant filters). This type of s.a.w. filter has a bandwidth range of 0.01% to 1% of centre frequency. These filters are normally suitable for communications channels of bandwidth 12.5kHz or greater.

Because these are resonant devices they have a more restricted range of parameter values than do transversal filters. They are usually specified in the same way as low-frequency LC filters with requirements of frequency, loss, bandwidth, response type (e.g. Butterworth) and order of filter. Table 2 summarises achievable resonator filter characteristics.

The ranges quoted in Table 2 are again not intended to suggest that all the extremes of range can be met simultaneously. For example a very narrow bandwidth filter (say 0.02% of centre frequency) at 500MHz with a third order response would have more than 6dB of loss.

Frequency range. The lower frequency limit is determined by the maximum substrate size. The upper frequency limit is set, as with transversal filters, by lithographic techniques – currently about 500MHz for demanding specifications and up to 1.5GHz for relaxed specifications.

Bandwidth. Within the range quoted (in Table 2) the filter loss decreases significantly as the bandwidth is increased. Whereas a filter with a bandwidth of 0.01% of centre frequency may have 6dB of loss, if the bandwidth is increased to 0.03% then the loss may be reduced to approximately 2dB. Above 0.1% the loss is mainly due to external components and stray capacitance (approximately 1dB).

Bandwidths of up to 0.05% can be achieved using a quartz substrate without any external temperature compensation but for broader bandwidth filters it is necessary to use a different material which means that ovening (control of the filter temperature) is usually required.

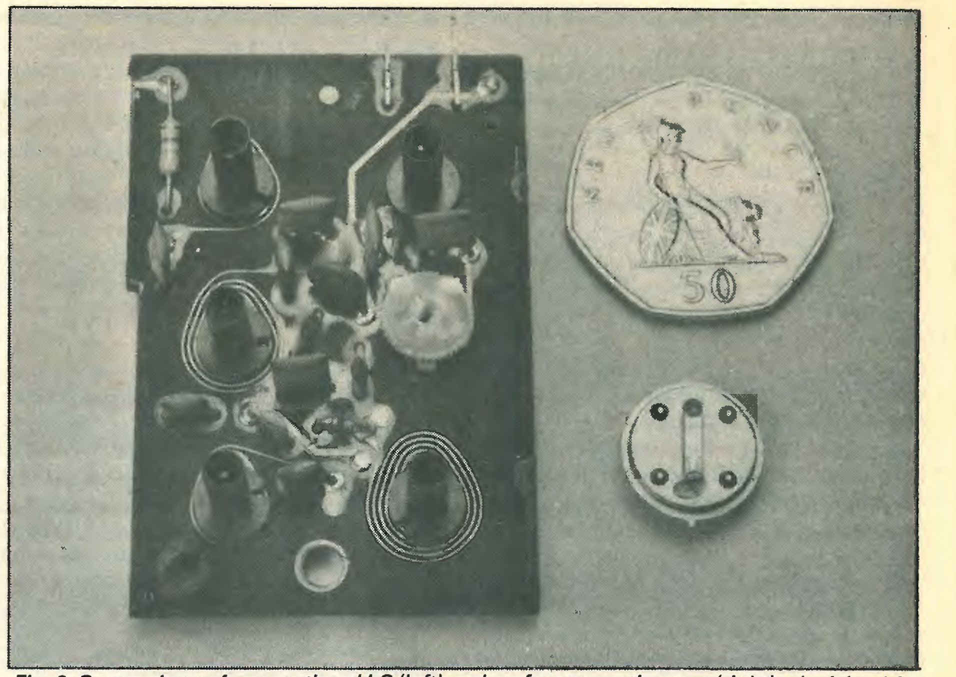


Fig. 2. Comparison of conventional LC (left) and surface acoustic wave (right) television i.f. filters showing the considerable size reduction.

Table 2: Surface wave resonator filter capability

Centre frequency range	50MHz–1.5GHz
Bandwidth (–3dB) minimum	At 50MHz 0.01% of centre frequency At 500MHz 0.02% of centre frequency
Bandwidth (–3dB) maximum	0.05%: temperature stable 1.0% not temperature stable but can be ovened
Insertion loss	Typically <6dB
Stopband	Typically 60dB
Size	Typically 25x15mm at 50MHz 10x5mm at 400MHz

Response type and order of filter. Standard response types such as Butterworth, Chebyshev etc. can be synthesised. Higher order filters can be produced but, for very narrow band filtering at frequencies above about 150MHz, there is likely to be a severe insertion loss penalty with orders of three and above.

Resonator filter example. Fig. 5 shows the response of a resonator filter which has been developed for an i.f. applications. The specification that is achieved is:

Centre frequency	149.950MHz
Bandwidth (–3dB)	30kHz
Insertion loss	3.5dB
Stopband level (f_0±0.5MHz)	–60dB
Order	third
Package	d.i.l.

Delay lines

Using surface acoustic wave techniques it is possible to make delay lines (tapped or fixed) with delays in the range 400 nanoseconds to 30 microseconds which are accurately defined and highly reproducible. The substrate length required is of the

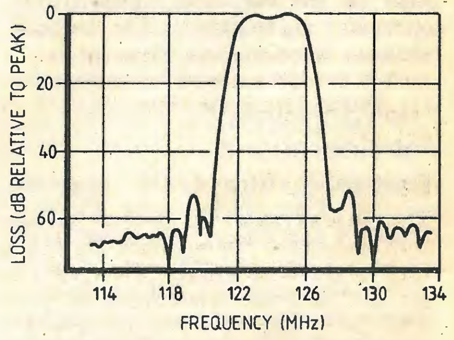


Fig. 3. Measured frequency response of 124MHz s.a.w. transversal filter showing low ripple and very steep sides. The bandwidth is 3.7MHz.

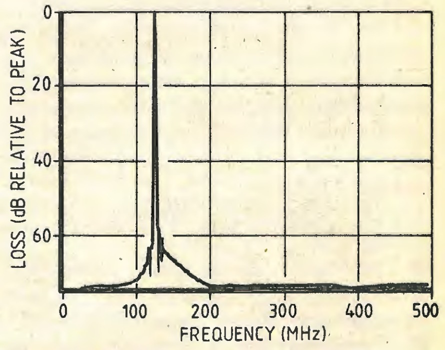


Fig. 4. Measured wideband frequency response of 124MHz s.a.w. transversal filter highlighting the excellent stopband properties of these filters.

order of 3mm per microsecond. Bandwidths of up to 100% can be achieved at centre frequencies from 10MHz to greater than 1GHz. A linear phase response can be achieved within the passband.

Oscillators

For low noise stable oscillators operating at high frequencies, surface wave devices are rapidly becoming recognised as the best control elements available. Using s.a.w. resonators or delay lines it is possible to make oscillators operating at fundamental frequencies between 50MHz and 1.5GHz, eliminating costly multiplier chains and spurious modes of oscillation.

A typical surface wave oscillator might have a fundamental frequency of 400MHz, a long term stability of better than 3p.p.m. /year and a short term (<10s) stability (Allen Variance) of 10^{-9}. Frequency variation with temperature is small (illustrated in Fig. 6) and it is possible, with compensation or ovening to improve this still further. A typical oscillator noise figure at 400MHz is −140dB/Hz at 10kHz from carrier. Some f.m. capability may be provided (sufficient for most audio communications purposes) using a voltage controlled element.

The small size and weight of s.a.w. oscillators are two of their particularly attractive features and, if the oscillator is made as a module (including s.a.w. component and amplifier), the whole device will usually fit into a space of approximately 2cm × 2cm × 1cm. Recent developments have resulted in an oscillator which fits into a volume of $1cm^3$.

Typical applications include local oscillators for telemetry applications (radiosondes etc.) and fixed frequency, low noise oscillators for communication purposes.

Environmental considerations

Temperature characteristics. In general the centre frequency and delay of a s.a.w. device are temperature dependent. There are several materials available for use as surface wave substrates and the choice of material depends on the required temperature characteristic, bandwidth and insertion loss. Substrate materials suitable for narrowband devices generally have a better temperature performance than those for wideband devices. However, it is possible to have temperature stable wideband devices if high insertion loss is acceptable.

Two of the most popular substrate materials are quartz (with good temperature stability), which is normally used for narrowband devices, and lithium niobate (with a linear temperature variation of frequency and delay), which is used for wideband devices.

Typical temperature variations of frequency and/or delay are:

- Transversal filters with bandwidth greater than approximately 5%: 94 parts per million per degree Celsius (p.p.m./°C).
- Transversal filters with bandwidth less than approximately 5%: less than 80 p.p.m. over range $t_0 \pm 50$°C (t_0 is reference temperature).
- Resonant filters, bandwidth greater than 0.05%: 94 p.p.m/°C.
- Resonant filters, bandwidth less than 0.05%: less than 80 p.p.m. over range $t_0 \pm 50$°C.
- Oscillators: less than 80 p.p.m. over range $t_0 \pm 50$°C.
- Delay lines: as transversal filters.

For very narrowband devices a different quartz substrate is available with a frequency or delay variation of less than 50 p.p.m. over the temperature range $t_0 \pm 50$°C.

Size. Standard or custom-designed packages may be used for s.a.w. devices. Typical package dimensions are:

- Transversal filters: size depends on filter skirt steepness, stopband level and passband ripple. In general the package size will be less than 25mm × 12mm × 6mm. TO-8 packages are commonly used.
- Resonant filters: size depends on centre frequency and bandwidth. The size is generally less than 25mm × 12mm × 6mm.
- Oscillators: total module size (including maintaining amplifier) less than 20mm × 20mm × 10mm.
- Delay lines: substrate length depends on delay. 1 microsecond of delay requires a substrate length of approximately 3mm, thus a packaged delay line with a delay of 7 microseconds would be approximately 25mm long.

Ageing. Ageing is only of importance for narrowband filters, oscillators and delay lines. Current quoted ageing rates are 1-2 p.p.m./year.

General. S.a.w. devices are made using standard photo lithographic techniques. Although it is possible to make devices operating at frequencies up to 1.5GHz, the fine geometries required may impose some restrictions on the range of achievable performances at these higher frequencies.

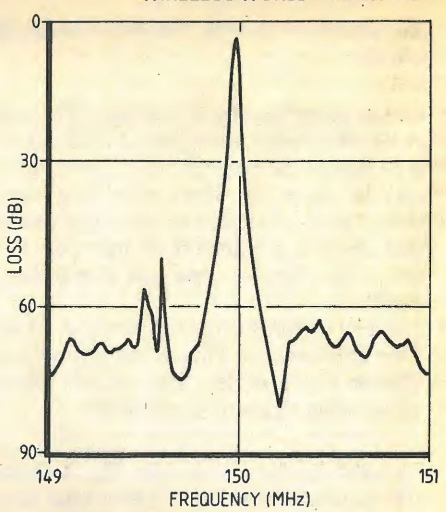


Fig. 5. Measured frequency response of 149.950MHz s.a.w. resonant filter with a bandwidth of 30kHz and insertion loss 3.5dB.

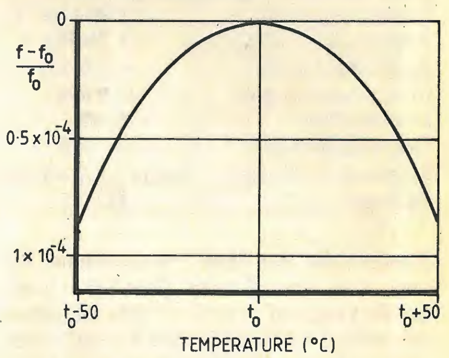


Fig. 6. Frequency of s.a.w. oscillator on quartz as a function of temperature variation about reference temperature (t_0).

Dr Phil White was born in rural Oxfordshire where he lived until 1969. He went from there to the University of Kent at Canterbury and obtained an Honours Degree in Electronics in 1972. After experience in a microwave development group he returned to university in 1973 to do research into microwave dielectric waveguides and was awarded a Ph.D. In 1976 he joined Philips Research Laboratories and since then has worked on various aspects of surface acoustic wave devices.

Bob Murray was born in London in 1952. He attended Southampton University where he obtained an Honours Degree in Physics in 1974. He then joined the GEC-Hirst Research Centre and worked on the physics of bulk and surface acoustic wave devices. In 1979 he joined Philips Research Laboratories where his current interest is the application of surface wave devices in television and communication systems. His hobbies include bridge, squash and gardening.

The surface wave substrate is mounted flat on the base of a package, giving a rugged planar device. Consequently vibration and g-sensitivity is low. Furthermore, the package type lends itself ideally to modern assembly methods. Alternatively, the surface wave substrate may be used directly in a hybrid circuit.

S.a.w. devices are readily obtainable at certain popular frequencies and custom design facilities are widely available. The major features which are attractive to systems designers can be summarized as follows: In many cases signal processing functions can be implemented which would be impractical with other technologies (e.g. high frequency, steep sided, linear phase filters). No adjustment or setting up is necessary. Large time delays may be implemented within a small volume. The devices are small, lightweight and rugged, and they are capable of meeting military electrical and environmental specifications. The technology is familiar and has been well proved in other areas. Finally, narrowband devices are temperature stable.

To be continued

Further reading

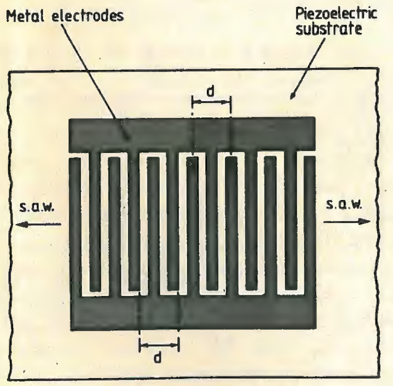
1. Topics in Applied Physics. Vol. 24. Acoustic Surface Waves, A. A. Oliner, Springer-Verlag, Berlin, Heidelberg, New York.
2. Surface Wave Filters: Design, Construction and Use, H. Matthews, Wiley, New York.
3. Reprints entitled "Key Papers on Surface Acoustic Wave Passive Interdigital Devices" edited by D. P. Morgan and published by the IEE. These cover the range of transversal filters, oscillators and delay lines, also other s.a.w. components. S.a.w. resonators are described more fully in White, P.D., Stevens, R. Surface Acoustic Wave Resonator Filters, IERE Conference on Radio Receivers and Associated Systems 1978, pp.93-100.

Principles: surface acoustic waves and the interdigital transducer

The most commonly used surface acoustic wave is an elastic wave which travels on a piezoelectric substrate with most of the energy confined close to the surface, the motion decaying exponentially into the volume of the material. Usually, more than 95% of the energy is contained within one wavelength of the surface and the propagation is essentially lossless.

The surface wave velocity is typically $3000ms^{-1}$. This is five orders of magnitude smaller than the electromagnetic wave velocity and therefore relatively large delays can be achieved with a small path length. It allows the realisation of compact tapped delay lines and the many signal processing components which employ them.

The best technique for launching and detecting surface acoustic waves on piezoelectric substrates is by means of the interdigital transducer (i.d.t.) illustrated at **A**. It consists of two sets of interspersed electrodes, each set being connected to a 'busbar'. The electrodes are photo-etched



A: Uniform interdigital transducer used for launching surface acoustic waves on a piezoelectric substrate. This transducer has a sin x/x frequency response

in a thin film of metal (usually aluminium of 500Å-5000Å thickness) on the polished surface of a piezoelectric substrate. Application of an alternating potential difference between the two sets of electrodes produces electric fields below the substrate surface which, for a piezoelectric material, create a periodic mechanical distortion on and within the substrate surface. This effect may be interpreted by assuming that each individual electrode launches a surface wave. These waves add in phase at the synchronous frequency f_o, given by: $f_o = v/d$, where v is the surface wave velocity and d is the transducer period (see **A**).

Thus, reinforcement occurs and a surface wave is launched from both acoustic ports of the i.d.t. as shown. The i.d.t. is capable of transmitting a band of frequencies centred upon f_o, where the wavelength (λ_o) of the surface wave is equal to d.

The transducer frequency and impulse responses are determined by the geometry of the i.d.t. The position of each electrode determines the point in time of the contribution of that electrode to the i.d.t. impulse response; the length by which the electrode overlaps its neighbours determines the amplitude of its contribution. The i.d.t. frequency response is determined by the Fourier transform of the impulse response. Hence the uniform i.d.t. (where all electrodes have the same length) shown in **A**, has an impulse response which, to first order, is a sampled rectangular function and the frequency response $H(f)$ is centred upon f_o and is of the form:

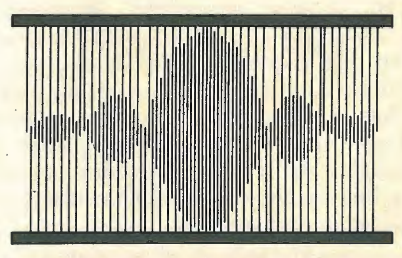
$$H(f) = \frac{\sin x}{x}$$

where $x = N\pi\,[(f-f_o)/f_o]$ and N = number of electrode pairs.

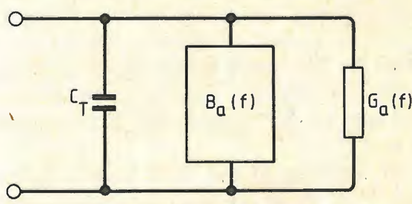
To design a transducer with a given frequency response, the required frequency function is Fourier transformed to give the associated impulse response. This is then truncated to a suitable length and optimised. The impulse response is built into the i.d.t. by variation of the electrode overlaps and phase inverted sidelobes are achieved by reversing the electrode busbar connections. As an example, diagram **B** shows an i.d.t. which has a sin x/x time function built into the structure and hence an almost rectangular frequency response. If the i.d.t. pattern is either symmetric or antisymmetric about its geometrical centre then the phase response is linear. To realise non-linear phase designs the i.d.t. structure must be asymmetric.

Electrically the i.d.t. can be represented by the lumped equivalent circuit shown at **C**. C_T is the static capacitance between the electrodes. $G_a(f)$ is the frequency dependent radiation conductance which represents the transfer of electrical energy into surface waves. $B_a(f)$ is the radiation susceptance and is given by the Hilbert transform of the radiation conductance.

The transducer may be driven directly or additional components may be used to provide an electrical power match to the i.d.t. impedance. In wideband filters and delay lines it is not usually advisable to operate the i.d.ts completely matched, since this maximises the level of unwanted acoustic reflections. These s.a.w. devices are, therefore, often operated under mismatched conditions. When analysing



B: Overlap weighted interdigital transducer. The frequency response is determined by the weighting pattern.



C: Lumped equivalent circuit of interdigital transducer.

or synthesising the frequency response of the i.d.t., the effect of the terminating electrical circuit must be taken into account.

Several piezoelectric materials are suitable for use in s.a.w. devices. The best substrates, in terms of centre frequency reproducibility and low propagation loss, are single crystals. Quartz generally has good frequency/temperature stability for the propagation of surface waves; one particular cut (ST-X) has a parabolic characteristic with a parabola constant of −31.25 × 10^{-9} degree^{-2}. ST-X quartz has a relatively low piezoelectric coupling co-efficient and is best suited for narrowband temperature stable devices. Y-Z lithium niobate has a high coupling co-efficient but a relatively poor linear frequency/temperature characteristic with a slope of −94p.p.m./°C. This material is thus more suitable for use in wideband low loss devices. Several other materials and orientations are suitable as substrates, e.g. lithium tantalate.

Provided the substrate is several wavelengths thick, the lower surface may be fixed to the base of a package with adhesive, producing a rugged structure capable of withstanding severe acceleration and vibration. Electrical connections to the i.d.t. are made using standard bonding techniques. Packages are also standard: TO-5, TO-8, d.i.l. and flatpacks are common.

Modular frequency counters

Flexible instrumentation based on the ICM7216 i.c.

by M. Voznjak

With the introduction of the 7216 family of l.s.i. frequency counter i.cs, the design of a counter/timer has been greatly simplified. However, construction of a high quality instrument still requires a number of important external circuits. This article describes a frequency counter module based on the 7216, and provides a selection of add-on modules which can be combined in one instrument or built as separate units.

There are four devices in the 7216 family, which are identified by suffixes A to D. Types A and B provide frequency measurement and most other features found in a modern frequency counter, while C and D are for frequency measurement applications only. The pin connections and general features for types A and B are shown in Fig. 1.

All versions of the 7216 have 28 pins, and 25 of these are used for inputs, outputs, reset and hold. The remaining three pins select six different modes, four different gate times and a number of other features. The function pin selects frequency counter, period counter, frequency ratio counter, time interval counter, unit counter or crystal oscillator test. The range pin selects four different gate times, and the control pin activates display blank, display test, crystal select, external oscillator enable and, for C and D versions, external decimal point enable.

All of the circuits to be described use the 7216B and associated components shown in Fig. 2. This module can be connected to various preamplifiers, shapers and prescalers to form an instrument with as many facilities as required. If a 1MHz crystal is used D_4 should be connected, and if an external crystal oscillator is to be used, D_3 should be connected. Alternatively, the internal oscillator can be used, but care must be taken to ensure stability. Both fixed capacitors should be silvered mica and the trimmer capacitor should be a multi-turn air dielectric type for improved temperature stability.

Decimal point display with the 7216B is achieved by connecting pin 23 to all of the decimal points wired in parallel. The i.c. automatically places the decimal point in the correct position for function and range so that frequency is displayed in kHz and period in μs. An overflow condition is indicated when the decimal point of digit 7 turns on. If the counter is used with a prescaler, the decimal point position will not be valid, and extra contacts on the prescaler switch are necessary. By breaking the connection from pin 23, the decimal points can be switched off when the prescaler is used. If the counter is to be used mainly for frequency measurement and has a display with separate decimal points, a range switch with two extra sets of contacts can be used. One set has the decimal points wired in accordance with the gate time, and the second set is wired for prescaler operation. Selection of the relevant contacts is made by the prescaler in/out switch, and if two prescalers are needed, a third set of contacts can be used. For other modes, additional contacts must be provided on the function switch, which can then supply power to the decimal points in the frequency mode and disconnect it in all other modes. Decimal point connections for the 7216D are much simpler because there is no function switch. A wiring diagram for the decimal point switching is shown in Fig. 3.

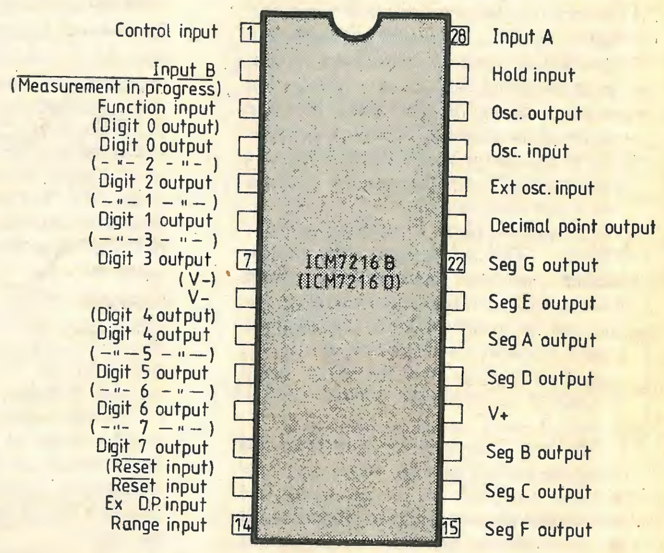


Fig. 1. Pin connections for the 7216B and D. The i.c. operates as a frequency counter, period counter, unit counter, frequency ratio counter and time interval counter. Four gate times from 10ms to 10s are provided in the count mode and 1 to 1000 cycles for period, frequency ratio and time interval modes. Frequencies up to 10MHz and periods from 500ns to 10s can be directly measured, and the i.c. will directly drive an 8-digit multiplexed l.e.d. display. A 1MHz or 10MHz crystal can be used with an internal oscillator, or the device can be controlled by an external oscillator. Other facilities include leading zero blanking, overflow indication, display off, hold and reset, and a test speed-up function.

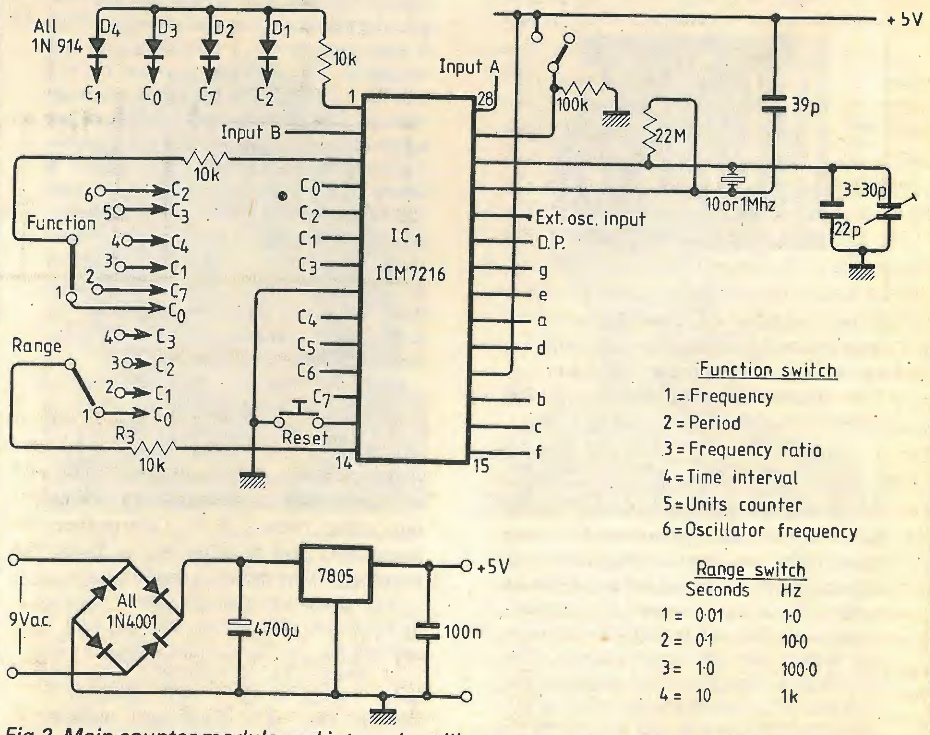


Fig.2. Main counter module and internal oscillator connections. Pin 1, via D_1 to D_4, selects display blanking, display test, external oscillator enable, and 1MHz external oscillator respectively. Break-before-make switches should be used for function and range, non-latching push-to-make for reset and display test, and single-pole toggle for hold and display blank. The voltage regulator must be insulated and mounted on the chassis.

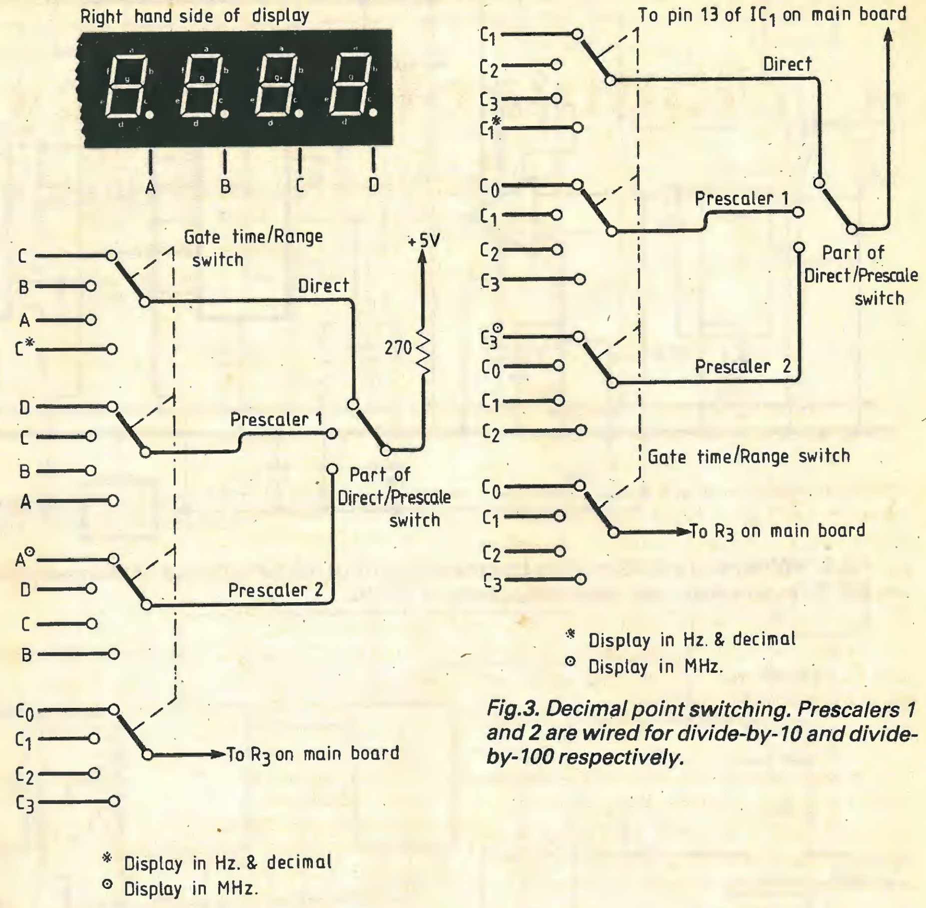


Fig.3. Decimal point switching. Prescalers 1 and 2 are wired for divide-by-10 and divide-by-100 respectively.

Crystal oscillator

If a general purpose counter is required, the internal oscillator is adequate. However, for more critical measurements an external crystal oscillator is recommended. Fig. 4 shows a simple but reliable and stable external oscillator which uses a 5MHz crystal and a divider. The 1MHz output is fed to the counter module, which must have D_3 and D_4 connected. This circuit, although simple, provides a stability of a few parts in 10^{-7} at room temperature.

Input preamplifier

Both inputs of the 7216 require digital signals, so a waveform to be measured must be brought to the logic level and shaped to produce a square wave. This requires a preamplifier which must provide a suitable frequency range, input sensitivity and input impedance. As the maximum operating frequency of the 7216 is 10MHz, the input amplifier should have a frequency range from 10Hz to 10MHz.

Most commercial counters offer input sensitivities between 10 and 100mV. Achieving a 10mV sensitivity is not difficult but, if combined with multiplexing i.cs, the scanning oscillator can cause interference when measuring low level signals. Using the 7216 and a preamplifier with an input impedance of around 10kΩ, the maximum sensitivity is limited to about 40mV. If the input impedance is reduced to around 1kΩ, the sensitivity can be increased to 10mV. Although many counters have an input impedance of 1MHz, this seems to have been passed on from valve equipment, and in practice a lower impedance is suitable for most applications. The design in Fig. 5 provides a frequency range of 10Hz to 10MHz with an input impedance of 20kΩ. The input amplifier stage is based on the LM733 video amplifier, which has an externally selectable gain/bandwidth, and the 100kΩ attenuator is normally set to minimum attenuation. If pins 3, 4, 11 and 12 are left open, the gain is 50 and the bandwidth is around 200MHz. With pins 3 and 12 connected together, gain increases to 100 and the bandwidth is reduced to 100MHz. Connecting pins 4 and 11 together increases the gain to 200 and reduces the bandwidth to 50MHz. By connecting a resistor between pins 3 and 12, intermediate values for gain can be selected, and 1kΩ provides a sensitivity of 40mV which, with the chosen input impedance, eliminates interference from the scanning oscillator.

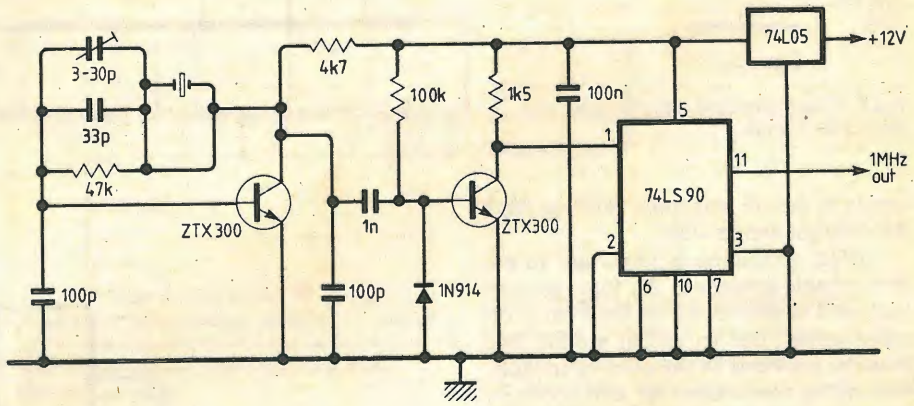


Fig.4. External crystal oscillator.

The amplified signal is shaped by a 74LS13 Schmitt-trigger to obtain a square wave. Because the signal from IC_1 is not large enough to trigger IC_2, dc is added by a resistive network and R_1 sets the threshold of triggering. To improve shaping, the second half of IC_2 is also used.

The output on pin 8 of IC_2 can be fed directly to input A of the counter chip if measurement to 10MHz is required. If a prescaler or p.l.l. frequency multiplier is to be added, it is useful to have a dc controlled logic selector and this is provided by IC_3 and IC_4. Three inputs are provided and are selected by switching their corresponding control lines as shown in Fig. 6. Transistor Tr_1 is necessary to improve the signal shape at higher frequencies and without this transistor the 7216 will not operate above about 9MHz. The complete circuit has a separate 5V

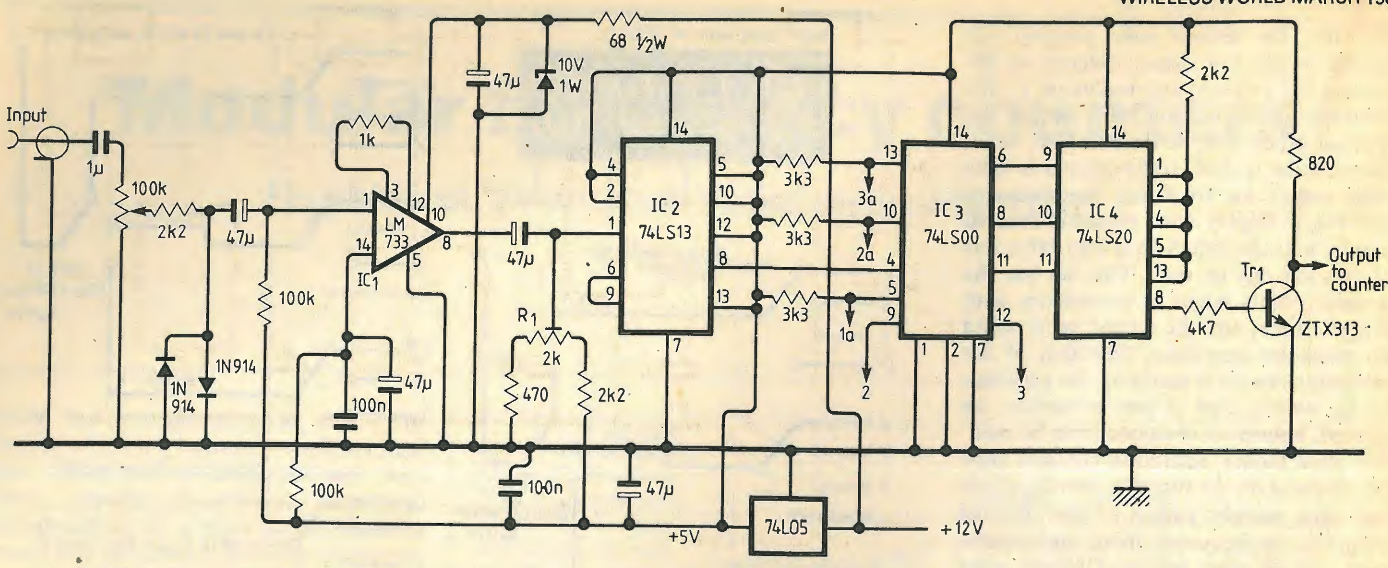


Fig.5. 10MHz input amplifier. Note that the output must not be fed to the 7216 unless power is applied to the counter. R_1 is adjusted for optimum sensitivity and reliable triggering at 10MHz.

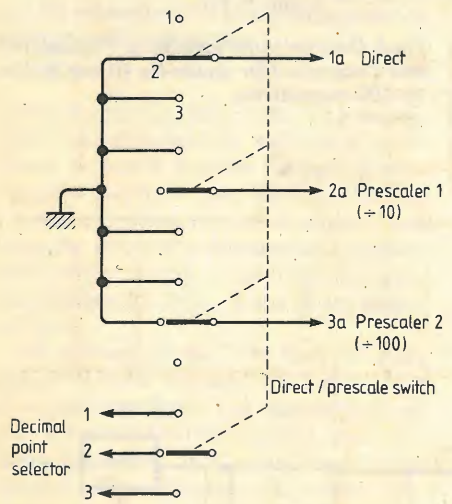


Fig.6. Enable/disable switching for the prescaler signals.

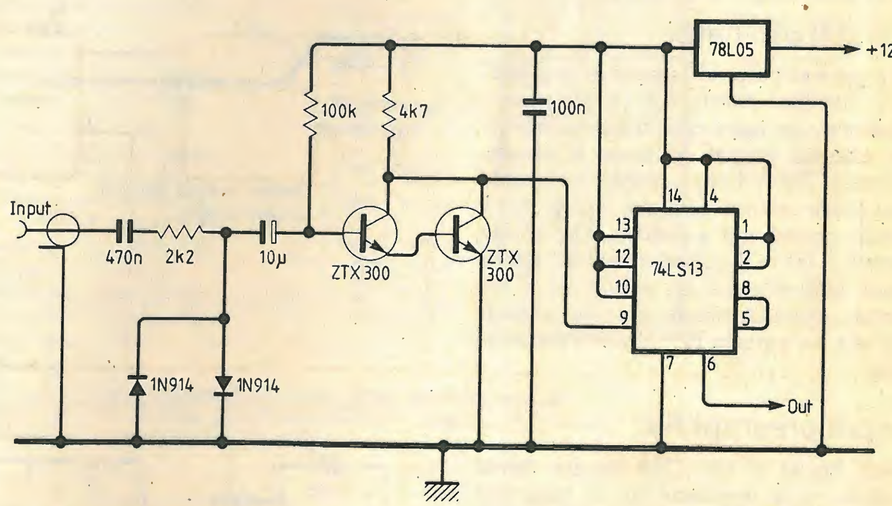


Fig.7. Simple preamplifier for input B of the 7216.

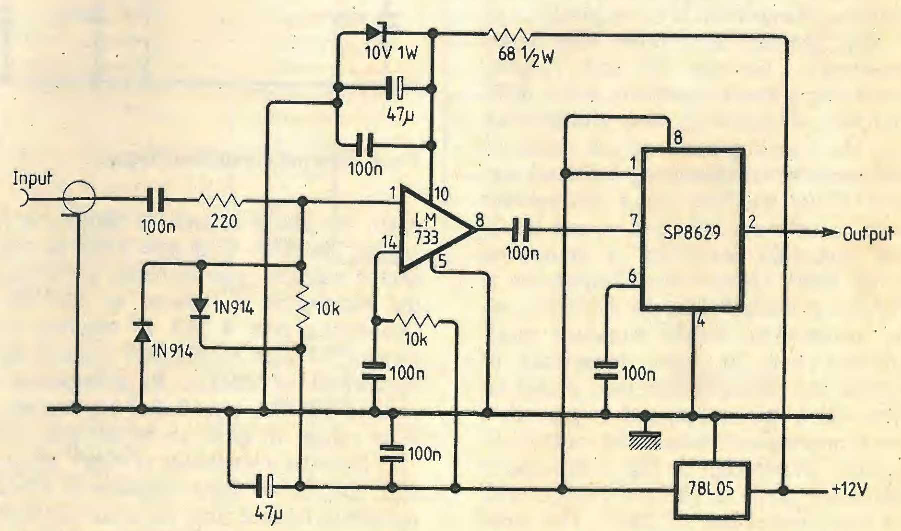


Fig. 8. 200MHz prescaler with an input impedance of about 2kΩ.

supply to ensure maximum isolation from the main counter module.

Correct grounding is important to ensure reliable operation. An input ground lead must be provided from the front panel input socket, and an output ground lead must be provided to the counter module. Grounding connections are also necessary for inputs 2 and 3 of the switching logic. There is also a ground lead which goes with +12V to the power supply on the main board. In all cases it is best to use two single wires and not screened cable.

For frequency ratio and time interval measurements, input B of the main counter is used, which also requires a logic signal. As the frequency limit is 2MHz, a simpler preamplifier is shown in Fig. 7 can be used. Input sensitivity is around 200mV and the input impedance is around 100kΩ. This module has a separate 5V regulator.

Prescalers

If frequencies above 10MHz need to be measured, a prescaler must be used as shown in Fig. 8. The Plessey SP8629 divide-by-100 prescaler i.c. is used which comprises an e.c.l. divide-by-ten circuit followed by a t.t.l. divider. The i.c. also contains a differential preamplifier which gives a sensitivity of around 500mV peak-to-peak. Because this sensitivity is not sufficient, an LM733 is used in the maximum bandwidth mode which increases the sensitivity to 20mV at 160MHz and approximately 80mV at 200MHz. This circuit must be built on a board with an earth plane which is insulated from the chassis, and i.c. sockets should not be used.

For higher frequencies a 500MHz prescaler can be used as shown in Fig. 9. This design is based on a Philips hybrid amplifier, type OM335, which provides a gain of 27dB from 40 to 860MHz. The input circuit of the prescaler uses a Shottky-diode bridge as an input limiter which is biased with a 10 turn potentiometer. After amplification, the signal is fed into an e.c.l. divider which brings the signal to 50MHz. This is followed by a high-speed transistor level translator which feeds a t.t.l. divider to bring the signal to 5MHz.

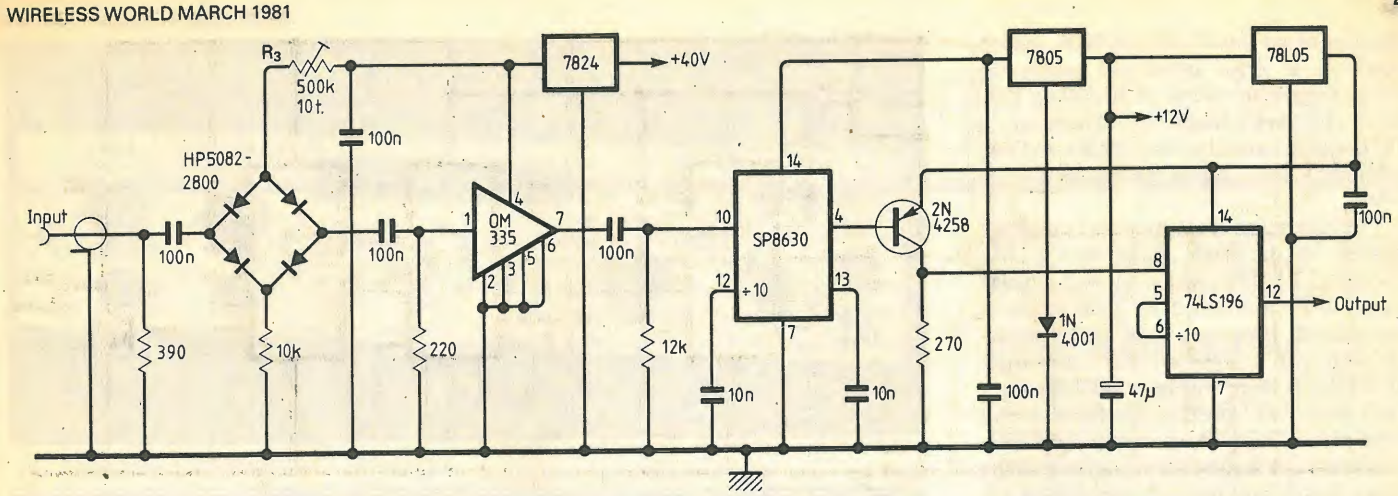


Fig. 9. 500MHz prescaler. R_3 should be adjusted for optimum sensitivity at 500MHz. To reduce stray capacitance it is important to use a ground plane on the circuit board and to use the i.cs without sockets.

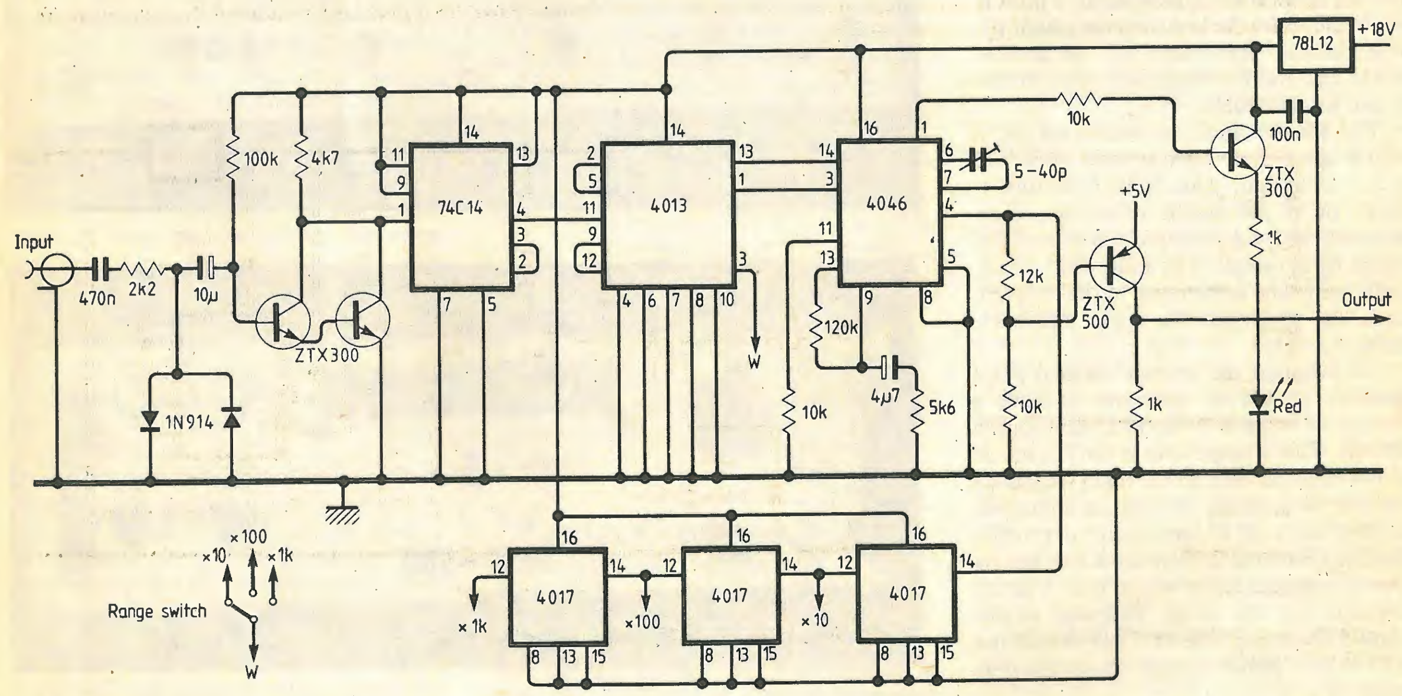


Fig.10. Phase-locked loop frequency multiplier. A secondary winding of 15V at 200mA is needed for the 18V supply, but the 5V rail can be taken from the main counter module.

The prescaler has an input sensitivity of 15mV up to 520MHz, and the power supply requirements are 24V at 35mA, 5.6V at 250mA, and 5V at 50mA. A transformer with two secondaries rated at 9V 1A and 30V 100mA can be used for the 24V and 5V rails, and the 5.6V for the e.c.l. divider can be obtained by connecting a silicon diode in the ground path of a 7805 regulator.

As the regulators dissipate a fair amount of power, a heatsink must be used and it is advisable to include a power on/off switch linked to the prescaler switch. Construction of the 500MHz prescaler is similar to the 200MHz version, however, the hybrid amplifier has rather short pins so care must be taken when soldering the grounded pins to the earth plane.

Pll frequency multiplier

In some circuits it is necessary to accurately measure low frequencies. Although a longer gate time, e.g. 10s, can be used, this method is very time consuming and not very precise because the frequency under measurement may change. Frequency multiplication is a superior method because the gate time can be relatively short without losing accuracy. Fig. 10 shows a suitable p.l.l. frequency multiplier which uses the simple preamplifier/shaper described earlier.

This preamplifier operates satisfactorily because the maximum frequency to be multiplied is about 350kHz. A c.m.o.s. 4046 p.l.l. is fed by a flip-flop which provides a symmetrical square wave at half the input frequency. The remaining half of the 4013 is used to divide the comparison signal by two. The v.c.o. output from the p.l.l. is fed to the counter and also to a chain of dividers. Frequency multiplication is achieved by dividing the comparison frequency by 10,100 or 1000, which produces an error signal and causes the v.c.o. to give an output of 10,100 or 1000 times the input frequency. Therefore, a 1kHz signal could become 1MHz and any frequency error would be multiplied by 1000.

The maximum frequency and range are limited by the 4046. Most devices will operate at 3.5MHz, which allows a 3.5kHz signal to be multiplied by 1000 in about one second. However, if the range capacitor of the p.l.l. is chosen to give an output of 3.5MHz, there will be a low-frequency limit which will prevent operation over a large part of the audio spectrum. One solution is to use a second set of contacts on the range switch and connect appropriate capacitors. Alternatively, a compromise can be made by restricting the maximum frequency to obtain a reasonable low-frequency limit. If the high-frequency limit is set to 2.5MHz, the low frequency limit for one decimal resolution will be around 60Hz and the low-frequency limit of the v.c.o. will be 600Hz.

The 4046 provides an output on pin 1 which goes from high to low when the v.c.o. is locked. This is used to turn a l.e.d. on if the circuit is not operating correctly because the input signal is out of range or too small. The multiplier circuit only needs one adjustment to the trimmer capacitor which sets the upper frequency limit to 2.5kHz.

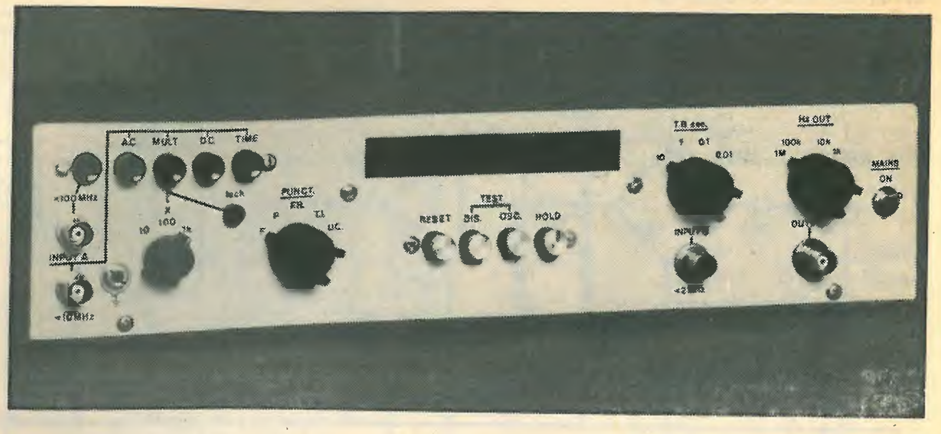
To maintain the correct decimal point position it will be necessary to have a second set of contacts on the multiplication switch. This arrangement is only practical if the circuit is built in the main frequency counter case. If the circuits are combined in one case, it is important to provide further switching so that the v.c.o. has its power removed when not in use. A set of contacts on the main function switch should disconnect the 18V rail so that the 12V v.c.o. signal cannot cause interference.

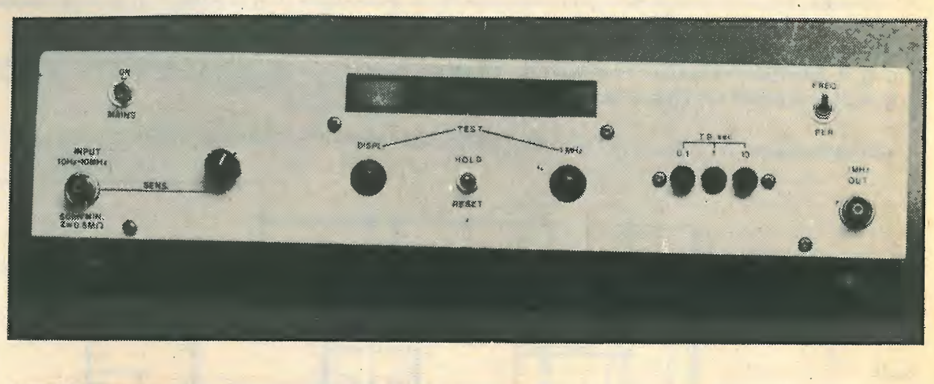
Display

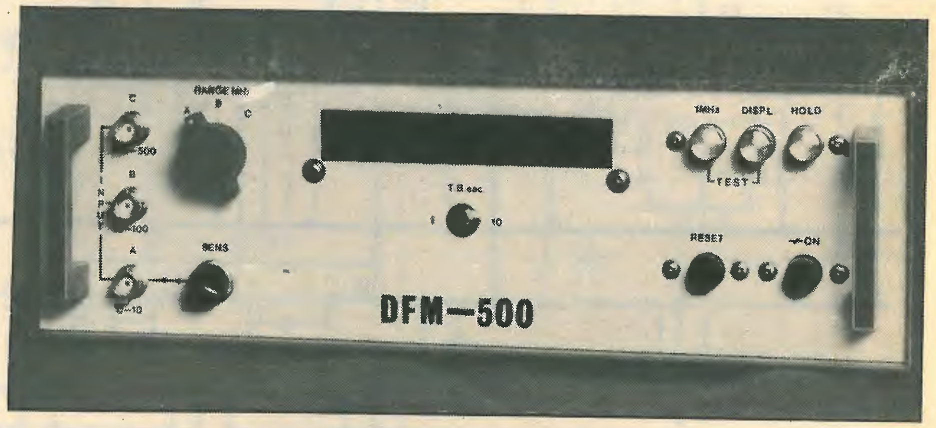
Although there are 4-digit multiplexed displays available, such as the NSB3881, their decimal points are wired in parallel and therefore cannot be separately switched. A better solution is to use four dual-digit 0.6in common cathode displays mounted on a printed circuit board (type C, BY69Y Maplin supplies). This provides a large display with decimal points that can be switched as necessary for correct placement.

If all of the modules are used there are quite a few interconnections to be made, especially for the decimal point wiring. Multicoloured ribbon cable is a great help and makes the wiring much neater. Although the construction of each module is reasonably straightforward, it is important to have separate ground connections from each module to the main counter circuit, and to provide the modules with separate regulators.

A set of 8 p.c.bs for the circuits described and a display will be available for £20 inclusive of v.a.t. and U.K. postage from M. R. Sagin, 23 Keyes Rd., London N.W.2.







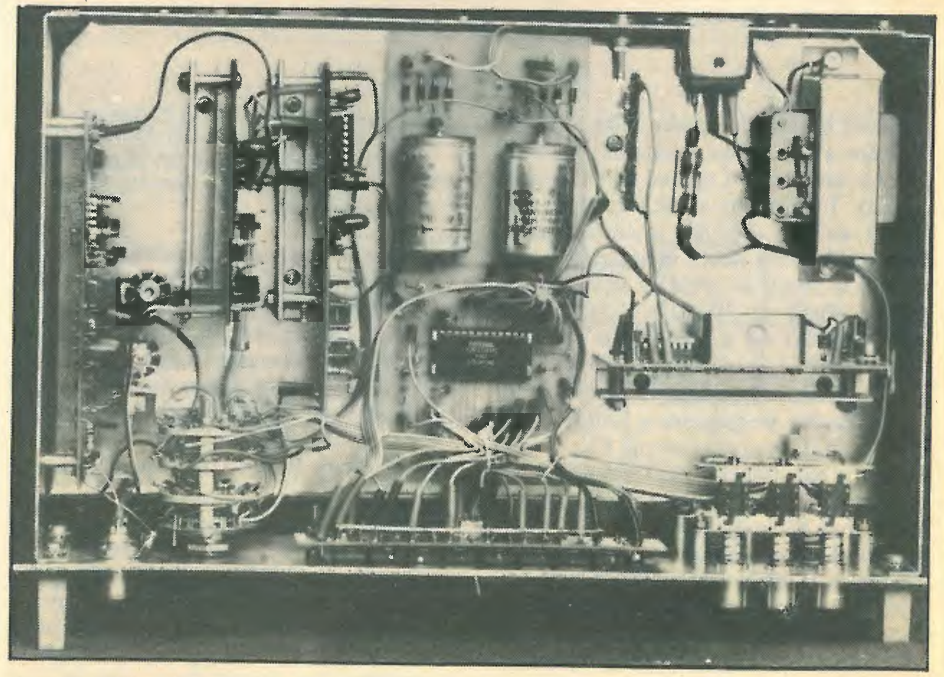


Fig. 11. Three of the author's instruments. A universal 100MHz counter (top) with frequency multiplier, a 200MHz frequency counter, and a 500MHz counter using the input amplifier two prescalers and an external oscillator.

WORLD OF AMATEUR RADIO

Using rain on 10 GHz

The scattering effect of rain on microwave signals has, for a number of years, been recognised as a potential cause of co-channel interference to s.h.f. communication systems, including satellite communications. In 1978, J. A. Lane of the Appleton Laboratory (*Electronics Letters,* Vol 14, No 14, pp.425-427) showed that although in general tropospheric and precipitation scattering are of less consequence for over-the-horizon s.h.f. propagation than super-refraction and ducting, this is not true over very rough terrain or where there is local screening by hills.

On the 10GHz amateur band, Clive Elliott, G8ADP who lives in a heavily screened location at Alresford, Hampshire, can work regularly over paths of up to 150km by means of tropo scatter and is convinced that signals are quite often enhanced by rain scatter. Over a particularly difficult 40km path to G3JVL located at sea level near Portsmouth, effective contacts are largely dependent on rain scatter, with signals maximum when there is heavy rain virtually overhead ("drizzle" is not sufficient) and in such circumstances signals from G3JVL can often be received regardless of which direction G8ADP's aerial is pointing.

He feels that this form of over-the-horizon 10GHz propagation is still seldom recognised or used by amateurs, since much of the effort tends to be concentrated on portable operation in conditions where heavy rain is not welcome. Under normal conditions (if the absence of rain can be termed normal) the signals from G3JVL are about −6dBn (in 2.5kHz bandwidth) but in heavy rain may rise to 30dBn, or about 5 to 15 dBn with the aerial pointing in other directions, including straight up into the sky.

Electrical interference levels

The latest radio interference report (covering 1979) of the Home Office Directorate of Radio Technology underlines the continuous increase over the past decade of complaints by the public of interference to sound radio: from 6492 in 1971 to 23,782 in 1979. During this period complaints relating to v.h.f./f.m. reception have gone up from 1773 to 7214 while those relating to l.f./m.f. from 4719 to 16,568. During the same period there has been an almost equally dramatic fall in the number of complaints relating to television reception (from 58,305 to 20,482) almost certainly the result of the changeover from v.h.f. to u.h.f. tv.

It would thus seem that levels of interference from l.f. to v.h.f. are still rising, largely due, one suspects, to the greater use of contact devices in the form of central heating thermostats. Certainly in suburban London, I find more and more electrical interference to h.f. reception (not covered by the Post Office investigations since these are confined to reception of the local radio and tv broadcast stations): small motors (d-i-y power tools, refrigerators) and radiation from the switched-mode power supplies and timebases of colour tv sets, still compete powerfully with contact devices, particularly where, as in my case, a long-wire aerial runs in places all too closely to domestic mains wiring.

The amount of interference to radio and tv reception ascribed to amateur transmitters is now very low (although possibly helped by a change in the way the information is provided): only 127 complaints arising from 91 sources from "fundamental" radiation; and 8 complaints arising from 6 sources for "harmonic" radiation. But one suspects that a certain percentage of the 16,669 complaints ascribed to "conditions at the receiving site" and including such causes as "inadequate system immunity", spurious responses etc, were initially investigated as complaints of amateur interference.

The Home Office report indicates that domestic contact devices account for an impressive 11,423 of the overall total of 46,031 complaints from 35,500 complainants.

"Typical" enthusiast

The longer one has held an amateur licence, the more convinced some of us become that there is no such person as a "typical" amateur – nevertheless it can be interesting in trying to put together such a composite being. The Cornish Amateur Radio Club recently attempted this by analysing over 80 replies to 200 questionnaires sent to their members. It produced the following picture of their "typical member": About 42 years old and has held an amateur licence for 7 years after being a shortwave listener for about five years . . . became interested in amateur radio as a result of listening . . . interested in all amateur bands from 1.8 to 432 MHz but favours 14MHz for home use and 144MHz for mobile operation, often using the local repeater on his journey to and from work . . . initial transmitter was a new factory-built rig used with a dipole aerial but has subsequently become interested in home construction of equipment . . . spends roughly 5-6 hours per week on his hobby, including operating . . . generally admires the friendships and spirit of the hobby but is appalled by the bad manners and discourtesy of a small minority of operators . . . also interested in music, hi-fi, household d-i-y-, sailing and fishing . . . reads thoroughly but seldom contributes to the club magazine . . . believes c.w. (Morse) operating is on the way out but would welcome more Morse classes run by the club . . . feels Raynet (emergency scheme) is 'for the younger ones' . . . feels he is a good amateur and club member.

From near and far

The propagation-mode that enabled the Canadian station VE1ASJ in St John, New Brunswick to make crossband 50/70MHz contacts last November/December with at least four British 70MHz amateurs is still uncertain: F2 layer reflection; ionospheric forward scatter; and "double-hop" Sporadic E all have their supporters, although a patch of intensive F2-layer ionisation seems the most likely. Crossband 50/28MHz contacts were made during December by British stations with VS6BE and VS6FX in Hong Kong. The Irish amateur EI6AS, who is licensed to use 50MHz, made two-way contact on the band with VS6BE.

Six radio-equipped Land Rovers, each with a Raynet operator as a member of the crew, spent about a fortnight in the Italian earthquake-disaster area during December to help rescue operations. An Italian "young lady" operator (I8YCT) maintained radio contact with the Land Rovers during the outward journey and German amateurs also rendered assistance.

It came as a shock to those of us who have for long advocated the introduction in the UK of a "novice" licence (akin to those available in the USA and many other countries) that would permit limited c.w. operation on some segments of the h.f. bands after passing a Morse test of perhaps 5 to 6 words per minute and a "conditions of licence" type of technical examination to find that the RSGB, in proposing such a facility to the Home Office, has added the off-putting rider that all operation should be under the direct supervision of a licensed amateur. This would reduce the system virtually to club or family stations and, on the lines proposed, would cost the novice as much as a Class A or Class B licence.

In brief

The 1981 National Amateur Radio Exhibition is now scheduled for May 28 to 30 inclusive in the Palm Court hall at Alexandra Palace in North London. This hall escaped damage in the fire last year . . . The Home Office has raised the amateur licence fee from £6.40 to £8 per annum as from January 1 . . . Members of the Cornish Radio Amateurs Club have formed a Computer Club which meets monthly at Pool between Redruth and Camborne (details R. M. Frost, Trecarne, Alexandra Road, Illogan, Redruth (Tel. Portreath 842583) . . . RSGB membership by December 1, 1980 had risen to 27,235 including over 60 per cent of all British amateur licence holders.

PAT HAWKER, G3VA

NEWS OF THE MONTH

Britain ahead in computer networking

According to a recently published report, Britain probably leads the world in linking its university computers and in the introduction of compatible data communications facilities among universities and research institutes. The report is the first review of the activities of the Joint Network Team (JNT) which was established by the Computer Board and the Research Councils in April 1979, and covers the period from its inception to August 1980.

Cornerstone of the team's programme is the adoption of standards for computer to computer and terminal to computer communications to ensure the greatest possible integrated use of equipment by universities and research establishments. International standards are applied where available but where no formal standards yet exist the team is ensuring that a uniform approach is adopted. The following is a list of standards to which the academic community will adhere:

– X25, X3, X28 and X29 (as defined in the Technical Guide to the Packet-Switched Service (PSS) of British Telecom).

– The Network Independent Transport Service (from Study Group Three of the PSS User Forum).

– The Network Independent File Transfer Protocol.

– The Network Independent Job Transfer and Manipulation Protocol (both published under the auspices of the Department of Industry's Data Communication Protocols Unit).

These protocols cover the connection of computers to packet-switched networks, terminal access to services, and facilities for the transfer of files and jobs.

The aim is to ensure that users can "talk" to computers at their own and other establishments by means of a communications hierarchy with local campus networks attached via gateway machines to wide area communications facilities. Twenty-four universities and Research Council sites are among the early subscribers to PSS. For local communications, work is being funded to explore and develop several technologies including campus packet-switches, Ethernets and Cambridge Rings.

The report notes that the activities described have resulted in extensive co-operation among computer centres in universities and research institutes. That degree of collaboration "is probably unparalleled in any other country and may be regarded as a measure of the lead which the British academic community has in implementing communications facilities among heterogeneous machines."

Breaking into the male-dominated world of service engineering, Pauline Cameron started by winning an Electrical Industries Training Board scholarship to train for a year at the EITB training school. After a further year of training with Marconi, she now services and repairs research equipment at LKB Instruments Ltd at Selsdon, Croydon.

Is VLSI just too much?

Semiconductor manufacturers are likely to face severe difficulties not only in making very large scale integrated circuits (v.l.s.i.) but also in persuading people to buy and use them, according to one senior man in the industry. Leslie Vadasz, president of Intel's Microcomponents Division in the USA, stated at an IEEE conference that by 1990 the v.l.s.i. device will have over a million transistors on a chip. "The question really is: what do you do with all that complexity? This will pose a serious software crisis as well as a marketing problem. As our products get more complex the software needed to develop and market them will grow exponentially. Unless we can put a million software people into the workforce by the mid-1980s I don't see how we can really exploit our capabilities".

On the question of manufacturing the v.l.s.i. devices in the first place, Mr Vadasz said: "Where are the engineers who will do the designing, fabricating and programming? The semiconductor industry has a relatively small base of key technical talent. The current shortage of such talent – and the predicted future shortages – are major for both our industry and educational institutions".

Speakers at the conference – on circuits and computers and reported in the December 1980 issue of the IEEE's newspaper *The Institute* – also discussed the future pattern of the industry and what sort of products it will offer as a result of further development in v.l.s.i. Mr Vadasz felt that as commercial success depended on sales volume the job ahead for the manufacturers was to minimise the use of custom-designed i.cs. "We must provide more complete solutions for users, so that they do not need a different chip for every job." One way of achieving this, according to Bernard List of Texas Instruments, was to make standard pieces of silicon that could be programmed by on-chip software to perform different functions for different customers. L. Saehn of Siemens felt that the increasing complexity of integrated devices would change semiconductor firms into systems firms, and he also expected "a narrowing market for new families of microprocessors, because of the need for software compatibility".

Such discussion among the experts must raise in users' minds the question of how far the process of integration can continue along the present lines. Presumably there is some physical limit set by the natural characteristics of the materials, radiation wavelengths etc. used in manufacture, but before that is reached constraints set by the market could take effect.

European business satellite

Plans for a business satellite communications service with Europe, with messages beamed direct to small aerials close to users' premises, were outlined by Peter Benton, Managing Director of British Telecom.

The service, due to start in 1983, is intended primarily for large business organisations, with their own internal telecommunications networks, and for other businesses with specialist requirements.

Mr Benton said that "The service will exploit the very latest transmission techniques, This will not only offer our customers additional facilities for sending telephone speech, telex, facsimile or computer data quickly between premises; it will also allow us to act swiftly in adding more advanced services, such as video-conferencing, high-resolution facsimile, high-speed data and multi-destination broadcasts, whenever the customer wants them."

The service was made possible by an agreement reached at a meeting in Paris of the Eutelstat ECS Council, of which British Telecom is a member. The council decided to modify the European Communication Satellites (ECS) so that all but the first five being built will be able to link up with small dish antennae. This function will augment their original role in providing new communications links through large earth stations like those at Goonhilly and Madley.

British Telecom will install small earth station aerials – about 4m in diameter – at locations appropriate for the users. They will also install ground-level links (conventional telephone cable, optical fibre, or microwave) to connect the aerial to a user's internal communication system.

Both ECS and Telecom 1, the French satellite system, will have extra transponders fitted to operate at the internationally-agreed small-dish frequencies of 12 and 14 GHz, supporting transmissions at 2Mbit/s.

A transportable dish aerial, about 4m in diameter, is one of two supplied by Ferranti Electronics to British Telecom for use in trials for the business satellite.

Earthquake simulator uses p.c.m. data links

Consisting of a 6×6 metre vibration platform which is capable of supporting structures weighing as much as 50 tonnes, the seismic simulator for the Moscow Hydroproject Institute is to be used to analyse the effects of earthquakes on building structures, especially on nuclear power station reactor equipment.

The hydraulic actuators can reproduce the full frequency range of seismic disturbances with vertical acceleration of 1.0g and horizontal acceleration of 1.2g at maximum payload. Double these values can be achieved with lighter test pieces. The platform is moved in the three linear and three rotational modes of freedom by twelve acutators operating under digitally programmed analogue control. The programming system accepts siesmic data in analogue form from magnetic tape or digitally from a curve digitiser, paper tape or card readers, or from magnetic disc. Signals derived from real earthquakes can be used as input signals.

The effects of the seismic shock on the structure are sensed by strain gauges and accelerometers, and are collected by two independent 104-channel high speed data acquisition systems, each comprising thirteen Johne & Reilhofer 8-channel p.c.m. links, based on high resolution J & R System 8K13 modulators and demodulators, with associated 8-channel analogue input signal conditioning units, filters and output interfaces.

Each 8-channel p.c.m. signal is recorded on a single magnetic tape track by a thirteen-track instrument tape recorder, having the capacity to record the data from all 104 input channels. The outputs from the recorder are converted by Johne & Reilhofer demodulators to 12-bit parallel data signals and fed to the central computer for analysis. Each of the 104 analogue systems can have frequency components up to 300Hz involving a sampling rate or 1.5kHz and an ultimate encoded bit rate of 19.5kb/s.

The seismic simulator was designed and built for the Hydroproject Institute by Servotest Ltd.

Selling hi-fi to the British

. . . is the title of a market research report published by Research Associates of The Radfords, Stone, Staffs. It contains few surprises: purchasers are willing to pay over the odds if they can get reliability, technical excellence and good appeareance; British-built equipment is less reliable than Japanese; hi-fi dealers think that Sony offers the best reliability while JVC is top value for money; exterior design is considered important by the consumers interviewed, younger respondents preferred a 'space age' style while older customers still like a wood finish; rack systems are liked by the dealers and the consumers; remote controls and track selection mechanisms are rejected as being gimmicky but the prospect of 'genuine' technical advances, such as laser-scanned discs and smaller, high quality loudspeakers, are looked forward to eagerly. The report also looks into brochures and advertising.

We noted that the report included few surprises. It is complete in 21 pages and there are also appendices on the sampling methods, the questions asked, and some of the responses. The really big surprise is the price. The report is available from Research Associates for £265.

Fifteen years of Pioneer 6

Originally designed to have a working life of six months, NASA's Pioneer 6 interplanetary spacecraft is still sending back useful data after 15 years of circling the sun in a planetary orbit. The craft has measured the sun's corona; returned data on solar storms and measured a comet's tail. It has made discoveries about the sun and about the solar wind, solar cosmic rays and the solar magnetic field, all three of which extend far beyond the orbit of Jupiter.

Since the launch in December 1965, the 64kg Pioneer has circled the sun 17½ times, covering just over nine billion miles and has sent back about four billion data bits. Together with Pioneers 7, 8 and 9, a network of solar weather stations circling the sun was set up, sending back data to many sun-watchers.

In August 1980 it was found that Pioneer 6 had turned itself off due to a momentary power shortage. Mission controllers at the Ames Research Center were able to command it back on again by radio signal and the instruments continued their observations. The Mission Manager, Richard Fimmel believes that they may get another ten years data from Pioneer 6.

New minister

In the editorial comment in our February 1981 issue, it was pointed out that information technology was likely to be given 'official' status by the appointment of a **Minister of Information Technology.** The appointment has now been made within the Department of Industry. Mr Kenneth Baker has been given the post and has responsibilities for telecommunications, computer systems, microelectronics applications, robotics and all aspects of information technology.

Prestel to control editorially its information providers

In an attempt to boost the present meagre use of Prestel by the public, the British Telecom bosses of this national viewdata service have reversed their earlier decision not to impose editorial control on the material offered by the information providers. For some time British Telecom and the information providers have been worried by the poor response of the public to this new service (News, November 1980 issue, p.54). Hitherto this has been blamed on the slow availability of the Prestel television sets which form the users' terminals. But now there are plenty of sets available, and some retailers are inviting people into their shops to see Prestel demonstrated, attention has been switched to another scapegoat – the alleged unattractiveness of the information on offer.

Recognising that "the real product being sold is the information", Frank Burgess, the head of Prestel's UK marketing organization, writes in our sister journal *Viewdata and TV User* (January issue) that although British Telecom remains committed to the principle of editorial freedom for the information providers, nevertheless "for commercial reasons" its hitherto neutral editorial position will not be continued. "In future database pages will not be allocated on a first come first served basis but will only be leased to organisations who can demonstrate the ability to set up information services which will lead to increased set sales and set usage. Conversely, information providers who have shown little inclination to provide an acceptable standard of service may not be given the opportunity to continue renting pages once contracts expire".

Thus British Telecom intends to exercise the authority and responsibility of a publisher, instead of being just a common carrier, which is its normal role in telecommunications. In this respect it will be performing more completely the function of electronic publishing to a mass audience, which was the original idea motivating the development of its viewdata system. Indeed, Mr Burgess confirms in his article that Prestel's aim in the 1980s continues to be to establish itself as "a mass-market medium of communication . . . . . ".

Further pressure on the information providers mentioned by Mr Burgess will be to involve them in sales promotion of the service. Here British Telecom are offering a material incentive in the form of rebates on their charges to the information providers – £25 for every directly attributable sale to a business customer and £10 for every such sale to a residential customer.

The comedians Eric Morecambe and Ernie Wise in full wisecracking form as they telephone, via satellite, Captain Douglas Ridley, Master of the QE2, as the liner sailed off Puerto Rico over 4,000 miles away. They made the call on a radiotelephone from British Telecom International's stand at London's Boat Show.

Prestel terminal specification

A joint programme of work between the private and public sectors of industry has resulted in the publication of the Prestel terminal specification.

The specification draws together the three technologies involved in the Prestel viewdata service – television, telephone and computer – by specifying the safety, interworking protocol, signal and display requirements of the Prestel system. It imposes the minimum of restrictions so that the maximum freedom of design is available to those working on new Prestel terminal equipment.

The specification is the result of more than a year's work involving consultation between engineers from all over the electronics industry and Prestel's own technical staff.

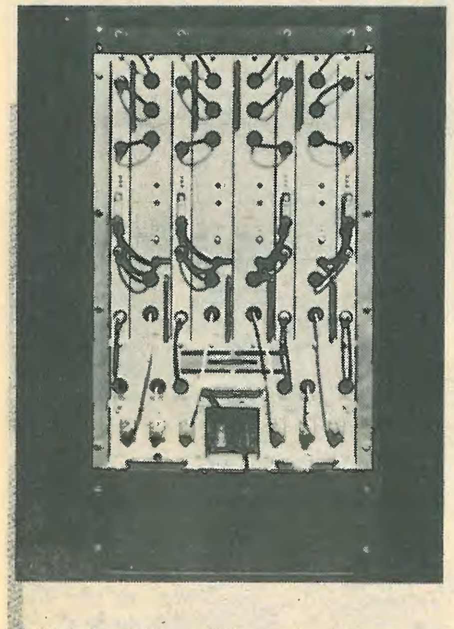
Copies of the Prestel Terminal Specification can be ordered directly from Prestel 4.1 CSU, Prestel Headquarters, Telephone House, Temple Avenue, London EC4Y OHL. The price of £10 includes provision for a year's amendments and extensions to the specification.

World standard for video recorders

To unify the recording conditions of the three broadcasting systems, PAL, SECAM and NTSC, the IEC has issued an internationally agreed standard for high-quality video and stereo-audio recording.

The standard, IEC Publication 602, is suitable for reel-to-reel and cassette recorders in all television standards. Specifically it defines in detail the electrical and mechanical parameters for the professional 'segmented field' video recording system on one-inch tape. The recording format in the standard is known commercially as the BCN-recording system which has been accepted by IEC member countries as the international system. Its application will ensure the interchangeability of recording be it for the 525-line, 60-field system or the 625-line, 50-field system.

The standard video signal is recorded in segments of 52 lines, which leads to 6 segments for each of both video heads for composing a complete television frame in PAL or SECAM. For NTSC only 5 segments are needed. For recording high quality audio signals, three tracks are available of which the third has been standardized as that for time code recording.



British firms have a chance to make the BBC's new u.h.f. transposer (four are shown) intended to bring television to small communities.

TV transponders for small communities

A new television transposer will be much used for the extension of u.h.f. television coverage down to population groups as small as 200. At this level the cost per head can be critical and the new BBC design sets out to achieve a high standard of performance combined with low cost. It will eventually supersede the previous BBC transposer design, still entering service at the rate of 140 a year, and which will remain operational at about 300 relay stations.

Unveiled at a recent BBC designs department exhibition, the new transposer uses the latest components and techniques to reduce the overall cost of providing television coverage. The basic manufacturing cost has been reduced by 40 per cent and size and weight reductions will bring significantly lower installation costs.

The transposer embodies two u.h.f. synthesizers. Digital frequency synthesis features in some modern television receivers but for transmission the resolution and stability requirements are more stringent. The new design uses a novel phase-locking system that provides the required performance at low cost and, as the necessity for setting-up adjustments has been avoided, restricts on-site maintenance to simple replacement of faulty modules. Another advantage of this system is that the output frequency is fixed even though the input frequency may vary, which can reduce the risk of visible co-channel interference in some situations.

Although specifically designed with small pockets of population in mind, the equipment can be used in larger service areas with an add-on power amplifier. Many such stations are still to be built and this arrangement has the advantage the BBC say of improving operational efficiency by limiting the variety of equipment in service.

The picture shows *four* of the BBC's new transposers mounted in a 19 inch rack.

NPL forms speech recognition club

In collaboration with a group of electronics firms and systems houses the National Physical Laboratory has formed a speech recognition club to develop and exploit the technology of direct control of machines by human speech. The idea is to make use of the basic research and technology already available in the NPL and for the club subscribers to help formulate a continuing research and development programme to meet their own commercial requirements. Each club member pays an annual subscription of £8000 to the NPL for these benefits of belonging. At present there are five members: Plessey, Ferranti Computer Systems, System Designers, Quest Automation Research and Nexos Office Systems.

The essential feature of the NPL technique of speech recognition is that it operates with continuous speech in real time and in the presence of noise – as distinct from other methods which will recognize words spoken in isolation or audibly connected words restricted to fairly short sentences. The researchers feel that the ability to use natural continuous speech – in which the spoken words follow on from each other without distinct breaks between them – has big advantages for the average human operator. The awkwardness of enunciating with pauses between words is avoided. And because a continuous input of spoken numbers, words and phases is possible the data capture rate is as high as that achieved by keyboards, according to the NPL. Their system also has a 'key-word' facility which automatically detects particular commands in conversational speech. Consequently the equipment does not have to be switched on and off when the user is talking both to it and to another person. Prototype systems based on this approach have been tested for medical and avionics applications. The NPL would not reveal details of these applications, but it is known, for example, that the Royal Aircraft Establishment, Farnborough, have been studying how direct voice input command systems might be used to improve the performance of pilots flying aircraft.

The NPL technique is able to operate with continuous speech in real time because it does a maximum amount of identification in analogue circuits before using a digital computer. Instead of doing extensive audio signal processing by computational means as in some advanced recognition systems – which would be a slow business in cheap and simple processors – it uses pre-processing hardware for this purpose. The computer that completes the recognition task can be quite a small machine – actually the Digital Equipment LSI-11. This pre-processing hardware analyses the speech in terms of particular phonetic features which relate directly to the control of the articulatory mechanism. Sixteen of these features are used, though about 40 are known altogether. They are not exactly the phonemes as understood by phoneticians but are features such as whether the sound is produced by the larynx (voiced) or not (unvoiced), whether it is a nasal sound, and which of eight possible vowel type sounds it is. Since these phonetic features relating to the articulatory mechanism are common to all human beings the technique is intrinsically suitable for a variety of speakers. There are no problems, for example, because of the differences between male and female voices. The method, too, is designed for telephone-type bandwidths of about 300-3000 Hz.

In the analogue pre-processing hardware the speech sounds are first converted into audio signals by a microphone. As these analogue speech signals pass through the system in a stream there is a continuous identification taking place of the phonetic features (from the range of 16) which they contain, and a complete analysis in terms of these 16 features is produced every 10 milliseconds. The analysis is done in two parts: a qualitative analysis which distinguishes between, for example, voiced sounds and fricative sounds; and a quantitative analysis which measures the strengths of the different constituents, such as the amplitudes of frequency components. For frequency analysis, in addition to conventional filters the system uses autocorrelation analysis as this is more effective with the very short duration frequency components which occur in speech. The information so produced, which consists of identified features in audio signal form, then passes into an interface which, under microprocessor control, converts it into the binary digital form of 64 bits in parallel every 10ms.

Reduction of the data rate to this level allows real-time decision making when the binary information is fed into the digital computer. Here the basic process is to compare the incoming list of phonetic features with stored phonetic specifications for each acceptable word in a pre-determined vocabulary. At another level a list of matches with attached probabilities is stored. Overall match is deduced in terms of the permitted syntax. The vocabulary of the system is 64 different words.

Team leader Brian Pay (standing) and colleague John Yardley with the NPL automatic speech recognition equipment.

Ceefax sub-titles improved

Sub-titles on television are, of course, very useful to the deaf, but until recently have been very difficult to provide. The BBC has improved its production facilities for Ceefax sub-titling so that a half-hour programme can now be sub-titled in about a day, where previously it would have taken at least two and a half days.

The programme is copied onto a video cassette and the master EBU time-code is recorded in a spare audio track. When the cassette is played on the sub-titling equipment, a mini-computer senses the time code, and displays it on a monitor. Also displayed is the programme and a purpose-built keyboard may be used to sub-title the programme by adding the text to the monitor display. Fixed points in the programme may be selected and added to the sub-titles and all the information is stored on a floppy disc through the mini-computer.

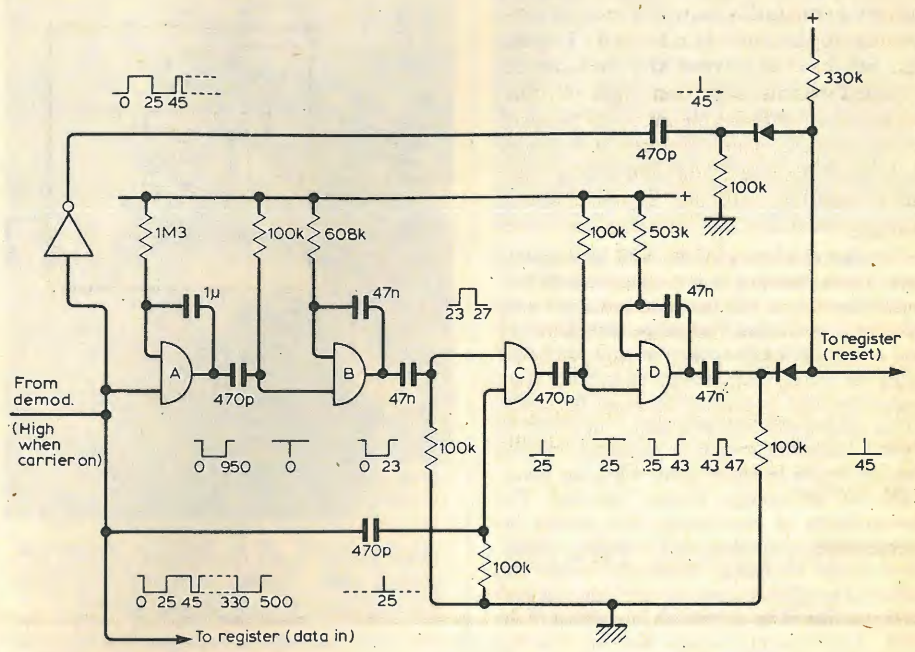
On transmission, the time code on the master tape is fed to a mini-computer where it can be synchronised with the time code on the floppy disc. The sub-titles are then fed to the main Ceefax computer and transmitted at the same time as the programme.

The Thander SC110 portable oscilloscope from Sinclair Electronics Ltd was the only British product to win a gold medal at the Brno Trade Fair in Czechoslovakia. The oscilloscope weighs less than 2½lb and has a 2in c.r.t. It has a basic specification of single trace, 10MHz bandwidth and 10mV sensitivity.

CIRCUIT IDEAS

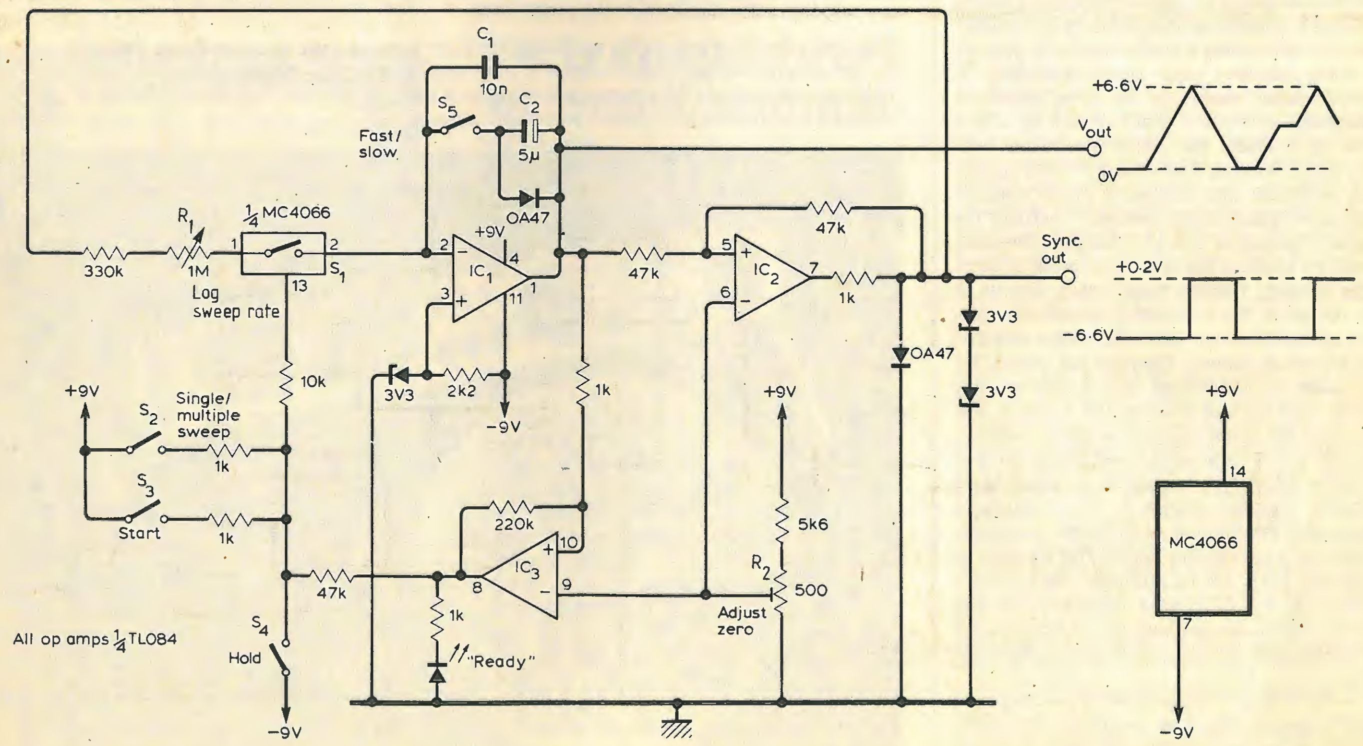
MSF pulse recognition

This circuit is an adaptation of the design by A. F. Cross in Feb. 1976. By using an unorthodox form of monostable, only one 4081 i.c. is used with one inverter. The pulse timings are shown in milliseconds from the carrier on-off transition which marks each second and minute. The 470pf/100kΩ networks at the gate inputs produce short pulses at the times shown. Operation of the RC networks depends on the protection diodes at the gate inputs. Because monostable A is not retriggerable, it protects the circuit from noise signals for 95% of each second. The pulse durations must be measured to determine the values of the monostable timing resistors as these will vary with the regulated supply voltage and the particular 4081 used.
G. Jackson
Cardiff



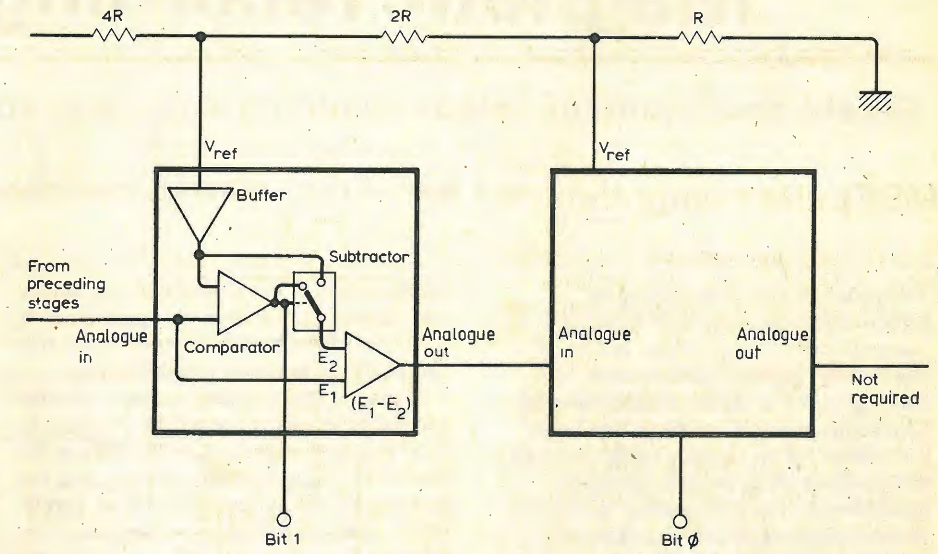
X-Y plotter time-base

With S_2 closed, S_1 is closed and the circuit operates as a standard unipolar triangle-wave generator which is controlled by R_1 and C_1/C_2. When S_2 is opened, the output of IC_1 stays at approximately 0V until S_3 closes S_1. IC_1 then ramps up and sets the Schmitt-trigger output, IC_3, to 9V which holds S_1 closed. The output of IC_3 remains high until the output of IC_1 ramps below the reference level set by R_2. At this point the output of IC_3 goes to −9V and opens S_1. The output of IC_1 then remains at 0V until S_3 is momentarily pressed again. The ramp can be halted at any point in the cycle by closing S_4, and the cycle can be completed by opening the switch. Resistor R_2 is adjusted for minimum zero output level before oscillation starts.
S. Kirby
York



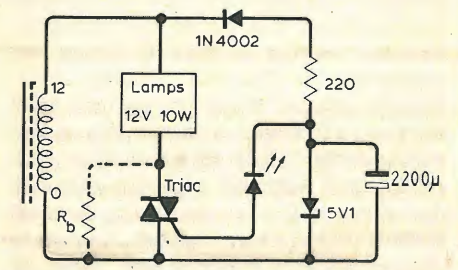
Fast a-to-d converter

By cascading several of the comparator circuits shown, fast conversion times, 8μs for an 8-bit word, can be achieved, and no convert command is required because continuous ripple conversion is used. The design can have any word size and can be extended without alteration to the existing circuit, and the full-scale input voltage can be any required value through one input. Each block has a reference and an analogue input, together with one digital and one analogue output. The reference input is fed from a R-2R-4R ladder and if the analogue input voltage is less than this reference, the digital output goes to zero and the analogue output becomes equal to the input voltage. However, if the analogue input is greater than the reference, the digital output goes high, and the analogue output becomes equal to the analogue input minus the reference voltage.
P. McChesney
Wirral
Merseyside



Simple a.c. lamp flasher

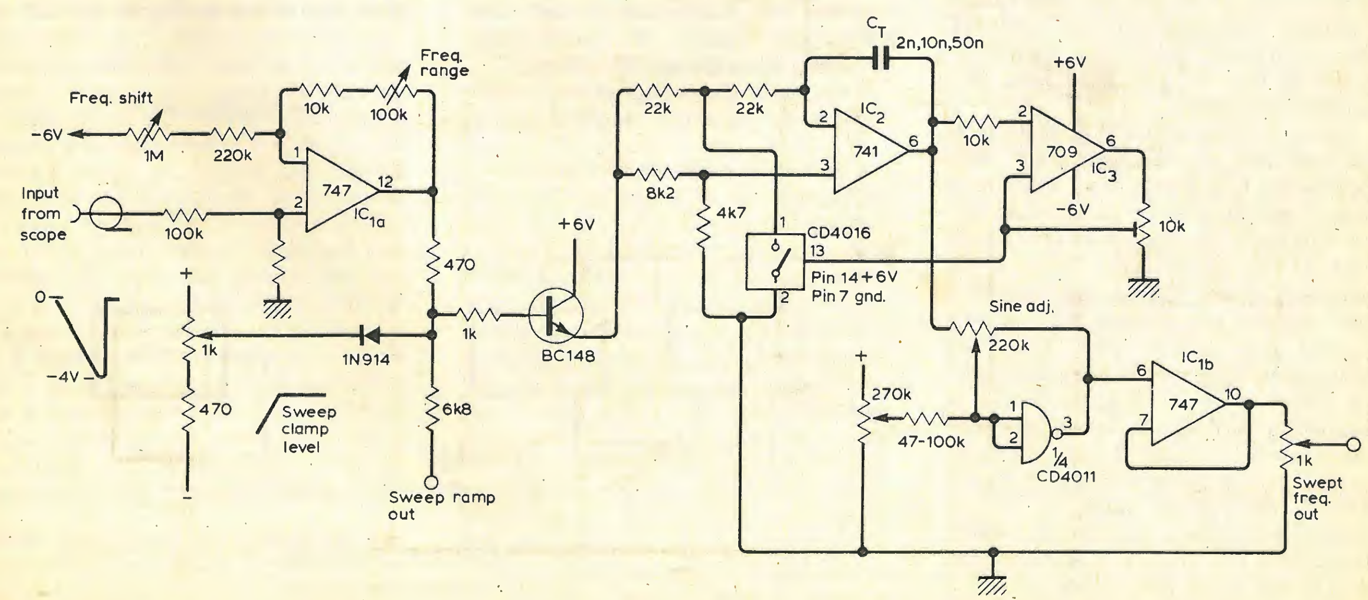
A flashing l.e.d. can be used to produce negative trigger pulses for a full-wave triac switch. To prolong the life of the lamp, a small quiescent current can be provided by R_b.
M. J. Bonner
Feltham
Middlesex



Audio sweep generator

A useful sweep generator can be built by combining the voltage controlled generator by J. W. Howden in Nov 1972, with the single i.c. function generator by J. W. Richter in Nov 1976.

Op-amps IC_2, IC_3 and the c.m.o.s. switch form a voltage controlled triangle-wave generator which operates over three ranges by switching C_T. The triangle wave is shaped by IC_4 to produce a sine wave which is then buffered. An oscilloscope time-base signal provides a convenient sweep input and IC_{1a} converts this to a positive ramp which can be clamped at any point. Frequency range and shift controls are also provided so the sweep range can be set, for example, to 1-100Hz or, by altering the shift control, 10-100Hz. The complete circuit operates from ±6V.
K. Padmanabhan
Madras
India



Magnetic recording review

Recent developments in tape recording in general, and cassettes in particular

by J. Moir, F.I.E.E., James Moir and Associates

The storage and processing of information on magnetic tape or magnetically coated discs is one of the fastest growing industries throughout the industrialized world. Since the growth in usage has been paralleled by an equally rapid growth in the performance of record/replay equipment, it is interesting to review the developments of recent years, paying particular attention to improvements in tape coatings.

Information, in this sense, means both data and audio signals, but the audio aspect takes precedence in this discussion. However, though there are differences in the hardware, problems of storing analogue signals on tape do not differ in any basic way from the problems of storing information in digital form on magnetic discs. The review commences with a summary of the fundamentals of magnetic storage.

For those readers with only a limited involvement in magnetic recording, a resumé of the subject may be useful.

Basics. The recording process is outlined in Fig. 1. The magnetically coated tape passes across the gaps of three ring-type heads in sequence. The first head is energized by a high-frequency 'erase' waveform which magnetically saturates the tape coating, effectively eliminating any previous recording. The second head carries the signal to be recorded, plus a high-frequency bias waveform that 'linearizes' the intrinsically non-linear magnetic-recording process. The third head provides a replay signal.

Signals in either analogue or digital form are impressed on the magnetically clean tape coating by a ferrous head, which has a narrow gap to produce an external field through which the tape passes. The data is read off the tape by replay heads of the same basic design; indeed, many domestic tape recorders use the same head for both record and replay. The tape/head relationship results in the information being stored as variations in the density of magnetization along the tape, a sinusoidal signal leaving a variable density-of-magnetization pattern rather like that in Fig. 2. Though the primary variations in tape magnetization exist along the tape it will be appreciated that there is a small component normal to the tape surface. This is of little significance when recording low signal frequencies, but it is one of the factors that limits the achievable performance when recording high-frequency audio signals or digital inputs at a high bit rate.

Magnetic information storage of this general type was proposed by Poulsen in 1900 and applied practically by Stille in the late 1930s, using steel wire or tape, but the techniques had no real commercial significance until Telefunken in Germany developed ferrous-coated PVC tape as the recording medium, a development that did not become widely known until the end of the last war.

In the thirty years since the commercial appearance of magnetic recording, tape speeds have fallen from the 30in/s (76cm/s) used in the original professional equipment to 1⅞in/s (4.7cm/s) in current domestic cassettes. The performance of a professional cassette recorder is in most respects better than that of the 30in/s professional recorder of 1945/50. Using the appropriate tape, a good example of a modern cassette record/replay machine will have a frequency response that is flat within about ±1dB between 30Hz and 15kHz, with harmonic-type distortions in the region of 2%, a signal/noise ratio of around 50dB and speed-modulation distortions of under 0.1% at a tape speed of 4.7cm/s. Professional ¼in tape recorders of current design have a frequency response flat to within ±1dB from 30Hz to 22kHz, harmonic-type distortions in the range around 1%, a signal/noise ratio in excess of 75dB and speed-modulation distortions in the .05% class. Digital tape recorders will comfortably exceed these performance figures, the speed-related distortions being almost non-existent.

If the maximum amount of data is to be stored in the minimum length of tape at the minimum cost (the logical commercial target) it is obvious that the wavelength of the signal recorded on the tape must be as short as can be achieved. This implies, among other aspects, that the tape speed be low.

The lower limit to the wavelength that can be recorded and reproduced is set by the practical difficulties in producing heads with gap lengths in the region of 2 microns (about .0001in) and on the molecular scale by the dimensions of the smallest magnetizable particle that it is possible to produce for the tape coating. These limits are being approached in television and data recordings, where signals of 15MHz and bit-storage densities of 4-8kbits per mm are now in use.

Though a low tape speed obviously minimizes the cost of the tape, it focuses attention on the transport mechanics. Lack of contact between tape and replay head introduces a high-frequency loss of 55dB per wavelength of separation. At a tape speed of 4.7cm/s, a 10kHz signal has a wavelength of approximately 0.0047mm: a head-to-tape spacing of this amount would result in a loss of 55dB, an intolerable loss even in a machine having no particular pretension to high fidelity.

This is a convenient point at which to outline the changes in the magnetic state of the tape during the recording process. Neglecting the effects of the high-frequency bias in linearizing the recording process, the flux variations in the tape coating follow the usual B/H relation for ferrous materials familiar to all power engineers and shown in Fig. 3, the flux increasing along the curve as the magnetizing force H is increased, but falling back along a different path as the magnetizing force decreases. When H has fallen to zero it will be seen that the flux still has the value B_2, the remanence expressed in gauss (in SI units). To reduce this residual flux to zero, the magnetizing force has to be reversed and increased to H_3, this value being known as the coercivity (expressed in oersteds).

Remanence and coercivity are of primary importance in indicating the performance of recording tape. The remanence is significant in indicating the flux amplitude that remains in the tape coating after magnetization under the record head, for the signal obtained on replay is directly proportional to it. The importance of high coercivity is less obvious. It indicates the extent to which the recorded tape coating will resist demagnetization, in the record head gap, by the high-frequency bias field and by the low-frequency signal field that extends well beyond the point in the tape path at which the higher-frequency signals are impressed on the tape. From this aspect, high coercivity is essential in ensuring a good high-frequency performance, but it does necessitate the dissipation of higher power in the erase head and a higher value of bias current in the record head to ensure the optimum performance.

Heads. Record and replay head design and construction have greatly improved in the last few years and both aspects justify some brief comment. Narrow head gaps are necessary if a good frequency response is to be achieved. This is made particularly obvious if the replay process is considered. At the frequency at which the gap in the core covers two half cycles of the recorded flux wave there will be no net flux through the core and no output will be generated. The relation between gap width and output voltage has been investigated by several workers and it has been shown that the output voltage/gap-width relation is as shown in Fig. 4, the first zero appearing at the frequency at which the gap is two half wavelengths wide.

$$V_{out} = \frac{\sin \pi d}{\lambda} \Big/ \frac{\pi d}{\lambda}$$

A similar relation applies if the head gap is not at right-angles to the edge of the tape, the first zero in the response occurring at the frequency at which one edge of the recorded track is two half waves ahead of the other edge. Thus the basic design requirement for a good frequency response is a narrow and dimensionally uniform gap which is at right-angles to the guided edge of the tape. Small size, high resistance to abrasion by the tape coating and of course, high efficiency obtained by a magnetic design that ensures that a high percentage of the available short circuit tape flux passes through the head core are all desirable design targets. A high degree of rejection of external magnetic fields is an additional practical advantage in minimizing hum pickup.

The same general requirements apply to the design of the record head, complicated to some extent by the need to dissipate several watts from the record amplifier without undue temperature rise. Recent developments in tape coatings have generally increased the power required to produce magnetic saturation of the coating, and have thereby increased the head designer's problems. Achieving gap widths of around 10 microns in record heads is not quite as important as in replay heads, because it has been shown that the magnetic signal remaining on the tape following its passage over the record head is due to the combination of record and bias fields that exist at a point beyond the record head gap. Thus, the recorded frequency response is largely a function of gap uniformity rather than mean gap width. However, it is common practice in domestic machines to use a single head for both record and replay and to accept the compromises that are then necessary.

Apart from the need to achieve narrow and uniform gaps in the heads, there is an obvious requirement to minimize hysteresis and eddy-current losses in the head material. Designers usually achieve this by assembling the head from thin laminations of a low-loss, nickel-iron alloy, or increasingly by moulding the heads from one of the ferrite derivatives.

Bias It is necessary to 'linearize' the basic magnetic recording process, because the relation between the applied magnetizing force and the resultant flux density is non-linear. The basic relation between the applied magnetizing force H and the resultant magnetic flux B, in any iron circuit has the familiar form shown in Fig. 3. If no remedial measures were taken this would result in the transfer characteristic, the input/output relation outlined in Fig. 5. The gross non-linearity of this relation is not acceptable in an analogue system, so it is standard practice to linearize the transfer characteristic by applying a high-frequency (80-150kHz) bias waveform to the record head, in parallel with the signal. The linearizing process is very effective, for the overall distortion generated by the system non-linearity can be in the region of 3%-5%, with a signal/noise ratio exceeding 55dB, even with domestic cassette record/replay equipment. Experience suggests

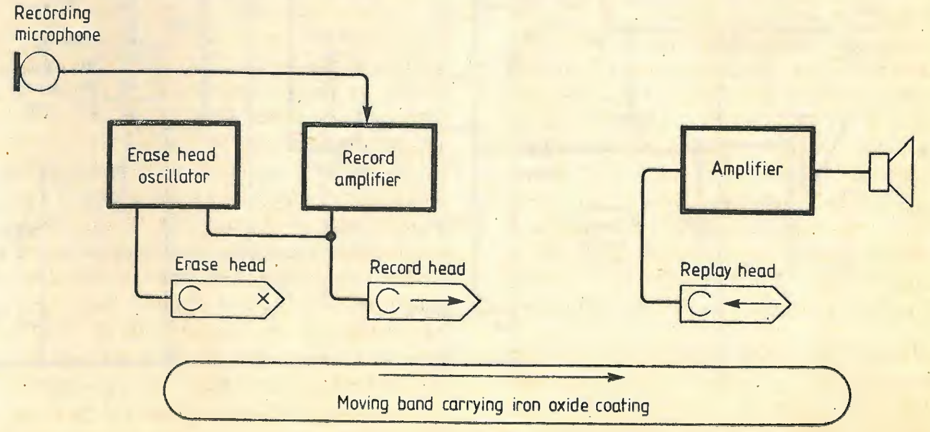
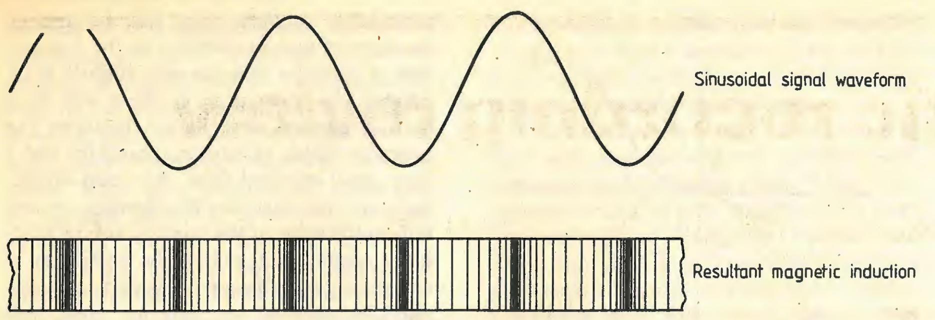


Fig. 1. Tape/heads schematic



A sinusoidal signal produces a sinusoidal distribution of flux along the tape

Fig. 2. Density variations along the tape

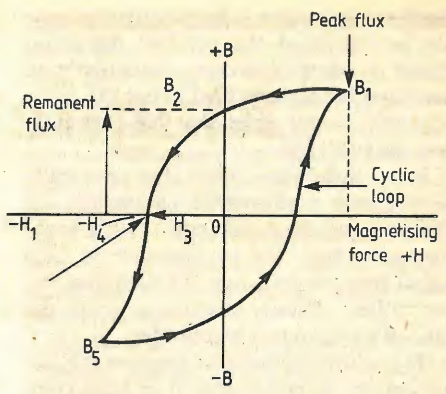


Fig. 3. B/H curve

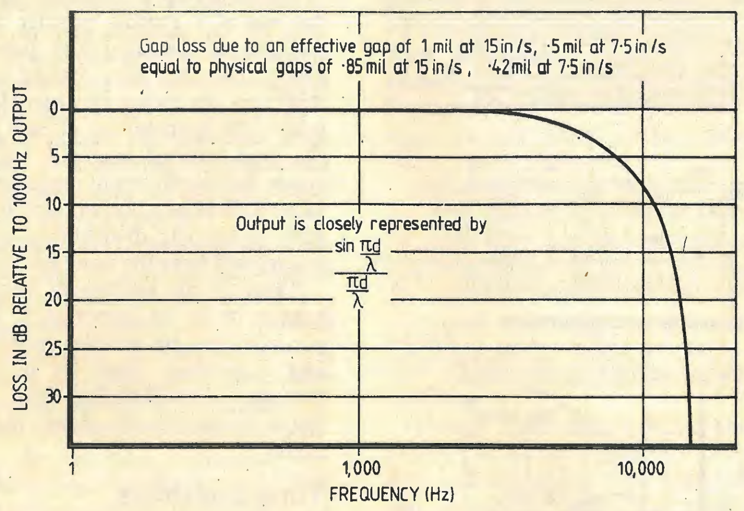


Fig. 4. Gap loss

that head saturation is frequently responsible for much of the residual distortion found in many recorders, particularly in machines hastily modified to use the high-coercivity metal tapes that have recently become available.

It is a little unfortunate that practically all the tape performance parameters are functions of the amplitude of this high-frequency bias, the performance at high signal frequencies (short wavelengths) being rather critically dependent upon the ratio of signal to bias amplitudes.

Though the actual bias frequency is not important, provided that it is high compared to the highest signal frequency, the waveform of this bias signal is very significant. Any waveform assymmetry implies the presence of a d.c. component, which results in some residual magnetization of the tape and an increase in the intensity of the magnetic component of the total system noise. It appears almost impossible to avoid some increase in tape noise due to this residual magnetization, bulk erased tape generally being at least 2dB quieter than tape erased on the recorder.

Noise. Finally comes the question of the noise generated by the tape system. There are two main sources of noise; that generated by the electronic circuitry in the early stages of the replay amplifier and that due to the magnetic characteristics of the tape. Circuit noise will not be considered in any detail, for the mechanism that produces the noise is no different to that producing noise in any amplifier system. In good machines, the thermal agitation and $1/f$ noise due to the amplifier is at least 10dB below the magnetic noise produced by the tape coating, and is therefore of no great consequence.

The basic system signal/noise ratio can be improved either by an increase in the amplitude of the signal recorded on the tape or by a decrease in the residual 'magnetic' noise that results from the passage of magnetically clean tape across the replay head gap. An increase in the amplitude of the recorded signal can be achieved by increasing the width of the recorded track, or by an increase in the remanent flux density of the tape coating, the limit to any increase in the flux density being set by the consequent harmonic distortion. Doubling the track width doubles the replay signal amplitude but magnetic noise, being a random phenomena, is proportional to the square root of the track width. In consequence halving the track width decreases the signal level by 6dB but reduces the tape noise by 3dB. Thus, halving the track width decreases the s:n ratio by only 3dB but may double the amount of data that can be stored on a given area of tape.

It will be seen from a later discussion that much of the s:n performance lost by the track-width reduction that has occurred in recent years has been regained by improvement in head design and tape coatings during the same period. Narrow tracks are now in very widespread use. The internationally standardized audio cassette employs four tracks, each 0.66mm wide with guard bands 0.35mm wide between adjacent tracks. A s:n ratio of about

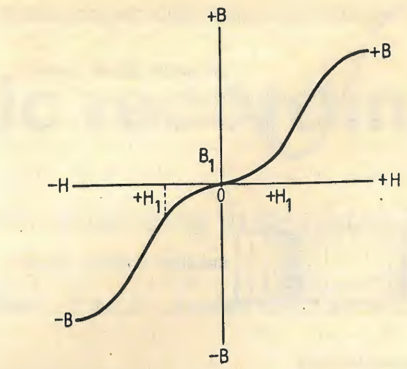
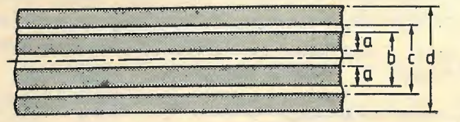
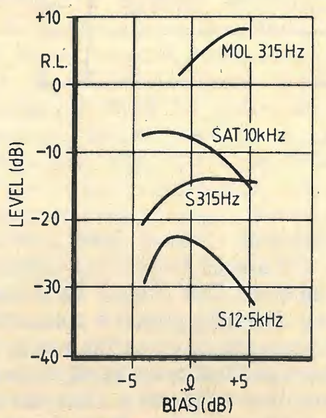


Fig. 5. Transfer characteristic



	mm		in	
	max.	min.	max.	min.
a	0·66	0·56	0·026	0·022
b	2·00	1·80	0·079	0·070
c	2·70	2·40	0·106	0·094
d	3·81	3·66	0·150	0·144

Fig. 6. Cassette track geometry



Measuring Conditions

Reference level : 250nWb/m
Tape speed : 4·76cm/s
Recording head : FXC 4μm
Replay head : FXC 2μm
Trackwidth replay head : 1·5mm
Equalisation : 3180μs + 120μs
DIN calibration tape : T 308S

MOL 315 : Maximum Output Level at 315Hz with 5% distortion
SAT 10k : Maximum Output Level at 10kHz

Fig. 7. Typical tape performance data (Philips Super-Ferro Data)

58dB can be achieved without the use of any electronic enhancement techniques (Dolby, etc.). The track geometry employed in audio cassettes is illustrated in Fig. 6. Data storage designs can achieve an adequate s:n from tracks that are only 0.1mm wide.

The tape-coating noise is a combination of several effects. For minimum noise, the unrecorded virgin tape coating should be a smooth, uniformly dispersed aggregate of small magnetic particles, displaying no external magnetic field. In virgin tape it is assumed that all the individual particles will either be completely free of residual magnetism or that there will be a large measure of non-uniformity in the orientation of particles that are only slightly magnetized. In consequence, there will be a high degree of cancellation between the magnetic fields of adjacent particles and a very small external field. Any non-uniformity in the particle distribution or any non-uniformity in the particle size or magnetic moment will reduce the possibility of complete cancellation between adjacent particles and a net external field will appear at the coating surface and induce a noise pulse in the replay signal as the magnetic non-uniformity passes the replay head gap.

The first step in reducing tape noise is the use of a coating powder which has particles of the minimum possible size, evenly dispersed as a coating of uniform thickness. Reducing the particle size will reduce the contribution of each particle to the total external magnetic field and increase the possibility of cancellation of the external field of adjacent particles. Uniformity of particle distribution in the tape coating will have the same advantage.

There is an additional factor that is known to be of importance. Producing particles that are acicular (needle-shaped) and orientating them by passing them through a magnetizing field has been found to significantly reduce the magnetic noise.

Tape coatings

At this point it is convenient to consider the magnetic properties that are desirable in a good tape and to consider just what properties should be improved by the tape coating formulators.

Tape coating design has been aimed at increasing the value of the remanence, since this results in a proportional increase in the output voltage from the replay head, but this has resulted in coatings of higher coercivity. The increased erase and bias field strength that are necessary cannot be provided by the majority of earlier machines, a limitation that will be covered in greater detail later in this review. The ideal tape coating would have a 'square' BH curve, the remanent value of flux density closely approximating the peak value —an ideal that is being approached by recently introduced tapes. 'Squareness' values, the ratio of remanent to peak flux, is now in the region of 0.80 to 0.9.

In summary, the tape coating designer is attempting to develop a magnetic medium that has high remanence, 'highish' coercivity and low noise and in this they are being very successful. The coating design technique justifies extended consideration.

The first coatings, developed in Germany during the second World War, were gamma-phase iron oxide, (Fe_2O_3) or jeweller's rouge, a material widely available, with grinding techniques well developed by the paint industry and, in consequence, available at low cost. The magnetic characteristics were reasonably good, remanence values around 900 gauss being achieved with coercivities of about 300 oersted. Development for tape-coating purposes was aimed at improving the magnetic characteristics of this gamma-phase oxide, the grinding process being modified to produce smaller particles. Techniques for shaping the particles to give a high length-to-width ratio were developed and coating uniformity procedures improved. All this initial work was devoted to improving 0.25in tape for use in open-reel machines running at speeds of 7.5in/s and higher.

There are obvious cost advantages in running the tape at lower speeds, but this results in recordings of shorter wavelengths that are increasingly difficult to record. The introduction of the cassette format, using tape running at 4.76mm/s (1⅞in/s) with tracks only 0.66mm wide, and the rapid acceptance of this format by the public focused attention on the need for tape formulations that had an adequate performance at these low speeds, implying wavelengths of 0.02mm at 20kHz. As a parallel development, the basic noise from the tape had to be reduced to compensate for the reduction in the s:n ratio produced by the reduction in trackwidth.

The standard gamma-phase ferric oxide was further improved by the addition of small amounts of other metals such as cobalt, manganese, nickel, zinc, titanium and chromium and by improvements in the process of precipitating the basic ferric oxide from the primary salt solutions. Cobalt proved to be the most effective additive, but the resulting compound was found to have magnetic characteristics that were rather temperature dependent. At room temperatures, remanence values of 1500 gauss and coercivities around 300 to 600 oersted could be achieved. Coatings in which the cobalt was absorbed into the surface of the oxide particle proved to be less temperature dependent than ferric/cobalt compounds and could be designed to achieve coercivities in the range of 300 to 600 Oe with remanence values of 1500 gauss.

Chromium dioxide has magnetic properties that are superior to those of ferric oxide and has been widely used. Coercivities as high as 500 Oe, with remanent flux densities around 1500 gauss can be achieved, but the basic material is expensive. The oxide particles are very uniform in size, have a high ratio of length-to-width, have an excellent short wavelength performance and are easily orientated. In the early products the magnetic properties exhibited some time dependence, but this aspect has been greatly improved in the current product. While the basic oxide is highly abrasive and indeed is the cleaning material used as the coating on several head-cleaning tapes, head wear has proved to be of little significance in programme tapes. This is partly due to the high binder content in most coatings and to the use of lubricants and various surface smoothing and polishing treatments.

Two-layer tapes have been developed in which the top layer is chromium dioxide, 1μ thick, over a 5μm coating of ferric oxide. This locates the expensive chromium dioxide coating adjacent to the head gap where its excellent short-wavelength performance can be of maximum value in ensuring high output at high frequencies, but experience suggest that the boundary of the two layers appears to be magnetically-visible, raising some problems of frequency-response equalization.

This brief survey brings us almost up to date with comment about metal particle tape and metal coating tapes. Tape coatings consisting of pure iron particles have outstandingly good magnetic qualities and can be produced with particles having a high ratio of length-to-width, but — and it is a rather large 'but' — the particles tend to react very actively with moisture in the air, forming ordinary rust. There are similar adverse reactions with the tape base and with the conventional tape binder. It is this chemical activity that has delayed the appearance of pure iron coatings in spite of their outstanding magnetic properties. However, the problems are being rapidly overcome and several pure-iron-coated tapes are now commercially available, albeit at a fairly high price.

Experimental tapes have been produced in which the metal coating is plated on to the tape. Very thin coatings with excellent magnetic properties can be produced and the basic tape noise is very low, but corrosion, poor wear resistance and poor reproducibility are problems yet to be solved before such coatings become commercially available.

Though these metal particle and metal coatings have high remanence, low distortion and low noise, their high coercivity and consequent high bias requirements make it difficult to take advantage of the excellent intrinsic performance.

Coating thickness has a significant effect on the magnetic performance, particularly in respect of frequency response and 'print-through', the tendency for recorded signals to be transferred to the adjacent layers of a spooled tape. Print-through will obviously be reduced by any increase in the thickness of either the tape base or the coating.

The effect of coating thickness change on the overall frequency response is relatively obvious. The short-wavelength, high-frequency flux is confined to the surface of the tape in the vicinity of the gap, whereas the long-wavelength, low-frequency field extends more deeply into the tape coating. At the highest frequencies, the flux only penetrates into the surface layer adjacent to the head. In consequence, an increase in the coating thickness only increases the signal output at low and middle frequencies. This is of value in those applications where the required frequency response is limited, but where the frequency response of the system must extend to frequencies of 15 or 20kHz, it is advantageous to use a thin coating to minimize the amount of electrical equalization that would be necessary to achieve a uniform and wide frequency response from a thick coating.

Bias settings

As noted earlier, the recent developments in tape coatings have resulted in tapes of increased coercivity, requiring higher bias levels for optimum performance. Many millions of machines were in the hands of the public before the high-bias tapes became available and so very few of these earlier recorders can provide the bias and erase field strengths needed for optimum performance from the metal tapes now available. This is very significant, for not only are these early low-bias machines incapable of realising the performance that is intrinsically available from the high bias tapes, but they can only achieve a performance that is significantly inferior to that obtainable from the much cheaper low-bias ferric tapes.

Analysis of the performance of several hundred samples of currently available cassette tapes shows that they can be divided into five classes based on their bias requirements, with the simple ferric oxide coated tapes requiring the minimum bias and the metal tapes requiring the maximum bias. It is not possible to express the bias requirements of coated tape in any really fundamental units because it depends on coating thickness in a manner that cannot be adequately indicated by magnetic measurements on the bulk coating material, so indirect methods are in universal use.

Two types of 'reference' tape have been standardized and the major tape manufacturers have agreed to keep samples of these available as secondary standards. The bias requirements and many other performance parameters of all the commercially available tapes are then indicated by comparing them with 'reference' tapes. Note that the 'reference' tapes are not examples of tapes with an ideal performance. Tape producers are perfectly free to market tapes with a performance that they consider advantageous in any respect, but the performance is indicated in the data sheets by quoting each parameter in dB with respect to the equivalent performance of the specified 'reference' tape.

The catalogue performance data is usually presented as a series of curves relating bias and the various performance parameters, a typical example of published data being shown in Fig. 7. A brief explanation is all that is necessary. The top curve show the variations in the maximum output level (m.o.l.) with bias for an arbitrary distortion content of 5% at a frequency of 315Hz. It will be seen that the m.o.l. has its maximum at a bias value 5dB above the bias required by the reference tape, in this instance the DIN T308S example. The next curve SAT 10kHz illustrates the variation in the m.o.l. at a frequency of 10kHz and it illustrates one of the major limitations in low-speed tape recording, for the maximum available output at 10kHz is falling off fairly rapidly with increase in the bias field above −2dB, whereas the m.o.l. at the low frequency of 315Hz is continuing to rise with increases in the bias. These relations imply that any applied signal having a fairly flat frequency spectrum will have the high-frequency, high-amplitude components in the signal compressed by magnetic saturation of the tape coating. It also implies that the overall frequency-response relation will vary with the level of the applied signal, a limitation that is unusual in any other component of a hi-fi system. The performance of a typical tape at two different signal levels is illustrated in Fig. 8.

Because all the performance parameters are so bias dependent it will be appreciated that some compromise is necessary in selecting the value of the working bias. This particular tape, with the characteristics quoted in Fig. 7 and biased to +5dB, will have a maximum low-frequency output some 5dB higher than if biassed to 0dB, but the output will be down by 3-4dB at frequencies in the 10kHz region and some 5dB down at 12.5kHz. Thus, the bias point can be chosen to achieve high mid- and low-frequency output and a relatively high value of signal/noise ratio by biassing at +5dB, but at the expense of a frequency response that is falling off significantly at frequencies above 10kHz. Alternatively, the bias can be set to −2dB to give a frequency response substantially uniform to well above 12.6kHz at the expense of a s/n ratio that is 4-6dB below the maximum obtainable at the higher bias level of +5dB.

There is little uniformity of practice among tape manufacturers in selecting the bias they quote as the optimum, or in specifying the process to be followed in determining the optimum bias. Each supplier has his own ideas about the relative importance of wide frequency response and a high signal/noise ratio and it is

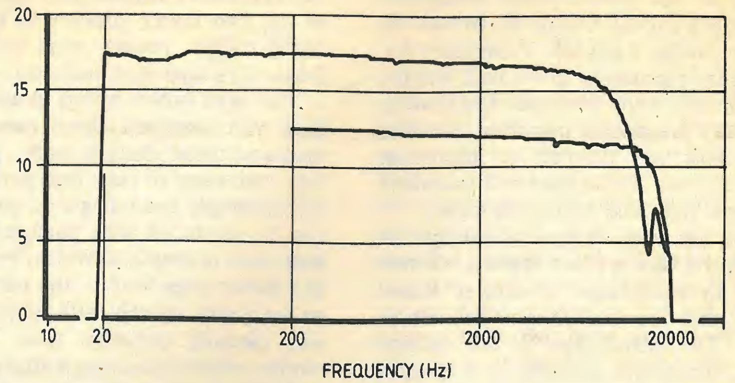


Fig. 8. Response at two signal levels

the relative importance of these factors that controls the bias that is recommended. Simple practice, such as the adjustment of the bias until the maximum undistorted 315Hz or 1kHz output is achieved and then reducing the bias by anything from two to five dB, is a common recommendation with some reasonable foundation when applied to tape such as that in the previous discussion about the effects of bias. However, the low-frequency output changes only slowly with change of bias, a result that has led to some designers suggesting that the bias be adjusted to give a flat frequency response by alternately switching between a 330Hz and a 10kHz or 15kHz signal and selecting the bias point that gives an equal output of both frequencies.

Some tape recorders include facilities for setting the bias in this way, a simple signal generator switchable between two frequencies being included. The previous discussion will have suggested that though the final bias will ensure a flat frequency response to the chosen high frequency, the s/n ratio is likely to be lower than can be achieved by selecting a higher bias setting and tolerating some loss of the higher frequencies. □

Weather effects on orthogonal polarization

Finding ways of making satellite communications more efficient is one of the many tasks of the research laboratories of British Telecom, who made it clear at their open days in September that they do not intend to reduce their efforts in research and development even though times may be hard.

Current satellite communications systems use mainly the 4 and 6GHz bands which are, even now, showing signs of becoming saturated. The obvious answer to the problem of band saturation is to use another frequency, but as frequencies lower than 4GHz are not practical for present systems, higher frequencies are the only option. However, signal attenuation effects caused by certain weather conditions become even more of a problem as the frequency is increased. British Telecom are investigating these effects at their Martlesham Heath laboratories, not only because of their increasing influences on the 12 and 14 GHz bands, which will be used for the next generation of satellite communications systems, but also because of their depolarizing effects on orthogonally polarized satellite communications signals.

Research into the feasibility of channel sharing by means of orthogonal polarization is well under way and involves studies into various aerial designs and interference problems, as well as the weather aspect. The main questions to be answered are: can the cross-polar isolation between two channels sharing the same frequency be maintained with sufficient reliability; and how does the choice of either linear or circular polarization (the two methods at present being compared in the evaluation) affect the reliability?

Radar and radiometer stations together with rain gauges etc., are used to collect data on the actual weather conditions, but the main data is obtained from observations on a link between the ESA Orbital Test Satellite and an earth station to determine the amounts of signal attenuation and depolarization using a 29.9° elevated path. Under good weather conditions both circular and linear methods of polarization exhibit cross-polar discrimination figures of

Monitoring of trans-horizon interference provides information needed to improve predictions for transmission/reception quality between earth stations and to study the effects of weather on microwave radio paths. Ron Fitzgerald is seen here lining up an aerial to measure over-the-horizon interference at British Telecom's Martlesham laboratories.

around 35dB. Rainfall in the signal path seems to have the greatest effect on reliability due both to its frequent occurrence and its attenuating and depolarization properties.

According to the researchers, a peak rainfall rate of around 140mm/h caused a maximum of −20dB attenuation using circular polarization at 14.5GHz, accompanied by a minimum cross-polar discrimination of 10dB. Using linear polarization, the same rainfall rate caused only a maximum of −12dB attenuation and a minimum of 25dB isolation using 11.6GHz. High altitude ice crystals give hardly any attenuation but a significant phase shift. During juperiod of high ice-crystal levels, 14.5 and 11.8GHz frequency signals which were circularly polarized showed minimum cross-polar discriminations of 15dB while a linear polarized 11.6GHz signal at the same time gave a minimum of 28dB. Dry snow showed similar characteristics to those for ice crystals, causing worst cross-polar discrimination figures for circular and linear polarization of 18 and 29dB respectively. Wet snow on the aerial caused figures of 10dB cross-polar discrimination for circular polarization and 15dB for linear.

Results obtained using linear orthogonal polarization have proved that the required reliability of channel isolation can be obtained almost constantly and this method will be used as a means of channel sharing in future communications satellite systems, says British Telecom. Circular polarization does have an advantage, though, in that it allows greater aerial misalignment. In fact, the ITU plan for European satellite broadcast systems indicates that circular polarization of broadcast signals will be used, mainly because of aerial alignment problems in domestic applications. □

LETTERS TO THE EDITOR

INTERFERENCE FROM MICROS

With microprocessor based toys, trainers and home computers being such a growth industry, I have recently become rather disturbed by the radio frequency radiation produced by these devices.

For some years I have been using a scientific programmable calculator which has considerable radiation at 500kHz – certainly this would create havoc aboard a ship with the poor radio operator listening out on the international calling and distress frequency of 500kHz.

Probably 500kHz radiation is an exception, but most microprocessors use a clock frequency around 1MHz with the internal signals having very fast rise times. Consequently there is not inconsiderable radiation of harmonics.

Both my Pet computer and my microprocessor evaluation system are unscreened and radiate significant energy right up to 200MHz or more – a most irritating form of interference on v.h.f.

I wonder if other readers would care to comment on this problem and if there should be some legislation to control interference from microprocessor based systems?

Hugh D. Ford
Richmond
Surrey

SATELLITE ORBIT PREDICTIONS

I was interested to read the brief article by M. L. Christieson in the December 1980 issue concerning orbit predictions from satellite images in view of our experiences in this field at the University of Surrey.

We have been involved for several years in the day-to-day tracking of US and USSR meteorological spacecraft and the derivation of orbital parameters and ephermeris as this data is not always readily available to the experimental observer. We have also expended considerable effort on a similar exercise concerned with the AMSAT-OSCAR series of amateur communication satellites in order to provide medium term (three months) orbital calendars primarily intended for the amateur radio fraternity, distributed through AMSAT-UK.

The determination of accurate orbital ephemeris using the transponders and beacons on board the Oscar spacecraft has proved to be awkward and imprecise; however, the real-time imaging facility on board the meteorological spacecraft has certainly provided a most convenient means for accurate observation. We have regularly used the observation of sub-satellite imagery correlated with a Droitwich time source which has enabled us to determine and predict the ephemeris of the meteorological spacecraft to within a few seconds in the course of several weeks and certainly to within two minutes over the entire three month period covered by the orbital calendar. We have found in practical terms that probably the most precise measurements can be taken by observing the image in real time and choosing a well defined landmark as near the equator as possible. Gibraltar has proved particularly useful in this regard and it has enabled timing to be established with ease to within one second.

The data thus collected and correlated over the course of the lifetime of the various spacecraft observed has enabled us to define the primary orbital parameters and their derivations to some accuracy. Analysis of the effect of atmospheric drag on the low orbiting spacecraft NOAA-6, TIROS-N and OSCAR-8 (all around 800-900km altitude) over several years has resulted in a good estimation of effect of the drag on the orbital ephemeris.

In conclusion, I would certainly recommend this method as it has on occasions proved more accurate than the official data and has been particularly useful during the observation of USSR meteorological spacecraft where no official data has been forthcoming.

M. N. Sweeting, G3YJO
Department of Electronic & Electrical Engineering
University of Surrey
Guildford

Reference
'Orbit' Nov/Dec 1980 (AMSAT). The Project OSCAR Orbital Calendar pp. 75/26.

I enjoyed reading M. L. Christieson's article on weather satellites in the December 1980 issue and no doubt it will be of interest of many budding satellite trackers.

We would however like to correct the statement that "Further data can be obtained from outside but this is not always easy to obtain". Rubbish! M. L. Christieson has not done his homework in the amateur sector. We issue TIROS-N and NOAA-6 orbital predicts every three months which are very accurate. These are computer derived to 10^{-19}. From the tables enclosed you will see that these also include the two Oscars currently in orbit. Other information on the activities of the Radio Amateur Satellite Organisation of the United Kingdom can be obtained from myself by sending a stamped and addressed envelope to the address below.

The cost of the three months Oscar calendar (TIROS-N/NOAA-6 included) is £1.27 postage and packing included.

Ronald J. C. Broadbent
AMSAT-UK
94 Herongate Road, Wanstead Park
London E12 5EQ

The author replies:

I am glad that Mr Broadbent enjoyed reading my article and that it has been the means by which his service has been publicized. He is quite correct that I was unaware that AMSAT publish predictions for TIROS-N and NOAA-6. I fear, however, that he may have missed the point. Long term predictions, as I explained, are subject to considerable drift. Those issued by AMSAT are no different, and are at present approximately 11 minutes (E.C.T.) and 8 degrees (longitude) in error. The method I described can correct this.

Mr Broadbent also mentions "10^{-19}" in connection with prediction accuracy. This refers, of course, to the *precision* to which the computer can calculate, and bears little relationship to the accuracy of the prediction; these two are often confused.

M. L. Christieson

THE DEATH OF ELECTRIC CURRENT

If Ivor Catt had read physics instead of engineering he might not now spend so much time agonizing over the "right" mechanistic model for processes which, in reality, are outside the area of our everyday perception (December 1980 issue, p. 79). It is true that what is happening in a charged capacitor can be considered to be the result of interference between two waves travelling in opposite directions, but it is easier to consider it as a distribution of charged particles. Similarly, one can map the currents flowing in the walls of a waveguide, but only a fool would treat waveguide theory in terms of current electricity.

E and H have no more physical reality than do ρ and J, being merely constructs in mathematical models. The usefulness of any mathematical model is measured by the accuracy of its predictions and the ease with which those predictions may be obtained. Although there is no real difference between predictions from the two models, in general it is easiest to use current theory for low frequency, or long term or continuous situations and e-m waves for high frequency or short duration or quantized situations. (This is a broad generalization and, like all such, has exceptions, so please don't rush to quote them at me!)

The proof of any pudding is in the eating. The machines on which much of our civilization is founded (that is, alternators and motors) are designed with the aid of the electric current model. You may not like civilization, but, clearly, the designs work. However, to say that electromagnetic theory has been ignored and suppressed is blatantly untrue. Where appropriate, e-m theory has been used in design; the delay-line modulator, developed to pulse radar magnetrons, is an example. That there are few others that spring to mind is indicative only of the historical superiority of the current model.

To use an overworked phrase, it is simply a matter of horses for courses. Mr Catt and his colleagues believe that the digital microelectronics course is one for which their rediscovered horse is superior. This may be so, but it is hardly justification for an attempt to put down the other contestant.

R. T. Lamb
Post Office College of Engineering Studies
Milton Keynes, Bucks

The author replies:

In his first paragraph, I think Mr Lamb has reversed physicists and engineers. I find the "charged particles" in a capacitor very difficult to consider, in view of their apparant need to shoot off instantaneously at the speed of light (for the dielectric) from a standing start when the capacitor is discharged. I wonder if our brothers the electrons consider it easy; or can a TEM step advance down a transmission line at the speed of light *without* any electrons being required to change velocity so abruptly? I consider such questions far from easy – hence Theory C.

Regarding para. 2, if neither E, H, ρ nor J have physical reality, then what does have physical reality? You seem to have ruled out the physical reality of electromagnetism – a far bolder step than my modest Theory C, which merely gets rid of ρ and *J*. The first sentence of this paragraph could come straight out of Osiander or Berkeley, and is discussed by Popper under the title "The Science of Galileo and its new betrayal" (K. Popper, "Conjectures and refutations", RKP 1963, page 97. See also M. Polanyi, "Personal Knowledge", RKP 1958, pages 145-147). I agree with Kepler that "It is indeed a most absurd fiction to explain natural phenomena by false causes." Bruno was burnt alive for taking this stand against the mediaeval church (and Lamb).

Regarding para. 3, as with Lamb, my work on alternators, motors etc. never led me to question Theory N. However, my work on high speed logic *did*. I have not said that electromagnetic theory has been ignored and suppressed. As to the suppression of Heaviside, you will not find mention of him in books on electromagnetic theory published during the last fifty years. It is scandalous the way he has been ignored and suppressed, in view of the great contribution he made to the subject. (Lamb seems to call Theory N "the current model" and Theory H "e-m theory".)

Regarding para. 4, I am perfectly happy to see people use Ohm's Law and current meters far into the future. I shall do so myself. This is not the same question as fundamental theory. Theory H has been re-discovered and found valuable in digital electronics. Theory C has only recently been discovered.

SAWTOOTH KEYING FOR ORGANS

Dr Colin Pykett's articles on organ-stop filters (October, December 1980) have been interesting and informative. Clearly the filter method ought to be the most direct way of getting complex tones, but no one seems to have tackled the problems systematically before. He did tell me though that my flutes were a little better than his!

He mentions the difficulty of keying a sawtooth without distortion; a typical d.c.-controlled keying circuit takes successive slices, so that during attack and decay the output has the shape of Fig. 1. Only a pulse or square-wave keeps its shape, and an attempt to key a sine-wave results in heavy clipping. Nevertheless, some constructors have used this form of keying for sawtooth sources, and it may be interesting to look at it more closely.

Since all harmonics are present in the sawtooth, clipping it can introduce no new ones (as it does with a rounded waveform), but can only alter their proportions. One method of analysing the Fig. 1 waveform is to consider it as the sum of pulses of decreasing width, all finishing together, and use the standard pulse formula. I conclude that the amplitude h_n of the nth harmonic in a truncated sawtooth of height k is:

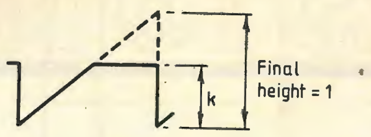
$$h_n=\frac{1}{n\pi}\sqrt{a^2+b^2}$$

where $a=\frac{1}{2\pi n}(1-\cos 2\pi nk)$

and $b=k-\frac{\sin 2\pi nk}{2\pi n}$

In a true sawtooth of height k (ideal keying) the nth harmonic amplitude is $k/n\pi$, and the ratio of the two is $\sqrt{a^2+b^2}/k$. Fig. 2 shows this ratio in dB for the first three harmonics; higher harmonics continue the same pattern, having more oscillations, and settling sooner to within 2dB of the sawtooth amplitude. Thus it seems likely that the output timbre, from a filter using many harmonics, will not be much affected during attack and decay; the initial delay of the

Fig 1



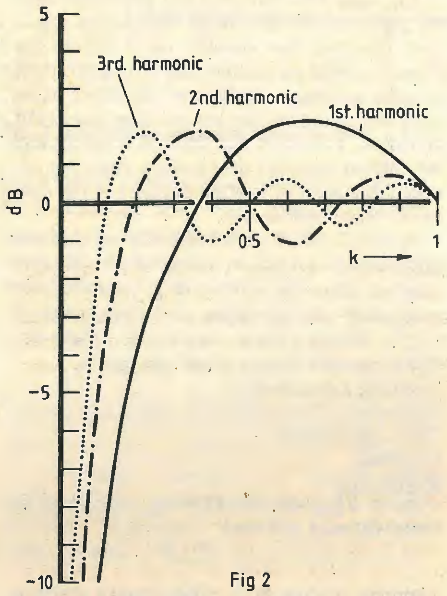


Fig 2

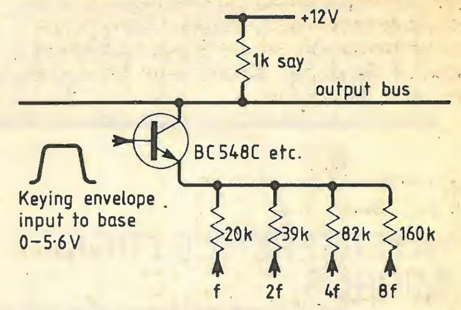


Fig 3

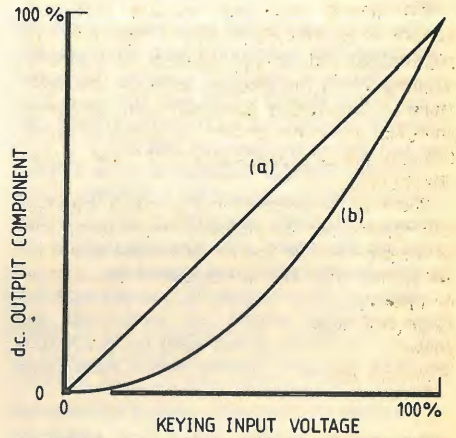


Fig 4

fundamental represents a kind of chiff at no extra cost. The effect on flutes and fluty diapasons may be more noticeable.

The concept can lead to considerable economies, and Fig. 3 shows a possible circuit using a single transistor to mix (at the emitter) the components of a staircase wave, and key them into the common busbar load. More information on this form of keying is given in part 2 of my organ series, *WW* Nov. 1978. Using a transistor for each square-wave frequency (bases and collectors commoned) would preserve the true staircase shape during keying; the single transistor produces truncation, as it would if the emitter source were a true sawtooth from 0 to 5V.

Unbalanced transistor (or diode) gates output a d.c. component as well as the signal. For a square-wave input, the d.c. is proportional to the keying input, as at (a), Fig. 4, and this also applies to a staircase keyed by separate transistors. For a single transistor with sawtooth input its growth is parabolic, curve (b), reaching twice the slope; this is also substantially true of Fig. 3, and might limit the usable attack rate in some cases; however, the final d.c. is the same as for a square-wave input giving the same a.c. levels for odd harmonics.

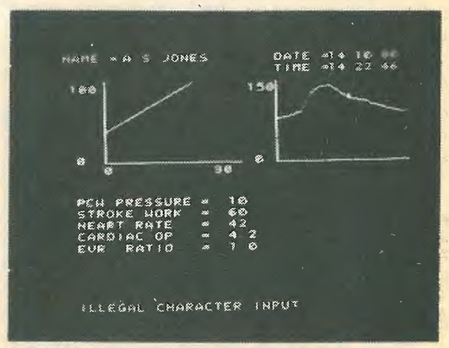
Further inputs, possibly useful in the lower octaves, can be added to Fig. 3, doubling the input resistor each time; adding 16*f* for example gives all the sawtooth harmonics in the right proportions except for 32,64,96, etc. Builders of my design who wish to add a staircase for 'Pykett' filtering should find it possible to accommodate the Fig. 3 components in the p.c. board spare positions. Preferably the staircase inputs should be taken from the third harmonic series, cross-keyed (*WW* March 1979), so unlocking the tones from the rest of the system; a separate speaker channel might be considered.

David Ryder
Bolton
Lancs

GRAPHICS AND MICROCOMPUTERS

I feel that Dr Witten's article "Graphical communication with micro-computers" in your August 1980 issue is worthy of some comment.

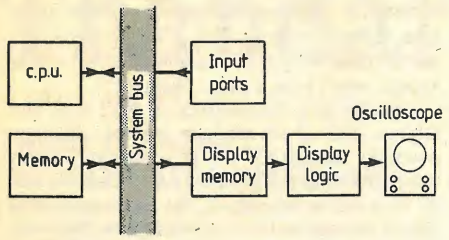
I have just completed work on a point plotting display which enables simultaneous display of graphics and text on an oscilloscope screen. A photograph of the display during development is shown here.



Dr Witten suggests that the use of a long-persistence phosphor reduces the necessity for a high screen refresh rate. In my experience this is not so since any event occurring regularly at less than the noticeable flicker rate (say 40Hz) produces a pronounced effect on the display. In the case of a long-persistence phosphor, the progress of the refresh is visible as a modulation of the brightness of the display unless the persistence is so long as to preclude any change in the information due to after-images.

My experiments involving display processors have indicated that, with a resolution of 256 × 256 points and a display of 4096 points, the

speed of both 8080 and 6800 processors is insufficient to perform just the display function without introducing flicker and therefore a hardware read of the display memory must be used as shown in the diagram.



The system uses two standard DAC0800 converters to provide X and Y inputs to the oscilloscope and, as can be seen from careful examination of the display, these are just adequate. It was found that continuous intensity modulation is not necessary. A simple 16-input OR function detects the address 0,0 and blanks the display.

There are no problems of bus contention with this system, though *prolonged* write operations by the processor do lead to a slight patterning on the screen. This can be eliminated by attention to software.

Leighton J. Man
Ilkley
West Yorkshire

DIGITAL ELECTRONICS TEACHING

In common with other academics, I am very willing to receive comments on syllabus content from industrialists. I therefore read with interest the letter from Ivor Catt in your November issue. Mr Catt believes that colleges and faculties refuse to teach the rudiments of digital electronic design, by which he does not mean the programming of microprocessors or other trivial surrogate activities.

I read on, hoping to learn what rudiments I should be teaching, but no such luck. It appears that at a seminar held at Hull University to discuss college electronics syllabuses Mr Catt issued a challenge to industry, but no rudiments.

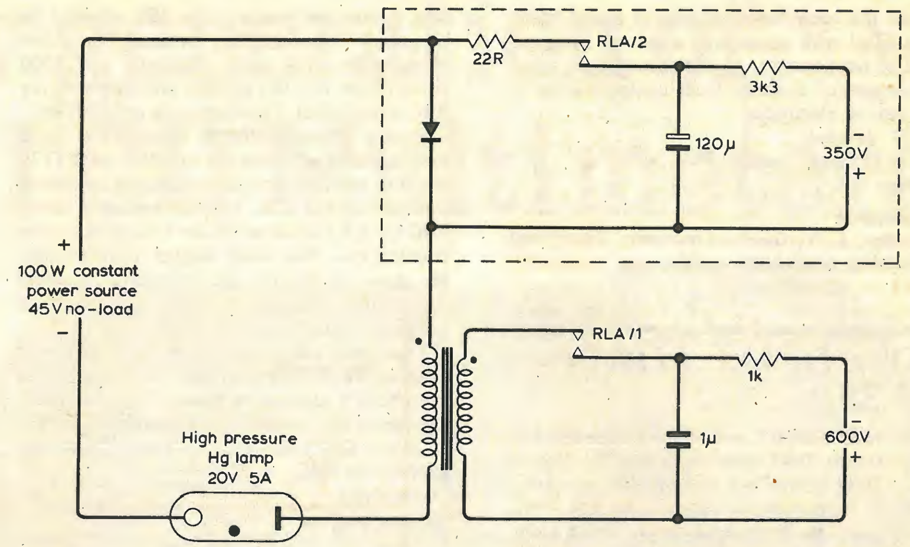
Please Mr Catt, reveal all about the rudiments of digital electronic design.

F. D. Cocks
Department of Engineering
Bristol Polytechnic

PICKABACK DISCHARGES FOR E.H.T

I was most interested to read Mr J. T. Lloyd's account (November letters) of his observations of the ability of a high-voltage discharge in air to allow the more or less complete dissipation of much higher energy stored at lower voltage in a capacitor in series with the discharge. I am reliably informed that an application of a discharge acting in this way may be found in the design of certain welding equipment in which the need to establish the arc by brief contact with the work is obviated by a supplementary source of r.f. power which triggers the main arc.

My own experience of the effect has been in the design of u.v. oscillographs in which it was required to strike a two-electrode high-pressure mercury lamp. A common method was to apply



Basic e.h.t. ignition circuit for mercury lamp as used in u.v. oscillographs, with supplementary h.t. source (in broken-line box). All resistors serve as current limiters. RLA is designed to operate repeatedly until ignition is obtained.

e.h.t. to the lamp by means of a capacitive discharge into a transformer having a secondary (necessarily of very low resistance) directly in the lamp current path, as shown in the lower right-hand of the diagram. It became evident that the energy of such a source was often insufficient to re-strike a hot lamp. Experiment showed that much greater energy stored at lower voltage in an electrolytic capacitor could be 'carried across' by placing it in series with the e.h.t. source, as shown in the upper right-hand of the diagram. The silicon diode is reverse-biased with respect to the h.t. source, and serves to conduct the lamp current after ignition. Thus the anode of the lamp receives a positive first half-cycle of e.h.t. (about 10kV peak) to which is added 350V from the electrolytic capacitor. The respective input energies of the h.t. and e.h.t. sources are in the ratio of about 40:1. A considerable improvement in lamp ignition was achieved by this method.

When this idea was first tested using a gap in place of the lamp a loud, vigorous spark was obtained. Thin paper placed in the gap would be readily charred or even ignited.

Like Mr Lloyd, I have considered the application of such a supplementary source to the improvement of automobile ignition. It must be said that under damp conditions such a system would be just as prone to degradation as a conventional one. Suppression could become a problem, as the use of highly resistive e.h.t. leads would nullify the advantages, and thorough screening would be required.

Finally, I would advise experimenters in this field of the risk of damage to fine-gauge secondary windings of e.h.t. transformers and ignition coils, in view of the high transient currents flowing in such a circuit. Shock will also be more severe than that experienced with car ignition.

My thanks are due to the directors of Bell & Howell Ltd, Basingstoke, for permission to describe details of the lamp ignition circuit of their model 5-137 oscillograph, for which I had some design responsibility while in their employment.

J. D. Southon
Farnborough
Hants.

MICROCHIPS AND MEGADEATHS

MY experience of *Wireless World* as a paying customer is not quite as long as that of Mr Scroggie (December 1980 letters), though I have read back numbers going back to 1915 or so, but I cannot remember any previous attempt at politicising its editorial and correspondence columns.

The trouble about persons with the wet-liberal outlook is that they readily respond to and propagate again the propaganda put out in the interests of the USSR. What a letter from the Campaign for Nuclear Disarmament is doing in your pages is far from clear. Has it not occurred to you that the wet-liberal outlook is no longer fashionable?

Philip Short
Gateshead
Tyne and Wear

I have just read your November 1980 editorial "Microchips and megadeaths" which must have the approval of all responsible readers. During the last war at least one man went to prison for pointing out that "Wars will cease when men refuse to fight". That was forty years ago. Now the picture is somewhat different: it is no longer the fighting man who will do the killing but the engineer who produces the hardware. Therefore if we are to avoid further Hiroshimas it is the technician and engineer who must now do the refusing.

Congratulations to *Wireless World* for stating what so many turn a blind eye to!

John Willmot
Bexhill-on-Sea
Sussex

SPARK GAPS

In his article "Spark Gaps" in the November issue Mr J. Dearden states that the pressure variation produced within a sealed spark gap by a change in ambient temperature will cause a corresponding change in breakdown voltage (he quotes a temperature variation of 40°C as causing a change of 2.3kV in breakdown voltage). However, the important parameter in determining the breakdown voltage of a gap of given length is the gas density [1], since this decides the number of collisons made by an electron in traversing the gap. The density is proportional to the pressure at a *fixed* temperature, but in a sealed gap it does not change with temperature.

Hence the breakdown voltages of sealed spark gaps filled with permanent gases are independent of temperature, in addition to their other advantages of freedom from contamination of the gas or electrodes.

R. G. Mitchell
The M-O Valve Company Ltd
London W6

Reference

1. Cobine, J. D., Gaseous Conductors, Theory and Engineering Applications, p.163.

LOSS OF M.F. POWER AT SEA

It was with interest I read the correspondence in your August 1980 issue regarding the loss of power from marine m.f. transmitters, operating to aerial under gale or storm conditions. The letter from Mr R.R. Venekamp, Eindhoven, who confirms by experiment, findings and a solution, that these losses are occurring, and that dielectric constant loss (*W.W.* May 1980) is a factor not to be ignored, was of particular interest.

One important aspect of the effects referred to, which should not be overlooked, is that the reports of loss of radiated power on m.f. which have been made are confined to instances when using a vessel's main or reserve transmitter to the ship's main or reserve aerial, the main transmitter having a power output up to 500 watts on m.f. m.c.w.

It therefore would seem likely that the losses referred to would have even greater effect when it becomes necessary to use the low power transportable transceiver, which all merchant vessels carry for use from the lifeboat. This unit operates on 500 kHz, 2182 kHz and 8364 kHz, with a power output of approximately 3 watts, using either a telescopic whip or wire antenna rigged to the lifeboat mast. These antennas would virtually be at sea level and in storm conditions, lifeboat, equipment and crew would literally be drenched in salt water. In such conditions, little or no r.f. energy may be radiated.

It is on record, though, that these small low power lifeboat transceivers have, in the past, saved lives. It would be interesting to know the sea and air conditions at the time they were successfully used.

A. K. Tunnah
Sydney, N.S.W.
Australia

COMMERCIAL AND PUBLIC SERVICE BROADCASTING

Congratulations on airing an important topic in your December editorial – but I notice you omit some of the argument. ITV is a federal structure – 16 independent organizations each with its own engineering and administrative staffs etc. The BBC is but one organization. So is it fair to compare costs directly? What's more, ITV companies make about the same number of programme hours as BBC1 and 2 combined – because of ITV's regional nature.

The commercial greed of ITV? All IPC journals (including *WW*) rely on advertising and not just counter sales. Is *WW* tainted by the need to sell soldering irons? Of course not.

Salaries are higher in ITV? Many say the reverse is true in some areas. You can't just pick on those examples that support your case.

Un-equitable comparisons, no more meaningful than yours, can be made in the reverse direction. As just one example, the BBC's licence fee is greater than NHK's who operate two tv channels, three radio channels, over 6000 transmitters, etc. But nobody says therefore the BBC is inefficient. Circumstances are different.

Having studied different countries, I think the standards achieved by the BBC (and ITV) are due to the particular plural broadcasting structure in the UK. Isn't it irrational to say BBC (or UK) broadcasters are biologically more talented etc. than their foreign counterparts? My guess is that it's the unique-in-the-world competitive situation. It keeps them *both* on their toes. Remember BBC tv before ITV?

I could add many other counter points to your analysis, but it isn't necessary. Of course it's in everybody's interest to have a buoyant BBC. But please don't argue for it by omitting half the evidence. Base your case on the quality and real needs of the BBC.

David Wood
Brussels, Belgium

AUDIO KITS

I feel that Mr Evans's letter on audio kits (November letters) cannot pass without some comment.

The kit-form hi-fi market contains many integrated stereo amplifiers with a wide range of available power outputs. I fail to see why Mr Evans knowingly chose one which was four times too powerful for his needs. Since, at least, materials costs increase with power output, it seems a needless waste of money.

I would also be interested to learn how he estimates the amount of time required to build a piece of equipment – whatever criteria he employs are obviously inadequate by a factor of two.

Regarding the opportunity costing of such ventures, perhaps he should have pondered the financial implications of kit building before he undertook the task. A major factor in constructing a kit is surely to obtain an item equivalent to commercial costing several times more – one's own labour charges are waived, as is true for any domestic d.i.y. project.

Mr Evans has apparently conducted a statistical survey of the kit-buying public (I was not included in his sample) since he knows that 60% of these "are not very fussy people". The other 40% obviously offered their beliefs regarding kit construction as part of the survey.

The whole tenor of the letter suggests to me that Mr Evans overreached himself with what appears to be a fairly demanding exercise. A little independent 'homework' regarding a kit product will produce a far better estimate of the work involved than the supplier's advertisements which are admittedly a little optimistic.

Anyone investing over £100 in a hi-fi kit should be convinced that they are capable of managing the complexity of assembly that such a price tag usually indicates, or they should patronise a kit supplier who offers an after-sale repair service. I have built many audio kits in the past few years and would agree with Mr Evans here: think about it very seriously before you buy.

A basic knowledge of electronics I consider is an essential pre-requisite (please forgive that sentiment appearing in this magazine) since suppliers instructions are, in the nature of things, seldom 100% comprehensive.

Few of us can design an integrated stereo amplifier but with painstaking care and patience many can assemble one from its components and obtain a working item first time.

M. G. Taylor
High Wycombe, Bucks.

DESIGNING WITH MICROPROCESSORS

Readers of Part 6 of "Designing with microprocessors" by Zissos and Valan in the December 1980 issue are advised not to take 8080 solution of PRINT problem too seriously, since apart from obvious syntax errors in the program listing of table 3 (p.73), the program, it would appear, won't print a single character.

Consider first the instruction MVI C at hex address 1003. If you load any non-zero character count in register C, the program will simply appear to jump to location L1 and Halt. Or will it? As a matter of interest, the instruction MVI affects no flags and even if you replace JNZ with JZ, it will be a waste of time trying it out. Furthermore, once you manage to patch-up this part of the code and somehow make the program counter reach L2, the printer will not stop printing garbage (unless man-handled) as the program will enter an endless loop by virtue of the JMP instruction at hex address 1013.

The authors are certainly not considering an abstract machine with a hypothetical instruction set but exemplifying the Intel 8080 with its well defined instruction set. How can well known authors produce such a piece of code? Perhaps they did not. One possible clue is the equivalent 6800 solution of table 1 and the authors' conclusion about the almost identical nature of solutions applying to all present-day microprocessors, and to all their modes of operation.

The architects of present-day micros certainly believe in non-compatibility with their competitors' products and a mere dictionary translation of mnemonics from one instruction set into another, or changing, for example, one accumulator of the 6800 to register C in the 8080, will not produce the same or even similar results, however trivial the program may be.

D. M. Vaidya
Westfield College
University of London
London NW3

COMMUNITY DATABASE

While considering a possible format for data transmission on an f.m. broadcast channel, it occurred to me that in order to control the stereo beacon, a receiver already contains a detector and pulse shaper. If the 19kHz signal were modulated at the transmitter then a simple opto-switch "Blu-Tack-ed" over the l.e.d. beacon could be directly connected to a serial input part of a microcomputer. This extremely cheap system would offer simple front panel connection and complete electrical isolation.

We thus have a method for programme marking, and to cater for the thousands of home computer owners, we could have a 'Computer Programme' with comment and programmes. We could also have community databases (cf. Community Memory Project in San Francisco) transmitting either on dedicated channels or perhaps 'piggy-back' on mono local radio stations. Items could be placed on the database via the telephone network. Access would be cheaper than for Prestel and the service would be more interactive than Ceefax or Oracle.

However, the maximum baud rate would depend on the reaction time of the beacon circuit and could be prohibitively slow. I cannot find any figures for this; perhaps somebody can help.

James Kidd
Warrington
Lancs

Linear modulator for radio control

Proportional 'digital' control with improved characteristic

by W. J. Hornsby, M.I.E.R.E

The story of this venture began when a friend of mine (who was building a model yacht) asked if I was aware of a suitable circuit for a model control system. I remembered having seen something a while ago, and a search in the archives of the local library turned up the item[1]. Not having given much attention to this subject before, the principles interested me, and, having most of the components to hand I decided to build the circuit. I found that it worked quite well, although it was rather large, occupying several circuit boards. I found also that although the control was proportional, it was not linear.

Having by now become more enthusiastic, I decided to explore the commercial field to see what was on offer and how the control was effected in proprietory equipment.

My findings were encouraging in some respects, I was able to buy control sticks with potentiometers in kit form and also the necessary basic items to produce small receivers. Servo mechanisms and the basic items for the servo decoders and amplifiers were also available[2]. I was disappointed, however, to find that the transmitter modulators still produced proportional but not linear control. I thought, therefore, that it would be an interesting exercise to try and develop an inexpensive, linear, proportional modulator, using readily available components, which was not expensive and occupied a relatively small space.

The use of integrated circuits sprang immediately to mind. I was advised to proceed with caution, however, since others had tried before and the problem had been to produce an economically priced unit consistent with the required performance. Whilst things could be done with t.t.l. i.cs, the current consumption was high and several devices were needed. On the other hand, the use of c.m.o.s. devices offered much lower power consumption, but there were snags here because charge storage in certain devices defeated the object of accurate control. Special timing devices such as the 555 were then considered, but these too proved unsuitable for direct control because, although they do have a modulation terminal, a very low impedance modulating voltage is required and this adds to the circuit complexity.

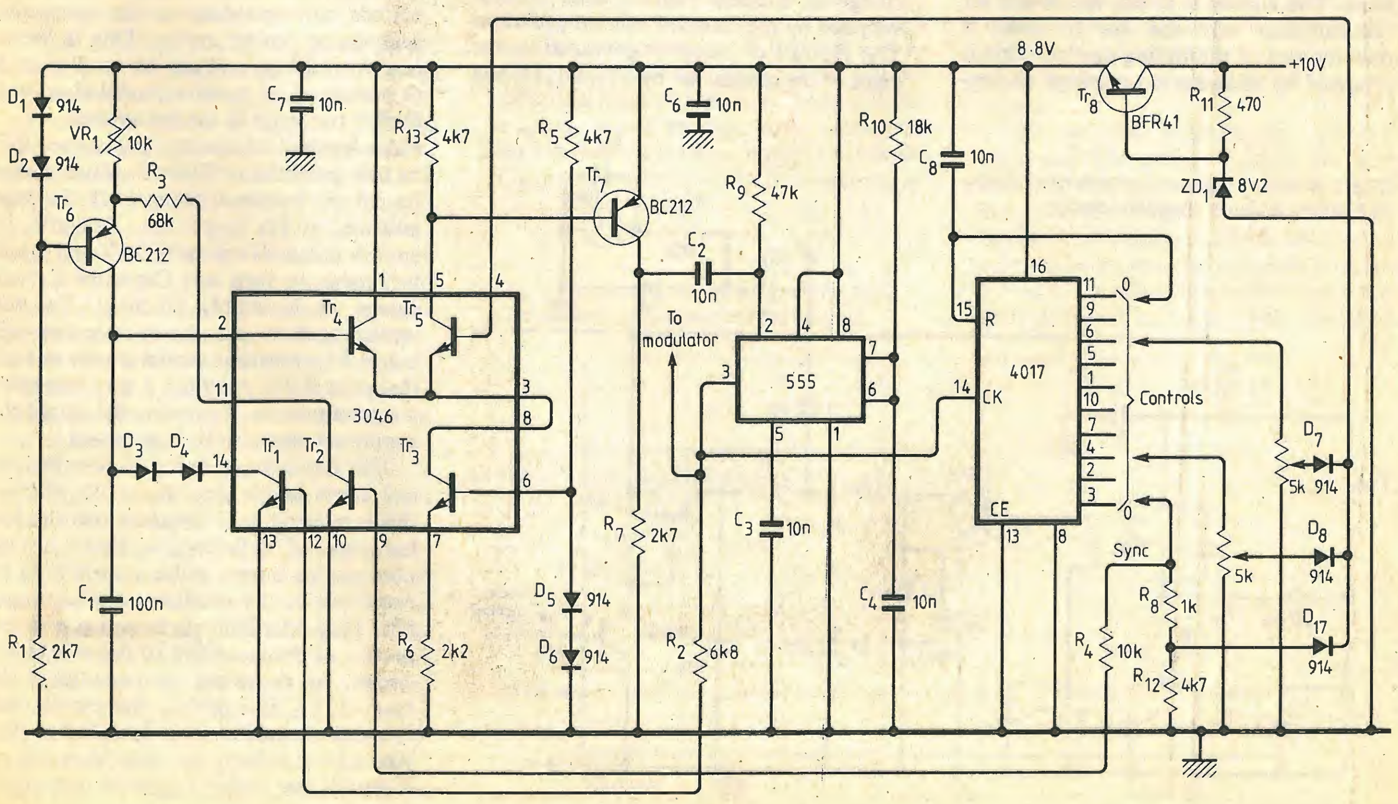
There had to be a compromise though, so the devices were permutated and several circuits tried until the arrangement shown evolved. It is a hybrid type, of small size and low power consumption, but it does produce extremely linear pulse-width modulation. The limiting factors are the quality of the timing capacitor, the linearity of the controls and the regulation of the supply voltage.

Circuit requirements

The modulator is required to produce a sequential train of pulses, each of which can be altered in width between 1 and 2ms, in a linear fashion, under the control of a manual actuator (control lever moving a linear potentiometer). A synchronizing pulse is needed to ensure that the correct pulse is fed to the appropriate servo at the receiving end, and each pulse should be separated from the next (and the sync. pulse) by a well-defined inter-pulse pause.

Types of circuit. Two methods are generally used. In one, the modulator consists of a number of individual timing circuits in series, each one triggering the following

Fig. 1. Complete circuit diagram of 9-channel modulator control unit. Capacitor C, should be polypropylene or, more economically, polycarbonate.



one after the end of its own cycle. The sync. pulse is derived from a multivibrator, which acts as the overall control section. The inter-pulse pause (i.p.p.) is produced by an individual element triggered by pulses produced when one timing element switches to the next. The i.p.p. therefore inhibits part of the next pulse and allowance must be made for this. This circuit in its usual form is not linear because it relies upon simple RC time constants. In addition, for a large number of channels it uses a considerable number of components and occupies a fair amount of space. It is, however, very economical on power.

The second common method uses only one timing element, and the various controls are offered to it in sequence (including one for the sync. pulse). This is known as the commutation method: it generally uses less components in larger systems, but a method of producing the i.p.p. must still be provided. Disadvantages of this method are that a very good timing circuit and comparator are required and that extreme care must be taken to ensure that there is no interaction between one control element and another. The modulator to be described shown in Fig. 1, uses the second method.

Devices

The ideal device for providing commutation is the 4017B. It was found, however, that if it was used directly to drive the timing element, the 'on' and 'off' times of one channel were influenced by what was happening on other channels. On closer examination it was discovered that only the timing and not the amplitude of the pulses was affected, possible due to charge storage. The 4017 does have much to recommend it, however, and to do the same thing with t.t.l. would require two devices. The answer is to use the device for commutation only and not to make it directly part of the timing circuit. This is arranged by allowing it to change its output state during the i.p.p. It now has time to settle down before it is required to provide the voltage for the next timing control and as already mentioned its voltage amplitude is reliable.

To generate the i.p.p. a 555 is ideal. It uses little current and few additional components to produce a very reliable output pulse. It can be used in the monostable (triggered) mode and its output can source or sink considerable current. This makes it ideal for the overall controlling element and it will provide the output modulating voltage.

The remainder of the circuit consists of an element which charges a capacitor linearly from a constant current source, and a voltage comparator. Discrete components appeared to be the only reasonable answer here but advantage was taken of the CA3046 in view of its compact form. Two discrete p-n-p transistors and an n-p-n voltage regulator are the only other active components.

Circuit operation

Assume for the moment that the circuit is already running and that we are looking at the start of one timing pulse. One of the control outputs of the 4017 will be 'high', thereby connecting the positive supply to the appropriate control potentiometer. (The other outputs of the 4017 will be 'low'). The slider of the selected control will connect the control voltage, via its isolating diode, to one side of the comparator Tr_5. The comparator is comprised of Tr_5, Tr_4 and a constant current source Tr_3, which is used to provide fast switching between Tr_5 and Tr_4.

At this point, Tr_5 will turn on and, in turn, will cause Tr_7 to conduct, raising the top of R_7 to supply potential. (This is ineffective at present but will be made use of later.) At the same time, C_1 begins to charge in a linear fashion from current supplied by the constant current generator Tr_6; the rate of charge is governed by the value of the current set by Vr_1, D_1, D_2 and R_1. The voltage across C_1 rises and this is applied to the other side of the comparator Tr_4.

When the voltage on Tr_4 base just exceeds that on Tr_5 base, Tr_4 turns on, causing its emitter potential to rise and pull all the current from Tr_3. This, in turn, causes Tr_5 to cut off. With Tr_5 off, Tr_7 ceases to conduct and the potential at the top of R_7 falls sharply towards ground, producing a sharp negative trigger pulse via C_2 to pin 2 of the 555. The 555 now begins its timing cycle, the length of which is governed by R_{10} and C_4. As soon as the 555 is started, pin 3 goes 'high'. This positive step is passed to the 4017 to advance its output to the next control potentiometer and the same positive rise is passed to Tr_1, which conducts and causes C_1 to discharge. (The function of D_3 and D_4 will be discussed later.) By the time the 555 has completed its timing cycle (the i.p.p.) the 4017 has settled down and presented the comparator with its next control voltage and C_1 is ready for its next charge.

All the circuit readjustments therefore are completed during the i.p.p. Furthermore, once the 555 has been triggered, it will ignore any other changes on its trigger pin until its timing cycle is complete. The circuit changes during the i.p.p. therefore will not affect the i.p.p. In addition, since the modulator output is taken from the 555, neither the timing nor i.p.p. pulses will be affected by the output circuit.

At the end of the i.p.p., pin 3 of the 555 goes 'low' and Tr_1 ceases to conduct, removing the discharge path from C_1. Tr_4 has already ceased conduction (when C_1 discharged) and Tr_5 and Tr_7 have turned on to prepare the next trigger pulse. The cycle now repeats as before, but with a new control potential applied to Tr_5.

The modulating output consists therefore of i.p.ps, separated by off periods corresponding to the successive comparator timing cycles. This is 'negative' modulation and can be used as such or converted to positive modulation by a further transistor in the transmitter.

Pulse limits. Considering Tr_3, we see that its base potential is 1200mV, which makes its emitter potential 600mV. If the base potential of Tr_5 is less than 1200mV, its emitter potential will be 600mV and it will be unable to turn on. Capacitor C_1 will charge up to supply potential, Tr_4 will remain hard on and circuit operation will cease. Tr_4 emitter potential is now so high that even if Tr_5 potential is now increased it cannot turn on. To restore the circuit the supply voltage must be interrupted.

This poses the question of how the circuit starts in the first place and whether this is reliable. It is fortunate that the 555 has a sort of 'deficiency' in that it always produces an output pulse at switch on irrespective of the condition of the trigger pin. This, in effect, gives an i.p.p. irrespective of the condition of the rest of the circuit. So, providing the potential of the base of Tr_5 is suitable, the circuit will function. Capacitor C_8 is included to provide a reset pulse to the 4017 at switch on to ensure that timing begins on output '0'.

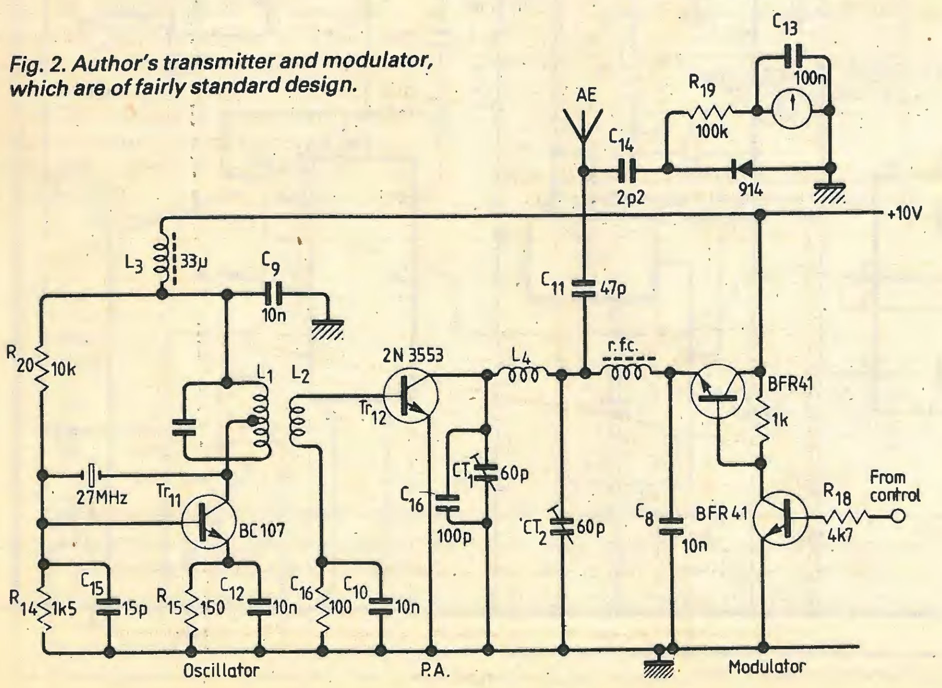


Fig. 2. Author's transmitter and modulator, which are of fairly standard design.

Without this, the 4017 occasionally tries to begin on another output and if this is not committed (i.e. no control connected to it) the circuit will not function. The reset pin of the 4017 is strapped to the output following the last pulse. If the sync. pulse is connected to output '0' then the channel numbers will correspond with the other output numbers. Note however that the 4017 pin numbers and output numbers do not correspond.

If the potential of Tr_5 base is raised to near the supply level, its emitter potential is so high that Tr_4 cannot turn on because C_1 cannot charge above the supply voltage less the saturation voltage on Tr_6 (about 200mV). If in this condition, however, the potential of Tr_5 base is reduced, Tr_5 will turn off, causing Tr_7 to produce a trigger pulse, and the circuit will continue to function. The maximum pulse length therefore occurs when Tr_5 base is near the supply potential and the minimum pulse occurs when Tr_5 base is about 1200mV.

Control adjustment

The charge graph of C_1 is shown on the graph as line A to D, the slope of which is determined by the value of charging current set by Vr_1.

The voltage at which the comparator trips is set by the control potentiometer and, in commercial control units, this is limited to only about one fifth of the potentiometer range. Plotting the control range on the same graph shows that there is only one angle of slope of the charging potential that will produce the 1 to 2ms range required. It was found that, in practice, using standard component values, this put Tr_5 base potential very close to the cut-off point mentioned earlier. Also, since C_1 began its charge from very close to ground, and the minimum trip voltage of Tr_5 is 1200mV, it was not possible to reduce the minimum pulse length much below 1ms.

This problem could be solved if the slope of the charge graph could be preserved but the point from which it began could be raised. In any case, C_1 can only discharge down to $V_{ce(sat)}$ of Tr_1 (about 200mV) so, if diodes D_3 and D_4 are added, the minimum point of discharge is now raised to 1400mV (600+200+200). This has two advantages: first it lifts the low control potential clear of the cut-off potential of Tr_5 (and incidentally puts the control potentiometers about mid travel, which should be the best part of the track) and secondly it allows Tr_5 base potential to approach very closely the minimum potential of C_1, so that very short pulses can be produced. As before though, if Tr_5 base potential drops below the minimum voltage of C_1, the circuit will cease to function as was the case when it dropped below 1200mV without D_3 and D_4 fitted. This will not occur, however, in normal operation because of the limited travel of the control sticks. If full rotation of a control is needed, e.g., to simulate the steering wheel of a car, the 5k control pot. can be replaced by a 1k pot. with a fixed 1k8 resistor at each end. The addition of D_3 and D_4, although slightly restricting the full overall range, has put the required range well clear of any difficulties.

Sync. pulse

Some systems require a short sync. pulse of the order of 600µs and this is simply arranged by putting a potential divider across one of the outputs of the 4017, thereby treating it as a normal pulse but of shorter duration (the addition of D_3 and D_4 assists this facility). Most modern systems, however, need a long sync. pulse of the order of 6 to 8ms, which is well outside the normal control range. An alternative means must therefore be employed.

The obvious answer in this case is to alter the slope of the charging potential of C_1 for this pulse alone. This is accomplished by adding Tr_2, which, when it receives its 'high' output from the 4017, turns on and adds R_3 to the constant-current circuit. R_3 brings the emitter potential of Tr_6 closer to its base potential and thus reduces the charging current of C_1, thereby extending the length of the sync. pulse. Adjust R_3 to give the length of sync. pulse required: R_8 and R_{12} provide the trip voltage for the pulse end.

Setting up

The unit was designed to work from a 10 volt NiCad battery and, to ensure stability over the duration of the discharge, the supply was regulated down to about 8 volts by Tr_8.

Again, the voltage is not too critical, but it is essential that it is stable because it is applied directly to the timing elements. NiCad cells have a fairly steady voltage during discharge, but if ordinary dry cells are used than a more elaborate regulating circuit ought to be used.

The best tool for setting up the circuit is an oscilloscope. It does not need to be too elaborate, but it should have a trigger facility because the waveform is not symmetrical. To make things easy, the best way to set up initially is to connect three control sticks and leave out the sync. pulse for the moment. It will be necessary to strap the 4017 to reset on the fourth step (i.e. output 3, since it counts 0-1-2-3). The position of each control potentiometer with respect to its stick, and the setting of Vr_1, are then adjusted alternately to give the required duration and variation. It is surprising how much variation in linearity there is between potentiometers; however, the expense of special units is not justified for the degree of improvement that would be obtained. During this process it will be necessary to trigger the 'scope from the top of each potentiometer in turn.

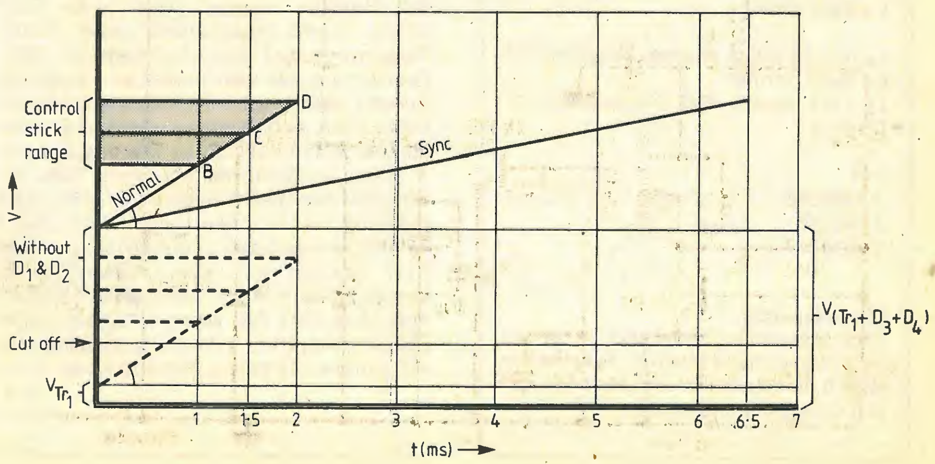
Check, at this stage, that the variation of one channel does not affect the others.

If C_1 is subject to excessive retentivity this will show up as the variation of one channel by the preceding one but not the succeeding one (this is because the voltage across C_1 is not stable when Tr_1 removes its discharge path. If all is operating correctly, there should be no noticeable interaction between channels even when one of these has been expanded to fill the 'scope screen.

The next job is to connect the sync. pulse into the circuit. Trigger the 'scope from the 4017 sync. pulse output and adjust the pulse length as appropriate. If a 'scope is not readily available then the circuit can be set up approximately by setting Vr_1 to 4.75k and all the control potentiometers to about half way. This is a useful

> Two of the circuits described have been in use for over 12 months by my sons and no failure has occurred. Although in recent times some dedicated i.cs have appeared, which offer a similar facility, these have usually only 6 or 7 channels, are available only from selected sources and are fairly expensive. All the components described in this circuit are readily available and cheap. It would be possible, by modification of the 4017 end of the circuit, to control more than 9 channels; however, the time between successive pulses on any one channel becomes longer as the number of channels increases. This makes the servos slower to respond and causes jitter unless the associated decoder and driver circuits are carefully designed.

Fig. 3. Addition of diodes D_3 and D_4 sets comparator range clear of difficulties.



method when the modulator is incorporated into existing equipment; the controls can then be adjusted to match the existing servo mechanism.

A simple transmitter following standard practice is shown in the diagram. The complete assembly of modulator and transmitter can be built on a printed circuit board 10.125 by 6.5cm.

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2. Micron Radio Control, 31 Hayworth Road, Sandiacre, Nottingham.
3. Ambit International, 200 North Service Road, Brentwood, Essex.

Components

Resistors

1	2k7	11	470
2	6k8	12	4k7
3	68k	13	10k
4	10k	14	1k5
5	4k7	15	150
6	2k2	16	100
7	2k7	17	1k
8	1k	18	4k7
9	47k	19	100k
10	18k	Vr_1	10k preset

Capacitors
1 100n polycarbonate
2 10n
3 10n
4 10n
5 33μ
6 10n
7 10n
8 10n
9 10n
10 10n
11 47p
12 10n
13 100n
14 2p2
15 15p
16 100p
T1 60p trimmer
T2 60p trimmer

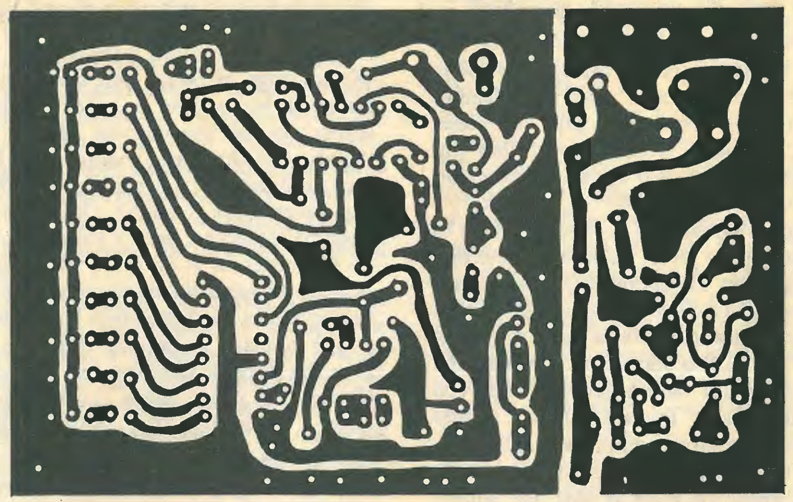
Inductors
L_1, L_2 113CNF2K209A (TOKO) with screening can } Ambit
L_3 33μH choke

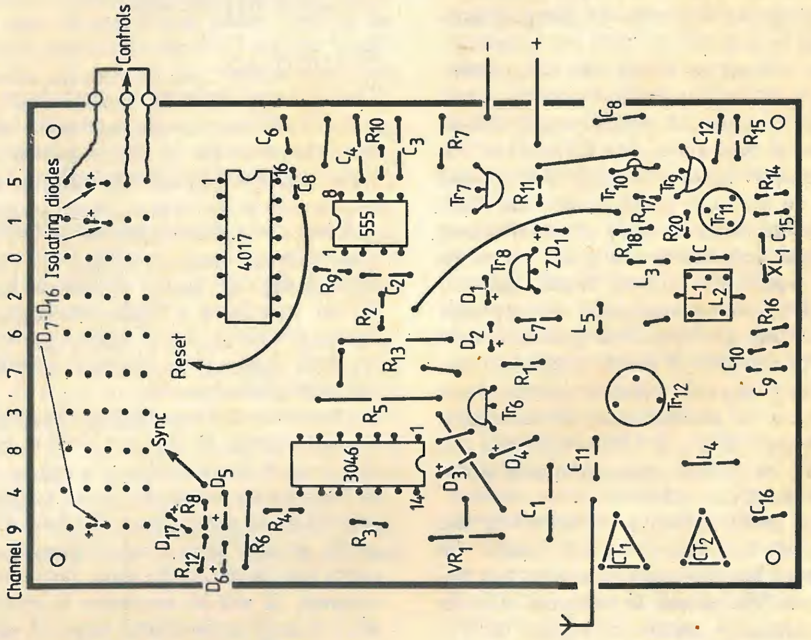
L_4 12 t. 16 s.w.g. enamel, close-wound on ⅜dia. former
L_5 V.h.f. choke. (R.S. Components or Doram)

I.cs
1 CA3046
2 NE555 } Ambit
3 CD4017

Miscellaneous
Stick assemblies, aerial and socket can be obtained from Micron. (Assemblies with 0.25in spindles are easier to adjust than miniature types).

Board layout for modulator and transmitter. Dimensions are 10.125×6.5 cm.





The author

Mr Hornsby started work with Post Office Telecommunications (now British Telecom) at the age of sixteen, in 1952. Following a two-year period as a National Service radio mechanic, he returned to the Post Office, and eventually took a post as lecturer at the Post Office Training College at Stone, in Staffordshire. Since 1968, Mr Hornsby has been involved in exchange planning in Sheffield and Nottingham (where he was head of the division) and is now working on network planning in Leeds. He is married, with three children and says that his spare-time activity is "pure" electronics, where, he thinks, to travel is more interesting than to arrive.

Current dumping analysis

Class B amplifiers without crossover distortion

by **H. S. Malvar,** M.Sc. University of Brasilia

The current dumping technique, as presented by P. J. Walker[1], is a very elegant solution to the problem of reducing the crossover distortion in class B audio power amplifiers, because it eliminates the requirement of a quiescent current on the output transistors, and the thermal problems associated with biasing. However, the amount of controversy[2-8] that followed Mr Walker's article denotes that the current dumping principle has not received a complete treatment. The purpose of this work is to show, by means of a more complete analysis, that current dumping does reduce the crossover distortion more than conventional feedback but it is not able to totally cancel this distortion on the output stage, even with a "theoretically perfect balance" or infinite feedback factor.

In Fig. 1, the general arrangement of the current dumping technique, the class A amplifier has a finite gain A, and it is shown later that the balance condition does not require A to be infinite. This configuration is general in the sense that all the current dumping circuits are possible realizations of it. The flow-graph of this configuration is presented in Fig. 2, and helps to understand how both feedback and feed-forward are employed.

The two basic equations for the amplifier in Fig. 1 are

$$V_1=A(V_s-kV_2), \quad 1$$

$$\frac{V_o}{R_L}=\frac{V_1-V_o}{R_3}+\frac{V_2-V_o}{R_4}\Rightarrow$$

$$\frac{V_o}{R_p}=\frac{V_1}{R_3}+\frac{V_2}{R_4} \quad 2$$

where $R_p=R_L//R_3//R_4$. These equations cannot be solved for V_o/V_s unless a third equation is introduced, if no particular values for R_3 and R_4 are assumed. The action of the dumper gives this equation. Its transference from V_1 to V_2 can be written as

$$V_2=BV_1$$

where B can be a highly non-linear factor, in which are present crossover effects. With this relation equations 1 and 2 become

$$V_1=A(V_s-kBV_1)\Rightarrow V_1=\frac{A}{1+kAB}\cdot V_s \quad 3$$

$$\frac{V_o}{R_p}=\frac{V_1}{R_3}+\frac{BV_1}{R_4}\Rightarrow V_o=\frac{R_p}{R_3}\left[1+\frac{R_3}{R_4}\cdot B\right]V_1 \quad 4$$

which, combined, finally lead to

$$\frac{V_o}{V_s}=\frac{R_p}{R_3}\cdot\frac{A}{1+kAB}\cdot\left[1+\frac{R_3}{R_4}\cdot B\right] \quad 5$$

which is the desired relation between input and output. The heart of current dumping is to make the denominator of the second factor equal to the third factor, which is attained if

$$\frac{R_3}{R_4}=kA \quad 6$$

Making this substitution in equation 5 the term $1+kAB$ is cancelled, and

$$\frac{V_o}{V_s}=\frac{R_p}{R_3}\cdot A \quad 7$$

This result is the reason for all the excitement that has involved the people that worked on current dumping, because it states that the output signal does not depend on the dumper transfer characteristic! This is highly impressive, because a look to Fig. 1 reveals that if $R_3 \ll R_4$ the dumper will be the main source of power to the load when it is on (i.e. at medium and high signal levels). But as the dumper gain B is not present in equation 7, its crossover distortion will not appear at the output voltage. (Remember that A is the gain of a class A amplifier, and hence free from crossover effects.)

The results of equations 6 and 7 were already known[1,3,5,6,7] in different forms but with the same meaning; that was well defined in an assertion by Mr Walker[4]: ". . . there is a theoretically accessible state where the output stage distortion will cancel to zero, without calling upon infinite loop gain . . ." Is this really true? Can one get the power of an amplifier (in this case, the dumper) without getting its distortion, too?

In fact, the situation is not so good as it may appear at first sight. A very important point was missed out of the analysis so far, and it was also missed from previous analyses of current dumping[1,3,5,6,7]: the distortion of the class A amplifier. This low-power amplifier must have a very low distortion level, because its distortion will appear at the output, which is clear from equation 7. Because it operates in class A, a very low distortion is not so difficult to achieve, and this problem was left out. However, if the gain A is distorted, even

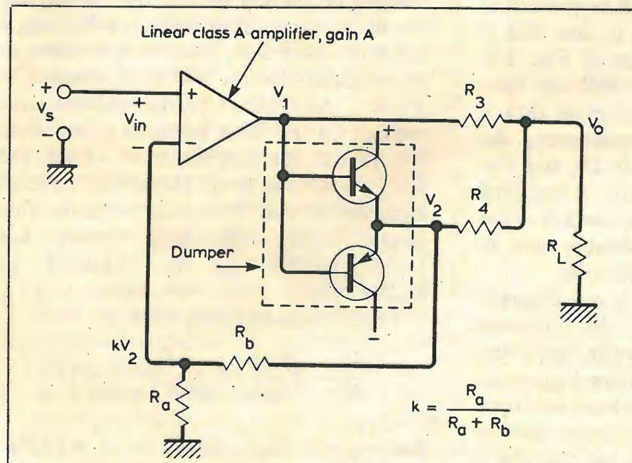
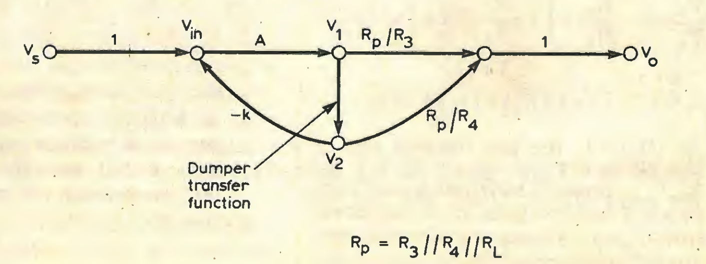


Fig. 1. General arrangement of the current dumping technique. For small signals (V_1<200mV r.m.s.) the output current is supplied by the low-power class A amplifier.

Fig. 2. As the principal power-carrying transmission is that from V_2 to V_o, the transmission from V_1 to V_o is feedforward. Hence it is clear that current dumping uses both feedback and feedforward.



by a small amount, the balance condition of equation 6 will not hold for all signal levels, and the term $1+kAB$ in equation 5 will not be perfectly cancelled.

To clarify the situation, the distortion of both amplifiers must be considered in the analysis. In fact, if one is looking for a distortion reducing scheme, all the distortion sources must be taken into account. A simple way to introduce these distortions is to write the gain of the low-power amplifier as $A(1+D_A)$, and that of the dumper as $B(1+D_B)$, where A and B are fixed constants and D_A and D_B are random variables with unknown distributions that represent the distortion of the class A amplifier and the crossover distortion of the dumper, respectively.

Hence, equation 5 becomes

$$\frac{V_o}{V_s}=\frac{R_p}{R_3}\cdot\frac{A(1+D_A)}{1+kAB(1+D_A)(1+D_B)}\cdot\left[1+\frac{R_3}{R_4}B(1+D_B)\right] \qquad 8$$

From this it is clear that there are no finite values for A and R_3/R_4 that can cancel the effects of D_B on V_o/V_s (remember that $R_3/R_4=kA(1+D_A)$ cannot be written, because a random variable cannot be always equal to a constant). An interesting result is obtained if the loop gain is made infinite, i.e. if $A\rightarrow\infty$. It is easy to see, from equation 8, that

$$\lim_{A\rightarrow\infty}\frac{V_o}{V_s}=\frac{R_p}{R_3}\cdot\frac{1}{kB(1+D_B)}\cdot\left[1+\frac{R_3}{R_4}B(1+D_B)\right]$$

$$=\frac{R_p}{kR_4}\left[1+\frac{R_4}{R_3}\frac{1}{B(1+D_B)}\right] \qquad 9$$

This shows that even with an infinite loop gain, crossover distortion will be present at the output. It is interesting to note that if the feedforward transmission in Fig. 2 is nulled, i.e. if R_3 is made infinite, thus converting the current dumping into a conventional feedback arrangement, the output will be free from both D_A and D_B. This is the classic result from conventional feedback: infinite loop gain (which is far from being physically realizable due to instability) means zero distortion.

Thus, current dumping is not theoretically able to cancel totally the crossover distortion of a class B power amplifier, either with a finite or infinite loop gain, but it is better than conventional feedback for a finite loop gain and assuming that the balance condition holds. To see that this is true, it is necessary to express the effects of D_A and D_B on V_o/V_s more clearly than in equation 8, which can be written as

$$\frac{V_o}{V_s}=A\cdot\frac{R_p}{R_3}\cdot\left[1+\frac{R_3}{R_4}B(1+D_B)\right]\cdot\frac{1+D_A}{1+kAB(1+D_A)(1+D_B)}$$

As $|D_A|\ll1$, the last fraction can be expanded in a Taylor series for D_A, around the point $D_A=0$. Using the fact that

$$\frac{\partial^n}{\partial D_A^n}\left[\frac{1+D_A}{1+kAB(1+D_A)(1+D_B)}\right]=\frac{(-1)^{n+1}\,n!\,[kAB(1+D_B)]^{n+1}}{[1+kAB(1+D_A)(1+D_B)]^{n+1}}$$

the final result is

$$\frac{V_o}{V_s}=A\frac{R_p}{R_3}\left[\frac{1+\frac{R_3}{R_4}B(1+D_B)}{1+kAB(1+D_B)}\right].(1+hD_A-h^2nD_A^2+h^3n^2D_A^3-\ldots) \qquad 10$$

$$\text{where } h=\frac{1}{1+kAB(1+D_B)}$$

$$\text{and } n=\frac{kAB(1+D_B)}{1+kAB(1+D_B)}.$$

If the balance condition 6 is satisfied, a simplification can be made in equation 10 (and this is the reason for the minimum in the output distortion already verified experimentally[5,8])

$$\frac{V_o}{V_s}=A\frac{R_p}{R_3}(1+hD_A-h^2nD_A^2+h^3n^2D_A^3-\ldots) \qquad 11$$

which shows that current dumping generates high-order distortion, as does conventional feedback.

At this point, it is useful to separate the analysis in two cases, corresponding to the off and on conditions of the dumper.

Dumper off. This condition corresponds to $B(1+D_B)=0$, which implies $h=1$ and $n=0$. Thus equation 11 becomes

$$\frac{V_o}{V_s}=A\frac{R_p}{R_3}(1+D_A). \qquad 12$$

So, when the output power transistors are off the transmission from V_1 to V_2 is nulled, which breaks the feedback loop. Therefore, as the output signal is supplied by the class A amplifier only, with no feedback, the distortion factor of V_o/V_s must be D_A, as stated in equation 12.

Dumper on. When the dumper is on, i.e. one of the output transistors conducting, it has little distortion, because it is acting as an emitter-follower, which is implied in $|D_B|<1$. As $kAB\gg1$ (which follows from the fact that R_3 must be much greater than R_4, the balance condition is $R_3=kAR_4$ and $B\approx1$), $h\ll1$ and $n\approx1$. Hence the series of equation 11 can be truncated to the first power term with little error, and $1+kAB(1+D_B)$ can be replaced by $kAB(1+D_B)$.

These considerations lead to

$$\frac{V_o}{V_s}\approx A\frac{R_p}{R_3}\left[1+\frac{D_A}{kAB(1+D_B)}\right].$$

$$\text{As } \frac{1}{1+D_B}=1-D_B+D_B^2-D_B^3+\ldots\approx1-D_B$$

for $|D_B|\ll1$, it follows that

$$\frac{V_o}{V_s}=A\frac{R_p}{R_3}\left[1+\frac{D_A}{kAB}-\frac{D_AD_B}{kAB}\right]. \qquad 13$$

So the output has two main distortion components: one due to the distortion of the class A amplifier, which is D_A reduced by kAB (this was expected because D_A is generated within a feedback loop with loop gain kAB), and the other due to the intermodulation of the two distortions, that is

$$D_A\frac{D_B}{kAB}$$

With $|D_A|\ll1$ the effect of the distortion D_B is reduced by an amount greater than the feedback loop gain kAB. Therefore, the current dumping technique can reduce the effects of the crossover distortion more than conventional feedback, given the same loop gain, even though this reduction cannot be total anyway.

Hence the current dumping allows the design of a power amplifier with output transistors in true class B, avoiding the well-known thermal problems in conventional AB output stages. Further, it is also correct to say that the performance of a current dumping power amplifier is dictated mainly by two factors: the linearity of the low power class A amplifier (see equation 13) and the precision of the balance.

Another important point is the effect of the output impedance of the amplifier A, and the input and output impedances of the dumper, the last two being highly dependent on whether the dumper is on or off. The variation in the output impedance of the dumper can be accommodated by the distortion factor D_B, and then does not affect the results. But its input impedance will affect the term D_A, as the amplifier A cannot have zero output impedance. Therefore the transistors of the dumper must have very high current gains, to minimize the effects of the loading of the linear class A amplifier by a non-linear load. In practice, these transistors must be Darlington pairs or triplets.

From a practical viewpoint, the balance condition is not too difficult to achieve, if the class A amplifier has its gain stabilized by local feedback[1,3,7]. Even if this condition is not satisfied, a balance can be obtained by adjusting k^5 or the resistors R_3 or R_4. As the gain A must have some kind of compensation in frequency due to the feedback loop around it, it seems that the equation 6 will not hold for all frequencies. However, if the compensation is made by a single pole in A, the resistor R_4 can be replaced by a series connection of an inductance and a resistor, in order to create a zero that can cancel the pole[1,8].

Finally, it seems that the current dumping technique is "the state of the art" for reducing to very low levels the crossover distortion (see, for instance, the results of the evaluation of a power amplifier that uses current dumping[8]), thus allowing the construction of high-fidelity amplifiers in true class B.

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Unified circuit theory

Millman's theorem does not seem to be widely known, yet most of the usual circuit theorems stem from it

by E. H. Pollard*

Millman's theorem deserves to be better known than it is; not only do all the usual circuit theorems stem from it as corollaries but it is extremely useful in its own right, often saving a fair amount of work. It certainly merits the detailed formal proof given.

Is disaster but a flick of the switch away in the bistable circuit of Fig. 1? True, the values beside the resistor symbols are rather unusual in an otherwise conventional circuit. But rest assured the circuit is practicable, and the transistors safe. Appending conductance values to resistor symbols might be considered to be an unfortunate peccadillo, but is not conductance as much an intrinsic property of a resistor as resistance, just as resistance is an intrinsic property of a conductor? Certainly it is not at all uncommon to see resistance values appended to conductors.

Whatever the circuit may lack initially in clarity due to an unexplained ambiguity, there is no shortcoming in logic. Undoubtedly it would have been helpful if the units were mentioned. In this case it is the recently introduced siemens, abbreviated to S.

As well as being clarified by the appending of their units, the conductances could have been more readily recognized if they had been allotted their own symbol. As this article is largely concerned with conductance and admittance in their relationship to electric theory, it is helpful if they are adequately differentiated from resistance and impedance, their respective reciprocals, or duals[1], by this means. Many years ago *Wireless World* printed a diagrammatic shorthand for valve circuits, which is not without relevance today. As shown in Fig. 2, the shorthand symbol given then for a resistor will be used to denote an admittor or admittance. The letters R, G, Z, and Y will be used as usual to signify resistance, conductance, impedance and admittance.

It is common practice to keep conductance and admittance in reserve, so to speak, and swing them into action when one is faced with an awkward array of parallelism. But there is no earthly reason why one should not keep conductance in the forefront for serially arranged conductors, just as one often sticks exclusively to resistance calculations for resistors connected in parallel. For example, Fig. 3(a) shows one cell of a battery, taken to be a pure source of e.m.f. in series with its internal resistance, whilst Fig. 3(b) shows the same cell in series with its internal conductance, the terminals A and B defining the boundaries of the cell. The batteries are, of course, equivalent; and the same load connected across the terminals A and B of either figure, whether it is reckoned in ohms or siemens, will give exactly the same current out of the battery. Similar conditions apply, and the outcome is the same, in the circuit of Fig. 3(c) where the battery has been replaced by the voltage generator – with its internal impedance – in series. Then, as $z=1/y$ where y is the internal admittance of the voltage generator, Fig. 3(d) with its internal admittance in series is equivalent to Fig. 3(c).

It can be shown by experiment, but not so easily as might be imagined, that if two different cells say are connected in series, as in Fig. 4(a), the resulting battery can be depicted simply as in Fig. 4(b), with the battery e.m.f.s summed, and the separate internal conductances replaced by a single conductance, given by the reciprocal of the sum of their individual reciprocals, a simple enough operation on a pocket calculator with a reciprocal key.

What happens if the cells are connected in parallel instead of series? Well, again, careful experiment will give the answer, in certain conditions. The separate internal conductances can always be summed directly to give the total conductance; and that as long as the voltages are the same or the conductances the same, or both the voltages are the same and conductances the same, the total e.m.f. will always be the mean e.m.f. of the individual cells. This still leaves one other possibility, where both the cell e.m.f.s and internal conductances are unequal. In this case the answer is less straightforward; one would have to be a Prince of Serendip to plump for the correct solution as a result of experiment: there is no simple equivalent. The total e.m.f. is $(V_1g_1+V_2g_2)/(g_1+g_2)$, which will always lie between the voltages of the individual cells. The different cases are shown in Fig. 5.

It can be easily deduced, however, that the expression for the total voltage of the two cells shown in Fig. 5(d) is correct. Remember that the internal conductances of the cells were set in series with the e.m.fs which were defined to be perfect voltage sources; that is, the sources have infinite conductance.

To make the matter more general Fig. 5(d) has been redrawn as Fig. 6(a), where the internal conductances g_1 and g_2 are included respectively in the series conductances G_1 and G_2. If V_2 is reduced to zero, the circuit of Fig. 6(b) results, which is a voltage divider having across its terminals a voltage $V_{AB1}=V_1G_1/(G_1+G_2)$. Similarly, if the voltage source V_1 is reduced to zero to give the circuit of Fig. 6(c), the voltage across the terminals is now $V_{AB2}=V_2G_2/(G_1+G_2)$. Adding these two voltages, which act in the same direction across terminals AB, gives the terminal voltage as $V_{AB}=(V_1G_1+V_2G_2)/(G_1+G_2)$.

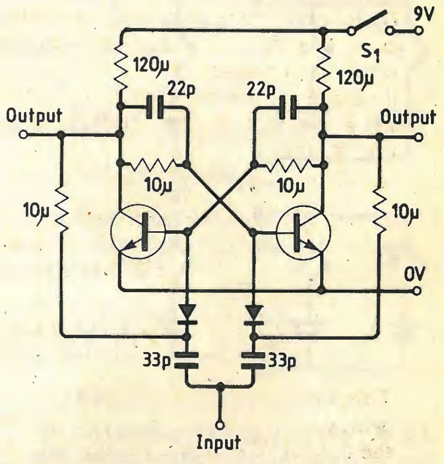


Fig. 1. Practicable bistable or a disaster?

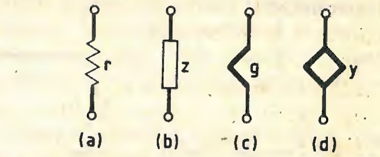


Fig. 2. Symbols used for resistor or resistance (a), impedor or impedance (b), conductor or conductance (c), admittor or admittance (d).

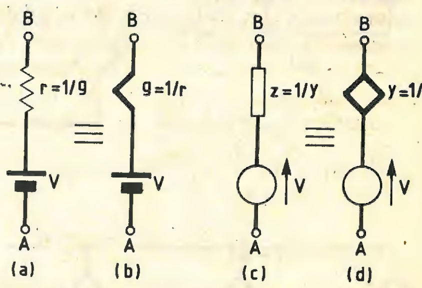


Fig. 3. Batteries of e.m.f. V in series with its internal resistance (a), internal conductance (b), and generators of e.m.f. V with internal impedance (c), and internal admittance (d). (Batteries and generators are equivalent).

* Submitted postumously by K. J. Pollard.

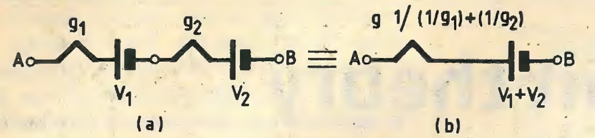


Fig. 4. Two cells represented by pure voltage sources in series, each with internal admittance (a). When the cells are connected in series, (b) + (a). Cell voltages add directly; the total admittance of the battery is the reciprocal of the sum of reciprocals of the separate internal admittances.

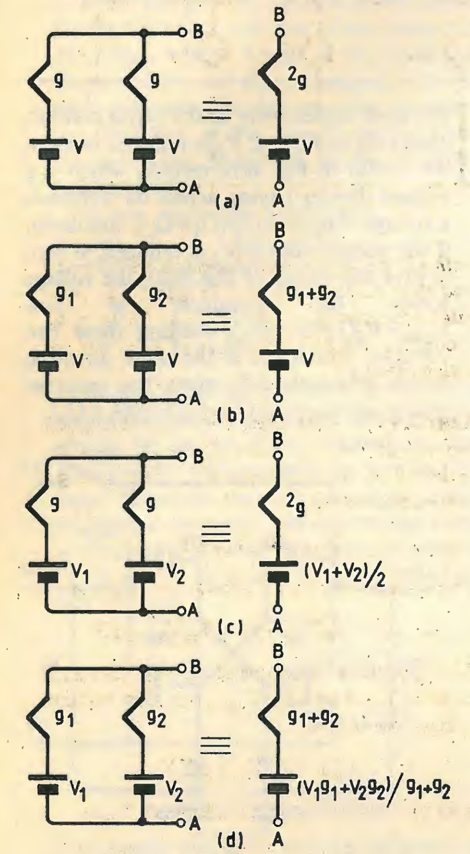


Fig. 5. Cells connected in parallel (a), (b) and (c) show that the voltage across the terminals is the mean of e.m.f. of the two cells (case (d) is not so simple) In each case internal conductances are added to give total conductance.

What happens if one adds further (say n) parallel arms comprising pure voltage sources and conductances in series? Well one can carry on reducing all voltages across the terminals AB for each source in turn, afterwards adding them all together. The voltage across terminals AB given by voltage V_1 for example will then be $V_1G_1/(G_1+G_2+G_3+\ldots+G_n)$. Addition of these individual contributions to the terminal voltage will give

$$V_{AB}=(V_1G_1+V_2G_2+\ldots+V_nG_n)/(G_1+G_2+\ldots+G_n).$$

You may recognize this as a statement of Millman's theorem, which deserves to be better known than it seems to be; for not only do all the usual circuit theorems stem from it as corollaries, but it is extremely useful, having wide application in its own right, and often saving a fair amount of work. It certainly merits the more detailed formal proof that follows.

Millman's theorem and its corollaries

A network of n parallel arms, each consisting of a voltage source in series with an admittance is shown in Fig. 7, the internal admittances of the voltage sources, V_1, V_2 $\ldots V_n$ being included respectively in the associated admittances, $Y_1, Y_2, \ldots Y_n$.

Millman's theorem states that the voltage across the rails, that is between terminals A and B, is

$$V_{AB}=\sum_{k=1}^{n}V_kY_k\Big/\sum_{k=1}^{n}Y_k$$

The proof is as follows. There are $n-1$ loops formed by the parallel arms. Let the currents in each of these loops, in accordance with Maxwell's cyclic-current rule be $I_1, I_2, I_3, \ldots I_{n-1}$, as shown. Note that in each of the n arms, except the first and nth, there are two currents acting, the one shown on the left adding to the generator voltage by its action with the associated admittance, and the one on the right subtracing by a similar action from the generator voltage. The first and last branches will be acted on by only one current each, I_1 and I_{n-1} respectively.

As the voltage V_{AB} is the same wherever terminals A and B happen to be connected to their respective rails, Ohm's law applied to each of the arms will give the following equations for the resultant loop currents through the branches

$$I_1=(V_1-V_{AB})Y_1$$
$$I_1-I_2=(V_{AB}-V_2)Y_2$$
$$I_2-I_3=(V_{AB}-V_3)Y_3$$

and generally

$$I_{n-2}-I_{n-1}=(V_{AB}-V_{n-1})Y_{n-1}$$
$$I_{n-1}=(V_{AB}-V_n)Y_n$$

By subtraction of the individual terms on each side

$$I_1-(I_1-I_2)-(I_2-I_3)-\ldots-(I_{n-2}-I_{n-1})-I_{n-1}=0$$
$$=(V_1-V_{AB})Y_1-(V_{AB}-V_2)Y_2-\ldots-(V_{AB}-V_n)Y_n$$

i.e. $V_1Y_1+V_2Y_2+V_3Y_3+\ldots+V_nY_n-V_{AB}(Y_1+Y_2+\ldots+Y_n)=0$

$$\therefore\ V_{AB}=(V_1Y_1+V_2Y_2+\ldots+V_nY_n)/(Y_1+Y_2+\ldots+Y_n)$$

or

$$V_{AB}=\sum_{k=1}^{n}V_kY_k\Big|\sum_{k=1}^{n}Y_k \qquad (1)$$

Absence of a voltage source in any one of the branches implies that a generator of zero volts is in series with the admittance in that branch. The term containing their product in the numerator is therefore zero and need not be present, although the admittance will appear in the denominator. For example, if all the generators except V_1 in Fig. 7 are reduced to zero, the terminal voltage becomes

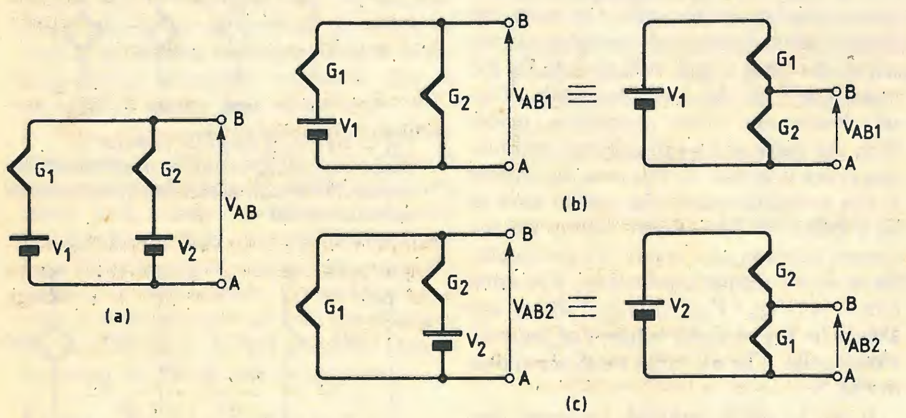


Fig. 6. How the rail voltage V_{AB} is calculated. Considering it as two simple voltage dividers each with a single voltage source, V_1 gives rise to a rail voltage $V_{AB1}=V_1G_1/(G_1+G_2)$ (b), and V_2 gives rise to a rail voltage $V_{AB2}=V_2G_2/(G_1+G_2)$ (c). Total voltage $V_{AB}=V_{AB1}+V_{AB2}$.

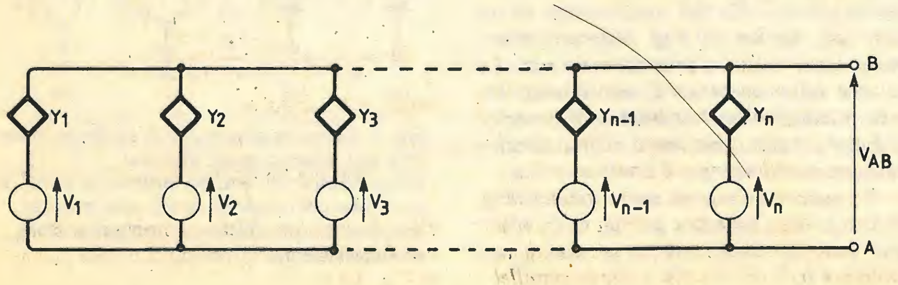


Fig. 7. Circuit having n parallel arms each with a voltage generator and series admittance. Terminal voltage is given by Millman's theorem, $V_{AB}=\Sigma VY/\Sigma Y$.

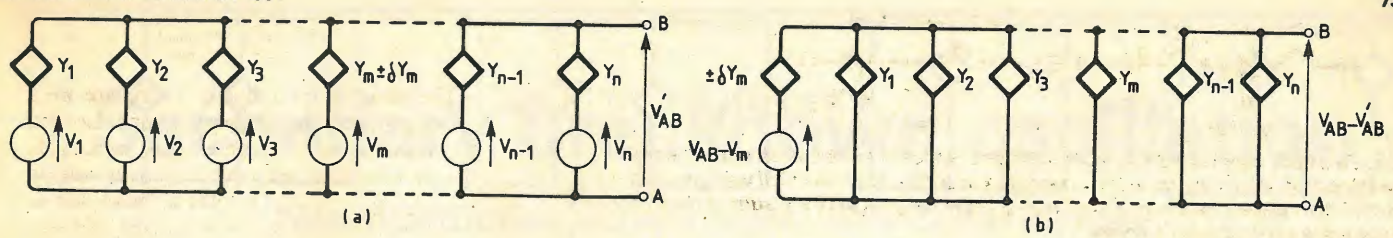


Fig. 8. In the compensation corollary, admittance Y_m of Fig. 7 is changed by $\pm\delta Y_m$ as in (a) changing V_{AB} to V'_{AB}. Change in current through any arm is given by (b).

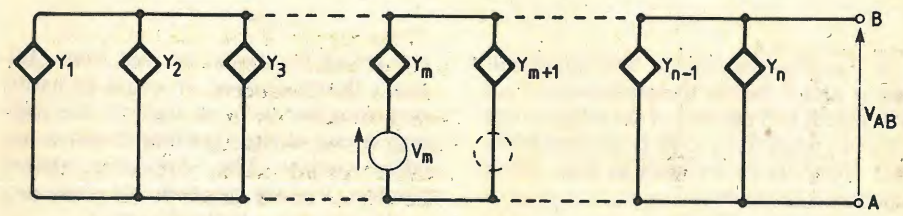


Fig. 9. Fig. 7 modified, illustrating reciprocity corollary. The current following in Y_{m+1} is $V_{AB}Y_{m+1}$. Same current will flow in Y_m if V_m is in series with Y_{m+1}.

$$V_{AB}=V_1Y_1\Big|\sum_{k=1}^{n}Y_k.$$

Corollary 1. If $Y_1=Y_2=Y_3=\ldots=Y_n=Y$ say,

$$\sum_{k=1}^{n}V_k/n=\overline{V},$$

where $\overline{V}$ is the arithmetic mean of the generator voltages.

Corollary 2. If $V_1=V_2=V_3=\ldots=V_n=V$ say, $V_{AB}=V$ the common generator voltage; it is also the arithmetic mean of the generator voltages.

Corollary 3. The expression for Millman's theorem given in equation 1 can be expressed as

$$V_{AB}=V_1Y_1/\Sigma Y+V_2Y_2/\Sigma Y+\ldots+V_nY_n/\Sigma Y$$

Each numerator on the right-hand side of this expression contains only the product of the branch voltage source and the admittance associated with it, whilst each denominator consists of the sum of all the admittances. Therefore, if all the voltage sources except one in turn are reduced to zero so that the voltage between terminals A and B then becomes consecutively

$$V_{AB}=V_1Y_1/\Sigma Y,\ V_{AB2}+V_2Y_2/\Sigma Y,\text{etc., then}$$
$$V_{AB}=V_{AB1}+V_{AB2}+V_{AB3}+\ldots+V_{ABn} \qquad (2)$$

Suppose one wishes to find the total current through the mth arm where the admittance Y_m is in series with the voltage V_m. Well, V_m is reduced to zero for the acquiring of all the voltages on the right-hand side of equation 2, except for the one that is used to obtain V_{ABm} when the current through Y_m is $(V_{ABm}-V_m)Y_m$. The total current through Y_m is therefore

$$I_m=V_{AB1}Y_m+V_{AB2}Y_m+\ldots+V_{AB(m-1)}Y_m+\ldots+(V_{ABm}-V_m)Y_m+\ldots+V_{AB(n-1)}Y_m+V_{ABn}Y_m$$

$$=\Big(\sum_{k=1}^{n}V_{ABk}-V_m\Big)Y_m \qquad (3)$$

From equation 2, equation 3 can also be expressed as $I_m=(V_{AB}-V_m)Y_m$, as would be expected.

Equations 2 and 3 express the *principle of superposition:*

In a network containing two or more sources of electrical energy the voltage across, or the current in, any branch may be found by setting in turn all sources except one to zero, and calculating the voltage or currents due to this single source, afterwards adding the results algebraically.

Corollary 4. Suppose it is desired to find how the current through a particular branch, say that containing V_j, and Y_j, is altered if an admittance in some other branch is altered.

The current through a branch containing V_j and Y_j is $I_j=(V_{AB}-V_j)Y_j$. Suppose now that one of the other admittances is changed by a small amount; say Y_m is changed to $Y_m+\delta Y_m$ causing V_{AB} to change to V'_{AB}. The current now flowing in the jth branch is

$$I'_j=(V'_{AB}-V_j)Y_j.$$

The change of current in the jth branch is therefore

$$I_j-I'_j=(V_{AB}-V'_{AB})Y_j=$$

$$\left\{\frac{\sum_{k=1}^{n}V_kY_k}{\sum_{k=1}^{n}V_k}-\frac{\sum_{k=1}^{m-1}V_kY_k+V_m(Y\pm\delta Y_m)+\sum_{k=m+1}^{n}V_kY_k}{\sum_{k=1}^{m-1}V_k+Y_m\pm\delta Y_m+\sum_{k=m+1}^{n}Y_k}\right\}.Y_j$$

which after a little manipulation becomes

$$I_j-I_j=\frac{(V_{AB}-V_m)(\pm\delta Y_m)}{\Sigma Y\pm\delta Y_m}Y_j \qquad (4)$$

There may not be an independent voltage acting in the branch containing Y_m, in which case this last expression reduces to

$$I_j-I'_j=V_{AB}(\pm\delta Y_m)/(\Sigma Y\pm\delta Y_m).$$

So, *if the admittance Y of a branch of a network* such as that represented by Fig. 7 *is altered by an amount δY, the change in current flowing in another branch of the circuit due to the alteration would be that which would arise if the original terminal voltage were a generator voltage acting in conjunction with any voltage generator previously present in the branch* (containing Y) *in series with an admittance δY, this branch being in parallel with all the pre-existent branches which have had their voltage generators replaced by infinite admittances.* See Fig. 8. You may recognize this as a statement of the *compensation theorem.*

Corollary 5. Consider Fig. 9, a modified version of Fig. 7, where all the voltages are reduced to zero except V_m. The voltage across the rails is

$$V_{AB}=V_mY_m/\Sigma Y$$

The current through admittance Y_{m+1} is

$$I_{m+1}=V_{AB}Y_{m+1}=V_mY_mY_{m+1}/\Sigma Y$$

If V_m is now transferred to the $(m+1)$th branch in series with Y_{m+1} the new voltage across the rails is

$$V_{AB}=V_mY_m+1/\Sigma Y$$

and the resultant current through Y_m is

$$I_m=V_{AB}Y_m=V_mY_mY_{m+1}/\Sigma Y$$
$$\therefore I_{m+1}=I_m. \qquad (5)$$

Hence *the current produced in any branch $(m+1)$th of the network by an e.m.f. V_m in any other branch (mth) equals the current in the other branch (mth) which would arise if the e.m.f. V_m, was transferred to the first branch, $(m+1)$th.* This result is usually known as the reciprocity theorem.

Corollary 6. The rail voltage in Fig. 7 was found by Millman to be

$$V_{AB}=\sum_{k=1}^{n}V_kY_k\Big|\sum_{k=1}^{n}Y_k.$$

Suppose an $(n+1)$th parallel branch consisting of an admittance Y_{n+1} is added across the rails of Fig. 7. The new rail voltage becomes

$$V'_{AB}=\sum_{k=1}^{n}V_kY_k\Big|\sum_{k=1}^{n+1}Y_k$$

$$\text{hence } V'_{AB}/V_{AB}=\sum_{k=1}^{n}V_k\Big|\sum_{k+1}^{n+1}Y_k$$

which is the same as

$$V'_{AB}=\sum_{k=1}^{n}Y_k\Big|\Big(\sum_{k=1}^{n}Y_k+Y_{n+1}\Big) \qquad (6)$$

Fig. 7, with its added admittance Y_{n+1}, can therefore be represented by the circuit of Fig. 10(a).

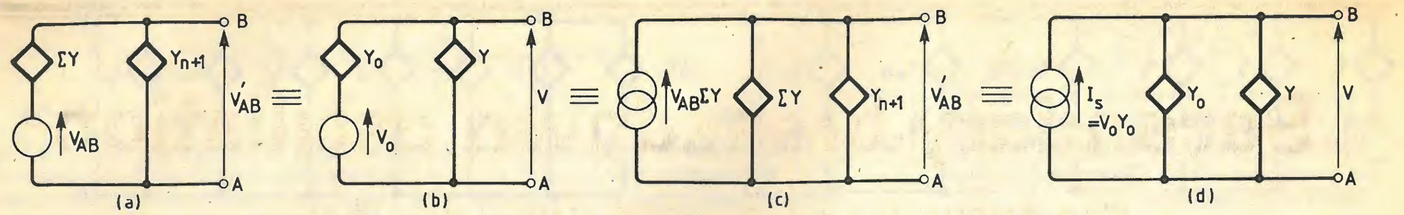


Fig. 10.

If $V_{AB}=V_0$, the voltage between terminals A and B before the admittance Y_{n+1} was added, and the sum of the admittances $\Sigma Y_k=Y_0$ while Y_{n+1} itself is denoted by Y, Fig. 10(a) can be redrawn as Fig. 10(b); equation 6 becomes

$$V=V_0Y_0/(Y+Y_0) \qquad (7)$$

where $V=V'_{AB}$ is the new voltage across the terminals.

Therefore, *if a parallel admittance Y is added to a network, the voltage V across Y is the same as if the original voltage V_0 across the terminals were acting in series with an admittance equal to the sum of the original admittances, in parallel with the added admittance.*

From Fig. 10(b) and eqn 7 V_0Y_0 is the current which would flow if the parallel admittance Y were short-circuited. If this short-circuit current is denoted by I_0 then

$$V=I_0Y+Y_0 \qquad (8)$$

This expression is known as the dual[1] of Thevenin's theorem, as it is usually expressed in its impedance form.

Corollary 7. The numerator $V_{AB}\Sigma Y$ in equation 6 is the current that would flow as a consequence of a generator voltage V_{AB} placed directly across the rails of the Millmann circuit of Fig. 7, when all the voltage sources are zeroed. Fig. 10(c) can therefore be drawn to show the terminal voltage V'_{AB} derived from a current source generating a current I_s; and if the same changes in voltage and admittance are made as in the case of corollary 6, the generalized circuit of Fig. 10(d) results. Clearly Figs 10(a) and (b) are equivalent to Figs 10(c) and (d) with respect to the added admittance Y: the current I_y through Y is V_y in both cases, or $I_y=V_0Y_0Y/(Y+Y_0)=I_0Y/(Y+Y_0)$. Therefore a Thevenin-equivalent circuit may be converted into the constant-current form shown. This is usually known as Norton's theorem, stated as

The circuit of Fig. 7, is identical to a circuit consisting of a current generator developing the short-circuit current in parallel with an admittance equal to the sum of the individual admittances, the voltage across the terminals being the same. Corollaries 6 and 7 are sometimes known as the Helmholz equivalent-source theorems.

All these corollaries have been enunciated as theorems – embracing the concept of impedance rather than admittance – and proofs will be found in any good textbook on electrical theory, for example, reference 3. There is a wealth of articles on Thevenin's theorem alone to be found in previous issues of *Wireless World*[4,5] and elsewhere.

The question arises as to just how general are all the results derived from Millman's theorem above. It would be hardly surprising for them to apply to the Millman circuit of Fig. 7 as they would to any other circuit. The Appendix shows, however, that any two-terminal active network reduces to a Millman circuit.

Appendix

Any two-terminal active circuit reduces to a Millman circuit.

In an circuit containing admittances and voltage sources, any single voltage source or combination of such sources can be replaced completely or partially by another single pure voltage source or sources, provided any sources dispensed with leave their internal admittances behind. New sources retain the admittances of the superseded sources, whilst any new sources are placed in series with pre-existent source admittances or other admittances, so that with regard to the placement of any pair of terminals, the admittance looking into, and the voltage across, those terminals are the same as before any changes were made.

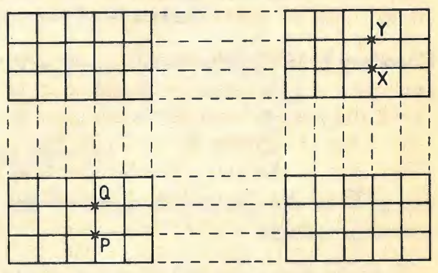
I call this the principle of equivalence and transferability of pure voltage sources, and it is self-evident. For the terminal voltage will always depend, according to the circuit configuration, on definite fractions $m_1, m_2 \ldots m_n$ of the voltage sources $V_1, V_2 \ldots V_n$ such that their summation equals the terminal voltage. That is, terminal voltage is

$$V=\sum_{i=1}^{n} m_iV_i$$

But notice that it may not be at all easy in any particular circuit to see what those fractions are. As a simple example of the principle consider Fig. A(a) to (e). Each is equivalent; each has the same voltage $V_{AB}=4V$ across the terminals; and each has the same admittance $Y_{AB}=10S$ looking into the terminals, that is with the voltage sources short-circuited. The internal admittance of the original source, Fig. A(a), is included in the $2S$ admittance, and remains there throughout. A voltage source cannot be placed in parallel with an admittance, say across $8S$ in Fig. A(a), for Y_{AB} would then be infinite when this source was shorted out. The figures show just five of the infinite variations possible.

A second proposition is a theorem due to J. L. la Cour, which states: *with respect to any two pairs of accessible terminals any passive linear network may be replaced by a two-mesh or T-network, and in general no more simple form can be found.*

Now consider a circuit, represented below, that possesses admittances and voltage



sources to any degree of complexity (the branches shown may consist of either or a combination of both). By the first principle, stated above, all the voltage sources that are present can be replaced by one. Therefore, let all the voltage sources be transferred to the branch PQ prior to this branch being removed temporarily, Fig. B(a), to leave the passive four-terminal network whose input and output ports are defined by the two pairs of terminals PQ and XY respectively.

J. L. la Cour's theorem applied to the remaining network between the two ports results in reduction to the T or star connection of admittances shown in Fig. B(b). Returning the temporarily detached branch PQ to its rightful place, the input port terminals, and replacing the combination of Y_{pq} and Y_1 by the single admittance Y_a results in the circuit of Fig. B(c).

The circuit of Fig. B(c) can be transformed to that of Fig. C(a), T or star to π or delta. Admittance Y_B is clearly redundant as far as conditions at the output terminals X and Y are concerned; for its removal makes no difference to

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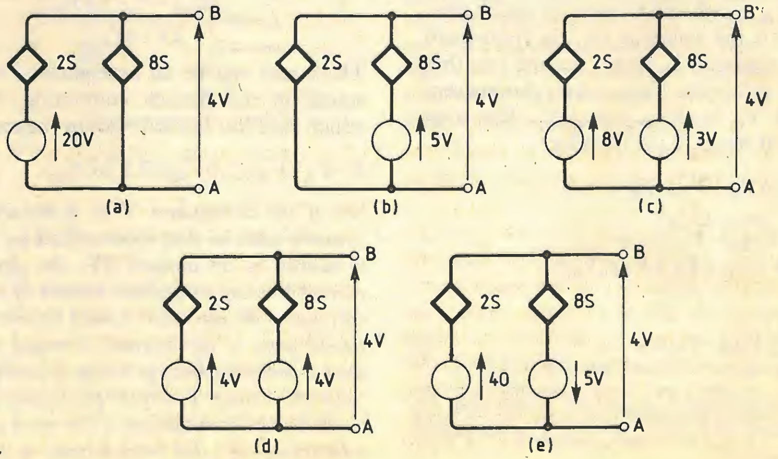


Fig. A.

Secondary breakdown oscillation

Non-destructive experiments with Zener diodes

by D. Di Mario

Although secondary breakdown exhibits a negative resistance characteristic, its application has been limited by the destructive nature of the phenomenon. This article describes some investigations carried out with specially prepared Zener diodes in oscillator circuits.

If the graph of collector current versus voltage for a transistor is considered beyond the normal working range, there are two areas of interest as shown in Fig. 1. The first, known as the primary breakdown region, is where a slight increase in voltage causes a large increase in current, which can damage a transistor. However, Zener diodes operate in this region and a special manufacturing process insures their safe operation.

The secondary breakdown region is catastrophic and normally causes the destruction of a semiconductor in a similar way to the discharge through a dielectric. Although it is imperative that a semiconductor never operates at or near this point, the area is interesting because it exhibits negative resistance and could therefore be used to oscillate or amplify. For secondary breakdown experiments, Zener diodes are preferable to transistors because they already operate safely in the first region. After much inspiration and desperation I found that glass-encased Zener diodes can operate in the second region provided that a certain procedure is observed. All low-power types seem suitable at every voltage above 12V. Below 12V it is increasingly difficult to obtain correct operation, and with 8V diodes it is impossible. This shows that the phenomenon is due to avalanche breakdown and not the true Zener effect.

For correct operation, certain permanent changes must be made in the junction. 400mW 16V diodes, type IN966, or 18V type 1N967 were used as a compromise between ease of operation and the need to work with a low voltage. The diode is fed from a constant current generator, see Fig. 2, and the current is slowly increased so that the device has time to warm up.

At about 30% above the normal Zener voltage the potential drop across the junction starts to decrease and the temperature reaches about 200°C. At this point the diode is left for at least 10 minutes, which seems to insure a certain constancy in the modified devices, and the temperature coefficient changes from positive to negative. During the preparation period there are occasional and spontaneous bursts of high frequency oscillation, therefore monitoring with an oscilloscope is advisable.

An alternative method for preparing a diode is to use a variable power supply connected across the Zener and a 150Ω resistor in series. The voltage is increased until the diode is in the secondary breakdown region and then the voltage is quickly decreased. With some practice a diode can be safely prepared and then tested as shown in Fig. 3. A supply of about 18V is initially required for a few seconds to bring the junction to the operating temperature of over 200°C. This temperature is critical and small metal clips holding the leads can be used as adjustable heatsinks to achieve maximum efficiency, which reached 10% in the audio oscillator shown in Fig. 4. Prepared diodes were tried as amplifiers, but the amplification factor was generally just above unity and, although one sample reached 10, it did not operate for very long. A tendency for the semiconductor to oscillate rather than amplify made measurements difficult and frustrating. The high frequency circuit shown in Fig. 5 was tried successfully, but an intrinsic capacitance in parallel with the Zener was difficult to estimate because the samples varied considerably. The frequency limit was found by using the circuit in Fig. 6, and during experiments frequencies up to 100MHz were observed.

Oscillation takes place with a current between 70 and 250 mA. At this point the normal power dissipation of the device is exceeded, but this should not cause any problems. The oscillation stops if the current is increased, and then returns from 0.5 to 1A. This secondary oscillation is at a higher frequency and careful control of temperature and current is required to obtain steady operation. With a current above 1A the junction glows in darkness

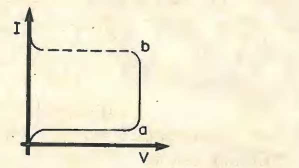


Fig. 1. Collector current versus voltage showing primary breakdown at a, followed by secondary breakdown at b. Region b exhibits negative resistance.

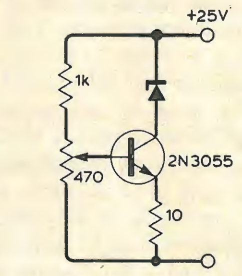


Fig. 2. Constant current generator.

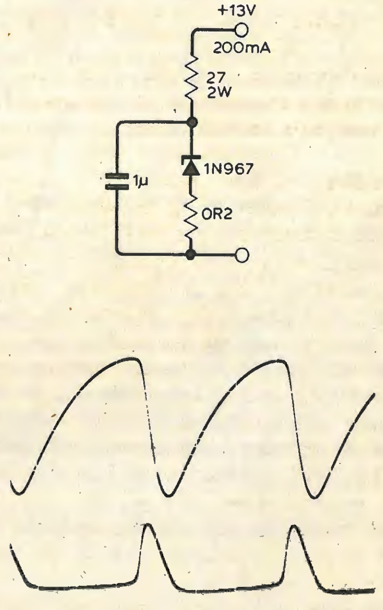


Fig. 3. (a) Test circuit to measure dynamic voltage and current, F = 18kHz, (b) Waveforms across 0.2Ω resistor, 3V (top) and pk-to-pk.

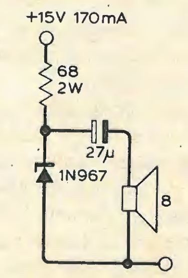


Fig. 4. Audio oscillator with an efficiency of 10% at 1kHz.

and a third oscillation region occurs, but operation is extremely difficult and sporadic. Once the junction has been subjected to the higher current range it will not operate reliably in the lower range, probably due to irreversible changes taking place at the higher temperature. However, the current has been increased until a voltage drop of only 1V was present without destroying the junction. It should be noted that with very high currents the glass case starts to melt and physical breakdown can occur before junction breakdown.

In most samples, when secondary oscillations were triggered, a change of state took place at a 3.3ms interval. The junction oscillates at a certain frequency, then changes to another for the next 3.3ms, followed by a state of self-modulation where the modulating signal is similar to noise. The sequence then repeats and produces a cycle with four different states. Once steady operation is achieved, the 3.3ms interval appears to be independent of temperature and current, but differs between samples. This behaviour has not been investigated further and cannot be explained.

Although it is unlikely that a standard Zener diode will be used in the secondary breakdown region, further study might result in some interesting devices. □

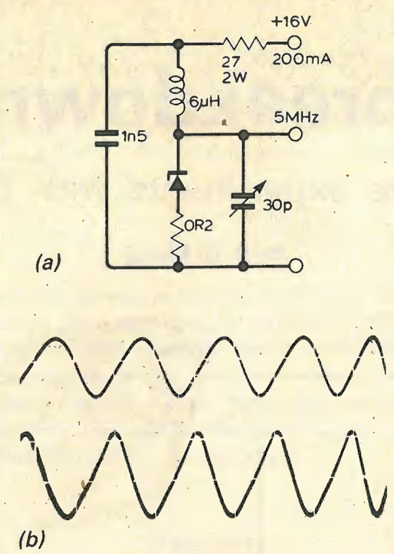


Fig. 5. (a) High-frequency test circuit. Current and temperature must be stable to avoid frequency drift. (b) Waveforms, 3V (top) and 10mV pk-to-pk.

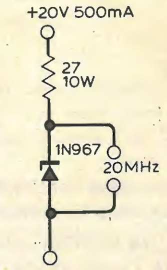


Fig. 6. Test circuit to measure highest oscillation frequency.

BOOKS

The Cosmic Ecosystem, by Alan Johnston, gives one the impression of having been written by a desperate man, frustrated because the scientific establishment closes ranks at the sight of him. It contains many ideas which are quite likely to goad a 'respectable' scientist into a fit of the vapours, a reaction which, however often the ideas are repeated, never seems to bring intelligible responses.

One of the notions put forward by Johnston is the one about gravitational repulsion, in which the force of gravity is due to radiation pressure, two bodies screening each other from the pressure and being forced together. This is not a new idea (Alex Jones came up with it a long time ago) but it has yet to receive any kind of considered reply, positive or negative. The author is similarly iconoclastic on the subjects of the Big Bang, continuous creation and the red shift, and several other theories enshrined in scientific litany. One wishes Mr Johnston well and can only hope that his shoulders are broad. The Cosmic Ecosystem is published at £7.95 by Wildwood House Ltd, 1 Prince of Wales Passage, 117 Hampstead Road, London NW1 3EE.

Sound Recording, by John Eargle, is a text book of sound studio techniques, intended for both technicians and arts students. The author points out in his preface that those students of sound recording who approach the subject from a musical background often understand how to use studio equipment, but are unclear on how it works. Since, however, there is little 'basic' information in the book on electronic circuitry, one feels that such students will continue to be unclear, at this level at least.

At a more peripheral level of description, the book is excellent. It begins with two chapters on sound and hearing, in 50 pages, and two more on stereo and surround sound (called 'quad') which describe most of the systems known, although there are no references later than 1973.

The rest of the book covers all the techniques and equipment employed in a studio, from microphones to monitor loudspeakers, with chapters on audio control systems, tape and disc recording, signal processing and a section on digital techniques. The final chapter advises on the economics and technical aspects of establishing a low-cost recording studio. Published at £16.45 by Van Nostrand Reinhold.

Unified circuit theory

continued from page 74

either the voltage at the terminals, or the admittance looking into the terminals. It may therefore be removed to leave the Millman circuit of Fig. C(b) which has a generator of zero voltage in one of its branches (but which by the first principle could easily be furnished with a more active generator). □

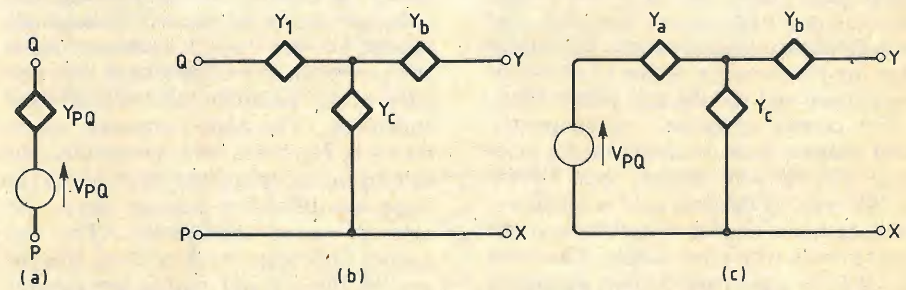


Fig B.

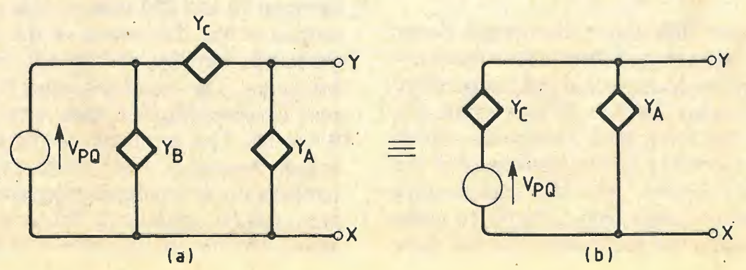


Fig. C.

References

1. Duals. *Wireless World* 1952, vol. 58 pages 152-5.
2. Millman's theorem, and parallel generator theorem, see Electric Circuit Theory by F. A. Benson & D. Harrison, Arnold 1959, page 116, and Useful network theorems, by J. Millman, *Proc. IRE* vol. 28, 1940 page 413.
3. Circuit Theory Analysis, by J. Mittleman. Hayden 1964 and Iliffe 1965.
4. *Wireless World*, vol. 55 1949, pages 109-12.
5. *Wireless World*, vol. 71 1965, letters page 144.

Wind speed and direction meter

2 – Display circuitry and testing

by N. Pollock

Circuits to show wind direction by both analogue and digital displays, and to indicate wind speed conclude the two-part article.

Display

For maximum ease of use, a digital direction display in the form 'port or starboard, 0 to 180°' was considered best. To obtain this display format, three decades of 4029 presettable, up-down, b.c.d. counters were used as in Fig. 9. The PRESET ENABLE pulse presets the counters to 180 and sets them to count down at the start of each encoding cycle. The GATED CLOCK × 10 pulses then cause the counters to decrement. If the counter output reaches 000, the counters are set to count up and further GATED CLOCK × 10 pulses increment the counters. This pulse train, which contains from 0 to 360 pulses, therefore causes the counter output to go from 180 to 0 to 180. The port/starboard information is given by the counters counting either up or down at the end of the pulse train. The middle and high-order decades and the count-up-or-down information is latched with two 4042s to drive the 'analogue' direction display, which is described later.

To keep the cockpit uncluttered and to minimize power consumption, it was decided to use a single digital display and switch it between direction and speed. To facilitate this switching, the direction counter outputs drive a display bus via two 4503 tri-state buffers.

The speed measuring system (Fig. 10) is quite straightforward. The CLOCK signal is gated by the LM 556 timer into three decades of b.c.d. counter, whose outputs drive tri-state buffers. A 4042 latch, a 74C85 digital comparator and an RS flip flop are used to generate the SLOW/FAST signal required by the clock-frequency multiplying phase-lock loop in Fig. 8 of part 1 of the article.

The prototype used 7-segment l.e.d. displays, as shown in Fig 11, because of their low cost. Leading-zero blanking logic was included to conserve power and to make the display more easily readable. Red and green l.e.ds, which only operate in the direction mode, are used to indicate port or starboard. The displays and the segment dropping resistors (R) will depend on the size and brightness required.

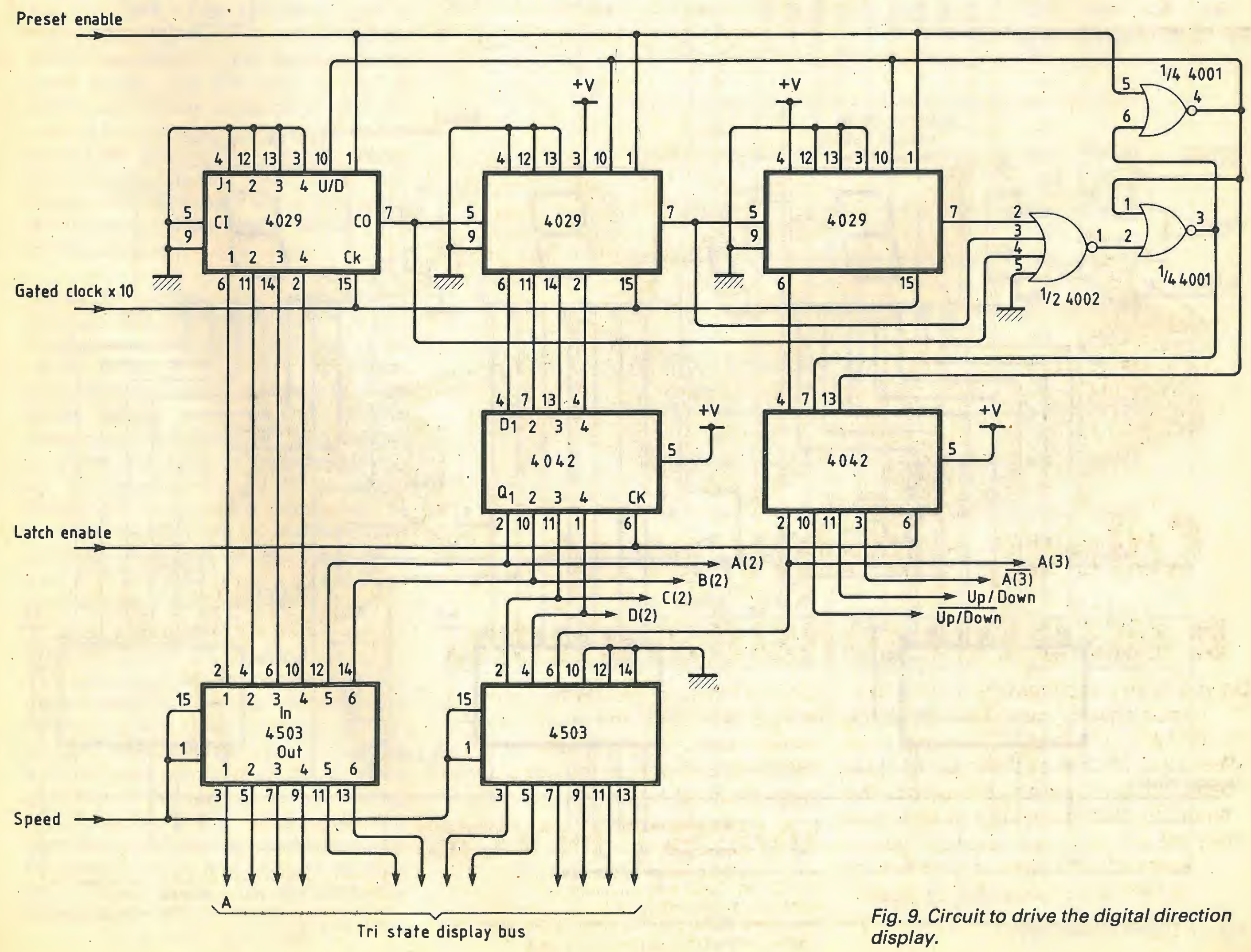


Fig. 9. Circuit to drive the digital direction display.

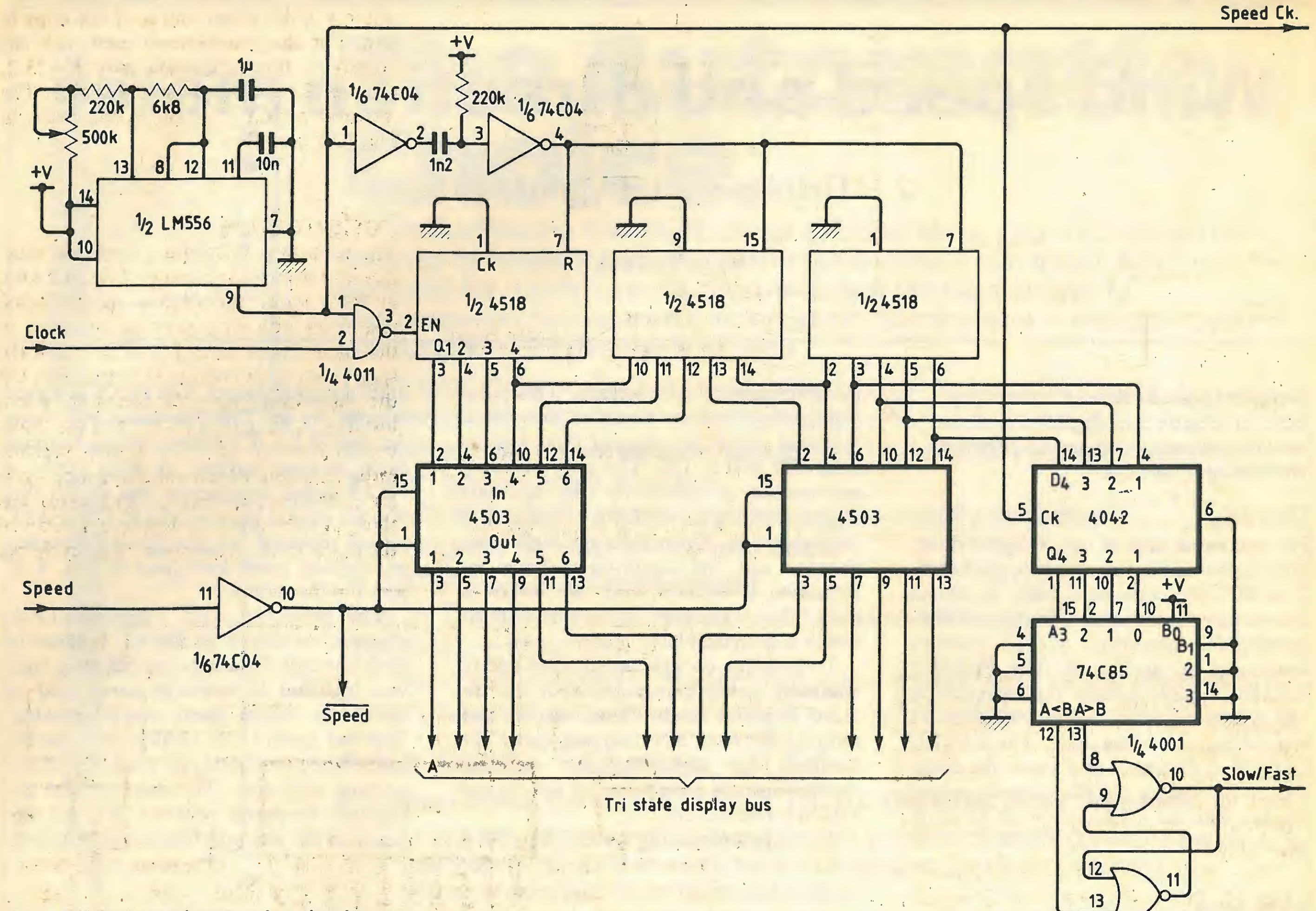


Fig. 10. Wind speed measuring circuit.

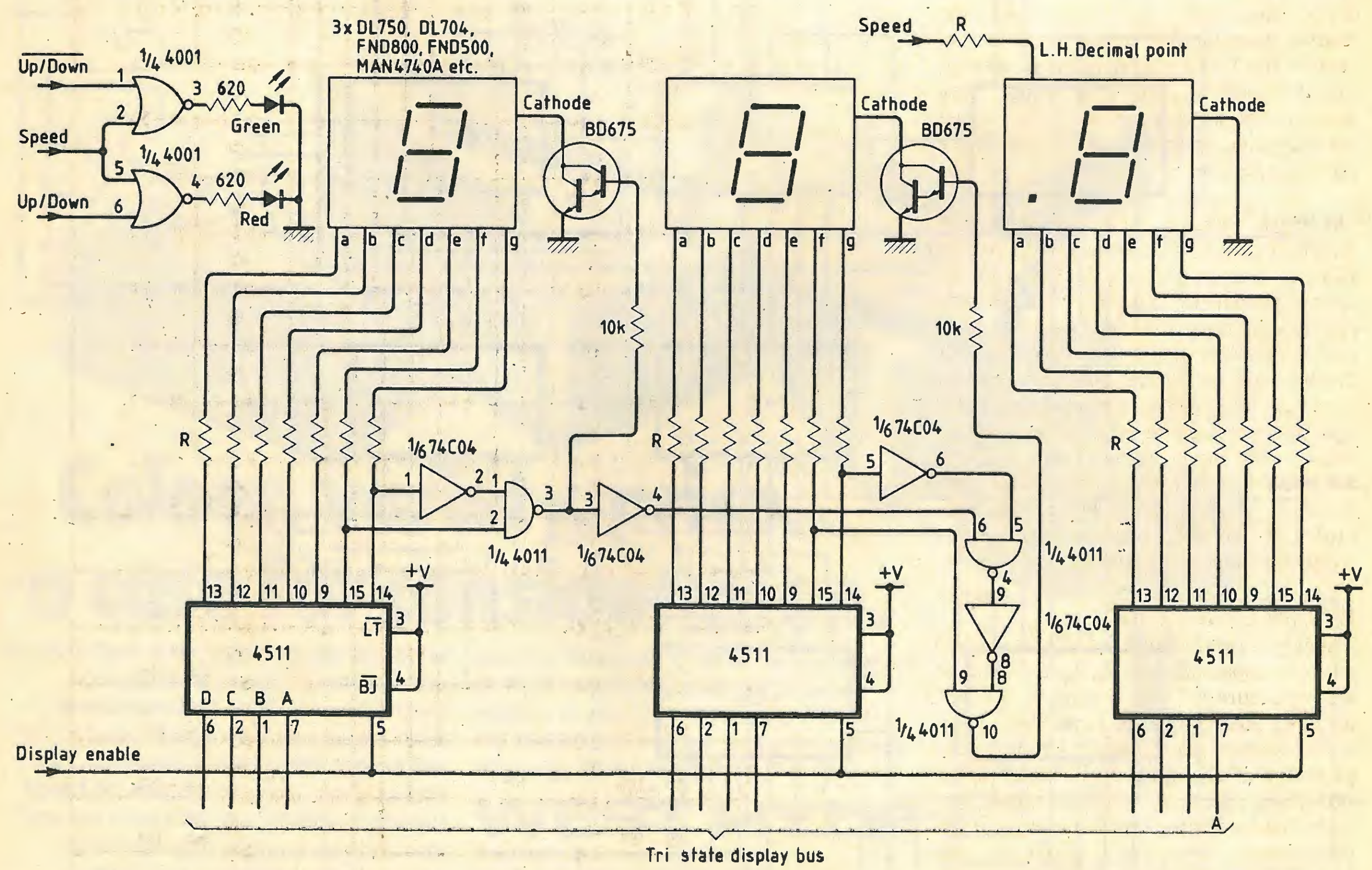
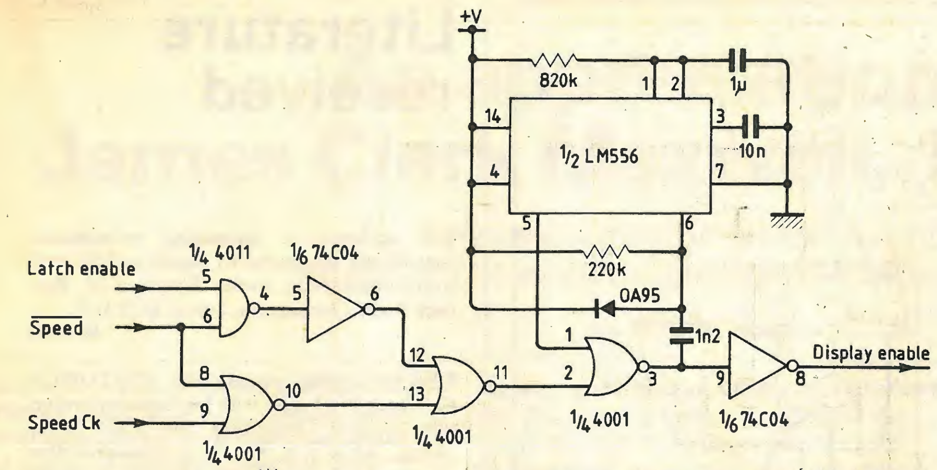


Fig. 11. Digital display circuit.



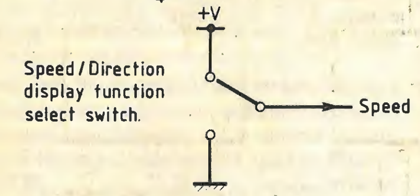
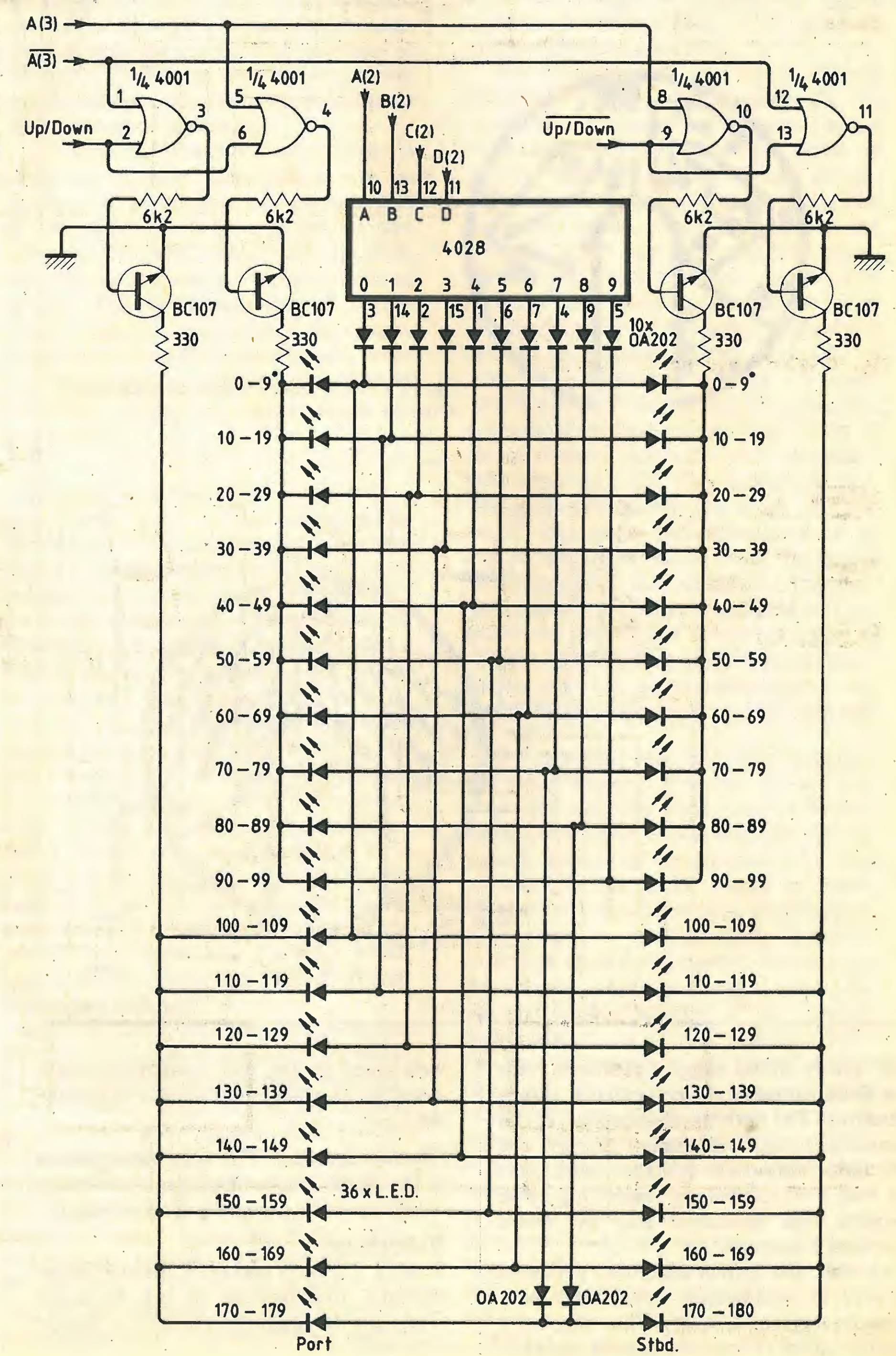


Fig. 12. Circuit of display control logic.

Fig. 13. Circuit of l.e.d. 'analogue' display seen in Fig. 14. ▼



If reasonably priced liquid-crystal displays were available their use would be preferred because of their lower power consumption and superior visibility under varying lighting conditions. If l.c.d. displays were used, the drivers would have to be changed from 4511s to 4543s and a display oscillator added.

The display control logic is shown in Fig. 12. The LATCH ENABLE signal is derived from the direction circuit and the SPEED and SPEED CLK signals from the speed circuit. The 556 timer is used to disable the display updating so that the reading is updated at a maximum rate of about 2 Hz. Without this update disabling, the direction update rate would vary directly with windspeed and at high speed the flickering of the display would make reading difficult.

Analogue display. The analogue-type direction display discussed earlier is provided by a circle of 36 l.e.ds driven by the circuit shown in Fig. 13. The display operates continuously, independent of the digital display mode selected. If this display, and the digital direction display, have port and starboard interchanged this can be overcome in one of two ways. Either the anemometer cups can be mounted the other way up to reverse their direction of rotation or the UP/DOWN and $\overline{\text{UP/DOWN}}$ inputs can be interchanged in Figs. 11 and 13.

Speed calibration

The wind-speed calibration can be carried out by comparison with an anemometer of known accuracy or by mounting the unit on a *long* pole in front of a car. Failing this a fair estimate can be made based on the geometry of the anemometer rotor using the information in Ref. 3. This report suggests that for a three-cup anemometer with the geometry typical of commercial units and a 36-hole timing track, the clock calibration factor (K) should be given by:

$$K = 1032/r \text{ hertz/knot}$$

where r is the mean radius of the cups in mm. For the anemometer used with the prototype, this expression gave K=23.2, whereas the measured value was 22.5. The required period T of the speed clock in Fig. 10 is given by:

$$T = 10/K \text{ seconds}$$

Performance

The prototype instrument operated satisfactorily over a speed range from 0.7 knot to 99.9 knot, the higher speeds being simulated with an electric motor drive to the anemometer disc. Below 0.7 knot the speed indication continued to function but the phase-lock loop in the direction system lost lock. The maximum error of the direction indication was ±2° with a periodic error function which exhibited one cycle for 360° direction change. This error was thought to be due to a slight eccentricity of the clock track. Any two or four cycle per

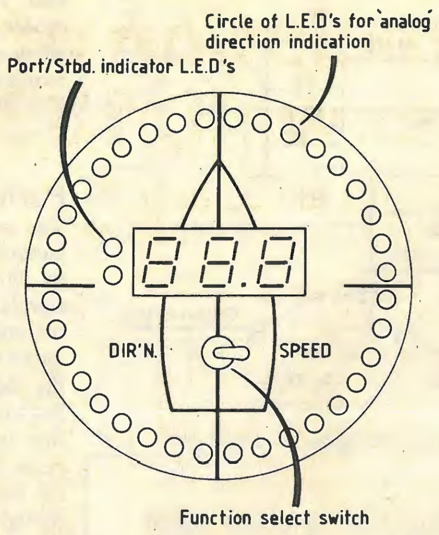
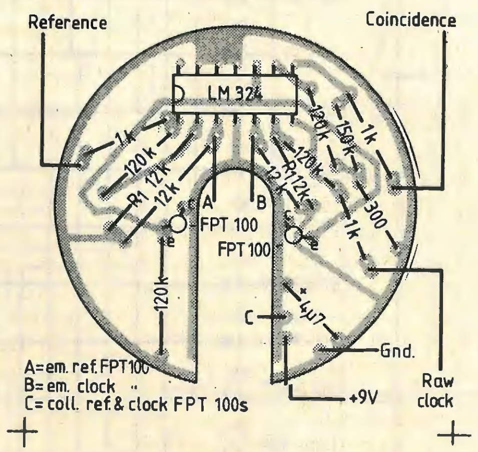


Fig. 14. Possible cockpit display.

Fig. 15. Printed-board layout of mast-head unit.



360° errors would suggest problems with the illumination of the coincidence photo-detectors. The current consumption of the complete system less digital display was 290 mA. The system operated satisfactorily with a 40 metre long connecting cable between the masthead and the main electronics assembly.

Overall, the system described performs as well as commercial instruments at a considerably lower cost. The inherently digital nature of this instrument makes it well suited for use with a microprocessor-based yacht data system or tactical computer.

Aknowledgement. The valuable assistance of Dr. B. D. Fairlie throughout the design of this project is gratefully acknowledged.

References

3. M. J. Brevoort and U. T. Joyner, Experimental Investigation of the Robinson Type Cup Anemometer. N.A.C.A. Report 513, 1935. □

Literature received

Full catalogue of **measuring instruments,** systems and accessories by Rohde and Schwarz is obtainable from Aveley Electric Ltd, Roebuck Road, Chessington, Surrey KT9 1LP. WW401

KEF have published two new KEFTOPICS, Volume 4, No 1 deals with **loudspeaker protection** methods, describing the KEF S-Stop system, and Vol. 4, No 2 is on **crossover filter** design. KEF Electronics Ltd, Tovil, Maidstone, Kent. WW402

Leaflet from Eraser International describes a range of flameless **heat guns** for shrinking tubing, drying, paint stripping, etc. Available from EI at Unit M, Portway Industrial Estate, Andover, Hants. SP10 3LU. WW403

Catalogue of Bulgin **components** is now available, covering an extended range of hardware and now including displays and digital panel meters. A. F. Bulgin and Co. Ltd, Bypass Road, Barking, Essex IG11 0AZ. WW404

Salford has produced a short catalogue to describe the SEI range of **measuring instruments,** including those for the heating and ventilating industry, and also showing the new digital panel meters. WW405

Booklet with the self-explanatory title "Everything you need to know about **cartridges**" is published by Ortofon Manufacturing A/S, 11B Mosedalvej, DK2500 Copenhagen-Valby, Denmark. WW406

Catalogue entitled "Non-contact sensing devices" details a range of inductive **proximity switches** made by Warner Electric Ltd, St Helen Auckland, Bishop Auckland, Co. Durham DL14 9AA. W407

Set of leaflets on Zycor **Prestel terminals** describe an adaptor for use with an existing television receiver, a cassette player to display recorded pages for demonstration purposes and a dedicated Prestel monitor. Leaflets are obtainable from Zycor Ltd, 33 Fortress Road, London NW5 1AD. WW408

The ZIP585 **paper tape punch and reader** are described on two leaflets, available from Data Dynamics Ltd, Data House, Springfield Road, Hayes, Middx. WW409

Designers' guide to the use of high-resolution **a-to-d, d-to-a converters,** sample-and-hold amplifiers, v-to-f converters and digital meters is produced by Analog Devices Ltd, Central Avenue, East Molesey, Surrey KT8 0SN. WW410

Brochure from Brown Boveri Kent contains details of a range of **process control** instrumentation, liquid metering equipment and systems, including laboratory equipment. BBK are at Biscot Road, Luton, Bedfordshire LU3 1AL. WW411

Leaflets on an 8-channel, 12-bit **data acquisition** equipment, based on 8K13 p.c.m. modulator and demodulator units, are available from Johne and Reilhofer (UK) Ltd, Oddstones House, Thompsons Close, Harpenden, Herts. AL5 4ES. WW413

An appreciation of James Clerk Maxwell, 1831-1879

Have we got the allocation of honours between Einstein and Maxwell right?

by M. G. Wellard

In the introduction to his book Electric Waves, Hertz wrote ". . . we can best characterise the object and the result of our experiments by saying: The object of these experiments was to test the fundamental hypothesis of the Faraday-Maxwell theory, and the result of the experiments is to confirm the fundamental hypothesis of the theory". The fundamental hypothesis of the Faraday-Maxwell theory was that space was not empty. This article attempts to explain why we should turn the cuckoo clocks back to Maxwell and start again.

Twice in 1979, BBC television broadcast a long programme celebrating the centenary of Einstein's birth. The production, a serious attempt at education, included a scene of several eminent scientists toasting "Einstein, the hero." A more lighthearted musical tribute was also broadcast. Both programmes highlighted two facts about Einstein's world. Because it is beyond our experiences on this planet, most of us cannot understand it; we can find no analogies. Furthermore, his world is one source of the modern fantasies of science fiction books and films. Once the size or space and the life-span or time of an atom depend on its velocity, the imagination is freed from the constraint of commonsense.

That year was also the centenary of James Clerk Maxwell's death, a fact largely ignored. He died in the prime of life aged 48, leaving science a legacy: his laws of electricity and magnetism and his Treatise on the same subjects. His laws are companions to Newton's law of gravity. These laws, governing the behaviour of the only three forces capable of acting across a distance in space, are based on Newton's laws of motion, governing the dynamic behaviour of all forces. Overseeing all scientific laws is the law of the conservation of energy which Helmholtz first derived mathematically from Newton's laws of motion.

Considering the magnitude of Maxwell's contribution to science, which stemmed from his theoretical interpretation of data and equations derived by his predecessors from thousands of experiments, while Einstein's fame rests on his theoretical interpretation of data and an equation derived by his predecessors from a single experiment – a theory that is safe from experimental verification – the ratio of the allocation of honours between Maxwell and Einstein is difficult to understand.

If Ampere is the Newton of electricity, Maxwell is the Newton of electromagnetism. His treatise, written by Maxwell the teacher, master of the understatement, for the benefit of students of electrical engineering, can, with a little perseverance, take anyone into the mind of this great man of genius. Obviously not everyone's choice today. Maxwell's world is basically one of plain commonsense and geometry, a world filled by a gifted accountant of engineering with the actions and reactions of Newton's bodies in space, where the paramount scientific law is the principle of the conservation of energy.

The tone of Maxwell's treatise is one of confident optimism that physics was on the verge of a complete understanding of all laws of nature. A far cry from the attitude today. Physics is filled with statistics, the analysis of countless opinion polls of energy. The strict principle of energy conservation has given way to the Uncertainty Principle of Heisenberg, and a particle of today may or may not be a wave of tomorrow. Energy has a choice and is ungovernable.

* * *

Looking back, the cause of the lawlessness can be traced to the theoretical interpretation of three experiments: Michelson and Morley's unsuccessful attempt to measure the velocity of the "ether wind", the source of Einstein's Special Theory of Relativity; Thomson's experiment with a form of television tube interpreted as confirmation of Lorenz's theory that electricity is carried by positively and negatively charged particles; and Planck's experiment with a black box, the "ultra-violet catastrophy", which took the wave out of Maxwell's waves and led to the quantum theory. These theories are unified by a common belief that the laws of electricity, magnetism, gravity and motion have a few loopholes. And therefore by inference the law of the conservation of energy which unifies all laws of physics, and if ignored disintegrates physics.

When Maxwell died, the physical nature of atoms was still a complete mystery. While the theories of modern physics were in their infancy at the turn of the century, some atoms were found that emitted electromagnetic waves – they were radioactive – and when Rutherford caused an atom to disintegrate, the data he collected showed that an atom was a form of electricity. He naturally assumed that the constituent parts of an atom were particles, and his planetary model of the atom was amalgamated with the particle theory of Lorenz and the packet theory of Planck to form the basis of the present quantum theory. This theory is used in the investigation of the nature of electricity and magnetism, and by those concentrating on the electromagnetic actions confined to the microscopic spaces between the atoms of solids, liquids and gases. Where electromagnetic energy is transmitted across greater distances, Maxwell's equations without his theory are used.

The ultimate role of science in any society is to forecast mathematically the result of the interaction between two or more identified and specified forms of energy. Scientists are the natural successors to those who consulted the oracle. The scientists' oracle is the principle of the conservation of energy, a principle fundamental to all laws and their equations of physics, and every scientific law from Archimedes' law of the lever to Hubble's law of the galaxies fulfills this role of science. Einstein's theory and the quantum theory were developed to explain the action of identified and specified forms of energy which Newton's and Maxwell's laws could not describe mathematically. The quantum theory has stuck fast for nearly fifty years, and if the cause of its lack of progress is due to faulty theoretical interpretation of data, clearly the interpretation and its associated equations will not satisfy the principle of the conservation of energy. Any attack on modern theory must use the principle as its major weapon deployed under the guidance of Newton and Maxwell.

This principle says that there is a fixed and finite amount of energy in the universe, and therefore energy cannot be generated, destroyed or wasted, and that an infinite amount of energy cannot be concentrated in one finite volume in space. Maxwell defined energy as something with the capacity to perform work. If energy cannot be wasted, all interactions between the various forms of energy must take place at 100% efficiency. If two Newton bodies always act equally and in opposition, the magnitude of the forces of their equal and opposite actions cannot increase indefinitely because an infinite amount of energy cannot be concentrated in one point in space. There must therefore be a safety valve, a limit to the magnitude of the two equal and opposite forces, and when that limit is reached, a transformation of energy from one form to another must take place, the objective being to stop the activity of the forces involved in the interaction.

Early in his treatise, Maxwell emphasised that an infinite concentration of electricity is impossible, and that when the concentration reached a certain limit, something snapped. He quoted the breakdown of capacitors, various glows and discharges as examples. These phenomena must be the result of the forces of equal and opposite actions exceeding a certain limiting value, the transformation of a local electric charge into a wave, for instance. This safety valve, the point of evolution, is the point of catastrophy of Thom's theorem, the two bodies acting equally and in opposition his "attractors". There is a close affinity between the minds of Maxwell and Thom; a fascination with maps, concrete geomentrical pictures of abstract algebra.

If energy operates at 100% efficiency it will always take the line of least action or strain during the interaction of forces, wasting nothing. The attractive forces of gravity always act in a straight line between two centres of force, the line of least action. Although the forces of electricity and magnetism must also be taking the line of least action, that line is often far from straight. The lines of least action in the space surrounding a magnet take some explaining. The fact that Maxwell did explain them theoretically and mathematically is indicative of his genius and the validity of his theory. Modern theory can offer no explanation, whatever of the lines of least action surrounding a magnet. In the case of gravity, there can be only one straight line of least action between centres of force. Energy does not have a choice if it is to conserve itself. A line of least action is not the manifestation of an uncertainty principle. If the line follows the changing energy levels of a half cycle of a wave, that is exactly what the line is describing.

The loophole found by Planck is not difficult to repair if the principle is applied to his theory and experiment. An infinite amount of energy cannot be concentrated in one volume in space.

Planck's theory that electromagnetic energy is emitted in packets or quanta rather than as waves grew from his theoretical and mathematical interpretation of the results of a single experiment, the ultraviolet catastrophy. Because the frequency of Maxwell's waves is a function of the energy to be radiated, the higher the concentration of energy at their source the higher the frequency, Planck thought he could introduce an infinite amount of electromagnetic energy into a black box simply by increasing the frequency of the waves he pumped into it. He was surprised to find that when the frequency of the waves reached the ultra-violet spectrum, the wave energy in the box began to disappear. Planck had made himself a cavity resonator and the waves in his box were beginning to resonate. The waves were satisfying the principle by making sure that an infinite amount of active energy could not be concentrated in the finite volume of his box. There is no sign of little packets of energy in his box, only the resonance of Maxwell's waves, an evolution of energy.

Planck concluded that electromagnetic energy was emitted not as a wave, but as individual packets of energy. The cyclic variations in the energy level of a Maxwell wave can be followed from its point of evolution to point of devolution. The energy of the wave gradually alternates from nothing to a maximum, and then to nothing. Planck's packets do not show themselves until they are completely full, and the amount of electromagnetic energy in every packet is always the same. This quantity, Planck's constant, forecasts the total amount of work the packet would perform by the expenditure of its whole energy. Planck's packet was later transformed by Einstein into a particle, the photon, the unit of the electromagnetic energy of light or to be more precise, the total electromagnetic energy of one cycle of a Maxwell wave, data not available to Maxwell before he died. In Planck's black box there are two "bodies" acting and reacting, and one of them is Maxwell's ether.

Planck's work is very important if the data he cleverly collected from his experiment is interpreted correctly and applied to the relevant equations of Maxwell. To be fair, Planck was always dubious about accepting Einstein's photon. He and many of his contemporaries who were experimenting with electromagnetic waves in a rapidly expanding spectrum, suspected they might be dealing with a very unusual kind of wave which might require a very unusual description. Before Hertz had experimentally verified Maxwell's mathematical proof of the nature of light two experimenters, Michelson and Morley, were planning an experiment that was to shake the foundations of physics and shatter the emerging picture of the universe so carefully and painstakingly pieced together by a lot of dedicated men and women.

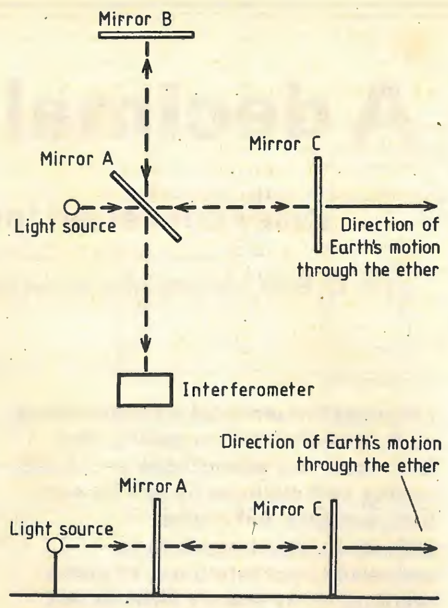
Experiment had revealed, prior to Hertz' experiment, many similarities between the behaviour of both sight and sound, and as sound was known to be a wave of energy passing through a medium, the air, the analogy was taken to its logical conclusion. Light must be a wave of energy passing through a medium, the ether. The earth as it passed through the ether might or might not cause the ether to interact, and it was decided that an "ether wind" blew past the earth. Michelson and Morley set up an experiment to measure the velocity of the ether wind.

They attempted to detect differences in the velocities of two beams of light shining at right angles to each other and travelling along paths of equal length, one beam pointing in the direction of the earth's movement around the sun. They believed the sun was stationary in the centre of the universe and they therefore had a point of reference for calculating the earth's direction of travel through the ether. They did not plan to measure the actual velocities of the two beams of light, only changes in frequencies and wavelengths due to the effect of the ether wind which caused one beam of light to travel more slowly than the other. A wave's velocity equals its wavelength times its frequency.

Their experiment was completely unproductive and they issued a communique warning physicists that light did not obey Newton's laws of motion. If two bodies travel towards each other along parallel straight lines, their passing velocity is the sum of their individual velocities. Michelson and Morley warned that light's velocity was so constant the other body's velocity was always nil. Light always passed you at its constant velocity, no matter how fast you approached it. They never discussed the case of someone travelling in the same direction as light. Newton's first law of motion says that a body travels at a constant velocity unless acted upon by an unbalanced force, and Newton's interpretation of Michelson and Morley's findings would be that light became a temporarily unbalanced force during the act of passing another moving body and brought that body to a halt. Whether one is to believe Newton or Michelson and Morley must remain a matter of individual judgement. The word unbalanced means polarized, and there is no evidence of polarization of light in Michelson & Morley's experiment.

The velocity of a sound wave travelling through air that is physically moving, a wind, is the arithmetic sum of the velocities of both the wave and the wind when measured from the wave's point of origin. With the wind the wave's velocity increases, against the wind it decreases when measured from the wave's point of origin. Waves always travel at a constant velocity in a medium of constant density. If the medium is stationary but the course of the wave is moving, the wave's velocity when measured from its point of origin is unaffected by the motion of the source. Ahead of the source the wave's frequency increases, astern of the moving source the frequency decreases. Only in the direction at right angles to the source's direction of travel is the frequency of the wave unaffected by movement of the source. This change in frequency is the Doppler effect. Under constant conditions the source's direction of travel only affects the orientation of the Doppler effect. When the air is still and a moving sound source such as an aeroplane exceeds the velocity of sound, the sound barrier velocity is independent of the plane's direction of flight.

Look at the plan view of Michelson and Morley's experimental set-up. Light from the source was split by the semi-transparent mirror A along two paths at right angles to each other, and reflected by mirrors B and C back to mirror A where it was reconstituted into one beam. The second diagram clearly shows that the light source and mirror A are moving directly into the face of the ether. If the ether is still, the only effect on the light wave would be to increase its frequency, the Doppler effect. Mirror C is moving away from the light source and the beam of light also sees mirror C moving away. When the light wave hits mirror C, the mirror implants a Doppler effect equal and opposite to the Doppler effect due to the forward movement of the light source. The frequency of the light wave is back to normal. A reverse action of Doppler effects occur on the light beam's return journey. When mirror C reflects the light wave it implants a Doppler effect, reducing the light wave's frequency. Mirror A is moving towards the light wave and therefore cancels the Doppler effect produced by mirror C on the returning wave. No change in frequency or wavelength takes place on this arm of the experiment. On the other arm of the experiment the same sequence of events take place. Because the light wave is travelling along a path at right angles to the earth's direction of travel through the ether, the magnitude of the Doppler effect is almost nil, taking place along two sides of a triangle whose base line equals the distance of the earth's travel during the time the light wave takes to complete its journey. No changes in frequencies or wavelengths occur on this arm of the experiment. Although this experiment was pivoted through 360°detect the directions of maximum interference, it proves nothing, or that the ether does not interact with the earth, and Einstein's theory stands or falls on Michelson and Morley's statement that light was a form of magic. The effect on physics of this one experiment was devastating. Scientists had no alternative but to consult the oracle.



"The effect on physics of this one experiment was devastating. Scientists had no alternative but to consult the oracle."

Lorenz consulted it first and came away with an equation that solved the puzzle. He reasoned that the ether must push against any moving mass and foreshorten it. The faster you travelled the more it flattened your face. The ether in this theory had the capacity to perform work. Because he was trying to explain the strange behaviour of light waves, he remembered that the movement of a sound source did something strange to sound waves called the Doppler effect. So he took the Doppler formula and using his knowledge of squares and roots produced an equation to forecast the amount of foreshortening. When an observer moved, you multiplied the observers' sides by $\sqrt{1-(v^2/c^2)}$ where v is the moving observer's velocity and c the light's velocity. Although the value of v^2/c^2 is insignificant at normal velocities of atoms found on earth, when v approaches the velocity of light v^2/c^2 approaches unity, and when $v=c$ the equation reads $1-1=0$. When the moving observer reaches the speed of light its sides disappear.

Einstein's attack on the problems of the moving observer were much more thorough. He could see that on the one hand light had an effect on the moving observer's velocity. On the other hand, light always travels through space at its constant velocity, the speed of light, although Maxwell said that it did not, it depended on the resistance of his ether. He eventually came to the sensible conclusion that the only way he could make sure that light always passed the moving observer at its constant speed was to agree with Newton and stop the observer moving. So he slowly brought the observer to a halt. He did this by multiplying both the frequency and wavelength of light by Lorenz's equation. This reduced light's speed to nil when the moving observer travelled at the speed of light. Add the two velocities together and he had the constant speed of light. He then called the frequency of light TIME, and its wavelength SPACE, and they became variables dependant on the moving observer's velocity. To confuse the issue he then transferred the variables to Newton's absolute units of time and space of the atoms of the moving observer.

This is admittedly an ingenious solution to the problems raised by Michelson and Morley, but somehow his equation mushroomed into the wonderful world of relativity, an interpretation of Lorenz's equation which in Einstein's own words "is not justifiable by any electrodynamical facts" (his paperback, page 51).

His paperback has been reprinted in the UK at least 22 times. Many people throughout the world have bought the translation of his book hoping to understand something of the physical world around them. Most of them put it down more confused than when they picked it up. Einstein said in his preface that he would do his best to explain his ideas as simply as possible and "in the sequence and connection in which they actually originated." "Actually" he never did. He starts by advising us we are fools. To save ourselves we should contemplate mysterious measuring rods, cuckoo clocks, and gravitational stopcocks for the use of moving observers who abruptly change course without stopping, and he eventually leads us back to Lorenz's equation which just happened to fit his theory. But there he stops. Whether he simply forgot to mention how it all started we shall never know.

Einstein's theory is not an easy target for the principle of the conservation of energy because he avoided tampering with the wavelength and frequency of light, but changed instead the dimensions of the space occupied by the moving observer's mass, and the time that elapsed while the observers dimensions were being changed. All units of dynamics such as mass, velocity, acceleration, density, force, work and energy, can be expressed simply in terms of Newton's fundamental and absolute or universally constant units of time (T) and length (L), the one dimension of space or a volume L^3. The density of a mass (M) described in units of time (T) and lengths (L) is M/L^3, an acceleration L/T^2. A force is the product of a mass and an acceleration, and if the concrete units of time and length are reduced, so are all forces applied to an atom's mass, and the total amount of work the atom will perform, its total energy, is also reduced. Where does the lost energy go? In Einstein's theory, nowhere. When his moving observer approaches the "constant" speed of light, its total energy is almost destroyed; it is a physical wreck. Einstein could not reduce the equal and opposite reactions of Maxwell's ether to the actions of the light wave, because his idol, Lorenz, in his other theory, the electron, had already forbidden the ether to perform work. Two equal and opposite laws for the one ether. Lorenz will one day find his way into the Guinness Book of Records as a greater destroyer of energy than King Canute.

Einstein was forced to limit the velocity of an atom to the speed of light because once his moving observer mathematically exceeded the "constant" speed of light, he would mathematically become a centre of negative energy and vanish down a black hole, proof that you can prove anything with figures. Einstein found himself with mass that tended to shrink when it moved by the value of his equation. To balance his books and satisfy the principle of the conservation of energy, he discovered rest mass, which allowed him to unshrink mass by the same value. He called his theory relativity because the total energy of each atom in the universe depended on the atom's velocity relative to a fixed and motionless point in space, all atoms being connected by bendable springs and flexitime clocks to conserve energy. Maxwell would have immediately dismissed Einstein's theory using the argument in Article 852 of his treatise, that the force acting between two "bodies" must be a function of their distance apart only, and if the force is a function of time or the velocity of the bodies, theory would not satisfy the principle of the conservation of energy.

Einstein's favourite occupation was performing what he called thought experiments. In the portable laboratory of his mind he could prove, without fear of contradiction, that scientific history was bunk. His laboratory was the envy of a few second rate accountants in a hurry. An analysis of debits and credits and their equal and opposite actions can be very time consuming, far easier to cook the books and make yourself a quick profit. They renamed Newton's laws the three laws of non-motion. If it moved you either saluted it or multiplied it by Lorenz's equation. Never in the understanding of a field of force, was so little owed by so many, to so few.

To be continued

A decimal Gray code

Easily converted for shaft position coding

by K. G. Barr, Faculty of Natural Sciences, University of the West Indies

For some incremental measurements, such as shaft position coding, the Gray scale has advantages over b.c.d. coding as it changes by one bit only between adjacent codes. Unfortunately, it needs to be converted back into b.c.d. to give a read-out. Gray scale is difficult and therefore expensive to convert and this decimal Gray scale overcomes the difficulty.

The author has recently designed equipment to monitor wind speed and direction. The wind vane drives a slotted disc whose position is sensed by l.e.d./photo-transistor pairs. A code is required to sense the position of the disc and transmit this position to the display and recording equipment. The reflected binary, or binary Gray code shown in Table 1 has the required property that only one bit changes in adjacent codes, but is an expensive code to convert to a decimal form for display.

Table 1: It will be noted that the most significant bit in the fifth column changes to 1 at a count of 16 whereas in the decimal Gray code it changes at a count of 10. In Table 2, D_0 to D_3 represent the decoded decimal number, while g_0^0 to g_1^2 is the decimal Gray coding

The decimal Gray code also shown in Table 1 is much cheaper to convert. It is "reflected" after each decade, and the low order bit of the next higher digit (b_0^1) is required for conversion. The b.c.d. digit ($b_3\ b_2\ b_1\ b_0$) corresponding to the decimal Gray digit ($g_3\ g_2\ g_1\ g_0$) is

$$b_0 = b_0^1 \oplus g_3 \oplus g_2 \oplus g_1 \oplus g_0$$
$$b_1 = g_3 \oplus g_2 \oplus g_1$$
$$b_2 = (g_3 \oplus g_2)\, b_0^1 + g_1\, b_0^1$$
$$b_3 = g_3\, \overline{b_0^1} + \overline{g_2\, g_1}\, b_0^1$$

Whatever the code used, it will in general be non-reflective at the zero point in a scale used to measure shaft position. If the angle is to be measured in 1° steps the codes at the zero point are as shown in Table 2. Additional logic must be added so that the offending bits g_2^1 and g_1^1 are set equal to g_1^2 before conversion when $D_0 = 0$ and the last value of D_2D_1 when $g_0^0 = 1$ was 00 or 35. The logic required for the complete conversion is shown in Figure 1.

It should be noted that the zero point logic is more likely to be simple if a decimal rather than binary based code is used. For example, in the example described D_0 is reflective because the transition occurs at 360° which is an even multiple of ten.

Fig. 1. The logic required to convert a decimal Gray code for the digits 0 to 360 (g) into a BCD code (b). The elements on the left remove the ambiguity at the zero point ▶

TABLE 1

	Binary Gray	Decimal Gray
0	0 0000	0 0000
1	0 0001	0 0001
2	0 0011	0 0011
3	0 0010	0 0010
4	0 0110	0 0110
5	0 0111	0 0111
6	0 0101	0 0101
7	0 0100	0 0100
8	0 1100	0 1100
9	0 1101	0 1101
10	0 1111	1 1101
11	0 1110	1 1100
12	0 1010	1 0100
13	0 1011	1 0101
14	0 1001	1 0111
15	0 1000	1 0110
16	1 1000	1 0010
17	1 1101	1 0011
18	1 1111	1 0001
19	1 1110	1 0000

TABLE 2

	D_2		D_1				D_0			
	g_1^2	g_0^2	g_3^1	g_2^1	g_1^1	g_0^1	g_3^0	g_2^0	g_1^0	g_0^0
349	1	0	0	1	1	1	1	1	0	1
350	1	0	0	1	1	0	1	1	0	1
351	1	0	0	1	1	0	1	1	0	0
.....										
358	1	0	0	1	1	0	0	0	0	1
359	1	0	0	1	1	0	0	0	0	0
000	0	0	0	0	0	0	0	0	0	0
001	0	0	0	0	0	0	0	0	0	1
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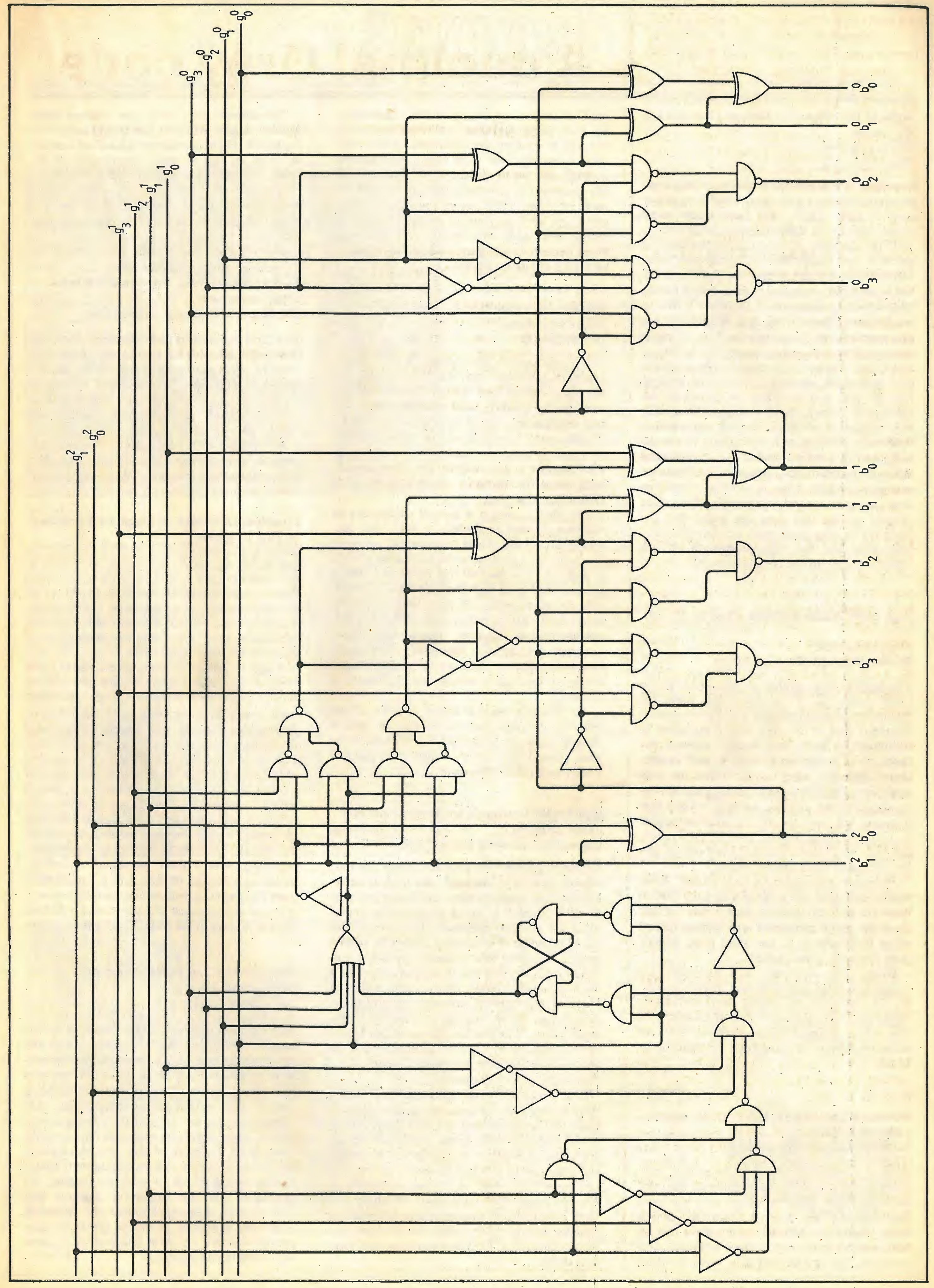
009	0	0	0	0	0	0	1	1	0	1
010	0	0	0	0	0	1	1	1	0	1



BOOKS

Myth of the Micro, by Rodney Dale and Ian Williamson
919pp, paperback.
Star, £1.50

Reasoned, restrained and responsible treatment of microprocessors and microcomputers is not easy to find. There have been many small books, with clever titles and remarkable covers, with the declared intention of explaining the micro to the many. Most of these have been successful in describing how the devices work, but in the main, uncritical in their crystal gazing and in their examination of the micro's part in our society. Messrs Dale and Williamson are also successful at explaining the ways in which chips work and what they are able to do, but in addressing themselves to micro's effects on the next generation, attempt to show that computers are still, and are likely to remain in the foreseeable future, hard-working idiots; that any suggestion of these devices approaching humans in intelligence is premature, to say the least; and that a great deal of the conventional wisdom on computers, industrial performance and jobs is based more on modern mythology than experience and observable fact. The "Myth" of the title does not mean that the authors are against microelectronic development, but that they disapprove of both computer worship and computerphobia.

High Performance Loudspeakers, by M. Colloms.
246pp, hardback.
Pentech Press, £8.95.

The loudspeaker, among all the links in the audio chain, still offers the widest scope for discussion, experimentation and measurement. Amateurs find it the most rewarding piece of equipment to build and modify, perhaps because of its deceptive simplicity, and professional engineers have not yet come up with perfection, for all the computer-assisted carpentry they now use. Indeed, it seems unlikely that they ever will, since the room in which a speaker is used effectively forms a part of the system, although Mr Colloms includes a few paragraphs on 'integrated room/speaker design'.

This book is intended for both groups. Amateurs should find in it most of what they need to know to work intelligently (and if they do not, there are useful references and bibliographies) while professionals in any field often benefit from a resumé of their subject.

Drive units, crossovers and enclosures are comprehensively dealt with and there is a chapter on methods of testing and evaluation, subjective and objective. This is the second edition, which includes additional information on several subjects. A paperback is available at £5.95.

Newnes Book of Video, Ed: K. G. Jackson.
128pp, paperback.
Newnes Technical Books, £5.95.

After 35 years of watching television programmes on a take-it-or-leave-it basis, the mass audience is not quite as captive as it was even five years or so ago. A tv set is now seen, by the more imaginative at least, as simply a display unit, which can be used to show television programmes, but which will also serve in other ways. Magnetic and disc recorders, computers, videotex, games machines – all these use a tv in one way or another, and this book, which consists of contributions from several well-known authors, sets out to explain to the layman what has happened to television since it changed its name to video. There are pieces on recording, on using a television camera with a recorder, on aerials and cable distribution, on videotex, home computers and games and a final chapter on how to spy on shoppers.

Videotex, by Roger Woolfe
184pp, hardback.
Heyden, £7.00

Few developments have given rise to such a tangle of offshoots and variations as has the television information system. The CCITT's name for the whole mass of systems in use or on trial is videotex, which is adopted by Roger Woolfe as the title for his book. This introductory text, intended for readers with a non-technical interest in the subject, is concerned only with interactive systems, typified by the Post Office Prestel service.

The first section is a general description of videotex systems as a whole – what they are, how they are used, where the information comes from and some guesses on the marketing of such systems. Prestel, the videotex used in the UK, is then oulined from both the users' and information-providers' points of view, with a look at costs and future prospects. Finally, developments in Europe, Japan and North America are briefly examined. The book concludes with a little crystal-gazing on the future of this type of system, and a glossary of terms.

Mr Woolfe's book is aimed squarely at the potential business user or participant, and it should prove extremely useful as a guide to the past, present and future of these new and still poorly understood systems.

The Art of Electronics, by Horowitz and Hill
716pp, paperback.
Cambridge University Press, £12.00

Although intended for possible college use and, indeed, so used at Harvard, this book is determinedly non-mathematical, and shows just how far it is possible to go in teaching electronics using the practical approach. The authors' declared intention is to conduct the reader from a state of innocence to a reasonable proficiency in circuit design. Since this is recommended for a one-year college course, the said reader is more likely to find himself following at a fast trot rather than being graciously conducted, but nevertheless, if he reaches the standard of competence implied by the content of the book, he will be more valuable than a great many, more theoretically based engineers.

It is impossible to list the contents here but there is very little the average engineer needs to know that is not covered in a varying amount of detail. Fundamentals, linear and digital design are treated in detail (with many practical circuits and examples of how not to do it), as are mini and micro computers. A section is devoted to h.f. techniques and high-speed switching and there is a chapter on measurements. Appendices contain, among other things, sections on relevant mathematics, i.c. types and some specimen data sheets.

Throughout the book, the authors have avoided jargon; when its use is necessary, it is explained. There is evidence of a certain amount of strain on this account (descriptions of RC filters, for example, without much more lavish reference to poles and zeros, must have been a problem) but this is a beautifully written and produced book, which can be highly recommended.

Hi-fi Year Book 1981, Ed: Kenneth Ellmore
256pp, paperback.
IPC Electrical Electronic Press, £3.00

This familar annual is now available, albeit in a new format and, for the second year, additional contents. Page size is increased to A4, and the subtitle "and Home Entertainment" recognizes the inclusion of five sections on organs, colour tv, radio/cassette recorders, games and video cassette recorders. All the information on high-fidelity equipment traditionally presented is also there, in directory form, and there are a number of articles on developments in home electronics.

Complete Handbook of Magnetic Recording, by Finn Jorgensen.
448pp, paperback.
Tab Books, $10.95

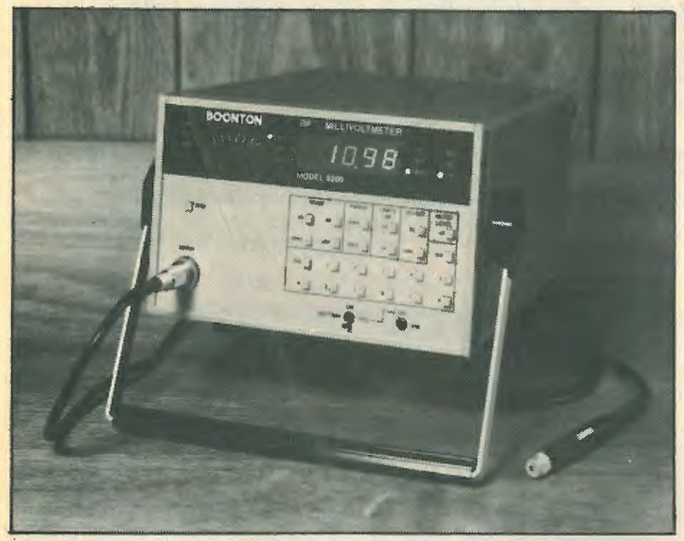
The title of this book seems, at first sight, to be an optimistic one for a paperback. On investigation, however, it is difficult to see how a book of this nature could be called anything else: it lives up to the title remarkably well.

A first couple of chapters gives a fairly brisk view of the subject as it is now and how it arrived at its current position, after which the book covers the whole of magnetic recording in three stages. Firstly, basic magnetic theory and the mechanics of recording and playback from tape are examined in some detail. As in the rest of the book, mathematics are used here only *in extremis* and could probably be ignored by determined innumerates. Two comprehensive chapters on heads, tapes and discs then follow, and the remaining six chapters are concerned with the design and use of recording equipment, beginning with a piece on digital recording, including a section on f.m. and p.c.m. techniques and finishing with advice on maintenance. A useful nine-page list of suggestions for further reading is given at the end of the book.

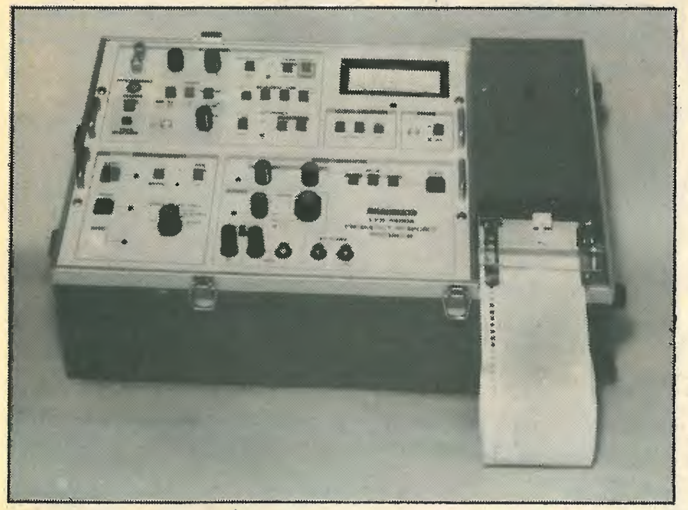
Early Wireless, by Tony Constable
160pp, hardback.
Midas Books, £8.50.

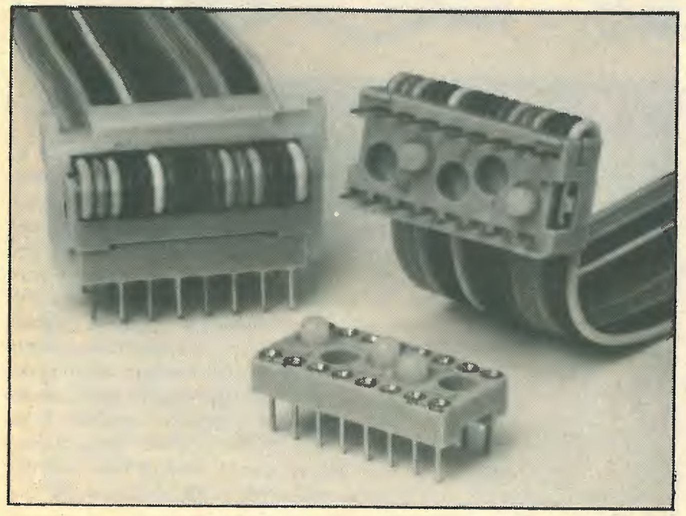
Not, perhaps, quite as comprehensive as its title implies, the little book is concerned with the development of domestic receivers from crystal-and-headphone times to the 1930s and is written with the interests of the collector in mind, being one of a collector's series from Midas. Mr Constable mentions most of the developments, people, places and companies that have contributed to the evolution of the wireless receiver, and the book is very well illustrated with photographs and drawings of early equipment. An appendix lists the important features and original price of several hundred sets. Although somewhat expensive at £8.50, the book does provide a single source of information on some very collectable equipment.

NEW PRODUCTS



WW301


WW302


WW303

R.f. millivoltmeter

Voltages from 200μV to 3V over the frequency range 10kHz to 1.2GHz can be measured using the model 9200 r.f. millivoltmeter from Boonton Electronics. This meter, claimed to be the first commercially available with microprocessor control, has automatic ranging and zeroing and can give readings in either mV, dBmV, dBV or dB relative to an arbitrary reference via its four digit display. An optional second input channel allows two voltage probes to be used. The instrument will then display either channel or their instantaneous difference in dB to simplify the measurement of voltage gain or loss. Outputs from the 9200 include a 0V to 10V linear signal proportional to the input signal level for driving pen recorders etc. and t.t.l. outputs for the high and low voltage limits which can be set to any desired level via the programming pad. Out of limit conditions are also indicated on the front panel. The digital readout is supplemented by an analogue panel meter to aid peaking and nulling. Other options include a rechargeable battery and a field installable IEEE-488 bus interface. A range of accessories is available such as a 100:1 voltage divider. Euro Electronic Instruments Ltd, Shirley House, 27 Camden Rd, London NW1 1YE.
WW301

A.f. response recorder

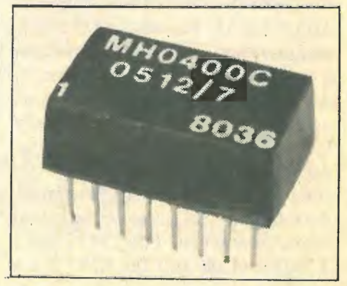
Frequency responses of various audio equipments can be plotted on chart paper using the Leader LFR5600A or LFR5601 frequency response recorders from Sinclair Electronics Ltd. Each instrument contains a sweep oscillator running from 20Hz to 30kHz and a pen recorder with four chart speeds which can both be operated independently. These recorders can be used to measure frequency response, wow and flutter and, since the instrument can also be used as a d.c. recorder, voltage and temperature drift parameters. They can also be used in conjunction with a speaker analyzer or an equalizer amplifier for cartridge response etc. Standard signal frequencies of 1kHz and 333Hz may be selected for magnetic-tape recorder measurements. The 5600A uses 50mm chart paper (linear or log. scale) and costs £1395 + v.a.t. The 5601 can use either 50 or 100mm paper and costs £1449 + v.a.t. Sinclair Electronics Ltd, London Rd, St Ives, Huntingdon, Cambs PE17 4HJ.
WW302

Ribbon cable connectors

Up to 20 identical d.i.l. plug/socket combinations can be made non-interchangeable by use of the Siemens BK-DIL 368/390 ribbon cable connector system. Each plug and socket has a number of positions into which 'polarization posts' may be inserted by the user. If any of these posts coincide on both plug and socket, the two cannot be joined. The plug is fitted to the cable by the insulation displacement method and is only 8.2mm high with strain relief fitted. Both plug (BK-DIL368) and socket (BK-DIL390) are available in 14, 16, 18, 24 and 20 pole d.i.l. versions. Plugs are available with either gold or tin plated pins. Siemens Ltd, Siemens House, Windmill Rd, Sunbury-on-Thames, Middx TW16 7HS.
WW303

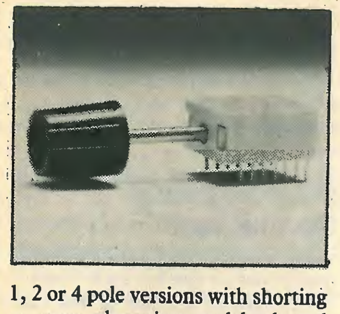
Active filters

Hybrid active filters which allow the customer to specify the frequency response, and which offer precise adjustment of cut off frequency, small size and low cost, are available from Menvier Hybrids Ltd. These 4th order filters are available in either low or high-pass types. Each module is trimmed during manufacture and requires no external frequency determining components. Customers can specify the cut-off frequency in the range 50Hz to 5kHz and select either Butterworth, Chebyshev or Bessel response approximation. Amplitude and phase response errors are less than 0.2dB and one degree respectively. Samples with either 16 pin d.i.l. or 8 pin s.i.l. packages can be supplied within two weeks of ordering. Higher order filters can be supplied. Menvier Hybrids Ltd, Southam Rd, Banbury, Oxon OX16 7RX.
WW304



D.i.l. rotary switch

Contact ratings of the FEME 5940 d.i.l. 12-position rotary switch are 100mA, 120V a.c./d.c. and 20mΩ maximum resistance. These flat-pack switches are available from Quiller Components Ltd in either



1, 2 or 4 pole versions with shorting or non-shorting gold plated contacts. The 5940 switch is said to be suitable for aplications involving automatic soldering and ultrasonic cleaning and is claimed to have a life of 50,000 operations when switching low power levels. Quiller Components Ltd, St Leonards Rd, Bournemouth BH8 8PA.

WW305

Encoder/decoder

With a single 5V supply, power consumption of the HD6409 c.m.o.s. 'Manchester' encoder/decoder is typically 5mA. Harris Semiconductor, the manufacturers, say that the device has a performance of 1Mbit/sec from −40 to +85°C. The 'Manchester' encoder/decoder is intended for low-noise, high-speed serial data communication and eliminates d.c. and l.f. components while allowing for clock recovery from the received signal. In 100+ quantities the 0.3in wide 20 pin d.i.p. version costs £4.73 per unit. Harris Systems Ltd, Semiconductor division, 145 Farnham Rd, Slough, Berks.

WW306

Microcomputer boards

Any combination of standard 24-pin compatible memory devices can be accepted by the latest addition to the Mostek MDX Z80 range of microcomputer boards via six sockets. The Z80 based MDX-CPU 2, from VSI Electronics Ltd, can be strapped to accept any combination of pin compatible r.a.ms, r.o.ms and e.p.r.o.ms and a decoder p.r.o.m., supplied with the board, programmed to implement them. Flexible memory decoding enables the memory device to be configured within any 1K boundary of the 64K memory map. The CPU 2 contains a 4-channel counter/timer circuit, the trigger inputs and zero count outputs of which are externally accessible. For long counting sequences, the four counter/timer circuits can be cascaded. Address, data and control buses of the board are bi-directional. The CPU 2 has a 2.5MHz clock and the CPU 2-4 a 4MHz clock. VSI Electronics (UK) Ltd, Roydonbury Industrial Park, Horsecroft Rd, Harlow, Essex CM19 5BY.

WW307

Digital multimeter

Capacitance, temperature and conductance ranges are included on this touch-operated digital multimeter from Non Linear Systems Inc. The Touch Test 20 has a 0.55in high 3½ digit l.e.d. display and can measure 10 parameters, 20 functions on 44 ranges. Audible continuity and diode test facilities are included. Some measuring limits of the multimeter are 10µV-1kV (0.2%) d.c., 10µV-750V r.m.s., 0.01µA-10A d.c., 10µA-10A a.c., 10mΩ-20MΩ resistance, 1pF-200µF capacitance, −40 to +150°C temperature and 0.01nS-1.999nS conductance. The mains-only version of the TT20 measures 2.9×6.3×7.5in, weighs less than 3lb and costs £195. A version with rechargeable batteries and charger is available for around £20 extra. Lawtronics Ltd, 139 High St, Edenbridge, Kent TN8 5AX.

WW308

P.c.b. guillotine

Designed for cutting p.c.bs and laminates, the Circuitape guillotine can handle workpiece widths of up to 16½in and incorporates an acrylic safety-guard. Both imperial and metric scales are attached to the cast baseplate of the tool and an adjustable stop is provided for use either in front or behind the cutting edge. Circuitape Ltd, New St, Aylesbury, Bucks HP20 2NL.

WW309

8-bit d.a.s.

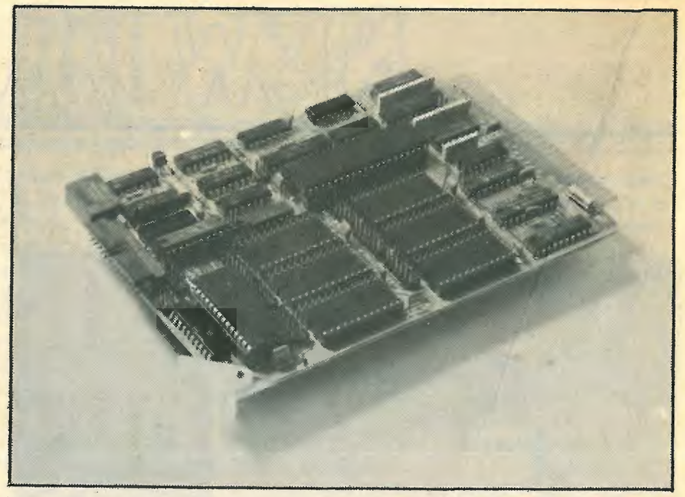
Power dissipation of the AD7583 9-channel 8-bit data acquisition system is 250mW maximum, making it suitable for battery powered remote data acquisition applications. This c.m.o.s. device from Analog Devices converts the addressed channel's analogue input to an equivalent 8-bit word in 4ms with an error of 1 count maximum over the temperature range −25 to +85°C and with supply voltage variations of +12V to +15V. Supply requirements are +5 and +15V. The device offers t.t.l. or c.m.o.s. compatibility, requires six passive components and two op-amps for operation and can be interfaced directly to microprocessors. Analog Devices Ltd, Central Avenue, East Molesey, Surrey KT8 0SN.

WW310

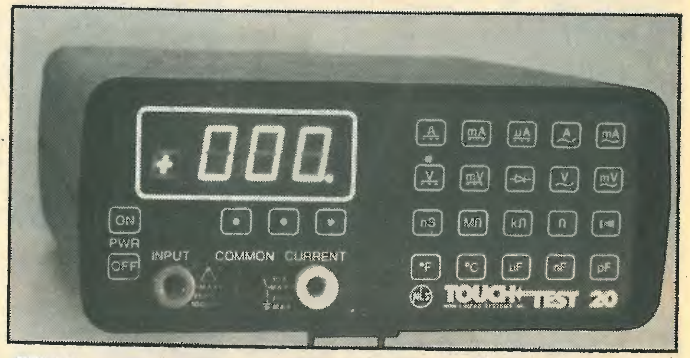
Pulse relay

The type RPC pulse-operated hybrid relay switches for a pre-set period, adjustable between 0.1 and 3 seconds, when an input pulse is received. This unit, from Appliance Components Ltd, comprises a power relay, a monostable multivibrator and an adjustable timer and is intended for use in delay, control and sequencing circuits for machinery and to replace more complex devices in pulse output circuits. Four basic models are available to cater for operating voltages from 100V to 240V a.c., 24V to 110V d.c. The d.p.d.t. relay contacts are rated at 7A (resistive), 240V. Life expectancy of the relay is said to be over 10 million operations. Appliance Components Ltd, Cordwallis St, Maidenhead, Berks SL6 7BQ.

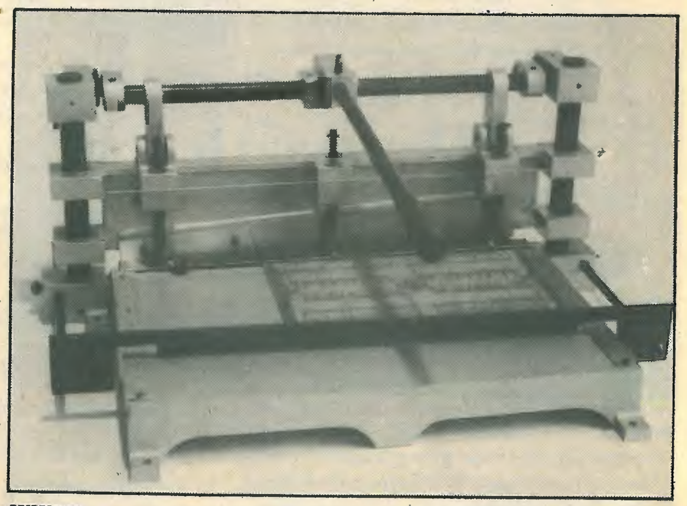
WW311



WW307



WW308



WW309

H.v. switchmode transistors

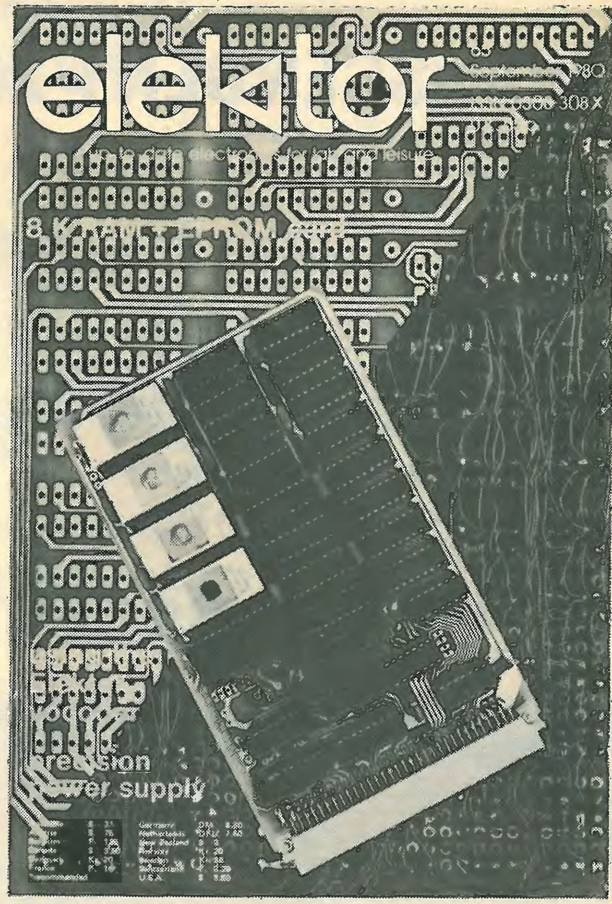
Extensions have been made to Motorola's MJE1300 fast switching transistor range to include the MJE13003A, 5A, 7A and 9A versions with maximum collector currents of 1.5A, 4A, 8A and 12A respectively. Voltage ratings of 400V (V_{ceo}) and 850V (V_{cev}) are quoted for these glass passivated plastic power transistors which are suitable for use in flyback type switchmode power supplies. Motorola Ltd, Semiconductor Products Division, York House, Empire Way, Wembley, Middx HA9 0PR.

WW312

WW—033 FOR FURTHER DETAILS

elektor

PRINTED CIRCUIT BOARDS



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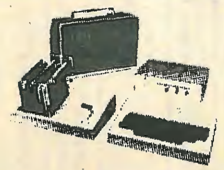

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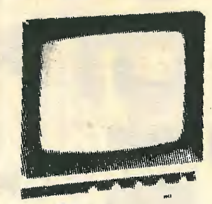

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Model	Output Power RMS	Distortion Typical at 1KHz	Slew Rate	Rise Time	Signal/Noise Ratio DIN AUDIO	Price & VAT
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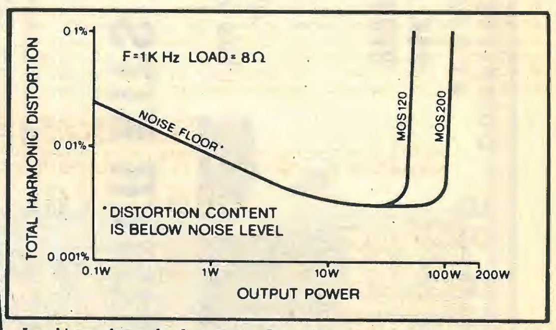

Load impedance both models 4Ω-∞ Input sensitivity both models 500mV
Input impedance both models 100KΩ Frequency response both models 15Hz-100KHz - 3dB

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(Standard O-P Transistors)

CHOOSE AN I.L.P BIPOLAR POWER AMP where power and price are first consideration while maintaining optimum performance with hi-fi quality and wide choice of models. From domestic hi-fi to disco and P.A., for instrument amplification, there is an I.L.P Bipolar to fill the bill, and as with our new Mosfets, we have encapsulated Bipolars within our New Profile extrusions with their computer-verified thermal efficiency and improved mounting shoulders. Connections are simple, via five pins on the underside and with our newest pre-amps and power supply units, it becomes easier than ever to have a system layout housed the way you want it.

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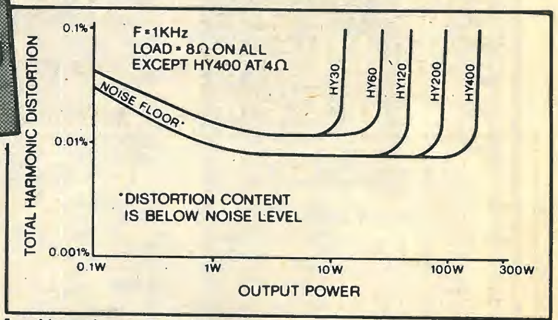
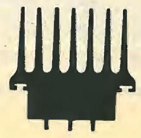
HY30	15W into 4-8Ω	0.015%	15V/μs	5μs	100dB	£7.29 + £1.09
HY60	30W into 4-8Ω	0.015%	15V/μs	5μs	100dB	£8.33 + £1.25
HY120	60W into 4-8Ω	0.01%	15V/μs	5μs	100dB	£17.48 + £2.62
HY200	120W into 4-8Ω	0.01%	15V/μs	5μs	100dB	£21.21 + £3.18
HY400	240W into 4Ω	0.01%	15V/μs	5μs	100dB	£31.83 + £4.77



Load impedance all models 4Ω-∞ Input sensitivity all models 500mV
Input impedance all models 100KΩ Frequency response all models 15Hz-50KHz - 3dB



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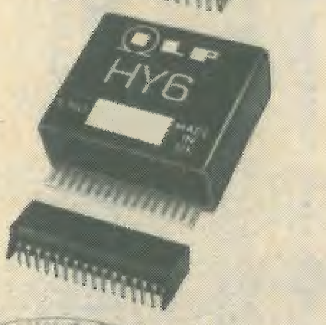

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HY6 (mono) and **HY66** (stereo) are new to I.L.P.'s range of advanced audio modules. Their improved characteristics and styling ensure their being compatible with all I.L.P power-amps both MOSFET and BIPOLAR, giving you chance to get the best possible reproduction from your equipment. HY6 and HY66 pre-amps are protected against short circuit and wrong polarity. Full assembly instructions are provided. Mounting boards are available as below.

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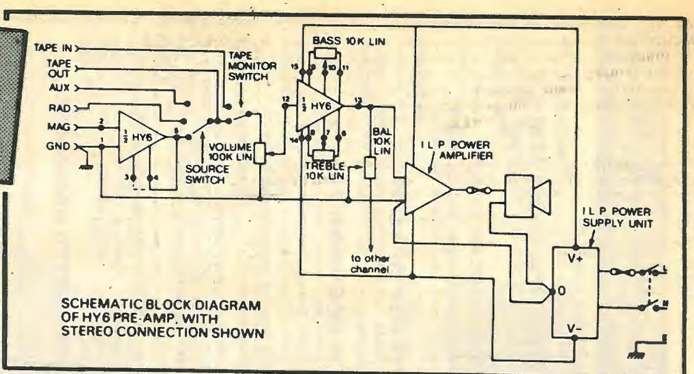
HY6 mono £6.44 + 97p VAT Connectors included
HY66 stereo £12.19 + £1.83 VAT Connectors included
B6 Mounting Board for one HY6 78p + 12p VAT
B66 Mounting Board for one HY66 99p + 15p VAT



SCHEMATIC BLOCK DIAGRAM OF HY6 PRE-AMP. WITH STEREO CONNECTION SHOWN

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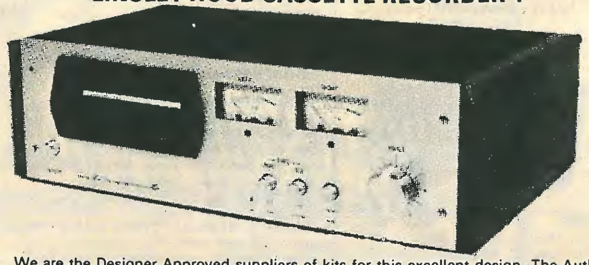
LINSLEY HOOD CASSETTE RECORDER 1



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 new HS15 Sendust Alloy record/play head, available separately at **£7.60** plus VAT, but included FREE as part of the complete kit at **£75** plus VAT.
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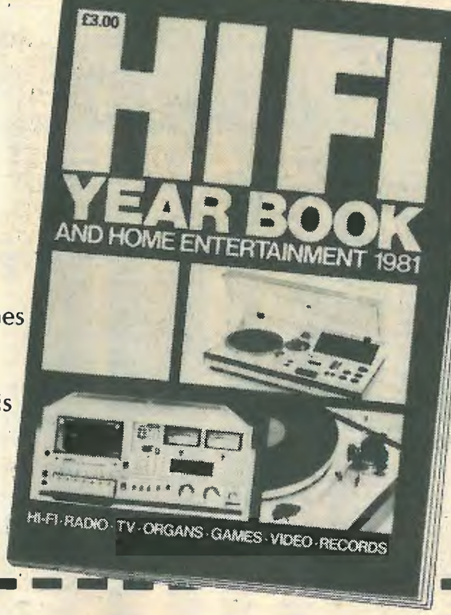
HI FI YEARBOOK AND HOME ENTERTAINMENT 1981



Published again in November, this new 1981 edition in larger magazine size means more comprehensive coverage of the whole range of home entertainment equipment, from aerials to headphones, from microphones to video recorders and from radios to electronic organs.
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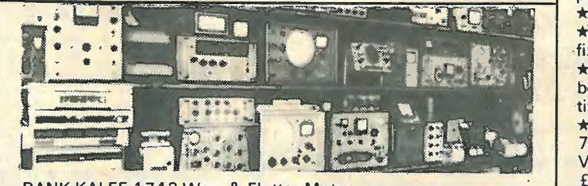

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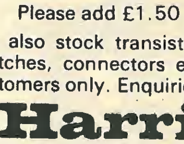

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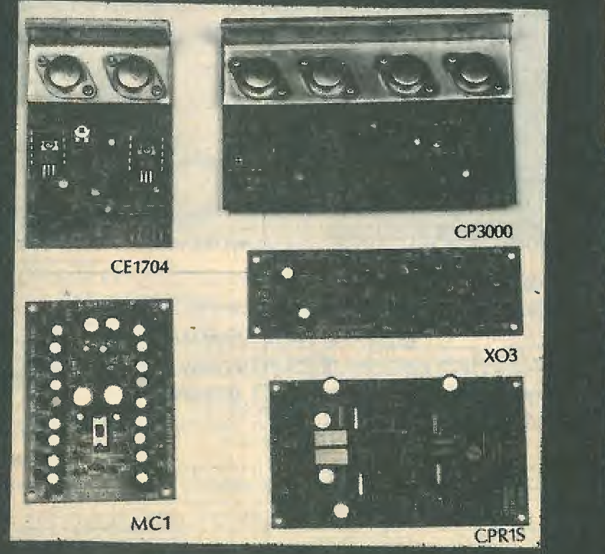

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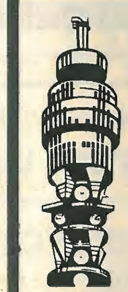

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(931)

Royal Marsden Hospital
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Medical Physics Technician Grade III

required in the Radiotherapy and Physics Electronics Workshop of the above hospital. The person appointed will work in a small group responsible for the maintenance of radiotherapy equipment, including three Cobalt units, a Philips 10 MeV Linear Accelerator and orothovoltage X-ray equipment. Applicants should have experience in electronics and in electrical and mechanical servicing.

They should hold ONC, HNC or similar qualification in electrical engineering or electronics and have at least 3 years, relevant technical experience to obtain salary on scale £5750-£7277.

Application form and job description available from Miss E. M. Bewley, Group Personnel Officer, Royal Marsden Hospital. Tel: 01-352 8171, Ext. 446.

(950)

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(938)

UNIVERSITY OF SURREY

Electronics Technician

Our Department of Electronic & Electrical Engineering seek a Grade 6 Technician to assist with the design, development, production and maintenance of a wide range of electronic equipment. Current projects include control and measurement modules for the UOSAT satellite to be launched in 1981, underground communication systems, instrumentation for ion implantation research and various applied microprocessor systems. Applicants should have broad based practical experience and HNC equivalent qualification.

The University offers good working conditions and environment, and facilities for a wide range of sports and social activities. Salary is on scale £6020 to £7191 per annum. For further details or to arrange a visit contact Dennis Pollard (Guildford 71281 ext. 392), or write to the Staff Officer, University of Surrey, Guildford GU2 5XH, quoting reference (8WW).

(920)

LEICESTERSHIRE AREA DEPARTMENT OF MEDICAL PHYSICS AND CLINICAL ENGINEERING

MEDICAL PHYSICS TECHNICIAN III/IV

(Clinical Measurements)

Applications are invited from suitably qualified persons for the above post based at the Leicester General Hospital.

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i) Urodynamic Measurements
ii) Bloodflow Studies
iii) Audiometric Electrodiagnosis
iv) Ophthalmic Electrodiagnosis

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For appointment at Grade III, three years' experience is required in at least two of the above categories, preferably i) and ii).

Salary IV — £4404-£5790   Salary III — £5223 - £6750

New entrants to the N.H.S. will normally commence on the minimum point.
Further information may be obtained by telephoning Mr. S. Bentley, Leicester 541414 ext 469. Application forms are available from the Personnel Department, Leicester General Hospital, Gwendolen Road, Leicester. Tel: (0533) 730222.
Closing date: 3rd March, 1981.

(917)

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ENFIELD

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(945)

UNIVERSITY OF SURREY

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Foreign and Commonwealth Office (936)

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(913)

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Department of Aeronautical Engineering

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Department of Civil Engineering

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Written applications, giving details of age, qualifications, experience, present salary, together with copies of any formal qualifications held and the names and addresses of two referees to Professor E. L. Blair, Department of Physiology, Medical School, The University, Newcastle-upon-Tyne NE1 7RU.

(918)

NORTH MIDDLESEX HOSPITAL,
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LONDON N18 1QX

MEDICAL PHYSICS TECHNICIAN GRADE III/IV

The Medical Physics department at the above hospital is responsible for providing an electronic maintenance service for the Haringey District Hospitals. The work load includes maintenance of equipment used in Intensive Care Unit, theatres, labour wards together with Radiotherapy and Nuclear Medicine equipment. Some development work is also undertaken.

The successful applicant should be experienced in electronic maintenance and some experience of design work would be an advantage.

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For further details contact Mr D. W. Bailey (Senior Physicist) on 01-807 3071 Ext. 678.

Application forms obtainable from The Personnel Office at North Middlesex Hospital, Sterling Way, Edmonton, London N18 1QX Ext. 571.

(955)

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MARCONI TF2333. M.F. transmission measuring set, as new, £250.

MARCONI TF801D/88, £150.

MARCONI R.F. RADIATION & POWER METER. OA1430 (CT477), as new, in grey metal case with full manual, power meter FX range, 10mc/s to 10gc/s, complete with X-S-L band aerials, £50.

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MUIRHEAD FACSIMILE EQPT. TX D901 D/S £50. RX D900 D/S or P/D £75 each.

WEATHER SATELLITE. RX D900 S/1A or S/2A £125 each with manual.

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EXTEL TRANSTEL MATRIX PRINTERS. 5 level Baudot Code. Accepts speeds up to 300 bauds. Supplied set to 50 and 75 bauds switched. Tested with manual, £165.

All items are bought direct from H.M. Government, being surplus eqpt. Price is ex-works. S.A.E. all enquiries. Phone for appointment for demonstration of any item. **JOHNS RADIO, 424 BRADFORD ROAD, BATLEY, YORKS. TEL: 0924 478159** (9.30 a.m.-1 p.m.).

(848)

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Products that help you make a better job of it.

Arax Multicore Solder.
Economy pack for general non-electrical use. Replaces solid wire and stick solder. (B.S. 219 Grade L). Econopak 200g reel of 3mm dia. Size 16A.
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Multicore 5-core solder for radio, T.V. and similar work. Reduces copper erosion. Suitable for service engineers and manufacturers using small quantities of solder. 1.2mm dia. Size 12. **£3.91 per reel.**

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Multicore solder-wick for removing solder from virtually any joint.
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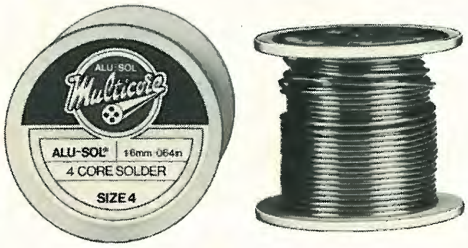

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Alu-Sol Multicore 4-core solder for soldering most types of aluminium. No extra flux needed.
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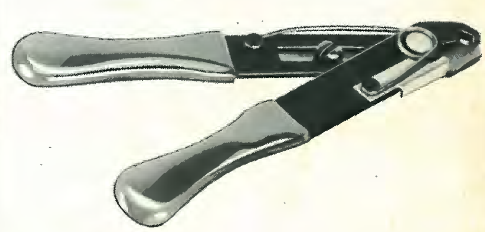

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Bib Hi-Fi Accessories Ltd., (Solder Division),
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