wireles world

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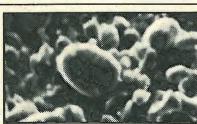
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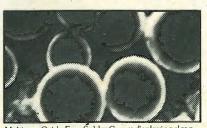


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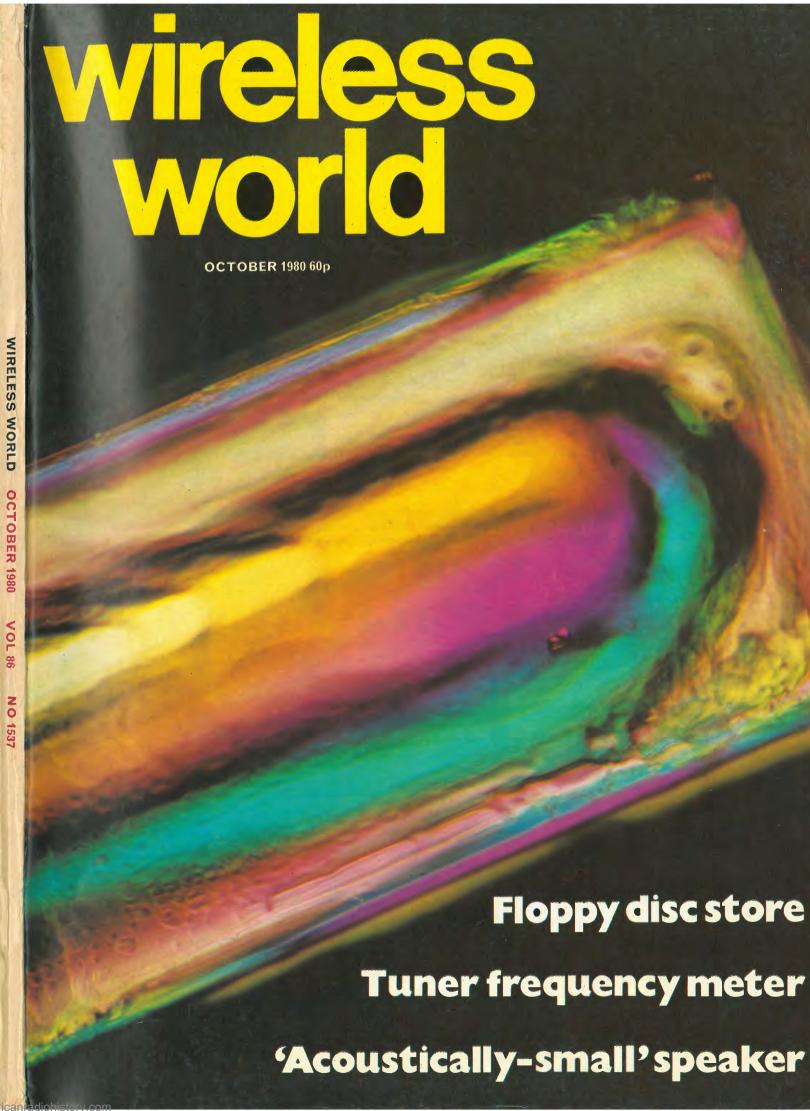
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Editorial & Advertising offices:

Dorset House, Stamford Street, London SE1 9LU.

Telephones: Editorial 01-261 8620. Advertising 01-261 8339. Telegrams/Telex: Wiworld Bisnespres 25137 BISPRS G. Cables Ethaworld, London SE1.

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USA mailing agents: Expediters of the Printed Word Ltd, 527 Madison Avenue, Suite 1217. New York, NY 10022. 2nd-class postage paid at

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wireless

ELECTRONICS/TELEVISION/RADIO/AUDIO

OCTOBER 1980 Vol 86 No 1537

35 Personal hygiene or public health?

36 Radio tuner frequency counter by J. L. Linsley Hood

42 Tone filters for electronic organs by C. E. Pykett

46 Floppy disc system for the scientific computer by J. H. Adams

49 Letters to the editor

V.h.f. programme labelling tests Designing with microprocessors Maxwell's equations revisited

> 52 Novatexts — Pulse control of analogue functions by F. Williams

> > 54 The 'Twins' paradox of relativity by Herbert Dingle

> > > 57 Audio gain controls by Peter Baxandall

65 An acoustically small loudspeaker by B. I. Harcourt

C.b. on 928MHz

68 News of the month Postal monopoly

Satellite television

71 Designing with microprocessors by D. Zissos and L. Valan

74 World of amateur radio

77 Satellite broadcasting in the eighties by G. J. Phillips

> 82 The floating bridge by R. M. Brady

> > 87 New products

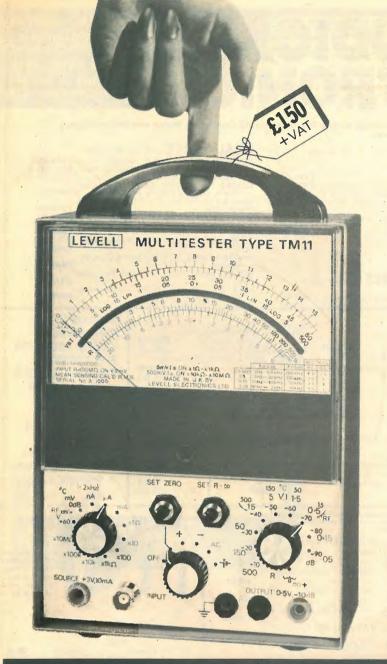


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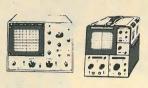
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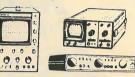
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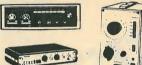
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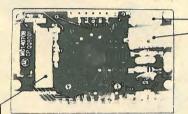
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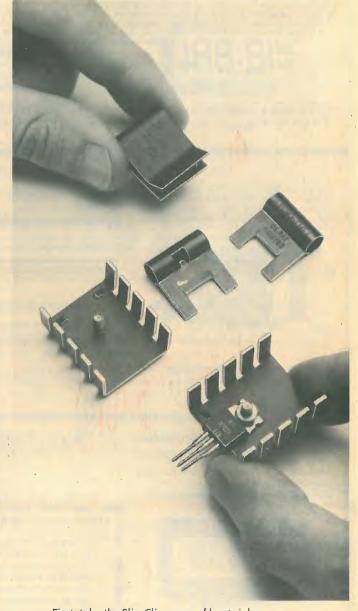
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WIRELESS WORLD, OCTOBER 1980



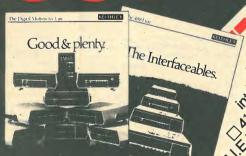


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HEWLETT PACKARD 8640B Precision AM-FM Signal Generator 141T Spectrum Analyzer — Mainframe 8552B Spectrum Analyzar — IF Section 8553B Spectrum Analyzer — RF Section 8555A Spectru

8556A Spectro 1600A 16 Cha PHILIPS PM 3212 Dua

Oscilloscope PM 3214 Du Oscilloscope

m Analyzer — IF Section um Analyzer — RF Section um Analyzer — RF Section	£1650 £2150	7A22 Differential 1 (12 month guarantee)
al Trace 25 MHz 2mV/Div	£525	10µV-10V/DIVACE Plug-in. 7A26 Dual Trace Plug-in. DC-150 MHz 5mV-5V/Div. 7B53A Dual Timebase 5ns-5s/Div CRT Read
e ual Timebase Dual Trace DC-25 MHz	£625	7853A Dual Tiffeboos
e .		

£3550

			200	
		Cable Test Equipment	Prices	PHILIPS
		MARCONI	from £	PM5127, O.1 Hz-1 MHz, Sine/
		TF2333 Transmission Test set	575	Square/Triangular/Pulse outputs.
			3/3	External sweep facility 30Vp. p max
		HEWLETT PACKARD		output
A	Prices	3556A For psophometric		
Acoustic/Vibration	from £	measurements from 20 Hz-20kHz.		Logic Analysers
BRUEL & KJAER		0.1mV-30V input level	475	HEWLETT PACKARD
1613 Octave filter set 31.5 Hz		NEC		1601L Logic state analyser
31.5 kHz	.250	TTS 37B. Noise, level and VU		12 channel display
2203 Precision sound level meter	400	measurement. Sensitivity -80dBm		1600A 16 channel 20 MHz clock
2608 Measuring Amplifier. Hi pass	400	up to + 20dBm	275	MAP A & B store
	375	STC		1607 16 channel 20 MHz clock
filter			240	(Display scope required)
4135 Microphone	60	74216A Noise Generator CCITT		TEKTRONIX '
4220 Pistonphone calibrator	175	74261A Psophometer CCITT	475	7D01F 16 channel up to 50 MHz
ZR0020 Integrator Converts 2203 &		WANDEL u. GOLTERMANN		clock MAP
09 for vibration meas.	56	DLM 1. Send/receive system for		
ZR0001/2/4/5/6 Attenuators	52	measuring phase jitter random noise		Mains Monitors
4424 Noise Dose meter built in mic.	250	and frequency shift on data		COLE
2215 Prec. Sound Level Meter A&C		transmission lines	1500	T1007 200-260V. 35-65 Hz
Oct. Filter	1095	LDS-2. 200Hz-600kHz sender for		Thresholds 10V, 50V, 100V, 200V
2218 Integrating Sound Level Meter		measuring group delay and		
A/LIN	1525	attenuation variations	3250	DATALAB
4230 Sound Level Calibrator for 1"		- LDEF-2. Filters for DLM unit	250	DL019 Power line interface for
and ½" mics	95	Counter Timers		transient recording
CFL .		HEWLETT PACKARD		DL905 Digital Storage Unit DC-3
112 LEQ meter digital readout	450		040	MHz 10mV
144 LEQ Meter 72dB Dynamic Range		5300A /5303B DC 520 MHz 6 digits	210	DRANETZ
	1005	5300A Display Module. 6 Digits.		606-3 Disturbance Analyser Avg,
Batt Optd	1325	3 × 10 ⁷	. 90	Sag/Surge
DAWE		5300B Display Module. 8 Digits. 2 × 10 ⁸		GAY
1461. CV(M) Portable Vibration			140	LDM Records + ve/ - ve transients
Analyser Kit	350	5302A DC-50 MHz. 100mV sens.		of 50ns on AC or DC Lines
Accelerometers		Time interval. Period. Ratio.		
		Totalise.	75	Modulation Meters
BRUEL & KJAER		5303B DC-525 MHz. (Plug-on)	200	AIRMEC
4343S Uni-gain type. Gen Purpose	55	125mV sens. 50Ω	120	210 1-300 MHz. AM/FM
4338S Uni-gain High sens med	440	5308A 0-75 MHz. Universal Module.		409 3-1500 MHz. AM/FM
frequency	110	50mV sens. 1MΩ	100	MARCONI
NARDA		5267A Time Interval Plug-in 10ns	120	TF2300A 1-1000 MHz. AM/FM
3044B-20 Dir. Coupler 3.7-8.2 GHz		5345 DC-500 MHz Time Int. Ave.		Multimeters-
17dB	155	Burst Total Ratio	1225	
Attenuators		10590A Adaptor converts 5245		Analogue
MARCONI SANDERS		Plug-ins to 5345	225	AVO
	470		223	8MKIII AC/DC V.AC/DC Amps.
6593 VSWR Indicator, Batt/Mains	175	RACAL		OHMS
Bridges		835. DC-15 MHz 6 digits		
CINTEL		Time interval/Period/Ratio	100	Oscilloscopes
277 Measures iron core inductances		9024 10 Hz-600 MHz 7 + 1 digits	250	ADVANCE
0.01H-1000H (with a Q value not		'9835 6 Digit DC-20 MHz 10mV	100	OS1000A DC-20 MHz. dual trace
less than 2)	- 130	9837 DC-80 MHz 6 digits	130	3300B Dual Trace DC-50 MHz
DAWE		S.E. LABORATORIES		5mV/div. Dual Timebase
		SM202 DC 150MHz. 8 Digits.		COSSOR
210B Decade Capacitance box	20	50Mv. A, B, C, Input. Time Interval		3100 Dual Trace DC-40 MHz
0.1µF-1mF 0.1 µf step	20	and Totalise	220	DYNAMCO
HEWLETT PACKARD			220	7210, DC-15 MHz, Dual Trace 1 mV
4342A 'Q' Meter QLC complete	1250	SYSTRON DONNER		
MARCONI		6053 9 Digit 20 Hz-3 GHz BCD O/P	850	sensitivity on CHI. Delayed Timebase
		6054A/04 11 Digit 20 kHz-18 GHz BCD O/P	2000	And the second second
TF1245 'Q' meter. Freq. range 1kHz-	350		2800	GOULD ADVANCE
300MHz using external osc.		Function Generators		OS1000B DC-20 MHz Dual Trace
TF868A Universal Bridge	250	HEWLETT PACKARD		X-Y TV Sync
-TF1313A Universal LCR Bridge 0.1%	375			HEWLETT PACKARD
WAYNE KERR		3310 0.0005 Hz 5 MHz. Multi-Mode.	050	1703A Storage 1000Div/ms.
B224 Wide range LCR Bridge	475	10V/50Ω sine, square, triangular	250	DC-35 MHz. Dual trace Mains/Ext
B500 Log LCR Bridge	225	INTER-STATE		DC
B601 RF LCR Bridge	125	ELECTRONICS		1707B/020 DC-75 MHz. Dual trace.
(Detector and Oscillator not incl).		F51A Multi-Mode. + and offset:		Dual Time Base.
B641. Measures L/C/R/G Accuracy		0.0005 Hz to 10 MHz 10/15V/50Ω	250	1707B/012 As 1707B/020 with
of 0.1%	450	FEEA Mark: Manda O COOF Har 10	200	Internal Devices Consider

F55A Multi-Mode, 0.0025 Hz-10 MHz. 10V/50Ω. Ext. VGC. Burst

O/P up to 100k bursts/sec

Q801. Y parameter test set. Plus

		Prices
	182C DC-100 MHz Mainframe, large screen	from £
	MEDELEC	020
	M-scope 4 channel DC-100 kHz U/V Chart	1225
iring	PHILIPS	425
	PM 3211 DC-15 MHz Dual Trace 2mV PM3233 Dual Beam DC-10 MHz	
	2mV/div. SCOPEX	400
	40-10B Dual Trace DC-10 MHz	180
3	TEKTRONIX 475 Dual Trace DC-200 MHz 2mV	1125
	485 Dual Trace DC-350 MHz 50Ω 1 MΩ 250 MHz	2100
75	556/1A1. True dual beam. DC-50 MHz. Can display 2 separate	
3/3	signals at different sweep rates.	700
205	Includes trolley 545B/1A1, DC-30 MHz, dual trace.	700
995	Delayed timebase 561A/3A6/3B1, DC-10 MHz, Dual	325
699	Trace. High persistence tube.	0.775
	Delayed Timebase 585A/82. DC-80 MHz. dual trace	275
1900	10 mV sensitivity 547/1A1, DC-50 MHz, dual trace	525
0070	DTB	525
£670	547/1A4. DC-50 MHz. four trace DTB	625
£780	7403N DC-60 MHz 3 Plug-in Mainframe	450
£650	7704A DC-200 MHz. CRT Readout.	
	Mainframe for 4 Plug-in TELEQUIPMENT	1200
	D53A. DC-25 MHz. dual trace.	
	10mV sensitivity with C-2 plug-in DC-15 MHz with JD plug-in	250
	D63/V1/V3 DC-35 MHz. Depending	
Prices	on sensitivity, 50 µV or 1 mV Sensitivity	675
from £	D34 Dual Trace DC-15 MHz 2mV Mains/Batt	525
	D75 Dual Trace DC-50 MHz Dual	
325	Oscilloscope Plug-ins	600
	HEIMI ETT DACKARD	
	1804A DC-50 MHz Four channel 20 mV-10V/div.	575
250	1825A Dual Timebase 50ns-1s/div. 1805A Dual Trace DC-100 MHz 5mV.	525
1850	1ΜΩ/50Ω	550
1500	TEKTRONIX Type R. Transistor R.T. tester. Pulse	
	rate 120 pulses/sec. R.T. Less than	
2650	5 mμs Type G. Differential amplifier, 100:1	100
	CMR DC-20 MHz. 50 mV sensitivity Plug-ins for 500 series	50
	1A1 dual trace Plug-in DC-50 MHz	225
75	1A2 dual trace Plug-in DC-50 MHz 1A4 four trace Plug-in DC-50 MHz	180 375
	1A5 Differential Plug-in Z Differential Plug-in	175 140
350	81 Adaptor Plug-in 1A Series to 580	
1055	Series 7A12 Dual Trace DC-105 MHz	75
	5mV/div. 7A18 Dual Trace DC-75 MHz	410
2625	5mV/div 7A22 High gain diff, amp.	370
	0.1 Hz-1 MHz 10µV 7A26 Dual Trace DC-150 MHz	450
1250	5mV-5V/div. 7B53A Dual Timebase 5ns-5s/div.	525
	For 7000 Series Mainframes	440
150	7B53AN as 7B53A Less CRT Readout	405
295	TELEQUIPMENT	.55
450	D67 DC-25 MHz. Dual trace. Dual Time Base. TV sync.	325
	D83 DC-50 MHz. Dual trace. Large	
	6½" CRT. Dual Time Base Oscilloscopes (storage)	650
	DYNAMCO	
60	7110. DC-30MHz. Dual trace. Writing speed 20 µs / Div.	525
	TEKTRONIX	
310	549/1A1. DC-30 MHz. 5mV sensitivity. Dual trace. Storage	
600	sensitivity. Dual trace. Storage scope, Writing speed: 5cm/µs with	0.00
400	enhancement. Includes trolley 564/3A74/3B4. DC-2MHz, four	675
400	channel. 20 mV sensitivity. Writing speed up to 500cm/ms	650
	564B/3A6/2B67, DC-10 MHz, Dual	0.00
300	trace 10mV sensitivity, split screen storage oscilloscope	750
	466 Storage 1350 cm/µs Variable	
400	Persist DC-100 MHz 7313 Split screen 4.9 cm/µs. DC-	2225
	25 MHz (M/F for 3 Plug-ins)	1650
,	TELEQUIPMENT DM64 Storage 250 Divs/ms.	
1200	DC-10 MHz Dual trace.	400
700	Phase Meter DRANETZ	
750	301A 5 Hz-500 kHz. Z in 100kΩ.	
650	Accuracy ±1° to ±2°. Analogue O/P	400

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Power Meters

Recent massive stock investment means-Today Carston Value makes even more sense

BRYANS SOUTHERN

Power Meters	Prices	BRYANS SOUTHERN	Prices	39.5 & 41.5 MHz
DYMAR	from £	29000 X-Y Recorder A4 0.25mV-	from £	Spectrum Analysers
2081/100 True RMS, DC-500 MHz.		10V/cm BS314 4 channel 1mV-10V	525	HEWLETT PACKARD
30mW-100W	425	16 speeds	1650	8443A Tracking Gene/counter
HEWLETT PACKARD		BS316 6 channel 1mV-10V	1030	100 kHz-110 MHz
432A 10µW-10mW, 10 MHz-10GHz	195	16 speeds	2350	8445A Automatic pre-selector
478A Thermistor Mount for 432A	40	29300 X-Y Single pen A4 0.25mV- 10V/cm 0.1s-50s/cm		10 MHz-18 GHz 8555A RF Plug-in 10 MHz-18 GHz
435A 0.3µW to 100mW 5 MHz-			545	1 kHz Res
18GHz	275	HEWLETT PACKARD		8558B For 180 Mainframe 100
8481A Power Sensor for 435A	225	680M. 5 inch. Stripchart Single Pen		kHz-1.5 GHz 1 kHz-res
MARCONI SANDERS		5mV-120V I/P 20cm/min 2.5 cm/Hr	275	3580A/001/002 Digital Storage
6460 10 MHz-40 GHz (Depending on		7046A Two pen A3 0.25mV-5V/cm	995	5 Hz-50 kHz. X-Y o/p
Head)	300	KUDELSKI		NELSON ROSS
6420 10 MHz-12.4 GHz 10mw	110	Nagra 4.2 LSP Professional Audio		011. DC-20 kHz. 80dB dynamic
6421 10 MHz-18.4 GHz 100mw	110	Recorder (Batt optd)	1215	range. Dispersion: 100 Hz-6 kHz
6422 10 MHz-12.4 GHz 1mw	85	NAGRA		022. DC-100 kHz. Dynamic range
6428 26.5-40 GHz 10mw	150	Mains Unit for 4.2 LSP	95	60dB fits into various 500 series
MARCONI		PHILIPS		CRO's
TF2512 DC-500 MHz 0.5-30w 50Ω	130	PM 8251 Single pen 10in chart		TEKTRONIX
TF 893A 10 Hz-20 kHz. 20µW-10W.	120	10mV-50V FS	450	3L5. Plug-in unit fits into various
Power Supplies		RACAL		500B series CRO's. 50 Hz-1 MHz.
		Store 4. Uses D/4 inch magnetic		Greater than 60dB dynamic range
ADVANCE		tape. Will record 4 F.M. channels.		Sweep Generators
1VI 12V DC to 240V 50 Hz, 150w		Operates at 7 different speeds.	1950	HEWLETT PACKARD
Inverter	125	S E LABORATORIES		8690B Mainframe. Int/Ext AM. Ext
BRANDENBURG	٠.	6150/6151 12 channel UV		FM
475R 10-2100V 5mA DC Stab.	150	1250 mm/s-25 mm/min 6 in chart	1400	8693B/100 3.7-8.3 GHz.5mW. PIN
FARNELL		994 6 Channel Pre-Amp ± 1% ± 1V		levelled 'N' connectors
L30B 0-30V 1A DC Stab.	55	o/p	450	8699B / 100 0.1-4 GHz.6mW. (20mW
FLUKE		6008 25 Channel µV 8 in 4m/sec to	000	to 2 GHz). PIN levelled. 'N'
415B 0±3100V 30mA 0.005% reg.		25mm/min	895	connectors
Protected *	350	SMITHS INDUSTRIES		TEXSCAN
ITT		RE541.20 Single Pen. 0.5mV-100V	1 444	9900 Sweep Generator 10-30 MHz
Power Lab. up to 30V Dual Supply	90	FSD. 3-60cm/min and hour	350	CRT Display
	30	YOKOGAWA		VS60 Sweep Generator 5-100 MHz
MARCONI		3046. 10 inch Chart Single Pen. 0.5		Rate 60 Hz
TF2154/1 0-30V 1A. 0±15V 2A 0±7.5V 4A	60	mV-100 VI/P2.60cm/min and/hr	350	LN40A Log Amplifier
a land and a second a second and a second and a second and a second and a second an	80	3047. 2 Pen Version of 3046	425	T.V. Test Equipment
SMITHS	-00	Signal Sources and		PHILIPS
4701 5-7V o/p Power Pack	32	Generators		PM5508B Pattern Generator, 625
SORENSEN				lines PAL. UK Systems
DCR 300-2.5 0-300V 2.5A DC Stab.	375	BOONTON		
Pulse Generators		102B 4.3-520 MHz Int/Ext FM/AM 0.1μV-1V 50Ω	1725	Voltmeters-Analogue
DB ELECTRONICS			1/25	AVO
150. I.C. pulse generator	50	DYMAR 1525 100 kHz-184 MHz Int/Ext		8 Mk IV
EH RESEARCH	•	AM/FM Batt/Mains	525	BOONTON
			525	92AD/01/09 10 kHz-1.2 GHz 1999
122, 1 KHz-200 MHz 5V/50Ω RT 12ns	220	GOULD ADVANCE SG70 5 Hz-125 kHz 600Ω 4w	85	FSD 10 _µ V Res
139(L). 10Hz-50 MHz 10V/50Ω	220		. 85	92C 10 kHz-1.2 GHz 500μV-3V. 1%
RT 5ns	175	HEWLETT PACKARD		of FS
1221. Timing Unit 6 Channel	175	204D 5 Hz-1.2 MHz. 600Ω. 80dB att.	470	BRADLEY
0-10 MHz 5V/50Ω RT 8ns	50	O/P 5V RMS	150	CT471C. AC/DC/Ω/current
G710. 5V/50Ω 30 Hz-50 MHz RT 5ns	100	204D/001 As for 204D (Battery operated)	175	multimeter and RF
132AL. 50V/50Ω 5 Hz-3 MHz		608E. 10-480 MHz AM	410	HEWLETT PACKARD
RT 12ns	175	620B 7-11 GHz 50Ω FM/PM 1mw		400E Millivoltmeter
HEWLETT PACKARD		8614A 800 MHz-2.4 GHz + 10dBm	1100	10 Hz-10 MHz B/W 1mV FSS
214A 100V/50Ω. Double pulse O/P.		to —127 dBm 50Ω AM/FM	1950	427A, AC DC/Ω multimeter
W50ns-10ms. 10 Hz-1 MHz. 15ns RT	350	8616A 1.8-4.5 GHz Ext AM/FM/PM		3406A. 10 kHz 1.2 GHz
MARCONI		10 mw	925	8405A Vector Voltmeter
TF2025 0.2 Hz-25 MHz ±10V/50V		MARCONI		1-1000 MHz B/W
RT 7ns	350	TF144 H/4S HF Generator		Auto Phase Lock 3400A 10 Hz-10 MHz 1mV-300V
PM5776 3V/50Ω. 1 Hz-100 Mz.		10 kHz-72 MHz AM	550	True RMS
Rise/fall Times less than 1ns.	275	TF791. FM Deviation Meter	95	
Recorders and Signal		TF801/D1. 10-470 MHz AM. FM.	255	KEITHLEY
Conditioning Equipment		TF995A/2. 1.5-220 MHz AM, FM.	350	610C Electrometer DC 1mV-100V,
AMPEX		TF2171 Digital Synchroniser for		Amps 10 ⁻¹⁴ Recorder o/p
		TF2015	525	LEVELL
PR2200 Instrumentation Recorder up to 16 channels. FM/DR. Record		TF2002/AS 10 kHz-72 MHz FM/AM	020	TM3B 5µV-500VAC 1 Hz-3 MHz +
replay all speeds. 1" tape FM/DR		0.1-1V o/p	625	50 to 100 dB
I.R.I.G. DC-40 kHz FM. 100 Hz-		TF2012 UHF, FM 400-520 MHz,		LINSTEAD
300 kHz DR	6500	0.03μV. Counter o/p	650	M2B. DC AC 10 Hz-500 kHz
BRUNO WOELKE		RACAL		MARCONI
ME102B. Wow and flutter meter	75	9081 5-520 MHz LED Display O/P -		TF2603. AC voltmeter to 1 5 GHz
ME102C. Wow and flutter meter	90	130dBm AM/FM	1875	PHILIPS
BRUEL & KJAER	50	ROHDE & SCHWARZ		PM2454B 1mV 300V. 10 Hz 12 MHz
2305B Bench type. Mains operated.		SWOB 11. 0.5-1200 MHz. 50Ω	850	Z in 19MΩ. DCO P
Log recording of AC: 2 Hz-200 kHz		SCHAFFNER		RACAL
and DC.50 or 100mm paper width.	750	NSG101 Mains Interference		9301 RMS Millivoltmeter
ZR0001 Linear Pat DC: 10-35 mV	59	Simulator, Superimposes Pulses on		10 kHz-1.5 GHz with carry case
ZR0002 Linear Pat DC: 10-33 11V	79	mains for testing immunity of		Voltmeters-Digital
ZR0004 25 dB Potentiometer	52	equipment to interference. Pulse		ADVANCE
ZR0005 50 dB Potentiometer	59	amplitude: ±800V. Rise Time 0.25µs.	200	DMM 7A/01 1999 FSD
ZR0006 75 dB Potentiometer	69	Width 50 & 200µs	300	AC/DC/Ω/Current
				FARNELL
				DM131B. 1999 FSD AC DC Ω
				Current Temperature
	_			

01-267 5311/2

Contact Brian Hollingsworth

or Noel Jennings

luumw	110	Recorder (Batt opto)	1215	range. Dispersion: 100 Hz-6 kHz	350
: 1mw	85	NAGRA		022. DC-100 kHz. Dynamic range	
٧ ,	150	Mains Unit for 4.2 LSP	95	60dB fits into various 500 series	
		PHILIPS		CRO's	350
5-30w 50Ω	130	PM 8251 Single pen 10in chart		TEKTRONIX	
20µW-10W.	120	10mV-50V FS	450	3L5. Plug-in unit fits into various	
Loper Torr.	120	RACAL		500B series CRO's. 50 Hz-1 MHz.	
		Store 4. Uses D/4 inch magnetic		Greater than 60dB dynamic range	475
		tape. Will record 4 F.M. channels.			4/3
Hz, 150w		Operates at 7 different speeds.	1950	Sweep Generators	
112, 15000	125		1000	HEWLETT PACKARD	
	125	S E LABORATORIES		8690B Mainframe. Int/Ext AM. Ext	
		6150/6151 12 channel UV		FM	600
Stab.	150	1250 mm/s-25 mm/min 6 in chart	1400	8693B / 100 3.7-8.3 GHz.5mW. PIN	
		994 6 Channel Pre-Amp ± 1% ± 1V		levelled 'N' connectors	600
).	55	o/p	450	8699B / 100 0.1-4 GHz.6mW. (20mW	
2 %		6008 25 Channel µV 8 in 4m/sec to		to 2 GHz). PIN levelled. 'N'	
005% reg.		25mm/min	895	connectors	1200
	350	SMITHS INDUSTRIES		TEXSCAN	
		RE541.20 Single Pen. 0.5mV-100V		9900 Sweep Generator 10-30 MHz	
ual Cumplu	00	FSD. 3-60cm/min and hour	350	CRT Display	525
ual Supply	90	YOKOGAWA		VS60 Sweep Generator 5-100 MHz	323
		3046. 10 inch Chart Single Pen. 0.5		Rate 60 Hz	950
15V 2A		mV-100 VI/P2.60cm/min and/hr	350	LN40A Log Amplifier	105
	60	3047. 2 Pen Version of 3046	425		100
			42.0	T.V. Test Equipment	
ck	32	Signal Sources and		PHILIPS	
		Generators		PM5508B Pattern Generator, 625	
		BOONTON		lines PAL. UK Systems	225
A DC Stab.	375	102B 4.3-520 MHz Int/Ext FM/AM		Voltmeters-Analogue	
S		0.1μV-1V 50Ω	1725		
S			1723	AVO	
,	50	DYMAR		8 Mk IV	70
	30	1525 100 kHz-184 MHz Int/Ext	FOF	BOONTON	
		AM/FM Batt/Mains	525	92AD/01/09 10 kHz-1.2 GHz 1999	
/50Ω		GOULD ADVANCE		FSD 10µV Res	525
	220	SG70 5 Hz-125 kHz 600Ω 4w	85	92C 10 kHz-1.2 GHz 500µV-3V. 1%	
V/ 50 Ω		HEWLETT PACKARD		of FS	350
	175	204D 5 Hz-1.2 MHz. 600Ω. 80dB att.			-
innel		O/P 5V RMS	150	BRADLEY	
ns	50	204D/001 As for 204D (Battery		CT471C. AC/DC/Ω/current	
MHz RT 5ns	100	operated)	175	multimeter and RF	75
MHz		608E. 10-480 MHz AM	410	HEWLETT PACKARD	
	175	620B 7-11 GHz 50Ω FM/PM 1mw	1100	400E Millivoltmeter	Labor.
ARD		8614A 800 MHz-2.4 GHz + 10dBm	1100	10 Hz-10 MHz B/W 1mV FSS	99
pulse O/P.		to — 127 dBm 50Ω AM/FM	1950	427A. AC DC/Ω multimeter	275
IHz. 15ns RT	350	8616A 1.8-4.5 GHz Ext AM/FM/PM		3406A. 10 kHz 1.2 GHz	345
		10 mw	925	8405A Vector Voltmeter	
10V/50V		MARCONI		1-1000 MHz B/W	
100/300	350	TF144 H/4S HF Generator		Auto Phase Lock	850
00.14-	330	10 kHz-72 MHz AM	550	3400A 10 Hz-10 MHz 1mV-300V	
00 Mz.	075	TF791. FM Deviation Meter	500	True RMS	350
1ns.	275	4-1024 MHz	95	KEITHLEY	
ignal		TF801/D1. 10-470 MHz AM. FM.	255	610C Electrometer DC 1mV-100V,	
uipment		TF995A/2, 1.5-220 MHz AM, FM.	350	Amps 10 ⁻¹⁴ Recorder o/p	350
		TF2171 Digital Synchroniser for			000
0		TF2015	525	LEVELL	
Recorder		TF2002/AS 10 kHz-72 MHz FM/AM	020	TM3B 5µV-500VAC 1 Hz-3 MHz +	00
DR. Record		0.1-1V o/p	625	50 to 100 dB	80
e FM/DR		TF2012 UHF, FM 400-520 MHz,	220	LINSTEAD	
100 Hz-	CEOO	0.03μV. Counter o/p	650	M2B. DC AC 10 Hz-500 kHz	25
	6500		000	MARCONI	
		RACAL			300
er meter	75	9081 5-520 MHz LED Display O/P -	4075	TF2603. AC voltmeter to 1 5 GHz	300
er meter	90	130dBm AM/FM	1875	PHILIPS	
		ROHDE & SCHWARZ		PM2454B 1mV 300V. 10 Hz 12 MHz	
s operated.		SWOB 11. 0.5-1200 MHz. 50Ω	850	Z in 19MΩ. DC O P	300
Hz-200 kHz		SCHAFFNER		RACAL	
per width.	750	NSG101 Mains Interference		9301 RMS Millivoltmeter	
0-35 mV	59	Simulator, Superimposes Pulses on		10 kHz-1.5 GHz with carry case	. 475
	79	mains for testing immunity of		Voltmeters-Digital	
0-110V neter	52	equipment to interference. Pulse		ADVANCE	
neter	59	amplitude: ±800V. Rise Time 0.25µs.		DMM 7A/01 1999 FSD	
neter	69	Width 50 & 200µs	300		115
				AC/DC/Ω/Current FARNELL	115
	The same			DM131B. 1999 FSD AC DC Ω	05
				Current Temperature	85
				All	-1.
0-	F	Assessment to the second		All equipment for s	ale
Carsto	on Elec	tronics Limited		in fully roturbished	to
		nden Road, London NW1 9NR, Telex	:23920	is fully refurbished	lO
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NSG330 Ignition Interference

Attachment NSG200B Mains Interference

TEXSCAN

STC 74216 Noise Generator 20 Hz-4 kHz Flat/CCITT Wtg

9900. 10-300 MHz. Sweep generator with CRT display

TV Markers set of 5: 31.5, 32.5, 35, 39.5 & 41.5 MHz

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612 kHz SPM3 Selective Level Meter 300 Hz-612 kHz Temperature Measuring Equipment

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475

225

COMARK 1601 BLS Analogue Thermometer for Type A thermocouples. 4 ranges -87 to + 1000°C 1604 BLS As 1601 except ranges -60 to +170°C in 23 steps 1625 BLS Type T thermocouples

100° to +300°C in 40 steps Batt/optd 1642 BLS Type J Thermocouples -120° to 800°C in 9 steps Batt/optd.

Radio Telephones & Test Equipment DYMAR BC282 4 Port Battery Charger 50 Hz-240V AC 883 0.5W VHF R/T. 3 Freq. Selector

Phase Meters DRANETZ PA3001 Phase module 2 Hz-700 kHz \pm 0.1 to 0.25° 301A 5 Hz-500 kHz Z in 100kΩ \pm 1°-2°

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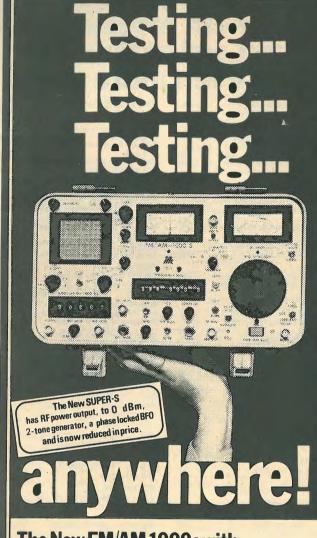
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For further information contact Mike Taylor



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AUTO UNIT DISPLAY CONTINUITY TEST

(6110 and 6100 only)

* 10 AMP AC/DC (6110 and 6220 only)

Introducing the latest professional state-of-the-art 31/2-digit DMM - at really oldfashioned prices! From just an unbelievable £39.95 inc. VAT, plus £1.15 p&p!

***** AUTO 'BATT'

WARNING

	6100	6110	6200	6220
RESOLUTION	.lmV, 10μΑ, 0.1Ω or	all models		
FULL AUTO RANGING	V .	~	V .	-
RANGE HOLD	<i>y</i>			
UNITS OF MEASUREMENT DISPLAYED	mV, V, mA	mV, V, mA, A	mV, V, mA	mV, V, mA, A
FUNCTIONS DISPLAYED	Ω, $ΚΩ$, $ΑUΤΟ$, $ΒΑΤ$	T, ADJ, LO, - and AC		
MEASURES DC VOLTAGE TO:	1000V	1000V	1000V	1000V
MEASURES AC VOLTAGE TO:	750V	750V	750V	750V
MEASURES AC/DC CURRENT TO:	200mA	10A	200mA	I0A
ZERO ADJUSTMENT	Zeros out minute te	st-lead resistances for precis	e measurements	
ACCURACY	0.5%	0.5%	0.8%	0.8%
LOW POWER OHM RANGES	For in-circuit resistar	ce measurements on all mo	dels	
BUZZER - Continuity Test		~		
BUZZER - Over Range Indicator		<i>ν</i>		
COMPLETE WITH	Batteries, pair of Tes	st Leads, Spare Fuse, One Y	ear's Guarantee	
PRICE	ONLY £64.95	ONLY £74.95	ONLY £39.95	ONLY £49.95
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I believe you! Please send me the DMM/s as marked.	ACCESS orders taken. Please write card no: and signature.
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uA709PC	0.36	HA12002 1.45	74LS21	0.24	7490N	0.33		0.54	74LS367 0.43 74LS368 0.49
uA710HC uA710PC	0.65	HA12017 0.80	7423N	0.24	74LS90	0.90	74LS155		74LS374 1.80
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.55	74LS379 1.30	
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MC2708

2102 2112 2513 HM4716 81LS97

Pairs rm	3.2
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	_
CRYSTALS	
32,768 kHz 2,70	- 1
100kHZ 3.85	
455kHZ 5.00	
1.0MHz 3.00	
3.2768MHz 2.70	
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4.19439MHz 2.30	
6.5536MHz 2.10	
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100MHz 3.00	
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	-				BF395	0.18
UCT	S	LEDs			BF479	0.66
		RED	0.12		BF679S	0.55
.49		RED CLEAR		١	BFR91	1.33
		RED	0.15	ı	BFW92	0.60
.50		X 5MM RED		Ł	BFT95	0.99
.50	5MM	GREEN	0.15	1	BFY90 40238	0.90
.20	3MM	GN CLEAR	0.16	1	RF POW	
*20	3MM	GREEN	0.16	1		
.00	2.5	X 5MM GN	0.20	ı	DEVICE	S
	5MM	YELLOW	0.15	1	VN66AF	
TALS	3MM	YELLOW CL	0.16	L	2N3866	0.85
		YELLOW	0.18	П	SMALL	SIGNA
)	2.5		0.20	ı	RF FET	/MOSF
		ORANGERED		1	BF256	0.38
65		ORA CL	0.29	ı	2SK55	0.28
		ORANGERED		П	2SK168	0.35
65		X 5MM ORA		ł	J310	0.69
85		INFRA RED		ı.	J176	0.65
25		41 IR DET	1.51	1	40823	
10		OPT CPLR	0.04	П	40673	3SK51
10	5MM	CLIP	0.04	1	3SK45	0.49
		LCDs			3SK51	0.54
	3.5		9.45	1	3SK60	0.58
			8.95	ı	MEM680	0.75
	5 d	igit	8.95	ł	BF961	0.70
					BF960	1.24
					3SK48	1.64
O.Ou	OTTERS?	DIODE BAL			_	_
		SBL1=MD108				
			4.25	,	LCD Mo	dule
		1-200MHz		1	CM161.	
			5.75	1	Miniatur	o clock
			8.45	1	12/24 hi	
			9.25	1	day, date	
		5-500MHz 1		1	backligh	
		5-200MHz 1		1	All for	
Divi	.02	J ACCUATE A		-		

TRANSISTORS	CAPACITORS
AUDIO DEVICES	All 5mm or less spacin
BC237 0.08 BC238 0.08	CERAMIC 50v

CERAMIC 50v 2P2,3P3,4P7,6P8 8P2,10P,15P,18P..0.04 22P,27P,33P,47P 56P,68P,82P,100P.0.05 150P,220P,270P 330P,390P,470P...0.055 LNO,2N2,3N3,4N7..0.06 LON (0.01uF)...0.05 22N,47N.....0.06 100N,220N.....0.09 MONOLITHIC CERAMIC 10N,100N.....0.16 FEEDTHRU 1NO SOLDER IN....0.09

POLYESTER (SIEMENS) LOmm LEAD SPACING 10N,22N,33N.....0.17 47N,68N,100N....0.19 220N,470N.....0.22 luF......0.29

POLYESTER (GENERAL) 10mm LEAD SPACING 10n,15n,22n,33n..0.06 47n,68n,100n....0.08 220N.....0.11 20mm LEAD SPACING 220N,330N,470N...0.18 AUDIO POWER

STUDENT LEAD SPACING
1N0,10N,22N,33N.0.08
100N....0.09
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100,102,105,108..0.11 202,207,303,309..0.12 407,506,608,100..0.13 TANTALUM BEAD CAPS TANTALOM BEAD CAPS
16v: 0.22,0.33,
0.68,1.0......0.18
16v: 2.2,4.7,10.0.19
6v3: 22,47....0.30
10v: 22,100....0.35

ALUMIN ELECTROLYTICS RADIAL (VERT, MOUNT)

(uF/voltage) 1/63,2,2/50,4.7/35 10/16,15/16,22/10 33/6.3.....0.08 22/16,33/10, 47/10.....0.09 10/63,22/50,33/50, 47/16,100/16....0.10 47/63,100/25,220/16 470/6.3.....0.12 1000/63,470/16, 1000/10.....0.18 1000/16,470/63...0.23 1000/63,2200/16.0.30 3300/25...0.69 1000/100...0.88 10000/70...3.00

AXIAL (HORIZ. MOUNT) 1/25,4.7/16,6.4/25 10/16.....0.08 10/16......0. 4.7/63,22/10,22/16 33/16.....0.09 47/25,100/16....0.10 100/25.....0.11 1000/16.....0.25 2200/16,1000/25..0.36 1000/35,4700/16..0.45 1000/50.....0.58

RESISTORS 0.25W, 5% El2 CARBON lohm-10M......0.02 0.25W 1% El2 METAL FILM

HORIZ CARBON PRESETS 10mm TYPE 100ohms-2M5.....0.12 HORIZ CERMET PRESETS 1k, 10k.....0.27

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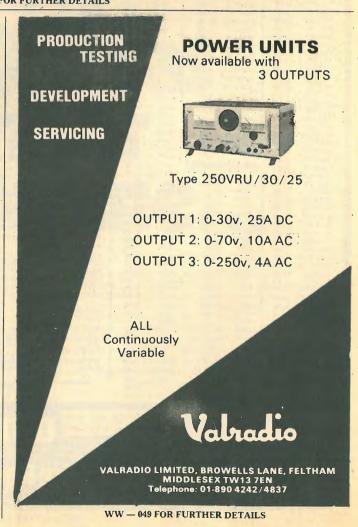
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time-consuming nuisance and even dangerous to replace, but more importantly they
are responsible for 'envelope
distortion' i.e. dynamic compres-

are responsible for 'envelope distortion' i.e., dynamic compression of the signal, even fuses in the feedback loop suffer from the first two disadvantages, and the latter to a lesser extent.



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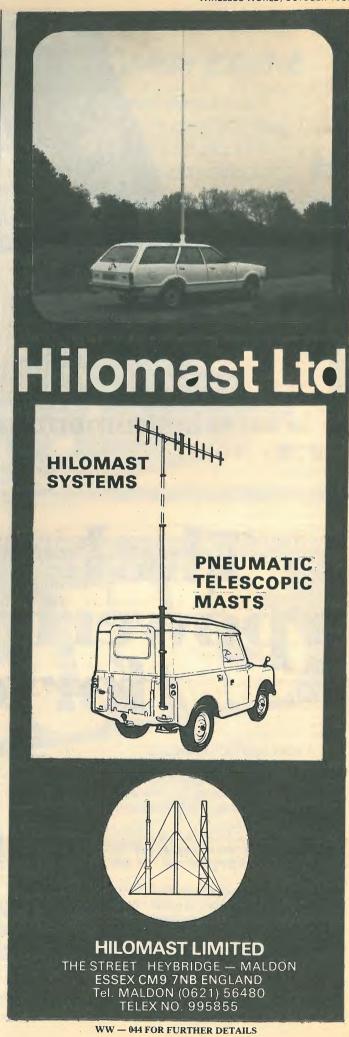
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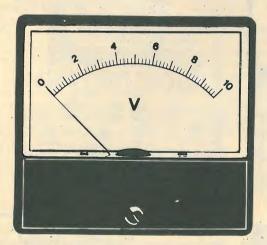
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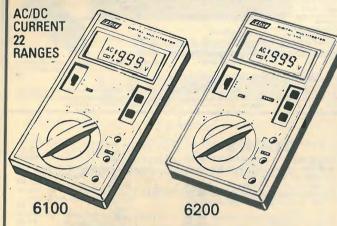
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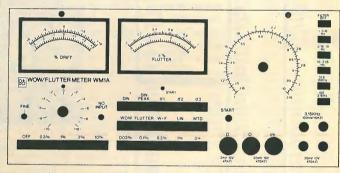
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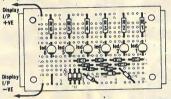
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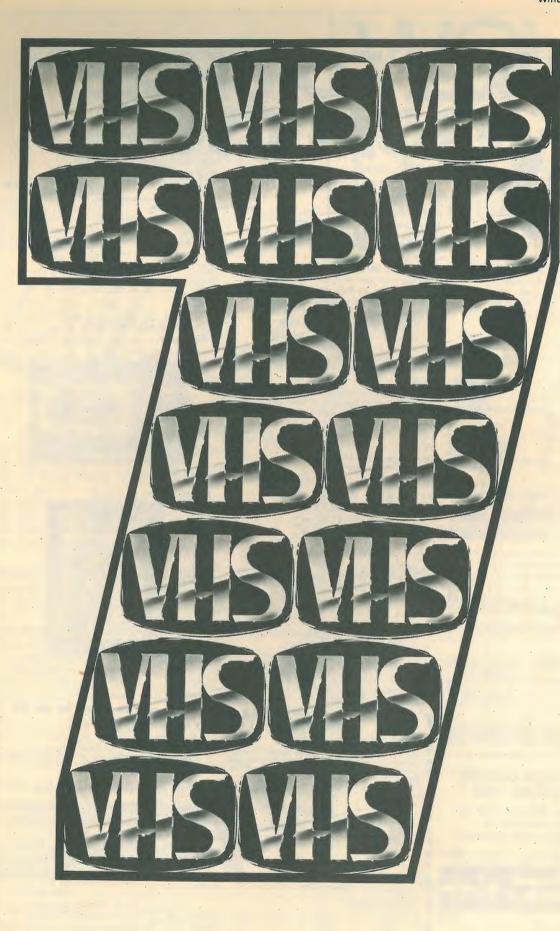
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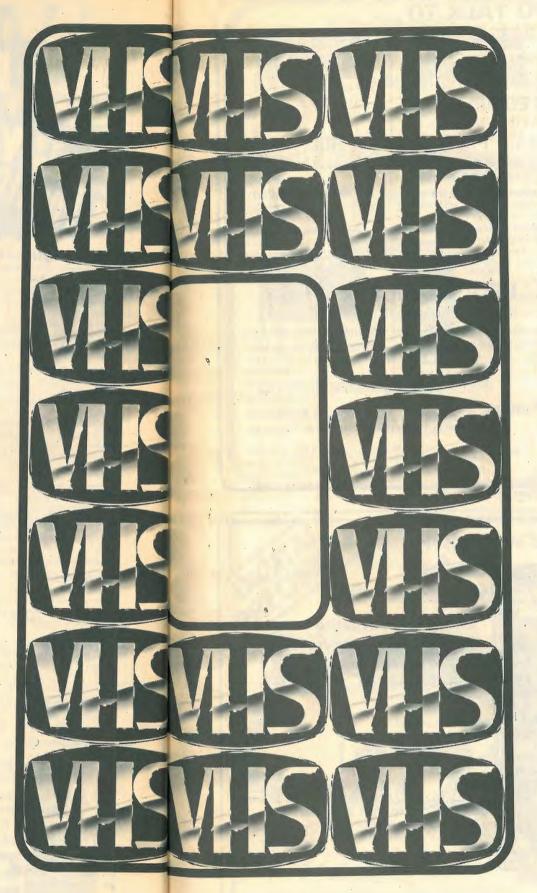
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Publishing Director: GORDON HENDERSON The director of the CCIR, Richard Kirby, made a good point recently when he said, in opening an IEE conference in London, that studies of spectrum utilization should be better recognized as a legitimate and challenging discipline of communication science (News, September issue). In spite of the fact that the welfare of peoples had "become intricately dependent on a great array of radio techniques and services" all sharing the common resource of the electromagnetic spectrum, only a few specialists were fully aware of "the precarious balance that is this matter of spectrum utilization and of its increasingly critical and complex character." This came well from the head of an international body. And the content of the IEE conference itself, on spectrum conservation, strongly reinforced his argument. No doubt from necessity rather than choice it had plenty of papers on particular techniques bandwidth efficiency, frequency re-use, station siting, reducing spurious emissions etc. — but not a single "overview" paper that tackled spectrum conservation as a general socio-economic requirement and analysed comparatively the different radio services' information handling needs. We had detailed results from specialists who are working away in separate compartments but not speaking to each other. Everyone is diligently practising personal hygiene in this field but nobody is concerned about public health.

The central fact that an "overview" paper would have brought out, of course, is that spectrum conservation is much more than the business of reducing frequency bandwidth to a minimum. Efficient use of the spectrum also depends on sharing frequencies in time or in geographic space - and also, less commonly, by different polarizations of wave propagation. This was at least implicit in the IEE conference. As one of our contributors, Leslie Berry, has pointed out, we should not be talking about spectrum

pure and simple but what he called "spectrum space" with the three dimensions of frequency bandwidth, time and physical space as area or volume (see "Measuring spectrum use", December 1978). And in his article Mr Berry proposed a measure for determining the efficiency with which radio systems use this quantity .— a ratio of communications output to spectrum-space input. Some such quantity (cf. Shannon's formula for the maximum capacity of a communications channel) should surely be the starting point for all studies in spectrum utilization. How far should communication

science go in pursuing fundamentals? It depends on what you understand by communication. Those of us who think that engineering should concern itself. with the human communicators as well as the hardware will claim that spectrum conservation should study both the demands and the real needs of the users of spectrum space. Any user who demands more space than he really needs is clearly planning to use the spectrum inefficiently. At present the goods are carved up arbitrarily and irrationally by authorities whose decisions are little more than passive responses to the demanders. He who shouts loudest gets most. Those with the loudest voices are the political and economic interests that determine the established order in any place. Because they benefit from maintaining the status quo these people do not want any other system of spectrum apportionment and least of all a system based on a rational assessment of human needs. They have a direct interest in continuing the piecemeal, divisive approach to spectrum studies and keeping engineers and scientists where they belong. This is why there is so little money available, as Richard Kirby noted, to support the fundamental study of this resource - a natural resource which, an international commission has rightly claimed (News, May issue), should be more equitably shared as the common property of mankind.

Digital frequency display for a receiver or for general use

by J. L. Linsley Hood

The addition of a numerical display of the tuned frequency can make a useful improvement to the ease of use of a radio receiver, especially in the case of broadcast reception on the short-wave bands, and a circuit is given for such a display designed for use with a Yaesu FRG 7 communications receiver. However, the circuit techniques employed for this purpose may be adapted with little difficulty to other applications ranging from l.f. frequency measurement to f.m. tuner station

One of the most attractive of the facilities offered by digital circuit components is the simple numerical display of voltage or frequency, with a substantial reduction in the ambiguities in the reading of either of these variables.

A particular area where the numerical display of frequency is of substantial value is in the display of the tuned frequency of a short-wave radio receiver, since the crowding of transmitters in the broadcast bands demands a degree of adjacent channel selectivity which makes an analogue tuning dial very difficult to interpret. The instrument described below was intended for use in the display of the frequency of the second, tuned, i.f. in a Yaesu Musen FRG 7 communications receiver, but the design was deliberately chosen so that it could be used equally well in other frequency counter applications with appropriate small modifications to the arrangement of the circuit.

Since it is the belief of the author that there is a wide, and growing, divergence between the areas of understanding of those electronic engineers whose interests and experience lie in 'linear' or 'analogue' electronics, such as amplifiers and radio systems, and those whose experience is mainly confined to 'digital' circuitry, as in numerical display systems and microprocessors, it is thought that any description of digital circuitry which is accessible to the former will appear very ingenuous to the latter. Apologies are therefore offered in advance on this score, to those whom it may offend.

Circuit arrangement

The method employed in frequency counting is shown in the block diagram of Fig. 1, and consists basically of five

parts. The first of these is a circuit designed to define an accurately determined time interval, during which some form of 'gate' will be opened to allow the frequency to be measured to pass through to a counter. This interval generating circuit is almost invariably quartz crystal controlled, and usually consists of a crystal oscillator, followed by an appropriate number of frequency divider stages. The 'gate' can be one of a number of logic elements, but an And or a Nand is usually the most con-

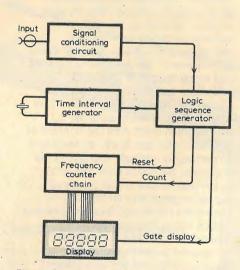


Fig. 1. General arrangement of frequency

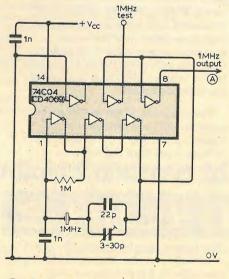


Fig. 2. Crystal oscillator, with test point output.

The second necessary part is some form of signal conditioning circuit, which will convert the probably small amplitude sinusoidal input signal at a high impedance into a well-defined square wave of adequate amplitude to swing cleanly between the '0' and '1' levels of the logic and counter elements.

The third essential section of the counter is a suitable logic-sequence generator which will perform the operations of resetting the counter, either to zero or to some predetermined number, opening the gate, and operating the display at the conclusion of the counting operation.

The two remaining stages are the frequency counter chain, which will normally have an output in binary coded decimal (b.c.d.), and the display section, which can be a b.c.d.-to-sevensegment decoder, some form of latch or display gating, and a seven-segment, light-emitting-diode, vacuum-fluorescent or liquid-crystal numerical indi-

By far the most cost-effective way of providing a numerical display of this type, is to use one of the many largescale integrated-circuit 'single chip' counters - available from Ferranti, National Semiconductor, Intersil, Oki and many other makers. The only drawback to this approach is that there is often very little scope for a change of intention once the integrated circuit has been acquired, and the choice of display, offset frequency, or operating range may be fairly circumscribed. Indeed, in using a single-i.c. frequency counter, there is little point in going to the trouble of building the counter for oneself, rather than buying a complete ready-built circuit, so long as the desired specification is available - but this necessarily precludes the possibility of a versatile unit.

For these, and other, reasons it was decided to put a frequency counter together from standard digital i.c. building blocks, while retaining as many options within the structure for other uses as appeared practicable.

For reasons of practical convenience, adaptability to supply voltages, and low power consumption within the counter circuitry, it was decided to use c.m.o.s. logic elements, of the 74C... series, which offer pin-for-pin interchangeability with the equivalent 74 ... (t.t.l.), 74S... and 74LS...

transistor-transistor, Schottky, and low-power Schottky 5V logic families. This would allow a subsequent increase in operating speed, if required, without the need for major redesign, by the simple replacement of some of the leading counter i.c.s and a reduction of the supply voltage.

Experience with the National Semiconductors 74C...c.m.o.s. logic elements, in the unbuffered types, has shown that at 12V supply-line levels, an operating frequency in excess of 7MHz can be assured, with 10MHz being typical. Also, in common with other c.m.o.s. logic circuits, the very high input impedance of the gates allows various quasi-linear operating modes to be employed, which are very useful in signal level translation.

Crystal-controlled oscillator. The circuit of this is shown in Fig. 2, and employs a 1MHz parallel-resonant, ATcut quartz crystal, of the type intended for use with a 30 pF load capacitance. The first element of the 74C04 hex. inverter is self biassed with a shunt 1M0 resistor, and a high-gain, phase inverted feedback signal is derived from the third of the series-connected stages.

A 1 nF capacitor from input to ground prevents spurious overtone modes. Two further stages act as buffers to the counter and test points respectively. The unused input is grounded to prevent uncontrolled action, a practice which should be observed, where appropriate, with all c.m.o.s. gates. A small, preferably ceramic, capacitor in the range 1 to 100nF — is connected from the h.t. line to the ground line as close as convenient to the supply to the i.c. to prevent spurious triggering of these or other stages.

The output from this circuit is a clean square wave at 1MHz frequency, and of about 0.8 V_{cc} amplitude, peak to peak. Precise frequency adjustment may be made by alteration of the 3-30pF trimmer capacitor.

Divider chain. The circuit of this is shown in Fig. 3, and consists of a chain of four 74C90 decade counters. These i.c.s are internally organized as a divide-by-five and a divide-by-two counter, connected in such a manner that the output is taken from the divide-by-two stage, which has an equal mark-to-space ratio squarewave output. Since the input is fed by a 1MHz signal, the output of the first i.c. in the divider chain, at pin 11, is a 200kHz signal. The crystal may be tuned to approximately 1MHz by adjusting for zero beat note between this and the 200kHz carrier from the Droitwich Radio 4 transmitter, or more accurately if a double-beam oscilloscope is available on which these two signals may be displayed simultaneously.

Signal conditioner. This circuit is shown in Fig. 4. The input stage is an f.e.t. amplifier with a gain of 6-10 in the range input

Fig. 3. Divider chain provides 100Hz for gate control.

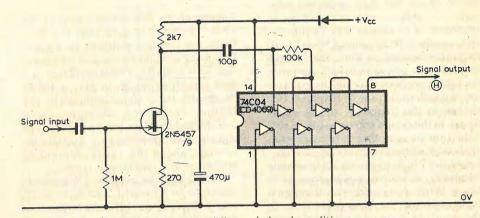


Fig. 4. Unknown-frequency input amplifier and signal conditioner.

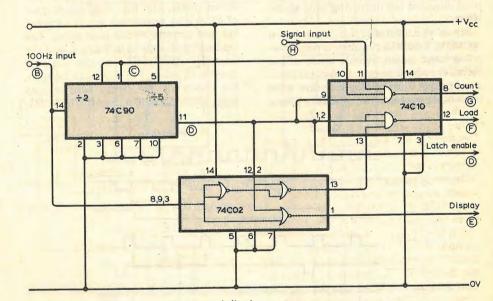


Fig. 5. Logic circuit to control gate and display.

1-10MHz, and capable of operating down to very low frequencies if the coupling capacitors are increased in value. Half of the six inverter stages of the 74C04 are used as a three-stage cascaded amplifier, with the input stage self-biassed to sit at a potential suitable for linear amplification.

Since the operation of this stage at high frequencies can be embarrassed by h.t.-line ripple being amplified by the cascaded stages, the input coupling capacitor is made small deliberately, and the h.t. supply to this i.c. is decoupled through a diode 'hook' and a large value electrolytic. The source of this problem is described later. The input sensitivity of the conditioner stage is better than 10mV at up to 5MHz.

Counter logic circuit. The design of this stage is of considerable importance in

the operation of the counter, and a number of variations of possible circuitry have been published, having varying degrees of complexity. The basic task is, however, a simple one. In order of required performance, an appropriate pulse must be provided to reset the counter, the gate must be opened to allow the signal to be routed to the counter, and after the count has been completed, the final count must be displayed; either by means of a pulse which turns on the display, or by a pulse which allows a 'latch' to transmit an input signal to its output, and then hold this signal until the next 'latch-enable' pulse is received.

In the particular application for which this unit was designed, that of displaying a 3-2MHz signal with a 100Hz accuracy, a gate open time of 10 milliseconds was required. This was achieved by dividing the 100Hz output signal of the counter chain by two. using part of a 74C90, giving waveform C in Fig. 6, and this is used to control one input of a 3-input Nand gate, as shown in the logic circuit diagram of Fig. 5.

A fundamental problem in all frequency counters is that posed by the statistical fluctuation of the count by ±1 digit due to the random sampling of the count during the gate open period. This causes an irritating flicker in the display of the last digit. A partial solution to this problem, normally employed, is to sample less frequently, and usually a free-running, slow, multivibrator is used to limit the sample frequency to one which will minimize the flicker without making the counter too sluggish in its rate of response to a change in the frequency of the input signal. In the design shown in Fig. 5, this reduction in sampling rate is accomplished by using the divide-by-five output of the 74C90 to give a 10Hz sample frequency, as shown in waveform D in Fig. 6. When combined with the signal input at H and the 10ms pulse from C, the result is a negative-going count waveform train as shown at G, having a 10ms duration but occurring only every

Since the counter i.c. chosen, a 74C192 (CD40192), requires a negativegoing reset pulse, which must occur before the count begins, two gates from a quadruple 2-input Nor gate are used to combine waveforms B and C to give a suitably timed 5ms duration pulse, which is gated by another 3-input Nand to give a negative-going 5ms pulse at 100ms intervals, as shown in waveform F. This is used to reset the 74C192 to a predetermined count number, in a binary-coded decimal form. This operation is described as 'load'.

Two options are available in the display of the count, which require different operating waveforms D and E. This choice is described later.

Counter chain. The circuit layout of the FRG 7 receiver is such that the first frequency changer is driven by a highfrequency oscillator, used in conjunction with a quartz crystal oscillator in a drift-cancelling mode, to give a 1MHz bandwidth 1st i.f. signal, reduced by the second frequency changer to a 3.455MHz-2.455MHz bandwidth slab of signals, corresponding to a 0-1MHz increment above the 1MHz interval to which the first oscillator is tuned.

The task of the frequency counter is therefore to represent 3.455MHz as '0' and 2.455MHz as '1000kHz', which is accomplished by the use of a 74C192 synchronous up/down counter in the 'down' mode, with the initial count level of 455.0'(the presumed 3rd i.f.) loaded into the counter by the load pulse. The loading table is given in Table 1, and the general organization of the counter chain is shown in Fig. 7. As can be seen. the choice of 'up' (from zero or any other chosen number) or 'down' count-

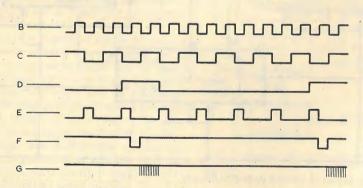
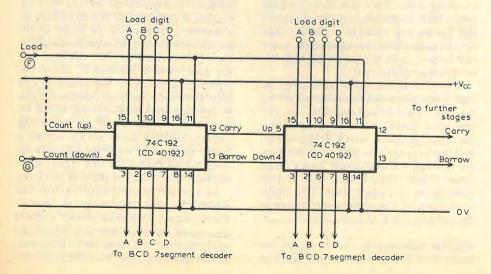


Fig. 6. Waveforms in logic circuit of Fig. 5.

Fig. 7. Part of four-stage down counter.



BCD

Table 1. Preloading.

ing is made by the selection of either the pin 5 or the pin 4 inputs to this i.c. The unwanted input is taken to the positive 'V c' rail.

Any number of counter stages may be cascaded in this fashion but, in the example shown, four are used with the signal input being taken to the least significant digit counter - which, in this case, will display the 100Hz number since the gating period is 10 milliseconds. Although a number of counts between 34,550 and 24,550 will be received during this period, the first digit is not of interest and is therefore not displayed. An input frequency of 3,455 kHz will therefore be represented simply as '000.0' and 3,355 kHz as '100.0', ascending to '999.9' as the input frequency decreases to 2.4551 MHz, which is the required condition.

Display. Two possible display modes are feasible, depending on whether leading-zero suppression is needed, based on the 74C48 b.c.d. to seven segment decoder - which permits leading zero suppression but not input latching - or the CD4511, which incorporates a latch but not leading-zero suppression. If both of these facilities are required, the 74C48 should be used with a separate quad or octal latch interposed between counter chain and decoders. An example of this using the 74C373 octal latch is shown in the Appendix.

The first of these two options is shown in the diagram of Fig. 8, using the 74C48 coupled to common-cathode l.e.d. seven-segment displays, via 180ohm, current-limiting resistors. The decimal point is permanently illuminated via a 1k2 resistor to the rail, at a position to the left of the least significant digit. Pin 4 on this i.c. performs the dual function of blanking input or leading zero suppression output, so that if pin 4 is always connected to the pin 5 of the next, less-significant digit, no leading zeros will be shown when the input pin 5 of the most significant digit is connected to the 0V line. Connecting this to the positive line will allow leading zero indication.

If pin 4 on these i.cs is taken to the 0V line, the display is suppressed, and this is used to prevent display during the count or reset periods by connecting these pins through small-signal silicon

diodes to the display pulse output E of the logic circuit. This causes the display to be illuminated at a 50Hz frequency on

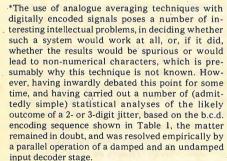
a 1:4 duty cycle. Persistence of vision

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prevents visible flicker.

The major snag with this arrangement is that a mean current of some 50mA for each seven-segment l.e.d. display is necessary for adequate daytime brightness, which means 200mA in total for four digits. Since this current is pulsed on a 1:4 duty cycle, the peak display current can be 0.8 amps at a 50Hz pulse frequency. This inevitably causes some h.t.-line ripple, and argues both the need for a separate power supply and some decoupling of sensitive portions of the circuit such as the signal conditioner stages, to prevent h.t.-lineborne interference with their operation. Nevertheless, with these precautions, this option is a satisfactory one.

As mentioned above, one of the inherent problems of any counter system is the inherent statistical uncertainty of the input count, which can cause a ± 1 digit flicker in reading. An amelioration of this problem which is possible with c.m.o.s. logic elements, because of their very high input impedance, is to put an RC input filter in the b.c.d. signal lines feeding the decoder, and this arrangement is shown in Fig. 8.* The only snag with this is that on changing frequency, the last digit (in this case the 100Hz one) tends to lag behind the others in its response. Since this digit is the least significant one, this is only a small penalty.



What was found in this trial was that occasionally the 'damped average' was biassed in one direction or another, by comparison with the visual estimate of the digit jitter, and that, very occasionally, unexpected numbers - i.e., more than one digit away from the central number - could appear briefly in the display. However, the visual comfort of a stable indication was thought, in this instance, to be of greater benefit than a possible ±1 digit averaging error. No non-numerical digits have been

Presumably, the result is predictable statistically if a Gaussian distribution is used to determined the weighting of the individual 'H's and 'L'. and if the gate is assumed to behave in an ideal manner in which any input >V_{cc}/2 is an 'H' and any input < V_{cc}/2 is an 'L'.



Fig. 10. Counter in use with communications receiver.

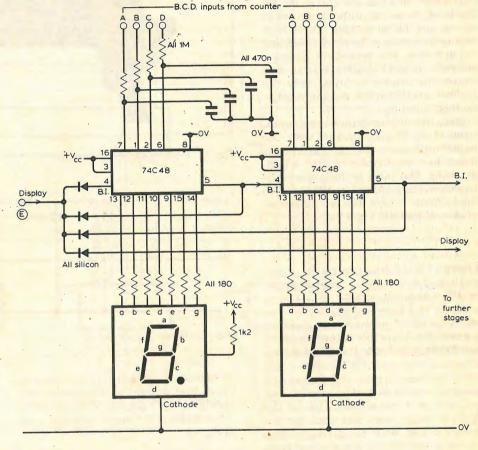


Fig. 8. Two sections of count and display circuit.

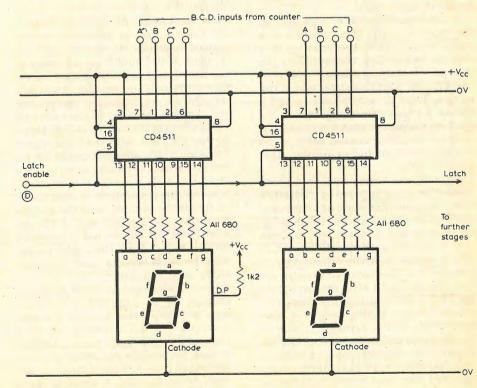


Fig. 9. Decoder and display using CD4511 latched decoder i.c.

CD 4511 option. This type of b.c.d.-toseven-segment decoder has a built-in latch circuit, which allows it to store the input b.c.d.-coded signal until such time as a refresh instruction is received.

The operation of this latch is such that no information can be transferred to the output while the input to pin 5 is. high (i.e. at V_{cc} line level). In this condition, the decoded output refers only to the last instruction received on its inputs while pin 5 was low (i.e. at the 0V line level). Since the output from pin 11 of the last 74C90 divider (D) is high during the whole of the reset (load) and count cycles, this waveform makes a convenient latch-enable signal, and causes the display to show only the number attained when the counter has finished counting.

Since this display is then continuous, there is no display flicker, with the important feature that the current demand from the decoder/display is nonpulsating. The value of the series resistors should be amended (to 680 ohms) to take account of this. The circuit connexions are shown in Fig. 9.

Use as FRG7 frequency counter

The complete circuit, as used, is shown in the photographs of Figs. 10 and 11. Inview of the high sensitivity of the receiver - if properly aligned in accordance with the manufacturer's instructions, the background sensitivity threshold is below 0.1 microvolt - it is very necessary that the frequency counter should be well screened, and enclosed in a metal case. One of the Vero G range cases was used for the author's unit, with an internal mains transformer having an electrostatic screen, and with a coaxial socket input for the i.f. signal input.

A small modification is necessary to the receiver itself. This consists of a mains outlet cable, terminating in a suitable free socket, connected to the internal transformer primary — so that the display extinguishes when the receiver itself is switched off - and a coaxial socket outlet on the rear of the receiver, connected with a short length of low-capacitance coaxial cable to the second oscillator buffer output (test point TP 404).

Since it was anticipated that there would be some unwanted radiation from the counter, in spite of all precautions, an on/off switch was included on the counter unit. In the event, this was only receivable at the 25, 27, 28 and 29 MHz frequencies, where it heterodyned with the small amount of stray radiation from the internal 1 MHz crystal within the FRG 7. By adjusting the frequency counter crystal tuning to give a zero beat on the 29MHz harmonic, it can be brought into concordance with the internal crystal to better than a few Hertz in 1,000,000. With this heterodyne removed, the total spurious radiation level on the prototype is so low that the

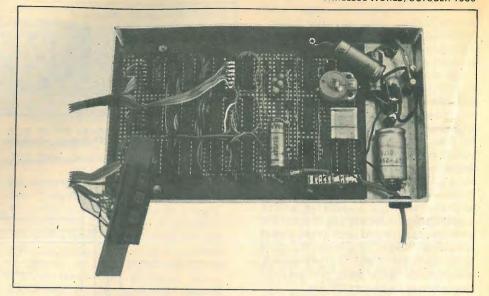


Fig. 11. Inside view of instrument.

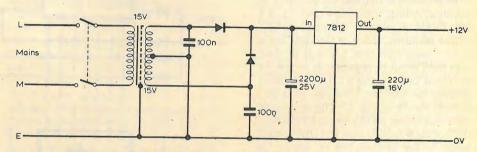


Fig. 12. Power supply circuit.

additional on/off switch is only used to check that occasional whistles on the tuning scale are not due to the counter.

A suitable power supply unit is shown in Fig. 12.

Appendix

Other applications. As shown in Table 1, the counter can be preloaded to any desired frequency offset. This can be 0, if inputs A, B, C and D are all taken to the 0V rail, so that the frequency read is that of the input. In this case, it will be a straightforward frequency meter, and will normally be used in the count-up mode. (Pin 4 of the leading 74C192 taken to the +V_{cc} rail, with the input signal fed to pin 5.) If a sampling rate of 100 Hz will give adequate display accuracy, the circuit can be used as it stands. If, however, the circuit is to be used for an l.f. frequency counter, with the sample frequency reduced to, say, 1 Hz, by additional 74C90s in the frequency divider chain of Fig. 3, the display flicker with the 74C48 decoder, used as shown,

would be unacceptable. Since leading zeros would normally require to be blanked (the least significant one never is, since it would cause the display to extinguish on a count of 0 — so if leading zero blanking is used, pin 5 of the RH 74C48 must be connected to +V_{cc}) the CD4511 is unlikely to be suitable. A separate latch will then be necessary. A convenient

system is shown in Fig. 13, using the 74C373 octal latch. Since the latchenable signal with this is high, the switching waveform from D requires to be inverted. The remaining Nor gate of the 74C02 in Fig. 6 is used for this purpose.

The remaining useful application of this circuit is in the display of the tuned frequency of an f.m. tuner, in the range 86-108 MHz. Since the oscillator frequency of the f.m. tuner head will be above the tuned frequency by 10.7 MHz, the counter will be used in the count-up mode, with a preloaded number equivalent to the 9's complement of 10.7 (89.2). If a 100kHz indication accuracy is adequate, a 4-digit counter will again be used, with the decimal point wired in ahead of the least significant digit.

Since leading-zero blanking will be needed, at least for the first digit, pin 5 on this 74C48 should be connected to the 0V rail.

As mentioned earlier, the upper reliable frequency limit of the c.m.o.s. counters is about 7-8 MHz on a 12 volt supply. The input frequency from the f.m. tuner oscillator will be well above this, and the most convenient way of solving this problem is by using an input divide-by-100 i.c., such as the RS 8629. This should be mounted with a transistor emitter-follower input, as close as practicable to the tuner head. The output frequency from this, in the range 1.187MHz to 0.967MHz, can then be

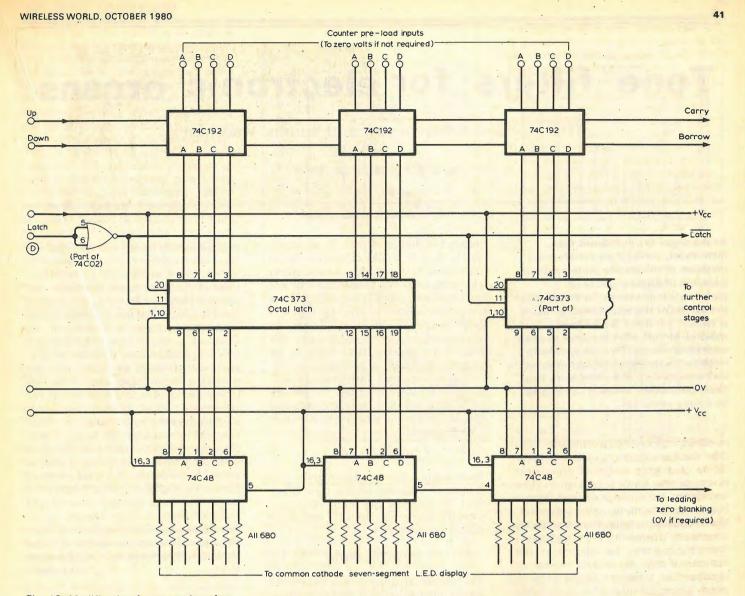


Fig. 13. Modification for use as low-frequency counter, using external latch.

taken to the counter by a screened cable. The count accuracy required will be lkHz, and will allow a sample period of 1 millisecond. One fewer 74C90 in the divider chain of Fig. 3 will be required. Apart from the modified input signal interface, as shown in Fig. 14, this reduction in the length of the divider chain and the change in the count mode and offset of the 74C192s, the circuit form of Fig. 10 is as required.

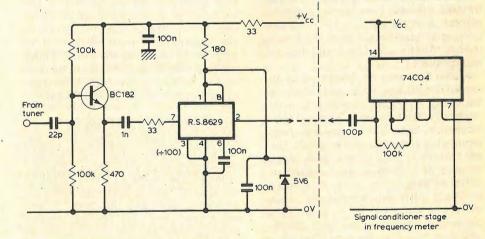


Fig. 14. Divide-by-100 prescaler for use with f.m. tuner.

Tone filters for electronic organs

Part 1: organ tone spectra and source waveforms

by C. E. Pykett, B.Sc., Ph.D.

As the organ is a sustained-tone instrument, achieving a satisfactory imitation of the steady-state acoustic emission of organ pipes is of paramount importance. In this respect the design of the tone-forming filters is crucial, yet there is a curious absence of definitive material dealing with filter design. This is apparently reflected in the range of commercial instruments on the market: with few exceptions their "voicing" seems to be mainly empirical.

To derive a simple expression for the frequency response of a tone filter consider the basic organ system, representative of a wide range of electronic instruments, shown in Fig. 1. The waveforms are initially derived from a continuously running tone generator. Waveforms at various frequencies are selected by depressing keys, and envelope shaping may be applied at the instants of key attack and release to simulate the transient phenomena of organ pipes. (Whilst of considerable importance, transients are not further discussed here). The signals are passed through various tone forming filters depending on the stops or tone colours selected and the output from the filters is then finally amplified and fed to loudspeakers.

A tone filter may be thought of as an amplifier whose gain varies with frequency. The gain can therefore be explicitly written as a function of frequency, G(f). Similarly, each harmonically rich waveform from the generators is equivalent to a large number of individual sine waves of different frequencies, each sine wave having a different amplitude. This waveform can also be written as a function of frequency, say H(f). Therefore the output from the tone filter, F(f), is the product of the input voltage and the gain just as with any amplifier

F(f) = G(f).H(f)

In general the tone filter will also modify the phase as well as the amplitude of each frequency component in the input signal. As the ear is insensitive to relative phase for present purposes, this does not matter, which

makes the design of tone filters much easier than it would otherwise be. It does mean, however, that the waveform emerging from the tone filter will not necessarily bear any resemblance to the waveform emitted by the organ pipe if both were to be viewed on an oscilloscope screen. It is only the frequency spectra that need to be matched as closely as possible.

If the frequency functions are expressed on a logarithmic amplitude scale then new functions are obtained that are related by addition rather than multiplication

$$P(f) = Q(f) + R(f)$$

Rearranging this equation gives the frequency response of the tone filter, Q(f), in terms of the input spectrum from the tone generator, R(f), and of the output spectrumP(f)

$$Q(f) = P(f) - R(f)$$

This simple equation shows that filter design involves three basic steps. First, the logarithmic spectrum of both the tone generator waveform and of the sound to be simulated must be available. Second, the frequency response of the required filter must be derived by subtracting one from the other. Third, the response so obtained has to be realised in hardware. Subsequent sections discuss each of these stages in detail.

Acoustic spectra of organ tones Before a filter can be designed to imitate the sound of a particular type of organ pipe the spectrum of that sound must be obtained. Following a careful search of the scientific and engineering literature extending back into the 1930s, it was discovered that very few systematic investigations into the acoustic spectra

Fig. 1. Basic electronic organ system considered in this article is the subtractive kind in which an harmonically rich waveform is filtered.

of organ tones have been reported. As this information is vital to the design of an imitative electronic instrument. three of the most useful references are appended here ^{2,3,4}. Boner's article (1938) describes one of the first attempts to use electronic techniques to analyse the sound of an organ pipe radiating in a free field (that is away from the reverberant conditions of an auditorium) by mounting organ pipes atop a 24 foot tower out of doors. From the three references quoted, spectra corresponding to the four main classes of organ tone can be extracted, viz flutes, diapasons, strings and reeds, and this goes some way toward providing a framework for the design of a wide range of filters. To augment this information I made recordings of organ sounds and analysed them. A large amount of information was obtained from a fourmanual instrument by Rushworth and Dreaper with some particularly fine solo

Recordings were made of organ pipes in situ using omnidirectional capacitor microphones with a frequency response from below 20 Hz to about 20 kHz. Two microphones were used, feeding separate channels of a tape recorder with a frequency response from 35 Hz to 16 kHz (±2dB). The recordings were subsequently replayed monaurally into a high resolution spectrum analysis system with a dynamic range of 60dB. The reason for using two microphones and then summing their outputs on replay was to reduce distortion of the spectrum through reflections from the surfaces in the auditorium. Because they set up standing waves, such reflections can result in a significant increase or decrease in the intensity of sound of a particular frequency at the microphone location. By using two microphones there is a reduced likelihood of an identical distorting effect occurring at both simultaneously. (A better method for averaging out the effects of reverberation would have been to use averaging in the frequency domain after phase information had been removed.) Recordings were made of four octavelyrelated samples from each stop on the organ, and the whole exercise has resulted in a library of some hundreds of pipe spectra.

WIRELESS WORLD, OCTOBER 1980

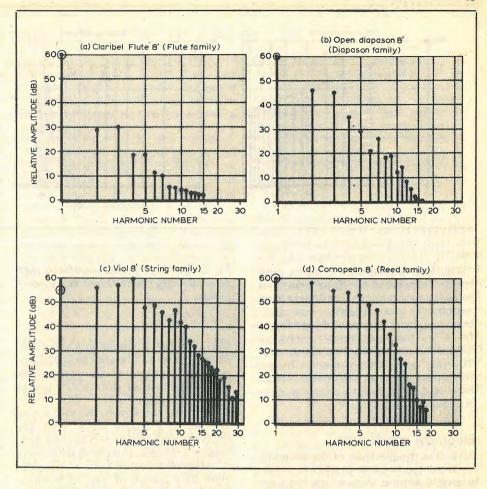
The steady state emission of a pipe is periodic at its fundamental frequency. This is the lowest frequency present in the spectrum in most cases and it defines the musical pitch of the pipe. Because the emitted waveform is periodic, the only other frequencies present in the spectrum are harmonics or integer multiples of the fundamental; there is virtually no acoustic energy lying between adjacent harmonics. Certain pipes, however, possess a significant noise component due to random fluctuations of the air. In other cases the amplitudes and phases of each harmonic fluctuate randomly to a significant degree. Both of these effects produce energy that is not confined to the harmonic frequencies in the spectrum. However assume for simplicity that the spectrum of an organ pipe consists only of equally spaced lines at the fundamental and harmonic frequencies.

This structure is shown in Fig. 2, with examples of spectra corresponding to each of the four classes of tone. These have been normalized to the frequency of the fundamental so that the abscissae represent harmonic numbers (on a logarithmic frequency scale). All of these spectra contain a large number of harmonics, at least 15, within the dynamic range of 60 dB. This is significant in that it clearly demon-

Table 1. Harmonic amplitudes of various pipe spectra in dB, corresponding to Fig. 2.

organ stop name

har.	claribel	open	viol	corno-
monic	flute	dia-		pean
		pason		
1	60	60	55	60
2 3	29	46	56	58
	30	45	57	55
4	18	35	60	54
5	19	29	48	53
6	11	21	49	49
7	10	26	46	47
8	5	18	43	42
9	5	19	47	37
10	4	12	42	33
11	4	14	40	27
12	3	8	34	25
13	3	5	32	16
14	2	2	28	15
15	2	1	27	10
16	-	0	26	7
17	-	_	25	9 .
18	_	_	23	6
19	_	-	22	_
- 20	_	-	22	_
21	_	-	18	_
22	_	_	20	_
23	- '	T	19	_
24	_		15	<u>.</u>
25	_	-	20	_
26		-	11	_
27	_	-	14	-
28	-	- '	13	_
29			_	
30	_		_	



strates that the flute is far from being a single sinewave as commonly stated. Nevertheless, as the amplitudes of the harmonics in this spectrum decrease rapidly with increasing harmonic number, it is possible to approximate to a reasonable flute tone using only a few harmonics. This is why additive sinewave instruments, which rarely have more than nine harmonics available, are able to provide good flutes whereas their performance at synthesizing almost any other type of tone leaves much to be desired. A glance at the remaining spectra in Fig. 2 shows why. For a subjectively satisfying imitation of these pipe tones, one should aim to embrace all harmonics within a dynamic range of about 60 dB. Therefore even the diapason requires about 15 harmonics and the other two spectra need more. Unless a very large number of harmonics is available in an additive instrument, the only cost-effective way to proceed is with the subtractive approach. (Whilst there are a very few additive instruments that have large numbers, perhaps in excess of one hundred, harmonics available for tonal synthesis, these are expensive experimental, developments using advanced microprocessor technology and as yet they are scarcely suitable for amateur construction.)

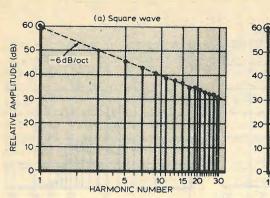
Returning briefly to the imitation of an organ flute stop of the sort illustrated by the spectrum in Fig. 2(a), this type of tone is in some ways the most difficult to simulate in spite of the apparent simplicity of the spectrum. Merely de-

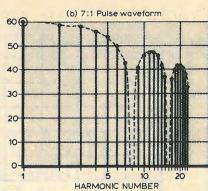
Fig. 2. Large number of harmonics in organ pipe spectra means high cost for additive

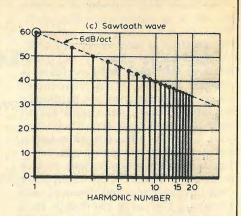
signing a filter to produce the same. overall spectral features often produces a tone that seems somewhat dull and lifeless compared to the original, especially on A-B comparison using tape recordings. Ladner³ made the same point, and it seems that the role of the low-amplitude high-order harmonics is not well understood. Sumner reports that physical features such as the "chimney" in the flute stop of that name are responsible for subtle formant bands in the spectrum, though he does not give further details.

Passing on to the other sounds, where imitation is much easier than for flutes, consider the diapason. The spectrum shows that the amplitude of the harmonics gradually falls off with increasing harmonic number. The viol, on the other hand, has harmonics that increase in amplitude up to the fourth, whereafter they fall. This is the result of a viol pipe being of smaller scale (narrower) than a diapason pipe of the same

Finally the cornopean has a spectrum in which the harmonic energy falls with frequency though the fall is not in excess of 6 dB until harmonics beyond the fifth are encountered. The relative smoothness of this curve compared to the previous three in which more scatter is apparent seems to be characteristic of many reed tones.







The four examples of organ pipe spectra represent the four principal categories of organ tone, and there is no reason why essentially the same spectrum should not be used to design filters for several footages, thereby producing a diapason chorus or a reed chorus, etc. Together with other examples in the references cited, a reasonably broad base of data is available for the construction of filters.

Electrical waveforms

As well as the spectrum of the sound to be simulated, we also need that of the source waveform from which the tone

Table 2. Harmonic amplitudes of various waveforms in dB corresponding to Fig. 3.

harmonic		waveform	
	square	7:1	saw
		pulse	tooth
1	60	60	60
2	*	59	54
3	50	58	50
4		56	48
5	46	54	46
. 6	•	50	45
7	43	43	43
8			42
9	41	41	41
10		46	40
11	39	47	39
12	0.0	47	38
13	38	46	38
14 15	37	42	37
16	37	37	37
17	35	20	36
18	*	36 40	35
19	35	42	35
20	*	42	35 34
21	34	41	34
22	,	38	33
23	33	33	33
.24		*	33
25	32	33	32
26	· (4)	37	32
27	32	39	32
28	-117 19	39	3.1
29	31	38	31
30		36	31

('denotes the absence of a harmonic)

Fig. 3. Easy-keying pulse waveforms such as in (a) or (b) are defficient in harmonic

filters are fed. It would be a short and simple matter to present the spectra of commonly used waveforms at this point but several other practical aspects require discussion first.

Probably the easiest waveform to

generate is a square wave. With the ready availability of top-octave synthesizers, dividers and envelope shapers in integrated circuit form a complete generating system of, say, 84 frequencies (seven octaves) can be contained on one card. Unfortunately the square wave is far from ideal for tone forming, except in a few cases, because it contains only the odd-numbered harmonics, whose amplitudes decrease at 6 dB per octave, Fig. 3(a). A square wave cannot therefore be used to derive any of the spectra shown in Fig. 2 as these contain even harmonics. It is, however, suitable for use where tones such as a stopped diapason or a clarinet are required, in whose spectra the odd harmonics are much more prominent than the even ones.

In a square wave multi-frequency generating system it is relatively simple to generate pulse waveforms of different mark-space ratios. These possess. in general, both even and odd harmonics and the spectrum of a pulse waveform with a 7:1 mark-space ratio has been discussed by Ryder⁵; this special case is of particular interest to those readers who may be building his organ. The spectrum, Fig. 3(b), shows that certain harmonics are missing. This effect is always obtained with pulse waveforms, including the square wave just discussed. This is merely a "pulse" waveform with a 1:1 mark-space ratio, where the nulls happen to coincide with the even harmonics. Whilst pulse waveforms again have the desirable advantages of simple generation and keying (envelope shaping) one possible problem concerns the low average energy of a waveform consisting of short pulses. This could give rise to noise difficulties at the output of the tone

filters which usually introduce considerable insertion loss.

The "classical" waveform that is often used when both odd and even harmonics are required is the sawtooth. This has a spectrum containing all harmonics, whose amplitudes decrease at 6 dB per octave as in Fig. 3(c). Unfortunately the sawtooth is not particularly economical to generate, and once generated it cannot be keyed by the simple non-linear envelope shapers commonly used for square or pulse waveforms without introducing distortion. One way to circumvent of pulse waveforms, and then combine them with appropriate weights so that a staircase waveform obtained. This is a good approximation to a sawtooth.

Another approach is to generate and key a single square wave and then convert it to a sawtooth using a discharger circuit of the type shown in Fig. 4. The square wave is first converted to a series of narrow pulses, for example by differentiation followed by rectification, which are then used to repeatedly discharge the capacitor C through the electronic switch S. Inbetween discharges the capacitor charges exponentially through R. A linear ramp is obtained if R is replaced by a constantcurrent source, though for musical purposes this would seldom be required. An exponential ramp produces little significant difference in the spectrum even at harmonics as high as the 30th. The source voltage V can be used to

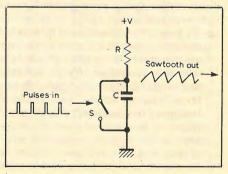


Fig. 4. It is easier to generate and key a rectangular wave and then convert it to a sawtooth than to operate on the sawtooth.

SOUND PRODUCTION IN THE PIPE ORGAN

Organ pipes emit sound when compressed air at a low pressure enters via a valve controlled from a keyboard. Various mechanical, electromechanical or pneumatic contrivances are used to control the valves. Each stop on an organ controls a whole rank of pipes, and has to be "on" before that rank will sound from the keyboard. In each rank there are as many pipes as notes on the keyboard (with a few important exceptions). Therefore even a very small organ will contain several hundred pipes, and a large one many thousands. It is this multiplicity of individually adjusted tone sources that gives the pipe organ its extraordinary richness of sound. (The origin of the term "stop" to denote a particular rank of pipes is of considerable antiquity, and is thought to derive from the great organs of the Gothic period which were originally built with no means of isolating one set of pipes from another. Not surprisingly, such means were soon introduced so that certain sections of the instrument could be "stopped" from sounding!)

Pipes fall into two categories known as flue pipes and reed pipes. Flue pipes are constructed in much the same way as a recorder or tin whistle in that the incoming wind is formed into a narrow sheet which then encounters a lip fashioned in the wall of the pipe. An oscillatory motion is imparted to the wind sheet whose frequency is controlled by the air column in the remainder of the pipe, acting as a close-coupled resonator. Thus the musical pitch is controlled by the length of the pipe. (This highly compressed description tends to hide the complexity of the physics of the flue pipe, a subject that cannot be adequately treated here). The tone quality of the sound is determined by the distribution of the energy in the frequency spectrum and for flue pipes this is to a large extent controlled by the relative proportion of length to breadth. This parameter is termed the scale of a pipe, and it gets numerically smaller as a pipe of constant length gets narrower. The smaller the scale, the greater the proportion of higher harmonics in the sound. You might think that the scale should remain constant across a rank of pipes if the tone quality is to remain constant. This is true, but in practice the scale is varied in a systematic manner so top end can remain subjectively the

Another factor controlling the timbre of the pipes is whether they are open or closed at the top. An open pipe encourages the formation of harmonics, whereas a stopped one has a sound that is dominated by the odd harmonics only.

Flue pipes are generally made of wood or metal and can have a variety of cross sectional shapes. They are used to generate three of the traditional types of organ tone (flutes, diapasons and strings) and the front pipes in an organ case often form part of a diapason rank. The physical difference in tonal structure between these types of tone is discussed later.

Reed pipes form the fourth type of organ tone and generate sound by means of a metal tongue (the reed) alternatively opening and closing an aperture that communicates with the rest of the pipe. Again, this is a closelycoupled generator-resonator system whose detailed physics is even more obscure than those of the flue pipe. An important factor however is that the shape of the resonator tube controls the timbre to a large extent. Pipes that are

flared reinforce all harmonics to a greater or lesser degree, whereas cylindrical bores emphasize only the odd harmonics. The names of reed stops, often fanciful, imply that they are attempting to imitate orchestral instruments such as the oboe, clarinet or trumpet. This imitation is usually in name only since the tone of organ reeds is unique and part of the tradition of organ building.

Perhaps the most artistic and subtle part of organ building resides in the hands of the voicer, who tunes and that the volume especially toward the adjusts the tone quality and volume of each pipe individually, a process which is the result of centuries of skill and craftsmanship. By basing the design of tone filters on the harmonic structure of actual pipes one attempts to endow the electronic instrument with some of the artistic virtues of the real one.

The relationship between the fundamental frequency of a pipe and its length has resulted in widespread use of the "footage" nomenclature to indicate pitch. An eight foot stop, for example, means that the frequency of the note two octaves below middle C (usually the lowest note on the keyboard in a church organ) is the same as that which would be produced by an open pipe as used on a piano. Stops of 16 foot pitch therefore sound an octave below this, and a four foot pitch an octave above, etc. The ability to control many ranks of pipes at once from one keyboard, or a variety of tone colours and pitches depending on the stops selected, contrinutes to the tonal variety and brilliant ensemble that is characteristic of a first rate pipe organ.

A complete account of the physical and aesthetic design principles of the organ can be found in the book by the late Professor Sumner.1

achieve envelope shaping during key attack and release.

Several filters are discussed in the next article, all designed assuming the availability of a sawtooth wave to feed them with. This has been chosen for the following reasons:

- # Its spectral structure is simple. Harmonic amplitudes decrease monotonically with increasing frequency rather than in the oscillatory fashion of a pulse spectrum. This results in a filter frequency response that is also much simpler than if a pulse waveform had been used. This is important because of the comparative ease with which an electrical implementation of the filter can be built.
- * A square wave has already been rejected as being unsuitable for all but a few special tones (though in these cases it is essential).

Sawtooth and square waves are available in the author's instrument. This meant that a subjective zudgment could be made as to the effectiveness of a filter design and in particular it was possible to make A-B comparisons of the electronically generated sounds against tape recor-; dings of the originals.

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Marconi Fellowship

The Marconi International Fellowship maintains the motto "to commission creative work in science, technology and humanism," and recognises such work with an annual award of \$25,000. The general criteria for eligibility include the importance of the candidate's contributions to communications, science or technology, and the degree to which the candidate's life exemplifies commitment to applying communications science or technology to bettering the human condition. For information of the award, write to Dr. Walter Orr Roberts, Marconi International Fellowship Council, Aspen Institute for Humanistic Studies, 1229 University Avenue, Boulder, Colorado 80302 USA.

Floppy disc system for the scientific computer — 1

8in disc stores 400K bytes

by J. H. Adams, B.Sc. M.Sc.

Storage of data in small computer systems is often accomplished by a 300-baud cassette tape recording. With a transfer rate of only 2K-bytes per minute, this method makes locating and transferring long strings of data a rather slow process. The introduction, through the users' club¹, of a more advanced operating system for the computer2, and the availability of memory expansion kits, has made a faster store very desirable. To solve this problem the author has developed a store based on an 8in flexible (floppy) disc, which can accommodate 400K-bytes of data and transfer 0.5K-bytes per

Recordings on disc are made by converting the data bytes into a serial stream of 1s and 0s at a rate of 250,000 bits per second, i.e. one bit every 4µs, truncating the 1s down to about 0.5 us pulses and then interleaving a regular stream of 0.5us pulses from the system clock as shown in Fig. 1. These pulses are used to reverse the current in the recording head and, hence, the sign of the flux recorded onto the disc. Converting the parallel input data to a stream of pulses is most easily achieved by one of the controller i.cs which are, in essence, dedicated microprocessors combined with programmed logic arrays to feed control information between the controller, the disc-drive electronics and the computer. Recordings are made on concentric rings, or tracks, 77 on an 8in disc, and a drive unit with two motors rotates the disc and steps the combined record/read/erase

head from track to track. Optical devices provide signals which indicate when the head is over the outermost track 0 and, using a small hole punched in the disc, an index pulse to indicate, when the disc starts each revolution.

The electronics in the drive unit convert t.t.l. levels to switching currents in the head and vice versa, operate the stepping motor and provide erasing' signals. Other functions may include door locking, motor-on indication, adjustment of recording current on inner tracks, separation of data, disabling the write operation on write-protected discs, and loading the head against the disc on read operations. This drive unit contains most of these features, although separation of data is achieved in the controller i.c.

Recording format

At 360 r.p.m. it is possible to record over 5000 bytes of data on each track of the disk. To allow the controller to identify recorded clock pulses from serial data pulses from the disc, the start of the decoding process is triggered by the index pulse, and the recording begins with a standard code which the controller can recognise and synchronize with. This code is often produced by repetitive recording of the byte 00, i.e. the clock pulses are recorded with no interleaved data pulses. The next task for the controller is recognition of the start of the first byte in the data stream. As all possible data bytes may start the stream, no single byte can be reserved for this purpose. Instead, a data byte with a few of the clock pulses missing is

used and is known as a mark byte. Normal bytes can be thought of as data bytes interleaved with the clock byte FF, i.e. all eight pulses. A typical mark byte is data byte FC, interleaved by the clock byte C7. After this index mark, about 5000 bytes of data follow and the recording runs to the start of the next index pulse with a final code of bytes. usually 00s or FFs. The total number of code bytes is determined by the accu-

Sectored tracks

If data transfers, which match the above, are all that is required of the disc, it is an efficient way of using the system in terms of bytes stored per disc. Usually, however, transfers are of variable length and, as it is not directly possible to access part of the way through a track, there is a limit of one recording per track, no matter how short the data block. To improve the potential disc capacity, each track is split into sectors which each require start and stop codes and identification marks. This leaves less space for data, but normally provides the most efficient mode. Such a format, now widely in use. is the IBM 3740 which fits 26 data sectors into each track, with 128 data bytes in each sector as illustrated in table 1.

In the present format, sectors consist of an identifying block followed by the data. Six 00s synchronize the clock/ data separator, an address mark (data FE, clock C7) indicates the boundaries, of the bytes, and, as previously explained, track and sector numbers are given. This is followed by a CRC, which is a two-byte cyclic recognition code used by the controller to check for errors when reading information. The sector then has a short code. immediately followed by six more 00s, a data mark (data FB, clock C7), the 128 bytes of data, a two byte CRC for the data, and a final code. Each track has 26 of these sectors end to end, prefixed by a large block of 00s, an index mark (data FC, clock C7), and trailed to the end of the track by a code. The copious supply of synchronizing bytes and CRC codes can, with suitable software in the computer, produce a very reliable sys-

Formatting all of this information onto the disc is a complicated operation,

racy of the clock and drive motor.

and a number of other formats.

the device, and the four control lines described above, there are two lines from the controller to the computer. One indicates that, either through natural completion or through a failure, the controller has finished an operation and wants servicing, INTRQ, the other, DRO, indicates that the controller dethis reason, part of the interface consires a data transfer either to or from it. sists of a simple but effective interrupt This information is present in the status register but, because of the high rates of data transfer taking place, these lines

Controller disc-drive interface

> Lines from the controller to the drive comprise step and direction signals for the head-position motor, data and gating signals for the write operation,

6 bytes 17 bytes Ident Track 00 Sector 00 Data Data CRC CRC 27 bytes 00 mark no. 00 128 bytes 2 bytes 2 bytes Ident field Data field

IBM 3740 track

IBM 3740 sector

WIRELESS WORLD, OCTOBER 1980

					 1 21 1	District Control of the Control of t
40 bytes FF	6 bytes 00	Index mark	26 bytes FF	26 sectors		approx. 240 bytes FF

must be used to enter the Z80, through

the interrupt line, in preference to the

much slower polling of the status reg-

ister. The Z80 can therefore keep up

with the steady demand for, or supply of

data between it and the controller. For

Table 1. Formatted disc arrangement. and most discs are supplied for-

matted with dummy data (usually byte

E5). Such discs are marked 128 bytes

per sector or per record, soft sectored.

The last mentioned term means that the

start of the sectors is indicated by soft-

ware recorded onto the disc, as opposed

to permanent hard sectoring, achieved

by punching index holes in the disc for

each sector (sometimes called 33 hole

media for this reason). One disadvan-

tage of soft sectoring is that, if format-

ting information gets magnetically cor-

rupted, the sectors affected become

useless. For this reason, even unused

discs should be treated with care. For-

tunately, if this should happen, the

controller can re-format tracks to this

Computer-controller interface To the computer, the controller looks like four input and four output ports. However, to address the controller, the computer only needs to supply one line each from IC, and IC, (the computer's input and output-port decoders) along with address lines A3 and A4 as shown in Fig. 2. Because neither of the address lines go to the decoding i.cs, they take no part in decoding the 8-bit port addresses which the Z80 sends along the bottom eight address lines during I/O instructions. Therefore, I/O commands such as IN(05), IN(0D), IN(15) and IN(1D) will activate the same line from IC2, the bottom three bits of each number being the same, 101, but each provides a different combination on A3 and A₄. By connecting the IN line to the controller's RE (read enable) input, and the two address lines to the Ao and A1 inputs, all four controller registers may be read by the computer. In a similar fashion, one line from IC3 drives the WE (write enable) line of the controller, which allows the computer to write information into any of the registers. For details of these see Table 2. As well as the data bus into and out of

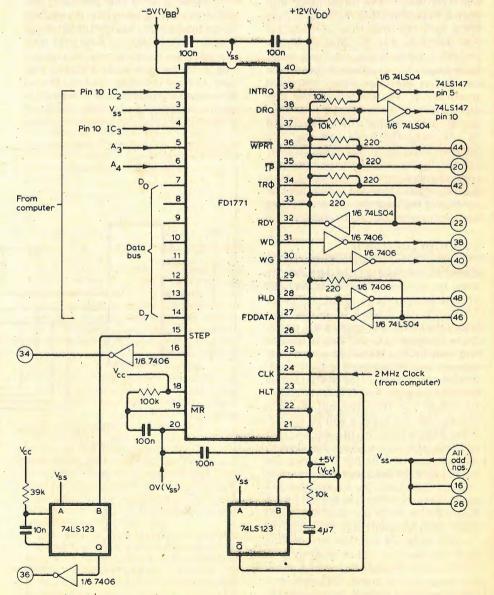


Fig. 2. Floppy-disc controller / formatter.

Fig. 1. Interleaved data and clock pulses.

The Index Mark, Data FCH Clock C7

Read on INTRO, checked for CRC IN A, (05) Status error and record not found bits during Read, Write and Seek. OUT (A0), A Receives commands from the Command computer. IN A, (OD) OUT (A8), A This register normally contains the 0 1 Track

current head position, 00 to 4C. It is reset to 00 on completion of the Restore command. 1 0 Sector IN A,(15) OUT (B0), A This register holds the desired

sector number for use during Read and Write operations

IN A,(1D) OUT (B8), A Used during Read and Write operations Data as the source and destination of data bytes. During Seeks, it holds the desired track number, towards which the head and the track-register step.

Table 2. Register structure of the floppy-disc controller. Note that these details refer to the controller in this interface. More details are given in the data sheet3.

and a head loading signal. In the opposite direction, the drive provides data in the form of interleaved clock and data pulses, the index pulse and signals to indicate that the head is over track zero, the disc drive is ready for use, and whether the disc is writeprotected. Some drives provide a headloaded signal, or can load the head onto the disc in the 10ms delay provided internally by the controller. As this drive does not, a monostable is used to provide a delay signal which is triggered by the outgoing headload line. To improve noise immunity when transferring signals to and from the electromechanical drive, the lines at both ends are pulled high via low value resistors, and high-current sinking buffers are used to drive signals to and fro.

Interrupt controller

Interrupts allow external hardware to divert the microprocessor temporarily from its stream of instructions, and accept instructions from, or more usually, to call a block of instructions which deal with the hardware's needs. In the computer, the INT line to the Z80 is driven by the MM57109 which only requires the Z80 to read (for Mk I and II systems) or transfer (for Mk III), data between the two. This is a fairly simple procedure and the interrupt mode is chosen, which causes a call to the address 0038 where, in the earlier two systems, there is a short routine to read the data. In the Mk III, there is a re-enable of the interrupt and return so that the interrupt line makes the Z80 pause until the MM57109 is ready for a data transfer. With the disc, faster and more complex responses are required because either of the two interrupting lines can become active separately or simultaneously and, depending upon the operation in hand, different responses may be required. To provide the extra flexibility without sacrificing

speed, Mode 2 interrupts must be possible, and this is carried out by the

interrupt controller in Fig. 3. In Mode 2, after the Z80 has completed its existing operation, it responds to the interrupt by asking the interrupting device to supply an 8-bit byte onto the data lines which it uses, in conjunction with the previously loaded I register, to form the address of the first of two consecutive memory locations where it will find the starting address of a subroutine to be executed. With these

two locations in the r/w.m., the disc operating system can make alterations to their contents and so alter the Z80's response to, in particular, a DRO interrupt to cope with read, write and verify operations which are at the heart of a disc system.

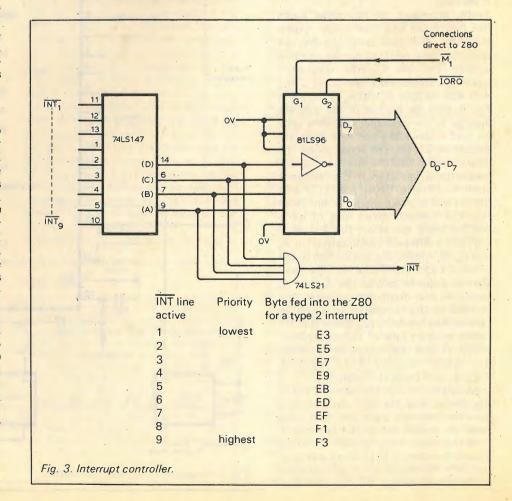
The circuit uses a 74LS147 priority encoder to generate a 4-bit code derived from the highest priority active input line. This code forms bits 1 to 4 of the byte which, when both MI and IORQ are active low, (a combination which only occurs during the Z80 interrupt acknowledge sequence) is gated onto the data lines. Direct connection to the Z80 is necessary because, during the interrupt response, both RD and WR are inactive and the bus transceiver isolates the Z80 from the main data bus.

The connection from the MM57109 to the INT pin of the Z80 must be broken and re-routed through one of the unused inputs to the interrupt controller. As the 57109 is now driving a low-power Schottky device, the pull-down resistor on this line must be changed to $10k\Omega$ connected to the -5V supply.

Part 2 will describe the controller circuit and software

References

- 1. Users' club, contact Mr P. L. Probetts, 50 Cromwell Road, Wimbledon, London SW19
- 2. Mk III monitor, contact the author at 5 The Close, Radlett, Hertfordshire WD7 8HA (Radlett 5723)
- 3. Data sheet, Western Digital FD 1771, Mar



IERS TO THE E

"CRANKY" VIEWS

WIRELESS WORLD, OCTOBER 1980

In May's letters Mr Williamson made two statements. Firstly he said that the millibel is rubbish, and secondly he said that a magazine of Wireless World's stature should not provide a platform for cranky views.

In respect of the former he may be right but in respect of the latter he is definitely not. Let us see what J. S. Mill has to say:

"To refuse a hearing to an opinion because they are certain that it is false is to assume that their certainty is the same as absolute certainty. All silencing of discussion is an assumption of infallibility. Though the silenced opinion be an error, it may, and very commonly does, contain a portion of truth; and since the prevailing opinion on any subject is rarely or never the whole truth it is only by the collision of adverse opinions that the remainder of the truth has any chance of being supplied.

Even if the received opinion be the whole truth, unless it is contested it will be held in the manner of a prejudice, with little comprehension of its rational grounds.

If there are any persons who contest a received opinion let us thank them for it. In an imperfect state of the human mind the interests of truth require a diversity of opinion." (On Liberty, abridged)

And let W. E. Weyl end the subject with a flourish:

"To every shade of thought, religious, scientific, political, economic, and social; to every craze, fad, dogma, heresy, and inspiration; there should be accorded a forum, a soap-box, a ton of type, and, subject to a subsequent responsibility for utterances, full liberty of speech and print."

(The New Democracy, 1912)

Long live cranky views.

S. Frost Edinburgh

VHF PROGRAMME LABELLING TESTS

I would be obliged if you could draw your readers' attention to the fact that the BBC are conducting experimental transmissions on their Radio 4 broadcasts from the Wrotham transmitter which can give rise to apparent interference when receiving stereo transmissions with some types of receiver, as the BBC seem to be keeping quiet about it.

Having experienced interference for the past few months which only affected Radio 4 and then only when a stereo broadcast was being received, which interference vanished when I disabled the stereo decoder, I rang the BBC's Engineering Information Department and was told that this trouble was caused by adjacent channel interference from a continental station, that I should fit an attenuator, and that there was definitely nothing wrong with their transmitted signal. I subsequently found out, quite by chance, that they are in fact making experimental transmissions from Wrotham, and on speaking to their Research Department was told that these were known to affect certain receivers, of which mine (an Alba UA800) was one. The noise on my receiver is somewhat like that of a distant diesel engine ticking over, which is

quite noticeable during quiet passages in music or speech

On ringing the BBC's Engineering Information Department a week or so later I was again told that my trouble was due to a foreign station. Only after I said that I knew that they were making experimental broadcasts which were affecting my receiver was the existence of these broadcasts admitted. and I was told that they would shortly be extended to Radio 2, as their Research Department had not had any complaints. Personally I cannot see how their Research Department could receive complaints when there has been apparently no publicity about these broadcasts, and it would appear that anyone making enquiries about interference is told that a foreign station is to blame, this despite the fact that the BBC's Research Department appear to know that problems will be experienced with certain receivers.

Some of my colleagues have reported noticing similar background noise with their receivers (various Philips and Ferguson models), but had attributed it to outside interference. It does not in any case make itself apparent very often due to the scarcity of stereo programmes on Radio 4. I would ask anyone who has noticed this effect and who has thought or been told that it is interference to contact the BBC.

I must say that I am somewhat puzzled by the BBC's approach. Although they have usually been quite open in the past about experimental transmissions, and requested feedback from the listeners, it seems that in this case feedback has been blocked for reasons best known to themselves.

R. Camp Romford Essex

The BBC replies:

May I fill in the background to the points in Mr Camp's letter?

As has been publicised in Wireless World and elsewhere, the BBC is investigating proposals for inclusion of data signals in radio broadcasts. If a suitable system can be established it could offer a number of facilities of considerable benefit to the listening public, including channel and programme identification; automatic receiver tuning and switch-on as pre-selected in advance by the listener: automatic receiver search for the type of programme desired (e.g. light music, news); and visual display at the receiver of simple text such as channel and programme title: clock time and news headlines.

Compatibility with existing receivers is clearly an important factor. BBC Research Department has carried out laboratory tests on a range of domestic receivers and these tests have been supplemented by broadcast trials wherein data signals on a 57kHz subcarrier have been included in Radio 4 v.h.f. transmissions from Wrotham

The results of these broadcast trials have been assessed by means of questionnaires issued to selected listeners, including receiver manufacturers' representatives. The trials have not been generally publicised since to do so would inevitably mean that many unrelated interference or other difficulties would be ascribed to the data signals and the trials largely vitiated.

Although the trials were unpublicised it would have been quite wrong for us to have wished to deny their existence or to give misleading advice to listeners complaining of interference. I very much regret that this happened to Mr Camp and freely apologise to him. In fact there was no intentional coverup, simply human failure in that Engineering Information Department engineers answering enquiries were insufficiently alerted to the possibility of interference from this source: that this was so is my responsibility and their suggestion of interference from other stations was reasonable since this is a common problem at this time of year.

The desirability of a system of data signalling on a subcarrier in v.h.f. transmissions has been recognised for some time in many countries. The international (CCIR) Recommendation for f.m. stereophonic broadcasting allows for the use of a 57kHz subcarrier in this way and receiver designs should be capable of handling signals according to this Recommendation: such a subcarrier is widely used in other countries. Some receivers, including Mr Camp's, are not so designed and some other older receivers may suffer some degree of interference unless their stereo multiplex decoders are very carefully aligned.

In September we shall assess very carefully the results of the current trials and Mr Camp's report on his experience will represent useful additional information in this context. We shall wish to be sure that no difficulty will be caused to receivers which are designed with the CCIR recommendation in mind, and to assess the degree of any problems experienced with older receivers. D. P. Leggatt

Head of Engineering Information BBC, London W1

MAXWELL'S EQUATIONS REVISITED

As mentioned in the May correspondence columns, we received a large number of letters commenting on Ivor Catt's article in the March issue. Our original intention was to present collectively the main points of all these letters. After discussions between the author and some of the correspondents, however, we finally decided to print one letter which was considered by an independent referee to be fairly representative. (This referee is a senior engineer in a large computer firm.) The letter chosen is followed by a direct reply and some general remarks by the author

Regarding Mr Catt's latest article, "Maxwell's equations revisited" in the March issue, I feel that he should be relieved of some of his pseudo-mathematical delusions. For example, what exactly does he mean by the equation

$$\frac{\partial h}{\partial x} \cdot \frac{\mathrm{d}x}{\mathrm{d}t} = \frac{\partial h}{\partial t} \tag{1}$$

One criticism is that dx/dt can only be used to represent the velocity of the train if x represents the x-co-ordinate of a fixed point

on it. Mr Catt originally introduced x and t as independent variables to define a point in space-time, so dx/dt is a meaningless quan-

Also, if Mr Catt had really performed a "careful analysis" he would have had great difficulty in deriving equation (1) in the first place, as anyone with even elementary knowledge of partial differential calculus could tell him. Equation (2)

$$\frac{\partial H}{\partial x} \cdot \frac{\mathrm{d}x}{\mathrm{d}t} = -\frac{\partial H}{\partial t} \tag{2}$$

falls into the same category of fallacies, Small wonder it never appears in the textbooks!

Mr Catt then goes on to say that "almost anything" is a solution to the equations

$$\frac{\partial E}{\partial x} = -\frac{\partial B}{\partial t} \tag{3}$$

$$\frac{\partial H}{\partial x} = -\frac{\partial D}{\partial t} \tag{4}$$

This, to put it mildly, is a slight exaggeration of the facts. It is a fact that a sinewave, or a number of sinewaves, is the solution of the equations given the correct boundary conditions. Mr Catt's train is also a solution of the equations but since it obeys a different set of boundary conditions it does not appear as a sinewave. More rigorously, the train profile can be considered as a Fourier series comprising an infinite number of sinewaves with different frequencies and amplitudes, and possibly also some exponential terms.

Having demonstrated the non-existence of any justification for the "theoretical" part of the article, I would like to ask the author if he has any justification for the abuse he proceeds to hurl at mathematicians in general. Mathematics is a tool for the scientist or engineer to enable him to concisely describe physical phenomena. Insight, or a "feel" for the phenomena, is built into the equations and a competent engineer should be able to "look inside" the equations and visualise what they represent. Visualisation of abstract concepts is more difficult but simply because mathematics is used as an aid in describing them does not make the theory "ludicrous and false".

Waveguides, antennae and the like are designed using Maxwell's equations, not by hit-and-miss methods, and behave as predicted by the mathematics. Electromagnetic theory is mathematical by its very nature and if Mr Catt abandons the mathematics he will be left with very little of any practical use.

R. C. Hayes University of Liverpool

The author replies: Equation (1) relates three things:

- (a) the slope of a surface,
- (b) its forward velocity,
- (c) the rate of rise of the surface.

If the slope is 1 in 4, the forward velocity 10 metres per second, then the rate of rise of the surface is 21/2 metres per second. This kind of



relationship is the stuff of which science and engineering is composed of. I think Mr Hayes knows full well what (1) means, since he has studied A-level mechanics.

Equation (2) says that if an unchanging TEM wave moves forward at the speed of light, the gradient of H with forward distance is related to the gradient of H with time. If it is a fallacy, then what is the correct formula?

Or are we not allowed to relate $\partial H/\partial x$ to $\partial H/\partial t$ for a TEM wave?

Let Mr Hayes tell mechanical engineers to convert their trains into a Fourier array of sinewayes, and see how they react! Thank God mechanical engineers are too practical to be sucked into the kind of quagmire that permeates electromagnetic theory! I do not want to travel in a train with some exponential terms designed into or out of it! Would Mr Haves recommend that the passengers be positioned so as to minimize their harmonic

Waveguides, antennae and the like are emphatically not designed using Maxwell's equations, any more than a tribal dance wins the battle that follows.

My successful pioneering attempts to interconnect high speed (1 ns) logic in Motorola in 1964 forced me to abandon all the maths that had grown like weeds to choke electromagnetic theory. A logic step is emphatically not a Fourier array of sinewaves, and you will run into all sorts of nonsense if you kid yourself that it is. Also, you can only successfully decouple the 5-volt. supply to sub-nanosecond logic because it is untrue that capacitors have stray series inductance. The regular abandonment, at vast cost, of high speed logic systems during development will only cease if we can infiltrate some common sense into electromagnetic theory, and it stops serving merely as a favourite stamping ground for physically ignorant, fancy maths obscurantists. We must take the blarney out of electromagnetic

The author also makes the following general remarks on the whole of the correspondence:

All twenty-two correspondents ignored the physics and concentrated on the mathematics. It seemed that whether Maxwell's equations mapped meaningfully and usefully onto reality mattered not. All that mattered was that the maths should be internally correct, or at least respected. An engineer like myself, who has sometimes worked as if through a blizzard of irrelevant, convoluted maths, takes the opposite view.

Some of the replies thought the minus sign should be there; some said it should not be. None noticed or contradicted my point, that the minus sign had no physical significance. (In fact it is an outgrowth of partial differentiation. Full differentiation has no minus sign, being a completely different operation from partial differentiation, in which the sign appears regardless of the nature of that which is being differentiated).

Always at a point on a surface in a three dimensional graph, the three slopes are related by

$$\frac{\partial z}{\partial x} \cdot \frac{\partial y}{\partial z} \cdot \frac{\partial x}{\partial y} = -1$$

The minus sign has nothing to do with electromagnetic theory. This contrasts with

$$\frac{\mathrm{d}x}{\mathrm{d}y} \cdot \frac{\mathrm{d}y}{\mathrm{d}z} \cdot \frac{\mathrm{d}z}{\mathrm{d}x} = +$$

which is always true of the gradients of lines in two-dimensional graphs.

IEEE 488 BUS

Mr Ellesfen's article in the June/July issue on the IEEE 488 bus standard is timely, but bitter experience convinces me that he has over-simplified things a little. The IEEE, IEC, GPIB and HP-IB systems are not all identical. Try interconnecting a 'strict' IEEE instrument (e.g. a Fluke 8502 d.v.m.) to a European GPIB bus instrument — lo and behold, the plugs are different. In fact Mr Ellesfen's Fig. 3 may show a typical GPIB rear panel but those aren't IEEE connectors. I do wish you engineers could agree on these things - it would make life a lot easier for us mere mortals! John Hennessy

Department of Physics University of Sheffield

IMPEDANCE MISMATCHING

The article "Impedance Mismatching" by Dr Lidgey in the March issue calls attention to an often overlooked point, because students fail to read the whole definition of equivalence in networks. For instance, one source says:1

"If one network can be substituted for another without change in the currents and voltages at the ports, the two networks are externally indistinguishable and are said to be equivalent at the ports. Nothing need be known of the internal network configura-

An equivalent network is not identical internally but has only identical values of external voltages and currents at the terminals. Thus comes our practice of substitution of "black boxes" to replace whole complex networks.

It is desirable to point out, as is done in the article, that power systems are not matched in impedance as 50 per cent efficiency rather raises the generation costs! But in communications where microwatts of power are very expensive, it is desirable to get out all we can, and so we match.

J. D. Ryder Ocala Florida, USA

1. J. D. Ryder, Introduction to Circuit Analysis p.175. Prentice-Hall, Englewood Cliffs, NJ 1973.

DESIGNING WITH MICROPROCESSORS

I would be grateful if you would allow me to draw attention to an error in Fig. 3 of "Designing with microprocessors" by Zissos and Valan in the May issue. The 8228 status latch shown would not produce the signals shown, since it decodes the status information placed on the data bus during the status pulse, STSTB, which is absent from the diagram. The outputs from the 8228 are much more akin to other processors, and comprise MEMR, MEMW, I/OR, I/OW, and INTA.

The diagram as shown would be correct if the status latch was a simple eight-bit latch clocked by the STSTB line.

P. B. Hodgson Grantham Lincs

The authors' reply: We thank Mr Hodgson for pointing out the omission of the STSTB strobe pulse in Fig. 3. This was intentional, for the sake of clarity. D. Zissos and L. Valan

DESIGNING WITH MICROPROCESSORS

WIRELESS WORLD, OCTOBER 1980

In their current series of articles "Designing with microprocessors" Zissos and Valan opened in the May issue with some remarks with which I profoundly disagree. They may be right, or I may be right; but either way, the questions involved are of such fundamental importance that I think both sides of the question need to be put.

To summarize, the authors put forward two views: that inefficient, machineindependent approaches are no longer justified, and that an understanding of the functioning of microprocessors is needed by their users. Let us look at this second point first.

It is obviously an advantage to understand how something works if you want to use it. It may also be more satisfying to know, but that is not the point. The question is, can a designer of something which incorporates a microprocessor do a better job if he understands how microprocessors work? I am sure that the answer is that he will probably use the microprocessor more effectively, but not by any means that he can produce a better design overall.

The problem is that designers are already overloaded with things that they ought to understand. Adding in the understanding of microprocessors is not a simple plus: it almost certainly means forgoing the attempt to understand something else.

Take the case where a microprocessorbased device is used as a circuit component - say an 'intelligent' filter which can discriminate between signal and noise far better than any array of passive components. The things that a designer could usefully come to understand better are how that filter itself works, or what its effect on the circuit as a whole will be. Understanding the filter may open up new opportunities, or avoid mistakes. Understanding the effects on the circuit will involve seeing the effects of a fixed, rather than frequency-dependent, delay, and may likewise open up opportunities or avoid mistakes. If the designer has time to take both on board, all well and good. More likely, though, something has got to go. Which? Well, the suppliers of the microbased filter can tell him the salient features of the device as a whole. But who can tell him the salient features of its impact on the circuit he is designing? Nobody. He must sort that out himself.

The situation gets even more acute where a micro is used as a systems component, say as part of office equipment. There the range of aspects that a designer needs to understand gets very wide indeed - including marketability, the psychology of the operator, safety legislation, and so on. The poor designer can hardly keep his head above water. Let's not load him even more.

What the designer needs to know is what micros can do, not how they do it. One day I really must learn how transistors work (I grew up on valves). Even so. I seem to be able to use transistors, because I can find documentation which tells me what the result is.

I am not objecting to the inclusion of articles describing how micros work - I, like most readers of Wireless World I suspect, am a person who likes to know how things work. But those articles should not contain the implication that such knowledge is the best way to approach their use.

That argument pales into insignificance beside the one of machine independence. It

took the 'big computer' world a decade or two to learn the lesson of machine independence. Now the microprocessor world seems determined to repeat the mistakes, against an economic basis that is far less tolerant of such mistakes.

Several years ago a fair size mainframe might cost £1,000,000. The application programs which used the machine probably cost somewhat less, but not much less. Originally, they were written to suit the characteristics of the machine. Provided that no major growth required intermediate change, replacement might be needed in, say, seven years. Then there was good news and bad news. The good news was that the replacement cost less (despite inflation) than the predecessor. The bad news was that modifying the software would cost far more, and would be such a major undertaking that the resources would not be available to make the change. Thus we see people buying ICL 2900s, which are something like a hundred times as fast as some machines they replace, but running programs slower than their predecessors, because the old programs have to be run, and when a 2900 pretends that it is an older machine it is very inefficient indeed.

This is actually rather a favourable example! As the 2900 is a replacement for earlier ICL machines, and as it is a major cost item, the incentive and the money to provide the software for it to pretend to be an earlier machine have been there. In other cases users have been left in a real mess.

How do micros compare? For a start, the cost of programs typically far outweighs the cost of the device itself. So replacing the hardware is almost trivial in cost terms. But if the software is tied to the hardware the resultant software change cost can be crippling. Microprocessors are only cheap if they are cheap to use; if the way in which they are used brings crippling costs, they are extraordinarily expensive.

The situation is made worse by the rate of change in microprocessors. With the mainframe instanced above the effective life was the life of the machinery. When micros are used in products, so that it is the product manufacturer who develops the software, the life is from the time he starts buying one version to the time that that version is superseded. Currently that timescale appears to be about nine months.

Put it another way. If you develop a prototype of a product incorporating microprocessors, the chances are that by the time you go into production that particular version will be obsolete. Later, when an assembly fails in service and requires replacement, the production chip originally incorporated will no longer be available. If the software is linked to a particular chip, you've got problems, and pretty severe ones. Among them is the fact that the people who designed the original software (which will no longer work on the latest version of the micro) might no longer be around, so that the new designers have to start from scratch. By the time that they have finished their work, a new version of the micro will already be on its way.

The large computer users have already learnt that there are considerable risks associated with machine dependent designs. The factors which have forced them to start using inefficient approaches to avoid being trapped are far more potent in the micro world. Far better endure the 50%, 100% or even 200% overhead of machine-independent software than be trapped in a short time by the 10,000% overhead instanced above.

What it comes down to is this. If you want

to use a microprocessor as an educational toy, then learn how it works, and write programs which exercise the individual capabilities of different chips. It is a fascinating pastime. But if you want micros as a serious tool, think again. If you have got some time to spare, use it on thinking how the tool can help you, how it will affect the system where you use it. And whatever happens, try to avoid getting so tied up with a particular version (which will probably be obsolescent by the time it comes on the market) that very shortly you will find yourself occupied full time in just keeping things working. R. M. O'Connor

Herts The authors reply:

St Albans

We must disagree with Mr O'Connor's contention that a better overall design cannot be produced by the understanding of how microprocessors work, as will be demonstrated in later articles. This understanding can be gained by first-year Computer Science or Electrical Engineering undergraduates in two to three 50 minute lectures.

Because of the methodology available today it is unnecessary for a designer to feel overloaded. Much of the stress experienced is normally caused by lack of knowledge of the idiosyncrasies of v.l.s.i. chips. Although this lack of knowledge may be tolerated in users, it should not be acceptable in doers.

The analogy drawn with computers of a few years ago is invalid, because we now have step-by-step procedures for designing and implementing digital systems that did not exist then.

A closer look at the 'new' microprocessors shows no fundamental changes in the basic structure, and therefore they can be easily accommodated in systematically designed systems. The figure of 10,000% overhead is simply unrealistic or a result of a bad design. D. Zissos and L. Valan

TV SETS FOR THE HARD OF HEARING

My hearing is poor and over the years I have found that I can best listen to the radio or television via a pair of headphones. Radio is, of course, no problem. But when I wished to purchase a television set with an outlet for 'phones I was just met with blank stares!

I am retired, and was in no hurry, so I went around all the television retailers in my area just to see what there was. The young assistants just did not want to know. Apart from one helpful dealer the only other trader willing to help was a stockist of exclusively Japanese products. The Japanese tv sets mostly had outlets fitted as standard for tape recording the sound, and for headphones. On some, inserting the headphone plug cut out the loudspeaker, on others there was a choice of cutting out or not cutting out the loud-

I bought one! What would you do? A circuit diagram was included and looking at it I concluded that the additional cost of fitting the outlets would be around £5.

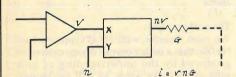
About ten per cent of the population have hearing problems, while five per cent have serious hearing problems requiring some form of assistance. On this basis it seems to me that someone is missing out on sales. Fred Holloway

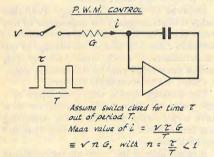
Essex League of the Hard of Hearing Rayleigh, Essex

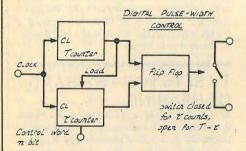
Pulse control of analogue functions

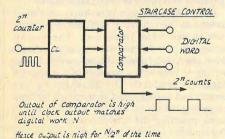
by Peter Williams, Ph.D. Paisley College of Technology

ANALOGUE MULTIPLIER CONTROL









BINARY-RATE - MULTIPLIER CONTROL

B.R.M. is a special purpose divider giving N/2" output pulses. Hence Switch is closed for N/3" of the time

Any other device interposed between one integrator and the next, having a controlled transfer function, will vary the frequency of the oscillator/filter. Analogue multipliers are designed to have an output proportional to the product of a pair of inputs. Interposing a pair of such multipliers with one input of each fed from a control voltage results in linear control of the frequency. The multiplier is being used for the restricted function of a gain-controlled amplifier, other forms of which may be substituted. Only two-quadrant operation is required as the control voltage is unipolar. Variable transconductance circuits can also be used. In some cases the output is in the form of a current and the following drive resistor may be omitted. Overall stability may be affected by the additional lags introduced by the multipliers though with standard operational amplifiers, the dominant lags caused by internal compensating capacitors are likely to affect the response first.

A completely different approach is possible if the nature of the integrator is reconsidered. The voltage across the capacitor depends on the total charge and not on the manner in which that charge is acquired. This suggests that the current may be allowed to flow in short controlled bursts provided that the switching rate is high enough to minimize the ripple voltage that is inevitably superposed on the output. This is essentially pulse-width modulation, and n is replaced by τ/T, where τ is the pulse duration and T the pulse period. In the first configuration an analogue switch is assumed to be repetitively closed at a frequency of 1/T with τ being varied to control the mean current over the cycle. Typically, for an oscillator frequency f then 1/T > 10f would be preferable for minimum ripple. As analogue switches operating up to the MHz region are readily available this places little constraint on the usable signal frequencies. As before any number of sections can be used with the switches drive in synchronism from a common generator — the last consisting typically of a fixed frequency astable driving a variable

The accuracy and resolution obtainable in the previous methods are restricted by the analogue sections of the circuit. Purely digital methods are possible for the control section which can increase the resolution without limit. One method is shown in which two interlinked counters are used to determine τ and T each as multiples of the period of a clock generator. The ratio T/T is thus independent of any timing circuitry, being the ratio of two integers either one of which may be preset or controlled by an external digital control word. The T counter sets the flip flop and starts the τ counter. At the end of the τ counter duration, e.g. counting down to zero from the previously loaded control word, the flip flop is reset until the start of the next period. In this particular case the switch is closed for counts and open for T - T counts, and any number of switches can in principle be driven from the given flip-flop.

As an illustration of the range of possibilities, consider the circuit shown, with a binary comparator driven from a 2rd binary counter. The output of the comparator is high until the counter output matches the control word N applied to the comparator. For the rest of the time the comparator output is driven low. Hence any switch activated by the counter will be on for $N/2^n$ of the time and the circuit would be capable of controlling a filter or oscillator such that f \propto N. A restriction on this and the previous form is the difficulty of filtering the switching waveform if the switching frequency becomes too low. It is desirable that $f_{clock}/\sqrt{2^n}>> f$. Any other switching circuit that can close a switch for a controlled fraction of the time can be employed, and the method has been employed as a precision voltage divider for measurement

One particular class of digital circuits seems particularly appropriate to this application binary-rate multipliers and dividers. In the former a digital control word N directly sets the number of pulses that are transmitted during a complete cycle of operation of 2ⁿ input pulses. Moreover these circuits can be combined to give output pulse rates that are complex functions of the input control numbers if required. Because the on and off states are distributed throughout the cycle the filtering problem is a little less severe, though they are not randomly distributed. The main advantage of these circuits is that no additional comparator or decoding action is required and the output can be applied directly to an analogue switch or switches. CMOS logic is the obvious family to use since analogue switches are readily available that are compatible with the b.r.ms. The restricted supply voltage range (typically ±7.5V) restricts the associated analogue circuitry to the same value

WIRELESS WORLD, OCTOBER 1980

Pulse control of analogue functions

THEORY

 The first method strictly belongs to the previous section in that the analogue multiplier produces a continuous variation in the scaling factor by applying a control voltage to one of the inputs

where n represents the control voltage on the Y input. The Y input may however receive a discontinuous voltage switching between zero and some maximum value but with a variable mark-space ratio: Provided the switching frequency is high the mean value of n is controlled by that mark-space ratio

If discontinuous control is adopted the analogue-multiplier can be omitted with the voltage applied to the integrator through a switch. Let the switch be closed for a time τ out of a period T.

Mean
$$\frac{i}{v} = \frac{\tau}{T}$$
. $G \equiv nG$
for $n = \frac{\tau}{T} < 1$

The ripple voltage across the capacitor is minimized for $T << 1/f_1$ where $f = \omega_L/2\pi$. is the lowest sinusoidal frequency to be controlled and $f_i \circ \circ_{\tau}$.

 The second counter, a down-counter, is loaded with a control word when the first counter fills and sets the flip-flop. Until the second counter empties the switch is held closed for a time τ proportional to the control word. For the remaining time until the first counter fills the switch remains open because of the resetting action on the flip-flop by the second-counter

Mark/space ratio is

$$\frac{\tau}{T-\tau}$$

 A related method compares the output of a single 2º counter with a control word N in a binary comparator such that the output is high for N/2ⁿ of the time. This is again available to control the on-off ratio of a set of switches.

For the bit-rate multiplier the internal logic causes an output pulse

or mark-space ratio

$$N: (2^n - N).$$

EXAMPLES

1. A two-integrator loop has the resistors switched into and out of conduction periodically — the period of the switching waveform is T and the conduction-time per switching cycle is τ. Given R is 100kΩ and C 10nF, calculate the frequency of oscillation for $\tau = T$ and $\tau =$ 0.1T assuming that that frequency for << 1/T i.e. that the ripple superimposed by the switching frequency on the sinusoidal oscillation

The effective value of each conductance is $G' = \frac{G'}{G}$

e. R' =
$$\frac{T}{\tau G} = \frac{TR}{\tau}$$

Therefore following the previous analysis, and assuming a unity gain

$$\omega_0 = \frac{1}{CR'} = \frac{\tau}{TCR}$$
i.e. $\omega_0' = \frac{\tau}{T}\omega_0$
For $\tau = T$ $f = \frac{1}{2\pi.10.10^{-9}.10^5}$
 $\therefore f = 159$ Hz.
For $\tau = 0.1T$ $f' = 0.1f = 15.9$ Hz

2. In the previous question a switching frequency of 1.59kHz is used Estimate the peak ripple as a fraction of the sinusoidal amplitude.

(i) $\tau = T$: the switch is closed permanently and there is no ripple due to the switching. If τ is finite but very close to T, then the switch is off for very brief periods and the capacitor voltages remain constant for that

(ii) $\tau = 0.1T$: the effect is that of a rectangular wave at a frequency of 10kHz with an amplitude equal to the value of the slowly changing output of the previous integrator i.e. the output Vsint of one is applied discontinuously to the second whose output approximates to -Vcos0. Hence the max. ripple on the output of the second occurs when $\sin\theta =$ 1 i.e. when the second integrator output is passing through zero. Note that Vsint itself contains a small ripple but this error is ignored. The peak ripple is thus

$$\frac{1}{R} \cdot \frac{\tau}{C} = V \tau \omega_0$$

or expressed as a fraction of the peak output, $V_x \omega_0 / V = 0.1 \omega_0 T$

But T =
$$\frac{1}{10^4}$$
 s

$$\omega_0 = 10^3 \text{ rad/s}$$

WIRELESS WORLD, OCTOBER 1980

the stationary one, a difference of ages

on reunion would require Peter and

Paul each to be the older at that event.

This is impossible, so asymmetrical

ageing cannot occur, and it is up to

those who claim that it can to discover

their error. (Though not among them I

once thought I had done this1, but later

found that my argument failed, though

not for the reasons alleged by my critics

at the time2. That left me with no

alternative to rejection of the theory.

However, we are for the present regar-

The arguments for asymmetrical

ageing - by far the most favoured

alternative - are legion but only one

calls for serious attention, namely that

given first by Einstein himself³ and

supported by Born⁴, Tolman⁵, Pauli⁶,

among others, and elaborated in detail

by Moller⁷ and Born & Biem⁸. No other

has a weight of authority behind it

comparable with this, or indeed when

examined carries any conviction at all.

while, granting the validity of the

theories, every step in this argument is

irresistible. I shall therefore consider it

The essence of this argument is that,

indeed, during the main part of the

whole journey - that at constant

velocity in both directions - the moving

twin, whether he be regarded as Paul or

Peter must be held to age more slowly

than the stationary one, but if Paul is

regarded as stationary, then the field of

force* must be assumed to exist every-

where during the period of reversal of

motion to keep Paul at rest, despite the

impulse given him by the working of the

engine of his vehicle, and also to bring

Peter back to him although no such

impulse is given to Peter. The effect of

this force-field is to make Peter, during

the period of reversal of motion, gain so

much in age, and Paul to lose so much,

as to far outweigh Paul's more rapid

ageing during the uniform motions, and

ultimately to give the same ages of the

twins on reunion as those following

from the assumption that Paul, and not

Peter, is the one who moves. The calcu-

lation in this last case is simple. No

force-field is required, since the engine

suffices to reverse Paul's motion and

Peter does not move, so Peter's gain

during the periods of uniform motion is

the sole effect. (Incidentally, when

Peter moves similar force-fields are

needed to accelerate him initially from

rest to velocity v and to bring him to rest

again at the end, and at these events

Peter and Paul are virtually at the same

place, and the general theory requires

that in such circumstances the dif-

ference in the effects on ageing is negli-

Now let us apply this to our example.

If Paul is the traveller he ages by 1.5

days during each of his outward and

return journeys and by a negligible

*It is called a gravitational field but this is mis-

leading because it must be granted properties not

found in natural gravitational fields.

gible.)

ding it as true.)

alone.

"What is long overdue is a general summing up of the whole matter, so that the source of the scandal can be located and removed without futile polemic."

by the late Herbert Dingle

In Nature, volume 269, page 284 (22 September 1977) I put a question to Dr Tom Wilkie concerning an often advanced suggestion he had repeated for disposing of the twin paradox of relativity. He did not reply, but added a note stating that he would be writing me "privately" on the matter. It was, of course, entirely proper that misunderstandings should be removed before the reply appeared, but although after considerable correspondence this seems to have been achieved, Dr Wilkie has not accepted my invitation to him now to publish his reply. It does not accord with recognised scientific practice that questions considered worthy of publication should remain without published answer, and it therefore become my duty to comment publicly myself on the implications of this incident.

But far more is involved here than the incident itself. The so-called "twin paradox" has been the centre of more or less continuous controversy for more than half a century, and still remains unsettled. Because of its peculiar - I. believe unique - character it is no exaggeration, but a considered temperate statement, to call this a scandal, for reasons which I shall show, and what is long overdue is a general summing-up of the whole matter so that the source of the scandal may be located and removed without further futile polemic. This is attempted here and the uniqueness of the problem in scientific discussions made clear, but first it is desirable, notwithstanding its familiarity, to state what the "paradox" is in its simplest form. For brevity and clarity a particular extreme example that given in my letter in Nature of 31 August 1973 - is chosen. There is no disagreement about the legitimacy and typicality of this example, and therefore no begging of the question in selecting

Peter and Paul are twins, of whom Paul travels at birth with uniform velocity v to a distant planet stationary with respect to the Earth, and immediately returns at the same ve.ocity having aged by three days, to find his Earthbound twin, Peter, 30 years old. In the general case, any two identical forms of standard clock may be substituted for the twins, and if t is "Peter's" age when "Paul" returns, "Paul's" age at that event is $t\sqrt{1-v^2/c^2}$, where c is the velocity of light. It is evident that in this example v must be very slightly less than c.

Now the peculiar nature of this "paradox" lies in the fact that this has never been observed in any form: the result is wholly a deduction from a theory. In all the traditional controversies or paradoxes of science - the Ptolemaic and Copernican theories of celestial motions, the wave or particle nature of light, etc. - the problem has arisen from observations, and what has called for decision has been the correct theory for explaining them. Here the reverse is the case. The dispute is not

"Failure to agree on the implications of our own constructed and accepted theories is not excusable. This is what makes the endless persistence of this controversy a scandal."

about what theory best explains the observations (which do not exist), but about what observation - equal or unequal ages of the twins on reunion - is required by an independently accepted theory. And indeed there is an allied peculiarity in that if the observation were made, its result, whatever it might be, would still leave the problem unsolved: the question would still remain, what is wrong with the deduction from the theory of the opposite result? The problem, therefore, lies within the theory itself. Failure to understand the course of nature is excusable and observation of nature might be expected to bring enlightenment: failure to agree on the implications of our own constructed and accepted theories is not excusable. That is what makes the endless persistence of this controversy a scandal.

Let us assume that the theory is true, and give the net results of the arguments for its requirement of symmetrical and asymmetrical ageing, respectively. The first can be stated very simply. The relativity theories, both special and general, require that it is equally true to regard Paul as moving with respect to a stationary Peter and planet, and Peter and planet as moving with respect to a stationary Paul. Therefore if, as the theories also require, the moving twin ages more slowly than

Herbert Dingle 2 August 1890 — 4 September 1978

scientist and philosopher who was Professor of History and Philosophy of Science at University College, London, from 1946 to 1955. Before that he was Professor of Natural Philosophy at Imperial College, from which he had graduated in 1918. His numerous scientific distinctions included the presidency of the Royal Astronomical Society (1951-53) and of the British Society for the History of Science (1955-57); he wrote several well-known books and an enormous number of scientific papers.

The early part of Professor Dingle's scientific career was a period of intense interest in relativity, and he became an expert on the subject. was sceptical about the well-known the paradox, during the 1950s, he impeccable, was physically imposanswer; were not always well

The accompanying paper is Professor Dingle's final summing-up of his views on the twin paradox. He sent me the manuscript a few months before his death, in the hope that I would be able to have it published, and I commend it to scientists in the hope that they will give it the serious

I am grateful to Mrs Pamela Dingle for giving her permission for this paper to be published, and to Wireless World for publishing it.

Ian McCausland University of Toronto

Herbert Dingle was a distinguished

Although an admirer of the theory, he clock paradox or twin paradox, and did not agree with its generallyaccepted resolution. After a prominent but inconclusive debate on became convinced that the special theory, though mathematically sible, and he spent much of his time and energy during the last 20 years of his life trying to persuade the scientific world that the theory was untenable. His criticisms of the theory, and his Socratic ability to ask questions that nobody else could received; I have suggested elsewhere (Canadian Electrical Engineering Journal, April 1980) that his thesis has not been satisfactorily answered.

attention that it deserves.

Why not discuss relativity?

After the accompanying article by Herbert Dingle had been submitted for publication, there appeared an article in New Scientist1, by Paul Davies, bearing the title "Why pick on Einstein?". This article defends relativity from its critics by presenting some of the evidence that is claimed to support the theory. As the only critic mentioned by name is Herbert Dingle, who is not able to answer back, I am briefly replying on his behalf. Because Professor Dingle has already presented the arguments in question himself, I shall not re-state them, but merely indicate the general nature of the arguments and cite appropriate references. I think the fact that Professor Davies does not even mention these arguments is evidence that they have not received enough attention from the scientific community.

Professor Davies mentions the Michelson-Morley experiment, stating that it consisted of comparing the times that light pulses travelling in perpendicular directions took to cross the same distance. But, as Dingle pointed out on various occasions2, 3, 4, the experiment did not involve a direct measurement of time; the time comparison has been inferred from interference patterns. Of the possible interpretations of the experiment, one is that Newton's laws of motion are wrong, another is that Maxwell's electromagnetic theory is wrong: the usual interpretation of the experiment, in which the time difference is deduced using Maxwell's theory, eliminates in advance the interpretation that that theory is wrong. This illegitimate elimination of one of the possible interpretations of the experiment rules out that experiment as evidence in support of the special theory.

Dingle has also pointed out 3, 4 that, in experiments that involve elementary particles moving at very high speeds, the speeds of the particles are not measured directly but are inferred from certain observations by a process that involves the use of Maxwell's electromagnetic theory; this fact also rules out experiments of this kind as evidence in support of special relativity.

Professor Dingle⁴ has also questioned observations of double stars as evidence supporting the special theory. Although one of his hypotheses - that light travels at constant velocity with respect to its own source, however the source may move - may seem rather difficult to accept, it is surely no more difficult to accept than some of the other phenomena that many physicists appear to believe in. The hypothesis is based on a suggestion already made by Faraday, and if it were true it would also, according to Moon and Spencer⁵, allow clocks to be synchronised regardless of their state of motion. Dingle has also suggested that more attention should be devoted to the work of Ritz, whom he mentions in his article and whose work has recently been discussed by Waldron6.

Furthermore, according to Dingle4, all the experimental evidence that is taken to support the special theory could with equal validity be taken to support Lorentz's quite different theory if Einstein's special theory had never been. conceived. In another New Scientist article, Roxburgh⁷ appeared to agree with this when he stated that Einstein's thoery and Lorentz's were "observationally indistinguishable."

In his book Space and time in the modern universe, Professor Davies8 makes the following statement in connection with two clock-carrying observers in uniform relative motion: "It is not that each observer merely sees the other clock running slow, it actually is running slow - a real physical effect." [Emphasis in the original]. This statement seems to me to provide strong, support for Dingle's claim that, if there are two clocks in uniform relative motion, the special theory requires each clock to run (not merely seem to run) faster than the other.

The heading of the New Scientist article1 uses the term "scientific malcontents" to refer to those who attack relativity. If being a relativist entails acceptance of all the mutuallycontradictory arguments (some of which I have recently documented9) that have been published in defending special relativity against the criticism of Herbert Dingle, then I prefer to be a scientific malcontent, and I accept that designation with pride. I think every scientist should be a malcontent; after all, what is the value of trying to contribute new knowledge unless one is dissatisfied with the present state of knowledge?

I could write at length about my encounters with what Davies calls the "special provision" that most editors of science journals make for coping with papers of the type he describes, but this is not the time or the place. In any case, Dingle has described his own experiences so eloquently4 that it is scarcely necessary to augment his description, but it is noteworthy that a supporter of relativity has now stated openly that most editors of scientific journals do make such special provision; it is not merely a figment of the critics' imaginations. Others who have encountered the "special provision" may tend to agree with me in thinking that the question in the heading of the New Scientist article should be amended to read: Why is criticism of relativity so resented?

Ian McCausland

- 1 Davies, P. Why pick on Einstein? New Scientist, vol. 87, 7 August 1980, pp.463-5.
- 2 Dingle, H. A Re-examination of the Michelson-Morley Experiment. Vistas in Astronomy, vol.9, 1967, pp.97-100.
- 3 Ref. 2 of the accompanying article
- 4 Dingle, H. Science at the Crossroads. Martin Brian & O'Keeffe, 1972.
- 5 Moon, P., and Spencer, D. E. On the establishment of a universal time. Philosophy of Science, vol.23, 1956, pp.216-29.
- 6 Waldron, R. A. The Electrodynamics of Ritz. Speculations in Science and Technology, vol.2, August 1979, pp.259-71.
- 7 Roxburgh, I. Is special relativity right or wrong? New Scientist, vol.55, 28 September 1972, p.602.
- 8 Davies, P.C.W. Space and time in the' modern universe. Cambridge University Press, 1977, p.39.
- 9 McCausland, I. Science on the defensive. Canadian Electrical Engineering Journal, vol.5, April 1980, pp.3&4.

amount during the three periods of acceleration, while Peter ages regularly by 30 years during the complete process; hence, when they meet again, Peter's age is 30 years and Paul's three days. On the other hand, if Peter is the traveller he ages by 1.5 days during each of his outward and return journeys, and by almost 30 years during the change from recession to approach with respect to Paul, while the stationary Paul ages by 15 years during Peter's outward journey, changes during Peter's reversal to a state nearly 15 years before birth, and then ages by 15 years during Peter's return, somehow getting born shortly before Peter arrives. Consequently, when they meet, Peter's age is 30 years and Paul's three days - exactly as in the former case.

We can hardly suppose that Einstein, Born and the others believed that these processes were both actual occurrances, the one entitled to claim reality depending on our preference in choosing to whom to assign the motion, nor did they. What they supposed was that the only observable events in the whole process were the separation of Peter and Paul at the beginning and their reunion at the end. Everything that happened in between was regarded as being beyond possibility of observation and therefore demanding compatibility only with theory, not with experience, with which it had nothing to do. This is obviously so important that it is necessary to confirm it by quoting Einstein's own words (in translation), all that needs explanation being that the clock U, is Peter and U, Paul and that "the right and left hand columns" give the descriptions of the process, as I have described them, when Peter and Paul, respectively, are regarded as moving. Einstein writes³:

"You must bear in mind that exactly the same process is described in the right and in the left hand columns, but the description on the left refers to the coordinate system K while that on the right refers to K'. According to both descriptions, at the end of the process the clock U, is retarded by a definite amount compared with U1. With reference to K' this is explained as follows: it is true that during the stages 2 and 4, the clock U1, moving with velocity v, works more slowly than U2, which is at rest. But this retardation is over-compensated by the quicker working of U1 during stage 3. For, according to the general theory of relativity, the clock works the faster the higher the gravitational potential at the place where it is situated, and during stage 3 U, is indeed situated in a region of higher gravitational potential than U2. Calculation shows that the consequent advancement amounts to exactly twice as much as the retardation during stages 2 and 4. This completely clears up the paradox."

What Einstein means here by "the same process" is, of course, everything that is observable, while "the description", which differs in the two cases, is wholly a mental construction. The first is unique, for it must be the one thing

that would actually occur; the last owes allegiance only to theory, not observation, and can vary within the limits allowed by the theory.

But it is clear, beyond possibility of question, that Einstein's "descriptions" relate to what is observable, and cannot therefore both be permissible; and furthermore, as the credentials of both are exactly the same, it is impossible to decide which must be rejected. Paul could be accompanied by a nurse, of such an age as to become 30 years younger without losing her power of intelligent observation, and she would report on return whether it was a baby or a teenage boy who arrived at the planet, and whether or not a baby was born during the return journey, even if she were unable to confirm the antenatal age of the being whom the planet left. The question I asked Dr Wilkie was, in effect, whether what the nurse would observe would admit of both of Einstein's "descriptions", or whether a

"... Mathematical consistency, though a necessary condition, is not a sufficient one for the truth of a physical theory."

theory that required it to do so must be abandoned. I am not surprised at his reluctance to commit himself to a choice; nevertheless, it is imperative that scientists shall make a choice if the ethical demands of science are not to be jettisoned.

What is the net result of all this? As I have said, it throws no light at all on what would happen if the experiment were made, for it is an analysis, not of a physical process that has never occurred, but of the requirements of a theory that purports to accord with physical processes, and I think it shows beyond doubt that the special relativity theory at least must be wrong. If the motion can be ascribed equally rightly to either twin, it cannot make them age at different rates; if it makes them age at different rates, there must be an absolute standard of rest to provide a criterion for distinguishing the faster from the slower developer. The special relativity theory requires different rates of ageing to result from motion which belongs no more to one twin than to the other: that is impossible.

It is impossible to exaggerate the importance of this result, for this theory is, by common consent, "taken for granted" in Max Born's words, in all modern atomic research, and it determines the course of practically all current developments in physical science, theoretical and experimental, whether concerned with the laboratory or with the universe. To continue to use the theory without discrimination, therefore, is not only to follow a false trail in the investigation of nature, but also to

risk physical disaster on the unforeseeable scale, modern atomic experiments being what they are. It should therefore be a point of honour with those on whose authority atomic research is now being conducted to acknowledge at once the untenability of the theory, and to take without delay the necessary steps to discover where the theory falls.

WIRELESS WORLD, OCTOBER 1980

That does not necessarily mean complete abandoning of its use, but it does demand the determination of the limits of its usefulness. It has already proved its effectiveness in many respects, and this has been mistaken by physicists for evidence of its truth. What the many successes of the Lorentz transformation equations have shown is that those equations are an effectual corrective of the imperfect classical electromagnetic equations within a limited range of experience. But it is now clear that the interpretation of those equations as constituting a basis for a new kinematics, displacing that of Galileo and Newton, which is the essence of the special relativity theory, leads inevitably to impossibilities and therefore cannot be true. Either there is an absolute standard of rest – call it the ether as with Maxwell, or the universe as with Mach, or absolute space as with Newton, or what you will - or else all motion, including that with the speed of light, is relative, as with Ritz. It remains to be determined, by a valid experimental determination of the true relation of the velocity of light to that of its source, which of these alternatives is the true one. In the meantime, the fiction of "space-time" as an objective element of nature, and the associated pseudo-concepts such as "time-dilation", that violate "saving common sense", should be discharged from physics and philosophy, and the fact realised that mathematical consistency, though a necessary condition, is not a sufficient one for the truth of a physical theory. Only thus can the scandal of more than half a century of confusion about the meaning of our' own creations be ended.

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Audio gain controls

A survey of the methods used to achieve acceptable control of gain in audio amplifiers.

by Peter Baxandall B.Sc. (Eng.), F.I.E.E., F.I.E.R.E., M.A.E.S., F.B.K.S.T.S.

The design of gain controls is by no means as simple as it might appear. Peter Baxandall examines the difficulties in design and comments on many circuits which have appeared over the years, from very simple types in which compromises must be accepted, to those used in high-performance equipment.

An ideal audio amplifier with variable gain would have the following characteristics:

(i) noise output voltage = (source Johnson noise voltage) × (gain)
(ii) ability to deliver its full output voltage even at very low gain settings, which may be less than unity. The amplifier is therefore capable of han-

dling very large input voltages at the lowest gain settings.

The simplest way to achieve (i) is that shown in Fig. 1(a), but this simple tech-

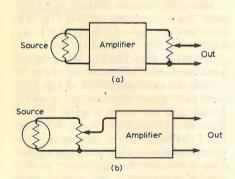


Fig. 1. Two small gain controls which do not fulfil both main requirements. Circuit (a) gives low noise, but will overload at low gain settings, while (b) introduces additional noise.

nique obviously fails lamentably with regard to (ii), for the maximum input voltage that can be handled without overloading is the same at all settings. The arrangement of Fig. 1(b), on the other hand, achieves (ii) perfectly, but fails with regard to (i).

By using sufficiently subtle gaincontrol systems, it is possible to satisfy (i) and (ii) concurrently and almost perfectly, but the simple and widely used arrangement shown in Fig. 2 provides a compromise solution which is very satisfactory for many practical purposes.

An ideal amplifier would give a variation of output noise voltage with

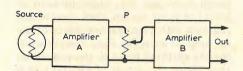


Fig. 2. This arrangement combines circuits of Fig. 1(a) and (b) to give compromise performance.

gain setting as shown by the full-line graph in Fig. 3, whereas the Fig. 2 scheme gives a characteristic as depicted by the broken line. Below a certain setting of P, the noise level from amplifier A at the output of P becomes less than the noise level of amplifier B referred to its input, so that the noise of amplifier B becomes the dominant contribution, establishing the level of the broken-line "plateau".

Now there is obviously no practical advantage in achieving an output noise level which is a long way below audibility at very low gain settings, so that a characteristic of the broken-line type is normally perfectly satisfactory, provided the level of the horizontal plateau is low enough. For a given overall maximum gain requirement, the product of the voltage gains of amplifiers A and B in Fig. 2 is fixed, but there is a choice with regard to the apportionment of this gain between the two amplifiers. The higher the gain of A is made, the lower is the position of the Fig. 3 plateau, but there is the disadvantage that the maximum signal input that can be handled without overloading amplifier A is reduced.

In domestic audio control units, the

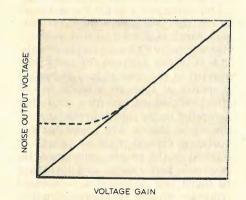


Fig. 3. Dotted line shows gain variation given by circuit of Fig. 2, where residual noise from amplifier B is predominant at low gain settings.

Fig. 2 arrangement is usually used. A suitable choice for the gain of amplifier B is normally such that full output level is delivered to the following power amplifier for an output level from the pot. slider of about 100mV r.m.s. If the: wideband noise of amplifier B, with P set to zero, is equivalent to a noise input voltage to B of 0.5 µV rms, which is fairly readily achievable, the zero-volumesetting noise output from B will then be 106dB below the full signal output level. (It may be added, however, that if the gain of B is made high enough to cope with the least sensitive of power amplifiers, which may require an input level of several volts, then the signal level at the pot. slider for full power output when used with a very-highsensitivity power amplifier, will be much less than 100mV rms, and a figure much less than the 106dB mentioned above will then apply. Thus, for versatile use, it is desirable to provide a preset gain adjustment within amplifier B, or in the form of a simple passive attenuator after this amplifier.)

A closer approximation to concurrently satisfying conditions (i) and (ii) at the beginning of the article may be obtained, on the same principle as in Fig. 2, by employing three amplifiers with ganged gain-control pots. between them, but in general it is much better, instead, to employ schemes in which variable negative feedback provides much of the gain variation.

Variable-feedback gain control offers advantages both with regard to achievable performance (noise and signalhandling) and often with regard to economy of circuit design. Variable feedback alone cannot normally reduce the gain to zero; for 100% voltage feedback reduces the gain to unity rather than zero. Thus, it is usual to combine feedback variation with passive gain control, sometimes using a ganged pot. and sometimes using the parts of the track either side of the slider, in an ordinary single pot., to perform these functions. There are many possible schemes, of which some have been known for thirty years or more.

One of the simplest schemes is that shown in Fig. 4. The pot. resistance can be made quite low, e.g. $1k\Omega$, since it is driven by the op. amp., not the signal source. This results in a good noise performance at all settings. Disadvantages of the circuit are:

(a) the minimum gain is unity, not zero,

(b) a floating signal source is required.

Disadvantage (b) is of little consequence when an input transformer is used, and (a) may be overcome by taking the signal output from the pot. slider. The latter change, of course, sacrifices the virtue of very low output impedance possessed by the Fig. 4 ver-

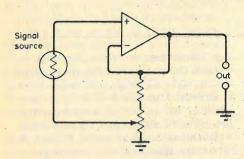


Fig. 4. Simple feedback gain control.

In assessing the pros and cons of various circuits, it is very helpful to appreciate the relationships between the circuits in the most vivid possible way, rather than relying purely on formal analysis. Very often the differences between circuits are much smaller than they appear to be, involving merely the choice of earthing point and/or the way of drawing the circuit diagram, rather than differences of more fundamental significance. Sometimes, in redrawing circuits employing op. amps. to facilitate better understanding of them, it is helpful to replace the op. amp. symbols by ordinary single-transistor symbols - an unfamiliar-looking circuit may then suddenly be recognised as an old friend! At other times, replacing a detailed transistor circuit by the op. amp. equivalent may reveal its true nature in the best way.

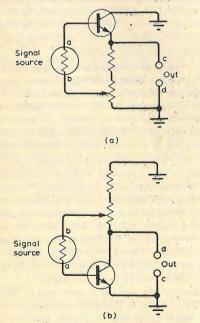


Fig. 5. Single transistor equivalent to Fig. 4, neglecting d.c. conditions. Rearrangement in (b) shows circuit to be easily recognizable.

On replacing the op. amp. in Fig. 4 by a transistor, the circuit of Fig. 5(a) is obtained. Though the collector would in practice be taken to a positive supply line, it is here shown as earthed, for in the present context we are concerned only with a.c. aspects and it is best to omit irrelevant details.

Shifting the earthing point to the emitter of the transistor, but making no other changes, leads to Fig. 5(b), which ios a simple common-emitter amplifying stage with adjustable feedback.

If the output in the Fig. 4 circuit is taken from the pot. slider instead of from the point shown, then the circuit, redrawn with a transistor in place of the op. amp., is as in Fig. 6(a). Merely shifting the earthing point to the pot. slider then yields the circuit of Fig. 6(b). It is now evident that moving the slider to the right has two separate effects — it increases the amount of resistance in the emitter lead, thereby increasing the amount of negative feedback, and it reduces the collector load resistance. Both these effects contribute to reducing the gain, which becomes zero with the slider fully to the right.

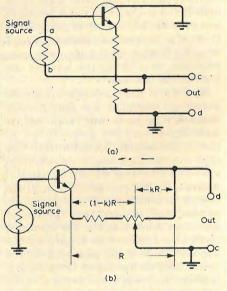


Fig. 6. Fig. 4. Circuit with output taken from pot slider and rearranged at (b) to show dual function - varying emitter resistance and varying feedback.

Employing just a single transistor, as in Fig. 6(b), will give a noise performance which is inferior to that achieveable with more elaborate arrangements. This is largely because the resistance inserted in the emitter lead is itself a generator of Johnson noise, which is effectively added in series with that generated in the internal resistance of the signal source. The transistor d.c. operating current must be chosen in relation to the source impedance, for good noise performance, and it will then be found that to obtain a substantial reduction of gain by inserting emitter resistance, the amount of resistance needed will give considerable degradation of the noise figure.

The above noise difficulty may be

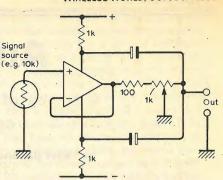


Fig. 7. Fig. 6(b) using an op. amp.

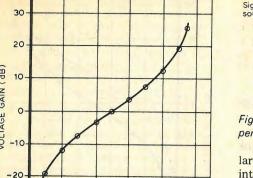
solved by replacing the single transistor by a suitable pair or triple, having a much higher mutual conductance than the single transistor but whose input stage operates at a similarly low current. The increased mutual conductance and output current permit the resistance values associated with the gain-control pot, to be made much lower, with a correspondingly reduced effect on the noise performance at low gain settings. The well-known configurations for pairs and triples as used in audio class 'B' output stages may be adapted to the present application, but an interesting alternative is that shown in Fig. 7. Here the supply connexions to the op. amp. are used as the equivalent of the transistor collector in the Fig. 6(b) circuit — a way of using an op. amp. which perhaps deserves to be more widely borne in mind.

Assuming infinite mutual conductance, the voltage gain of the Fig. 6(b) idealized circuit is simply k/(1-k), Expressing this in decibels gives the graph of Fig. 8(a). The Fig. 8(b) graph is a measured one for the circuit of Fig. 7.

With the idealized circuit of Fig. 6(b), unity gain occurs when the pot. is set for k = 0.5, and the curve is quite symmetrical about this centre point. With the Fig. 7 circuit, however, the curve is not symmetrical about the unity-gain point. This is because the right-hand part of the pot. is shunted by the parallel value of the two $1k\Omega$ resistors going to the supply lines.

Another very simple feedback gaincontrol circuit is shown in Fig. 9. With high forward gain in the op. amp. itself, this circuit gives a gain, between the input and output terminal pairs shown, accurately equal to k/(1-k). (This formula, as for the Fig. 6(b) case, may be prefixed by a minus sign if it is desired to allow for the fact that phase inversion occurs.)

The Fig. 9 circuit, unlike those previously discussed, has the feature that the current in the gain-control resistance chain is supplied by the signal source. This makes it impossible to achieve a good noise figure over a wide range of gain adjustment, no matter how the resistance values are chosen in relation to the signal-source impedance. That this must be so can best be understood as follows. Negative feedback as such never has any effect on the signal-



WIRELESS WORLD, OCTOBER 1980

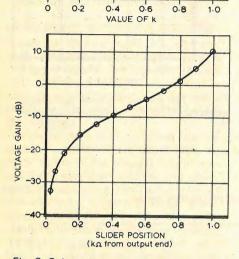


Fig. 8. Gain variation of Fig. 6(b) circuit is at (a). Measured performance of equivalent op. amp. circuit of Fig. 7 is shown at (b).

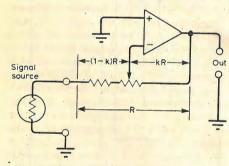


Fig. 9. Feedback gain-control circuit, which has disadvantage of source-fed resistor chain, giving poor noise figure over wide range.

to-noise ratio, at a given frequency, of an amplifier circuit to which it is applied, though the resistors introduced for the purpose of providing the feedback may do so. Thus the output signal-to-noise ratio of the Fig. 9 circuit is the same as that of the circuit shown in Fig. 10. If R is made low, say equal to the internal resistance of the signal source, it will degrade the signal-tonoise ratio at the source terminals*, whereas if R is made much higher, a

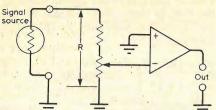


Fig. 10. Circuit of Fig. 9 gives same noise performance as circuit shown here.

large amount of resistance is introduced into the op. amp. input circuit at intermediate slider settings, with correspondingly large Johnson noise and maybe noise from the op. amp. equivalent current-noise generator.

Comparing Fig. 10 with Fig. 1(b) might suggest that the Fig. 9 circuit is no better than that of Fig. 1(b) as regards noise performance. This is not so, however, for to effect a given number of decibels reduction of gain below maximum, the slider in Fig. 9 has to be moved a smaller fraction of the way from the signal-source end of R than is necessary for the same gain reduction in the Fig. 1(b) circuit. The noise performance of Fig. 9 is better than that of Fig. 1(b), but is nevertheless not very good.

Another feature of the Fig. 9 circuit which makes it undesirable for some applications is that the loading of the signal source varies with the pot. setting. If the signal source has a complex internal impedance, the overall frequency response will vary with the gain setting.

This undesirable characteristic of the Fig. 9 circuit may, to a large extent, be overcome by inserting an emitterfollower (or op. amp. follower) between the signal-source and the left-hand end of the resistance chain. With a 50kΩ signal-source, for example, R could be made about 5kΩ, giving reduced Johnson noise from R but nevertheless subjecting the signal-source to negligible loading.

As already mentioned, the Fig. 9 circuit as it stands produces the gaincontrol characteristic shown in Fig. 8(a), which is symmetrical about the unity-gain point. Over a range of about 30dB, and using an ordinary linear pot., the scale shape obtained approximates fairly reasonably to the desirable one having uniformly-spaced decibel divisions, though for many applications a gain of more than unity would be preferred at the centre of this control range. The modification shown in Fig. 11 provides an increased gain at the point of inflexion of the control characteristic, but has the weakness that the gain cannot be reduced right down to zero. Provided R_a and R_b are made much lower in value than the pot. resistance, however, the minimum gain may be made sufficiently low for many pur-

If a stud type pot. is used, and assuming there is complete freedom in the choice of its law and total resistance value, the Fig. 11 modificiation gives no advantage, the required performance being obtainable with better economy of components by adopting the Fig. 9 arrangement.

The circuit of Fig. 12 possesses a combination of several good features. It employs only one op. amp., has a high input impedance, the feedback network can be of low resistance for good noise performance, and the values of Ra and R_b can be chosen, in relation to R, to make the point of inflexion in the control characteristic occur at a gain of much greater than unity, as sometimes

Analysis shows that the gain of the Fig. 12 circuit is given by:

$$\frac{V_{\text{out}}}{V_{\text{in}}} = \frac{R + \frac{R_{\text{b}}}{1 - k}}{R + \frac{R_{\text{a}}}{k}}$$

or
$$\frac{V_{\text{out}}}{V_{\text{in}}} = \frac{k}{1-k} \times \frac{R(1-k) + R_{\text{b}}}{R_{\text{k}} + R_{\text{a}}}$$
 2.

Thus, if R_a and R_b are each much greater than R, the gain is approximately proportional simply to k/(1-k), and is approximately equal to R_b/R_a when k=0.5. Thus the control characteristic is fairly closely as in Fig. 8(a) but shifted upwards. For lower values of R_a and/or R_b the characteristic is of modified form, covering a smaller number of decibels with reasonable

The curve shown in Fig. 13 is the result of a measurement using the Fig. 12 circuit with the following values:

$$R = 1 \text{k}\Omega R_a = 330\Omega R_b = 3.3 \text{k}\Omega$$

Comparison of this curve with Fig. 8(a) shows that it gives a poorer

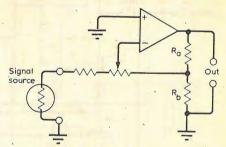


Fig. 11. Variation of Fig. 9, giving increased gain at halfway position of

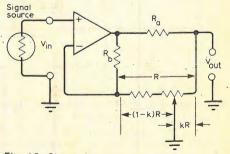


Fig. 12. Circuit featuring only one amplifier, high input impedance, low-resistance feedback chain for low noise and flexibility in choice of inflexion

^{*}When a resistive source of internal resistance R is shunted by a load resistance equal to R, the signal voltage is halved, but the Johnson noise voltage is reduced by a factor of only 2. The signal-to-Johnson-noise ratio is therefore worsened by 3dB.

ideal curve would not, of course, remain linear down to k=0, for this would make it impossible to fade a programme down to zero volume. For most audio purposes, the ideal characteristic would cover about 40dB linearly, curving down to "minus infinity dB" below about k=0.2.)

Another circuit combining feedback and passive gain variation by means of a single linear pot. is shown in Fig. 14.

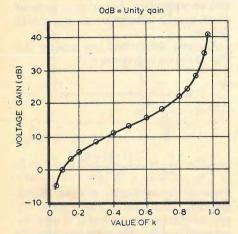


Fig. 13. Curve of circuit in Fig. 12.

This, in essence, is the circuit used by the BBC in their OBA9 outside broadcast amplifier, published in 1952. The gain is given by:—

$$\frac{V_{\text{out}}}{V_{\text{in}}} = \frac{kR + R_{\text{a}}}{R_{\text{a}}} \times \frac{R_{\text{b}}}{(1 - k)R + R_{\text{b}}}$$
or
$$\frac{V_{\text{out}}}{V_{\text{in}}} = \frac{1 + kR/R_{\text{a}}}{1 + (1 - k)R/R_{\text{b}}}$$
4.

The Fig. 14 circuit cannot give zero voltage gain, the gain with k=0 and

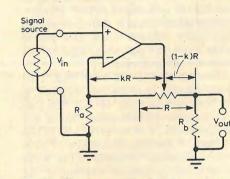


Fig. 14. Circuit providing feedback and passive control in one pot.

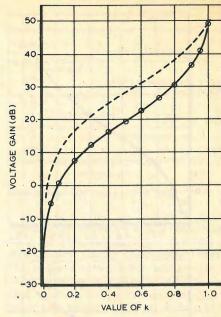
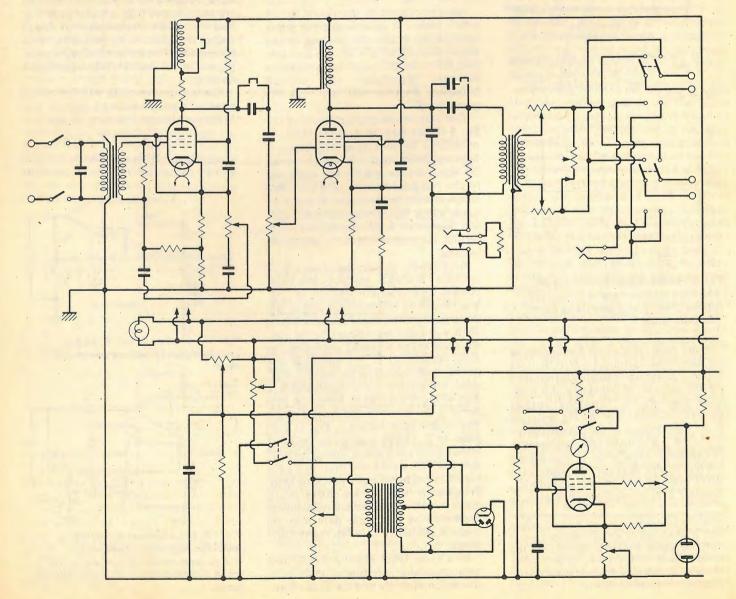


Fig. 15. Full-line curve shows calculated performance of Fig. 14 for two values of R_b.

Fig. 16. BBC OBA8 circuit of 1939, with peak programme meter.



WIRELESS WORLD, OCTOBER 1980

100% negative feedback being $R_b/(R + R_b)$. Though not an ideal feature, the minimum gain in the BBC design is nearly 90dB below the maximum gain, and is stated to be "effectively nil in normal conditions of use"².

The full-line curve in Fig. 15 is a calculated result for the Fig. 14 circuit, using the values $R=100\mathrm{k}\Omega$, $R_a=330\Omega$ and $R_b=3.3\mathrm{k}\Omega$. For the broken-line curve, R_b was changed to $10\mathrm{k}\Omega$. (The values in the BBC design were $R=1\mathrm{M}\Omega$, $R_a=390\Omega$ and $R_b=100\mathrm{k}\Omega$.)

Figure 15 shows that with an ordinary, linear $100k\Omega$ pot. in the Fig. 14 circuit, a control law not departing by more than 2dB from the ideal linear decibel scaling is obtained over an approximately 40dB range. In the BBC design², a stud type of $1M\Omega$ pot. was used, giving 38 steps of 2dB each and two larger steps at the low-gain end. Of course, if the luxury of stud pots. is allowed, any of the circuits here discussed may be given whatever control law is desired.

Though there is much to be said on grounds of economy, especially in stereo systems, for using a single pot. section to vary the feedback and effect passive attenuation, the use of ganged stud type pots, to perform these operations separately gives the designer greater freedom of choice in optimizing the design in all its aspects. This technique was used in the BBC OBA8 outside broadcast amplifier, designed well over forty years ago1. Starting at the maximum-gain setting, anticlockwise rotation of the knob first simply applied increasing negative feedback to the first stage, by raising the effective value of the feedback resistance in the cathode circuit. When this purely local feedback had been increased sufficiently to give a gain reduction of 16dB, further rotation of the knob maintained this first-stage feedback constant but proceeded to insert increasing passive attenuation between the first stage and the second (output) stage. In this way the twovalve amplifier was made capable of delivering full output level to line, at low distortion (abour 1%) for peak microphone input levels extending over a range of 56dB. (It is evident that the designers of this amplifier and the associated units gave high priority to keeping the number of valves used down to the absolute minimum necessary number. This is understandable enough, bearing in mind that the AC/SP3 television pentodes used were physically large and consumed four watts of heater power each. Now that high-gain devices are very small and cheap, and consume relatively tiny amounts of power, the designers of today are justified in adopting a very different outlook, often exploiting the plentifulness of gain to eliminate, or reduce the size of, transformers and also to achieve lower distortion levels in equipment of very much smaller size. Now that it has become fairly easy and cheap to obtain very low distortion

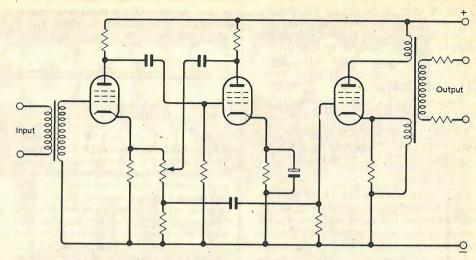


Fig. 17, BBC OBA9 circuit, designed in 1952.

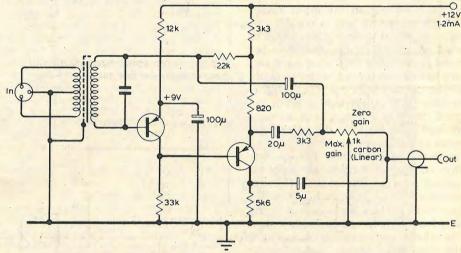


Fig. 18. Author's design of 1961.

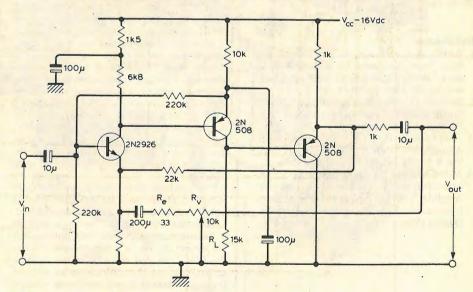


Fig. 19. Circuit by McWhorter of 1966.

levels, there is little argument for doing otherwise, whereas when the OBA8 was designed, lower distortion would have meant more valves, higher power consumption and shorter operating time on standby batteries. The designers were therefore justified in making the distortion just comfortably low enough, but no less, though they were doubtless

quite capable at that time of achieving much lower distortion levels had this been thought desirable. In most circumstances of use, it is doubtful whether the subjective quality of the OBA8 could be distinguished from that of the best modern equipment. The weakest feature of the design is that the secondary of the input transformer,

which stepped up to the exceptionally high impedance of $300k\Omega$, is shunted by a $300k\Omega$ resistor, thus sacrificing, in simple theory, 3dB of potentially-available signal-to-noise ratio. This point does not appear to have been appreciated at the time.)

Figures 16-19 show four practical amplifiers which use a combination of feedback and passive gain control. The McWhorter design⁴ of Fig. 19 employs the basic circuit of Fig. 12, which has also re-surfaced recently in a Philips tape recorder⁸.

My own circuit³ of Fig. 18 is the same in broad principle, but unlike Fig. 12 has the negative feedback and the signal output taken from different electrodes of the output stage. This permits injection of the feedback voltage in series with the transformer secondary, thus obviating the introduction of local emitter feedback in the input stage. Though this circuit was in regular and very successful use for some years, a weak point in its design ultimately became evident, but only after hard service had caused the pot. slider to make erratic contact with the track. Unfortunately, if the slider fails to earth the track, there is a signal path straight through the track from the output collector to the input base. This is positive feedback and is of greater magnitude than the negative feedback from the output emitter. Violent oscillation therefore occurs during moments of poor slider contact, with accompanying very loud noises from the loudspeaker! The other circuits described do not have this weakness - a point worth bearing in mind.

To be continued

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BOOKS

A Users Guide to Copyright, by Michael F. Flint, is intended to make clearer the subject of copyright "to enable people whose jobs or even hobbies - cover any copyright field, to acquire a general understanding". It is, however, only a reference book, and does not cover all the more complex legal aspects which may arise when dealing with this intricate subject. The book is laid out in a manner which will enable its reader to obtain the relevant piece of information quickly, and each chapter is sub-divided into well defined sections, each with a reference number and a bold sub-heading. Part 1, the first 14 chapters, is a general explanation of the copyright law, whilst the second part gives a more specified description of copyright in practice, with chapters directed at publishers and printers, advertising agencies, the music industry etc. The book is published by Butterworth Law Publishers Ltd, 88 Kingsway, London WC2B 6AB, and its price is £8.50 in limpback form.

Microcomputers are responsible for a great number of paperbacks, mainly from the USA, and the pace of publication does not appear to be slackening. Three such books have reached this office recently, among others too numerous to mention, each slanted in a different way.

The first is by a British author, Robin Bradbeer, and is entitled The Personal Computer Book, published by Input Two-Nine at £5.25 and distributed by MCB Publication, 198/200 Keighley Road, Bradford, West Yorks. BD9 4JQ. This one assumes no knowledge of computers - not even enough to know what computers will do - and, accordingly, the first two chapters are extremely basic. The rest of the book is an attractively written explanation of the more important aspects of computing techniques and of computers, a very useful feature being a survey of equipment currently on the market. Several appendices provide information which is quite difficult to find elsewhere in one place, such as bus standards, addresses of clubs, manufacturers and pub-

The second book, by E. A. Parr, is published by Bernard Babani (Publishing) Ltd., The Grampions, Shepherds Bush Road, London, W6 7NF at £1.75. This one is entitled **A Microprocessor Primer**, and approaches the subject by way of a hypothetical device, the DIM-1, so that the author can explain general features of microprocessors without being constrained by any particular design. Having gone through this process, he then sets out to study the Z-80. This is a small book (75 pages) but within its scope achieves its purpose.

Thirdly, there is Introduction to Microcomputers for the Ham Shack, by Harry L. Helms, Jr., published by Howard Sams and distributed by Prentice-Hall International, 66 Wood Lane End, Hemel Hempstead, Herts HPZ 4RG at £3.20. Also a small book, this is concerned with the application of micros to amateur radio. Three chapters are allocated

to the basics of micro operation and programming, after which two chapters describe present and future operations using micros to send and receive Morse, to convert slow-scan tv to fast-scan for ordinary viewing, to store frequencies, in digital modulation, and in several other roles.

Early Radio Wave Detectors, by V. J. Phillips, gives a comprehensive account of various radio wave detectors used before the advent of the crystal and thermionic valve. Among the types described are spark-gap, electrolytic, magnetic, thin-film and capillary detectors, as well as tickers, tone wheels, heterodynes and coherers, the type of detector which makes use of "a phenomenon which occurs in a poor electrical contact, the sort of contact which the engineers of today would call a 'dry joint'".

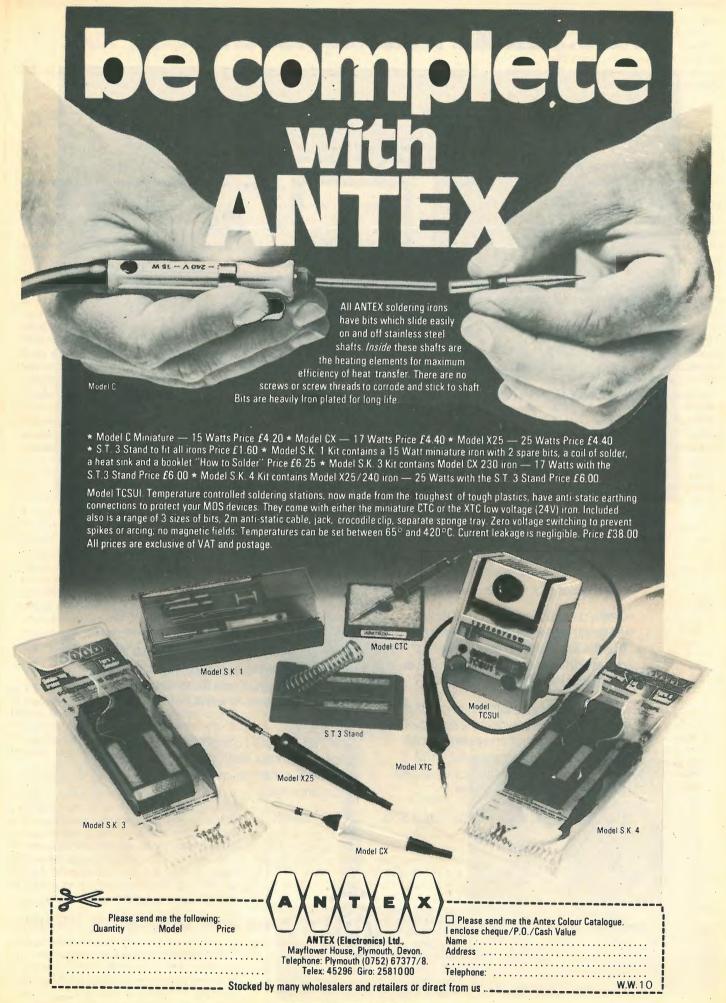
Among the items described under the heading "Miscellaneous detectors," are the 'physiological' receiver, which made use of the electrical sensitivity of a frog's leg to displace a pointer on the smoked surface of a rotating drum, and the use of a human brain as a coherer, the description of which is supplemented by a photograph for which an advisory note is given for the benefit of "readers of delicate sensibilities". Be forwarned, however, the note appears at the bottom of the page, and the photo at the top!

The last chapter, entitled "And so to the modern era," covers the early crystal and thermionic valve type detectors and how they were used — an appropriate finale to an interesting and well-illustrated book. The publishers are Peter Peregrinus Ltd, Marketing Dept, Station House, Hitchin, Hertfordshire SG5 1RJ, and the price of the book in hardback form is £16.

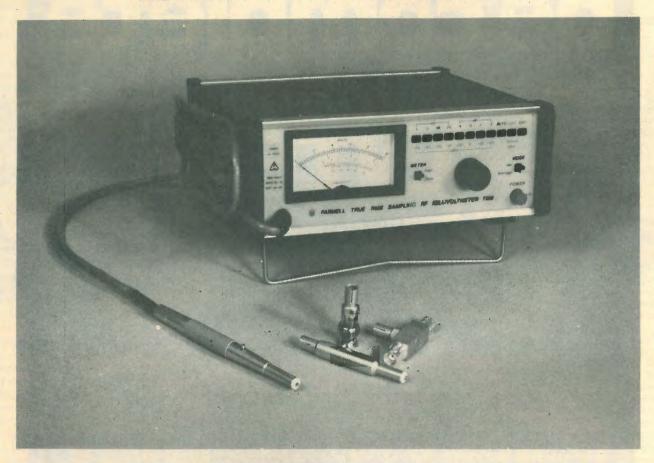
Digital Techniques and Systems, by D. G. Green, is intended as a first course book for students with a basic knowledge of electronics and telecommunication transmission techniques, but the combined coverage of basic techniques used in modern digital circuits, and elementary principles of data communication, laid out in a logical sequence, make it useful for anyone wishing to gain insight into this field.

Chapter 1 gives a concise introductory description of a few of the uses of modern digital applications to which he may put the knowedge that he is about to learn. The second and third chapters cover the operation of electronic gates of all kinds and the remainder of the book, which includes chapters on digital modulation, data-links and pulse code modulation, is devoted to the subject of data transmission over telephone lines.

Worked examples are included in the text, and each chapter concludes with exercises, some of the questions of which have been taken from past C and G examination papers. Multiple-choice questions are also provided at the end of the book, which is priced at £4.95 and published by Pitman Books Ltd, 39 Parker St, London WC2B 5PB.



NEW R.F. MILLIVOLTMETER



The TM8 is a new autoranging analogue true r.m.s. millivoltmeter with a specified operating range of 10kHz to 1GHz and useful indications up to 1.5GHz. It measures r.f. voltage from 1mV to 3V (or 300V using the 100:1 precision divider) and also has a logarithmic range which spans four decades—useful in setting-up tuned circuits.

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An acoustically small loudspeaker

Unusual design gives low colouration and good off-axis response

by R. I. Harcourt B.Sc., M.I.E.E.

This design for an active-crossover loudspeaker system is based on acoustic principles which are well established, and on psycho-acoustic criteria which are subjective in nature. As with all designs, trade-offs are possible. The acoustically small loudspeaker is designed to reduce colourations of the sound, below the limit of audibility where possible. This can be done at the expense of bass distortion, though since 40% second harmonic distortion is inaudible at 80 Hz¹, it is not considered important. In addition, a novel fourth-order, bandpass sub-woofer is described using an acousto-electronic crossover and feedback Q correction.

The basic aims of the design were low colouration and a uniform off-axis frequency response. A flat on-axis frequency response is the accepted criterion, but the off-axis response often compromised in commercial designs - determines the stereo imaging qualities. Colouration and off-axis response depend upon both the drive unit and its enclosure, particularly in the mid-range, where the ear is most sensitive: it is between 1-4 kHz that most of the image is found. To avoid compromising this part of the spectrum, the Jordan 50mm aluminium-coned unit was used for its small size, low colouration, and good transient response - this was the only drive unit found for which the impulse response is published. It must be emphasized that it was not designed as a mid-range unit, and is specified to 22 kHz: the booklet advocates using the unit, together with a bass driver, to form a two-way system3. However, to the author's and a colleague's ears, an improvement was obtained with the use of a dome tweeter above 4 kHz, making a three-way system. Whichever way the unit is used, there are no crossovers in the critical range 500 Hz to 4 kHz to detract from the imaging quality by giving rise to an uneven polar response around the crossover frequency2.

The design of a loudspeaker is often influenced by the ease, or otherwise, of its manufacture. For example, it is rare to find other than a cuboid of wooden construction used for the enclosure. But

even ideal drive units are at a disadvantage in a wooden cuboid. The shape, size, materials and construction of an enclosure all have audible effects on the response. The great advantage of home construction is that one is freed from many of these constraints, and this advantage is exploited in the design. A mid/high frequency enclosure is made of modelling clay which does not require firing. Thus, the shape and materials of the enclosure are optimized.

Cavity resonance

Distortions in musical sounds take several forms, and the total harmonic distortion is often quoted. More recently, it has been found that this measurement does not correspond well

with how a unit sounds: indeed, sometimes a valve amplifier with a high t.h.d. is preferred to a transistor design. It has become clear that steady-state measurements do not give a good indication of performance, and other measurements have been used. With pickup cartridges and loudspeakers, there is a large variation between units sometimes expressed as "detail" or "dynamics", perhaps due to the presence or absence of masking effects of one sound upon another. More complex effects have been found, which are time-dependent, such as the 1 millisecond forward inhibition of a sound upon a following one, and the 30-120 millisecond backward inhibition of a sound upon a preceding one4. These



could be stimulated by delayed resonances in transducers, which often have time-constants within these ranges. With this in mind, and the author having a particular dislike of the sound of delayed resonances, the design for the acoustically small loudspeaker sets out to minimize them.

The cavity resonances of an enclosure constitute an inharmonic series given by the solution to the wave equation for rectangular (or other) boundary conditions. The cuboid has resonances

$$f_{\text{nx. ny. nz}} = \frac{c}{2} \sqrt{(n_{\text{x}}/x)^2 + (n_{\text{y}}/y)^2 + (n_{\text{z}}/z)^2}$$

where c is the velocity of sound, n_x , n_y , n_{x} are integers chosen separately and x_{x} . y, z are the dimensions. These resonances can be heard because they present a widely varying acoustic load to the rear of the diaphragm and thus affect its motion. At high frequencies they can be damped using acoustic filling material, but this is not true at lower frequencies, nor necessarily for small enclosures; for both frequency and thickness of material affect the absorption. An acoustically small loudspeaker is of such a size that the lowest, and therefore all, the cavity resonances are outside the passband, and the loudspeaker is used below these frequencies. This applies, then, to a bass unit. Choosing $n_x = 1$, $n_y = 0$, $n_z = 0$ gives the lowest resonance at

$$f = c/2d$$

where d is the largest dimension of the enclosure, and this is the onedimensional half-wave standing wave.

Panel resonance

Many loudspeakers have a "boxy" sound while reproducing male speech. The box can produce sounds in various ways, one of which is given above. Another way is by the panels of the box vibrating. It has been found⁵ that at certain resonant frequencies, the output of the box is within a decibel or so of that of the loudspeaker. As an experiment, some enclosures were made after Linkwitz², constructed of 6mm plywood with a 10mm internal layer of roofpatching tar. The transmission of the cabinet side-panel was measured by placing two such units together, fed by sine-waves of equal amplitude but opposite phase, so as to null the sound from the loudspeaker. The microphone was placed 1cm from the side-panel so that the near-field response was measured. The results are shown in Fig 1. After correcting for the relative emitting area of the panel and allowing for two panels, the output from the box at 150 Hz was found to be about 8dB below that from the drive unit. Since the Q was measured to be 5, the box will continue to produce the sound after the drive unit has finished, which constitutes a delayed resonance. In this case the 40dB decay time will be Q/0.7f = 48ms.

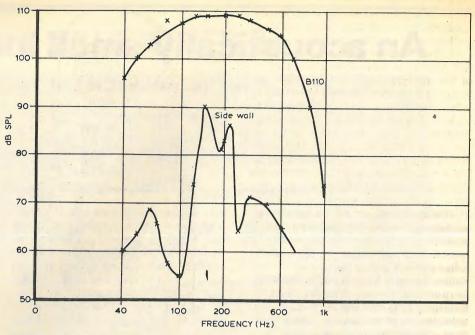


Fig. 1. Near-field transmission of cabinet side compared with that from B110.

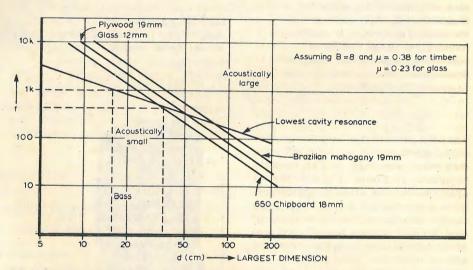


Fig. 2. Cavity and panel resonances for varying maximum dimension.

A panel has a series of resonant frequencies, the lowest one of which is

$$f = \frac{Bt}{2\pi a^2} \sqrt{\frac{E}{\rho(1-\mu^2)}}$$

where t is the panel thickness, a its dimension (for a square panel). E the Young's modulus of the material, p the density and µ the Poisson's ratio. B varies according to the construction, and is higher for a clamped panel than a freely supported one. For loudspeaker enclosures, B is taken as 8. An acoustically small loudspeaker can be made so that the lowest, and therefore all, the panel resonances are above the frequency of operation. The two resonance functions mentioned are plotted in Fig 2. From this can be determined the maximum dimension of an enclosure for it to be acoustically small. It can also be appreciated that most loudspeakers are acoustically large. The graph of the lowest cavity resonance

coincides with a criterion for determining the maximum frequency at which to operate a drive unit to ensure wide dispersion, and the maximum enclosure width.

The sound emitted by an enclosure depends upon its dimension and the degree of its motion. For a circular piston, the emitted sound pressure level increases by 12dB for a doubling of its diameter, which implies that as a box is made smaller, so the sound radiated from it decreases. However, the internal pressure within the box increases as the volume is decreased, so that the deflecting force on the panel increases. This is compensated by a decrease in the actual deflection with reducing dimension, according to a square law. The combined effect of all this is a decrease of about 6dB in the emitted sound with a halving of the dimension. All the above factors represent a confluence of ideas pointing to acoustic size as being an important parameter. It is therefore no coincidence that listening tests have

revealed a preference for small loudspeakers, provided that these are also well designed in other respects.

The panel transmission loss below the first resonance depends on the stiffness of the material used, not its mass or damping properties. The bass enclosure is best constructed of a thick material of high Young' Modulus. In this respect plywood is better than chipboard and hardwood better than plywood. Glass would seem to be an ideal material, for it has a Young's modulus 75 times that of chipboard, and an enclosure can be fabricated in the same way as an aquarium, using silicone rubber as an adhesive. This is a subject for further work.

Clay enclosure

Diffraction round an enclosure has been found^{6, 2} to have a bearing on the frequency response and stereo imaging qualities of an enclosure, and Fig. 3 shows the frequency response of differently shaped enclosures, other things being equal. The sphere was found to give the smoothest response of the shapes tested, since there are no discontinuities in the surface to give rise to frequency-dependent effects. A novel enclosure is made as close to a sphere as practicable, and consists of a short vertical cylinder with domed top, as shown in Fig. 4. The shape is achieved by using modelling clay, which has a high density and large internal losses - it is acoustically "dead". The clay used is sold under the trade name of "Das", and does not require firing. It is not possible to include the bass unit in this enclosure, so only the mid/high-frequency unit or units are placed in it, and it is stacked on top of the bass enclosure. Because of the rounded shape, advantages are obtained in suppressing cavity resonances. The top-to-bottom, onedimensional standing wave which normally occurs in a pipe is suppressed by

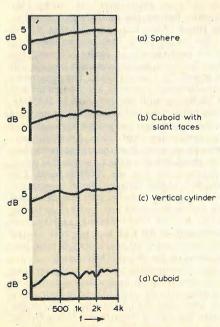


Fig. 3. Frequency response of four different cabinet shapes.

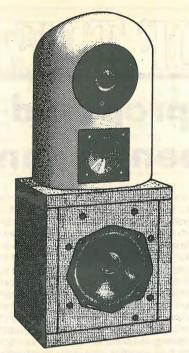


Fig. 4. Author's prototype. Clay enclosure on top contains mid and high-frequency units, while wooden bass enclosure is for B110. Single sub-woofer is not shown.

the domed top, and similarly the axial one-dimensional standing waves are suppressed by the cylindrical walls. This leaves the two-dimensional waves, and the lowest is calculated to occur at 1.4 kHz, where the damping material used has a high absorption.

Bass enclosure

The simplest way of making an acoustically small bass enclosure is to make it physically small, and for operation up to 500 Hz the lowest resonance is placed higher, at 1 kHz, where the response is 20dB down. This, combined with the high absorption of the filling material and the internal losses of the enclosure material, will give only very small amounts of unwanted sound. The maximum dimension for the bass enclosure is found from Fig. 2 and is 16cm. The volume of a 16cm cube is 4 litres, and it is clear that this will give insufficient bass extension. A cube is to be avoided, since resonances coincide to give a higher Q, and the bass enclosure

is best made with dimensions in the ratio 2.3:1.6:1, this being the ratio used for designing listening rooms. The maximum volume is around 1.1 litres, which is too small, and so a modification is called for. A 5in bass unit requires an enclosure dimension of 15cm on the front panel, and so the box is made square at 16cm: the other dimension is determined by the volume obtained from the design procedure for bass loading. The volume is divided internally by a partition placed to brace the magnet against the back panel, which also suppresses the offending double resonance caused by the square dimen-

An acoustically small loudspeaker does not have to be physically small, and this is achieved by a scheme of internal partitions, in which each subvolume is acoustically small, but is connected to the adjacent one by a low resistance path. The partitions simultaneously brace the panels, effectively sub-dividing them into smaller ones which are acoustically small. The smallest dimension of the box is the width to ensure wide dispersion, and this is equal to or slightly greater than that required to house the drive unit.

Sub-woofer

The bass extension in this design is obtained by a novel sub-woofer, the aim being to achieve economy in space and expense. A 12in bass guitar speaker is capable of producing high levels of bass below 100 Hz, and is inexpensive. However, it has a rather high resonant frequency, which was utilized by placing it in an acoustically small enclosure and using it below resonance, with a second-order filter, to give the required amount of bass boost. The closed-box enclosure acts like a second-order high pass filter, and the flat part of the response above resonance is made to fall off at 12dB/ octave using the filter. The portion of the response below resonance which was falling off at an ultimate slope of 12dB/octave is made flat with the same filter. The falling part of the new response is tailored to form half of a 12dB/ octave crossover, the other half being the natural fall-off of the bass enclosure

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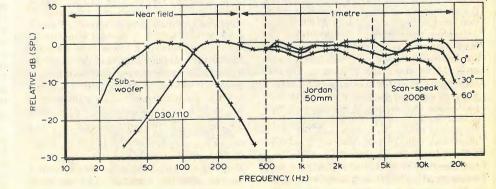


Fig. 5. Frequency response of system plotted using one-third octave pink noise signals.

NEWS OF THE MONTH

928 MHz proposed for UK's Open Channel

Concern about interference seems to be the main reason why the Home Office is proposing "just above 928 MHz" as a frequency for Britain's citizens' band service. The thinking behind this choice is published in a Green Paper discussion document entitled "Open Channel" - which is what the Government intends to call the UK service to disssociate it from the bad reputation of c.b. operation on 27MHz in some countries. By placing the service in this part of the u.h.f. band they would put it above all the television channels used in Britain and also above most of the communication and other systems - notably those of the police, the fire brigades and aircraft landing systems where interference could have serious consequences. Because of likely interference they have rejected the National Electronics Council's proposal for a band somewhere between 100MHz and 500MHz (August 1978 issue, p.38). Having looked particularly closely in this region at suggested bands in the neighbourhood of 225MHz and 450MHz. the Home Office remarks in the Green Paper that here the interference "could be so severe and intractable as to lead to the dropping of certain television channels. This could not be contemplated in the context of European television planning, quite apart from the effect on the broadcasting authorities".

The choice of the particular figure of 928MHz, above which the Open Channel would be placed, is determined by several factors. First, several other countries, including the USA Canada and some in Western Europe, are also considering setting up personal radio services in this part of the spectrum, notably in the new 900MHz mobile radio band (862-960MHz in Region 1) which was allocated by international agreement at the WARC in Geneva last year (February 1980 issue, p. 48). International standardisation would be a good thing, particularly to give manufacturers opportunities for economies of scale and larger markets in the design, production and sale of equipment. Secondly, in this respect, "just above 928 MHz" would avoid the ISM (Industrial, Scientific, Medical) band of 902MHz to 928MHz which is designated for use in ITU Region 2 (the Americas). Thirdly, there are the constraints of other, fixed communication services in this u.h.f. area which help to determine the figure of 928MHz. But the most important factor in fixing it is the possibility of image interference in television charinel 68 resulting from frequencies immediately below 928MHz. The Green Paper in fact envisages a band 1MHz wide containing a maximum of 40 channels, each of 25kHz. This implies the possibility of frequency modulation, although the document does not say this directly. The Home Office's Radio Regulatory Department sees the question of choice of modulation as a commercial rather than a technical matter.

The proposal for a frequency of 928MHz

raises the question of the likely range of Open Channel transceivers because of the relatively high absorption of r.f. energy by obstructions in this part of the spectrum. Of course, the Government doesn't want longdistance transmission in any case, and they consider a range of about 15km is enough. The Green Paper says that the frequency selected must enable this desired range of 15km "to be achieved in most environments without excessive transmitter power, thus minimising local interference, enabling frequency channels to be re-used and at the same time avoiding long range interference". After quoting some earlier studies of propagation at 900MHz and the ranges achieved, the document goes on to mention a limited series of tests done by the Home Office themselves in and near London to obtain additional data in this part of the spectrum: "From these tests it was estimated that with 25W e.r.p., a sensitive receiver and with aerial heights of 4 and 1.5m, the range in urban and suburban environments would be from 3 to 10km and that in open, flat country with no trees it might approach 20km."

To revert to the question of interference with other services, the Green Paper analyses what could happen with an Open Channel band near three different frequencies: 225MHz, 450MHz and 900MHz. Around 225MHz, the third harmonic of the transmitter "would come within the range 669-675MHz and reception of television channels 45 (662-670MHz) and 46 (670-678MHz) would be potentially affected." Up to 1.1 million television sets could be affected here by transmissions from equipment mounted in vehicles or from portable transceivers with integrated aerials. Below 450MHz, television reception of channels 64, 65, 66 and 67 would get interference from the second harmonic of the transmissions. Up to 1.7 million tv sets could be affected. Above 450MHz, the problem "is one of spurious responses in television receivers tuned to channels 23 and 24. It is estimated that 2.6 million installations receive a television field strength which is less than that necessary to protect them from the transmissions of fixed Open Channel equipment." At 900MHz, and up to 928MHz, Open Channel frequencies in this region "would potentially affect the reception of television channels 59-68 inclusive ... " and it is estimated that "... up to 1.8 million tv installations receive a lower field strength from main stations than would be necessary to protect them from nearby Open Channel transmitters." Here, and in the band above 450MHz, the interference could be dealt with by filtering "but the scale of the possible problem is daunting."

On the question of regulatory control of Open Channel, the discussion paper says that the Government proposes "to combine the simplest possible licensing system with a limited technical control." The licensing system "would be flexible, simply authorising a named user, or a person acting with his permission, or a person to whom he had hired equipment, to use Open Channel." Licences would be renewed annually "and unlicensed transmissions would constitute, as now, an offence under the Wireless Telegraphy Act," Revoking licences would be a way of applying sanctions, short of prosecution, against deliberate illegal use. The licence fee would be set to pay for the administration of the service.

For technical control, only minimum standards for equipment would be set, and the Government's responsibility to users would be "confined to ensuring that a certain standard of service can be obtained rather than ensuring that it is obtained." There would be no formal specification system as with p.m.r. equipment. Regulations under the Wireless Telegraphy Act could set out technical requirements - on modulation, power, frequency stability, spurious radiation etc. which equipment manufacturers would have to meet. These regulations "could make manufacturers liable to certify their products as conforming to those requirements; the onus would then be on the user to ensure that he used only certified equipment." In general the Green Paper makes it clear that the Government sees its responsibility as creating the technical conditions for a reasonable service but not in coping with

 Reactions to the Government's proposal have been mixed. The UK radio amateurs, for example, are quite pleased (for reasons explained in World of Amateur Radio this issue). So are those who concur with the Government's view of citizens' band as basically an amusement or hobby (Mr Timothy Raison, the minister concerned at the Home Office, has remarked that "it will be fun for people"). They see no reason to strive particularly to make life easy for, manufacturers, dealers and users in what will be essentially a luxury trade. The equipment manufacturers, however, are predictably not at all happy with the proposal of 928MHz. They think there will be insufficient demand to make the design and production of sets for this frequency profitable. American experience suggests that the transceivers may cost about 20% more than comparable p.m.r. equipment for, say, the 200MHz band. The president of the Citizens' Band Association pressure group, James Bryant, has described the Green Paper as just another delaying tactic by the Government, and Walter Stevenson, of Air Call Ltd, has commented that many potential operators will now just go ahead on the illegal 27MHz band.

The Government is, of course, inviting such comments on its discussion paper and has asked all concerned to send in their views not later than 30th November, 1980 to the Radio Regulatory Department, Home Office, Waterloo Bridge House, Waterloo Road, London SE1 8UA.

Government begins erosion of Post Office monopoly

Referring to "a transitional period of three years," Sir Keith Joseph announced on July 21 that British Telecom's monopoly would be limited by government changes in the way terminal equipment is used as well as in the supply of services to third parties.

The new provisions, for which legislation will be introduced in the next parliamentary session, are expected to make it easier for privately supplied equipment to be connected to the Post Office network, assuming that the equipment meets the required technical standards. Similarly, more freedom will be extended to people who wish to use British Telecom's circuits to offer services to third parties which are not currently provided by the company, data processing facilities, for example.

In announcing these changes, Sir Keith also mentioned the possibility of "allowing the private sector to provide telecommunications transmission services such as satellite business systems," He said that he expected the main changes to lead to a significant growth in information, data transmission, educational and entertainment services provided over telephone circuits and to the emergence of new business. He said that he would be commissioning an independent economic assessment of the implications of allowing complete liberalisation for what are commonly referred to as "value added" network services. These include database services providing archives, advertising and entertainment services, electronic office facilities such as word processors, verbal message services, etc. as well as facilities for the interconnection of normally incompatible apparatus such as computers, facsimile machines and word processors. Monitoring and security alarm services also fall under this heading.

The first telephone and associated wiring

connected to the main network will remain the responsibility of British Telecom, as will the maintenance of private branch exchanges (PABX) and associated wiring.

In theory, this should make available to the user a wider variety of equipment and sources and Sir Keith said that he is looking forward to seeing at an early stage approved extension telephones on sale in the shops, as well as greater competition in the installation and wiring of currently approved apparatus on business premises.

A spokesman of the Post Office Engineering Union, responding to the announcement, said that the changes would allow private operators to "cream off" the more profitable side of the business, leaving British Telecom to deal with the less profitable but necessary sector.

At the same time, Sir William Barlow, the Post Office chairman said that consumers' bills were likely to rise as a result of the changes.

IEC nuclear reactor standard published

The 70th standard produced by the International Electrotechnical Commission was issued late in July, and deals with periodic tests and monitoring of the protection systems of nuclear reactors. This standard, IEC publication 671, lays down principles for testing protection systems during both normal power operation and shutdown. Among details such as short interval or continuous surveillance checks the standard also considers the effect of test equipment failure on the reliability of reactor protection. The full publication can be obtained from the International Electrotechnical Commission, Central Office, Geneva, Switzerland, price 39 Swiss francs.

Pergamon Press makes first data deal with Russia

An investment of £10 million a year for the next ten years as part of a deal to provide western customers access (in English) to literature in Russian scientific and technical data stores, has been announced by Robert Maxwell, chairman of Pergamon Press.

The agreement made with Viniti, the Soviet Institute of Scientific and Technical Information, and Vaap (Soviet copywright) also includes the joint development of computerised information services. The guardian reports that these services would be immediately available on computer terminals, through the Infoline service, which was acquired recently by Pergamon and the agreement also includes the supply by the Russians of all documents in microfiche

The English language service is expected to begin in the first quarter of 1981 and will include material on information retrieval systems and the environment, with mathematics, energy and engineering following in 1982

Phone charges up

Price increase proposals have recently been put by British Telecom, the telecommunications part of the Post Office, to the government and the Post Office Users National Council. The increases, which are expected to take effect from November 1, include a 0.5p increase in the telephone call unit fee to 4p accompanied by a reduction of time in the inland cheap rate period, although the IDD (International Direct Dialling) cheap rate will be extended to 8a.m. Foreign affairs will clearly be easier to arrange! Telephone rental charge increases to £16.75 per quarter for a business line and £12 per quarter for a residential line. Installation and extension charges will also rise. Further details can be obtained from British Telecom, Public Relations Department, 23 Howland St, London

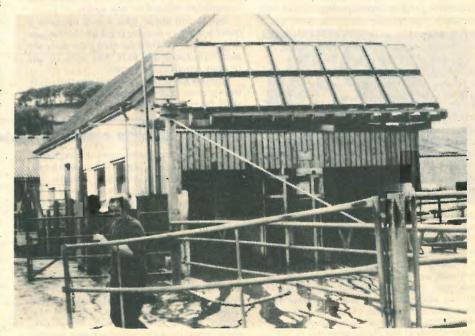
Data logger keeps an eye on the dairy

Two projects to determine economic use of energy are being run simultaneously at the Seale-Hayne College dairy unit in Newton Abbot, Devon, using a multi-channel data logging system, the Microdata M1600L.

One is aimed at energy conservation in the farm's milking parlour and associated dairy and involves a comprehensive study of energy input and consumption, while the other looks at the development of a solar energy system for use in the farm, this being linked to a study of dairy water requirements. More than 50 parameters are being monitored at regular intervals over several years involving the interfacing of a variety of transducers, both analogue and digital, with the data logger. This is achieved by the use of a separate plug-in signal conditioning module for each channel.

Where checking of fluid flow is concerned, turbine-type sensors deliver a pulse output with frequency proportional to flow rate and although this is an analogue signal, the data logger handles it digitally, with signal conditioning modules operating as tachometers to provide the pulse rate in digital form.

Recorded data is subsequently fed into the Plymouth Polytechnic Computer for analysis.



Government set to introduce cheap-rate engineering authority

The formation of a new body to govern the engineering profession, probably a 20-man committee with chartered status, is expected to be announced by Industry Secretary Sir Keith Joseph as we go to press. The new authority is likely to be only a small affair compared with the powerful engineering authority (which was to be directly responsible to government) envisaged by the Finniston Report of April 1980. This amounts to a compromise which could save the government about £8 million, reports the Sunday Times, and would also dispose of the present system of self-regulation by the institutions, through the CEI.

It is thought that the government will appoint members after recommendations made by engineering employers, unions and the institutions, with an initial expenditure level of £1 million to £2 million to get the new independent body started.

'tapping into" the current Engineers' Registration Board and accreditation will probably depend at first upon the goodwill of existing institutions, once members reach the standards agreed by the new authority.

An alternative form could be a chartered and the Engineering Employers Federation.

First small-dish digital video transmissions by satellite

Successful transmission of digital video colour tv signals through a European space satellite using small-dish terminals at both ends of the link, has been achieved by the IBA at Crawley Court, Winchester, reports Pat Hawker

Digital signals were passed through the 120 MHz transponder on the OTS satellite launched in May 1978 and the experiment was carried out with the co-operation of British Telecom and the European EUTEL-SAT organisation. The test signals, using the IBA-developed experimental 60 Mbit/s encoder/decoder, were both sent and received at Crawley Court using the 14 GHz 2.5 metre dish "up-link" terminal (at about 1.5kW transmitter power) and the 3 metre dish receiving aerial.

During preliminary tests using pseudorandom digital signals, error rates of the order of only one in 10 million bits were recorded. During transmission of 625-line



Despite one or two humorous suggestions as a caption to this picture, including

"off-resonance draught detector in action" and "an obsessive approach to stereo speaker positioning", the gentleman is in fact a fully-equipped boardroom "bug" detector. The equipment is the Scanlock Mark VB and the makers, Audiotel International, claim that its sensitivity and frequency coverage (100 to 1800MHz) make it possible to detect a bug automatically in less than a second:

Most of the savings will be effected by

body created by the Fellowship of Engineering (set up by the CEI in 1976), which would imply that membership would be decided entirely by the profession although the most likely authority is the former, partly because so many of the DOI's respondents to offer their views on the Finniston Report suggested that the impetus of an independent body was vital. This type of authority is also supported by the IEE (70,000 members)

colour tv pictures through the system no degradations, other than those introduced by the encoder/decoder system, were observed.

The techniques used here are for experimental purposes only and are not being proposed as an international standard, but the work has shown that digital video could provide useful advantages for news gathering and national and international links through satellites.

Landsat working again

NASA's five-year-old Landsat spacecraft is now back in service after a six month retirement caused by a malfunction preventing correct orientation in its orbit. The spacecraft developed problems on Nov. 5 1979 when the yaw attitude flywheel, part of the mechanism which kept it pointed towards earth ceased functioning, probably due to a lubrication breakdown.

University technicians get 12%, lecturers 17%

The standing commission's report on pay comparability, published at the end of July, recommends a salary increase of about 12% for university technicians, while university lecturers are expected to be awarded 17%, although the Association of university Teachers had originally made a provisional agreement for 19.6% with the university authorities.

The commission recommended that an additional lump sum should be paid to technicians, varying from £46 to £140 according to grade and the increases are back-dated to April 1980. The basic minimum salary for a trainee technician is now £2,367 (at age 16) with the grade 1A technicians minimum at £3,288. Grade 8 represents the maximum at

Under the 17% settlement, minimum lecturers' salary will be £5,505 with the maximum at £11,572, rates effective from October

Professors' average pay will go up from £14,148 to £16,765. An additional cost of living figure is to be added to lecturers' and professors' salaries after talks with the government in September.

Seeing and hearing things at Decca

London Print and Design, a relatively unknown company based in Northington Street, has bought the old Decca record manufacturing plant, lock, stock, and barrel. The plant, which is located at New Malden, Surrey, was taken over by Racal earlier this year (see Wireless World, April 1980) and immediately offered for sale again.

Speculation about who would eventually make a move to acquire the high technology disc business has been running high since the non-manufacturing part was sold to the German company Polygram.

Rumour has it that LPD will use the plant, through the co-operation of key technical staff from Decca, to press videodiscs in partnership with companies interested in the home video entertainment field. LPD has been advertising the sale as well as asking companies interested in the pressing equipment to approach them. The plant has an annual production capacity of 14 million discs, although there is a chance that the central matrix unit could be retained at New Malden, with up to four "satellite" pressing plants operating at other sites, each producing about 31/2 million discs annually.

For some reason best known to itself, LPD chooses not to reveal the nature of its current business, but informed guesses point to links with printing, designing and maybe some more pressing business in the near future.

Japanese satellite completes global telex link

An arrangement made recently between British Telecom and the Japanese telephone authority makes it possible for telephone and telex users in the UK to reach ships in the Indian Ocean, using the satellite earth station at Yamaguchi, south-west of Hiroshima.

This amounts to an extension of the Marisat system which already provides satellite links for ships in the Atlantic and Pacific and is the final link required to provide global coverage. The main advantage of the system, set against normal radiotelephone messages, is that calls are free of fading and distortion. About 320 ships throughout the world are now equipped to use the Marisat system.

To make a satellite telex call, British users should follow the dial/key procedure for making international calls to places outside Europe and North Africa. The keying codes to use are 581 for ships in the Atlantic and 582 for those in the Pacific. The caller then keys the ship's seven-digit call number, followed by a plus sign.

To make a satellite telephone call, British customers should dial 100 and ask the local exchange operator for Freefone 2187, the International service at Faraday Exchange, London. Callers should give the name of the ship it is wished to contact, its location and the vessel's satellite call number.

Designing with microprocessors

5 — Test-and-skip systems

by D. Zissos and Laurelle Valan

Department of Computer Science, University of Calgary, Canada

This and the following article describe step-by-step procedures for the design and implementation of microprocessor-based systems using the test-and-skip mode. In the second article the design steps will be illustrated by means of a fully worked out example.

In the previous article we explained the nature of the synchronization problem, which results from the fact that the microprocessor operation cannot be slowed down to the speed of slow peripherals by reducing the frequency of its clock. Two solutions, which do not involve adjusting the clock frequency, were outlined. One uses software and the other hardware. In the first case, the microprocessor executes a programming loop, during which the status of the peripheral is read and tested. If the peripheral is found to be busy, the process is repeated, that is the microprocessor skips execution of the next instruction. The test-and-skip process is repeated until the peripheral becomes ready, at which point the microprocessor exits the software wait loop. In the

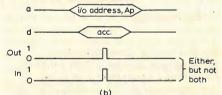
Fig. 1. (a) Microprocessor/peripheral links during execution of an i/o instruction. (b) Microprocessor signals during execution of an i/o instruction.

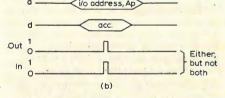
second solution hardware is used to put the microprocessor chip into an idling (wait) state while the peripheral is responding. When in the wait state all microprocessor activities are suspended without turning off the clock. Microprocessor-based systems using this method are referred to as wait/go and will be discussed in detail in a later

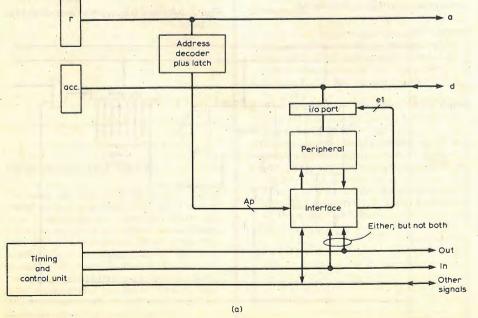
1/O instructions

Before we describe the philosophy and steps we use to design and implement test-and-skip, and indeed all types of microprocessor-based systems, it will be useful, particularly in the case of the inexperienced reader, to recall the stepby-step execution of i/o instructions, which was described in detail in an earlier article.

Briefly what happens is this. The op code is fetched from memory and copied into the instruction register (i.r.) during cycle M1. Next, the i/o address is fetched from memory and copied into







addressing register r. The i/o instruction is executed by connecting within the m.p.u. chip the address bus to the addressing register and the data bus to the accumulator, as shown in Fig. 1. In addition, the timing and control unit generates on specified pins of the m.p.u. chip either a read or write pulse, denoted by In and Out, depending on whether data is to be copied from the peripheral into the accumulator, or vice versa - see Fig. 1(b). The presence of an allotted address signal and an i/o pulse at the input of an interface causes it to activate the peripheral. In other words the input to an interface in a microprocessor-based system consists of software-generated electrical pulses.

In practice the relative timing of i/o pulses, addresses and data vary from microprocessor to microprocessor. However, in our design procedures it is not necessary to consider such signals until the implementation stage:

Design philosophy

The design philosophy adopted is one that allows the inexperienced user to produce sound and reliable systems simply, while at the same time providing the specialist with the tools to improve his technique in dealing with more sophisticated assemblies. As in the case of logic circuits, elegance of design is not sought but can be achieved.

In developing our design philosophy, we considered the following as import-

System reliability. All systems must function correctly.

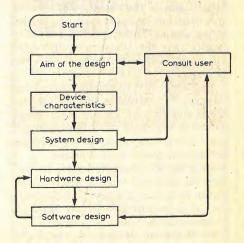


Fig. 2. Chart showing the successive steps in the design process.

Fig. 3. Configuration of a basic test-and-skip system.



Fig. 5. Left-shift through carry.

Circuit maintainability. The systems should be easy to maintain.

Design effort. This must be minimal to allow for greater creavity.

Documentation. This should be concise and to the point. Symbols and diagrams are preferable to verbal statements; they are more readily understood by non-English speaking persons and are likely to prove more attractive to the export market.

Design steps. These must be easy to apply. In our case no specialist knowledge is necessary.

Modifications. The systems should be easily modifiable to meet new conditions as they arise.

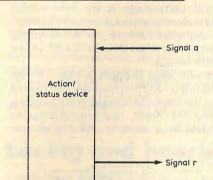
Design steps

Our design process is accomplished in five steps, listed below. See also Fig. 2. **Step 1: aim of the design.** The system specification is expressed in the designer's terms. This step is introduced to ensure that the system requirements are interpreted correctly by the system designer.

This stage is critical for successful co-operation between the system designer and the user. Failure at this stage is usually the cause of system misoperation which then produces the need for subsequent design modifications.

Step 2: device characteristics. In this step the designer studies the terminal characteristics of the devices to be used. Any consideration of purely internal characteristics should be avoided.

Step 3: system design. In step 3 the designer specifies the system characteristics in general terms by means of a block diagram and a system flow chart.



Interface

Fig. 6. Terminal characteristics of action / status devices. Signal a means that a 0 to 1 signal transition on the action terminal activates the device. No activation is possible when signal r = 0. Signal r indicates the availability (r = 1) or unavailability (r = 0) of the device.

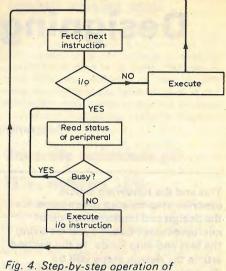


Fig. 4. Step-by-step operation of test-and-skip systems.

Step 4: hardware design. The fourth step involves the design and implementation of the system hardware. This step is provisional, and its results may well be modified in the light of the experience of the next step. It is accomplished conventionally, using well-established methods.

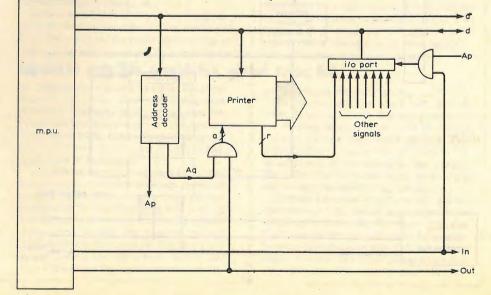
Step 5: software design. On the basis of the hardware design in step 4 and assuming the necessary machine code instructions, the basic software for the operation of the device is designed. This process may well indicate improvements to the hardware which was designed in step 4. In fact, steps 4 and 5 should be repeated until a satisfactory design is obtained.

It is advisable to use a programming model when executing this step.

Basic configuration

The basic configuration of a test-andskip system is shown in Fig. 3. The function of the interface is to provide the microprocessor with status information about the peripheral (indicating

Fig. 7. Interface hardware of test-and-skip systems.



whether it is ready or not), and to activate the peripheral at the correct time, that is when i/o instruction with address Aq in our case is being executed.

Peripheral status information is made

WIRELESS WORLD, OCTOBER 1980

available to the program through an input port. If the ready/unready state of the peripheral is indicated by the '0' and 'l' values of signal r in Fig. 3, to determine whether the peripheral is ready or not, the programmer proceeds in the following manner. He executes an IN instruction with address Ap. Execution of this instruction copies the signals rxxxxxxx in Fig. 3 into the accumulator. If r=0 the process is repeated, otherwise the next (i/o) instruction is executed, which allows the microprocessor to communicate with the peripheral, as shown in Fig. 4. The programmer has several options to determine the value of r. We shall describe two such options. He can AND the contents of the accumulator (rxxxxxxx) with 10000000 (80 in hex), which modifies them to r0000000. If r = 0the zero flag is set, otherwise it is reset. Alternatively, he can shift the accumulator left through the carry flip-flop, as shown in Fig. 5, which shifts the value of r into the carry flip-flop.

If we assume that our peripheral is an action/status device, that is a device whose terminal characteristics are shown in Fig. 6, the hardware implementation of a test-and-skip system is shown in Fig 7. Action/status devices are described in Appendix 1 of "System Design with Microprocessors", Academic Press, 1978.

In the next article we shall demonstrate design steps by means of a PRINT problem. This problem, which will also be implemented using the wait/go, interrupt, d.m.a. and d.d.t. modes, has been chosen, first because a printing operation can be readily visualized and secondly, the character printer used can be assumed to have been in existence in the 1940s, that is well before the era of computers and microprocessors.

WW index for 1979

The index for Volume 85 (1979) of Wireless World is now available, from the General Sales Department, IPC Electrical-Electronic Press Ltd, Room CP34, Dorset House, Stamford Street, London SE1 9LU, price 75p including postage. Cheques should be made payable to IPC Business Press Ltd.

We apologize for the unusually long delay in the production of this index. This was due to a combination of editorial staff problems and/more general industrial disputes.

An acoustically small loudspeaker

continued from page 67

containing the 5in unit. In this fashion, a two-way, second-order crossover is obtained for the price of a single-stage filter, and considerable bass extension with an inexpensive unit. The success of this design is best judged by the observation that, when placed in a corner, 102dB SPL peaks were measured while playing a recording of cannon shots during the "1812" overture, without any sign of stress. Naturally there is a price to be paid, and that is increased harmonic distortion. This could not be heard during music, but sine-waves or pink noise showed it up, and the result was that the source of sound could be located, which is not generally true for such low frequencies. Frequencies below 100 Hz were found to occur infrequently during most music, but bass guitar, bass drum, and organ enthusiasts may prefer some other solution below 100 Hz. That due to Linkwitz2 is an alternative.

The on-axis response of the units was measured above 300 Hz in situ using third-octave pink noise. That below 300 Hz was measured by taking the nearfield response of each unit to eliminate the effects of the room. The results are shown in Fig. 5. The off-axis response was measured above 300Hz in situ, rotating the loudspeaker. Curves are shown for 30 and 60 degrees horizontally off-axis, and show that an integration has been achieved between drive units, and that there are no large steps in the off-axis response to cause shifting or diffuse stereo imaging.

Design and construction will be described next month.

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IN OUR NEXT ISSUE

Unique pickup arm

By displacing the horizontal and vertical pivots of an arm from each other it becomes possible to increase the radius of the arc in which the pickup travels across the record and so reduce tracking distortion. This article describes a practical design for home construction.

Amplifierloudspeaker interface distortion

Matti Otala examines distortion caused by intermodulation between the signal and a delayed, frequency transformed version generated by the loudspeaker and propagated in the feedback loop.

Measurements on four power amplifier circuits are discussed.

Designing inductors carrying d.c.

It's difficult to select initially a core for a winding that is carrying d.c. A simple procedure allows different cores to be compared and the optimum one chosen for a particular inductor design.

On sale October 15

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Voice synthesiser

The "Wooden Fender" group of amateurs in and around Colchester, Essex including a number at the University of Essex — have built and installed what is thought to be the first computergenerated voice synthesiser on a local u.h.f. repeater, GB3CE, located in the Colne Estuary and using channel RB14 (output on 433.350MHz). According to Ian Dilworth, G3WRT, the computer has initially been programmed to synthesise the call-sign and "QRA locator" (ALO5E); in addition the repeater announces frequency and channel number. The basic system, however, has been designed to provide a voice output of the strength and frequency check of the incoming signals, although this has not yet been implemented.

The value of v.h.f. repeaters to provide relatively long ranges in conjunction with simple hand-held transceivers is being proved by experiments that have been carried out by the Canadian Department of Communications during recent years in a remote arctic area 500km north of Fort Chimo, Quebec. There an experimental system for "trail and remote camp radio" has been under test to enable an Inuit hunting community to keep in touch with their village by means of a speciallydeveloped battery-operated h.f./v.h.f. repeater installed on Diana Island, 280 metres above sea level and from the community village at Koartac. The Department acknowledges that the system uses technology drawn from North American amateur use of v.h.f./f.m. "autopatch" repeaters (shared hilltop facilities with automatic, mobile toneaccess to the public switched telephone network). The economic existence of many arctic communities depends on hunting, fishing and berry-picking requiring villagers to spend long periods away from their homes, on the trail or in remote camps.

Repeater abuse

In the UK and USA, unfortunately, the use of amateur v.h.f. repeaters continues to be the subject of controversy and abuse. Paul Essery, G3KFE, in a strongly-worded editorial in Shortwave Magazine writes provocatively "The outcome of the inept plan to quadruple the number of London repeaters can now be seen this appeasement of the deliberate interferers (of all kinds) has merely played into their hands and produced four times the abuse and misuse of these relays. The time is now well overdue for firm action to be taken, for the good of amateur radio . . . If the Home Office is unable or unwilling (as seems to be the case) to make a concentrated effort to find, close-down and

prosecute the offenders, then the RSGB - which holds the licences for these repeaters - has no choice but simply to close down the repeaters."

Not everyone will agree with this analysis but it is a fact that, in the USA, the owners of repeater licences are increasingly complaining that they are being held responsible by the FCC for the content of the communications, including the profanities and jamming, going through their repeaters. Under American regulations, both the repeater owner and the station originating a message are responsible for the content of any communications transmitted through the repeater. The real miscreant is clearly more difficult than the owner to identify and trace. The FCC, it has made clear, has no intention of relaxing regulations in this area.

Open Channel

While the general question of the recent Home Office discussion document on "Open Channel" is not a matter for WoAR, the reaction of radio amateurs, as such, seems generally favourable. It is of course recognised that it will not be easy for industry to provide low-cost, rugged base, mobile and handheld transceivers at the unexpectedly high frequency of 928MHz. Few existing inexpensive u.h.f. power transistors or varactor multipliers could provide 5 watts output, though it is possible that some use could be made of superregenerative receiver techniques and s.a.w. (surface acoustic wave) u.h.f. oscillators.

928MHz meets the RSGB request that Open Channel should not be placed: close to an amateur band; it is conveniently almost exactly mid-way between the 432 and 1300MHz bands. Amateur experience on these bands shows that 928MHz is not necessarily a short-range "line of sight" band, particularly during conditions of anomalous propagation or from hill-top sites.

There remains the danger that Home Office efforts to reduce illegal activity on 27MHz could result in more "piracy" in the amateur bands, particularly 28MHz. In the USA, despite the availability of the 40 channels around 27MHz, there is already increasing intrusion into the low-frequency (c.w.) end of the 28MHz amateur band. Similarly despite efforts by the FCC to stamp out the use of high-power "linears" by c.b. operators (including forbidding the sale of any linears covering the 28MHz band) there are still c.b. operators using 2kW p.e.p. s.s.b. equipment.

The Home Office makes the valid point that "if an individual wishes to use sophisticated equipment to communicate over long ranges and make

international contacts, he should become a licensed radio amateur by taking the appropriate radio examination." The introduction of "multiplechoice" questions in the Radio Amateurs' Examination since 1979, and the consequently higher "pass rate," has removed the argument that amateur radio is open only to those experienced in taking written examinations. But it is to be hoped that the Home Office will consider the possibility of introducing some form of "novice" licence.

The fact that the Home Office is not proposing to allocate "call-signs" for Open Channel should also prevent its becoming a "shamateur" band and so help keep it as a useful and welcome facility for the general public, while not ruling out its use for "fun" purposes.

Amateurs in hospital

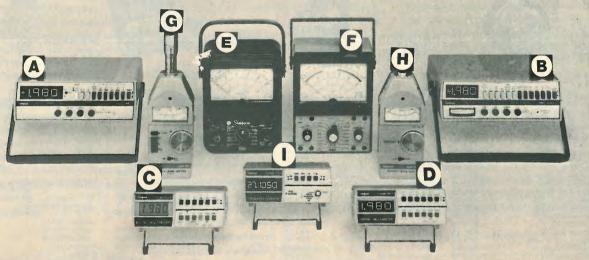
Fred Judd, G2BCX, points out that provided permission is obtained in advance, there is usually no objection to the use of amateur radio equipment in British hospitals. Permission needs to be obtained from the Unit Administrator and/or the District Works Officer of the hospital concerned and tests should always be made to ensure there are no electromagnetic compatibility (e.m.c.) problems with sensitive hospital equip-

In brief

Fee for the Morse test in the UK has gone up from £6 to £8... Danish amateurs now have permission to use 1720-1740 kHz and 1830-1850 kHz, 10 watts c.w. . . . "Rusty" Russell, G5WP, of Guildford, one of the only two British amateurs ever to have won the BERU Commonwealth Contest and a consistent "dx" operator on 3.5MHz, has died ... Australian amateur licences rose in 1979 from 10,587 to 12,596, of which 6,126 are "full" licences, 3,273 "limited" and 3197 "novice" . . . RSGB reports show that the number of RAE courses being run this season at local adult education centres is about 50 including many towns not listed last month Amersham, Birmingham (2), Borehamwood, Brentwood, Burgess Hill, Bury, Canterbury, Chester, Chingford (2), Cove, Crawley, Derby, Dudley, Exeter, Grafton, Harrow, Hemel Hempstead, Highbury, Huddersfield, Knottingley, Nottingham, Paddington, Southampton (2), Stockton-on-Tees, Stretford, Turnford, Wakefield and Walsall ... Forthcoming events include Welsh Amateur Radio Convention at Blackwood, Gwent on September 28 and the British Amateur Television Club Convention at Post House Hotel, Leicester on October 5 (from 11 a.m.).

PAT HAWKER, G3VA

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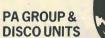


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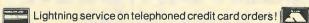
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Satellite broadcasting in the eighties

A report on technical progress in Europe

by G. J. Phillips, M.A., Ph.D., B.Sc., BBC Research Department

This article outlines the framework set by the 1977 ITU plan for satellite broadcasting in Regions 1 and 3 and reviews the work that has been done so far to implement it. After showing the coverage areas for different countries in Europe the author discusses the life expectancy and costs of broadcasting satellites, considers the design of domestic receiving equipment and aerials and, in a second article, will report on the current plans for building satellites by the European Space Agency and other groups.

We are in the decade of a new method of transmitting broadcast programmes television and sound - into the home. The transmitter is placed in an orbit above the equator at a height of 36,000km so that it moves round at the same rate that the Earth spins on its axis; it can thus remain at a fixed point in the sky. Doubts on the one hand and over-optimism on the other hand existed in the early seventies. Now, however, three points seem to be established: experiments have proved it works as expected; costs can be assessed and appear acceptable and, finally, it is increasingly appreciated that (whatever other methods of distribution such as optical-fibre cable may ultimately prevail in the nineties or beyond) it is a method of distribution that comes closer to the ideal concept of broadcasting than any other method. There is surely an elegant simplicity in a transmitter of about 100 watts being able to provide a television programme to any home within a moderatly-sized country provided there is access to a simple receiving system in line-of-sight from the satellite.

The ITU plan for 12GHz

A broadcasting satellite conference held in Geneva in 1977 agreed a plan for ITU Region 1 (Europe/USSR/Africa) and Region 3 (Asia/Australasia), in which orbit positions and frequencies were assigned to prescribed beams covering each country (or sub-division of a large country). This plan will not only ensure the orderly development of satellite broadcasting itself, but will also help to

avoid mutual interference problems with other services using the same frequency band, notably terrestrial microwave links. With a few exceptions, every country in Europe and Africa was assigned five frequency channels within the 11.7 to 12.5 GHz range. Polarisation (circular, clockwise or anticlockwise) was also specified. Each channel is suitable for a frequency-modulated television signal within a 27MHz bandwidth; the actual channel spacing was about 19.2MHz, giving 40 channels in all, but possible adjacent-channel as well as co-channel interference was allowed for in the plan. A regional conference to conduct a similar assignment plan for Region 2 (Americas) is due to take place in 1983.

In the 1977 plan a power was also specified for each transmission. This was derived on the basis of providing a power flux density of at least -103dB(W/m²), sufficient for good reception with an individual 0.9m

diameter antenna, for 99% of the time in the worst month*. Fig. 1 gives the areas of coverage to this standard for the UK and Ireland. Fig. 2 gives some examples of coverage on the same basis for four other cases: France, Luxembourg, Monaco and the large beam that was allowed on certain channels to cover the four Nordic countries as a group. These figures assume ideal pointing of the satellite antenna; it is seen the coverage areas allow some latitude for pointing error (0.1° maximum is assumed). Considerable overlaps occur: for example, the French beam covers southern England, Switzerland and northern Italy.

One can also indicate over what area

*Power flux density figures can be converted into the more familiar field strength figures by subtracting them from 146. The result is then in dB relative to 1 microvolt per metre. Thus $-103dB(W/m^2) = 43dB(\mu V/m) =$ 140uV/m. - Ed

Table 1: Channel assignments and orbit positions for countries of Western and Southern Europe

Orbit position	37* West	31* West	19* West	5% East
Lower half (11.7-12.1 GHz) Right-hand polarisation	San Marino 1,5,9,13,17 Lichtenstein 3,7,11,15,19	Ireland 2,6,10,14,18 United Kingdom 4,8,12,16,20	France 1,5,9,13,17 Luxembourg 3,7,11,15,19	Turkey 1,5,9,13,17 Greece 3,7,11,15,19
Upper half (12.1-12.5 GHz) Right-hand polarisation	Monaco 21,25,29,33,37 Vatican ¹ 23,27,31,35,39		Belgium 21,25,29,33,37 Netherlands 23,27,31,35,39	Cyprus 21,25,29,33,37 lceland ² 23,27,31,35,39
Lower half (11.7-12.1 GHz) Left-hand polarisation	Andorra 4,8,12,16,20	Portugal ³ 3,7,11,15,19	West Germany 2,6,10,14,18 Austria 4,8,12,16,20	Finland 2,6,10 Norway 14,18 Sweden 4,8, Denmark 12,16,20
Upper half (12.1-12.5 GHz) Left-hand polarisation		lceland 21,25,29,33,37 Spain ⁵ 23,27,31,35,39	Switzerland 22,26,30,34,38 Italy 24,28,32,36,40	Nordic ⁴ 22,24,26 28,30,32,36,40 Sweden 34 Norway 38

1. 0.6 degree beam except channel 23 which covers mainland Italy. 2. Covers Iceland, Azores and part of Greenland. Channels 27 and 35 registered under Denmark. 3. Same transmission channels also beamed to Azores (common programme). 4. Eight channels in a wide-beam covering Nordic countries; assigned to Finland (22, 26), Sweden (30, 40), Denmark (24, 36) and Norway (28, 32). 5. Same transmission channels also beamed to Canary Islands (common programme)

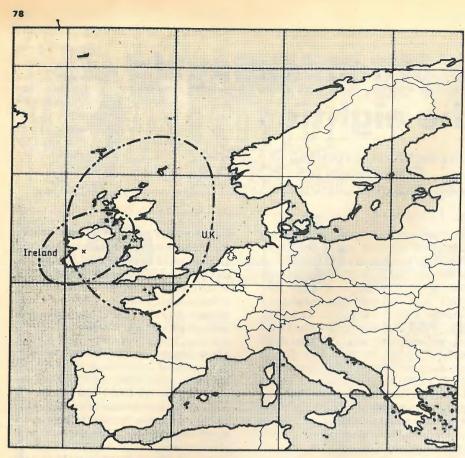


Fig. 1. Examples of coverage areas for individual reception in the UK and Ireland. Flux density is $-103dB(W/m^2)$.

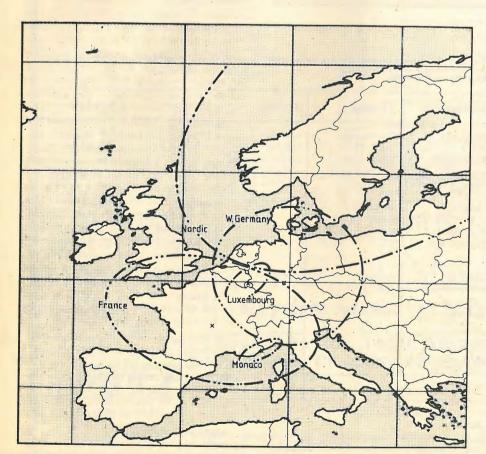


Fig. 2. Coverage areas for individual reception in W. Germany, France, Luxembourg, Monaco and the Nordic countries. Flux density is — 103dB(W/m²).

a satisfactory signal can be received with a somewhat larger receiving antenna. For convenience we can take the -111dB(W/m²) flux density limit corresponding to the level indicated at the 1977 conference for reasonably noisefree community reception; this is illustrated in Figs. 3 and 4 for the same countries. In some areas this coverage may not extend to the limits shown because interference from other satellites, while planned to be negligible for individual reception within the country to which the transmissions are aimed. may be slightly disturbing near the limit shown for community reception.

A summary of the allocations for all countries in Western Europe is given in Table 1. Where possible, a definite request by a country to have the same orbit position as another country was met in the plan; this facilitates individual reception of transmissions of two or more neighbouring countries in border areas, where there may be common interests, cultures or language.

In order to bring out the factors which affect the ease of reception in any country of transmissions other than those intended for that country, the table distinguishes groups of channels according to polarisation and whether they are in the upper or lower half of the 11.7 to 12.5 GHz band. This is discussed later.

Satellite life and costs

Transmitters in the sky are no new thing. We have employed geostationary satellites for more than a decade to relay telephone traffic and television signals between continents. The powers of the transmitters on a satellite for this purpose are generally below 20 watts; they also beam their signals over large areas. As a result a very large receiving antenna is needed on the ground (e.g. a 30m diameter reflector to receive 4GHz signals) in such point-to-point links.

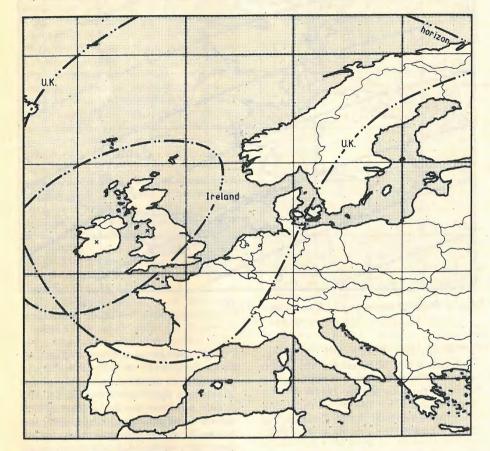
For broadcasting the available transmitter power is concentrated by beaming over the limited coverage areas, typically with a bandwidth of one degree for many European countries. A power of the order of 100-200 watts is then sufficient for individual reception with an antenna diameter of 0.9m.

The experiments in Canada and USA since 1976, with the CTS (Hermes satellite), have come closest to this concept and successfully demonstrated television reception with small terminals. The satellite employs a 200W repeater at 12GHz and the beam is about 2.5 degrees wide.

Two important points govern the costs of satellite systems. First, the reliability should be as high as that from current terrestrial services. Secondly, the satellite should remain accurately in its allocated position so that individual receiving antennae set up in fixed positions, pointing to the satellite, will remain effective. Because of orbit-

perturbing forces, station-keeping requires fairly frequent correction by gas jets on the satellite, and the quantity of fuel to operate these is the critical factor which governs the life of a satellite. A seven-year life is typical if a reasonable allocation of payload be-

tween the fuel and other essential items is made. Thus a reliable service requires a spare satellite in orbit and a third ready to be launched at any time, so the cost of making and launching a single satellite is not sufficient investment to provide a service for 10 or 20 years.



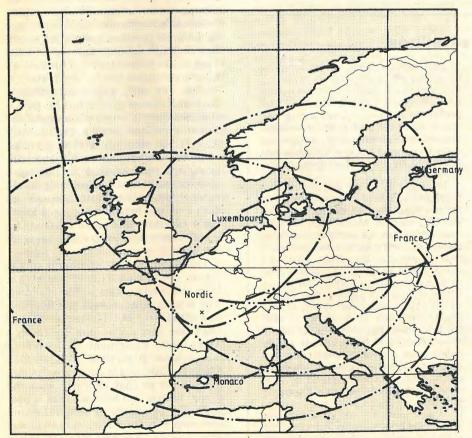


Fig. 3. Coverage area for reception on a larger antenna (1.8 to 2.0m diameter), for countries given in Figs. 1 and 2. Flux density is $-111dB(W/m^2)$.

For example, for 10 years' reliable service, allowance must be made for the provision and launching of five satellites. So if the cost of a satellite giving transmissions on four channels is £12, million, with a similar cost for launching, some £120 million is required over 10 years. 1 Nevertheless the cost of £3 million per annum per channel is actually less than total engineering cost to the broadcasters of providing a national service at u.h.f. by terrestrial transmitters. Of course, an overall national picture, taking into account the receiver cost, would show a somewhat greater total cost for a satellite system. National decisions to implement satellite broadcasting will have to take into account total costs but, if broadcasters' costs are not prohibitive, it is reasonable to expect that continuing development in receiver technology will provide receivers at a price acceptable to a steadily increasing proportion of the public.

The receiving antenna

For terrestrial television we are used to aerials which range from little more than the proverbial wet string near transmitters to large Yagis at the fringe which usually need mounting high up to get good signals.

Receiving antennae for domestic reception of satellite broadcasting signals, however, are uniformly sized because everyone will get a flux within the narrow limits of -100 to -103dB(W/m²) according to the standard mentioned earlier. Also, the 12GHz signals are such that a clear line-of-sight is usually essential, but with the angles of arrival of satellite signals for the UK, as shown in Fig. 4, almost everyone can find somewhere on their premises that meets the requirements. This has in fact already been confirmed² by asking the occupiers of several hundred homes to observe shadows when the sun was shining at 3 p.m. British Summer Time in mid-October this being a time when the sun has the same position in the sky for the UK as the assigned satellite position. The limited sample suggested that suitable sites for antennae could be found in 99.5% of cases. Furthermore in many, but not all, cases the most suitable site is low on the side of the house or on the ground rather than at roof level.

The requirement of a 0.9m diameter antenna is not a precise one. It depends on receiver noise performance, the available signal flux and the importance attached to a low-noise signal. As a guide, with the suggested antenna size, 8dB noise figure gives 14dB carrier-to-noise and a slightly noisy picture when the flux is just—103dB(W/m²). Manufacturers' developments now suggest that noise figures of 5dB with a mixer first stage, or 4dB with a f.e.t. amplifier, will be obtainable at modest cost. Some allowance for pointing error and reduced antenna efficiency should

be borne in mind when considering the likely performance of domestic equipment over a period of several years.

In order to exploit extensive frequency re-use in the 1977 plan, advantage was taken of the directivity of receiving antennae, corresponding to a 2 degree beamwidth (at -3dB) for a 0.9m dish. Antennae lined up on one satellite position have a poor response to signals from neighbouring positions. The plan used 6 degree spacing between adjacent allocated orbit positions, and at 6, 12 and 18 degrees respectively the antenna responses are assumed to be 20, 28 and 33 dB below the maximum. Furthermore it is assumed that in the direction of maximum response (i.e. for the same orbit position) the response to a signal with a polarisation opposite to that of the wanted signal is 20dB down. It is thus clear that an antenna, when set up for one orbit position, cannot be used by the viewer to receive from another. Receiver requirements are considered next, but clearly where a viewer in the UK wishes to receive transmissions planned for France or vice-versa, a first essential is either to have two antennae, or one that can be rocked between two carefully set-up aiming positions. Possibly, if there is a demand, some neater arrangement such as a single reflector with two feeds will be designed.

Domestic receivers

The ITU plan, as seen from Table 1, calls for a tuning range of 400MHz to receive

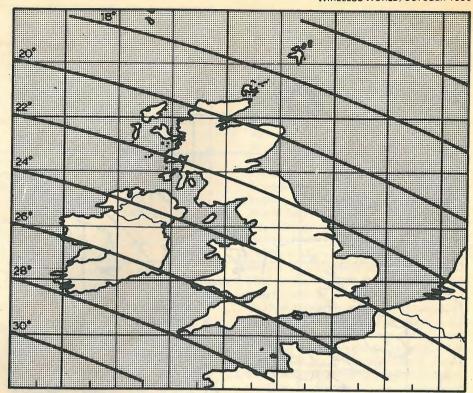


Fig. 4. Angle of elevation of UK satellite at 31 *W.

all channels of any one country, although it can be foreseen that, when several countries have begun using most of their channels, there will be some demand for means of receiving over the full 800MHz band. Considering the basic 400MHz receiving system first,

it might well be as outlined in Fig. 5. A down-converter to 900-1300MHz is placed on, or adjacent to, the aerial and employs a fixed-tuned oscillator, and avoids a microwave down-lead (which would be either lossy or expensive) but leaves the actual tuning in the room set. Secondly, terrestrial television must continue to be catered for since, in the foreseeable future, this would continue as the most practical system for television networks giving regional or local programme variations. Thirdly, an f.m./a.m. converter is not featured. Rather, the early appearance of sets with dual tuners (i.e. a.m./f.m. television receivers) is to be encouraged if the benefit from the picture quality with direct f.m. demodulation is to be obtained. (A f.m./a.m. unit demodulating and remodulating for feeding a conventional television receiver would have three intermediate frequencies with a change to base-band, i.e. a quadruple superhet; this could be prone to interference and suffer a.m. system distortion.)

The basic set-up in the home of the future could well expand somewhat from this modest start as Fig. 6 shows. The system now extends the distribution of the 900-1300MHz signals to more than one room. The one containing audio/radio equipment now uses an a.m. television receiver with a video input facility, and a separate f.m.-tovideo tuner so that interplay with the video cassette and television set is possible. The possibility of a digital sound multiplex in place of television on one of the satellite channels has also been anticipated. The assumption in this example is that the digital sound signal

UK's change of heart on satellites

In 1976 we reported that Britain's broadcasters were showing very little interest in the prospect of satellite broadcasting. Their general attitude seemed to be that the UK already had good coverage from terrestrial broadcasting and consequently there was little need for this new type of service. Since then they have shown a distinct change of heart. The BBC for example has stated that it proposes to take up two of the five satellite broadcasting channels allotted to the UK (see table in Dr Phillips's article), one for subscription television and the other carrying "the best programmes from and BBC2 Thames Television's director of sales has publicly discussed the interest of the tv programme companies. The IBA, though somewhat less positive, have said that if a national decision is made to establish satellite broadcasting in the UK they would not stand aside.

The reasons for this volte-face are largely commercial. First there is the fear of losing British audiences and advertising revenue to competition from Continental satellite broadcasters—notably from those countries whose satellite coverage areas "overspill" onto the UK. Secondly, programme companies in Britain see opportunities

to get revenue from advertisements broadcast into European countries by this means. In addition British space and electronics manufacturers see profitable markets in supplying the actual satellites and their associated ground equipment, and in seeking such business they are officially encouraged by the Department of Industry.

In response to all this the Home Secretary, who is of course responsible for the regulation of broadcasting in the UK, said in March that he had decided to launch a study of the implications of setting up a satellite broadcasting service by about 1985 which would be the earliest practicable date - or by about 1990. The Home Office is now conducting this study, which covers technical, financial and resource matters, in consultation with the BBC, IBA, Dol, other government departments and organizations which might have a direct interest. It takes account of the Government's plans for the fourth television network. The results of the study, which will present the various options and their implications as a basis for making a decision on satellite broadcasting, are expected to be published at about the end of this year. - WW staff.

would be approximately within the normal video bandwidth and would frequency-modulate the transmission in the same way as for television.

WIRELESS WORLD, OCTOBER 1980

A receiver for the international viewer can be considered with reference to Table 1. Here the situation may vary. Certain pairs of countries (UK and Ireland, France and Belgium, West Germany and Austria, for example) need no more than the basic receiver to see each other's programmes. In other cases the neighbour might have the other half of the 800MHz band, different polarisation, or both, A receiver to tune over 800MHz would be very convenient but could be difficult to design on the basis of a fixed oscillator and extending the range of the first intermediate frequency to 900-1700MHz. I would suggest a simple alternative: if the basic 400MHz units have a wide market and therefore reasonable cost, two such units could be attached to the antenna with two down-leads, each carrying 900-1300MHz, one for each half of the band. The receiver would have two input sockets and a two-way selector switch.

To change polarisation, a remotelycontrolled switched-polarisation feed could be fitted at the antenna. However, with more than one television set in the home, this solution could frustrate independent choice of viewing. An alternative would be an 'orthogonal feed' from which the left- and righthanded polarisations could be simultaneously connected to two basic first frequency-changers, again using two down-leads. Elaborating further, some 19° West satellite viewers in favourable locations might want to receive most or all of the eight European national services. They would require four 400MHz units and four down-leads in order to cover both polarisations over an 800MHz bandwidth.

To complete the picture on the receiver design the u.h.f. tuner must be considered. This will select channels within the first i.f. range and the conventional approach would be to have a second, tunable, frequency-changer and a final i.f. in the region of 125MHz. Image rejection would be necessary in this tuner. A surface-acoustic-wave filter could be a good choice for the i.f. filter in front of the f.m. discriminator which has to operate with a 27MHz

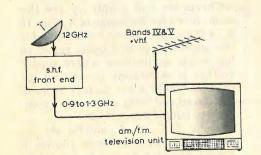


Fig. 5. Simple system with an a.m. / f.m. room television receiver unit.

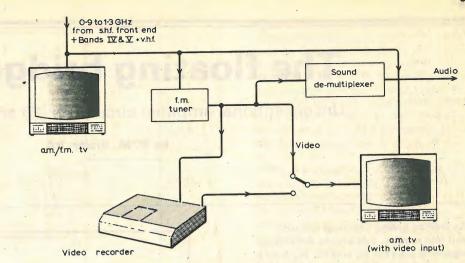


Fig. 6. Possible developments for unit video and unit aidio in the home.

bandwidth with low group-delay distortion. A less conventional approach, under study in France, is a phase-locked-loop f.m. demodulator which can operate directly on the required signal in the 900-1300MHz band and provide a video output directly.

The method of transmitting the television sound component in satellite broadcasting is under active study by the European Broadcasting Union. Although the starting point in 1977 was to consider a f.m. subcarrier compatible with the terrestrial system (e.g. a 6MHz subcarrier in the case of television Standard I, as used in the UK), serious consideration is being given to alternatives for enhancing the sytem to provide a pair of channels for stereo or second language. Digital modulation for the sound signal, which could give better quality and higher signal-to-noise ratio than is possible with analogue systems, is also being considered.

Feeder links (up links)

The system envisaged for a broadcasting satellite system is that a signal modulated to exactly the same standard as the downcoming transmission should be sent up to the satellite. This allows the satellite to be designed as a frequency-translating relay. The detailed assignments, or even the choice of the frequency band, for the up-links or feeder links were not dealt with at the 1977 conference because priority was given to attaining an agreed down-link plan. Studies having established that a bandwidth at least as great as the down-link broadcasting band would be essential, the 1979 World Administrative Radio Conference was able to allocate for world-wide use the band 17.3 to 18.1GHz for feeder links to broadcasting satellites transmitting in the 12GHz band. In limited geographical areas, alternative frequencies have also been allocated for feeder link use, if required, including the 10.7 to 11.7GHz band in the European area.

Although interim arrangements can easily be made at early stages when the band is relatively uncrowded, it is agreed that a detailed assignment plan for feeder links should be made and the expectation in the case of Regions 1 and 3 is that the frequency channelling and assignments could, as a starting point, be a carbon copy of the down-link plan translated from 11.7-12.5GHz to 17.3-18.1GHz. One advantage would be a constant frequency change of 5.6GHz for all transponders, which would lead to some economies in design. If changes or adjustments had to be made to the assigned frequencies, they could be made without enormous repercussions (as would be the case in attempting to change the closely interwoven downlink plan). This is because up-link antennae at Earth stations are expected to be so directional that the choice of frequencies for transmissions aimed at one satellite would have little effect on the choice for transmissions aimed at other orbital positions. The main task is to agree on a series of mini-plans, each acceptable to a group of countries assigned to one orbit position, but with an overall check on the effect on other orbit positions. An ITU conference to be held in 1983 for planning down-links in Region 2 will also be asked to consider detailed feeder link planning for the same region. It is not yet decided whether the conference needed for feeder link assignments in Regions 1 and 3 will take place at the same time or later.

To be continued

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The floating bridge — 2

Unconventional amplifier circuits for 15 and 200 watts

by R. M. Brady, BA

As well as giving practical circuits and test results, this article describes a general plan for A₁ and for A₂ which makes full use of the unique way in which these circuits may be simplified. In particular by using an i.c. which is able to control A1 without loss of performance, and by exploiting the fact that A need be made only to poor performance specifications. It also takes a further, more quantitative, step in distortion analysis. The unconventional approach to these circuits, was outlined in the first part of this article, September issue.

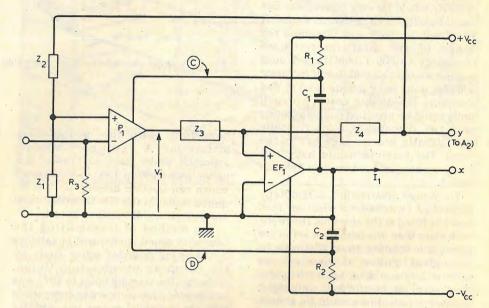
A design for A, which uses a B-type feedback loop with the simplest earthing system is shown in Fig. 11, and although the following analysis is based on this circuit, it applies equally well to A-type bridges, and to circuits incorporating a change-of-origin device. P₁ is a high voltage-gain i.c., and EF, is what would conventionally be called an emitter follower, containing the power transistors. C and D are points which are kept at a constant potential with respect to earth by C₁ and C₂, and they act as power supply points for P,* Impedances Z₁, Z₂, Z₃ and Z₄ are part of the feedback loops of the circuit.

Closer inspection of the circuit shows that EF, is arranged in an A-type feedback loop, and its complex voltage gain is Z_4/Z_3 (not unity, so that "emitter follower" is probably a misleading name for this part of the amplifier). It is this ability to extract voltage gain from EP1 which makes this circuit uniquely suited to be controlled by an i.c. - it may be arranged that the voltage gain demanded of the i.c. is around unity, so that a high bandwidth may be obtained.

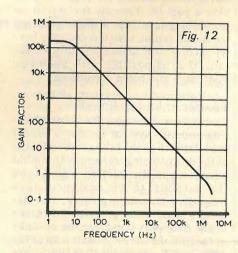
If G is the gain of the i.c., and if Z₃ and Z4 are small enough that the gain of the emitter follower approximates to Z₄/Z₃, then the open-loop gain of the whole amplifier becomes

$$\frac{GZ_{4}Z_{1}}{Z_{3}(Z_{2}+Z_{1})}$$

In the actual circuits, described later, $Z_3/Z_4 = Z_1/(Z_1 + Z_2)$, so that this expression reduces to G **. The 741



amplifier is used and as this is frequency compensated the whole amplifier is stable. Fig. 12 shows the loop gain of the 741 amplifier as a function of frequency.



To quantify the coupling between A₂ and A1, imagine that A2 produces a distortion signal of V volts. At low frequencies this couples into A1, largely via the extra current 2V/R, which is injected into point x through R1 and R2 $(R_1 = R_2)$. If the impedance of the emitter follower (V_1/I_1) in Fig. 11 is Z, and the loop gain of P, is G, then this current results in a distortion voltage across y-x of

$$\frac{2VZ\left(Z_{1}+Z_{2}\right)}{R_{1}GZ_{1}}$$

Fig. 11. Circuit for amplifier A, based on version B. Z, Z, and R, are part of the feedback loop of Fig. 5 with Z = R, C, $Z_{1} = R_{2}$, $R_{3} = R_{5}$. EF, is arranged in an A-type feedback loop with voltage gain Z₄/Z₃ so only low voltage swings are needed in the operational amplifier P ..

Taking typical values of Z = 0.1 ohm, (Z. $+ Z_1)/Z_1 = 30$, and $R_1 = 220$ ohm the distortion coupling is around 3/G percent at low frequencies. Inspection of Fig. 12 shows that this coupling is hardly worth considering. At high frequencies, the value of $(Z_2 + Z_1/Z_1)$ falls, to some extent counteracting the fall in gain of the i.c., and also the distortion produced by A2 diminishes because of the low response cut-off of this amplifier.

Cheapness and simplicity are the main criteria for the design of A2. Figs 13 & 14 show two alternative designs for this amplifier, the first being suited to low-power applications where supply voltage is accurately controlled for example where a car battery is used, and the second being a little more complicated but far more robust.

In Fig. 13 C₁, R₁, C₂ and R₂ are the same components as those shown in Fig. 11 Resistors R₃ and R₄ are set so

Robert Brady is researching in superconductivity at the Cavendish Laboratory, Cambridge.

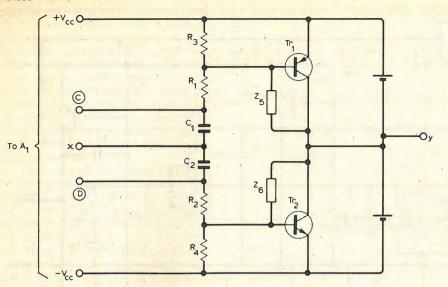


Fig. 13. Simple design for A₂ — cutting all corners because distortion in this amplifier does not affect the output. So that y = -x. Gain should be about -1 and is $Z_5/R_1 = -x$. Z_6/R_2 At high frequencies, the gain must fall and so a capacitor is included in each of Z_s and Z_s . Components C_1 , C_2 , R_1 and R_2 are needed for A_1 also and so serve a double

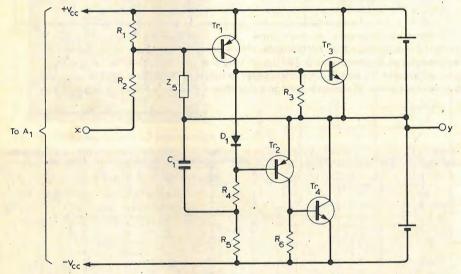
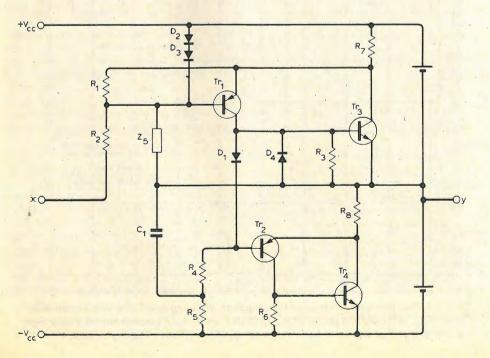


Fig. 14. High power design for A_2 . This is a more conventional design, with gain Z_5/R_2 .



that in the quiescent state Tr, and Tr, are only just turned on (one of R3 and R4 may be a preset if desired). As x rises with respect to point A, C, and C, pull C and D up with it, thus switching off Tr₁ and switching on Tr2, so that y falls. The gain of A₀ is (y-A)/(x-A) and is a little less than Z_6/R_2 ; it is arranged to fall off at around 5 to 10kHz where full power output is not required from

A₁.
Fig. 14 shows an emitter follower version which is controlled by Tr, so that the gain is Z_5/R_2 . To prevent large distortion at low signal levels due to effects at the crossover point, R₃ is included.

A large open-loop gain is made possible by low values of Z_5 and R_2 , so that Tr, and R3 are capable of helping substantially at the crossover point.

Fig. 15 shows this circuit with current protection incorporated. When Tr. and Tr₃ are conducting, a voltage is showed across R₇ which is proportional to the current flowing. When this voltage becomes about 0.6V D₃ and D₂ conduct, preventing large currents from passing through Tr3; a similar mechanism around D₁ and D₄ protects the Tr₂-Tr₄ pair. If desired, more accurate control may be obtained by inserting a p-n-p silicon transistor in place of D₂ and D₂, with its emitter to the + supply, its collector to the base of Tr, and its base to the emitter of Tr₁; and a second in a similar configuration around Tr₂.

12 volt 15 watt amplifier

A circuit for a type-B bridge amplifier which is suitable for operation using a 12 volt power supply is given in Fig. 16. Comparing this with Fig. 11 components R₁ and C₁, R₃ and C₂, R₄*, and R_5 and C_3 are represented by Z_1 , Z_2 , Z_3 and Z4 in the block diagram representation. Transistors 2 to 5 are represented by EF, and C, C, R, and R_{10} have their counterparts in C_1 , C_2 , R_1 and R2. Capacitors 6 & 7 are represented in Fig. 13 by Z₅ and Z₆. The circuit shown differs from the plan in that Z₅ and Z₆ are largely reactive, so that at low frequencies the gain of A2 is dictated mostly by the current gain of Tr₆ and Tr. Notice P, is being operated with a low supply voltage, so that a potentiometer is needed to control the

Fig. 15. High power design for A2 with current protection. This protects both amplifiers A, and A, but it is easier to put current protection into A, as quality output is not required from this amplifier. Current must pass through R, - if too much passes D, and D, conduct switching off the drive, or through R₈ when D4 and D1 conduct it too much

^{*}C is the "semi-stabilized" voltage point referred to

^{**} The actual situatiuon is a more complicated, but this analysis is a good approximation

^{*} In series with the output impedance of P1 which is

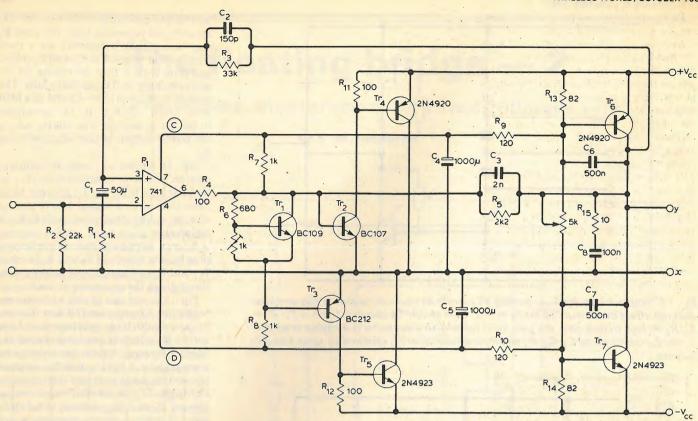


Fig. 16. 12-volt version of the floating bridge gives 12 watts output, distortion-free, into 4 ohms; more power is available if a lower impedance loudspeaker is used. High quality amplifier A_1 comprises a 741 i.c. driving the "emitter follower" stage Tr_2 and Tr_3 , gain R_3/R_1 . Low-quality amplifier A_2 is based around Tr_6 and Tr_7 . Potentiometer is adjusted until the output voltage y is half way between the supply rail voltages in the quiescent state. If a supply voltage other than 12 volts is used R_{13} and R_{14} may need adjustment.

quiescent value of y accurately. Components R_{15} and C_8 are included to by-pass inductive loads.

A computer analysis of the response of this amplifier is shown in Fig. 17 in which G is the signal gain, stretching across the whole audio bandwidth. It is dangerous to attempt to restrict this response because the loop gain L is already falling at 6dB per octave and so any interference would almost certainly result in instability. Also on this graph is the calculated A2 rejection factor, which does not fall below 55dB hence distortion induced in the output by A2 is negligible. The distortion to be expected of this amplifier at medium signal levels is discussed in the appendix. The performance to be expected of the amplifier at high and very low signal levels is not very different from that discussed here.

Fig. 18 shows a change of origin suitable for this amplifier; in the circuit tested it was mounted on the same board as the main amplifier, though this is not necessary. Point F is suitable for use as a power supply point for a preamp, though this may need slight modification if the pre-amp takes significant current. If a positive earth is to be used, the change of origin may be modified by changing ${\rm Tr}_{101}$ and ${\rm Tr}_{102}$ for similar p-n-p devices e.g. BC212, point C for point D, and ${\rm +12V}$ supply for ${\rm -12V}$ supply, so that the input signal is now compared to the ${\rm + rail}$.

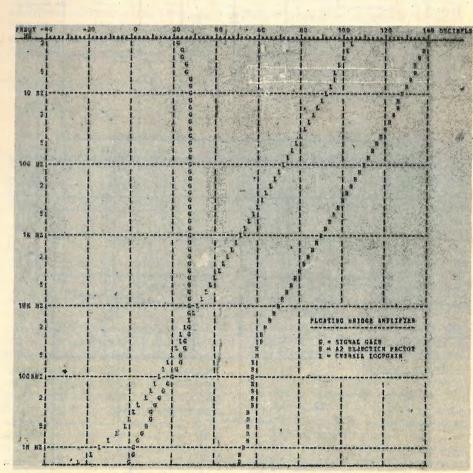


Fig. 17. Predicted performance of the amplifier, showing open-loop and closed-loop gain of A_1 , and rejection factor of A_1 , to distortion in A_2 . Rejection factor is very high, showing that A_2 can indeed be of very poor quality without affecting output.

Before switching on for the first time, set the preset resistor to its maximum resistance, and set the potentiometer to the middle of its resistance range. When power has been switched on, adjust the pot so that x=Y=V/2 where V is the power supply voltage. Now insert an ammeter in the power rail, and adjust the preset until the current begins to increase rapidly as it is turned, and then adjust back a little. The amplifier is now ready for operation.

If the change of origin is to be mounted on a separate board, use the following screening system. Connect the screen of the coaxial lead from pre-amp to main amplifier to point x at the amplifier end only; leave the other end flying. (If convenient provide a track in the pre-amp to which the screen may be physically mounted). This cable is at x potential, so keep it well away from the sensitive components of the pre-amp.

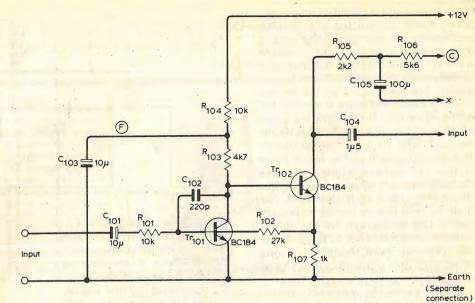


Fig. 18. Practical circuit for a "change of origin" for the 12 volt amplifier allows earth to be connected to the negative power rail. If a positive earth is required, use BC212 for Tr_{101} and Tr_{102} and replace point C by point D, so that a "mirror image" may be built.

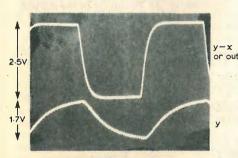
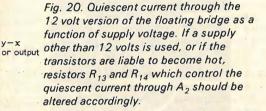
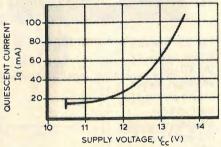


Fig. 19. Response of the 12 volt version to a 3kHz square wave input. Upper trace shows the 30kHz cutoff due to C_2 in Fig. 15. Lower trace is the output of the poor quality amplifier. A_2 showing cutoff at about 5kHz. Note that the output, y-x, is decoupled from the A_2 output, y.





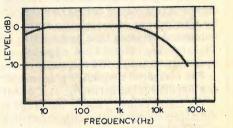


Fig. 21. Signal response of the A₂ amplifier as a function of frequency.

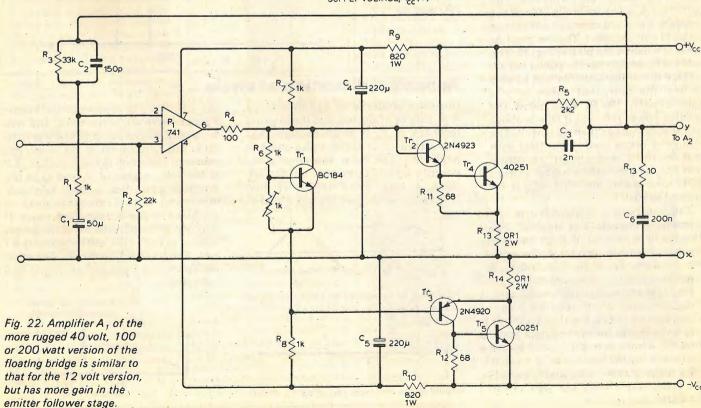


Fig. 20 shows the quiescent current taken by the amplifier as a function of supply voltage. Below 10.5 volts the i.c. becomes unstable and low frequency oscillation sets in, whereas above 13.5 volts A₂ begins to take an unacceptable current. (This may of course be changed by altering the value of the resistors connected between base and emitter of Tr₆ and Tr₇. The i.c. is capable of operating with up to 30 volt supply.)

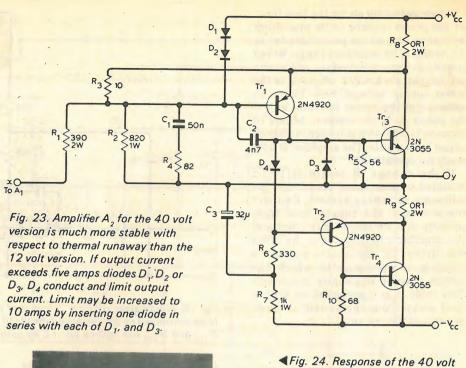
The measured frequency response of the amplifier is shown in Fig. 21. Cut-off at the high frequency end of the range is didctated mostly by the characteristics of the change of origin device, and these are fairly "safe" if changed.

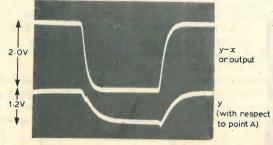
40 Volt 200 Watt amplifier

A 40-volt amplifier design is suitable for high power applications, and is capable of driving an eight-ohm load at 100 watts or four-ohm load at 200 watts, with high fidelity performance. The circuit for A₁ is shown in Fig. 22 and the feedback loops are identical with those in the 12 volt amplifier. The design of A2 in Fig. 23 follows the plan of Fig. 15. The value of R₃ determines the power supply voltage which the amplifier will be able to use: the quiescent value of y is approximately 190/R3 volts below the positive power rail, and if this is about half supply voltage then all is well. The change of origin used is identical with that for the 12-volt circuit, excepting that Tr₁₀₂ is replaced by a higher voltage BC147 transistor; and that R₁₀₄ is increased to $27k\Omega$.

The distortion analysis given in the appendix is valid for this amplifier also Though finer control of the crossover region may be had because of the emitter resistors. Tr₄ is now included in possible sources of crossover distortion.

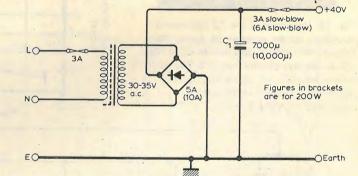
Fig. 24 is an oscilloscope photograph of the response of this amplifier to a 3 kHz square wave input. Hideous crossover distortion in the lower trace does not couple into the signal across the load. A further point to notice is that the y signal is more accurately controlled, being very closely one half of the y-x signal.





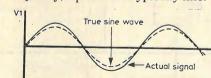
version to 3kHz square wave input. Upper trace is the output, y—x, and appears to be distortion-free despite the very bad crossover distortion in the output y of amplifier A₂ (lower trace).

Fig. 25. Suitable power supply for the 40 volt amplifier. Values shown are for 100 watt version; alternative values in brackets are for 200 watt use.

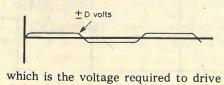


Appendix: distortion analysis

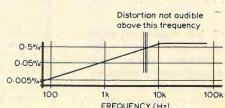
Distortion produced by A_2 and induced in A_1 may be neglected, and the internal feedback in P_1 is large enough that distortion of D G/3P volts peak is neglected. The main source of nonlinearity is the EF_1 system, see Fig. 11. With a sine wave input at low frequency, V_1 will look typically like:



This may be regarded as the sine wave plus an extra:



the $\mathrm{EF_1}$ if $\mathrm{Z_4}$ is removed. Its major Fourier harmonic is the third, and this has a peak magnitude of a little less than D/3. This component is the dominant source of distortion in the amplifier. If P is the voltage gain of $\mathrm{P_1}$ and G is the amplifier gain, then a third harmonic distortion of D G/3P volts peak is produced at the output. The value D depends on the adjustment of the preset resistor in Fig. 16 but will be typically 0.1 volt. The resulting harmonic distortion for a pure sine wave and an output of 5 volts peak is:



NEW PRODUCTS

New computer products

The fourth Microcomputer show, which this year moved to the Wembley Conference Centre, attracted 52 exhibitors and increased the attendance by 500 to 8,500 over three days. Although the show specializes in small business systems and personal computers, James Scott Electronic Developments reported a sale of v.d.us, worth around £1/2 million, to a German client. Comart were demonstrating a North Star Horizon hard-disk system which provides 18M bytes, expandable to 72M bytes, of storage on a 14in Winchester disk/drive. The Byte shop, now a subsidiary of Comart, announced a Prestel board which uses their own software to create Prestel compatible colour displays that can be stored, edited and transmitted.

A new serial printer from Mannesmann Tally features a print mechanism and print head which are claimed to offer a double life of 200 million characters. The printer, type MT1602, looks like the T1602 but offers improved performance. A microprocessor selects the shortest print path and accelerates the head across blank spaces to provide a printing speed of about 160 cns

Micro byte's new "stereo S100 sound computer board" is an interesting add-on for the hobbyist. This card uses two AY-3-8910 sound i.cs which accommodate three tone channels, three amplitude controls, a noise generator, a 16-bit envelope period control, two parallel I/O ports and three d-to-a converters. The card is supplied in kit form with four parallel I/O ports, two amplifiers and a prototyping area. A 60 page manual supplements the card, and Basic and assembly language programming examples are provided.

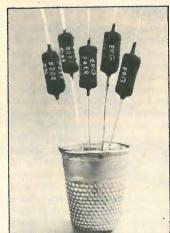
A single-board computer kit was also announced by Micro byte, who expect deliveries to start in September. The board comprises a Z80 operating under CP/M, 64K of r.a.m., one floppydisk controller, serial and parallel ports, and a 24×80 v.d.u. Very little literature was available at the show, but the advertised price of £395 + v.a.t. makes it worth investigating.

BMG Microsystems, a wholly British computer company, has developed a production management system, based on their MS5000 microcomputer, for scheduling and cost control.

BMG say that the system provides small and medium sized engineering companies with a versatile tool that is easy to use and does not require computer expertise. The standard system uses an 8in floppy-disk and a single v.d.u., but this can be expanded to include several v.d.us and up to 20M bytes of exchangeable disk storage. The makers are currently working on the implementation of a 16-bit processor and larger disk capacities.

Miniature power resistors

Wirewound resistors of small physical size, yet with good surge capabilities and power ratings up to 3W at 20°C, are introduced by Erg Components. With features such as an all welded construction for good reliability, a black silicone resin coating and a temperature coefficient of typically



60 p.p.m./°C for values above 1Ω, the Erg 74ER series is said to fill in a gap between metal-oxide and vitreous enamel w.w. type resistors. Their performance complies to BSE 9114 N001, and they measure 11.8mm long with a diameter of 5.25mm. Resistance values down to 30milliohm are available. Erg Industrial Corporation Ltd, Luton Rd, Dunstable, Bedfordshire LU5 4LJ.

WW301

Btu meter

A portable infrared sensing device, the Thermoflow, which enables the measurement of heat losses and gains without actual contact, is now available from Unity Power Systems. A large digital display gives a direct readout in Btu/ft²h when the unit is pointed at the source, making it a



simple matter to determine heat flow over a certain period of time. Two modes of operation, selected by means of a trigger switch, enable either the checking of heat losses from steam, pipes, walls, windows etc, or the checking of heat loading from such items as lighting units and electronic appliances. Features are battery operation, automatic ambient temperature compensation and an 8-14 micron filter which eliminates potential errors due to water vapour, carbon dioxide, sky radiance and reflected sunlight. The instrument is supplied complete with carrying case, weighs about lkg and costs in the region of £600. Unity Power Systems, Pembroke House, 44 Wellesly Rd, Croydon, Surrey CR9 2BU.

WW302

Eurocard boards

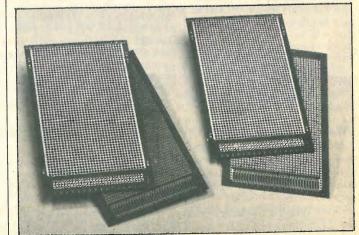
Extensions to Vero Electronics Ltd's Eurocard range have been made to include fully pierced Veroboards, which are available with or without maximum copper colander ground planes. Positions are provided for linking directly to board or pins from connectors, and either soldering or wire-wrapping techniques can be used for wiring. Grid reference numbers are silk-screened onto

the component side for the convenience of the user, and their size makes them compatible with subracks to DIN 41494, IEC 297 and SC 48D specifications. Also introduced are single-height, square-pad boards with maximum copper colander ground planes. These boards have been designed primarily to accommodate wire-wrap sockets. With holes accurately positioned on a 2.54mm matrix, they will accept any size of integrated circuit. Vero Electronics Ltd. Industrial Estate, Chandlers Ford, Eastleigh, Hampshire SO5 3ZR

WW303

Digital storage oscilloscope

Storage of waveforms in a digital memory of 8-bit × 1024-word capacity is the main feature of the MS-1650 oscilloscope from Trio, which, when used as an ordinary oscilloscope, has a frequency bandwidth of zero to 10MHz, and an input sensitivity of 10mV/div. Maximum write speed is lus/word, and analogue input signals may be sampled at any time, converted into 8-bit digital signals by an a.-to-d. converter, and then stored temporarily in the memory. The stored signal may be displayed immediately on the oscilloscope screen, or used to drive a pen recorder via the memory output. Ability to store the signal generated prior to the trigger pulse, facilitates the storage and display of one-shot, transient and repeated signals, and simultaneous display of stored and real-time waveforms, which may also be overlapped, enables comparisons to be made. Retention of the memory data when the power is removed, is possible by means of an optional NiCad battery for which space is provided inside





the standard commands to which

it can respond are "carriage

return", "line feed", "form feed"

and "shift out" (cursor home).

Cursor addressing can be carried

out by using device control

channels one and three. Standard

panels are manufactured in a

variety of forms, from a single

line of 40 characters, to a 480

character (40×12) message

panel, all having 120° viewing

angles in both planes and 5×7

dot characters. Operation at data

rates from 150 to 19 200 baud is

possible. Other features of the

new Argus display, which is ex-

pected to be of particular use in

applications where weight, size

and power-consumption are

critical factors, are its high-

brightness and dead background.

optimum line-to-line spacing,

and a flicker-free display during

updating, made possible by the

use of superior data organization.

Perdix Components Ltd, 98 Crof-

Angular displacement can be

measured directly using the

CllK capacitive sensor from

Jackson Brothers (London) Ltd.

Error in linearity is less than 1%

f.s. of the 100pF devices, which

have a differential arrangement

consisting of two sets of stator-

vanes, and one set of rotor-vanes.

enabling their use in bridge

circuits for improved accuracy

and cancellation of the effects of

environmental changes. The

standard unit costs around £10

with electrical characteristics

such as an insulation resistance

of 10°Ω, 500V d.c. breakdown

voltage, 50 p.p.m./°C tempera-

ture coefficient and a Q of 1000.

Continuous rotation can easily be translated, with minimal step

inaccuracies, into a triangular

waveform, which has a pitch

corresponding to 180° of angular

movement. A virtually unlimited

life is claimed for the C11K, the

only points subject to wear under

normal conditions being the

constant-contact wiper and the

low-torque bearings. Maximum

ton Park Rd, London SE4.

Capacitive sensor

the cabinet. The MS-1650 incorporates a 118×98mm c.r.t., weighs only 9kg, measures 284×138×400mm, and costs £1440 with a two year guarantee. House of Instruments, 34/36 High St, Saffron Walden, Essex CB10 1EP.

WW304

Multi-turn encoders

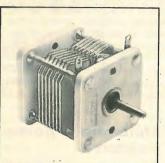
Tracking absolute encoders with resolutions of up to 1 part in 500 000, and a choice of 10, 64 or 100 turns for full-scale count. have been introduced to the UK by Techmation Ltd. These units. manufactured by Computer Conversions Ltd, convert any shaft input to 5 or 6 digit b.c.d. or 19 bits of binary information, corresponding directly to the shaft angle, with an error of less than ±1 part in 105. Output data is continuously available, accurate up to input rates of 104 degrees/second, and in addition, readout units are available with either 4, 5 or 6 digit, 0.5in high displays, and have t.t.l.compatible data, busy and inhibit outputs for interfacing with a computer. Resetting of the zero point to any value is possible via. an offset adjuster, and any output scale factor can be provided (such as pounds, feet, etc.). Rack or panel mounting versions exist, either with an internal power supply or without, in which case external supplies of ± 15V and +5V d.c. are required. Other specifications include 0 to 70°C (or -55 to +85°C) operating temperature range and a maximum transducer/readout-unit cable length of 1000ft. Techmation Ltd, 58 Edgware Way, Edgware, Middlesex HA8 8JP.

WW305

Matrix panel

ASCII compatibility enables the new version of the Argus gas plasma display panel, from Perdix Components Ltd, to be used as a direct replacement for a c.r.t. It can accept ASCII data in one of three ways, 20mA Loop, RS232-C or differential t.t.l., and among

operating torque is less than 20z-in and the sensor, with vanes made from silver-plated brass, measures 1.3cu.in without shaft and solder-lugs. Its drive shaft protrudes from both ends, one end having a plain diameter of 1/8in, the other stepped, with diameters of 1/8in and 1/16in, and mounting into transducer heads, linear/rotational translator units etc., is possible via two thread bushes, with one inch



centres, in each ceramic endplate. Jackson Brothers (London) Ltd, Kingsway, Waddon, Croydon, Surrey CR9 4DG.

Microprocessor

Developed with the aid of the Newcastle Science and Technology Education Centre, this desk-top unit, called the Microprocessor Tutor (MPT), is manufactured by Welwyn Electric Ltd, and has been designed to teach the uses and applications of the microprocessor simply and cheaply in schools, colleges and universities. It is claimed that the MPT has already been tested in more than fifty educational establishments. Program instructions and data are entered in the form of 8-bit words by means of eight switches with "0" and "1" positions, the program and data entry being indicated by a row of eight l.e.d.s, and the instruction set is intentionally limited to load,

store, add, substract, and, complement, branch (always), branch (if accum=0) and halt, for reasons of simplicity. A 'step' button allows programs to be run, one instruction at a time, with the l.e.d.s displaying the current data or the address at each step. Among the operations which can be demonstrated using the MPT, are the entering, storing and recalling of digital data, the addressing of memory locations and stepping from one instruction to another automatically. For practical demonstrations, a "traffic lights" simulator is provided along with instructions for writing programs to control their speed of operation, and program modifications to enable the delay time of the lights to be automatically changed as "traffic" builds up. Also included is a small d.c. motor for use in conjunction with motor-speed control programs. Welwyn Electric Ltd, Bedlington, Northumberland NE22 7AA.

WIRELESS WORLD, OCTOBER 1980

Wire-wrap kits UK-manufactured kits, each

comprising base board, precision screw-machined socket terminals, 96-way DIN connector(s), ejector keys, solder clips and a pin insertion/extraction tool, are made by Cavac Systems Ltd, and offered in single, double or triple Eurocard and maximum I/O double Eurocard styles. They feature maximum-power and ground-plane areas and are developed for use in prototype and pre-production applications. An alternative version can be supplied with a selection of discrete i.c. sockets instead of loose socket/terminals. A full data sheet, detailing the component parts and giving ordering information, can be obtained from Cavac on request. Cavac Systems Ltd, Unit 15, Suttons Park Avenue, Suttons Industrial Estate, Early, Reading RG6 1AZ.





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* FULL OPEN AND SHORT CIRCUIT PROTECTION GUARANTEED STABLE INTO ANY LOAD.

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* INTERLOCK CAPABILITY FOR UP TO EIGHT UNITS.

* 3-YEAR PARTS AND LABOUR WARRANTY.

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The fact, for example, that a portable. JVC colour camera costs little more than an ordinary black-and-white camera.

And the further fact that by adding a JVC VHS you have a complete colour recording system for as little as £1,300 plus VAT. For playback, a standard TV receiver is all you need.



At these prices every user can benefit from colour. Training will be easier to understand; publicity more compelling; management communications more interesting; rôle-playing more effective. After all, we live in a coloured world.

PUSH-BUTTON FEATURES

Don't think for one minute that the low price has been achieved at the expense of useful features. Among other things the camera has an iris control which automatically adjusts lens aperture to match lighting conditions; a 6:1 power or manual zoom, giving close-ups as close as 50 mm; TTL indicators which automatically show exposure level. auto-white balance, operating mode and power level.

BETTER STILL

Or, if you feel inclined to make even fuller use of the camera's capabilities, couple it to a JVC 3/4-inch U-format recorder.

The picture will be improved. You'll have another

sound track to use for foreign-language commentaries or question-and-answer training routines.

On 34-inch, moreover, you'll be in the right format to edit and duplicate - or add in library material. And still



the cost of the system needn't exceed £2,700 plus VAT. Alternatively, at very attractive rates, it can be leased.

SEE FIRST, THEN DECIDE

You can, of course, spend more. At any Bell & Howell Video Centre vou'll see more expensive cameras, video recorders and electronic editing equipment that wouldn't be out of place in a national network.

But do you need them?

Let the Video Centre, or Bell & Howell, help you decide. Whatever your decision, two things are certain.

One, colour video now costs a lot less than it used to (as well as being highly dependable and very easy to use).

Two, every unit in the system you choose qualifies for the Supershield warranty unique to Bell & Howell.

Under Supershield, all adjustments, repairs and replacements (except for tubes and tapes) are free for two years after purchase. And if a job can't be done on the spot we also provide free transport anywhere in mainland Great Britain to and from a fully equipped Supershield video workshop.

Convert to (or start with) colour. With JVC video equipment. And the Bell & Howell Supershield guarantee.

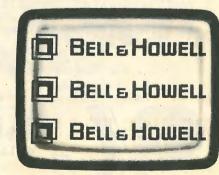
Let Bell & Howell show you the answer.

To Pieter Glas, Bell & Howell A-V Ltd., Freepost, Wembley, Middlesex HAO 1BR. Please send me more information about video equipment and a list of your Video Centres.

Organisation

JVC CAMERAS. JVC RECORDERS. JVC STUDIO EQUIPMENT. JVC MONITORS. ELECTROHOME MONITORS. FUJI VIDEO TAPES.

WW - 079 FIR FURTHER DETAILS



8K ON BOARD MEMORY! 5K RAM, 3K ROM or 4K RAM, 4K ROM (link selectable): Kit supplied with 3K RAM, 3K ROM. System expandable for up to 32K memory. 2 KEYBOARDS! 56 Key alphanumeric keyboard for entering high level language plus 16 key Hex pad for easy entry of machine GRAPHICS! 64 character graphics option — includes transistor symbols! Only £18.20 extra! MEMORY MAPPED 0 High resolution VDU circuitry using discrete TTL for extra flexibility. Has its own 2K memory to give 32 lines for 64 cha-KANSAS CITY PSI COMP 80 **Z80** Based powerful

scientific computer. Design as published WIRELESS WORLD

Cabinet size 19.0" × 15.7" × 3.3"

The kit for this outstandingly practical design by John Adams published in a series of articles in Wireless World really is complete!

Included in the PSI COMP 80 scientific computer kit is a professionally

finished cabinet, fibre-glass double sided, plated-through-hole printed circuit board, 2 keyboards PCB mounted for ease of construction, IC sockets, high reliability metal oxide resistors, power supply using custom designed toroidal transformer, 2K Basic and 1K monitor in EPROMS and,

KIT ALSO AVAILABLE AS SEPARATE

(10 × 12.3) 143.20. Pair of revposaros £34.80. Firmware in EPROMS £30.00. Toroidal transformer and power supply components £17.60. Cabinet (very rugged, made from steel, really beautifully finished) £26.50. P.S. Will greatly enhance any other single board computer including OHIO SUPERBOARD for which it can be readily modified. Other packs listed in our FREE CATALOGUE.

PSI COMP 80 Memory Expansion System

By carefully thought-out engineering a mother board with buffers and its

own power supply (powered by the computer's transformer) enables up to 3 8K RAM or 8K ROM boards to be fitted neatly inside the computer cabinet. Connections to the mother board from the main board expansion socket is made via a ribbon cable.

Fibre glass double sided plated through hole PCB 8.7"×3.0" set of all components including all brackets, fixing parts and ribbon cable with socket

Fibre glass double sided plated through hole PCB 5.6"×4.8" £12.50

5.6" x4.8" £12.50
Set of components including IC sockets, plug and socket but excluding RAMs £11.20
2114L RAM (16 required) £4.50
Complete set of board, components, 16 RAMS £89.50

Fibre glass double sided plated through hole PCB 5.6"×4.8" £12.40 Set of components including IC sockets, plug and socket but excluding ROMs £10.70
2708 ROM (8 required) £6.00

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€68.50

Complete set of board, components, 8 ROMs

OUR CATALOGUE IS FREE! WRITE OR PHONE NOW!

Expansion up to 32K all inside the computer's own cabinet!

PACKS For those customers who wish to spread their purchase or build a personalised system the kit is available as separate packs e.g. PCB (16"×12.5") £43.20. Pair of keyboards £34.80. Firmware in Television not included in price

ETI VOCODER COMPLETE KIT ONLY £195+VAT

Being published in Electronics Today International

COMP 80



Penel size 19.0" x 5.25". Depth 12.2"

14 CHANNELS! NOISE GENERATOR! 2 OSCILLATORS! voiced / unvoiced detector!

Kit includes FREE foot control and test oscillator!

MANY MORE KITS

ON PAGES 93, 95

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add £2.50 (VAT inclusive) per kit.

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Call at Sales Counter. Open 9 a.m.-12 noon. 1-4.30 p.m. Monday-

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TRANSCENDENT DPX

DIGITALLY CONTROLLED, TOUCH SENSITIVE, POLYPHONIC, MULTI-VOICE SYNTHESIZER

Another superb design by synthesizer expert Tim Orr — published in Electronics Today International

The Transcendent DPX is a really versatile new 5 octave keyboard instrument. There are two audio outputs which can be used simultaneously. On the first there is a beautiful harpsichord or reed sound — fully polyphonic, i.e. you can play chords with as many notes as you like. On the second output there is a wide range of different voices, still fully polyphonic. It can be a straightforward piano or a honky tonk piano or even a mixture of the two! Alternatively you can play strings over the whole range of the keyboard or brass over the whole range of the keyboard or should you prefer — strings on the top of the keyboard and brass at the lower end (the keyboard is electronically split after the first two octaves) or vice versa or even a combination of strings and brass sounds simultaneously. And on all voices you can switch in circuitry to make the keyboard touch sensitive! The harder you press down a key the louder it sounds — just like an acoustic piano. The digitally controlled multiplexed system makes practical touch sensitivity with the complex dynamics law necessary for a high degree of realism. There is a master volume and tone control, a separate control for the brass sounds and also a vibrato circuit with variable depth control together with a variable delay control so that the vibrato comes in only after waiting a short time after the note is struck for even more realistic strong sounds.



Cabinet size 36.3" × 15.0" × 5.0" (rear) 3.3" (front)

COMPLETE KIT ONLY £299 +VAT

To add interest to the sounds and make them more natural there is a chorus / ensemble unit which is a complex phasing system using CCD (charge coupled device) analogue delay lines. The overall effect of this is similar to that of several acoustic instruments playing the same piece of music. The ensemble circuitry can be switched in with either strong or mild effects.

As the system is based on digital circuitry digital data can be easily taken to and from a computer (for storing and playing back accompaniments with or without pitch or key change, computer

composing, etc., etc.)
Although the DPX is an advanced design using a very large amount of circuitry, much of it very sophisticated, the kit is mechanically extremely simple with excellent access to all the boards which interconnect with multiway connectors, just four of which are removed to separate the keyboard circuitry and the panel circuitry from the main circuitry in the cabinet. The kit includes fully finished metalwork, solid teak cabinet, professional quality components (all resistors 2% metal oxide), nuts, bolts, etc., even a 13A plug!

MANY MORE KITS ON PAGE 95. MORE KITS AND ORDERING INFORMATION ON PAGE 93.

TRANSCENDENT 2000 SINGLE BOARD SYNTHESIZER

LIVE PERFORMANCE SYNTHESIZER DESIGNED BY CONSULTANT TIM ORR (FORMERLY SYNTHESIZER DESIGNER FOR EMS LIMITED) AND FEATURED AS A CONSTRUCTIONAL ARTICLE IN ELECTRONICS TODAY INTERNATIONAL.

construction is so simple it can be built easily in a few evenings by almost anyone capable of neat soldering! When finished you will possess a synthesizer comparable in performance and quality with ready-built units selling for many times the price.

COMPLETE KIT ONLY £168.50+VAT!

Comprehensive handbook supplied with all complete kits! This fully



Cabinet size 24.6"×15.7"×4.8" (rear) 3.4" (front)



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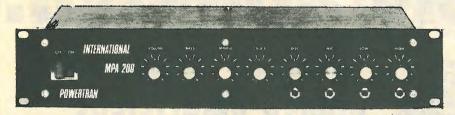
The Advertising Standards Authority. If an advertisement is wrong, we're here to put it right. A.S.A. Ltd., Brook House, Torrington Place, London WC1E 7HN.

WIRELESS WORLD, OCTOBER 1980

MPA 200 100 WATT (rms into 8Ω) MIXER/AMPLIFIER

Featured as a constructional article in ETI, the MPA 200 is an exceptionally low priced — but professionally finished — general purpose high power amplifier. It features an adaptable input mixer which accepts a wide range of sources such as a microphone, guitar, etc. There are wide range tone controls and a master volume control. Mechanically the MPA 200 is simplicity itself with minimal wiring needed making construction very straightforward.

The kit includes fully finished metalwork, fibreglass PCBs, controls, wire, etc. — complete down to the last nut and bolt.



Panel size 19.0" × 3.5". Depth 7.3"

COMPLETE KIT ONLY

£49.90 + VAT!

MATCHES THE **CHROMATHEQUE 5000** PERFECTLY!

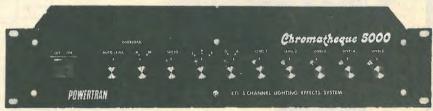
CHROMATHEQUE 5000

5 CHANNEL LIGHTING

This versatile system featured as a constructional article in ELECTRONICS TODAY INTERNATIONAL has 5 frequency channels with individual level controls on each channel. Control of the lights is comprehensive to say the least. You can run the unit as a straightforward sound-to-light or have it strobe all the lights at a speed dependent upon music level or front panel control or use the internal digital circuitry which produces some superb random and sequencing effects. Each channel handles up to 500W and as the kit is a single board design wiring is minimal and construction very straightforward.

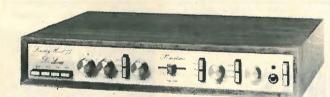
Kit includes fully finished metalwork, fibreglass PCB controls, wire, etc. - Complete right down to the last nut and bolt!

COMPLETE KIT ONLY £49.50 + VAT!



Panel size 19.0" × 3.5". Depth 7.3"

SYNTHESIZER KITS ON PAGE 93. MORE KITS AND ORDERING **INFORMATION ON PAGE 91.**



DE LUXE EASY TO BUILD LINSLEY HOOD 75W STEREO AMPLIFIER £99.30 + VAT

circuit boards interconnected with gold plated contacts resulting in minimal wiring and construction delightfully straightforward. The design was published in Hi-Fi News and Record Review and features include rumble filter, variable scratch filter, versatile tone controls and tape monitoring while distortion is less than 0.01%.

All kits also available as separaté packs (e.g. PCB, component sets, hardware set etc.). Prices in our FREE CATALOGUE.

T20 + 20 20W STEREO AMPLIFIER £33.10 + VAT

This kit, based upon a design published in Practical Wireless, uses a single printed circuit board and offers at very low cost, ease of construction and all the normal facilities found on quality amplifiers. A 30 watt version of this kit (730+30) is also available for £38.40+VAT.

MATCHING TUNERS — See our FREE CATALOGUE!

Above 2 kits are supplied with fully finished metalwork, ready assembled high quality teak veneer cabinet, cable, nuts, bolts, etc. and full instructions — in fact everything!

MUSIC EFFECTS DEVICE - AS FEATURED IN ELECTRONICS TODAY INTERNATIONAL!

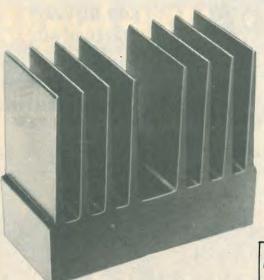
The BLACK HOLE designed by Tim Orr, is a powerful new musical effects device for processing both natural and electronic instruments, offering genuine VIBRATO (pitch modulation) and a CHORUS mode which gives a "spacey" feel to the sound achieved by delaying the input signal and mixing it back with the original. Notches (HOLES), introduced in the frequency response, move up and down as the time delay is modulated by the chorus sweep generator. An optional double chorus mode allows exciting antiphase effects to be added. The device is floor standing with foot switch controls, LED effect selection indicators, has variable sensitivity, has high signal/noise ratio obtained by an audio compander and is mains powered — no batteries to change! Like all our kits everything is provided including a highly superior, rugged steel, beautifully finished enclosure.

COMPLETE KIT ONLY £49.80 +VAT (single delay line system)

De Luxe version (dual delay line system) also available for £59.80 + VAT

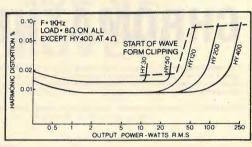
Cabinet size 10.0" x 8.5" x 2.5" (rear) 1.8" (front)





POWER AMPLIFIERS

encapsulated within heatsinks designed to total heat dissipation needs. They are rugged and made to last a lifetime. Advanced circuitry ensures their suitability for use with the finest loudspeakers, pick ups, tuners, etc. using digital or analogue sound sources.



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HY30	15 W into 8 Ω	0.02%	100 dB	-20 -0- +20	105×50×25	155	£6.34 + 95p
HY50	30 W- into 8 Ω	0.02%	100 dB	-25 -0- +25	105×50×25	155	£7.24 +£1 09
HY120	60 W into 8 Ω	0.01%	100 dB	-35 -0- +35	114×50×85	575	£15.20 + £2 28
HY200	120 W into 8 Ω	0.01%	100 dB	-45 -0· +45	114×50×85	575	£18.44 + £2.77
HY400	240 W into 4 Ω	0.01%	100 dB	-45 -0- +45	114×100×85	1.15Kg	£27.68 + £4 15

ILP PRE-AMPS ARE COMPATIBLE WITH ALL ILP POWER AMPS AND PSUS

Load impedance - all models 4 Ω - ∞ Input sensitivity - all models 500 mV Input impedance - all models 100K Ω

Frequency response - all models 10Hz - 45 KHz - 3dB

POWER SUPPLY UNITS



ILP Power Supply Units with transformers made in our own factory are designed specifically for use with ILP power amplifiers and apart from PSU 30 and 36 which are smaller PSUs - all the other ILP's own manufactured toroidal transformers are used which are half the size and weight of laminated equivalents. They are also more efficient and have greatly reduced fields of radiation.

PSU 30 ± 15V at 100mA to drive up to 12 x HY6 or 6 £4.50 + £0.68 VAT x HY66 THE FOLLOWING WILL ALSO DRIVE ILP PRE-AMPS £8.10+£1.22 VAT PSU 36 for 1 or 2 HY 30's

The following include toroidal transformers

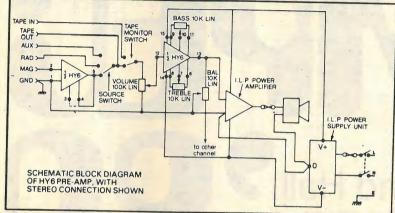
PSU 50 for 1 or 2 HY 50's £9.75 + £1.46 VAT PSU 60 for 1 HY 120 £9.75 + £1 46 VAT PSU 70 for 1 or 2 HY 120s £13.61 + £2.04 VAT

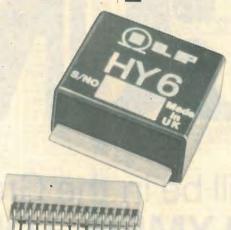
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Sizes - HY6 - 45 x 20 x 40 mm HY66 90 x 20 x 40 mm Active Tone Control circuits provide ± 12dB cut and boost. Inputs Sensitivity - Mag. PU - 3mV: Mic - selectable 1-12mV: All others 100mV: Tape O/P - 100mV Main O/P - 500mV: Frequency response - D.C to 100KHz - 3dB,



HY6 mono £10.60

WIRELESS WORLD, OCTOBER 1980

+VAT £1.59

Connectors included

B6 Mounting Board **78**p + 12p VAT **B66** Mounting Board **99p** + 15p VAT LOW DISTORTION - Typically 0.005%

S/N RATIO - Typically 90 dB (Mag. P.U. - 68 dB)

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ppe 1000 1001 1001 1002 1004 1004 1006 1007 1008 1009 1009 1001 111 111 111 111 111 111	price £0.10 £0.13 £0.13 £0.13 £0.13 £0.13 £0.13 £0.25 £0.25 £0.15 £0.15 £0.17 £0.20 £0.17 £0.28 £0.28 £0.28 £0.28 £0.28 £0.28 £0.28 £0.28 £0.28 £0.28 £0.28 £0.28 £0.28	Type 7427 7428 7430 7433 7437 7438 7440 7441 7442 7443 7444 7445 7446 7470 7451 7450 7451 7450 7451 7450 7451 7450 7450 7450 7450 7450 7450 7450 7450	Price £0.28 £0.30 £0.35 £0.25 £0.25 £0.24 £0.14 £0.58 £0.46 £0.81 £0.81 £0.81 £0.81 £0.81 £0.13 £0.13 £0.13 £0.13 £0.13 £0.13	Type 7472 7473 7474 7475 7476 7480 7481 7482 7483 7484 7485 7489 7490 7491 7490 7491 7490 7491	Price £0.23 £0.29 £0.29 £0.33 £0.51 £0.58 £0.98 £0.78 £0.78 £0.78 £0.25 £1.96 £0.37 £0.40 £0.40 £0.40 £0.40 £0.40 £0.40 £0.40 £0.40	Type 74105 74107 741110 741111 74118 74119 74121 74122 74123 74136 74141 74145 74150 74151 74160 74161 74161	Price £0.44 £0.28 £0.41 £0.67 £0.67 £0.68 £0.45 £0.46 £1.00 £0.63 £0.78 £0.55 £0.55 £0.58 £0.58 £0.67 £0.67 £0.67 £0.67	741 741 741 741 741 741 741 741 741 741	Price 63 £0.71 64 £0.78 65 £0.78 66 £0.90 74 £0.75 75 £0.71 76 £0.87 77 £0.67 80 £1.73 81 £0.67 82 £0.81 80 £0.87 82 £0.81 80 £0.81 90 £0.71 92 £0.69 99 £1.21 99 £2.13	AL10 AL20 AL30A AL60 AL80 AL120 AL250 PRE-AI PA12 PA100 PA200	JFIERS 3 watt Audio Amp 5 watt Audio Amp 5 watt Audio Amp 7-10 watt Audio A supply 15-25 watt Audio 80 watt Audio Am 90 watt	lifier Module : mplifier Module plifier Module plifier Module plifier Module mplifier Module -32v input ser /AL30 -55v inputs: -1 /80 / AL1 20 / / /65v inputs: Noutput 500mw	22-32v supply 22-32v supply 22-32v supply dule 30-50v 40-60v supply 60-70v supply le 50-80v supply rape, Tuner, Tape, Tuner, AL 250 Aag, P. U., Tape 65v inputs: 2 Guitar	£3.63 £4.11 £4.78 £5.92 £9.28 £15.11 £22.54 £9.83 £20.30 £20.98	2nd GRADE LED PAK A pack of 10 standard sizes and colours wh perform to their very rigid specification but ideal for amateurs who do not require the fu O/NO: 150 MAMMOTH I.C. PA 16223 — Approx. 200 pieces assorted integrated circuits including Logic 74 seri Audio and DTL. Many coded devices b unmarked you to identify SOCKETS 1611 8 pin DIL 1612 14 pin DIL 1612 14 pin DIL 1720 18 pin DIL 1721 20 pin DIL 1722 22 pin DIL 1721 24 pin DIL 1614 24 pin DIL 1615 28 pin DIL 1615 28 pin DIL 1616 28 pin DIL 1723 40 pin DIL 1616 10 18 transistor 1617 103 transistor 1726 14 pin DIL Wire wrap gold plated Can
04001 04006 04006 04008 04008 04009 04011 04011	Price 0 £0.16 £0.23 £0.18 £1.06 £0.20 £1.06 £0.52 £0.55 £0.23 £0.22 £0.48	CD401 CD401 CD401 CD401 CD402	0 £1.04 1 £0.94 2 £0.94 3 £0.22 4 £0.75 5 £0.22	Type CD4026 CD4027 CD4028 CD4039 CD4031 CD4035 CD4037 CD4040 CD4041 CD4042	£0.58 £0.78 £0.98 £0.55 £2.30 £1.38 £1.09 £1.01 £0.87 £0.83	Type CD4043 CD4044 CD4045 CD4046 CD4047 CD4050 CD4050 CD4056 CD4056 CD4069	£0.94 £1.61 £1.50 £1.00 £0.52 £0.52	CD40 CD40 CD40 CD45 CD45 CD45 CD45 CD45	Price 70 £0.20 72 £0.20 72 £0.20 81 £0.20 82 £0.25 10 £1.27 11 £1.44 16 £1.15 18 £1.15 14 £0.92	PS12 SPM80 SPM120 SPM120 SPM120 SG3Ò	PA100 to 25 watts 0/55 PA200 FA200 FA200, 1×AL250, 15-0-15 Stabilised 2×GE100MKII LLANEOUS Stereo Magnetic Ca input 3.5mv Output Stereo FM Tuner SL	x AL10, 2×AL S, 450 S, 450 Stabilised supp Stabilised supp Stabilised supp PA200 power supply	.20, AL60, bly — suit 2×AL60, bly — suit 2×AL80, bly — suit 2×AL120, for mplifier- 20-30v —	£1.90 £5.57 £7.34 £7.34 £7.34 £4.37	G.P. SWITCHING TRANSISTORS 1018 sim to 2N706 /8 BSY27 /28 / 95A ALL usable devices no open & shorts. ALSO available in PIPP sim. to 2N290 /BCY7 20 for 57 - 50 for £1.15 = 100 for £2.07 or £9.20 - 50 for £1.15 = 100 for £2.07 or £9.20 - 500 for £16.10 - when of state NPN /PNP. SILICON DIODES G. 300mW 40PIV (min) sub min FULLY TESTE for Organ builders. 30 for £72 - 100 for £1.72 - 500 for £1.
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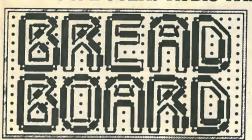
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COMPUTERS AUDIO RADIO MUSIC LOGIC TEST GEAR CB GAMES KITS





26th Nov - WEDNESDAY - 10am-6pm 27th Nov - THURSDAY - 10am-8pm

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COMPONENTS DEMONSTRATIONS SPECIAL OFFERS MAGAZINES BOOKS

Royal Horticultural Halls Elverton Street

Westminster London SW1 November 26-30 1980 It's all at Breadboard '80

This is **the** exhibition for the electronics enthusiast. From November 26 -30 there is only one place in the universe for the electronics enthusiast to be — Breadboard '80, at the Royal Horticultural Hall in London. The majority of leading companies will be exhibiting, including all the top monthly magazines in the field. There will be demonstrations on most stands and many feature special offers that are EXCLUSIVE to Breadboard!

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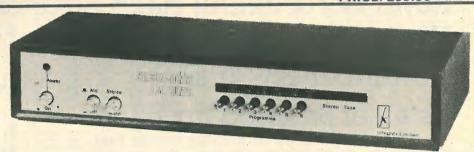


Brief Spec. Amplifier Low field Toroidal transformer, Mag. input. Tape In / Out facility (for noise reduction unit, etc.) THD less than 0-1% at 20W into 8 ohms. High Slew Rate. Low noise op. amps used throughout. Power on / off FET transient protection. All sockets, fuses, etc., are PC mounted for ease of assembly. Tuner section uses UM 1181 FET module requiring no RF alignment, ceramic IF, INTERSTATION MUTE, and phase-locked IC pilot cancel, stereo decoder, LED tuning and stereo indicators. Tuning range 88-108MHz 30dB mono S/N @ 0.7 µV. THD 0.3%. PRICE: £69.95 + VAT

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A very high performance tuner with dual gate MOSFET RF and Mixer ready built front end, triple gang varicap tuning, linear phase I.F. and 3 state MPX de-

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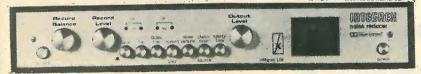
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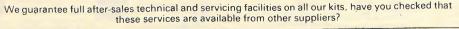
Calibration tapes are available for open-reel use and for cassette (specify which)

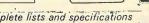
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Dynamic range >90dB





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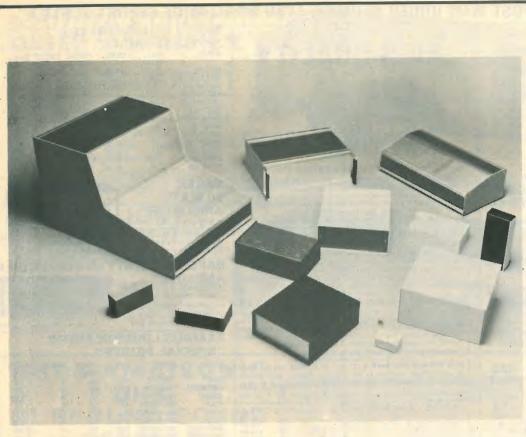
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0-8-9, 0-8-9
0-15, 0-15
0-20, 0-20
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	the Control of the Control		-	10000	2,000	30, 30
	12 OR	24V OR 1: Amps	2-0-12V	9	*	Ref. 124 126
Ref. 111 213 71 18 85 70 108 72	12V 0.5 1.0 2 4 0.5 6 8	24V 0.25 0.5 1 2 2.5 3 4 5	Price 2.42 2.90 3.86 4.46 6.16 6.99 8.16 8.93	P&P 0.52 0.90 0.90 1.10 1.10 1.10 1.31 1.31	uous Ratings	127 125 123 40 120 121 122 189
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2	0. 24. 30V or	12V-0-12V or 1	5V-0-15V	
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6.5 mid. 280V AC
6.5 mid. 280V AC
1.0 mid. 250V AC
1.0 mid. 250V AC
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X011		9+9	4.44			
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X013		15+15	2.66			EACH
X014		18+18	2.22			£5.76
X015		22+22;	1.81			LJ./U
X016		25+25	1.60	,		+£1.20 P&P
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X030		240	0.33			
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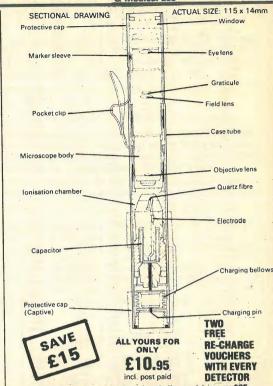
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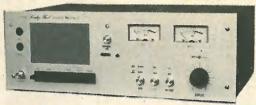
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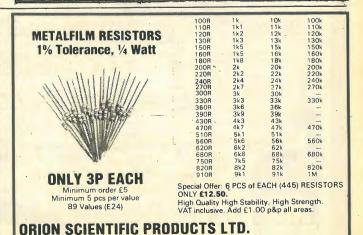
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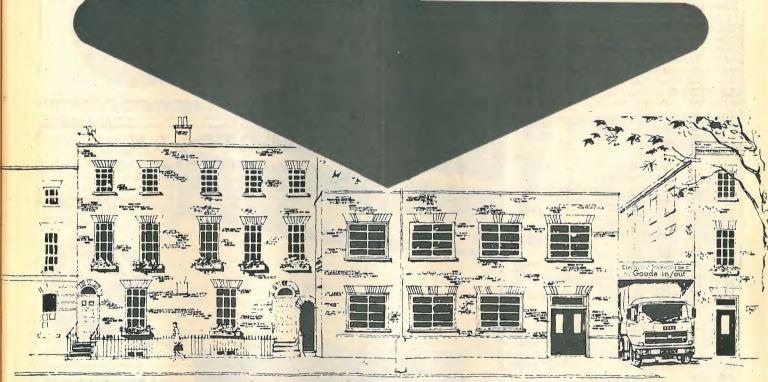
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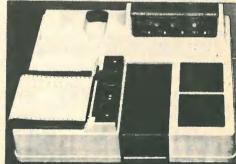
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Hazeltine H-1000

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Specification SCREEN SIZE - 12" diagonal.

SCREEN CAPACITY - 960 characters; 80

CHARACTERS - 5 x 7 Dot Matrix; 625-line

CHARACTER SET - 64 ASCII alphanumerics and symbols.

KEYBOARD - TTY format

INDICATORS - Power On. Parity Error. PARITY - Parity error indicated by Parity Light and question mark (?) displayed in cha-

TRANSMISSION - Asynchronous. Switchselectable for any two standard rates up to

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REFRESH RATE - 50 fields per second.

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Specification
SCREEN — 12" diagonal. 1998 characters: 74 per

CHARACTERS — 5 x 7 Dot Matrix; 625 lines raster CHARACTER SET — 64 alphanumerics and symbols. 32 ASCII control codes.

KEYBOARD - Detachable, solid state. TTY design 10-key numeric cluster plus editing and cursor control keys.

TRANSMISSION — Asychronous.

Switch-selectable, for combinations of 5 standard rates, 110 to 9600 baud.

OPERATING MODES — Switch-selectable, full duplex, half-duplex or batch.

MEMORY TYPE — 2048 x 8 RAM.

FDITING FEATURES - Full Cursor Controls plus Insert/Delete Character, Insert/Delete Line, Clear Screen, Clear Foreground Data Only, Tab. STANDARD INTERFACE - CC ITT V-24 (EIA

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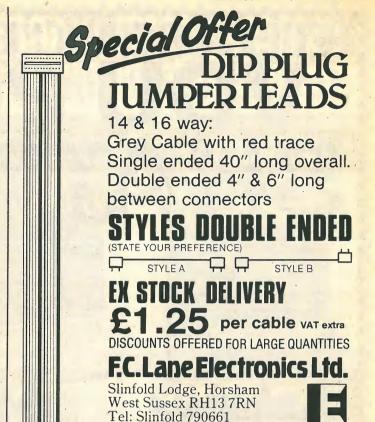
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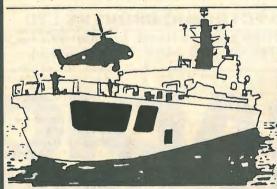
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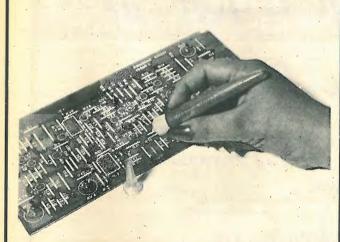
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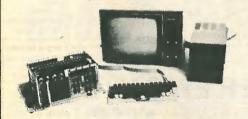
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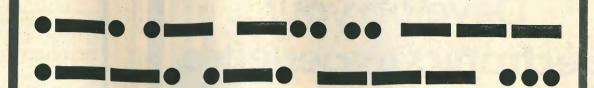
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INDEX TO ADVERTISERS OCTOBER

Appointments Vacant Advertisements appear on pages 131-143

PAGE	PAGE	FAGE
Acoustical Mfg	GEC M-O-Valve	OMB Electronics
A.E.L. Crystals	GMT Electronics	Orion
Ambit International	Guide to Broadcasting Stations	Pascall
Antex 63		P.B.R.A. Ltd. 25
Apex 14	Hameg	P.M. Components 104
Airamco 26	Hall Electric Ltd 2	Powertran Electronics 91, 93, 95
Amateur Radio Assoc 23	Happy Memories 94	Practical Computing
ASA 94	Harris Electronics (London) Ltd	Tractical Computing
Aspen Electronics Ltd 32	Harrison Brothers 92	Quantum Electronics 8
Audio Elec	Hart Electronics	
Austerfield Clark 105	Hi-Fi Y/Book 106	Radio Components Specialists 103
	Hilomast Ltd	Ralfe, P. F
Bach-Simpson 75	House of Instruments 24	R.C.S. Electronics
Bamber, B. Electronics 107		Rediffusion Resitronics
Bang & Olufsen	I.L.P. Electronics Ltd	Royal Navy
Barnet Metal	ILP Transformers Ltd 110	Roxborough
Barrie Electronics Ltd 105	Integrex Ltd 102	RST Valves 100
Bell & Howell 90	Interface Comps	NOT Valves 100
BIB Hi-Fi Cover iv	Jackson Music	
Bi-Pak Semiconductors Ltd	Jackson Music 116	Sagin (Circuit Services) 108
Breadboard 80	** *** * * * *	Samsons (Electronics) Ltd 114
British National Radio	Keithley Insts	Scopex Instruments Ltd
. C	Kelsey Acoustics	Service Trading
Carston Electronics Ltd 10, 11	KGM Electronics	Shure Electronics
Catronics	Kirkham Amplifier 89	Sinclair Radionics
Chiltmead Ltd	Langrex 100 -	Sonic Sound Audio 107
Codespeed Elec	Lascar Electronics 22	Special Products Ltd
Colomor	Levell Electronics Ltd 3	Standard Pneumatic
Computer Appreciation	Lowe Electronics Ltd 6	Strumech Eng'g
Consumers Assoc. Loose insert	Maclin-Zand Elec. Ltd	Strutt Electrical & MSH Ltd 92
Continental Specialities	Mascot Elec	Surrey Electronics Ltd
Crimson Elektrik	Maplin Electronic Supplies Cover iii	Swanley Electronics Ltd
CHIRSOII Elektrik	Marshall, A. & Sons (London) Ltd 108	
Davies, A. R	MCP Electronics	Technomatic
Display Electronics	Microcircuits Ltd	Teleradio Elec. 23
Dondene	Microtime	Telefadio Elec.
Dondene 112	Mills. W	Valradio Ltd
Edicron	Milward, G. F	Vero Speed
Electronic Brokers Ltd	Minim Audio	Vero Systems Ltd
23, 25, 117, 118, 119, 120, 121, 125, 127, 129, 144	MTL Microtesting	VHS Committee
Electro-Tech Comps Ltd	Mullard	
Eraser Int'l	Multicore Solders Ltd Cover iv	West Hyde Developments Ltd
Faircrest Eng	Mura Electronics	Wilmslow Audio
Farnell Instruments Ltd Cover ii, 64 Readers card	MWM Co	
Field Tech	Newtronics	Zaerix Elec. 104
12		

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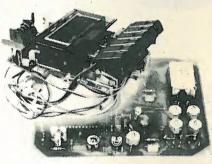
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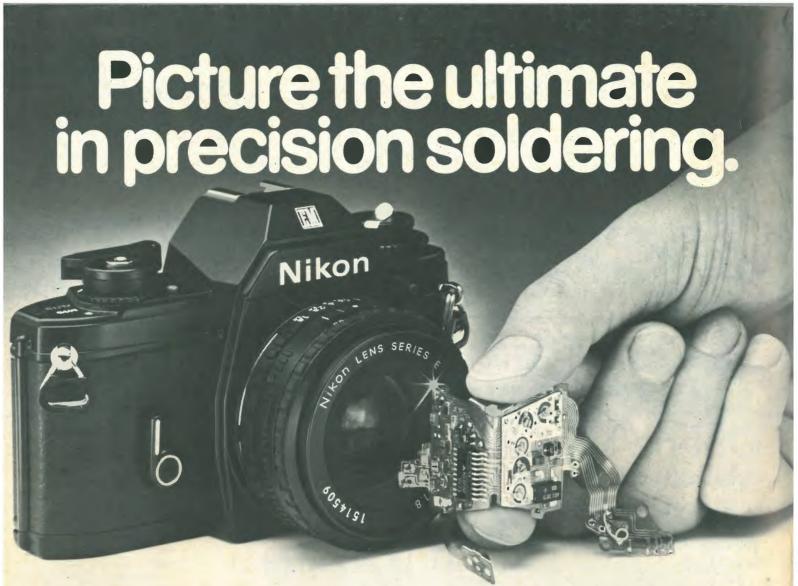
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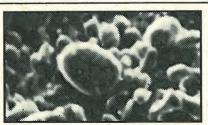
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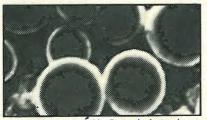


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