

THE JOURNAL FOR PROFESSIONAL ENGINEERS

ELECTRONICS & WIRELESS WORLD

JANUARY 1987

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**P.c.b.
connector
review**

**ATE – an
introduction**

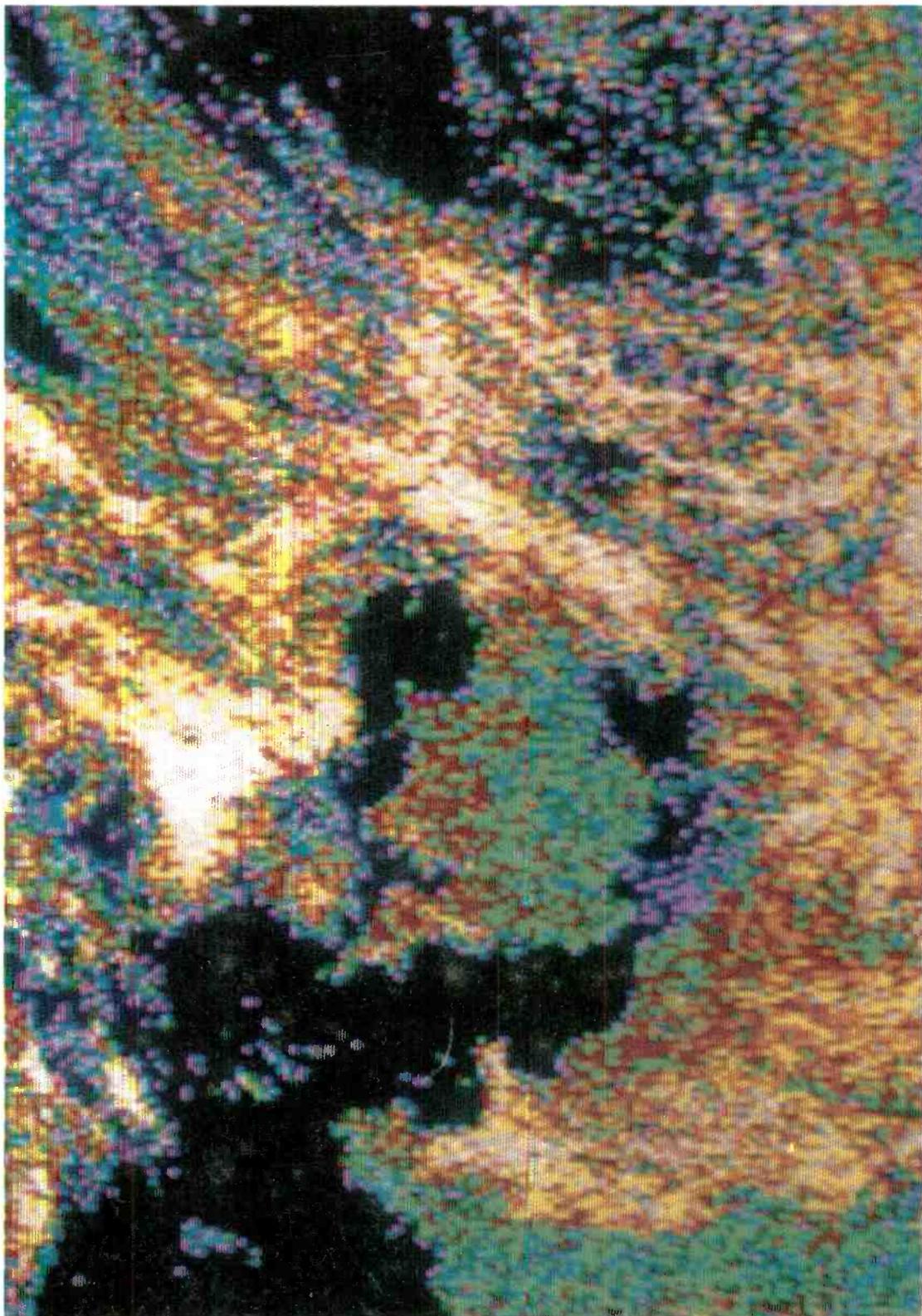
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standards
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CONTENTS

JANUARY VOLUME 92 NUMBER 1611

FEATURES

COVER STORY

Part of a picture received from NOAA-9, displayed on Don Clarke's video digitizer

CONNECTOR REVOLUTION

9

Hierarchical interconnection technology could be the basis of future techniques
by John Curry

DESIGNING A SURFACE-MOUNT CONNECTOR

14

Design of a low-profile but stable connector
by Adrian Hyner

MODELLING THE COCHLEA

19

Electrical analogue of the ear's hearing organ
by R.W. Guelke, J.P. Ramackers and A.E. Bunn

VIDEO FRAME STORE – 3

23

Design of the control and memory circuits
by D.E.A. Clarke

THE MAXIMUM POWER THEOREM

41

JW advises against too literal an interpretation
by Joules Watt

DOUBLE-NOTCH FILTERING SIMPLIFIES DISTORTION ANALYSIS

47

Canyon-shaped notch eases filter tuning
by B.J. Sokol

RANDOM-ACCESS PHRASE RECORDING

54

High-quality compressed speech data is easily recorded and quickly replayed
by A.L. Evans and J. Fenner



Unshrouded headers and shorting links for p.c.b. connection from Winslow Component Systems illustrates this month's feature on connectors, starting page 9.

VERSATILE OPERATIONAL AMPLIFIERS

58

Discrete-component designs provide up to 500Vpk, fast slewing and wide-band operation
by J.Barron

ELECTROMETER AMPLIFIERS FOR SUB-PICOAMP CURRENTS

65

Stitched-range and logarithmic amplifiers for current down to 20 femtoamps
by D.F.Caudrey

TWO NOVEL OSCILLATORS

67

Minimum-component op-amp designs
by M.T. Abuelma'atti and W.A. Almansoury

AUTOMATIC TESTING

69

An overview of techniques used in the automatic test of printed-board assemblies
by Graham Elliott

TELEVISION STANDARDS CONVERSION

75

Thirty years of development, from image transfer to digital methods and onwards
by S.M.Edwardson and C.K.P. Clarke

MOBILE RADIO IN THE LANCASHIRE POLICE

96

Development of a v.h.f. mobile system
by J. Davies

PIONEERS – 1. STEPHEN GRAY

100

The man discovered electrical conduction
by W. A. Atherton

WITHIN THE 68020

103

The architecture of this 32-bit processor with dynamic bus sizing
by D. Burns and D. Jones

DIGITAL COMPUTER MODELLING OF LINEAR SYSTEMS

107

Using analogue techniques to solve second-order differential equations with a microcomputer
by Richard Oxley

REGULARS

COMMENT 3

BOOKS 30

SATELLITE SYSTEMS 31

TELECOMMS TOPICS 33

CIRCUIT IDEAS 35

RESEARCH NOTES 37

APPLICATION NOTES 43

NEWS 52

FEEDBACK 62

NEW PRODUCTS 82-87

RADIO BROADCAST 89

TELEVISION BROADCAST 91

RADIO COMMUNICATIONS 93

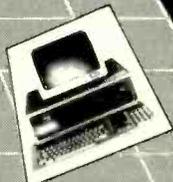
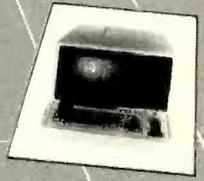
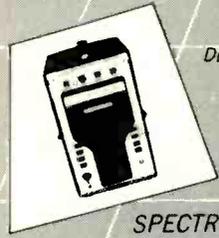
WORKFILE 115

ADVERTISEMENT INDEX 120

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SPECTRAL LINES

Frequency planning, which is now the responsibility of the Radio Regulatory Division of the DTI, is on the point of being privatized. A report commissioned by the government from CSP International Ltd recommends that the RRD should relinquish its activities in this field in favour of a number of Frequency Planning Organizations who would, in return for a fee, license areas of the spectrum to individual facilities.

Since the FPOs will be profit-making bodies, it seems likely that their decisions on the merits of license applications will not be totally unconnected with the size of the fees they can extract from the licensees, so that social considerations may possibly receive less than due weight.

In the years since it first became necessary to control the use of the radio spectrum, the Home Office and RRD have been targets for complaints of despotism, fascism, inefficiency, pig-headedness and probably illegitimacy from applicants for licences to operate services; nevertheless, the control of spectrum use has been in the hands of one body with overall responsibility which was not, at least, in it for the money. On the face of it, therefore, the new proposals do seem to be in line with the modern right-wing approach to government, which means that, to some eyes, they are automatically suspect.

And so they may be, but there could be a saving grace, in that if someone is trying to make a profit from allocating the spectrum the chances are that efficiency in its use will increase markedly. An under-used allocation, if someone has to pay a high price for it, will not be supportable and will surely be relinquished to some other user who needs it badly enough to pay for it. Technically speaking, it is probable that pricing the spectrum will be a powerful incentive to the development of equipment which will be a little less profligate in its bandwidth requirements. Whether this is the best way to share out the spectrum is a matter for debate, but it should at least avoid large, sparsely populated bands of frequencies being denied to other users.

The report claims that the predicted "explosion" in radio communications in the UK - from 400,000 to 1,000,000 mobile radios by 1995 is one of its forecasts - can be accommodated in less than 40% more of the spectrum, which can be obtained from reallocating existing frequency blocks to "commercially most productive forms of usage".

It is to be hoped that the RRD will not abandon its role completely, since the thought of a bunch of commercially motivated FPOs fighting each other for the use of the radio spectrum, perhaps without sufficient experience or expertise to handle the allocations responsibly, is not an attractive one.

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Allows an easy method to reconfigure pin functions without rewiring the cable assay. Jumpers can be used and reused. £22 (d)

Serial Mini Test

Monitors RS232C and CCITT V24 Transmissions, indicating status with dual colour LEDs on 7 most significant lines. Connects in Line £22.50 (d)

CONNECTOR SYSTEMS

I.D. CONNECTORS

No of ways	Speedblock Type		Edge Conn.
	Header	Pin	
10	90p	85p	120p
20	145p	125p	195p
26	175p	150p	240p
34	200p	160p	320p
40	220p	190p	340p
50	235p	200p	390p

D CONNECTORS

	No of Ways			
	9	15	25	37
MALE:				
Ang Pins	120	180	230	350
Solder	60	85	125	170
IDC	175	275	325	-
FEMALE:				
St Pin	100	140	210	380
Ang Pins	160	210	275	440
Solder	90	130	195	290
IDC	195	325	375	-
St Hood	90	95	100	120
Screw	130	150	175	-
Lock	-	-	-	-

TEXTTOOL ZIF

SOCKETS	24-pin £7.50
28-pin	£9.10
	40-pin £12.10

EDGE CONNECTORS

	0.1"	0.156"
2 x 6-way (commodore)	150p	300p
2 x 10-way	150p	-
2 x 12-way (vic 20)	-	350p
2 x 18-way	-	140p
2 x 23-way (ZX81)	175p	220p
2 x 25-way	225p	220p
2 x 28-way (Spectrum)	200p	-
2 x 36-way	250p	-
1 x 43-way	260p	-
2 x 22-way	190p	-
2 x 43-way	395p	-
1 x 77-way	400p	500p
2 x 50-way (S100conn)	600p	-

EURO CONNECTORS

	Plug	Skt
DIN 41612	230p	275p
2 x 32 way St Pin	275p	320p
2 x 32 way Ang Pin	260p	300p
3 x 32 way St Pin	375p	400p
3 x 32 way Ang Pin	400p	400p
IDC Skt A + B	400p	-
IDC Skt A + C	400p	-

For 2 x 32 way please specify spacing (A + B, A + C).

MISC CONNS

21 pin Scart Connector	200p
8 pin Video Connector	200p

AMPHENOL CONNECTORS

36 way plug Centronics (solder) 500p (IDC) 475p
36 way skt Centronics (solder) 550p (IDC) 500p
24 way plug IEEE (solder) 475p (IDC) 475p
24 way skt IEEE (solder) 500p (IDC) 500p
PCB Mtg Skt Ang Pin
24 way 700p 36 way 750p

GENDER CHANGERS

25 way D type	
Male to Male	£10
Male to Female	£10
Female to Female	£10

RS 232 JUMPERS

(25 way D)	
24" Single end Male	£5.00
24" Single end Female	£5.25
24" Female Female	£10.00
24" Male Male	£9.50
24" Male Female	£9.50

DIL SWITCHES

4-way	90p	6-way	105p
8-way	120p	10-way	150p

RIBBON CABLE

(grey/metre)			
10-way	40p	34-way	160p
16-way	60p	40-way	180p
20-way	85p	50-way	200p
26-way	120p	64-way	280p

DIL HEADERS

	Solder	IDC
14 pin	40p	100p
16 pin	50p	110p
18 pin	60p	-
20 pin	75p	-
24 pin	100p	150p

74 SERIES 74C SERIES LINEAR ICs COMPUTER COMPONENTS

7400	0.30	74273	2.00	74LS273	1.25
7401	0.30	74276	1.40	74LS279	0.70
7402	0.30	74278	1.70	74LS280	1.90
7403	0.30	74279	0.90	74LS283	0.80
7404	0.30	74283	1.05	74LS290	0.80
7405	0.30	74284	3.20	74LS293	0.80
7406	0.30	74293	0.90	74LS295	1.40
7407	0.40	74298	1.80	74LS299	1.40
7408	0.30	74351	2.00	74LS298	1.00
7409	0.30	74356A	0.80	74LS299	2.20
7410	0.30	74357A	0.80	74LS321	3.70
7411	0.30	74376	1.60	74LS323	3.00
7412	0.30	74390	1.10	74LS324	3.20
7413	0.50	74393	1.20	74LS348	2.00
7414	0.70	74490	1.40	74LS352	1.20
7416	0.36			74LS353	1.20
7417	0.40			74LS356	2.10
7420	0.30			74LS363	1.80
7421	0.60			74LS364	1.80
7422	0.36			74LS365	0.50
7423	0.36	74LS00	0.24	74LS366	0.50
7425	0.40	74LS01	0.24	74LS367	0.52
7426	0.40	74LS02	0.24	74LS368	0.50
7427	0.40	74LS03	0.24	74LS373	0.70
7428	0.43	74LS04	0.24	74LS377	0.70
7430	0.30	74LS05	0.24	74LS375	0.75
7432	0.36	74LS08	0.24	74LS377	1.30
7433	0.30	74LS09	0.24	74LS378	0.95
7437	0.30	74LS10	0.24	74LS379	1.30
7438	0.40	74LS11	0.24	74LS381	4.50
7439	0.40	74LS12	0.24	74LS385	3.25
7440	0.40	74LS14	0.50	74LS390	6.00
7441	0.90	74LS15	0.24	74LS393	1.00
7442A	0.70	74LS20	0.24	74LS395A	1.00
7443A	1.00	74LS21	0.24	74LS399	1.40
7444	1.10	74LS22	0.24	74LS445	1.80
7445	0.70	74LS23	0.24	74LS445	1.20
7446A	1.00	74LS26	0.26	74LS46	1.20
7447A	1.00	74LS27	0.24	74LS490	1.50
7448	1.00	74LS28	0.24	74LS540	1.00
7450	0.36	74LS30	0.24	74LS541	1.00
7451	0.35	74LS32	0.24	74LS608	7.00
7453	0.38	74LS32	0.24	74LS610	25.00
7454	0.38	74LS33	0.24	74LS641	1.00
7460	0.55	74LS37	0.24	74LS624	3.50
7470	0.50	74LS38	0.24	74LS626	2.25
7472	0.45	74LS40	0.24	74LS628	2.25
7473	0.45	74LS42	0.50	74LS629	1.25
7474	0.50	74LS43	0.50	74LS640	2.00
7475	0.50	74LS45	0.90	74LS641	3.00
7476	0.45	74LS49	1.00	74LS641	1.00
7480	0.65	74LS51	0.24	74LS642	2.50
7481	1.80	74LS54	0.24	74LS642	1.30
7483A	1.05	74LS55	0.24	74LS643	2.50
7484A	1.25	74LS57A	0.30	74LS643	1.30
7485	1.10	74LS74A	0.35	74LS644	3.50
7486	0.42	74LS74	0.48	74LS645	2.00
7489	2.10	74LS76A	0.36	74LS645	1.40
7490A	0.55	74LS78	0.42	74LS668	0.90
7491	0.70	74LS83A	0.70	74LS669	0.90
7492A	0.70	74LS85	0.75	74LS670	1.70
7493A	0.55	74LS86	0.75	74LS672	2.50
7494	1.10	74LS89	0.38	74LS673	2.48
7495A	0.60	74LS91	0.90	74LS678	3.50
7496	0.80	74LS92	0.35	74LS677	3.50
7497	2.90	74LS93	0.54	74LS688	3.50
74100	1.90	74LS95B	0.75	74LS783	16.00
74107	0.50	74LS96	0.90		
74109	0.75	74LS109	0.40		
74110	0.75	74LS109	0.40		
74111	0.55	74LS112	0.45		
74116	1.70	74LS113	0.45		
74118	1.10	74LS114	0.45		
74119	1.70	74LS122	0.70		
74120	1.60	74LS123	0.80		
74121	0.55	74LS126	0.50		
74122	0.70	74LS126	0.50		
74123	0.80	74LS132	0.65		
74125	0.65	74LS133	0.55		
74126	0.55	74LS136	0.45		
74128	0.55	74LS138	0.55		
74132	0.75	74LS135	0.75		
74136	0.70	74LS145	0.95		
74141	0.90	74LS147	1.75		
74142	2.50	74LS148	1.40		
74143	1.30	74LS151	0.65		
74144	2.70	74LS152	2.00		
74145	1.10	74LS153	0.65		
74147	1.70	74LS154	1.60		
74148	1.40	74LS155	0.65		
74150	1.75	74LS156	0.65		
74151A	0.70	74LS157	0.50		
74153	0.80	74LS158	0.65		
74154	1.40	74LS160A	0.85		
74155	0.80	74LS161A	0.75		
74156	0.90	74LS162A	0.75		
74157	0.80	74LS163A	0.75		
74159	2.25	74LS164	0.75		
74160	1.10	74LS165A	1.10		
74161	0.80	74LS166A	1.50		
74162	1.10	74LS168	1.30		
74163	1.10	74LS169	1.00		
74164	1.20	74LS170	1.40		
74165	1.10	74LS173A	1.00		
74166	1.40	74LS174	0.75		
74167	4.00	74LS175	0.75		
74170	2.00	74LS181	2.00		
74172	4.20	74LS183	1.90		
74173	1.40	74LS190	0.75		
74174	1.10	74LS191	0.75		
74175	1.05	74LS192	0.80		
74176	1.00	74LS194A	0.75		
74178	1.50	74LS195A	0.75		
74179	1.50	74LS196	0.80		
74180	1.00	74LS197	0.80		
74181	3.40	74LS221	0.80		
74182	1.40	74LS220	0.80		
74184	1.80	74LS241	0.80		
74185A	1.80	74LS242	0.90		
74190	3.00	74LS243	0.90		
74191	3.00	74LS244	0.90		
74192	1.10	74LS245	0.90		
74193	1.15	74LS247	1.10		
74194	1.10	74LS248	1.10		
74195	0.80	74LS249	1.10		
74196	1.30	74LS251	1.30		
74197	1.10	74LS253	0.75		
74198	2.20	74LS256	0.90		
74199	2.20	74LS257A	0.70		
74221	1.10	74LS258A	0.70		
74251	1.00	74LS269	1.20		
74252	1.50	74LS270	1.50		
74265	0.80	74LS266	0.60		

74C00	0.70	74C04	0.50	74C10	0.70
74C08	0.50	74C14	0.50	74C20	0.70
74C16	0.50	74C20	0.70	74C32	1.00
74C24	1.50	74C32	1.00	74C48	1.50
74C32	1.00	74C48	1.50	74C64	1.50
74C72	1.20	74C76	1.00	74C80	1.50
74C80	1.50	74C84	2.00	74C88	2.00
74C92	2.00	74C96	1.50	74C100	1.50
74C96	1.50	74C104	1.50	74C108	1.50
74C112	1.50	74C116	1.50	74C120	1.50
74C124	1.50	74C128	1.50	74C132	1.50
74C136	1.50	74C140	1.50	74C144	1.50
74C148	1.50	74C152	1.50	74C156	1.50
74C160	1.50	74C164	1.50	74C168	1.50
74C172	1.50	74C176	1.50	74C180	1.50
74C184	1.50	74C188	1.50	74C192	1.50
74C196	1.50	74C200	1.50	74C204	1.50
74C208	1.50	74C212	1.50	74C216	1.50
74C220	1.50	74C224	1.50	74C228	1.50
74C232	1.50	74C236	1.50	74C240	1.50
74C244	1.50	74C248	1.50	74C252	1.50
74C256	1.50	74C260	1.50	74C264	1.50
74C268	1.50	74C272	1.50	74C276	1.50
74C280	1.50	74C284	1.50	74C288	1.50
74C292	1.50	74C296	1.50	74C300	1.50
74C304	1.50	74C308	1.50	74C312	1.50
74C316	1.50	74C320	1.50	74C324	1.50
74C328	1.50	74C332	1.50	74C336	1.50
74C340	1.50	74C344	1.50	74C348	1.50
74C352	1.50	74C356	1.50	74C360	1.50
74C364	1.50	74C368	1.50	74C372	1.50
74C376	1.50	74C380	1.50	74C384	1.50
74C388	1.50	74C392	1.50	74C396	1.50
74C400	1.50	74C404	1.50	74C408	1.50
74C412	1.50	74C416	1.50	74C420	1.50
74C424	1.50	74C428	1.50	74C432	1.50
74C436	1.50	74C440	1.50	74C444	1.50
74C448	1.50	74C452	1.50	74C456	1.50
74C460	1.50	74C464	1.50	74C468	1.50
74C472	1.50	74C476	1.50	74C480	1.50
74C484	1.50	74C488	1.50	74C492	1.50
74C496	1.50	74C500	1.50	74C504	1.50
74C508	1.50	74C512	1.50	74C516	1.50
74C520	1.50	74C524	1.50	74C528	1.50
74C532	1.50	74C536	1.50	74C540	1.50
74C544	1.50	74C548	1.50	74C552	1.50
74C556	1.50	74C560	1.50	74C564	1.50
74C568	1.50	74C572	1.50	74C576	1.50
74C580	1.50	74C584	1.50	74C588	1.50
74C592	1.50	74C596	1.50	74C600	1.50
74C604	1.50	74C608	1.50	74C612	1.50
74C616	1.50	74C620	1.50	74C624	1.50
74C628	1.50	74C632	1.50	74C636	1.50
74C640	1.50	74C644	1.50	74C648	1.50
74C652	1.50	74C656	1.50	74C660	1.50
74C664	1.50	74C668	1.50	74C672	1.50
74C676	1.50	74C680	1.50	74C684	1.50
74C688	1.50	74C692	1.50	74C696	1.50
74C700	1.50	74C704	1.50	74C708	1.50
74C712	1.50	74C716	1.50	74C720	1.50
74C724	1.50	74C728	1.50	74C732	1.50
74C736	1.50	74C740	1.50	74C744	1.50
74C748	1.50	74C752	1.50	74C756	1.50
74C760	1.50	74C764	1.50	74C768	1.50
74C772	1.50	74C776	1.50	74C780	1.50
74C784	1.50	74C788	1.50	74C792	1.50
74C796	1.50	74C800	1.50	74C804	1.50
74C808	1.50	74C812	1.50	74C816	1.50
74C820	1.50	74C824	1.50	74C828	1.50
74C832	1.50	74C836	1.50	74C840	1.50
74C844	1.50	74C848	1.50	74C852	1.50
74C856	1.50	74C860	1.50	74C864	1.50
74C868	1.50	74C87			

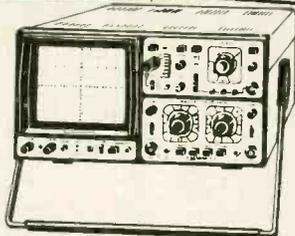
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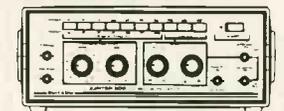
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Hameg, Crotech, Hitachi Stockists (UK TNT delivery
£7.00 plus VAT)

HAMEG Dual trace with component testers: 8 x 10cm green screen	
203/6 Dual 20 MHz (with probes)	£298.00
204 With sweep delay	£402.00
205 Digital storage dual 20MHz	£498.00
605 Dual 60 MHz plus sweep delay	£567.00
CROTECH with component tester	
3031 Single 20 MHz 9.5cm CRT	£195.00
HITACHI V212 Dual trace 20 MHz 8 x 10cm	P.O.A.
GENERAL OST5m single trace 5 MHz 3 CRT	£139.09
THANDAR SC110A Portable 10 MHz 32 x 26mm display	£175.00



GENERATORS 220/240V AC operation



FUNCTION - sine, sq, triangle, TTL D/P	
J500 0.1Hz to 500kHz	£110.00
TG101 0.02Hz to 200kHz	£110.00
TG102 0.2Hz to 2MHz	£160.00
PULSE Square, Complement, TTL D/P	
TG105 5Hz to 5MHz	£105.00
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GENERAL EQUIPMENT

X1/X10 Probe kits (250MHz)	£10.00
20D 100kHz - 110MHz R-C OSC	£65.17
220 20Hz - 200kHz R-C OSC	£78.22
CM200 Bench Digital Cap. Meter	£89.00
TH301 LCO Digital Temperature	£59.50
2018 Audio Power Meter	£220.00
YF501 Electronic Insulation Tester	£69.00
PFM200A 200MHz Pocket LED Counter	£75.50
DM 358 8 CH. Scope Multiplexer	£179.00
KOM6 200MHz TR. DIP Meter	£43.04

ANALOGUE MULTIMETERS

1015 15 range pocket 10k/V	£7.50
1035 19 range 20k/V 10A DC	£10.39
1028Z 19 range 10A DC + Hfe 20k/V	£12.61
HC201 19 range 20k/V 12A AC/DC	£15.61
2020 20 range 10A DC + Hfe 20k/V	£19.09
3010BZ 24 range 10A DC + buzzer 30k/V	£20.83
155mm 21 range 10A AC/DC 20k/V	£22.57
BANANA 15 range + buzz shockproof 20k/V	£23.60
* 5050E 41 Range FET	£28.65

ALL WITH FREE CASE (*no case)

EHT METER

0/40kV Direct Reading	£45.00
-----------------------	--------

PROBE/PULSER

610B 20MHz Logic probe & bleeder	£16.95
625 50MHz Probe & pulser 0.5Hz-500Hz	£22.95
07 20MHz Logic probe	£9.52

AC CLAMP METERS

With cases and leads	
300 300amps (5 ranges)	
600V (4 ranges) 1K OHM	£30.87
4300 Autorange Digital 300A	
500V - cont. test	£75.00

DIGITAL CAPACITANCE METER

8 range LCD 200pF to 2000nF	
Push-button selection (5013)	£59.00

COUNTERS AND TIMERS



220/240V AC mains. 8 digit LED	
Counters 0.1Hz resolution, 5mV sensitivity	
100MHz 2 ranges	£99.00
600MHz 3 ranges	£126.00
1000MHz 3 ranges	£175.00
1500MHz 3 ranges	£199.00
Timers 0.001Hz Resolution, 100MHz 5mV sensitivity	
OP10 Standard model	£219.00
OP100 Incl. display hold, etc. stopwatch/ RPM, etc	£285.00



DC POWER SUPPLIES
240V AC Input. Stabilised variable output voltages. Direct meter reading volts and amps.

243 0/24 volts 0/3 amps DC	£43.26
245 0/24 volts 0/5 amps DC	£56.26
154 5/15 volts 0/4 amps DC	£33.26
Atiso 13.8V DC, 3 amps max. Fixed (no meter)	£13.00
0/30V 0/2 amps DC LCD display	£150.00

DIGITAL MULTIMETERS

3 1/2 digit LCD displays	
508 15 range, 200mA DC, 2 MEG	£19.96
105 14 Range 2A DC	£21.50
528T 13 range, plus Hfe test	£26.04
55C 28 range 10A AC/DC	£35.65
6010 28 range, 10A AC/DC, 20 MEG	£33.50
578 20 range auto-memory hold 10A AC/DC	£34.50
615 19 range, plus Hfe test, 20 MEG 10A DC	£30.39
5010 31 range, 10A AC/DC + buzz 20uA AC/DC	£36.50
TM357 30 range 10A AC/DC + buzz	£39.00
5010 EC as 5010 + cap + Hfe + cond.	£52.13

ALL WITH FREE CASE

SWITCH MODE PSU's

Open frame models. 1/P 100/240V AC	
240watt -5v 25A; -12v 7A; -12v 1.5A; -12v 1.5A	£34.74
SX510 2 to 5v (adjustable) 10amp	£7.78

BENCH DMM'S

3 1/2 Digit LCD 10A	
AC/DC 20 meg. Buzzer etc.	
322 0.25%	£ 89.00
3210 0.1% Tru RMS	£119.00

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- 68000 two pass assembler with system monitor and BBC link software in EPROM.
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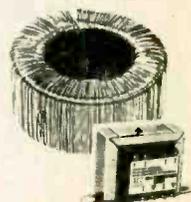
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Also available 1k2, 1k5, 2k, 2k5, 3k. Prices on request.
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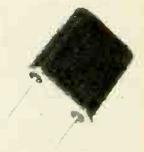


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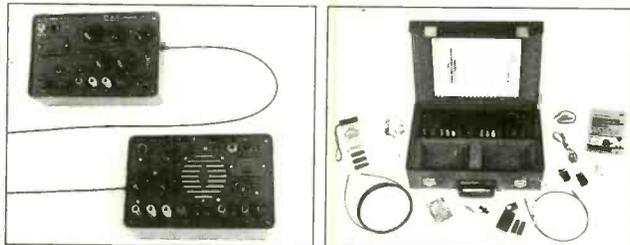
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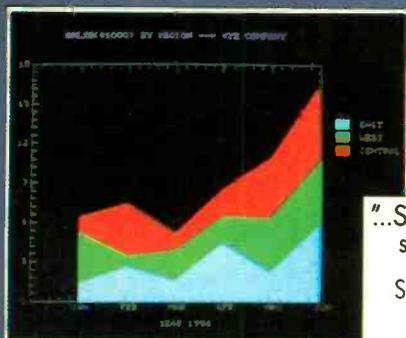
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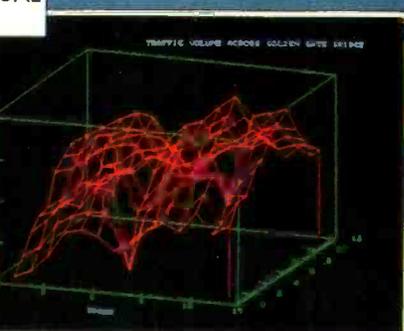
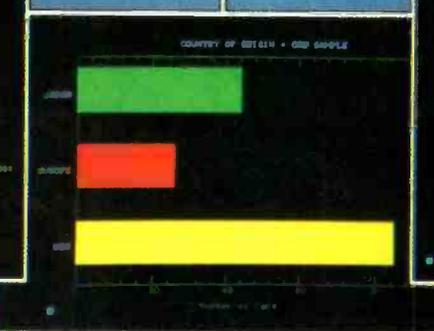
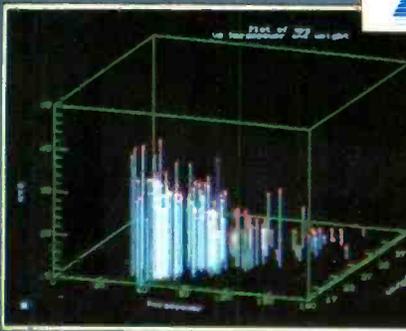
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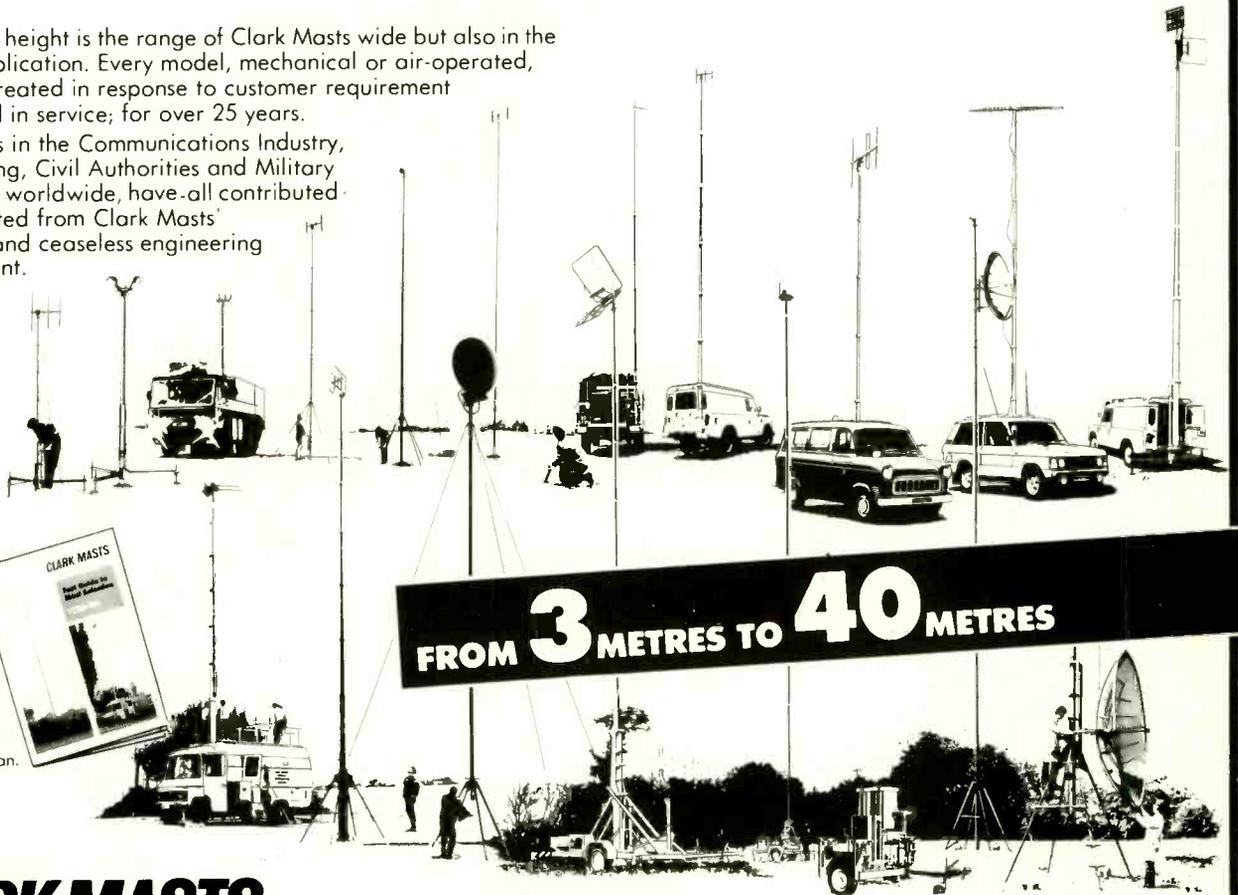
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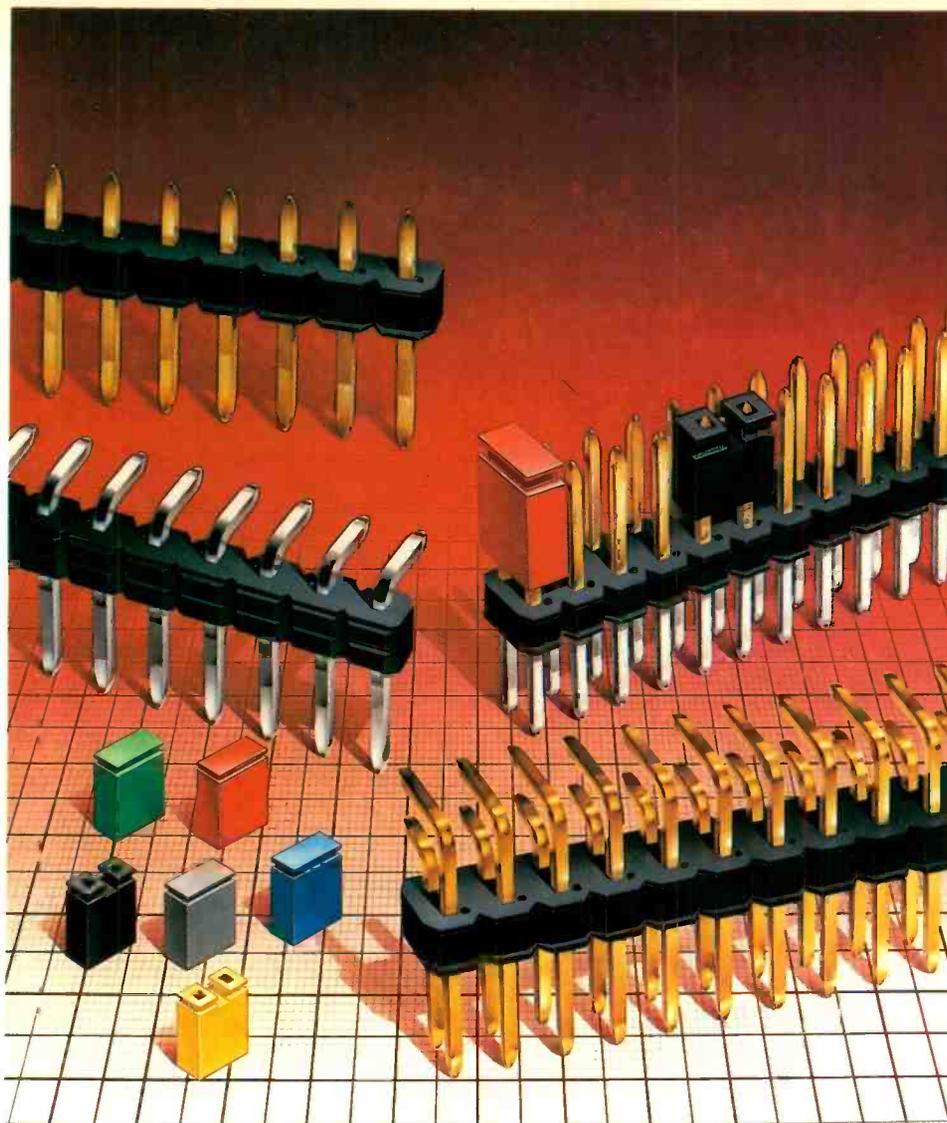


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Connector revolution

This round-up of connection techniques concludes that British-designed Hierarchical Interconnection Technology could be the basis of the next generation of p.c.b. interconnect

by JOHN CURRY

The semiconductor revolution which has transformed equipment design and construction methods over the past 25 years has in turn brought about a revolution in connection techniques. Methods of connecting components to p.c.bs, daughter boards to mother boards, mother boards to backplane boards, have all been streamlined to minimize costs and maximize equipment production, performance and reliability. Similar changes have taken place in cabling and interconnecting equipment.

Another factor that has helped transform connectors and connection techniques is the massive rise in price of traditional connector materials such as tin, brass, copper and gold. Most connector pins and sockets have therefore been redesigned to reduce the amount of material needed to produce them and cut

production time and costs wherever possible.

For example, most commercial connector pins are now pressed from metal sheet and folded, instead of being machined from solid metal, the pressing technique providing savings in material and allowing mass production at lower cost.

Costs have been further reduced by techniques such as selective plating of pins and sockets, and reducing gold and other plating thicknesses to a minimum.

The connector revolution has also been helped by the various industry recessions. These have made all sectors of the industry very cost conscious, and resulted in ever increasing demands for cost-effectiveness in equipment design, production testing and installation.

The need for greater cost-effectiveness has

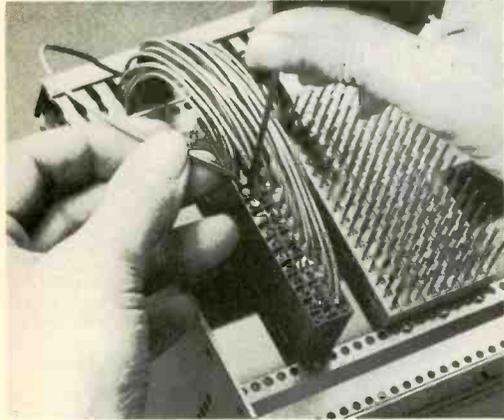
been complemented by what has become an obligatory requirement for shorter product development and manufacturing times brought about by increased use of computer aids. Manufacturers have to get their new products to the market as soon as possible to secure sufficient market share to cover high development costs. Thus productivity is a key requirement in all modern connection systems.

SOLDERLESS TECHNIQUES

All these factors have resulted in faster, cheaper and easier to use methods of connection and a massive swing to solderless techniques. In telecommunications, for example, the need to minimize operating costs has obliged organizations like British Telecom to adopt quicker and cheaper ways of

making connections in telephone system installations.

As a result, time-consuming terminations such as solder- and screw-terminal connections have largely been superseded by simpler and much faster solderless techniques such as insulation-displacement connection, which require the minimum of tools, little or no preparation of the wires to be connected, and therefore little skill.

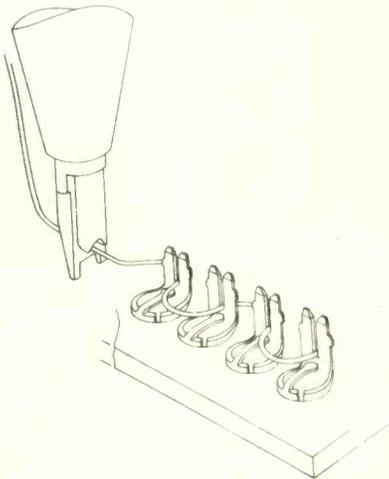


An everyday example is the now familiar white telephone plugs and sockets used in home and offices. These use a non-solder technique called cable-piercing which enables them to be wired with a single squeeze of a special crimping tool. The squeezing action causes the connecting contacts of the plug or socket to pierce the cable's outer insulation and make contact with the copper wire inside.

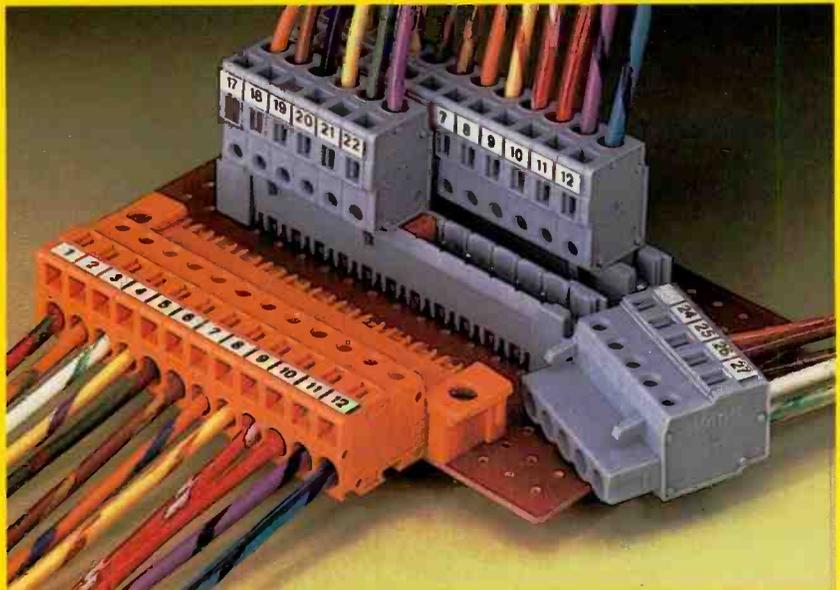
INSULATION DISPLACEMENT

Another widely-used solderless wiring technique is insulation displacement connection, a good example of which is the Rapid Wiring System that inserts and connects a wire in a single push action without the need for wirestripping.

Designed for wiring prototype or small production run p.c.bs by hand, or high density wiring of boards by machine, it



utilises a one-piece beryllium copper contact which pushfits into a p.c.b. and has an i.d.c. terminal at one end, and an i.c.-pin at socket the other.



2 to 24-pole p.c.b. connectors

A range of p.c.b. connectors comes from Hellermann Electronic Components. These are available in 2 to 24-pole and with a pitch of 5mm and 5.08mm. The female connector with cage-clamp connection is for solid, stranded and flexible conductors from 0.14mm² up to 2.5 mm², (AWG 26 to 14). Male connectors are available with straight or angled solder pins 1 by 1mm or 1.2 by 1.2mm. There is a test socket for test plugs 2mm and 2.3mm dia. Pole marking can either be with adhesive marker strips or direct printing. The female connectors can be operated from above as well as parallel to the conductors. The front entry system allows the connection or disconnection of conductors when the connector is mated. The male connectors are suitable for p.c.bs up to a thickness of 3.5mm. They can be screwed to the board with the aid of fixing elements to ensure that optimum guidance of the female connector is obtained. Depending on the mounting conditions, there are several possibilities for having the connectors mated in a non-reversible way.
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Screened D-connector

A modified version of the Ceep subminiature D-connector hood has been designed to minimize r.f.i. and e.m.i. and to offer continuous screening in conjunction with the cable braid. Metallized plastic construction reduces the cost and weight while remaining effective.
254 on reply card.

D.i.1. p.c.b. connector with i.d.c.

Du Pont Connector Systems has introduced an entirely new dual-in-line solder-to-board, and male pluggable, connector system. This is based on new i.d.c. technology and is suitable for high performance, high reliability applications in harsh environments, such as those encountered in the telecommunications industry.

Du Pont's "Quickie" d.i.1. p.c.b. connector has a patented i.d.c. contact with an integral anchor, which provides base-to-cover locking and ensures correct positioning of the conductor in the i.d.c. area. It also features a pre-loaded, bridge-type cover, with dual-sided cable entry, which makes it ideal for automated assembly operations and for daisy chain applications. Low profile strain-relief is available as an optional extra.

The new connector is supplied in grid sizes of 2.54 by 2.54mm, 2.54 by 7.62mm and 2.54 by 15.24mm, and with 4 to 64 contact positions. The male pluggable version has selective gold-plated contacts: the solder-to-board version has tin/lead-plate contacts.
259 on reply card

The wire is simply pushed into the slot created by the contact's two tines. The width of the slot is slightly narrower than the wire's conductor. Thus as the wire is pushed down, the two tines cut its insulation and firmly grip its conductor, making a secure gas-tight connection.

The single-push wiring action is three times faster than wirewrap and enables terminals to be daisy-chained with a common link, further reducing the number of wires that need to be cut. In addition, the i.d.c. tines are one third the height of comparable wirewrap posts, yet can accommodate three 30s.w.g. wires, and can be re-used up to 50 times without impairing the integrity of the gas-tight connections they provide.

The savings in time and space which i.d.c. connections provide, coupled with their reliability and re-usability, has prompted BT, and telecommunications utilities worldwide, to adopt these connectors as the standard method of connection for telephone wires and cables. BT is currently in the process of a massive conversion project which will result in their use in all UK telephone exchange and installation wiring. The telecommunications industry's recognition of its integrity as a dependable means of connection has helped spread the use of the technique to cables, including flat cable and coaxial cable. As a result, there are now i.d.c. versions of most cable connectors.

PRESS-FIT CONNECTORS

Solderless connectors are also superseding traditional soldered edge-connectors on printed-circuit boards in low- and medium-volume applications. Called press-fit connectors, they have tapered pins which are simply pressed into an appropriate pattern of plated-through holes on the p.c.b. This compresses and forces their outer surfaces against the plating so that they conform to the shape of the holes, rather than vice-versa, resulting in connections that are gastight, resilient and both electrically and mechanically sound.

The board requires no heat treatment. Thus the technique limits internal stresses and p.c.b. warpage. In addition, the pins can be readily replaced. It also enables a variety of terminations to be fitted, including plugs, sockets, and wiring posts such the RWS i.d.c. terminal.

For example, the widely used DIN41612 Eurocard connector is available with press-fit pins. Called "compliant" pins, they have a tapered c-shape cross section. Press fit connectors with a variety of other tapered-pin cross section shapes are also produced.

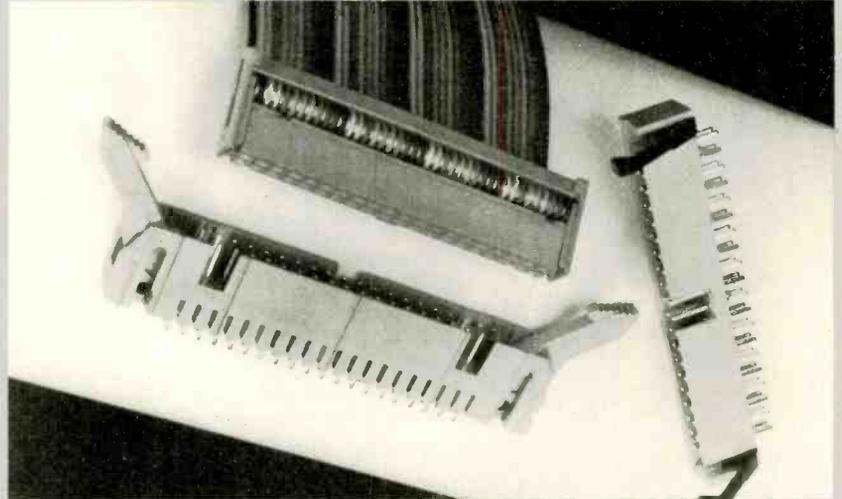
GTH CONNECTORS

Another popular solderless connector technology is the gas-tight high-pressure contact (g.t.h.) developed by Burndy which, as its name suggests, uses a contact-pressurising arrangement to make a positive gas-tight connection.

Connection is made by simply lifting the connector's outer body, inserting a pre-stripped wire into low insertion force g.t.h. socket, and then pushing the body down to

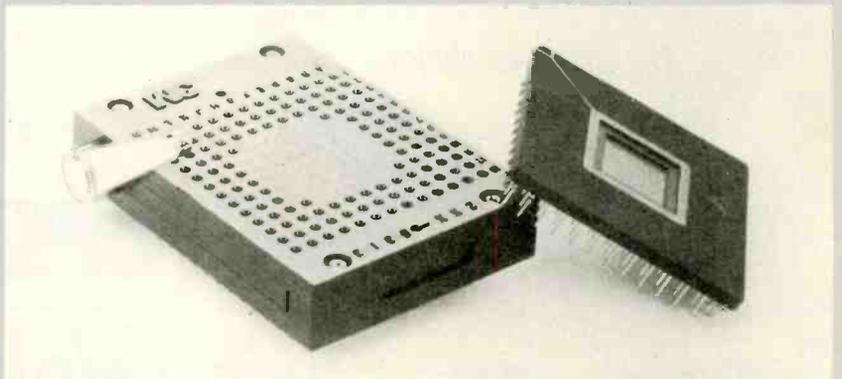
New latch for i.d.c.

What are claimed to be the only i.d.c. connectors in the world to offer both vertical latch and eject facilities are now available from Nortronic Associates. The top latching offers the option of considerable savings in p.c.b. area compared with equivalent side latch connectors. Both straight and right-angled versions are available and both of these styles are now stocked in 10 to 50-way DIN configurations. Each i.d.c. component has its pins fully protected and is centrally polarized as a guarantee against incorrect insertion. Insulation material comprises glass-reinforced thermo-plastic carrying a 94V0 rating, while each of the phosphor-bronze contacts has a gold plating over nickel. Current rating is 1A. A full range of mating female sockets is also available. 257 on reply card



Pin grid array sockets

A family of Textool zero insertion pressure (zip) test and burn-in sockets from BFI Electronics will pack tightly onto any p.c.b. surface and will handle a variety of P.G.A. (Pin Grid Array) devices. They will therefore be useful to quality assurance, test and programming departments. The zero insertion pressure feature of these sockets protects expensive P.G.A. devices from damage during insertion and withdrawal. The package is securely locked into place by lowering a level-operated cam. The sockets are designed to provide maximum use of board space. The cam level is positioned on the side of the socket, making the sockets end-to-end stackable. Optical locating holes on the sockets permit robotic loading and unloading, and because the lever remains above the edge of the socket when locked, manual unloading is easier. Seven matrix sizes, ranging from 10 by 10 to 21 by 21, are available. The pin pattern can be customized for any specific device on an average 7-day turn-around. 250 on reply card.



Handtool for i.d.c.s

An i.d.c. connector hand termination gun designed to provide the speed of assembly associated with i.d.c. but at price level normally realised by crimping tools is now available from Dage Eurosem.

The gun, manufactured by Robinson Nugent, has been designed so that certain elements of the connector assembly task, such as wire stripping and crimping, are no longer necessary. As a result the gun can complete twice as many terminations compared with traditional crimping tools during the same period. The hand gun, suited for the assembly of connectors used in computer, instrumentation and test and measurement applications, is capable of making typically 800 terminations an hour. The hand tool's ratcheting mechanism allows handle release only after the wire is seated correctly in the contact and also means that, if required, the connector can be reversed. A self-indexing system automatically positions the connector for wire insertion which speeds the assembly process. However, the tool can be manually indexed if required. The stuffer blade is adjustable to allow termination of 20 to 30a.w.g. wire with different insulation thicknesses. A range of metallized plastic or plastic hoods are available as required. 258 on reply card

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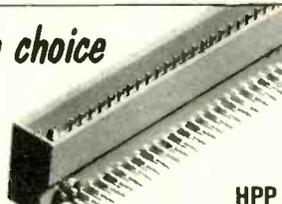
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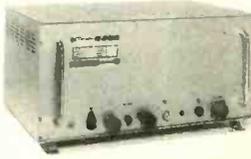
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its normal position. This forces the contact against the wire, and penetrates any oxidized deposits on the mating surfaces.

A key feature of the system is that it uses relatively inexpensive contact materials such as tin-plated phosphor bronze.

Burdny has applied the g.t.h. technique to a variety of different connection requirements including i.c. sockets, connectors for flat cables, connectors for terminating wires on p.c.b.s, and connectors for closely stacking p.c.b.s.

ZERO INSERTION FORCE SOCKETS

Another increasingly popular method of providing plug-in mountings for dip components on p.c.b.'s is the zero insertion force socket.

As its name implies, the zif socket enables dip devices to be fitted or removed without the need for any insertion or withdrawal force. Nor are any tools required. The socket's contacts are normally open, and are closed by a cam-action lever arm.

On being closed, the contacts grip the smooth flat sides of the dip device's legs, the last 15° of cam movement providing a sliding action that wipes the legs clean to ensure a positive contact.

Most zif sockets are press-fit devices with wire-wrap or similar pins, and therefore eliminate the need for board preparation.

HEIRARCHICAL INTER-CONNECTION TECHNOLOGY

Custom chips, v.l.s.i., hybrid circuits and surface-mount techniques collectively enable equipment designers to pack so much functionality and therefore complexity onto a p.c.b., that the value of printed circuit boards will rise significantly.

Double Eurocards for example can currently cost up to about £1500, but the increased complexity possible is likely to result in boards of this size containing functionality worth £50,000, i.e., over 30 times more valuable. Moreover their complexity would be so great that they will take several hours to test.

Thus there is a strong case for partitioning such designs into a number of small 'daughter' modules which plug onto a Eurocard, to simplify production and maintenance. A possible answer to this problem is a British-developed technique called Hierarchical Interconnection Technology (hit).

Hit is based on the use of 'daughter' modules whose profile is less than that of a DIN41612 connector, i.e. under 12mm. Up to eight can be mounted on one side of a double Eurocard. Single and double-sided designs in various sizes and substrates are being evaluated. They will fit into zif-type rectangular sockets having 130-140 contacts.

An IEE working group has been set up to prepare a proposal for a new equipment practice on Hit. Thus Hierarchical Interconnection Technology could be the basis of next generation of p.c.b. interconnect.

John Curry is managing director of Astralux Dynamics Ltd.

160 Way High-density DIN connector

A new design of DIN connector which meets the increasing demand for higher densities in interconnections is manufactured by Erni and distributed by Radiatron Components. The connector is fitted with five rows of 32 contacts, making 160 ways in total. Contacts are rated at 4A at 20°C ambient and are spaced on a 0.1 by 0.1in. matrix. Male connectors have right angle p.c.b. solder terminals: female connectors are offered with a choice of solder spills, wire wrap tails or compliant press fit terminations. Erni has also designed a special moulding which fits between two E160's on a double eurocard. This accepts up to nine of the special high current, co-axial or fibre-optic inserts already offered for the DIN 41612 M connections. This is part of a whole range of DIN 41612 connectors which include connectors for surface mounting and versions fitted with pressfit right angled spills.

261 on reply card

IBM PC-AT dual card slot connector

Viking Connectors (UK) Ltd has introduced a dual card slot, 18 + 31 position connector as part of their new low cost card-edge connector series. This new connector has been specifically designed for the motherboard i/o slots in the IBM PC AT computer or compatible products. The one-piece design replaces the need for two separate connectors, thus saving connector and production costs. Others of the Viking range of card-edge connectors include those suitable for all IBM PCs

262 on reply card

Backplane for industrial control

A pressfit backplane system for use in industrial control applications is announced by EMB Ltd. The board is flexible enough to form the heart of a wide range of complex multi-processor systems. Being DEC-compatible, it will be of interest in many industrial automation applications. The backplanes are typically multi-layer boards using universal bus configurations and termination networks suitable for address and data lines.

The design is based on DEC interface standards, with vertical headers for I/O. It is implemented using Pressfit solderless connectors.

263 on reply card

Microwave connectors

Norbain Technology Ltd has launched a range of high-reliability semi-rigid and flexible microwave cable assemblies designed for use in commercial communications and avionics applications. Designated the Seallex 2 and manufactured by Sealelectro, the cable assemblies are available in a variety of connector types including SMA, TNC and N types in either plug, right angle plug, bulkhead jack D hole and bulkhead jack 4 hole flange versions.

All microwave cable assemblies provide consistent microwave performance to 18GHz enabling low v.s.w.r. to be specified without spikes present in the upper frequency spectrum. In addition, they can be used up to 26GHz with minimal degradation using SMA connectors. All assemblies are 100% tested with each individual assembly supplied with v.s.w.r. test plots. The microwave cable assemblies are available ex-stock in half metre and one metre lengths in 0.125in, 0.18in and 0.250in standard cable sizes. Non-standard assemblies will be made to order.

264 on reply card

Surface-mount interconnection

Snap (sustained necessary applied pressure) is the name given by Dowty to their inter-board connection system. This uses an extruded flat conductor cable with a polyester coating which has gaps to expose the pre-tinned copper conductors at intervals. A simple tool is used to cut the cable and from the contact points in one easy operation. The Snap connector clamp provides uniform pressure across the connector strip. The tinned conductor tracks on the p.c.b. match the spacing of the connector cable and no soldering is required. The system requires two holes to be drilled in the p.c.b. (of different sizes for polarity). The connector requires little space on the board and is less than 13mm high, is gas-tight and vibration resistant.

252 on reply card

PLCC sockets

A new range of competitively-priced, rugged Textool plastic-leaded-chip carrier sockets offering ease of alignment and extraction of Jeduc C packages is now available from Microbusiness, Newbury. The design of these production sockets allows visual and mechanical polarization of both the device into the socket and the socket to the circuit board to ensure positive alignment and electrical integrity. Slotted corners provided greater ease of device extraction with a simple handtool. In addition, the socket offer 0.025in test access probe holes for each contact; 0.020in stand-offs for cleaning and ventilation holes for solvent removal; and sturdy solder tails with radiused ends for easy installation.

253 on reply card

Headers and links for p.c.bs

The Winslow range of unshrouded headers are available in single and dual rows with straight contacts or at 90°. They can have any number of contacts up to 72. Two thousand standard products cater for different pin lengths: but anything not in the catalogue can be made to order. Shorting links have open or closed tops and six colours. They offer a low-cost alternative to d.i.l. switches. Terminals can be tin or gold-plated.

256 on reply card

Designing a surface-mount connector

New 34-way s.m.c. incorporates key market features.

ADRIAN HYNER

A new surface-mount connector idea is aimed at meeting the need for an s.m.c. that can interface with existing component and substrate technologies but will offer the most critical benefit s.m.t can bring, namely savings in board and system space.

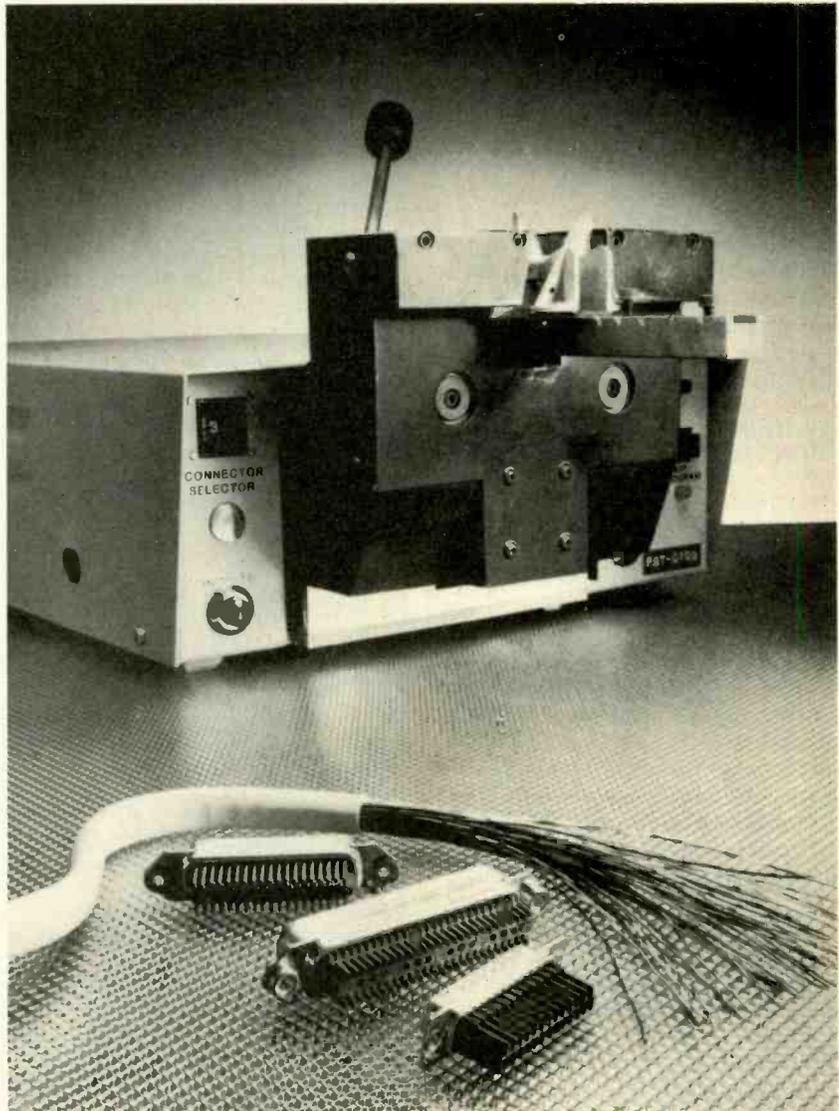
The majority of s.m.c. designs have been modifications of standard connector systems and as such suffer from severe disadvantages. These include too high a profile and they can suffer from structural failures resulting from a lack of sufficient mounting stability to cope with high insertion and removal forces.

Among the major features to bear in mind when developing a new purpose-designed low-profile s.m.c. is that, for the moment, the receptacle tails must terminate on one side of the p.c.b. Because surface-mounted components are not yet available in sufficient variety and at the right price to make wholly-s.m.t. designs a viable solution for many applications, most s.m.t.-based p.c.b.s currently feature a mixed-technology approach, in which both s.m.t. and p.t.h. components appear on the same board and are all soldered on one side of the board. A further limitation is presented by the fact that cable pitch of 0.005in is the current industry standard, restricting the p.c.b. header interface to that pitch.

Admittedly these are temporary restrictions. There is no doubt the next step will be to have the surface-mount connector on both sides of the p.c.b. because of the density advantages that can be accrued by terminating on two sides. Amphenol is already producing 0.0025in i.d.c. cable and the first products in a connector range to terminate it, with a view to producing an s.m.c. on 0.0025in pitch in due course. Nevertheless, for the interim period when these restrictions do apply, a connector solution must be considered.

A further requirement from a low-profile s.m.c. is that the receptacle should exhibit various mounting capabilities so that the connector can be either glued, solder tabbed, rivetted or even latched in place on the p.c.b. For military/aerospace applications, the often severe vibration problems dictate a different approach, requiring a variety of fixing operations to ensure that the connector stays in place in harsh environmental conditions.

The connector must be capable of assembly using a fully automatic machine, but



Machine speeds i.d.c. connectors

Latest in a line of semi-automatic connector assembly machines from Amphenol is the microprocessor-based Electro-Pierce Mark V. Its production rate is 30 to 40 connectors per hour. The machine is programmable for the entire range of connector sizes, from 14 to 64 way, the different connector sizes being accommodated by simple adjustment of a nest fixture. It can be programmed to skip selected positions. This skip function program is learned by the machine from a known-good connector and is retained until the machine is switched off.

Wires are selected by the operator, one pair at a time, and are positioned to activate the drive motor. The machine automatically cuts the wire to length and inserts it into the contacts, and then positions for the next pair. Optional accessories include an adaptor kit for top entry and rack and panel connectors, and a cable clamp closer for 157 Series connectors with cable clamp and plastic hood.

251 on reply card.

must also be of a design which is simple to use if manual assembly needs to be employed.

MARKET REQUIREMENTS

An in-depth study of these market requirements has highlighted eight key features that are critical to the success of a new concept in s.m.c. technology.

- Low insertion force contacts. A pull force retention system in the form of a connector latch or dimple to retain the connector in position is also a necessary feature.

- Low profile – the receptacle needs to have a profile of less than 5mm above the p.c.b. surface for most applications. A high profile can lead to the imposition of stress on the solder joint, leading to the possibility of tearing the connector completely off the board. A lower profile connector imparts less stress on the solder joint. Amphenol has designed a connector which, even if mounted on the surface of the p.c.b., has a profile less than 5mm. In fact, they recommend mounting onto the p.c.b. from the side to further reduce the profile (to 3.5mm) and improve rigidity. A rigid tongue incorporated in the connector design locates with the edge of the p.c.b, improving the rigidity of the board around the solder joint so that when the assembly passes through the soldering stage, which may be a vapour phase process running at a temperature of about 215°C, there are no inherent problems related to board flexing due to thermal mismatching of board/connector materials.

- Receptacle bodies must be made from a high temperature plastic capable of withstanding 230°C for 90 seconds.

- A 'snap-off' carrier strip is necessary to hold the contacts on 0.005in pitch and maintain vertical complanarity of contacts during assembly.

- Pitch 0.005in or less – already discussed.

- Small footprint.

- Reliability – it is considered that a two-piece connector along the lines of DIN 41612 is a better solution than edgcard connectors.

- Although the connector can be shrouded, this defeats the object of pursuing a low-profile design. Male half-plug contacts need to be recessed into the tongue to prevent shorting if the connector is to be unprotected.

- Built-in strain relief for the contacts, by incorporating a slot cut into the i.d.c. clip and folding the ribbon cable through it, is an added refinement.

The first low-profile s.m.c. incorporating all of these features is a 34-way product. Intended as a sampling product for examination, test and evaluation, it will be followed by a complete family based on the new concept. Versions already in the pipeline include double-sided connectors and mother-daughterboard types. The design results in a 60% space saving compared with, for example a 96-way DIN equivalent cable interconnection or mother-daughterboard connector.

Amphenol recognises that within the p.c.b. industry the need will remain for existing interfaces and, with this in mind, has developed s.m.t. versions of standard connectors such as headers, sockets, 57 series, D-type and DIN41612. However the company also believes that a new approach is essential to meet specific s.m.t. needs and that by working closely with p.c.b. equipment manufacturers, by using new materials and new contact technology, it has got the basic concept right. There is no doubt it can be developed to meet the spectrum of s.m.t. interconnection requirements, and Amphenol is hoping that it will lead to the generation of new standards idealised to s.m.t.

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NEXT MONTH

Displays for instruments. Liquid-crystal, led, plasma, c.r.t...? Our February feature reviews the applications and preferences for the various types of alphanumeric display used in instrumentation.

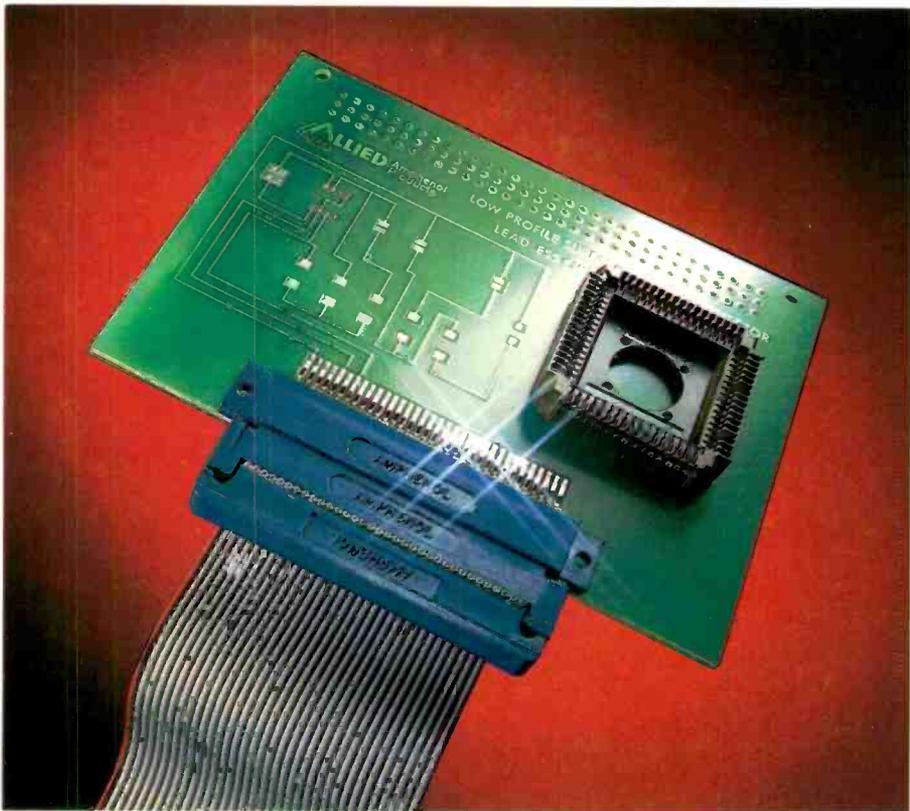
Desing of video output stages. L. Sage contributes a piece on the design and development of a high-performance video amplifier for medium and high resolution colour tubes in monitors fitted with RGB inputs. Bandwidth extends to 100MHz and slew rate is 9000V/μs at 25V swing.

What is a seg?. If the single-ended conductance can be thought of as a new circuit element, it should be possible to create novel circuit functions. Part 1 of the article sets the scene to demonstrate some interesting consequences.

VME developments. Multiple-bus architectures are by no means new, but recent specifications provide the means to quickly integrate multiple systems with off-the-shelf board products – in particular STE and VME bus systems. This second VME feature also discusses the impact of VME in the UK and introduces the latest subsystem standard.

Microprocessors – design for test. The time a manufacturer needs for volume testing v.l.s.i. devices is constantly increasing with the greater complexity of modern designs. Mike Catherwood describes the techniques of on-chip test facilities and structured testing.

SEE OUR SUBSCRIPTION OFFER ON PAGE 56



Hadamard versus Fourier transformation

The greater operational speed and simplicity of the Hadamard method is more suitable for use with microprocessors and ideally suited for real-time signal processing of speech

MARK S. VARNEY

Fourier analysis fits a series of sinusoidal waveforms to original data. Its transform algorithm requires a sine and cosine look-up table and special multiplication routines, adding greatly to the time taken to complete the transformation calculation. Sinewaves are used because they produce the same form on integration or differentiation, and linear systems give sinusoidal outputs for sinusoidal inputs.

It is equally possible to use any other waveform, so long as it can be described mathematically. All transforms are based on the fact that two linear combinations of two numbers contain the same information as the original numbers. Hadamard transformation (HT), for instance, involves the use of rectangular waves (otherwise known as Walsh functions) of differing periods and phases. The special advantage of the Hadamard method is that the wave has values of either +1 or -1 and so only additions or

subtractions are necessary - making it an ideal algorithm for writing on a computer, particularly eight-bit microprocessors such as the Z80 and 6502 which do not have hardware multiply features and would require additional code just for multiplying two eight-bit numbers.

MATRIX ALGEBRA

The Hadamard matrix H_n of order n is an $n \times n$ matrix of plus and minus ones, whose rows and columns are orthogonal to one another. The smallest order Hadamard matrix is two,

$$H_2 = \begin{bmatrix} 1 & 1 \\ 1 & -1 \end{bmatrix}$$

and by limiting the order to an integral power of two, symmetrical matrices will be obtained, for example

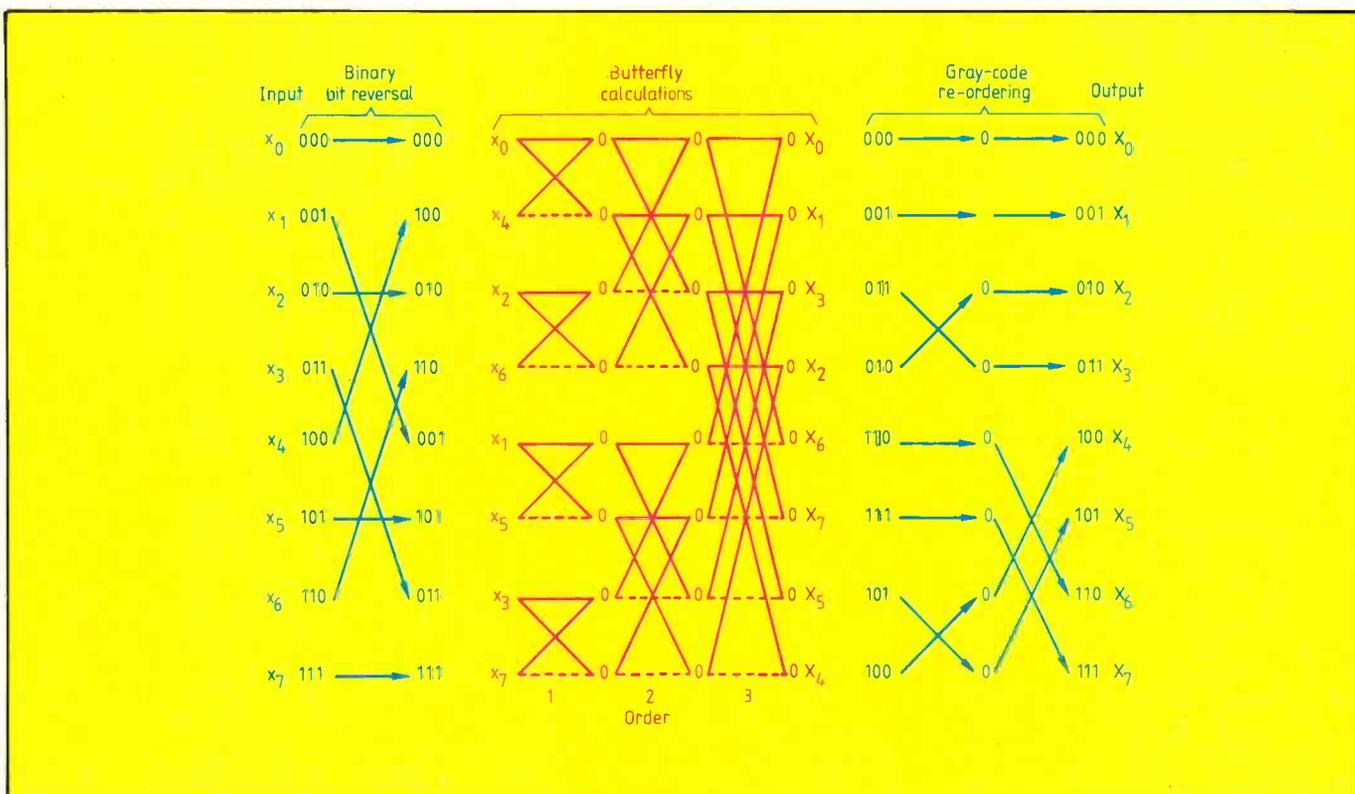
$$H_8 = \begin{bmatrix} 1 & 1 & 1 & 1 & 1 & 1 & 1 & 1 \\ 1 & -1 & 1 & -1 & 1 & -1 & 1 & -1 \\ 1 & 1 & -1 & -1 & 1 & 1 & -1 & -1 \\ 1 & -1 & -1 & 1 & 1 & -1 & -1 & 1 \\ 1 & 1 & 1 & 1 & -1 & -1 & -1 & -1 \\ 1 & -1 & 1 & -1 & -1 & 1 & -1 & 1 \\ 1 & 1 & -1 & -1 & -1 & -1 & 1 & 1 \\ 1 & -1 & -1 & 1 & -1 & 1 & 1 & -1 \end{bmatrix}$$

Because these forms of the Hadamard matrix are orthogonally symmetric, the procedure to compute the forward and reverse transform are basically the same, as with the Fourier transform. The one-dimensional Hadamard transform of a real signal may be defined as

$$X(i) = \frac{1}{N} \sum_{j=0}^{N-1} x(j) \text{Had}(i,j)$$

where $X(i)$ is i th (normalized) Hadamard coefficient, $x(j)$ is j th signal point, and

Fig. 1. Steps involved in an eight-point Hadamard transform include binary bit-reversal, butterfly calculations of the transform itself, and Gray code reordering. Circles represent intermediate products, solid lines represent additions, dashed lines represent subtractions, and arrows represent reordering.



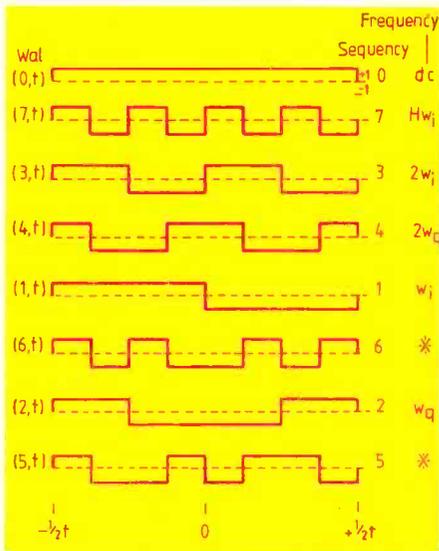


Fig.2. Walsh, or switching, functions used in the eight-point Hadamard transform are denoted by their frequency and phase angles (i =in-phase, q =quadrature). The function that crosses the time axis i times is called the i th Walsh function and denoted by $Wal(i,t)$. Those denoted by * are not regular square waves and have no exact physical representation.

$Had(i,j)$ the i th Hadamard function.

Like the fast Fourier transform (FFT), the Hadamard can be computed 'in-place' in order to save precious memory space. Similarly, the structure of the FFT can be exploited to give an algorithm for the fast Hadamard transform (FHT) without using complex numbers. Moreover, by integrating the normalization factor, $1/N$, into the algorithm, the size of the data block will be limited only by the available ram. The in-place method stores intermediate products from the butterfly products in the locations of the input samples, without the need for auxiliary storage. However, the transformed coefficients are then in their natural form (not in order of increasing frequency). To obtain the ordered sequence, the coefficients are firstly bit-reversed as in the FFT either before (time), or after (frequency) the transformation, and finally rearranged with respect to a Gray code permutation (Fig.1). In effect, the original signal waveform is multiplied by each of eight waves as in Fig.2, which correspond to each row of the eighth-order Hadamard matrix. The analogy with phase-sensitive detection is clear enough. For instance, the d.c. component is just the summed average of each of the original points multiplied by 1, the first waveform, labelled d.c. The in-phase and quadrature components merely use a single frequency wave shifted in phase by 90° . The first harmonic uses the fifth labelled w_o , the 90° out-of-phase component the seventh waveform, or w_q and so on for the second, fourth, etc harmonics.

SYNCHRONOUS SIGNAL DETECTION

In an analogue lock-in amplifier, a sinusoidal input signal $Asin(\omega_o + \phi)$ of amplitude A , angular frequency ω_o and phase angle θ is actually multiplied by a square wave of unit amplitude and angular frequency ω . The

PROGRAM 1. 8086 assembly language version of the fast Hadamard transform uses the FIG-Forth assembler and Forth variables to hold current values of the loop counters, etc. Binary bit-reversal is the same as for the FFT, and the Gray code reordering is not included for brevity. The 16-bit data is assumed to be in two's-complement.

Due to circumstances beyond our control we are unable to include the programs in this issue. They will appear in next month's issue, but readers in a hurry may obtain a copy from the editorial office.

technique is however limited to the measurement of only one frequency at one phase angle. The principle of operation is of a binary nature – it involves multiplying successive points in the waveform by $+1$ and -1 , i.e. alternately adding and subtracting each point. The digital version of this allows the simultaneous acquisition of the fundamental and subsequent harmonics, and the in-phase and quadrature components.

To extract information from a specific frequency and at a specific phase angle from 2^N input data points, a total of 2×2^N additions and subtractions are needed, i.e. 2^N at two orthogonal phase angles. For N different frequencies the number is $2N \times 2^N$. As a consequence of the Nyquist limitation, information on only $N-1$ frequencies can be extracted from 2^N points, that is at $n=0,1,2,\dots,N-2$. Digital synchronous detection of both the in-phase and quadrature components at $N-1$ frequencies would require $2(N-1) \times 2^N$ operations, whereas the Hadamard transform only requires $N \times 2^N$. Thus, another advantage of the Hadamard transform over straightforward digital detection is a factor of $2(N-1)/N$, which approaches 2 as the size of the data array increases.

There is no dispute that the FHT is a much faster method than the FFT by up to two orders of magnitude in my experience. However, although the fast algorithm has been used in this article, the slow version is just as simple to write and implement. It

would even be possible by simple extension of the above ideas to write an algorithm to determine only those frequencies of interest, and increase the speed of calculation still further.

MULTIPLEX FREQUENCY ANALYSIS

A disadvantage of this technique is, however, that it can be more restrictive in the choice of measurement frequencies that can be multiplexed. In fact, all the integral frequencies are an exact power of two i.e. 0,1,2,4,8,16,etc. The other coefficients in between these frequencies are not true regular squarewaves, and cannot represent actual frequencies but they are necessary in the calculation of both the forward and reverse transforms.

A frequency interpretation can be given to the whole of the Hadamard matrix. Along each row in Fig.2 the "frequency" is represented by the number of changes in sign, the term "sequency" has been used instead to describe frequency domains. It is possible to construct a Hadamard matrix of order $N=2^N$ that has (sequency) components at every integer from 0 to $N-1$. This may be of consequence to those who wish to avoid (or promote!) the effects of intermodulation between frequencies. However this is not such a disadvantage in areas such as speech recognition since only the major coefficients

continued on Page 42

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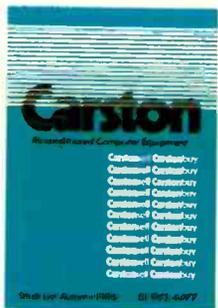
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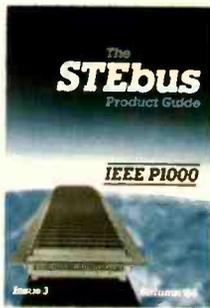
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Modelling the cochlea

The ear's hearing organ is too small to model mechanically. But an electrical equivalent mimics its behaviour well enough to provide a useful study tool.

R.W. GUELKE, J.P. RAMACKERS AND A.E. BUNN

Our sense of hearing can distinguish between sounds of different frequencies with a high degree of accuracy. The mechanism responsible for this frequency discrimination is understood to be the shell like structure known as the cochlea.

It consists of two galleries or tubes – the scala vestibuli and the scala tympani – separated by an elastic membrane known as the basilar membrane. Incoming sounds vibrate the tympanic membrane (eardrum) and the ossicles (malleus, incus and stapes) and these in turn excite vibrations at the oval window of the cochlea. These sound waves then proceed towards the helicotrema, the opening between the two galleries at the end of the cochlea (Fig.1). It has been established that high audio frequencies excite the basilar membrane close to the stapes end of the cochlea whereas low audio frequencies excite the membrane near the helicotrema.

Recently the pattern of vibrations along the cochlea has been determined and demonstrated on a mechanical model^{4,6}. The mathematical treatment involves the use of electrical-acoustical analogies, transmission line theory and the idea of transverse resonance.

This mechanical model was larger than the cochlea and so its frequency range was correspondingly lower. A mechanical model to cover the same range as the cochlea would have to be approximately the same size as the cochlea itself. This would present problems that would be difficult if not impossible to solve.

However, the action of the cochlea can also be demonstrated by a model in which the mechanical and hydraulic quantities are replaced by their electrical equivalents. Such a model can be designed to cover the

range of frequencies perceived by the ear yet does not need to be the size of the cochlea.

To cover the audible range it is necessary to replace the continuous line by a number of lumped impedances. This can be done by dividing the cochlea into a number of sections and replacing each section by an electrical circuit which corresponds to the mechanical or hydraulic system of that particular section.

ELECTRICAL-ACOUSTICAL EQUIVALENTS

Before developing an electrical model it is necessary to choose the equivalents that are to be used and to establish a numerical relationship between them.

Equivalents for an element Δx (Fig.2) are given on the initial assumption that we are dealing with a continuous system. The element can then approach the infinitesimally small element dx and the usual methods of the calculus apply. Figure 2 also shows differential equations applicable to both systems, from which the correspondence between the quantities chosen is obvious.

If lumped impedances are to be used the element Δx will be finite and each section will be represented by a number of individual capacitors, inductors and resistors, forming a circuit which can be dealt with using normal circuit theory. Each section will consist of:

1. An inductor L_1 to represent the series inductance of the galleries. Using the definition of inductance as

$$\text{this will be } \frac{\text{mass}}{\text{area}} \frac{\rho \Delta x}{A_1} + \frac{\rho \Delta x}{A_2}$$

where ρ is the density of the liquid in the galleries (endolymph and perilymph); A_1 and A_2 are the cross section of the scala vestibuli and the scala tympani at the point in question.

2. A capacitor C to represent the compliance of the basilar membrane equal to $C\Delta x$. C is the volume compliance per unit length and according to Bekesy² can be represented approximately by the expression

$$C = 5.1 \times 10^{-11} \exp 0.01355x$$

where x is the distance from the stapes end of the cochlea (metres).

3. An inductor L_2 to represent the transverse inductance proportional to $1/\Delta x$ and given by the following expression⁶:

$$\frac{\rho n}{2\pi(r_2^2 - r_1^2)^2} \left\{ r_2^4 \ln \frac{r_2}{r_1} - \frac{3}{4} r_2^4 + r_2^2 r_1^2 - \frac{1}{4} r_1^4 \right\}$$

The same expression calculated using r_2 corresponding to A_2 :

$$r_1 = \frac{nW}{2\pi} r_2 = \sqrt{\frac{nA}{\pi} + r^2} n = \frac{2\pi}{\theta}$$

θ is the angle subtended by equivalent sector (radians) W the width of basilar membrane, and A_1, A_2 the cross section of each gallery.

4. Resistance associated with each inductance (neglected).
5. A conductance to represent losses in the compliant element (hysteresis losses).

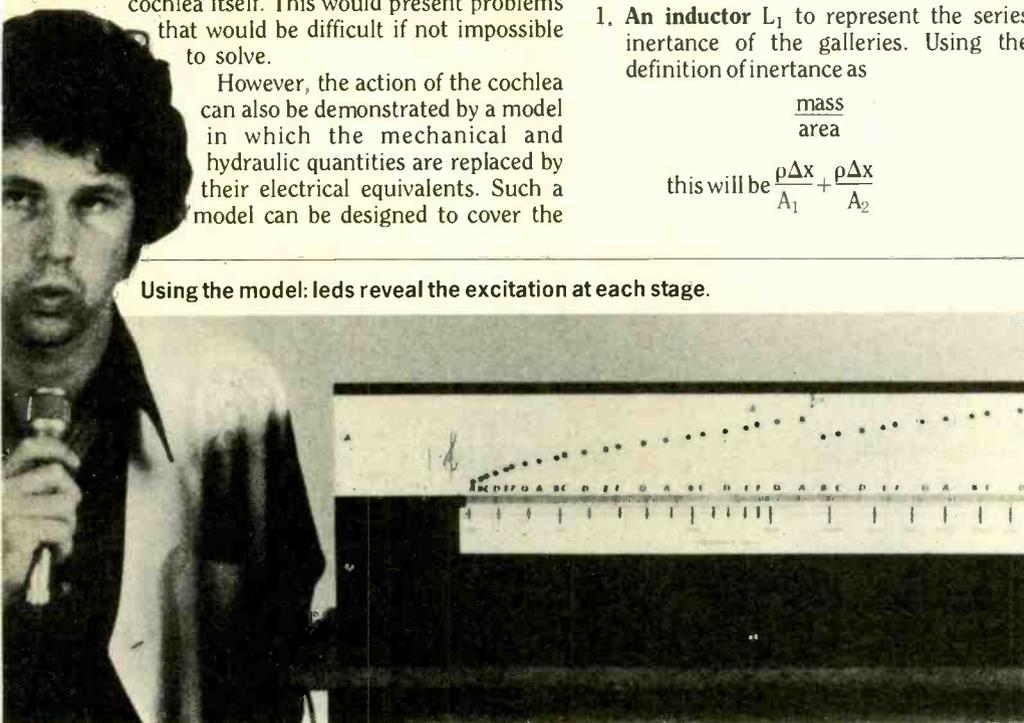
DESIGN OF THE ELECTRICAL MODEL

The larger the number of sections in the design, the closer it will be to conditions in the cochlea itself. The structure of the cochlea involves over 4000 sections if the arches of Corti represent discrete elements. A compromise has to be reached so that the essential behaviour of the model is comparable to the cochlea.

Previous models have used 40 to 65 sections^{1,5} and have recorded the distribution of voltages in accordance with the place theory, i.e. high frequencies exciting the sections at the stapes end and low frequencies towards the helicotrema end. If, however, this circuit is examined more closely it will be seen that the individual sections act as low pass filters, producing an analysis of the incoming signal which is not related to the way in which the cochlea works.

It can be estimated that to avoid any effect

Using the model: leds reveal the excitation at each stage.



due to low pass filter action for frequencies up to 5000Hz, at least 100 sections should be used. A model similar to the one described here used 175 sections³. Details of the magnitude of the individual components were not given, however, and so no comparisons can be made.

For the model described in this article 190 sections were used; this number also provides for adequate resolving power. If the range chosen is from 200 to 5000Hz, or 4.6 octaves, we have about 3.4 stages for every semitone.

CHOOSING THE ELECTRICAL COMPONENTS

Deciding the number of stages establishes the equivalent electrical components for each stage. It has been shown⁶ that all the elements in the cochlea vary from the stapes end to the helicotrema end. However, the main variation is provided by the variation in volume compliance due to the varying width of the basilar membrane.

For simplicity we ignored the variations in series inductance and transverse inductance and chose a mean value of these two parameters. Only the equivalent of the volume compliance was varied. Bekesy² measured a ratio of about 100:1 for the volume compliance from the stapes to the helicotrema. In the model a larger ratio of 575:1 was used for the corresponding capacitors, to extend the response down to about 200Hz.

Having chosen the number of stages to be used and the ratio of capacitors over the range there is one more arbitrary choice to be made concerning the electrical equivalents. If the chosen range of frequencies is to be covered, the ratio of the capacitors to the inductors is determined. An arbitrary choice has still to be made of either a capacitor chosen to be the equivalent of a particular compliance or an inductor to be the equivalent

TABLE 1: hydraulic quantities and their electrical equivalents.

Hydraulic (acoustic) quantity	Corresponding electrical quantity
pressure (pascals)	e.m.f. potential (volts)
flow (cubic metres per second)	current (amperes)
inertance (pascals per cubic metre per second squared)	inductance (henrys)
compliance (cubic metres per pascal)	capacitance (farads)
resistance (cubic metres per second per pascal)	resistance (ohms)

of a particular inertance. Table 2 indicates the actual values selected. Parameters for the cochlea are taken from measurements on the geometry of the cochlea and from Bekesy's measurements of the volume elasticity of the basilar membrane². No measurements of the losses in the cochlea are available. It was assumed that the resonances were associated with a Q factor of about 20. This factor was easily achieved for the inductances representing the transverse inertance. Figure 3 gives the basic circuit used in the model and Fig.4 shows the detailed circuit.

The choice of a particular equivalent inductance for the transverse inertance and the consequent choice of capacitors to provide resonance corresponding to the audio range then determines the characteristic resistance of the line. To avoid reflections the line was terminated by a resistor equal to the characteristic resistance of the last section, given by

$$\sqrt{\frac{L_1}{C_1}}$$

If resonance occurs before the end of the line, the attenuation after the point of

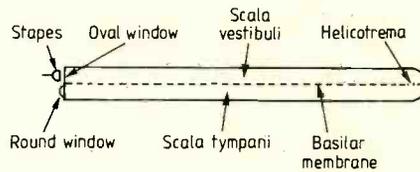
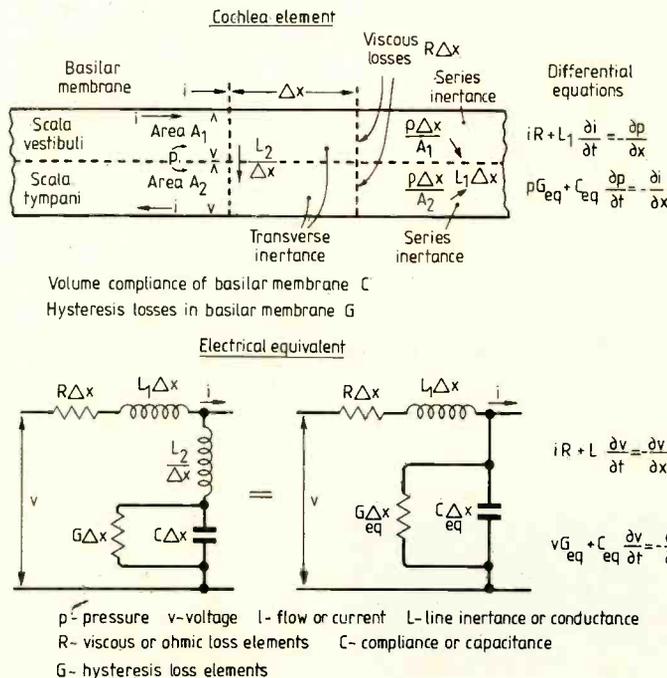


Fig.1. Diagrammatic representation of the unwound cochlea.

Fig.2. Elements of a section of the cochlea and their electrical equivalent.



resonance is so great that reflections are of no consequence and the termination of the line is unimportant.

TRANSVERSE RESONANCE

In previous models the transmission line representing the cochlea consisted of series inductances and shunt capacitances only. An inductance in the shunt arm was omitted altogether.

Calculation of shunt inductors to simulate the shunt inertance at any particular point has shown that these shunt inertances are due mainly to liquid loading: combined with the volume compliance at the point of question, this produces a resonance which explains the function of the cochlea as a frequency analyser.

In addition to this liquid loading the mass of the basilar membrane will also contribute to the transverse inertance. This mass, however, contributes much less than the liquid loading and for this model it is neglected.

Note that this inertance is inversely proportional to Δx whereas the volume compliance is proportional to Δx . The frequency of resonance is therefore independent of the element of length that is being considered but will vary along the cochlea depending on the variation of the volume compliance from one end to the other. Bekesy², as mentioned previously, found a variation of 100 to 1 in the volume compliance when processing from the stapes to the helicotrema. It is quite possible that even larger variations exist because Bekesy carried out his measurements on cadaver specimens and during life there may be differences.

Other investigators have suggested,

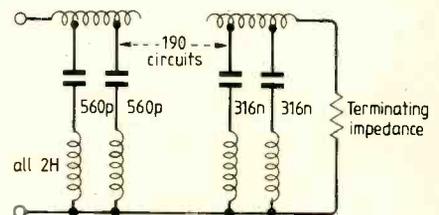


Fig.3. Basic circuit of the model.

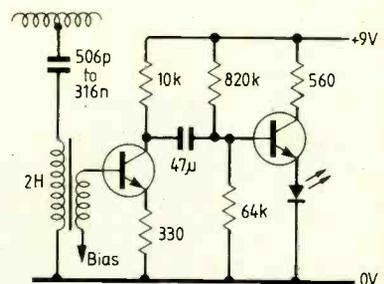


Fig.4. Detailed circuit of each stage.

TABLE 2: parameters of the cochlea and (right) electrical equivalents used in 190-section model. Note that the values of inertances per section are double the values for each gallery. Capacitance at the low frequency end was increased above the value corresponding to Bekesy's measurement to extend the range to 200Hz.

Compliance		Series inertance (each gallery)		Transverse inertance (Each gallery)		Capacitance	Series inductance henrys	Transverse inductance henrys
per metre	per section	per metre	per section	per metre	per section	farads		
from 5.1×10^{-11}	9.1×10^{-15}	6.67×10^8	210×10^3	803	9.0×10^6	0.5×10^{-9}	0.025	1.8
to 5.1×10^{-13}	9.1×10^{-17}	3.33×10^6	120×10^3	962	10.8×10^6	56×10^{-9}	(average)	2.3
						316×10^{-9}		

however, that this ratio should be increased^{3,7,8,9}. There is therefore some justification for the increase which has been chosen for the model.

PRACTICAL DESIGN

In this electrical model a light-emitting diode display indicates the response at each section to the incoming excitation. It was found necessary to incorporate an amplifier so that the inductors did not have to carry currents which would run into the non-linear region of their magnetic cores.

The 190 sections (Fig.4) were accommodated on 12 printed circuit boards. The leds were placed in line so that a frequency analysis of any complex sound could be observed.

RESULTS

The model demonstrates effectively the various patterns of excitation which are to be expected from different sounds. Figure 5 illustrates these patterns: the formant patterns of the various vowels are clearly indicated and the filtered white noise from the sibilants is also clear. It is interesting to follow the pattern produced by a soprano voice from a recording. The vibrato which is normally used plainly visible.

An instrument of this type could probably be used effectively in voice training, particularly for the hard-of-hearing.

Professor Guelke and colleagues are in the ear, nose and throat department, Faculty of Medicine, at the University of Stellenbosch, Tygerberg, South Africa.

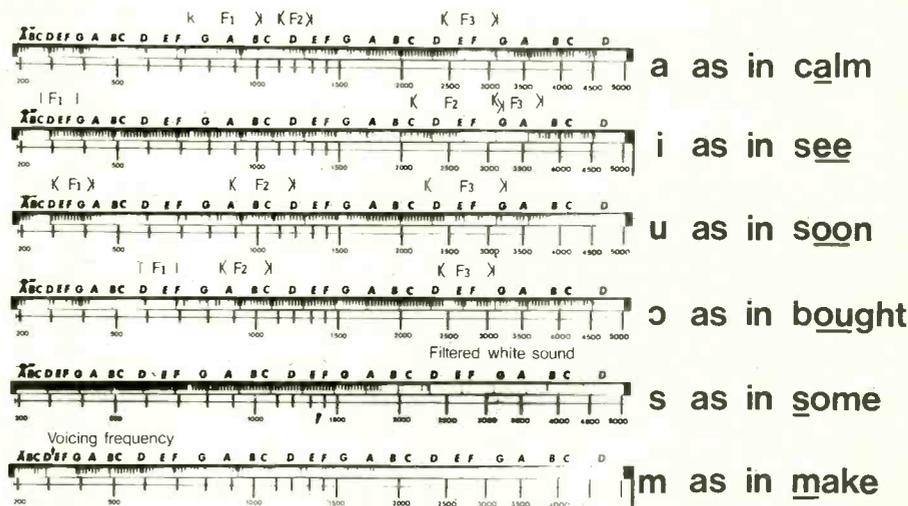
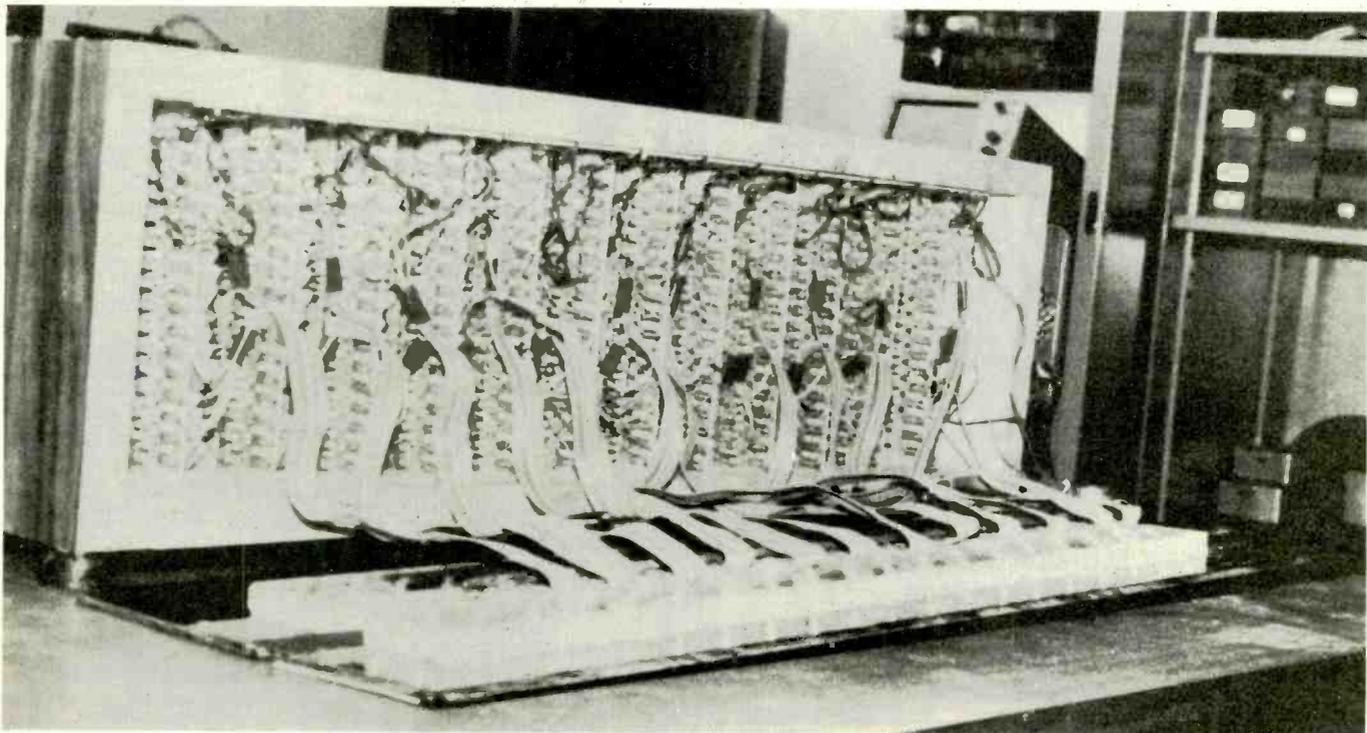


Fig.5. Speech sounds recorded on the model. Formant regions F_1 , F_2 and F_3 indicated for the vowels concerned. Although these regions are not precisely coincident with the observed patterns, the general agreement is clear.

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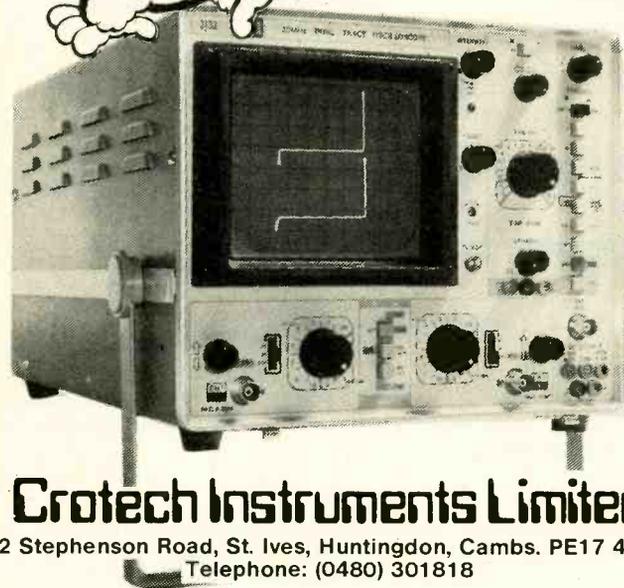
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Behind the panel: 190 sections over a range of $4\frac{1}{2}$ octaves, 200-5,000Hz.





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Video frame store

Further details of the control board which, with the memory board described next, is all that is needed to set up a basic system.

D.E.A. CLARKE

Figure 15 shows the relationship between line sync and the clock. The trailing edge of the sync-derived waveform from Link 2 is differentiated by C_{103}/R_{106} and the resulting narrow negative-going pulse squared by Schmitt trigger IC_{112} . The resulting pulse, of about 500ns duration, is applied to a D-input of flip-flop IC_{103} . On the next positive clock transition the output changes state and gates the oscillator, terminating oscillation in the high state. Since there is now no clock to reverse the state of the flip-flop, the trailing edge of the gating pulse is differentiated by C_{104}/R_{108} and used to reset IC_{103} (pin 6) thereby re-enabling the oscillator. After the delay determined by the LC time constant the clock restarts.



The resulting clock waveform is phase-locked to line with an integral number of clock cycles and a short gap when the clock is always high.

DIGITAL SYNC SEPARATOR

In the previous article I described the field sync separator on the analogue board. This waveform, designated FS1 in Fig.9, is useful when non-standard video sources are used: but there is an unavoidable tendency to vertical (plus or minus one line) jitter under some circumstances. An alternative sync separator on the control board derives a field sync waveform (FS2) digitally from the composite (line) sync waveform, and is completely stable. However, this sync separator can normally be used only with standard (CCIR) sync sources. Therefore either field sync waveform can be selected as required by Link 1.

Monostable IC_{101} produces a pulse which follows the trailing edge of line sync (Fig.16). For correct operation this pulse must be in the range 4.32 μ s: it is also used to provide a horizontal delay for positioning the display window. The pulse is fed to the D-input of flip-flop IC_{102} (2) which is clocked on the negative edge of line sync. In this way the wide pulses during the field sync interval are detected.

ODD/EVEN FIELD DETECTOR

When 512x512 (interlaced) mode is selected it is necessary to identify the order of fields, since the stored frame consists of two. Figure 16 shows that the line sync waveform toggles flip-flop IC_{102} which is

reset during the field sync interval. The output at pin 9 is applied to the D-input of flip-flop IC_{114} (2). The trailing edge of FS2 then clocks IC_{114} (3). The duration of FS2 is critical: it must be $148 \pm 16 \mu$ s in order to detect the odd number of lines between successive fields. These lines are identified by the output IC_{102} (9). Duration of FS2 is set by potentiometer P_{102} and the field identity waveform is available at IC_{114} (pin5).

DUAL IMAGE MODE

It is possible to update and display odd and even fields individually in 512x512 (interlaced) mode. This is useful for dual-image capture or for 3D display techniques and is achieved most economically by modifying

TABLE 1 Link options: all are two way shorting links.

	2	4	5	6	7	8	9	Notes
256x256								
Central screen repeat field	1	1	1	1				[1]
512x313								
Full screen repeat field	2	2	2	2	1	1	1	[2]
512x512								
Central screen interlaced	1	2	1	1	2	2	2	[3]

[1] 64K ram. Clock frequency = 5.8MHz approx. $C_{106}=220pF$

[2] 256K ram. Clock frequency = 9MHz approx. $C_{106}=0$. Connect IC_{111} (10) to IC_{115} (3,4,5,6), not V_{cc} .

[3] 256K ram. Clock frequency = 11.6MHz approx. $C_{106}=0$

Link 1=2 for non-CCIR interlaced-sync video sources
Link 3=1 for increased address hold-time after write (choose to suit ram-chip specification)

the length of the FS2 monostable pulse by $\pm 32 \mu$ s (tolerance remains $\pm 16 \mu$ s). The odd/even waveform then remains in a defined logic state to select permanently one or other of the images. The resulting three modes (odd field select, even field select and interlace select) are determined by switch S_2 .

Odd/even field identification is available only in 512x512 interlaced mode. For non-interlaced 512x512 mode, IC_{114} (5) is configured as the high order bit of a modulo-512 address counter by links 7,8 and 9 in position 1.

In 256x256 mode, field-detect circuitry is redundant unless required by the host computer for synchronization. Field-sync or field-ident is then available to the host via Link 10.

ADDRESS COUNTERS

The major job of the control board is to supply addresses and control signals to the memory. Two counters are employed for this purpose, one to count clock pulses along the line (the horizontal counter), the other to count lines down the screen (the line counter).

Line sync is always the initiator of a horizontal counting cycle (Fig.17), but it is desirable to be able to position the display window horizontally according to the display format and aspect ratio required. To achieve this, monostable IC_{101} (13) delays the start of the counting cycle by an amount determined by potentiometer P_{101} .

This start pulse triggers the toggle IC_{104} (3) via the nor gate IC_{110} , the toggle having been reset previously by the last line sync pulse at IC_{104} (1). Clock counters $IC_{104-105-106}$ are held in the clear position by IC_{104} until this point and are now enabled. For minimum propagation delay IC_{104} is a 74F74 and it divides the clock by two prior to the address counter IC_{105}/IC_{106} . This is because the memory is configured as 16 bits and two samples are stored per location; more detail will be given later.

Link 3 and IC_{109} provide a selectable delay element concerned with memory write timing. For $IC_{105-106}$, 74F series devices are used: these are superior to the LS series in that each output is individually buffered and can drive the capacitive load presented by the memory without malfunction. Propagation delay is also considerably reduced, allowing higher sampling rates with slower memories.

Fig.15 (right): waveforms in the phase-locked clock generator on the control board.

Fig.16 (middle right): digital field sync timing and odd/even field detection.

Fig.17 (bottom): timing for the horizontal counter.

Once a count of 128 or 256 (depending on Link 4) is reached, IC₁₀₄(3) is triggered again and the horizontal display cycle terminated until re-started by the next line-sync pulse. The counters are inhibited during the field.

The line counter works in an almost identical fashion to the horizontal counter except that the mechanism for positioning the display window vertically is necessarily different (Fig.18).

Line syncs are applied via IC₁₁₀ to the binary counter IC₁₁₅ which is held in the cleared state by field sync. When field sync ends, the counter increments until the inputs of the gate IC₁₁₁(9,10) are both high (after three lines or 35 lines depending on Link 5), at which time the counter input is inhibited by IC₁₁₀ pin 6. The inhibit output IC₁₁₁(8) then starts the line counter IC₁₁₆ via IC₁₁₄(8) in a similar manner to the horizontal counter. The terminal count is either 256 or 512 depending on Link 6 and the configuration of IC₁₁₄.

BLANKING GENERATOR

Blanking is derived from the address counter clear waveforms (Fig.19). The start of horizontal blanking is initiated by the first positive address transition at IC₁₀₅(12) clocking a logic 1 through flip-flop IC₁₁₃(3). This delay is necessary because the first few memory locations contain invalid data. Horizontal blanking generator IC₁₁₃(1) is reset at the end of the horizontal counting cycle. The horizontal blanking output IC₁₁₃(5) is then retimed by IC₁₁₃(12) and held in the cleared state by the vertical display window waveform at IC₁₁₃(13). Composite blanking appears at IC₁₁₃(9) and is combined with the c.p.u. blanking signal in IC₁₁₂(3) before being sent to the analogue board.

MEMORY CONTROL

The frame store has two operational modes: memory write (continuous display update) and memory read (display freeze).

Switching from one to the other must occur on a field sync transition (Fig.20). This is ensured by D-type flip-flop IC₁₀₃(8,9) which retimes the freeze command from the c.p.u. (if connected) or the freeze switch. When this signal is low the memory is continually updated and live video is seen on the screen: when high, memory is read continually and stored video is seen on the screen.

Memory control waveforms are derived from the clock divider IC₁₀₄(8,9) in conjunction with gates IC₁₁₁(3,6,11), and will be

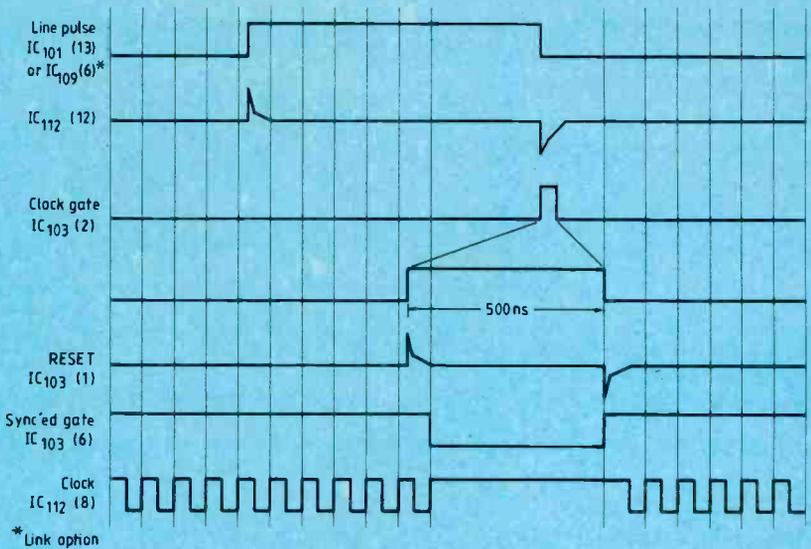


Fig.15

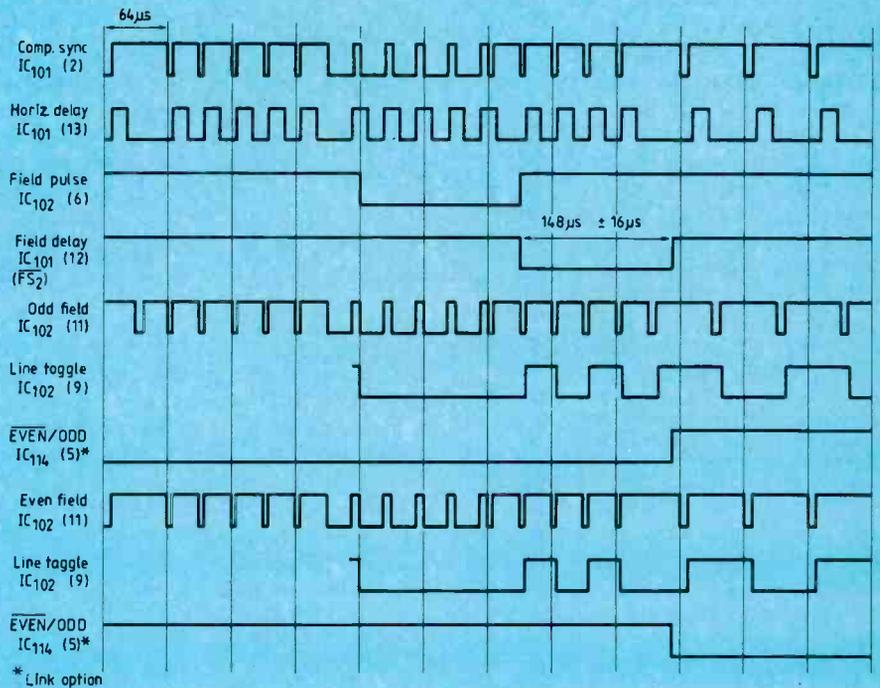


Fig.16

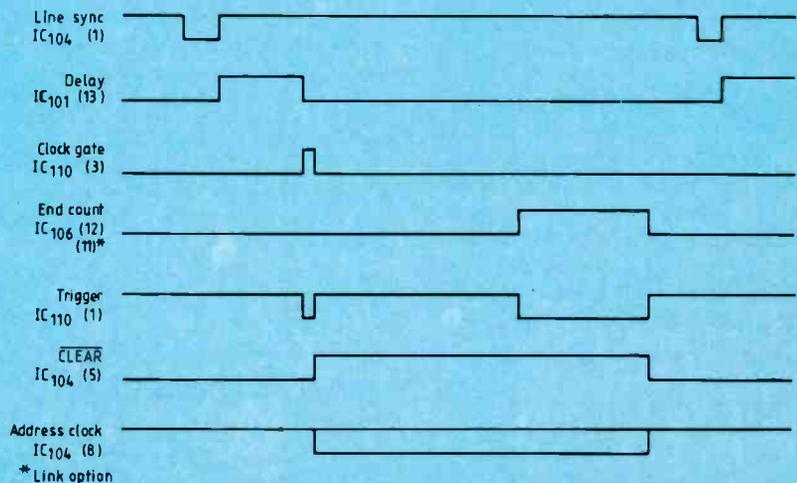


Fig.17

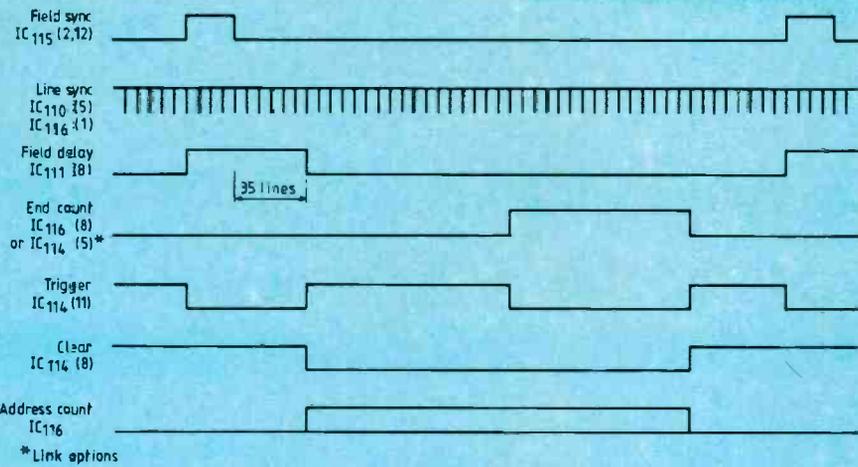


Fig.18

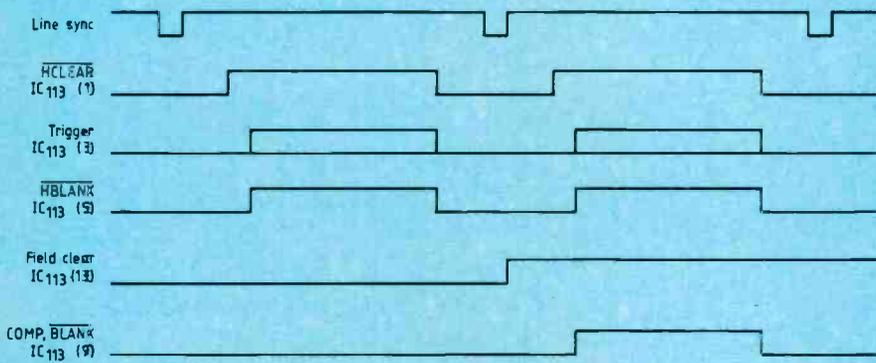


Fig.19

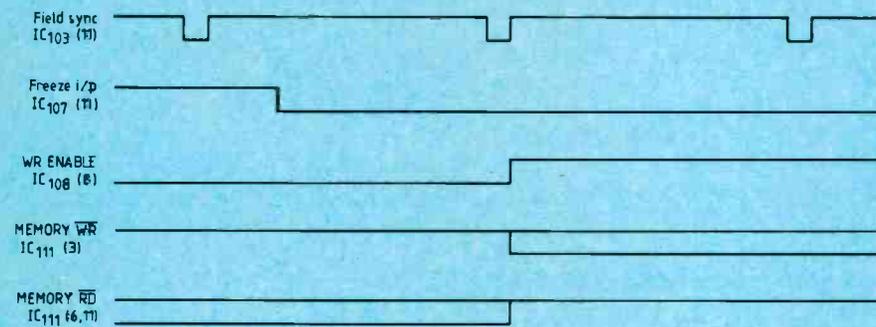


Fig.20

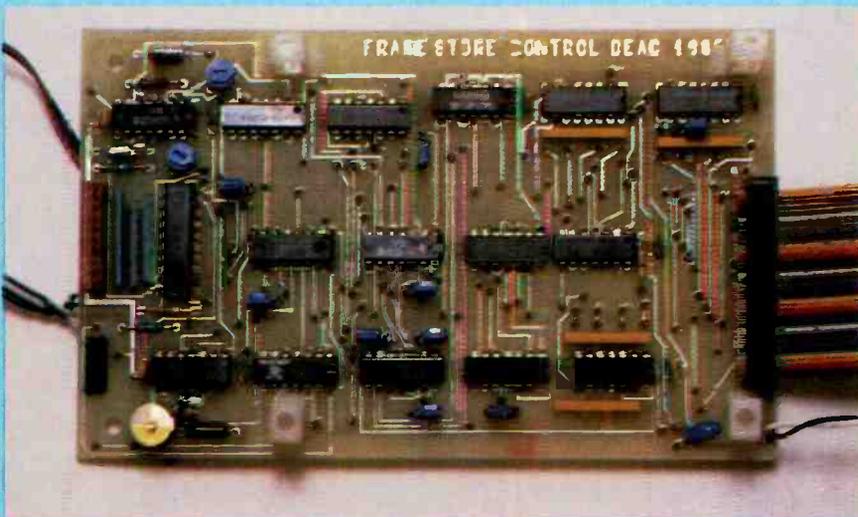


Fig.18 (left): timing for the line counter on the control board.

Fig.19 (middle left): generation of blanking waveforms.

Fig.20 (below, left): memory read-write mode timing.

discussed in detail in the next article. Waveform LE at IC₁₁₂(6) is used to tri-state the a.d.c. latch on the analogue board whenever data is read from the memory or c.p.u.

COMPUTER INTERFACE

The computer interface is the simplest possible, consisting of six control inputs applied via the buffer IC₁₀₇. Pull-up resistors ensure the correct operating state when the c.p.u. is disconnected, while limiting resistors prevent latch up. The use of a 74HCT series device results in simpler driving requirements because of its low input current.

The c.p.u. control input IC₁₀₇ pin 13 blanks the screen and selects c.p.u. access via the data selector IC₁₀₈. Complete control of the store is then gained by emulating the clock, line-sync, field-sync and mode inputs to increment and reset the address counters and select update or freeze mode.

Computer interfacing techniques will be discussed in a later article.

MEMORY BOARD

Next we turn our attention to the memory board and power supply. These two items are all that are now required to get the basic frame store up and running.

Display resolution of the basic unit is 256×256 with 256 grey levels. This translates to a memory requirement of 64K bytes. Until recently, dynamic ram chips would have been the only option for this amount of storage because the cost of static ram would have been prohibitive: but 64K static rams are now relatively cheap, and although more expensive, bit-for-bit, than dynamic, their advantages outweigh their extra cost.

The advantages of static ram in this application are:

- lower total power consumption because refresh is not required.
- reduced cycle time (equal to access time) compared with dynamic equivalents.
- minimal control circuitry and greatly simplified computer interfacing through the elimination of refresh.

Their increased board area occupancy is more than offset by reduced circuit complexity and chip count.

MEMORY ORGANIZATION

The 256×256 display format results in a pixel rate of 5.8MHz, assuming 44μs active horizontal display period and 11.6MHz for

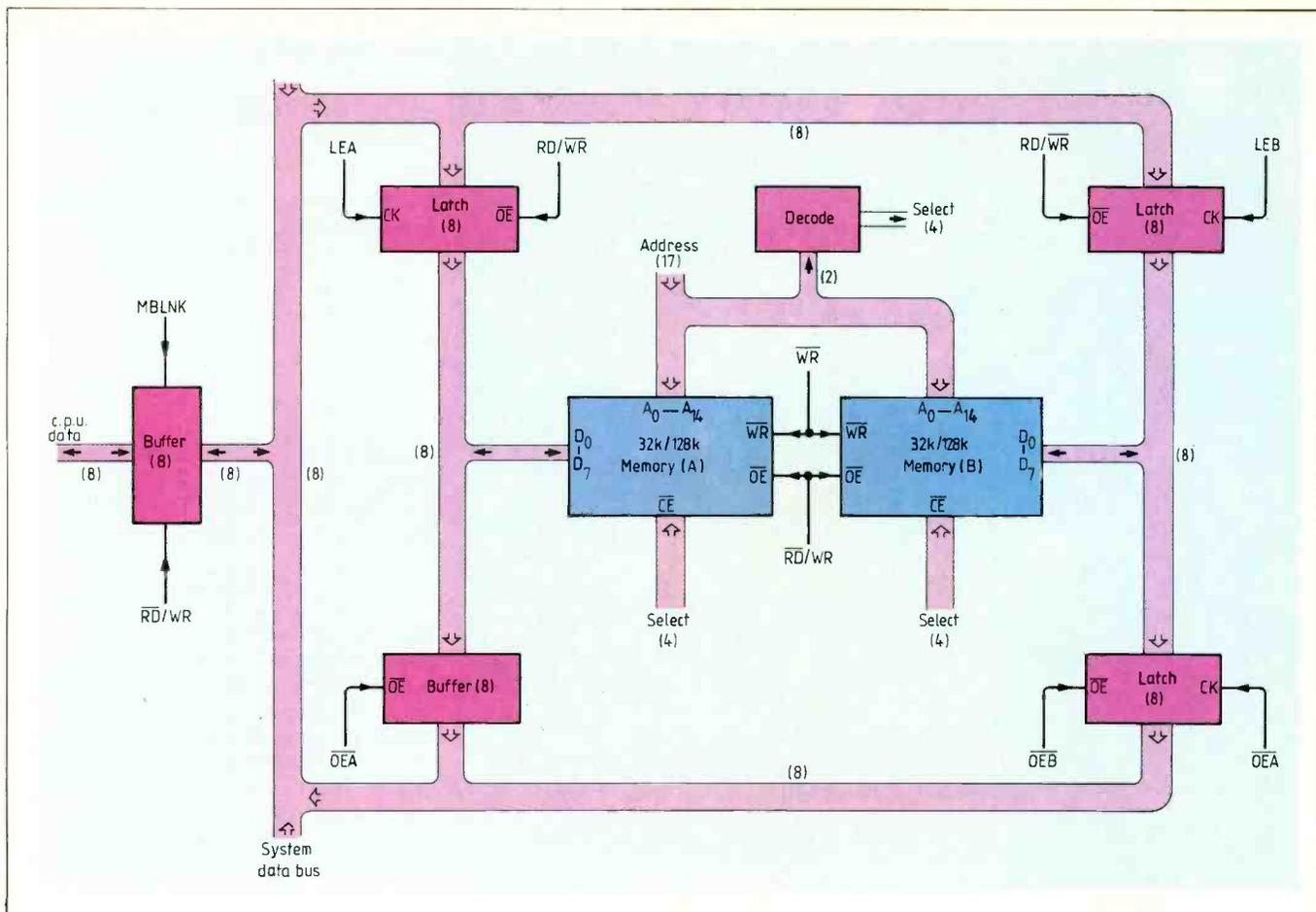


Fig.21. Outline of the memory board. A choice of ram chips may be used, for single-field or full-frame storage.

the 512x512 option. The latter is equivalent to an 86ns memory cycle. Using sub-86ns rams would be very expensive and wasteful, since by storing two samples/pixels per memory location (by organizing the memory as 16 bits per address location), the memory cycle-time can be doubled to 172ns and cheap 150ns rams used. A minor penalty is the requirement for latches to hold data temporarily as it is written to and read from the memory. The overall arrangement is shown in the block diagram. Fig.21.

The ram is split into two blocks, A and B, but only the data i/o pins are separate: all other inputs, address, $\overline{RD/\overline{WR}}$ and chip enable are commoned, so that data is read and written simultaneously as a 16-bit quantity. Latches are shown for holding input data during write-mode and a latch-buffer for holding-steering output data in read-mode.

There are four 16-bit memory banks (eight ram chips in total). The two high-order address bits from the line counter are decoded to select the appropriate bank.

To enable an external computer to access the memory, a bi-directional buffer is also on this board.

MEMORY CIRCUITRY

The memory circuit is shown in Fig.22. Links are provided to configure the board for eight 64K or two 256K devices (256x256 display) or eight 256K devices (512x512 display).

Address decoding (IC_{208}) takes place on line boundaries, when access time is least

critical. Access to the c.p.u. is via IC_{207} which is a 74HCT245 with pull-up resistors. The series resistors are to prevent latch-up and should be shorted when driving non-mos ports etc. on the host computer. Output data from block A does not require latching and an 8-bit buffer IC_{201} suffices.

MEMORY OPERATION

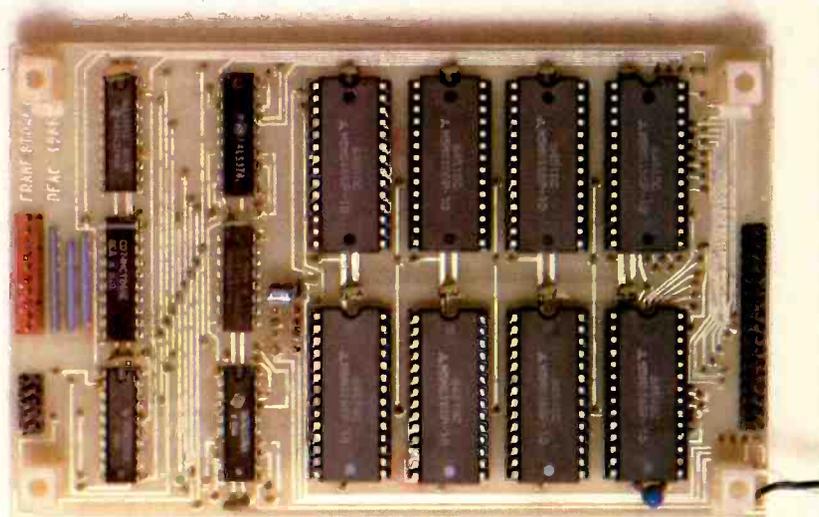
The memory operates in two distinct modes (controlled by the freeze signal), read-mode

TABLE 2: memory board link options

Resolution	Ram devices	Link 1	Link 2	Link 3
256x256	8x6264	1	1	1
256x256	2x62256	3	2	2
512x512	8x62256	2	1	1

and write-mode, $\overline{RD/\overline{WR}}$ determines which mode is active. During memory write, live digitized video is displayed on the screen.

To be continued



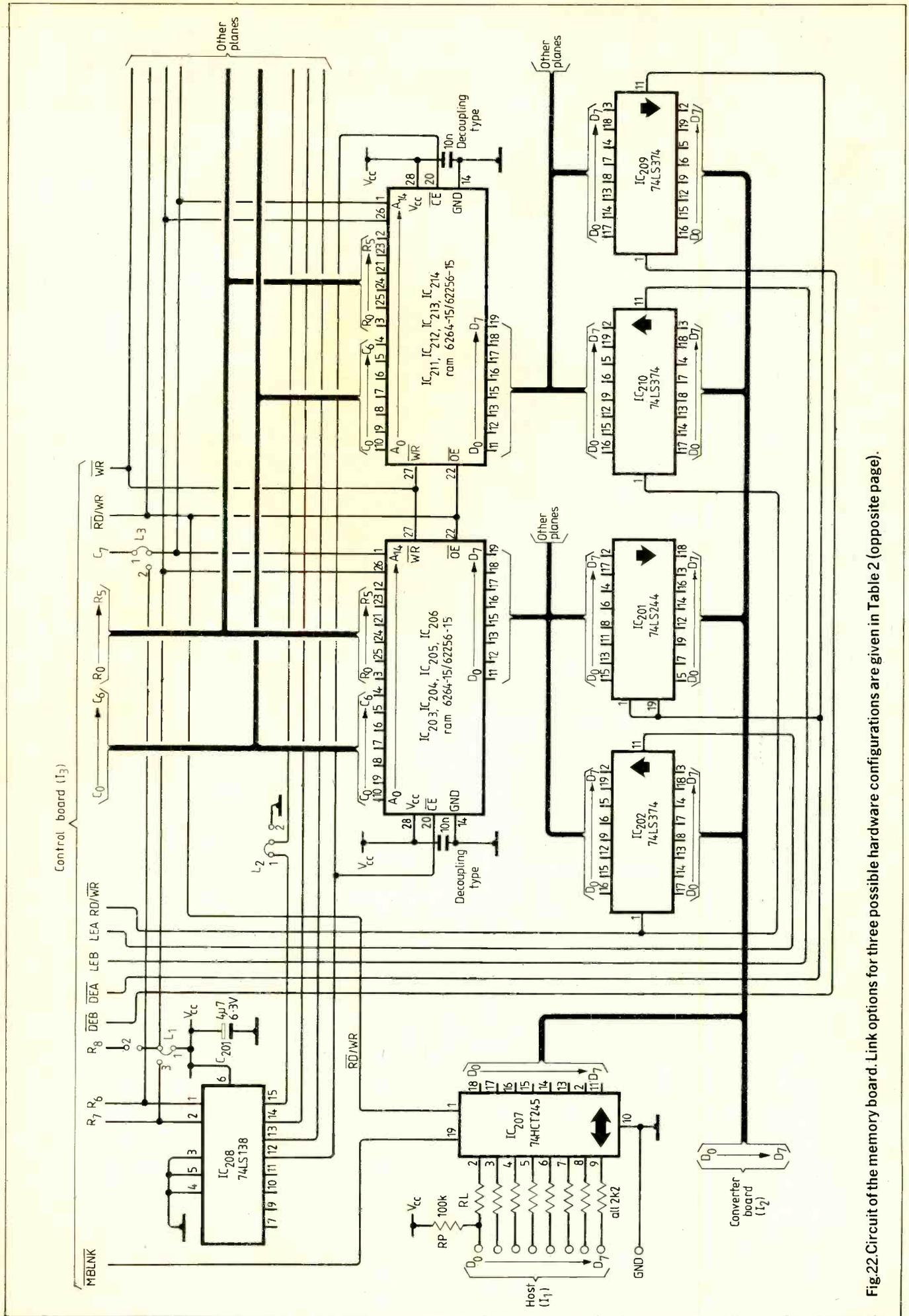


Fig.22.Circuit of the memory board. Link options for three possible hardware configurations are given in Table 2 (opposite page).

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21 Hanley Road, Shirley Southampton SO1 5AP Tel: 0703 780084



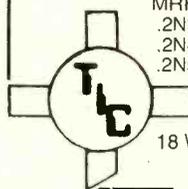
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| .2N4128 | .2N5070 | .2N5591 | .2N6080 | .2N4427 | .2N5090 |
| .2N5634 | .2N6083 | .2N4431 | .2N5102 | .2N5918 | .2N6084 |



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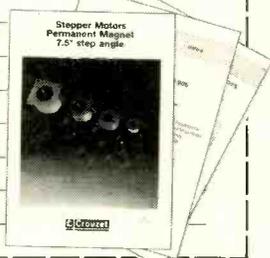
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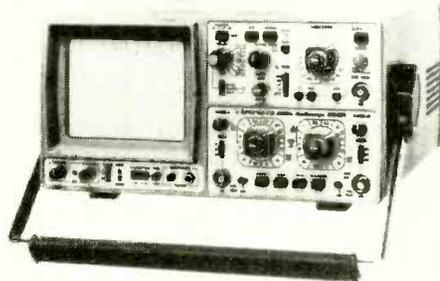
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<0.2% 50Hz-50kHz. TG152DM has an output meter.

LEVELL RC OSCILLATORS TG200D/DMP £139/175
1Hz-1MHz. 12 ranges, acc 1.5%+0.01Hz to 100kHz, 2%
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<0.05% 50Hz-15kHz. Sync output >1V.
TG200DMP has output meter and fine frequency control.

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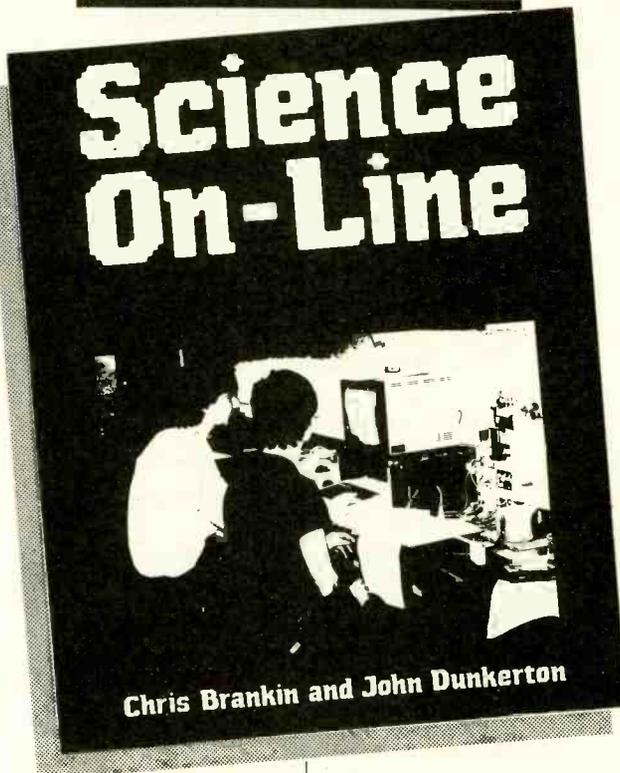
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BOOKS

A practical introduction to the new logic symbols, second edition, by Ian Kampel. Butterworths, hard covers, £11.50. User-friendly guide to the international symbols now appearing in manufacturers' literature and other documentation. There is no way of avoiding the new symbols, warns the author – and there is no wholly painless way of picking them up, since there is no one-to-one correspondence between the old symbols and the new. The new system, he says, is more a language than a mere set of shapes. But he softens the blow by observing that the symbols offer superb facilities for specifying designs at high level without the need to consider the exact form of implementation. The book is based on the formal and somewhat inaccessible IEC specification, with the IEC's collaboration. For a sample of Ian Kampel's approach, see his articles in E&WW in April and May 1985.

Computer organization: hardware/software, second edition, by George W. Gosline (professor of computer science at Virginia Polytechnic and State University). Prentice/Hall International, 623 pages, soft covers, £14.95. Textbook for computer science courses, extensively recast (according to the author's preface) to take account of developments in computer architecture, language and operating systems since the first edition six years ago. Some familiarity with assembly language, and a knowledge of a procedure-level language such as Fortran, Cobol or PL/1 is assumed. Section headings include the control unit; memories; input/output, data paths and interrupts; the Von Neumann computer; multiprocessors and multicomputers; special purpose systems (pipeline machines, parallel processors etc.); and computer networks. There is a 10-page bibliography.

Radio Database International, 1987 edition, edited by Lawrence Magne. International Broadcasting Services Ltd (P.O. Box 300, Penn's Park, PA 18943 USA), 351 pages, soft covers; available in Europe from Interbooks, Stanley, Perth, Scotland PH1 4QQ at £12.50 plus £1 inland postage (£6 airmail worldwide). Second



edition of this comprehensive and well-produced guide to the h.f. broadcasting bands. Coverage extends from 2.26 to 21.81MHz in the form of frequency-by-frequency computer-generated charts which show each station occupancy of its channel for every hour of the day. Graphic codes identify nine major languages at a glance. Information is derived partly from published schedules and official lists, though much comes from monitoring observations: further codes mark transmissions which are being jam-

med and times of fade-in and fade-out. Also included is a receiver buyer's guide and a number of feature articles.

Science on-line: eye-catching science experiments for the BBC micros, by Chris Brankin and John Dunkerton. Hodder and Stoughton, 72 pages, soft covers, £4.95. Attractive and encouraging guide for beginners to computer interfacing. Its aim is to enliven school science lessons by demonstrating real results on the screen: among the experiments it describes are a method

of recording signals from a toad's heartbeat and a discharging capacitor; oscillations from a vibrating metre rule; pH and conductivity meters and a colorimeter; an ingenious blood-flow monitor based on a slotted optocoupler fastened to the ear-lobe; and investigations of magnetic fields using a Hall-effect probe. There are many suggestions for further work. To encourage the faint-hearted, an early chapter introduces the black art of soldering. Authors are respectively head of Physics at Bartholomew School, Eynsham, Oxford and head of VIth Form biology at Trinity School, Leamington Spa.

X.25 Explained: protocols for packet switching networks, second edition, by R.J. Deasington (IBM UK Laboratories). Ellis Horwood (Halsted Press, John Wiley & Sons), 131 pages, hard covers, £16.50. Introduction of the bottom four layers of the seven-layer Open Systems Interconnection model, the system standardized by the ISO for enabling computers from different manufacturers to be interconnected.

Chapters deal with physical connections (X-21 and X.21 bis, alias RS232C); the link level; network level; transport level; and 'triple X', the three protocols defined by the CCITT to connect character mode terminals to packet-switched networks using X.25.

SOFTWARE

Commstar II: rom-based intelligent communications facility for BBC Model B, B+ and Master series computers. Pace Micro Technology, Allerton Road, Bradford BD15 7AG, tel. 0274-488211; price £29.57 excluding v.a.t. New version in a 16K rom provides several new features and a fresh look to old ones. In the scrolling terminal mode (for electric mail and subscription databases), features are selected from drop-down menus and sub-menus; data parameters and

other settings are displayed on a status panel at the foot of the screen and may be saved to file and re-loaded. Data can be buffered in memory or on disc and there is an Xmodem file transfer capability. Viewdata mode has extensive facilities for tagging and recalling pages and for saving them in memory or on file. A useful editor for preparing mailbox frames off-line is included and an on-screen timer keeps track of telephone line-time in both viewdata and terminal mod-

es. Facilities are easy to use and are fairly comprehensive. A modem-driver routine for Pace's own modems is to be added.

Spaceheights Linear Circuit Analysis Program: Spaceheights Ltd (6 Prospect Place, Chapelhey, Weymouth, Dorset, DT4 8JY), £129. Versatile MSDOS program for Apricot Xi, PC, F, intended for use in education or the laboratory. Handbook includes many case studies with much practical information.

SATELLITE SYSTEMS

TOM IVALL

Observing the Earth

Scientific bodies around the world are now beginning to make use of the geographical, geophysical and agricultural data being transmitted by an earth-observation satellite called Spot-1. Launched in February 1986 on the last of the Ariane-1 rockets, this small scientific spacecraft is orbiting the earth with 101-minute period at an altitude of about 830km. It takes optical images, both visible and infra-red, of the earth's surface in 60-km wide swath as it travels round.

Spot is similar in broad principle to the Landsat series of earth observation satellites (Nos 1 to 5) which have been in operation over many years, but it has several performance advantages. It offers a higher spatial resolution, with picture-element sizes of 20m×20m for colour 10m×10m for black-and-white. This com-

pares with a 30m×30m pixel size from Landsat. In other words, Spot can detect objects or features as small as 10m across.

Secondly it will provide stereoscopic images where Landsat does not. Thirdly it will give fast-response repeat images of pre-determined areas. But each Spot scene is smaller, with an area of 60km×60km as against the 185km×185km scene area from Landsat. Compared with aerial photography, which offers scales in the range of about 1:2000 to 1:80,000, Spot has a scale range of 1:25,000 to 1:100,000.

This new observation satellite was developed as a commercial venture by the French Centre National d'Etudes Spatiales in collaboration with Belgium and Sweden. Matra of France built the spacecraft and provided the imaging instruments. There are two identical but independent optical systems, together with their electronic units. These use green, red and near infra-red

sections of the optical spectrum, plus a wavelength range of 0.51 to 0.73μm for panchromatic images. Radiometric resolution is 256 levels. The optical information is electronically digitized and formatted and transmitted to the ground by an X-band radio telemetry system, using quaternary phase-shift keying and a data rate of 50 Mbit/s.

On the ground the data signals are received by various stations, where they are decoded, analysed and processed to give photographic-type pictures. The Spot system's main receiving station is at Toulouse, France, while others are in Sweden, Belgium, Canada, Australia and New Zealand. The survey information obtained by the system is sold through a company, Spot Image, which was set up specifically for the purpose and already has customers in more than thirty countries.

Recently the Spot organization has awarded six contracts to the American firm International

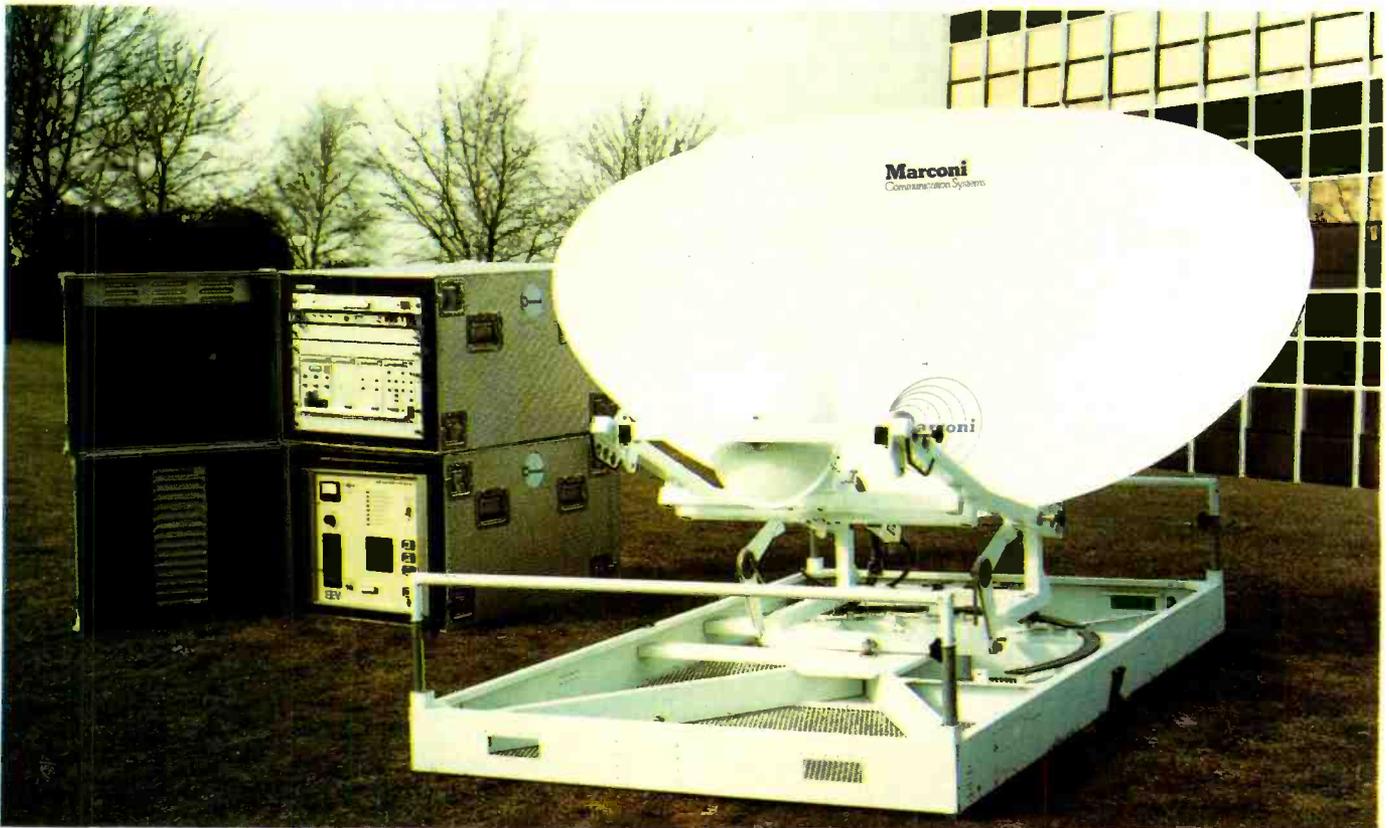
Imaging Systems to supply hardware, software and complete image processing installations to users of Spot data around the world. The equipment and software is based on the US firm's existing S600 image processing system. Customers for these products include China, Iraq and Italy. At the same time this firm is developing new software to enhance the observational data.

The satellite, which weighs 1800 kg on the ground, is expected to have a lifetime of three years. So a second one, Spot-2, is planned for launch by Ariane 2 or 3 in October 1988 and a third and fourth in the 1990s.

Mobile radio tests

Mark Nelson mentioned in the November 1986 issue (p.12) that u.h.f. bands below 1000 MHz were unlikely to be assigned for land mobile uses of satellite communications in North America.

This portable Ku-band uplink terminal has been used successfully by Independent Television News to send tv news reports from distance places back to the UK via Intelsat communications satellites. Made by Marconi Communication Systems, it consists of three units, each of which can be carried by two men from an aircraft or road vehicle. The 2m×1m elliptical antenna reflector swings down flat onto the carrier for transportation. Its offset-fed Gregorian arrangement complies with the 29–25 log θ sidelobe law for the current 2° spacing of comsats (where θ is the angle from the centre-line of the radiation pattern). Typical e.i.r.p. is 68 dBW. ITN has used this satellite news gathering (s.n.g.) terminal in Bali, Indonesia, to cover President Reagan's tour in the Far East, and in Burkino Faso, Central Africa, to report on the Sport Aid marathon, as well as for various news events in and around the British Isles.



SATELLITE SYSTEMS

This has, in fact, now been decreed in the USA. The Federal Communications Commission has rejected requests from twelve applicants, including the Canadian government, for 8 MHz of spectrum space for this type of service in the 806-902 MHz land mobile allocation of ITU Region 2.

As Mr Nelson pointed out, satellites would increase the coverage for land mobile radio and allow more users to be accommodated – a great advantage in the large-area countries such as USA or Canada. In addition, if the u.h.f. land mobile frequencies could be used with satellites it would be possible to have just a single set in the car or other vehicle to cover both the terrestrial (p.m.r. and cellular) and satellite services.

However, the FCC has authorized a spectrum of 27 MHz in the L-band 1.5-GHz downlink and 1.6-GHz uplink allocations for mobile radio systems based on satcoms. Unfortunately, these higher frequencies will require more expensive equipment to achieve r.f. power levels equivalent to those obtainable at u.h.f.

Meanwhile, as mentioned elsewhere in these pages, the possibility of land-mobile satellite systems is being explored through INMARSAT. Although the main function of this international organization is to offer maritime mobile communications by satellite it is now beginning to move into other areas. In 1985, for example, all the member countries agreed that the organization should also offer aeronautical satcom services.

On land INMARSAT has already been providing emergency and relief communications using land-based terminals for natural disasters like earthquakes and volcano eruptions. Now the organization will be participating in a pan-European test and demonstration of satellite-based land mobile communications planned to take place in late 1987. They say a low-density service should occupy only a small part of the communications capacity of their existing satellites and would have the advantage of an already developed transmission system. Use of existing coast earth stations would help to make such a scheme economical to operate. A mobile terminal

similar to the shoe-box size one also described in these pages should cost about the same as a 900-MHz cellular radiotelephone.

INMARSAT's decision to provide aeronautical services has also resulted in some L-band trials in aircraft being held in 1987. In the UK British Telecom International, British Airways and Racal-Decca Advanced Development will be collaborating on these. To begin with the trials will be done from Racal's Jetstream aircraft. Later on they will be extended to scheduled BA flights.

Air passengers will make radiotelephone calls by inserting a credit card into a specially adapted payphone. This will unlock the handset and connect the caller with a ground-based operator in the UK who will connect the call. To provide an earth terminal, BTI will be using one of the antennas at the Goonhilly, Cornwall, satellite station exclusively for the trial services. Initially the air passengers will be able to make calls but not receive them, though BTI think it will be able to offer all normal telecommunications services if there are enough customers wanting them.

Racal-Decca Advanced Development, in consultation with BA, are producing the airborne 1.6/1.5-GHz transceivers and developing specialized aircraft antennas for the scheme. This firm has already demonstrated an air-to-ground data communications link.

Multi-beam comsats

The Advanced Communications Technology Satellite (ACTS) mentioned briefly in the October issue (p.7) will be a precursor to the next generation of commercial communications satellites when it is launched in 1989. Operating with uplinks in the 29.5-31 GHz band and downlinks in the 19.7-21.2 GHz band, this NASA spacecraft is intended as a test-bed for future comsat systems using multiple spot beams.

The project grew out of a realization that satellite communications have become so popular that slots in the geostationary orbit and frequencies in the fixed satellite service allocations are rapidly being used up.

Conservation is obviously needed, in the three dimensions of space, time and frequency spectrum. The method to be tested in ACTS is a combination of time sharing by time division multiple access (t.d.m.a.), transponder sharing by frequency division multiple access (f.d.m.a.), space sharing by multiple spot beams directed at selected areas, and frequency re-use in the spot beams.

Two kinds of spot beam are to be used in ACTS. Fixed beams are directed at three large conurbations for high-volume communications traffic. Scanning beams are used to move across a large area such as a whole country and link with earth terminals using a low rate of digit bursts in the t.d.m.a. system. The scanning pattern is controlled by information held in a memory. Each spot beam covers an area of only about 240 km in diameter.

UK space plan

1987 should be a significant year for Britain's activities in space science and technology. The government is expected to announce how much money they will allocate to a co-ordinated programme for the whole country, covering all aspects. Since the autumn of 1986 they have been studying a National Space Plan put before them by the recently formed British National Space Centre (BNSC), which has headquarters in the Millbank Tower, London.

At the time of writing details of the plan are not yet available, but at a recent IEE meeting the director general of the BNSC, Roy Gibson, gave some idea of what is afoot. He said the BNSC had been set up in November 1985 as a focus for British space policy. Its purpose was to co-ordinate the various activities that had previously gone on separately in national scientific and other organizations and so make the total UK space effort more effective and economically efficient.

In the past individual organizations had looked at particular international space projects and decided on their own whether or not to take part. But now circumstances had changed. Space projects for different purposes were becoming much more interrelated. It was

no longer appropriate to make decisions in separate compartments. A national body with an overall view was needed to produce a long-term space strategy and to co-ordinate all future activities.

At present the BNSC has no independent funding. It is run by a staff mainly from four existing government bodies: DTI, the SERC, the NERE (National Environmental Research Council) and the MoD. There are about 40 people in the London headquarters. A technology arm is based on the Royal Aircraft Establishment, Farnborough, and the Rutherford Appleton Laboratory, Chilbolton, and this provides the equivalent of about 240 full-time staff. Other government organizations involved are RSRE and the Met Office.

British industry is taking part as well. In addition to the few big firms like British Aerospace and Marconi who are well known for their space products, there are about a hundred UK manufacturers with interests in this field. Mr Gibson said it would be a good idea if they could form a trade association, outside the existing UK Industrial Space Committee, that could speak with a single voice and help them overcome difficulties in selling their products to international bodies like the European Space Agency.

One UK company has already supplied equipment to the BNSC itself. Ferranti Microwave has custom-designed and manufactured a satellite communications earth terminal, which has been set up at RSRE's Defford establishment in Worcestershire. To begin with it will be used in an experimental programme with Ku-band satellites and ESA's Olympus-1 multi-purpose satellite due to be launched in 1988.

The terminal is based on a 5.6m dish aerial. Initial operation will be in the 12/14-GHz band and the system is adaptable to allow substitution of an 11/13-GHz and ortho-mode transducer (for orthogonally polarized signals). The antenna pointing system is compatible with the tracking requirements of the Intelsat standard E earth station and the Eutelsat ESC multi-service system specifications. The up/down converters use low-noise phase-locked oscillators and the receiving chain has a low-noise fet amplifier.

TELECOMMS TOPICS

Cable and Wireless in KDD2 consortium

Cable and Wireless and Japanese trading company C. Itoh are the leading participants in a consortium to provide an alternative international telecommunications system for Japan. Formed initially to carry out a feasibility study, it is the precursor to an operational company. The new company, with two-thirds Japanese and one-third western shareholding, will be known as Kokusai Digital Tsushin Kikaku (KDTK) – or more simply as KDD2 (because KDD is the existing international carrier).

The two major partners each hold 20%. Then, Toyota and Pacific Telesis, one of the Regional Bell Operating Companies i.e. telephone companies, (subject to US Court and Governmental approval) have 10% each and the remaining shareholders individually holding far smaller stakes.

The outline business plan, worked out in a previous C. Itoh and C & W study, will commence in 1987 with leased circuit services via Intelsat and the existing cables. Switched services will then start in 1989 to be followed by, beginning in 1990, a full range of digital leased and switched services via PPAC trans-Pacific submarine optical fibre cable jointly owned by KDTK (Japan) and Pacific Telecom Cable of the USA. Subsequently, connection of Japan with Hong Kong and Korea, Taiwan, the Philippines, Australia, China and other Pacific Basin countries by a further network of submarine cables or satellites.

This consortium provides reinforcement for C & W's plans for a Global Digital Highway "to provide the world with a choice of international communication facilities for the first time in history." It will concentrate on new services with the aim being to provide digital transparency so as to meet user needs and be able to support A and mu laws with 64 and 56kbit/s information channels that are in accordance with CCITT and Bell standards respectively.

It is expected that the Japanese government authorities will award the relevant licence next year. Even though there are at present two contenders for the license, KDD2 is confident of being the winner in view of its great technical strength as compared with that of its competition.

The investment by the consortium is expected to be hundreds of millions of dollars and "could well reach \$1bn eventually". KDD's international market is currently worth \$1bn and growing at 10% p.a. overall, but facsimile has a growth rate of around 35% p.a. according to Jonathon Solomon, C & W's director of corporate strategy. Sir Eric Sharp, C & W's chairman, said that Japan is under-developed in international communications and the business could be worth at least \$3bn in 5 to 10 years.

50,000 Vodafone subscribers

The Racal-Vodafone service, one of the two nationwide cellular radio services has just announced that it has connected its 50,000th subscriber.

According to Gerry Whent, chairman of Racal Telecommunications Group.: "The achievement of connecting 50,000 subscribers to the Vodafone network in less than two years strengthens greatly the profit forecasts for cellular radio made in January 1986. Our previous estimate of 60,000 subscribers by March 31, 1987 will now be well exceeded." The network, with 200 base stations servicing 400 cells, now covers some 82% of the UK population and is being expanded continually.

BT contract for Hewlett-Packard

Hewlett-Packard has won a multi-million pound contract for the supply of specialized remote access and test (Rates) equipment to British Telecom. This represents the third part of a four year programme worth more than £15 million in total.

The Rates equipment, designed and developed at H-P's Queensferry Telecommunications Division in Scotland, improves the maintenance of private circuit services by greatly reducing time spent in fault diagnosis. The system is, therefore, particularly suited to BT's private lines in Britain, which is expected to cover more than 2000 small and large exchanges nationwide by 1987.

BT dealer system for Japan

Teletrade, the export unit of British Telecom's International Products Division, and the Japanese trading company Mitsui & Co Ltd have signed an agreement to distribute the City Business System in Japan. Mitsui's specialist telecommunications subsidiary, Adamnet, will handle sales, installation and support.

CBS, a touch-screen dealing board for the financial community, has been specially developed for the Japanese market, including the use of Kanji characters for the v.d.us. The introduction of Kanji was seen as a necessary development to enable CBS to be sold successfully to Japanese financial institutions. Its standard version is already being used by the Japanese offices of international companies.

France/UK videoconferencing link

The first commercial videoconferencing link between the UK and France has been inaugurated between British Telecom International and Direction General de Telecommunications, the French PTT. The connection to France consists of a 2Mbit/s link via the French satellite, Telecom 1. The BTI service is due to be expanded this autumn, when a link to the Netherlands will be opened.

The service is available to users in the UK who have their own videoconferencing facilities or through BT's public conferencing rooms.

Plessey markets System X via Intelsat

Plessey Major Systems staged a live System X demonstration at the Plovdiv Trade Fair, Bulgaria as part of its drive to be selected as that country's public switch supplier. A Remote Concentrator Unit (RCU) at the show was operated, via Intelsat, under the control of a host exchange located at Plessey's Edge Lane, Liverpool factory.

The initial order for the Bulgarian network could be worth some £50million. Currently consisting of about a million lines, it is expected that the network will be expanded to some three million by the early 1990s. In view of the size of its network, it is expected that Bulgaria is looking for a single system rather than the two or more competing systems adopted by larger administrations. If System X is selected, the fact that it is offered by both Plessey and GEC would provide the customer with the advantage of competitive purchasing. However, because it would be subject to an export license, this is a long term prospect.

The ISDN capability was demonstrated to the State Trading Companies, PTTs, as well as Telecommunications Institutes using fast digital facsimile and slow scan tv equipment. User facilities also shown to visitors included call diversion, automatic reminders auto advice of call duration and other automatic or subscriber programmed services.

In addition to providing Plessey with a powerful international marketing tool, which will be used at exhibitions in Beijing, Indonesia and Geneva etc., these demonstrations also indicate the sophistication now practical in international telecommunications.

The link with Plovdiv was the first public demonstration of the use of CCITT No. 7 common channel signalling system via a communications satellite. The RCU in Plovdiv was connected, via a portable satellite terminal (provided by British Telecom International) and an Intelsat satellite, to a similar ground

TELECOMMS TOPICS

station at the Plessey factory where it linked into the System X host exchange. From Liverpool, the link proceeded over BT's digital trunk network to public System X exchanges in Manchester and London.

Plessey System X international Sales Director, Sidney Reeves, said "When a telecommunications administration is considering the substantial investment required to upgrade its telecommunications network it wants to make absolutely certain that it is choosing the right switch. A practical demonstration is a very powerful sales aid but to achieve a demonstration of a full working public telephone switching system at a trade fair has always presented a problem."

CGS videotex update

Cap Gemini Sogeti, the largest software company in Europe which was lead contractor for the French electronic telephone directory project, aims to expand into the UK videotex market.

CGS has introduced its Multitel product for IBM computers and already has a pilot project under way with General Motors and its dealer network. Multitel allows organizations in countries operating under different standards and in different environments to take advantage of the business potential of videotex. This applies whether the standard adopted is Prestel, or Teletel operating under the overall umbrella of CEPT protocols.

According to Jeff England, CGS UK General Manager, the company anticipates 5% of its revenues will derive from videotex installations by 1990. This, in conjunction with software conversion products, which eliminate the costs of re-writing software for videotex access, will become a market leader. He claims that the contribution made by CGS to cost-effective videotex applications lies in more than the software. Advice is given on aspects such as how to extend the number of terminals inexpensively, with easy connections, and how to reduce the operating cost of managing the network.

At the same time as this

announcement was made Meinhard Donker de Marillac, director of communications for CGS Group Europe, reported that by the end of 1986 it is expected that there will be 2.5 million Minitel terminals installed in France and that, in last May/June there were already over 20 million calls made with over 2.3 million connect hours per month.

ISDN chips and kits introduced

Both Intel and Siemens have introduced chips and kits for ISDN subscriber functions. Intel announced its 29C53 digital-loop controller together with support tools including terminal and line card evaluation kits. As well as this, the company has arranged availability of third party software based on the 80188 microprocessor. This software handles CCITT Layers 2 and 3 and will support multiple connections over a single communications controller. It organizes, under microprocessor control, the 144kbit/s aggregate of the 2B plus D channels.

Sample 29C53 chips are now available and "the unit price, in quantities of 10,000, is below \$15 in 1987". The 29C53 transfers data at the ISDN basic access rate of 192kbit/s using three channels - two 64kbit/s "B" voice/data channels plus the 16kbit/s signalling "D" channel - together with 48kbit/s (invisible to the user) for framing and synchronisation.

Intel's Subscriber-Line Data-link (SLD) architecture is the foundation of the company's family of Advanced Telecommunications Components (iATC). It provides a common standard for Intel's telecommunications system, such as Intel's Multibus, is a standard for microcomputer-based systems and is endorsed by a number of semiconductor manufacturers. According to Intel, SLD offers compatibility with current as well as future iATC products and is claimed to be flexible enough to be used in both centralized and distributed PBXs and central offices.

The Intel philosophy is to provide complete solutions that encompass the accepted standards

thereby freeing the o.e.m. to focus his resources on differentiating the end-product.

Siemens is also offering its STU 2000 kit made up of three boards; two of these are dedicated to subscriber functions and the third represents the switching system. It can be used to test all the ISDN-related functions in an environment approaching that of a "real" system. Siemens claims that, by using the kit, programming can be brought forward by six months.

Telephone socket liberalization

Following wide consultation between the Office of Telecommunications (OfTel), relevant trade associations, user bodies, and public telecommunications operators (PTOs), OfTel has announced that from last December 1st (1986) telephone users and independent contractors will be permitted to install extension telephone sockets and their cabling, and plug them into the public telephone networks.

Users with the new-style master sockets, installed by a PTO, will be able to take advantage of the new arrangements. The competitive supply and installation of the master socket itself, through which telephones are linked to the exchange line, is still under consideration.

Consequently, users who do not yet have the new-style sockets installed in their home or office must ask the PTO to convert their installations by fitting a master socket before they can go on to install and connect further sockets of their own. (The British Telecom charge for

converting an existing installation is currently £25 plus VAT.)

In due course, it is expected that the PTOs will start to offer a new pattern of master socket that incorporates provision for direct, rather than plug-in connection wiring. At the moment, however, the only method available is via a plug-in connection. This uses a length of approved cable which can be cut to length as required. One end of the cable will probably be ready attached either to a simple plug that will fit a new-style telephone socket or to a plug-in adapter to enable a telephone to continue to be plugged into the existing socket.

Mr Peter McCarthy-Ward, marketing manager in British Telecom's Customer Premises Equipment division has stated that: "We welcome this change which will stimulate demand for internal extension points and extension telephones." This is understandable as, after taking paperwork into account, it must be difficult to break even, let alone make a profit on this wiring.

Access to Horizon

Birmingham-based Horizon Holidays has signed a contract worth £1 million for the use of Istel's Infotrac network. This value added network (VAN) is already used by eight of ten of the top tour operators. It will give travel agents around the country local call access to its viewdata reservations computer. It will augment Horizons existing dedicated network and, because Infotrac has 65 nodes strategically sited throughout the UK, it almost doubles the number of travel agents who can access Horizon's viewdata booking system. Horizon has also doubled the speed of its processors which trebles the number of calls that can be handled simultaneously.

These factors, combined with the speed and efficiency of the Infotrac network, are claimed to cut down the average time taken to make an on-line booking from six to four minutes. In addition to reducing the cost of the agents' phone calls, it should reduce the difficulties in getting through to the reservation system, even at peak periods.



CIRCUIT IDEAS

Increased range for the 8910 sound chip

Although the AY-3-8910 is specified to run at up to 2MHz, it seems to work well at 4MHz. The disadvantage is that in most cases, output frequency is clock frequency divided by 16 to 65536 so a 4MHz clock produces a range of approximately 61Hz to 250kHz. Most of this range is inaudible.

As shown here, a dual 4-stage binary counter gives a range of clock frequencies down to 15.625kHz; 1MHz now provides a range of 15Hz to 62.5kHz and at the lower frequencies, sound envelope attack/decay times are extended to minutes.

Two 8-bit i/o ports are built into the sound generator chip. Take D_{0-2} from one of the ports and use them to drive the three address inputs of a 74LS151 eight-channel multiplexer. Outputs of the counter feed the multiplexer data inputs.

Now, sending a value between zero and seven to the generator i/o port selects one of eight different clock frequencies.

Mel Saunders
Leicester

D.s.b.-s.c. detector with p.l.l.

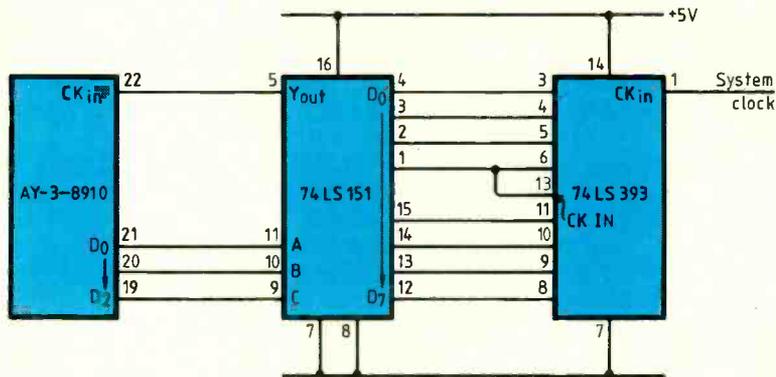
By adding a 'duty shaper' to the p.l.l. circuit shown on page 70 of the April 1986 issue, a signal synchronized with the suppressed carrier of a d.s.b.-s.c. signal could be generated. A duty shaper is a circuit which changes the 50% duty cycle of a sinusoidal signal to a new value (close to 25% in our case). Practically, its design depends on the carrier frequency.

I chose a circuit suitable for the 38kHz suppressed carrier found in stereo f.m. Its main purpose is to demonstrate that a simple RC active circuit could detect suppressed-carrier d.s.b. without the need for a reference such as a 19kHz pilot tone.

The LM324 op-amp acts as a class-C amplifier giving a train of pulses (but amplitude modulated). An LM311 comparator acts as a high-gain amplifier giving squared pulses with a duty cycle of around 25%, which is the optimum value to get the widest lock range as described in the April issue. Voltage-controlled-oscillator output, delayed if necessary by an RC network, could then be divided by two using a 4027 JK bistable device.

The analogue gate (4066) is used for synchronous detection. Finally, to recover the audio signal, gate output must be low-pass filtered.

Kerim Fahme
Autolight
Aleppo
Syria



Transformer design – a simple method

Following the excellent article by Baert* examining transformer design in detail, I felt that a simple evaluation method would be useful for the experimenters working with unknown laminations.

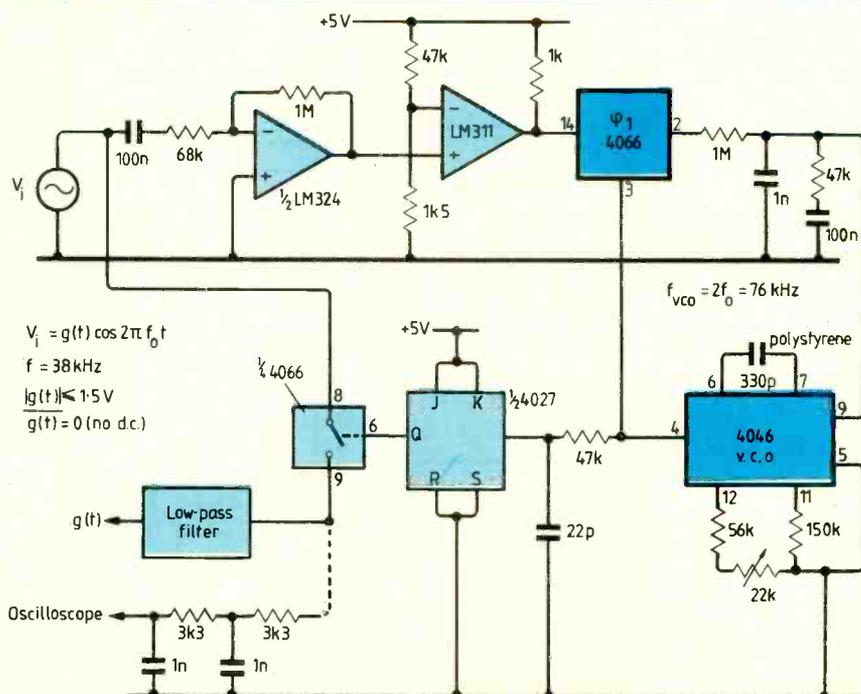
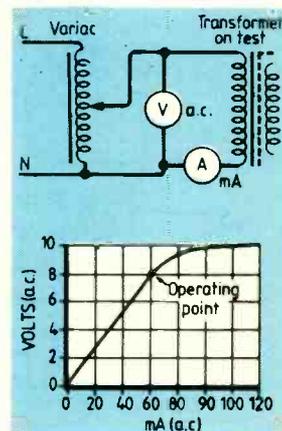
Wind 100 turns, or any convenient number, of insulated wire onto the former to be used. Insert the laminations and plot the magnetizing curve using the circuit shown. If a Variac is not available, then a tapped transformer or various power resistors may be used.

Draw a graph. As shown, the transformer should be operated just below the knee. From the graph, voltage at this point is read off. The number of turns used is divided by this figure to give the transformer's turns/volt rating, which will usually lie between five and ten turns/volt.

The winding wire gauge is found from wire tables using a figure of between 3 and 3.5A/mm² (1500 to 2000A/in²).

P.J. Dinning
Newcastle-upon-Tyne
Northumbria

* D.Baert, Designing small transformers, *Electronics & Wireless World*, Aug. 1985 pp17-19.



CIRCUIT IDEAS

One-to-four-way telephone exchange

By eliminating the possibility that more than one set can operate at once, this exchange produces four lines from one without overloading. Once one of the four telephone handsets has been lifted, the line will not be released until the handset is replaced in its cradle. It also ensures privacy.

Monitor leds in each holster indicate whether the line is available. The single-pole, switches on each set can be either

miniature microswitches or reed switches and magnets with the switch in the holster and the magnet on the handset.

Should a malfunction occur the system can be overridden so the line is always available.

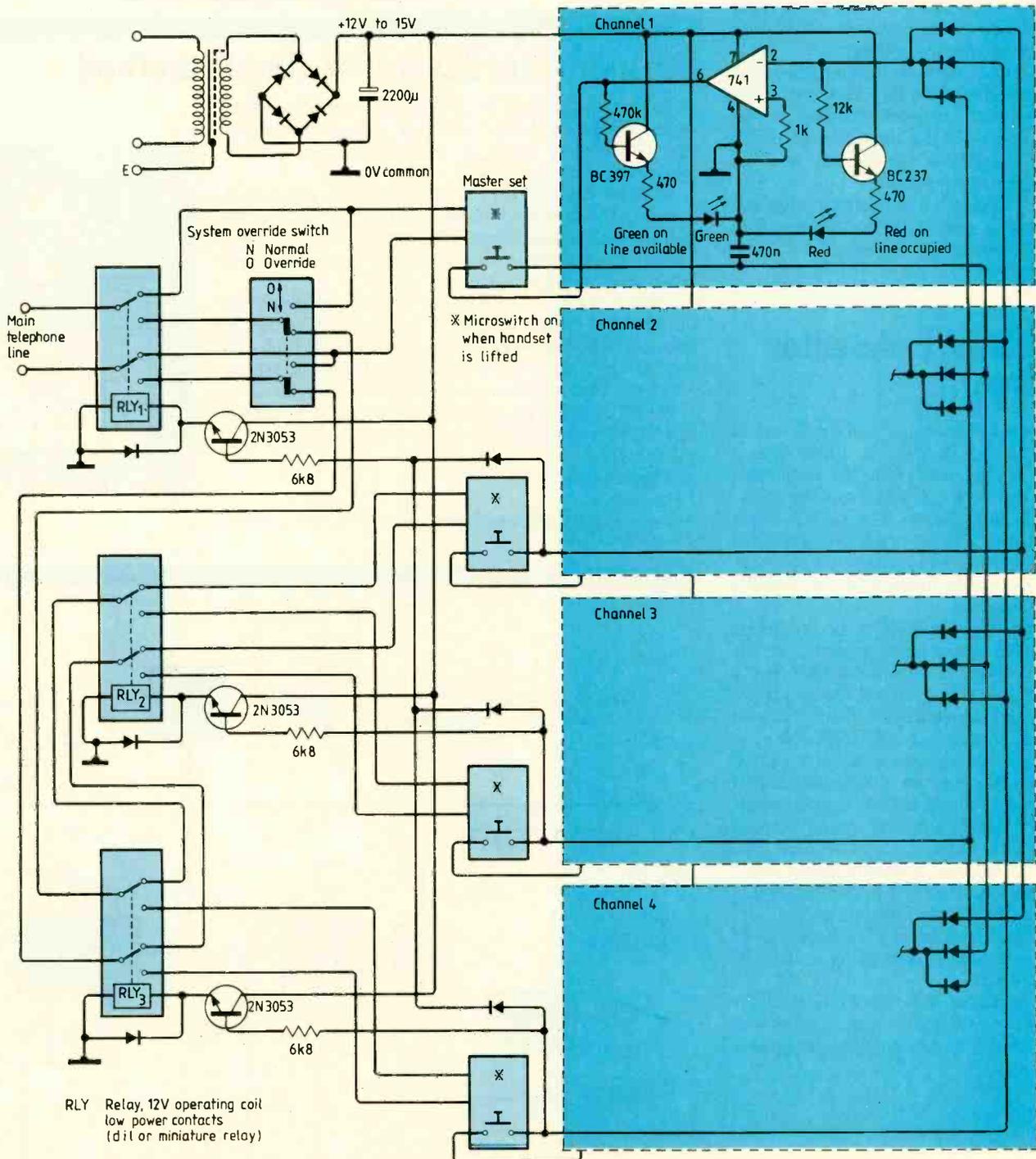
Raymond Farrugia
Zabbar
Malta

Don't waste good ideas

We prefer circuit idea contributions with neat drawings and widely-spaced typescripts, but we would rather have scribbles on "the back of an envelope" than let good ideas be wasted.

Submissions are judged on originality and/or usefulness so these points should be brought to the fore, preferably in the first sentence.

Minimum payment of £35 is made for published circuits, normally early in the month following publication.



RESEARCH NOTES

Fastest microwave transistor

In conjunction with the University of Illinois, the American General Electric Company has developed what it claims is the fastest ever microwave transistor. It's a modfet (modulation doped f.e.t.) with a maximum cut-off frequency of 230GHz – a shade faster than a similar device produced recently by MIT Lincoln Laboratory, the details of which were released earlier this year*.

The increased speed is a function both of the semiconductor material itself and also of the fabrication technology. Electrons travel faster in indium gallium arsenide than in both gallium arsenide and silicon, but unfortunately the former material is not readily available. Researchers at Illinois therefore used molecular beam epitaxy to create 40 atomic layers of indium gallium arsenide on a gallium arsenide substrate. As for device geometry, GE managed to fabricate with a gate width of one quarter of a micron (2.5×10^{-7} m), significantly reducing electron transit time.

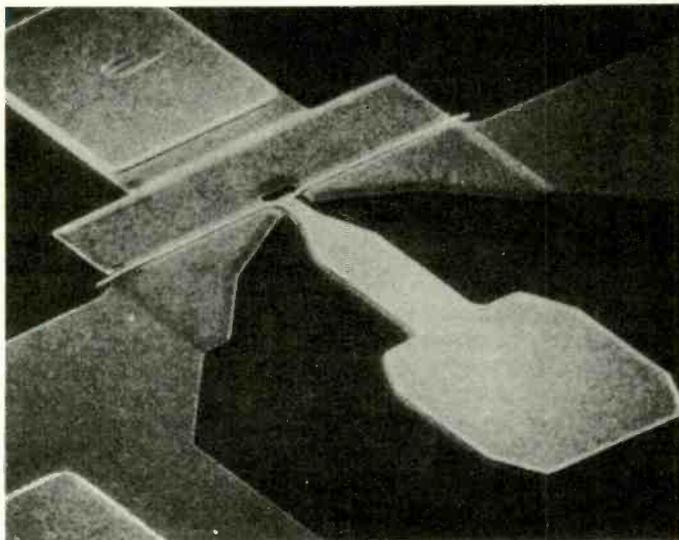
A noise figure of around 2.5dB at 60GHz and an efficiency of 25 to 30% has been achieved with the new transistor, but GE say they have no plans to market the device, and expect to refine it for incorporation into their own products.

* Communications Commentary, October 1986, page 9.

Conductive windows improve r.f. immunity

Recent tests at ERA Technology have shown that thin metal coatings on glass can provide an effective electromagnetic screen to radio-frequency interference. Glass panels coated with a thin layer of conducting material have potential application in the reduction of emissions from computing devices and in providing protection from incoming high-intensity radiation from radar transmitters.

The results of ERA's tests, carried out as part of a sponsored research programme on electro-



magnetic compatibility, agree reasonably well with calculated values based on accepted screening theory, and some simple expressions have been derived to allow designers to obtain a rapid assessment of screening performance from the sheet resistance value.

Two types of coated glass were tested, one having a layer of indium tin oxide deposited by a d.c. reactive sputtering process and the other having a tin oxide/copper/tin oxide sandwich layer deposited by magnetron sputtering. Tests were made on 1m^2 panels forming the front of a copper box which covered the range 1 to 30MHz (magnetic mode) and 30 to 1000MHz (electric mode).

Comparisons between the results with the coated glass panels in place and with the glass removed i.e. from the open box (the box is a typical computer terminal enclosure) showed that the 1m^2 enclosure on its own made little difference to the free-space attenuation in the electric mode except at 50-60MHz, due to multipath interference within the cabinet. However, because of such existing multipath loss, the additional benefits of adding the coated glass panel were negligible at these frequencies.

Elsewhere in the spectrum the glass samples gave attenuation values of 20-30dB. (Separate tests using ridged-guide horn aerials showed that these results held good up to 10 GHz at least.) These values agree reasonably well with theory and may be calculated according to

$$S(\text{dB}) = 20 \log_{10} \left(5 + \frac{100}{R_s} \right)$$

where S is screening effectiveness and R_s the sheet resistance in ohms per square. Results of tests in the magnetic mode showed that the attenuation of coated glass is minimal. ERA Technology say therefore that the application of such glass to v.d.u. enclosures is only of value at higher frequencies. It would not, they add, be much use in protecting line-scan radiation, when a mesh would be more appropriate.

First error-free microprocessor

Ferranti Electronics has produced samples of what it claims to be the world's first microprocessor with guaranteed error-free design. Known as Viper (verifiable integrated processor for enhanced reliability) the new device is a 32bit microprocessor designed by the Royal Signals and Radar Establishment for applications requiring high operational integrity such as aircraft autopilot systems, missile systems and nuclear power plants.

Viper's operation can be formally specified and verified using mathematical techniques to ensure that a completely predictable system can be implemented for such safety-critical applications. This 'provably' correct operation has not been possible with previous microprocessor systems. The 5000-gate logic design was simulated and implemented in silicon by Ferranti using the u.l.a. 'silicon compiler' software system.

Last November, Ferranti Elec-

tronics was one of two companies selected to manufacture the first Viper chips, now delivered to RSRE for evaluation. Although the device is designed to operate in a military environment and is resistant to high radiation levels, it is expected to have many civil applications, and Ferranti will be marketing the microprocessor commercially as the VIP1.

The device is fabricated using Ferranti's latest 1.5 micron double-layer metal advanced bipolar process and will be supplied in a Jedec standard 84-pad chip carrier.

Large-screen l.c.d. gets closer

A race is on between British and Japanese groups to produce the world's first commercially viable fast-switching liquid-crystal display. Here in Britain a collaborative programme under Professor George Gray FRS of Hull University's Department of Chemistry is expected to have a demonstration display running in two year's time. Flat-screen television is however just one application of the University's new work on liquid-crystal technology. Since liquid-crystal screens do not flicker, their use for visual displays in word processors would considerably reduce the eye-strain often associated with those machines. They could also be used to make a very fast camera shutter.

Professor Gray's team, which includes Ken Toyne, David Lacey and Mike Scowston, is continuing research into the tilted smectic liquid crystals required. These respond many times faster than ordinary smectic liquid crystals used for large panel displays or the nematic liquid crystals used for calculators, which are simply not fast enough for television applications.

Already the Hull team have come up with compounds that respond in less than $100\mu\text{s}$ – about a hundred times faster than the response time of nematic displays. Part of the secret lies in the viscosity of the material, but the real advantage comes from the tilted arrangement of the molecules. Unlike those of a nematic system which have to

RESEARCH NOTES

rotate and point in the opposite direction, the molecules of a tiled smectic system have only to sweep out half a cone.

First all-optical regeneration

The first all-optical light regenerator for use in optical communications has been demonstrated by British Telecom Research Laboratories at Martlesham Heath. The regenerator – still in the experimental stage – both amplifies and retimes light pulses directly without converting them from light to electricity, as occurs in conventional repeaters.

All-optical regenerators, when developed commercially, promise considerable savings in the cost of optical communications links which presently need regenerators every 30km or more, especially for undersea systems. Regenerators will be significantly cheaper and simpler to make, and their power requirements will be reduced.

The all-optical regenerator was developed by two British Telecom research engineers, Rod Webb and John Devlin. Its key component is a microlaser which under certain conditions can behave like an optical logic switch. An optical clock signal is fed to the laser to hold the switch state just in the "off" condition. When a pulse of light from the incoming fibre arrives at the laser it has sufficient energy to switch on the laser but only when the optical clock signal is present. This triggers the laser to generate a more powerful burst of light in synchronism with the clock which is then injected into the outgoing fibre.

The BT all-optical regenerator differs from previously demonstrated optical amplifiers in two important respects. Bistable operation leads to a signal output level that is relatively constant over a range of input levels; and secondly the signal is retimed by an optical clock.

It is based on the principle that a Fabry-Perot semiconductor laser has nonlinear transfer characteristics because its effective refractive index varies with optical power level. At some wavelengths this nonlinearity leads to bistability.

To form a regenerator, an optical clock waveform consisting of a train of pulses with peak power just below the bistable threshold is combined with the data stream and coupled into the amplifier. When a pulse is 'low', a slightly amplified clock pulse appears at the output, and when 'high' the additional power is sufficient to exceed the threshold and the output jumps to a higher level, which is insensitive to the data power, and reverts to low only at the end of the clock pulse. The output is the regenerated data in return-to-zero form, retimed by the clock.

Superdeformed nuclei

Results obtained recently from the Nuclear Structure Facility (NSF) at the Science and Engineering Research Council's Daresbury Laboratory represent the final step in a long search for atomic nuclei with a superdeformed shape. Theory predicts that under conditions of extreme stress caused by rapid rotation some nuclei will suddenly adopt a fixed, superdeformed shape. This shape, which is similar to a rugby ball, has a 2:1 major-to-minor axis ratio and is stable under these extreme stress conditions.

Over the past 15 years, confirmation has been sought by many groups worldwide and previous reports by scientists from Daresbury and Liverpool University have already hinted at a breakthrough. In a recent series of experiments at the NSF using high-resolution spectroscopy, scientists have now, for the first time, produced a spectrum showing the sequence of discrete gamma rays emitted as a rapidly rotating superdeformed nucleus (dysprosium 152) slows down.

Almost 2% of the dysprosium 152 nuclei, produced by bombarding a palladium 108 target with beams of calcium 48, were formed in the superdeformed shape. The spectrum shows a series of 19 gamma rays, each separated by an energy of 47keV, which slow down the nucleus from an angular momentum of 60 units to 22 units. This observation of nuclear states up to 60 units is itself a great leap forward: the previous record was

46 units and previous advances have been in steps of only a few units; in addition, 60 units is very close to maximum possible angular momentum for any nucleus before it breaks up under rotational stress.

In superdeformed nuclei the emitted gamma rays, which arise from transitions between a sequence of states, are predicted to have very short lifetimes caused by the large quadrupole moments of the highly deformed shape. The lifetimes of these gamma rays have now been measured by scientists from Daresbury and Liverpool University. These measurements established that the quadrupole moments are indeed extremely large, confirming that this nucleus is the most deformed nuclear shape found so far.

These observations open up new possibilities for studying the nucleus – nature's only strongly interacting, many-body quantum system – under novel conditions. Such studies are already being started at several European and American Laboratories as well as at Daresbury.

Thunder clash in the new Mexico

A bizarre experiment to try and create lightning involved stretching an electrified wire 2km long between two peaks in the Magdalena Mountains. A report by researchers at the New Mexico Institute of Mining and Technology claims not only to have caused clouds to form but to have started local thunderstorms. But spectacular though this may have been, it's nowhere near as significant as the renewed storm it has sparked off over the origin of cloud electrification.

The group believe their work supports a convection theory in which charges are carried into the atmosphere on rising plumes of hot air. This theory is, however, discounted by a rival group of scientists working at the same institute. *Their* theory, known as the particle-charge theory, suggests that charges are built up within clouds as frozen moisture particles collide with each other. On this hypothesis, the more

heavily negatively charged lighter particles rise. Charge separation is thus accomplished.

Which of the theories more nearly represents the true state of affairs remains to be seen; perhaps, as some workers have suggested, it could be a combination of both. Whatever else, these New Mexico experiments represent a spectacular and successful attempt to control the development of thunderstorms.

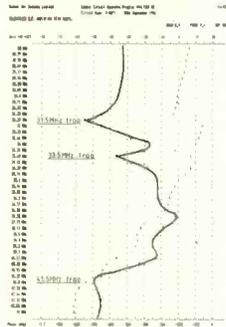
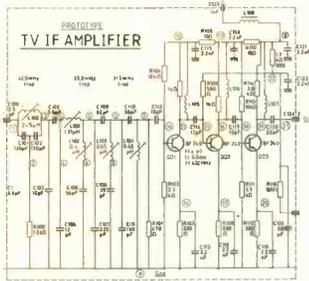
Radio telescope bigger than the earth

Aperture synthesis, common enough on a small scale as a means of reducing the beamwidth of a transmitting or receiving antenna, has recently been applied on the grand scale by radio astronomers in Butan and the USA. Not content with using radio telescopes in different countries linked by radio, they've taken advantage of a temporarily redundant tracking and data relay satellite (TDRS) put into orbit to keep shuttle vehicles in contact with Houston. This TDRS, with its 5m dish, has been hooked up by radio link to the Tidbinbilla tracking station in Australia and a similar receiver at Ousuda in Japan. Together they form an antenna with an equivalent revolving power of a dish 178,000km in diameter. In practice this means that it should be capable of resolving radio objects no more than 2 arc sec. in diameter.

The experiment, coordinated from the Jet Propulsion Laboratory in California, is extraordinary because of the extremely fine degree of control needed on the satellite. For the array to work it was necessary to measure the distance between the earth-bound tracking stations and the satellite to better than 13cm – the wavelength being used for the astronomical observations. In practice, the satellite also had to be controlled so that the 5m dish did not move by more than 1cm during the observations.

Remarkable though this is as a technical feat, it's by no means the biggest array envisaged. A project code-named Quasat is designed to make use of a 50m dish in space.

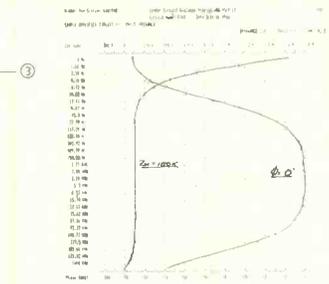
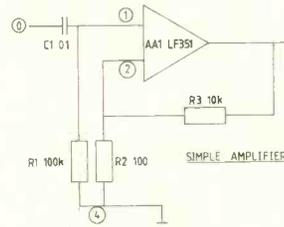
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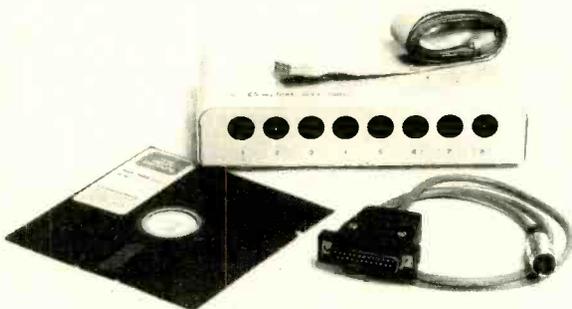
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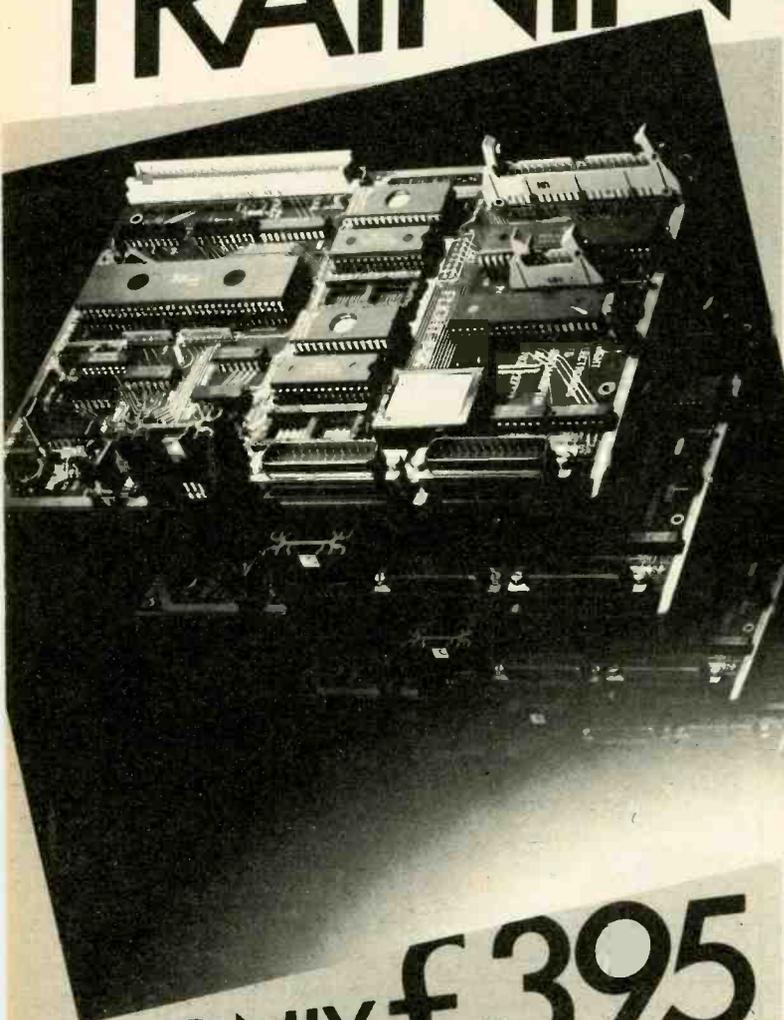
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The maximum power theorem

JW warns that some theorems can endanger your health

JOULES WATT

The transfer of power from a generator to a load is one of the most fundamental operations in electronics and electrical engineering. I have worked with students who, once introduced to the Power Transfer theorem, tend to offer it up with reverence as a kind of panacea.

But as usual the picture is not quite so straightforward. A blind belief that the conditions for maximum power transfer always apply could get you into trouble. I mention a car battery later as an example. If you tried to match the battery, you might well end up with molten copper, boiling electrolyte and buckled plates! So, automotive engineers do not attempt to match the battery.

In case your memory is a little rusty on the theorem, I will briefly review it. In the "d.c." form, the theorem states: "The maximum power is transferred from a source to a load when the load resistance is equal to the internal resistance of the generator." This is

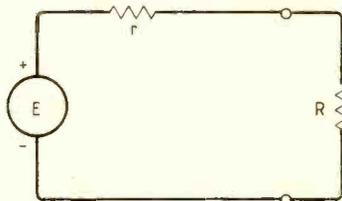


Fig.1. Power in load is a maximum when $r=R$.

easily seen from Fig.1 where, on finding the current and then the power in the load resistor, we have:

$$I = \frac{E}{r+R}$$

∴ power in R is

$$P_R = I^2 R = \frac{E^2 R}{(r+R)^2}$$

If there is a maximum when R varies, this occurs when $dP_R/dR=0$.

$$\text{Now } \frac{dP_R}{dR} = \frac{E^2[(r+R)^2 - 2R(r+R)]}{(r+R)^4} = \frac{E^2(r-R)}{(r+R)^3}$$

and for this to equal 0, $R=r$.

You can see intuitively that, if a short circuit is applied, the power available externally is zero. For an open circuit, no current flows, so the power dissipated is again zero. The simple analysis shows that in between these extremes the power available is a

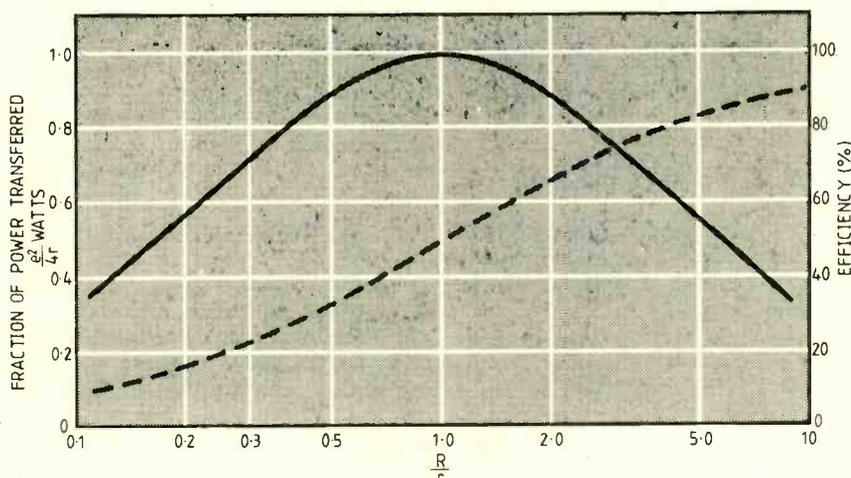


Fig.2. Although power is a maximum when $r=R$, efficiency is 50%.

maximum when $R=r$. A plot of power in the load as R varies, r remaining fixed, shows this. See Fig.2.

Communications and radio engineers often call this equality of source and load resistances the *matched condition*. From the equation for P_R , with $R=r$ inserted, the maximum available power from a source e.m.f. E with internal resistance r is,

$$P_{\max} = \frac{E^2}{4r}$$

Some authors confusingly state that P_{\max} is "independent of the load". It is, if the equation is written as I have shown without R in it, but you have to remember that the maximum power is only obtained in practice if R is adjusted for the matched condition.

The theorem also applies to a.c. circuits containing reactances as well as resistance. All active circuits that end up with a final pair of terminals supplying a load can be reduced to a voltage generator, plus an internal resistance in series with a reactance, as shown in Fig.3. This is an application of Thévenin's theorem. (Norton's theorem is equivalent, but reduces the circuits to a current generator with an internal conductance shunted by a susceptance. Norton is said to be the *dual* of Thévenin.)

If the source impedance of a single-frequency a.c. generator is $z=r+jx$ and a load impedance connected to it is $Z=R+jX$, then the magnitude of the current is

$$i = \frac{e}{\sqrt{(r+R)^2 + (x+X)^2}}$$

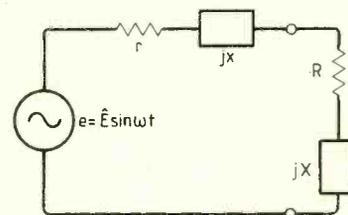


Fig.3. Any active circuit of the type supplying a load comprises a generator, an internal resistance and a reactance.

We are interested in the power dissipated in the resistive part of the load only, because the reactive part never dissipates any power. This is given by

$$P_R = \frac{e^2 R}{(r+R)^2 + (x+X)^2}$$

Without any differentiating, you can see that P_R is greatest for given resistances when $X=-x$. This means whatever the source reactance value, you can maximize the power in the load by 'tuning it out' using an external reactor, equal to the internal one but opposite in sign, as part of the load. In fact, you have simply set up a series-tuned circuit in the system.

But matters are not as simple as they seem. Not many people are interested in a continuous single frequency. In communications, a whole band of frequencies occurs. In digital systems, it is even worse. The band of harmonics is very wide for fast

switching edges. Many digital engineers have come to grief by not realising the 'analogue' nature of the networks, transmission lines and so on, in their systems.

The source e.m.f. and the real and imaginary components of its impedance in a Thévenin equivalent circuit vary with frequency. Even for fixed values of reactive components, L and C, the shifts of reactance with frequency are in opposite directions, relative to each other. For simple circuits this means that, at best, matching can only be obtained at one or two points in the band.

Usually, engineers cope with this by showing that a good compromise can be arrived at by equalizing the *magnitudes* of the source and load impedances. If Z is the magnitude of the load impedance and θ its phase angle, then the real and imaginary parts are $Z\cos\theta$ and $Z\sin\theta$ respectively. The power in the real part is then

$$P_R = \frac{e^2 Z \cos \theta}{(r + Z \cos \theta)^2 + (x + Z \sin \theta)^2}$$

and if you differentiate this with respect to Z you can show that the maximum power is transferred when $Z=z$. The amount of power obtained now becomes,

$$P_{\max} = \frac{e^2 \cos \theta}{4z \cos^2 \theta / 2(\theta + \theta_s)}$$

where z is the magnitude of the source impedance and θ_s is its angle.

To return to my opening remarks, one decidedly does *not* try to match car batteries, whose r may be a few milliohms. The obvious reason is that, if the maximum power theorem is attempted, many kilowatts become involved, half of which are dissipated inside the battery, which would soon boil, and explode, or something. The other half would melt the wiring harness.

Consider matching a power station to a city — an even more disastrous proceeding. The efficiency can never be more than 50% for the maximum power condition, which would mean that half the megawatts produced would have to be dissipated at the power station.

What this means is that power engineers do not show much interest in the maximum power theorem. Their aim is to increase power transfer efficiency. As you can see from curve B in Fig.2, high efficiencies require $R \gg r$ and this is certainly true in the application of car batteries and power stations.

So the discussion comes back to communications engineers, who do often require maximum signal power in a load. For example, the rather naive case of constant noise power in a system would require the

signal power to be maximized for best signal-to-noise ratio to be obtained. But even in our field, this maximum condition is not always sought. In power amplifiers — class B operation, for example — efficiencies of 70% or so might be obtained. In r.f. work with class C stages, even higher efficiencies are expected. 50% is not looked upon as very good, but that is the best you can do with maximum power matching as we have seen.

For good voltage regulation from a power supply, the source impedance must be very low compared to the load and any variations in it. Such supplies do not attempt the matching condition on this argument alone. The regulation from no load to full (matched) load would be,

$$\frac{\text{Voltage at maximum load}}{\text{Voltage at no load}} = 50\%$$

which is very poor.

In conclusion, it seems that in small-signal, high-source-impedance, systems, we often consider the maximum power condition. In these, the money is in the hardware and not the energy consumed. In power systems, where costs of the energy dominate, it is efficiency that counts and that requires $R \gg r$.

Hadamard versus Fourier

continued from page 17

would be used to store the original voice data, and be used in later speech reproduction. The 'frequency' data would not be used as such; that information is held in a number of these coefficients and these can be used to mathematically reconstruct the original waveform.

SPEECH RECOGNITION

As part of an experiment to test the applicability of the Hadamard transform, a small routine was written around the basic FHT algorithm (program 1) to input 10000 samples of speech in one second. The FHT was then used to transform 10 consecutive blocks of 1000 points into the frequency domain. A proportion of the major coefficients from each block were then stored in memory and used to reconstruct the original speech waveform and played through the internal speaker of an Apricot Xen. The program was run interactively through the keyboard within a Forth kernel.

First results found that the reconstruction of certain words containing the phonemes /l/, /w/ or /r/ were difficult to distinguish. Subsequent attempts increasing the number of coefficients from 16 to 32 or 64, enabled better audio resolution of the phonemes. There are still difficulties at the time of writing — for instance, the phonemes /th/, /g/ and /p/ are difficult to distinguish, and appear to depend on the exact instant when sampling is started. However, as a first approach the FHT technique gives excellent results and is capable of analysing speech or voice data in real time. An interesting little

diversion was to shift the transformed data up in frequency, and then playback the 'Mickey Mouse' version of the original speech!

The FHT technique has also been used in an Apricot Xen to control an electrochemistry module that transmits a multi-harmonic signal to an electrode immersed in a solution containing species that will react at particular potentials. The return of the signal through special potentiostat circuitry contains the (stimulated) response of the species and can be defined by frequency response analysis. Different species and electrode materials change and impedance of the system in predefined ways and can be used as such to identify the processes and mechanisms at the electrodes. FFT software was used in the past, but by changing to the FHT there has been a substantial decrease in the time taken to perform all the necessary calculations without loss in information content.

The same should be true of many other areas that use transforms, such as image coding, video techniques, speech and object recognition, radar and radio and data transmission and compression. Because of the general complexity of the FFT algorithm a suitable alternative has been found that is simpler to code, occupies less memory, does not resort to complex number notation, and is considerably faster in operation. There are other transforms, and some of these will undoubtedly be better than the FHT for certain cases, but this article demonstrates that the FFT is not the only transformation routine available, and that it is possible to

Gray codes are not natural codes for computers. They were first developed for use on shaft encoders, or for work with rotating parts where the angle of rotation had to be controlled accurately. It has the property that only one bit in the binary code changes from one transition to another. The rule is:

- begin with all zeros,
- always change the single l.s.b. that gets to the next state.

Gray codes can be generated with any number of bits. The machine-code routine to do this is rather complicated, but a Fortran subroutine looks like:

```
SUBROUTINE GRAY(M,N,X)
DIMENSION X(N)
C N = 2**M = number of samples
N2=N/2
N4=N/4
DO 1 I=1,M-1
INT=2**(I-1)
DO 2 L=1,INT
DO 2 J=1,N4
IY=N2+J+(L-1)*N/INT
IX=IY+N4
Z=X(IX)
X(IX)=X(IY)
2 X(IY)=Z
N2=N2/2
1 N4=N4/2
RETURN
```

tune a microcomputer to work in, or approach, real-time analysis.

Mark Varney obtained his doctoral degree at Liverpool University in chemical oceanography. He has spent three years in postdoctoral research and two years as director of his own research consultancy, designing and developing analytical instrumentation. In 1986 he joined the Oceanography Department at Southampton University where he carries on this work and also lectures in marine organic chemistry.

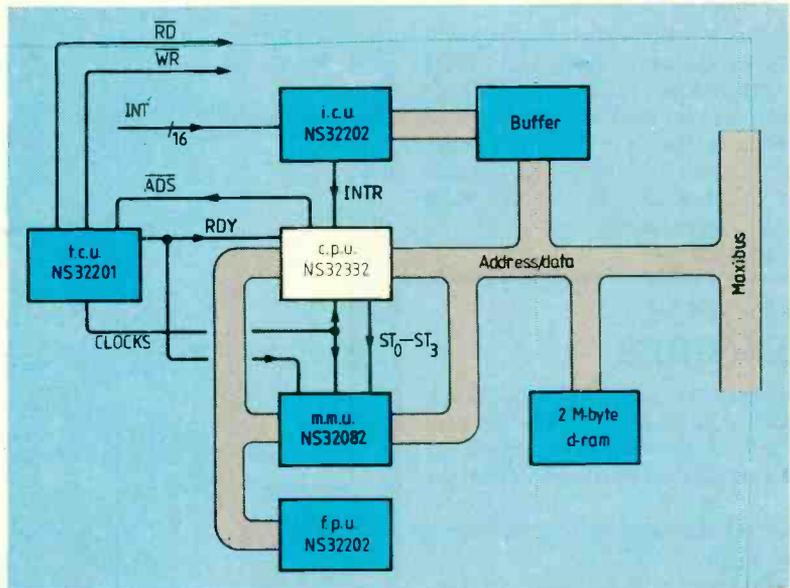
APPLICATIONS SUMMARY

32-BIT COMPUTER

Computer boards accessing memory across a non-standard bus designed to suit the microprocessor have advantages over systems using industry-standard buses, according to a note on National Semiconductor ICM332-1 integrated computer modules. These advantages are that price decreases because there is no need for backplanes and card cages, and performance increases because all memory is accessed through a local bus tailored to suit the processor.

Boards described in the note are based on 32-bit Series 32000 processors and peripherals which include the NS32332 c.p.u., 32201 timing controller, 32082 memory manager, 32081 floating-point unit and 32202 interrupt-control unit. These modules are intended for designing into office automation, workstation, graphics and process-control equipment.

The c.p.u. board has 2M-byte of ram with parity checking but can access a further



12M-byte through its 32-bit Maxibus. Further boards provide i/o processing, with s.c.s.i., and memory expansion. National's

Unix operating-system port Genix V3 is available for use with the boards. 301 on reply card

IMAGE SENSING D-RAM

When light falls on a dynamic memory cell its storage capacitor discharges at a rate proportional to both the light intensity and exposure duration. It has been known for some time that with certain types of d-ram, if you can successfully expose the top part of the chip, you can use it as a crude image sensor.

Now an American company is manufacturing specially fabricated dynamic rams with a glass window on top for image sensing. Unlike ordinary d-rams, these devices can be adjusted for sensitivity by varying voltage on an analogue input pin (pin 1 for the d-ram pinout experts).

As with ordinary dynamic rams though, these i.c.s have two or more separate memory matrices on the chip depending on the version. With the 64K-bit device there are two 128-by-256 element matrices for example so a picture sensed using the whole chip has a dead zone either along or down the middle. But the sensors are intended for computerized image sensing, in say robotics

applications and process control, so the dead zone becomes less of a problem. Of course just one matrix can produce a lower resolution picture without a dead zone.

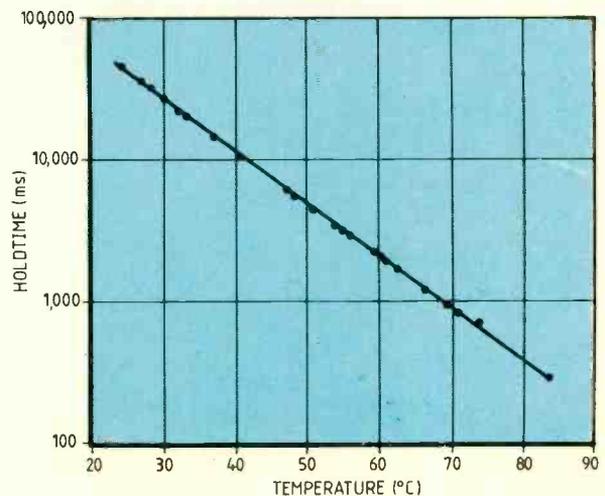
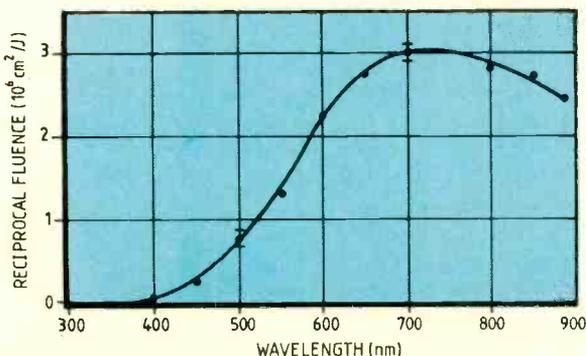
In order to achieve high chip density, cells within the chip are staggered so the address matrix does not correspond exactly with the physical matrix. Unless the image representation is rearranged, either in hardware or software, it will lack definition because of the staggering.

There are two possible ways of achieving a grey scale, one by varying the threshold voltage and the second by altering the scan rate. Although varying the threshold voltage seems most obvious, the manufacturer suggests that altering the scan rate is the best method.

Topology is described in detail in the

device data sheets. These graphs come from the IS32 OpticRam Spectral data applications note. This note discusses sensitivity, effects of threshold voltage (pin 1), dark current and pixel-to-pixel and chip-to-chip sensitivity. Reciprocal fluence versus wavelength illustrated in the first graph shows that peak sensitivity is about $3 \times 10^6 \text{ cm}^2/\text{J}$ between 700 and 800nm. The second graph is of more general interest, showing the maximum hold time for the unilluminated d-ram against temperature.

Three OpticRams are produced by Micron Technology Inc., one with two 128-by-256 image sensing elements, one with four 128-by-512 elements and one with twenty 128-by-256 elements. They are distributed in the UK by Joseph Electronics. 302 on reply card



APPLICATIONS SUMMARY

CRT CONTROLLER

Comprehensive design details starting with a comparison between raster and vector graphics are included in Hitachi's 118-page application note for the HD63484 advanced c.r.t. controller. Details of this frame-buffer interface include ram timings and full dot-clocking information. The device's many registers are also explained. 303 on reply card.

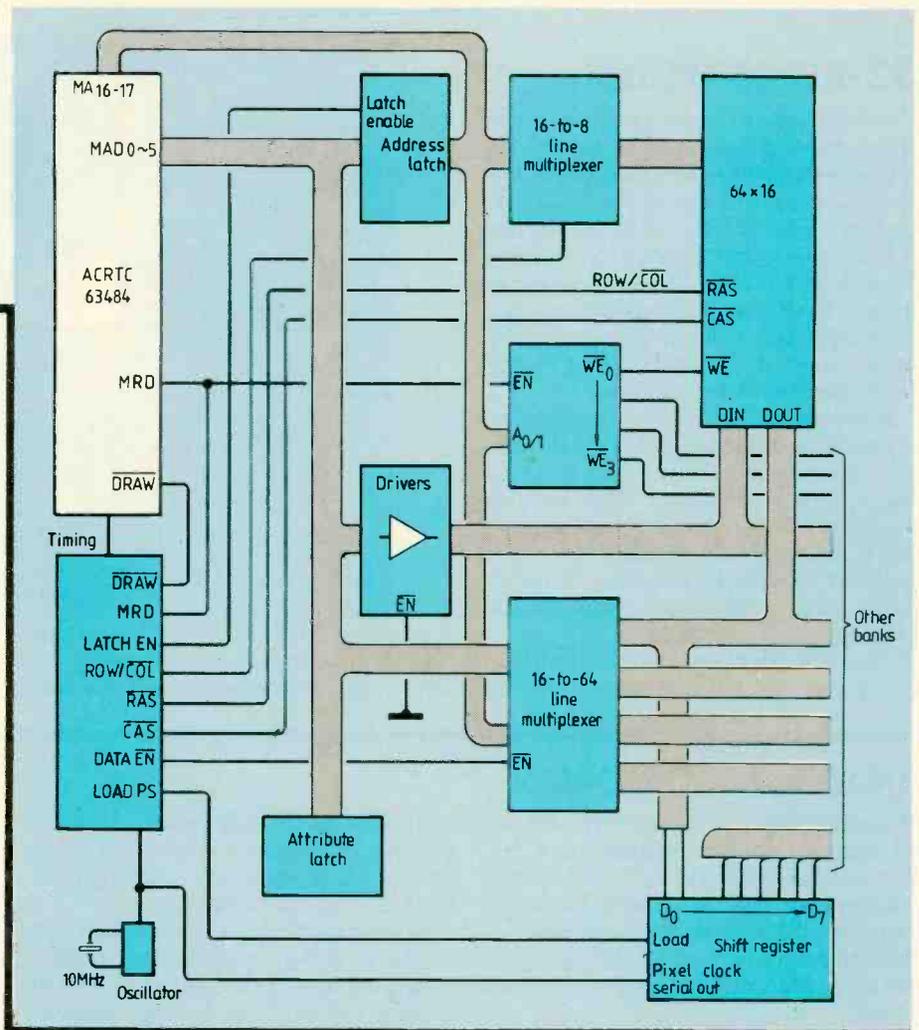
PCM/ADPCM TRANSCODER

Adaptive differential pulse-code modulation, or a.d.p.c.m., increases efficiency in p.c.m. telephone links. The device shown below is a p.c.m./a.d.p.c.m. encoder/decoder called the TEL72.

It has three-state outputs and runs from a single 5V supply.

In conformance with international standards, the c-mos 721 transcodes 64k-bit/s p.c.m. to 32k-bit/s a.d.p.c.m., or vice versa, on a single speech channel. Switching for either the CCITT G721 or draft UST1Y1 algorithm operation is included and both A-law and μ -law p.c.m. i/o can be handled. For digital-speech interpolation, the 721 can be operated at lower a.d.p.c.m. rates of 24 or 16k-bit/s.

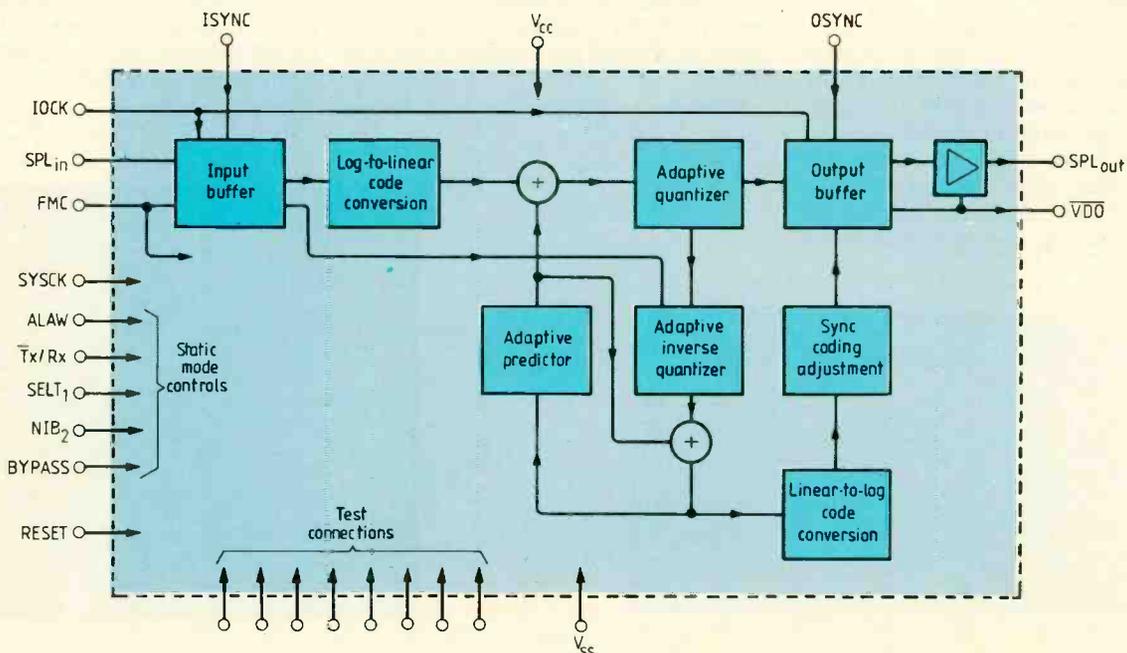
Several transcoders are shown connected to common input and output buses in the application example. Input in the form of p.c.m. or a.d.p.c.m. is fed to all devices and each device is enabled by a different input strobe so that it responds to only one speech channel. Likewise, each device is enabled by a different output strobe so that it inserts data for one speech channel into the p.c.m. or a.d.p.c.m. output data stream. A nibble selection input assigns the 4-bit a.d.p.c.m.



sample to the first or second half of the 8-bit multiplexed time slot.

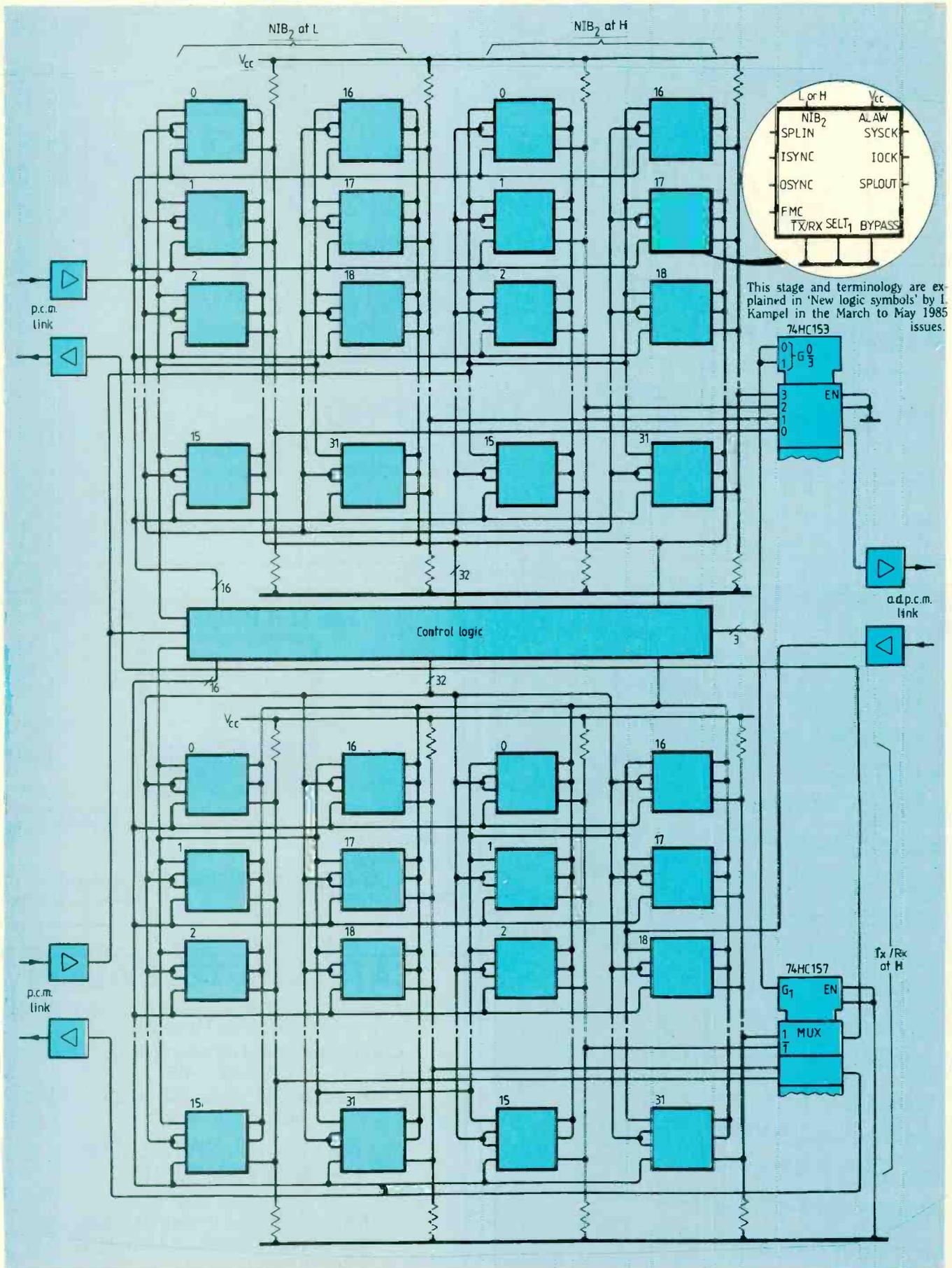
Currently the 721 complies with G721 1985 recommendations. When pending

changes to the G721 and T1Y1 algorithms have been ratified, STC plans to introduce a revised version of the chip. 304 on reply card.



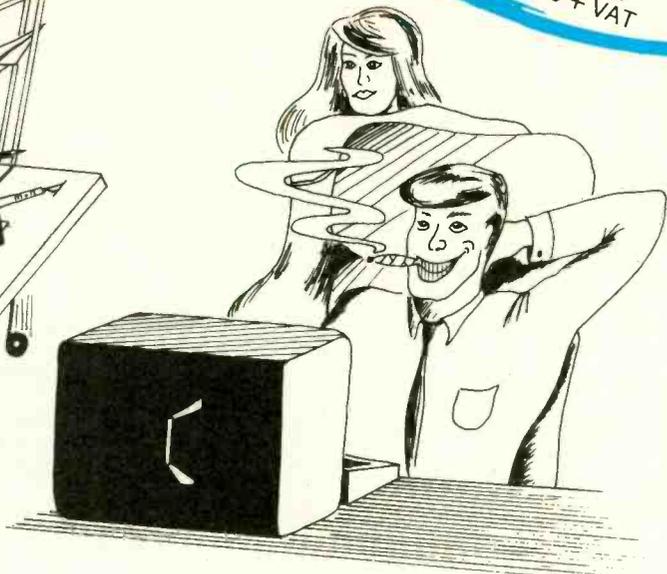
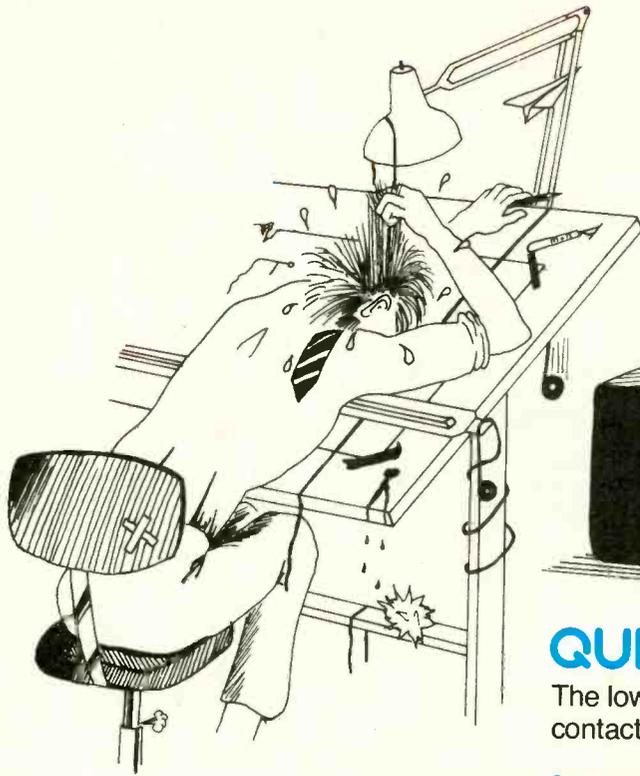
This issue carries an article describing an OK1 chip with a similar function - Ed.

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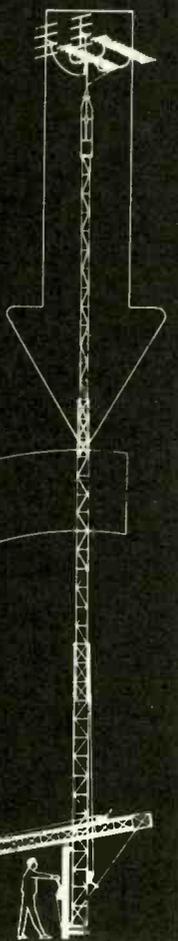
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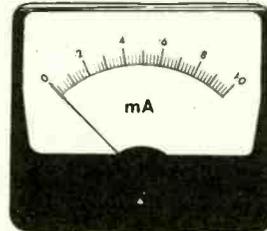


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ENTER 6 ON REPLY CARD

Double-notch filtering simplifies distortion analysis

Using a notch filter with a canyon-shaped characteristic makes t.h.d. analysers easier to tune and produces more consistent results.

B.J. SOKOL

Equipment for measuring total harmonic distortion is invaluable for checking amplifiers, tape-recorders and record styli and should be used regularly to check high-quality reproduction systems for degradation (see ref.). It is also useful for designers and builders of audio equipment.

An adjustable notch filter and simple circuit, Fig.1., will measure t.h.d. but manual adjustment of a deep-notch filter with reasonable Q can be difficult*. My aim was to design a band-stop filter with at least 60dB rejection to allow fundamental suppression to 0.1%, and yet avoid using the usual two interacting controls for frequency tuning and depth trimming.

My first prototype – a notch filter which is automatically tuned to the fundamental frequency – consisted of a 4046 phase-locked loop i.c. driving an MF10 switched-capacitor filter. This approach failed to meet my design requirements for the reasons shown in the panel but the design was capable of servo control of fundamentals down to 1% so it would be suitable for low-fidelity measurements.

Next I thought more carefully about the usually employed notch filter itself. This is a second-order filter with a pair of zeros on an imaginary axis. These zeros lie where the axis is cut by a circle around the origin passing through the poles. Thus both zeros and poles have the same resonant frequency and the filter transfer function is

$$H(s) = \frac{s^2 - \omega^2}{s^2 + (\omega/Q)s + \omega^2}$$

At the pole frequency, this filter has a hole in its amplitude characteristic of a theoretically infinite depth for when s is $-j\omega$, the numerator of $H(s)$ becomes zero. Narrowness of the notch surrounding this hole is what makes t.h.d. measuring instruments difficult to tune manually.

But infinite suppression of the fundamental is not essential for practical t.h.d. measurement so I thought about what could be done to achieve a broader notch by sacrificing depth. I needed in effect not a notch but a 'canyon' in the amplitude transfer charac-

* The need for tuning is especially acute if transducing equipment such as a record turntable or tape deck is being checked. Width of a 'canyon' filter floor allows a consistent and accurate measurement to be made despite moderate variations in the speed of the transducer motor.

SERVO NOTCH FILTER

This servo notch filter tunes the notch to input frequency by forcing bandpass output to be in phase with the input. When bandpass-output is in phase the notch is in quadrature and thus nulls the input. Null is set at $s = -j\omega = -j$ where

$$H_{BP} = \frac{s}{s^2 + \alpha s + 1}$$

and

$$H_n = \frac{s^2 + 1}{s^2 + \alpha s + 1}$$

The effect of a phase error $d\phi$ is equivalent to the bandpass output BP locking the loop at phase $d\phi$ at frequency $1 + d\omega$;

$$\phi_{BP} = \tan^{-1} \frac{1 - \omega^2}{\alpha \omega} \approx \frac{1 - \omega^2}{\alpha \omega}$$

giving

$$\frac{d\omega}{d\phi}(1) = -\frac{\alpha}{2}$$

This frequency error moves the notch amplitude function up from null according to the derivative

$$\left. \frac{d|H_n|}{d\omega} \right|_{\phi=0}$$

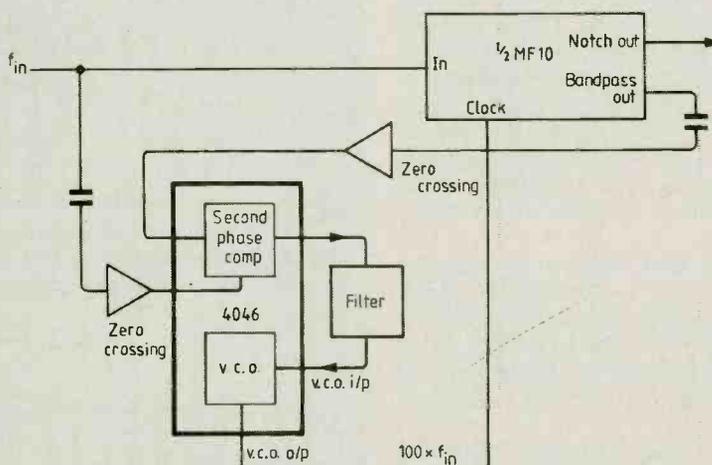
which works out to $-2/\alpha$. So the overall proportional effect of the phase error is

$$\left. \frac{d|H_n|}{d\phi} \right|_{\phi=0} = \left. \frac{d|H_n|}{d\omega} \right|_{\phi=0} \frac{d\omega}{d\phi} = \frac{-d}{2} \cdot \frac{-2}{\alpha} = 1$$

Thus to get -60dB (amplitude 0.001) requires 0.001 radian phase accuracy.

This translates to a minimum voltage accuracy in the zero crossing comparators of 1mV per peak volt input because $V = A \sin \theta$ has phase slope

$$\frac{d\theta}{dV}(0) = \frac{1}{A} \text{ rad/V}$$



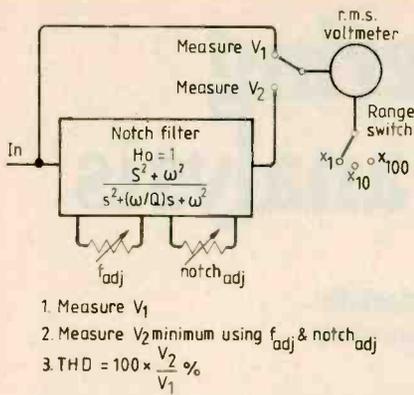


Fig.1. An adjustable notch filter and simple circuit measure t.h.d. but manual adjustment of a deep notch filter can be difficult.

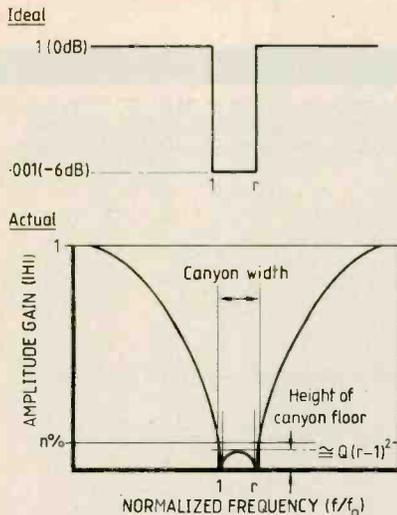


Fig.2. Using two notches placed close together results in a canyon-shaped transfer function. Using such a characteristic instead of a narrow notch filter makes t.h.d. measuring instruments easier to tune.

teristic; a gorge not necessarily infinitely deep but with steep walls and with a certain width at the bottom.

Experiments with a computer program of mine called Bode (a listing is available), which tabulates phase and amplitude characteristics of any filter or transfer function, soon showed that two notches placed close together will produce just this overall transfer function.

Figure 2 shows such a transfer function using a graphical version of Bode. Of course the canyon bottom is not perfectly flat but maximum height between the walls can be limited to comply with the required figure of -60dB. Several applications of Bode showed that there is a trade-off between maximum floor height, filter Q (or steepness), and the distance (in frequency) between the canyon walls.

Investigating how wide one can make the floor of this canyon filter without unduly sacrificing its depth requires a little calculation. (Normalized frequency ratio rather than absolute frequency determines the slope of all filter curves, hence for example 6dB/octave.)

Without losing generality, frequency (radian) of the lower of the two notches can be made 1 and of the upper notch, r. If

(a)
Do you want normalized frequencies? Type y/n <cr>.
Type: starting freq (relative to 1), number of steps/octave, number of octaves.
Do you want a dB amplitude scale? Type dB/lin <cr>.

Number of simple poles 0
Number of simple zeroes 0
Number of double poles 2
Number of double zeroes 2

Type coeffs B,C of (next) poles s^2+Bs+C
Type coeffs B,C of (next) poles s^2+Bs+C
Type coeffs B,C of (next) zeroes s^2+Bs+C
Type coeffs B,C of (next) zeroes s^2+Bs+C

(the programme BODE replies)

POLES
-0.5 + .8660254 j
-0.5 + -.8660254 j
-0.515 + .8920062 j
-0.515 + -.8920062 j

ZEROES
0 + 1 j
0 + -1 j
0 + 1.03 j
0 + -1.03 j

Frequency	Phase (rad)	Amplitude
0.25	0.5130143	0.9356178
0.3535534	0.7550781	0.8640671
0.4999999	1.153829	0.7023717
0.7071065	-1.27174	0.3522751
0.9999995	-5.905931E-02	5.628971E-08
1.414213	1.188121	0.3128393
1.999999	-1.19933	0.6813552
2.828425	-0.7814911	0.8548866
3.999997	-0.5297	0.9314614

(b)

Do you want normalized frequencies? Type y/n <cr>.
Type min W, max W, W step:
Do you want a dB amplitude scale? Type dB/lin <cr>.

Number of simple poles
Number of simple zeroes
Number of double poles
Number of double zeroes

Type coeffs B,C of (next) poles s^2+Bs+C
Type coeffs B,C of (next) poles s^2+Bs+C
Type coeffs B,C of (next) zeroes s^2+Bs+C
Type coeffs B,C of (next) zeroes s^2+Bs+C

(the programme BODE replies)

POLES
-0.5 + .8660254 j
-0.5 + -.8660254 j
-0.515 + .8920062 j
-0.515 + -.8920062 j

ZEROES
0 + 1 j
0 + -1 j
0 + 1.03 j
0 + -1.03 j

Frequency	Phase (rad)	Amplitude
0.99	-9.917195E-02	1.587483E-03
0.995	-7.907128E-02	6.916142E-04
1	undetermined	nil
1.005	-3.913313E-02	4.896358E-04
1.01	-1.930095E-02	7.797505E-04
1.015	4.364872E-04	8.729742E-04
1.02	2.007696E-02	7.721004E-04
1.025	3.961799E-02	4.800777E-04
1.03	undetermined	nil
1.035	7.839303E-02	6.649018E-04
1.04	9.762275E-02	1.51127E-03

Fig.3. These two dialogues with Bode give the effect of a double notch, with both Q values at 1 and with spacing of 3%. Dialogue (a) reports at 1/2 octave intervals to give an overall idea of the notch. In dialogue (b), a fine analysis of the double notch around the centre is produced by using closer frequency intervals of 0.5%.

$$H(s) = \frac{s^2+1^2}{s^2+s+1^2} \cdot \frac{s^2+1.03^2}{s^2+1.03s+1.03^2}$$

Note: the 'canyon floor' is really at \sqrt{r} but this is close to $(1+r)/2$ when r is approximately 1. If r is 1+d, $(1+r)/2$ is $1+(d/2)$ and $(1+(d/2))^2$ is $1+d+(d^2/4)$ which is approximately equal to $1+d=r$.

distance between the notches is, say, 3% then r is simply 1.03. Assuming that each notch is symmetrical and with the same Q of $1/a$ (you will see why later) the transfer functions of the two second-order notch filters are,

$$H_1 = \frac{s^2 + 1}{s^2 + as + 1}$$

and

$$H_2 = \frac{s^2 + r^2}{s^2 + ars + r^2}$$

Amplitude of these transfer functions is determined by substituting $-j\omega$ for s and then calculating the square root of the sum of the squares of the real and imaginary coefficients. Here, ω represents relative frequency in the ratio 1:r so the rad/s versus Hz distinction need not be made.

Manipulation becomes rather complex but the points of interest – depth and width of the canyon floor – have a relatively simple expression. When the two transfer functions are multiplied and the combined transfer function is assessed for amplitude it turns out that the maximum value of the amplitude between the two notches, the highest point of the canyon floor, is exactly.

$$\frac{(r-1)^2}{(r-1)^2 + ar}$$

For small values of $r-1$, for example when the notches are fairly close together, and Q is $1/a$, this is nearly equal to $Q(r-1)^2$.

This is a pleasing result for if r is 1.03 and Q is 1 the height of the canyon floor is close to 0.03^2 which is less than 0.1%. A canyon filter consisting of two notch filters spaced apart by 3% will therefore suppress all frequencies by at least 60dB across a band at least 3% wide (a musical quarter tone); it will also be easy to tune.

Using the program reveals an exact result showing that a 3% gap between notches actually produces a canyon floor with a maximum height of 0.000873 (relative to an amplitude of 1) and that around this canyon there is a 4% wide frequency band throughout which amplitude suppression is better than 60dB, Fig. 3. This result exceeds the approximation so there is room in the design for a slightly wider notch, which is useful as can be seen from Fig. 4.

The next issue to consider is steepness of the canyon walls. One must examine behaviour of the double notch at frequency $f=2$ and above, which is where the harmonic distortion products are found. A neat approximation is possible when the notches are close together, for then they can be considered to be coincident at $f=1$ looked at from $f=2$ or above. Hence the transfer function is approximately the square of the transfer function of the single notch H_1 above. If $-j\omega$ replaces s in the formula and the result is squared, the expression for the amplitude at $f=2$ simplifies to $9/(9+4a^2)$.

This approximate result indicates that for a canyon with Q values of 1, amplitude gain at the octave above the fundamental will be 9/13 or 69.2%. Using the program to find the exact values gives 68.1% for a notch spacing of 3% (Fig. 3), confirming the approxima-

```
Do you want normalized frequencies? Type y/n <cr>. n
Type min W, max W, W step: 0.98,1.05,0.005
Do you want a dB amplitude scale? Type dB/lin <cr>. lin
```

```
Number of simple poles 0
Number of simple zeroes 0
Number of double poles 2
Number of double zeroes 2
```

```
Type coefs B,C of (next) poles s^2+Bs+C 1,1
Type coefs B,C of (next) poles s^2+Bs+C 1.035,1.0712
Type coefs B,C of (next) zeroes s^2+Bs+C 0,1
Type coefs B,C of (next) zeroes s^2+Bs+C 0,1.0712
```

(the programme BODE replies)

```
POLES
-.5 + .8660254 j
-.5 + -.8660254 j
-.5175 + .8963224 j
-.5175 + -.8963224 j
```

```
ZEROES
0 + 1 j
0 + -1 j
0 + 1.034988 j
0 + -1.034988 j
```

Frequency	Phase (rad)	Amplitude
0.98	-0.1491927	4.384417E-03
0.985	-0.1289434	2.978065E-03
0.99	-0.1087736	1.779765E-03
0.995	-8.868631E-02	7.877373E-04
1	undetermined	nil
1.005	-4.876988E-02	5.856179E-04
1.01	-2.894603E-02	9.7147E-04
1.015	-9.215174E-03	1.160086E-03
1.02	1.042028E-02	1.154161E-03
1.025	2.995801E-02	9.565485E-04
1.03	4.939569E-02	5.702484E-04
1.035	6.873123E-02	1.597292E-06
1.04	8.796241E-02	7.557175E-04
1.045	0.1070873	1.68872E-03
1.05	0.1261039	2.797099E-03

Fig. 4. Fine-scale tabulation from the Bode program of a double notch allowing a 3.5% gap between notches. Q is 1.

$$H(s) = \frac{s^2 + 1}{s^2 + s + 1} \cdot \frac{s^2 + 1.035^2}{s^2 + 1.035s + 1.035^2}$$

tion. By the time that $f=4$, where f is 4, the program shows that the amplitude value is 93.1% so there is little loss.

If Q were increased to two the above equations show that gain at the first harmonic ($f=2$) would be 90% while amplitude response at the highest point of the canyon floor would be less than 0.002 or 0.2% of the fundamental level. That might seem a better compromise but it was rejected because sensitive t.h.d. measurements are for most purposes more important than absolutely accurate ones. That is to say, 60dB fundamental suppression seemed more important than the partial loss of harmonic products due to a lower Q .

Returning to the former design with a Q of 1 and the notches 3% apart, you will recall that amplitude response is 0.681 at $f=2$. This represents a loss of first-harmonic content of about 3dB. Loss of higher harmonics is negligible. For reasons discussed in my next article, Q of the canyon filter may vary over a range of about 1.5:1 when it is tuned. This will result in a slight degradation of fundamental suppression, offset by an improvement in canyon steepness.

On paper, or using Bode, better results seem to be possible if asymmetrical notch filters with ω_p not equal to ω_z or two notches with unequal Q values are employed. Such filters can pass frequencies higher than the fundamental (the harmonic) to a greater extent at the expense of suppression of frequencies below the fundamental[†].

However, such methods are undesirable. False readings of t.h.d. are likely to result from a seemingly useful steep rise of the

amplitude function up to $f=2$ and a very small amplitude response at and below $f=1$. For example if there were quite a lot of second-harmonic distortion the minimizing procedure outlined in Fig. 1 could well result in a frequency above the fundamental being chosen as $f=1$ so that a little less fundamental than 0.1% but quite a lot less second-harmonic than 68% reaches the supposedly nulled r.m.s. distortion measurement. In other words one could null out some of the harmonics which would of course give a falsely low t.h.d. reading.

Thus it seems that a symmetrical double-notch filter with notches approximately 3% apart and with both Q values at around 1 will serve the purposes of my design very well. It should also be easy to tune.

Putting the double-notch filter to use will be discussed in a further article. A copy of the Bode listing can be obtained by sending a large s.a.e. to E&WW's editorial offices at Quadrant House, The Quadrant, Sutton, Surrey SM2 5AS. Please write Bode clearly on your envelope.

Reference

Distortion on and off record by John Linsley Hood in *Hi-Fi News and Record Review* Oct. 1982 discusses how useful t.h.d. measurement can be in obtaining and monitoring high-fidelity equipment.

[†] The ultimate extension to this concept would be to use a high-order elliptic or Cauer high-pass filter to pass only harmonics and suppress the fundamental – but then the problem of false nulling by placing the high-pass 'brick wall' where it cuts out harmonics clearly arises.

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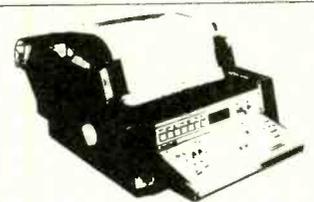
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ENTER 54 ON REPLY CARD

OS-9 for new euro-schools computer

A new 'European Education Standard Microcomputer' being developed by Thomson, Olivetti and Acorn, and expected next year, will use the OS-9/68000 operating system. According to Paul Dayan, technical director of The Soft Centre, UK distributors of Microware OS-9: "UK experience of school micros indicates that total European sales of the new machine, which will also double as a personal computer, could well be in the millions. Thus, we expect that a whole generation of European school-children will become familiar with the features and applications of OS-9, strongly influencing the adoption of this 'universal' operating system."

OS-9 is also implemented on the Atari 520ST and 1040ST computers and is to be adopted by Philips and Sony as the basis for the compact-disc interactive standard (CD-1). Says Paul Dayan: "These new licences are consolidating OS-9 in its position as the natural choice for 68000-based machines, from industrial VME board-systems to personal computers and beyond. OS-9 has the advantages of being modular, generally requiring less memory than alternatives and it is romable in many applications. In addition, it is a real-time system with multi-user facilities, which provide a Unix-style environment at a fraction of the cost. We claim that OS-9 is the operating system with the widest range of possible applications." The operating system is available with powerful languages and tools, including Microware Basic, Pas-

cal, C and Fortran, plus a growing selection of third-party applications programs. Since its introduction by Microware in 1983, OS-9 has been licensed by almost 300 manufacturers for use in a variety of industrial, scientific and consumer products, personal computers, industrial control systems, data processing equipment and telecommunications.

Machine vision at BAe

Built as a research project at the British Aerospace Sowerby Research Centre, Visive is a demonstrator that simulates aspects of biological vision. It is a hard-wired computing device for converting images from a sensor observing a natural scene into a digital data format suitable for processing by computer. The object of this research is to develop machine vision systems of general utility, capable of working in real-time, which are not subject to the restrictive operational conditions, and therefore limited applications, of current 'pattern recognition' systems. Biological vision is the model on which the design of Visive is based. Many of the principles involved in human vision can be applied in machine vision. The eye is not a particularly effective sensor, but human visual perception is good because the brain can interpret the image presented to it.

Analysis of the image can be divided, broadly, into three areas: imaging and formatting of raw data; manipulation of the data to provide a basis for interpretation; and interpretation of the information to initiate some useful action, for example, track-

ing a target, navigating a vehicle, inspecting a component, orienting a robot and the like.

Visive is designed to extract and format the image data and produce information at a resolution in the order of 10 times the sampling resolution. Video input is used in experiments with Visive. The input is from a c.c.d. (charged coupled device) camera that provides a digitized video output at the rate of 30 frames/s of an actual scene. Each frame comprises a 256 by 256 pixel matrix.

The selected image supplied to Visive enters its image input buffer store after first being converted into an 8-bit, 128 by 128 hexagonal pixel matrix. This is the spatial resolution Visive is designed to handle. Image data is stored in a fast ram array. From here the image data is relayed to a group of microprocessor computing units at a rate of 10 frames/s.

Algorithms used in these units have been developed from the principles underlying biological vision to assess the tonal variations between pixels to identify edge points, to determine the relative strengths and orientations of these edge points, and then to associate contiguous edge points with one another to establish the boundaries of objects. Boundary data is then sent to Visive's image output buffer where it is available for display on a video monitor, or on to further computing units for additional analysis and classification. Considerable computational power is required for scene analysis. A general-purpose computer could be programmed to perform the necessary computational tasks but the resulting system would be slow. Visive operates in real time, is hard-

wired and performs many of the data processing operations in parallel to ensure an adequate speed of response.

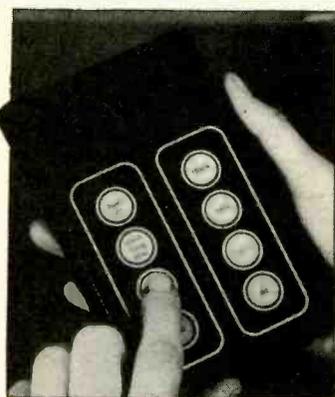
2nd generation (digital) cellular radio

Research into mobile telephone networks using digital technology promises improvements in system capacity and offers hope for a major manufacturing opportunity for UK industry. Early results from two experimental systems have been presented to the Nordic Mobile Radio Conference in Sweden.

The UK research is into a Europe-wide second generation cellular radio network destined to be introduced in the early 1990s. The creation of a European market is expected to further reduce the cost of cellular equipment. The research is being co-ordinated by the Cellular Radio Advisory group (CRAG), which is chaired by the DTI and includes representatives from British Telecom, GEC/Marconi, Philips, Plessey, Racal and STC. A spokesman for CRAG said: "In a very short space of time the UK has come a long way in cellular technology. We are now in a position to capitalize on our research expertise and secure a major role for the UK as a manufacturing base for digital cellular radio equipment." CRAG's work forms part of the UK contribution to the Conference of European Posts and Telecommunications' (CEPT) work on the development of a new standard for a European-wide cellular radio network. The CEPT has established a working group - the Group Spécial Mobile (GSM) - to co-ordinate the standards development work. The UK is represented on the GSM by the DTI, British Telecom, and Racal, who jointly sponsor Bernard Mallinder, recently appointed to lead the GSM's permanent team in Paris.

Research into second generation cellular equipment includes the following:

1. A joint digital cellular radio "Test Bed" experiment being developed by BT at Martlesham, GEC/Marconi at Baddow, and Racal at Reading. Jointly funded



Pocket Speech Aid is a small battery-powered immediate-access device with a natural sounding speech output. Developed to assist the communications requirements of the non-vocal, eight or sixteen phrases are available for selection, and male or female voices can be recorded with preservation of dialect and accent. Successful use of such a device requires a clear identification of situations where immediate speech production is essential, for example answering the telephone and attempting to initiate or control conversations. Pocket Speech Aid was developed by the West of Scotland Health Boards' Department of Clinical Physics and BioEngineering and is available from Nuclear Medical Electronics Ltd, 2 Hutton Square, Brucefield Industrial Park, Livingstone, West Lothian EH54 9BX, Tel: 0506 411974. Phrases are recorded using the adaptive differential p.c.m. compression technique described on pages 54-58.

by the DTI and the research partners, the experiment uses 125 and 312 kbit/s digital radio links with 16 kbit/s speech coding, and is specifically aimed at investigating the performance of such a narrow-band digital system. This experiment is now yielding results in representative urban and rural locations.

2. An experiment, now operational, being undertaken by Plessey Electronic System Research at Romsey on behalf of the DTI and BT. Supported in their work by the University of Southampton, this digital radio link experiment is aimed at quantifying the problems related to transmitting at different bit rates – from 100 kbit/s up to 4Mbit/s, and in particular seeks to identify the optimum degree of channel multiplexing.

3. Plessey were also commissioned by the DTI and BT to produce a top-down study analysing the alternative digital system proposals presented to GSM. The work examined the performance of these systems against the criteria established by GSM for second Generation system specification, to include: system capacity; base station cost; mobile cost, speech quality, ability to support hand portables, ability to support new (e.g. data) services. This report, representing the first comprehensive analysis of the various approaches presented to GSM, showed that while none of the systems held significant advantages in capacity, notable differences were revealed in the complexity and development risk factors between narrow band (low risk) and broad band (high risk) digital system.

4. GEC Research at Wembley have been co-ordinating a team of v.l.s.i. and system engineers from UK industry to examine the technological aspects of system production. Examination has focussed particularly upon the impact of cost and power consumption on a hand-portable product. The report shows that the complex v.l.s.i. necessary to meet the specifications can be available by the early 1990s. Meeting the prime criteria of portable units costing no more than first generation cellular equipment will be possible if production volumes consistent with a European wide market can be achieved.

5. STC has produced a report on the system requirements stemming from competitive operation of cellular networks within one country. At present the UK is unique in Europe (with a minor exception in Scandinavia) in operating its cellular network with two competitive operators. STC were also responsible for co-ordinating the selection of a 16-bit/s speech coder as the UK candidate to a GSM selection process taking place this autumn. As a result, a codec developed at BT Research Laboratories, Martlesham, has been offered to GSM for the European evaluations to be carried out in Turin later this year.

Landing on a microwave

A P-SCAN (Phase-Scanned Commutated Array Network) microwave landing system (m.s.l.) has been installed at London Heathrow Airport by the United Kingdom Civil Aviation Authority. The system will be used for trials in support of the work programme being undertaken for the International Civil Aviation Organisation (ICAO) to gain experience of m.l.s. in a busy airport environment. This forms part of an extensive UK technical and operational evaluation of m.l.s. Two Plessey P-SCAN systems have been delivered to the UK CAA and technical trials have already commenced at the Royal Aircraft Establishment, Bedford. m.l.s. is scheduled to become the preferred ICAO "standard approach aid" in 1998 but a considerable amount of work remains to be undertaken by ICAO and its working panels; particularly the All Weather Operations Panel, into technical and operational aspects of both the ground and airborne equipment. Significant emphasis is being placed on developing the design and test criteria for the high integrity Category II and III systems that provide the vital guidance to aircraft landing in conditions of poor visibility. This work will require a major appraisal of the integrity of computer-based software systems. The UK CAA is an acknowledged world leader in high integrity landing system development and pioneered the first use



This French medal commemorates the first regular broadcast service. It shows General Perrie, who was in charge of transmissions in 1921 and on the reverse, Sacha Guitry and Yvonne Printemps, who sang at the official inauguration of the system at the Eiffel Tower, February 6th, 1922. Copies are available in bronze (about £13) or silver (£178). Details from M. Jaques Campet, Le directeur, Monnaies et Medailles, 11 Quai de Conti, 75270 Paris 6me, France.

of fully-coupled automatic blind landing, using the STAN 37/38 i.l.s. which was also supplied by Plessey. This extensive experience had led to much emphasis being placed on the need for high system integrity.

The m.l.s. to be installed at Heathrow Airport comprises separate azimuth and elevation transmitter systems. The azimuth transmitter scans a two degree beam across a sector 40 degrees either side of the runway centreline. The elevation transmitter scans a one degree beam from 0 to +15 degrees. A British Airways Boeing 757 shuttle aircraft will be fitted with m.l.s. receivers and data recorders. The shuttle aircraft was chosen because of its frequent service to Heathrow which will maximize the amount of data collected. Secondary surveillance radar transponders and photographic techniques have also been developed by the CAA to establish the exact position of the aircraft on final approach. m.l.s. guidance signals will be displayed to the pilot and, in a later phase, coupled into the aircraft's flight control system for use only during clear weather conditions. At all times the existing instrument landing system (i.l.s.) facility will be displayed to the pilot and used as the primary landing aid.

It is planned to retain the m.l.s. at Heathrow Airport for at least twelve months before undertaking further operational trials at other international airports such as Manchester.

Plessey has worked closely with the UK government over the last two decades in evolving the new microwave landing system. Since 1978, when ICAO chose the Time Reference Scanning Beam as the m.l.s. standard, Plessey has been incorporating many new techniques into the P-SCAN m.l.s.. The techniques involved enable the generation of very precise scanning beams and also allow high integrity monitoring.

Quality a.m. radio

Readers of J.L.Linsley Hood's article in October's issue may be interested to know that specially selected i.f. coils are available from Hart Electronics. A set of six cost £2.80, whilst a set of oscillator coils for three bands costs £1.74. Semiconductor devices for the tuner section total £6.14. All these prices include vat and postage.

Hart say they will have a complete kit ready by January, including p.c. board. See advertisement on page 113 for address.

2D Fourier transforms. We omitted to point out that software to accompany the article is available on disc, together with programs that accompany previous articles on Fast Fourier transforms. Please send formatted disc indicating whether 40 or 80 track, together with return postage, marking your envelope 'FTT/Omer'.

Random-access phrase recorder

High quality compressed speech data is easily recorded and replayed on a BBC microcomputer using this speech digitization unit.

A.L. EVANS AND J. FENNER

The sequential nature of tape recorders limits the therapeutic and educational techniques that can be utilized. A random-access tape recorder would allow a variety of approaches to be tried, though rapid access of sequential tape/cassette recordings is expensive.¹

A useful system can be implemented by using a BBC B microcomputer and storing the speech data on floppy discs, though the impracticality of storing huge files on disc makes some form of data compression desirable. It can also be used to record a.d.p.c.m. data for prom-based synthesizers.

SYSTEM DESCRIPTION

The system components are shown in Fig.1, in which adaptive differential pulse code modulation, a waveform coding technique, compresses the data by factors of two to four. All units are standard, readily available items apart from the speech digitization unit, the design of which is described in this article. Data is exchanged between the digitization unit and computer memory via the 1MHz bus. Power is supplied via the 5V line in the analogue port, leaving the auxiliary power socket free for the disc drive. The system is

controlled by means of a touch pad (concept keyboard) which allows the recording or playback of a large number of files of data, limited only by the disc filing system (31 for the Acorn DFS). Each file can hold as much data as is allowed by the computer memory, the data capacity of the disc is soon approached unless an 80-track drive is used. In practice, a useful system has been implemented with the concept keyboard allowing a choice of nine files, each containing the speech data corresponding to a word or a short phrase. A small area of the touch pad is reserved for controlling the record or play-

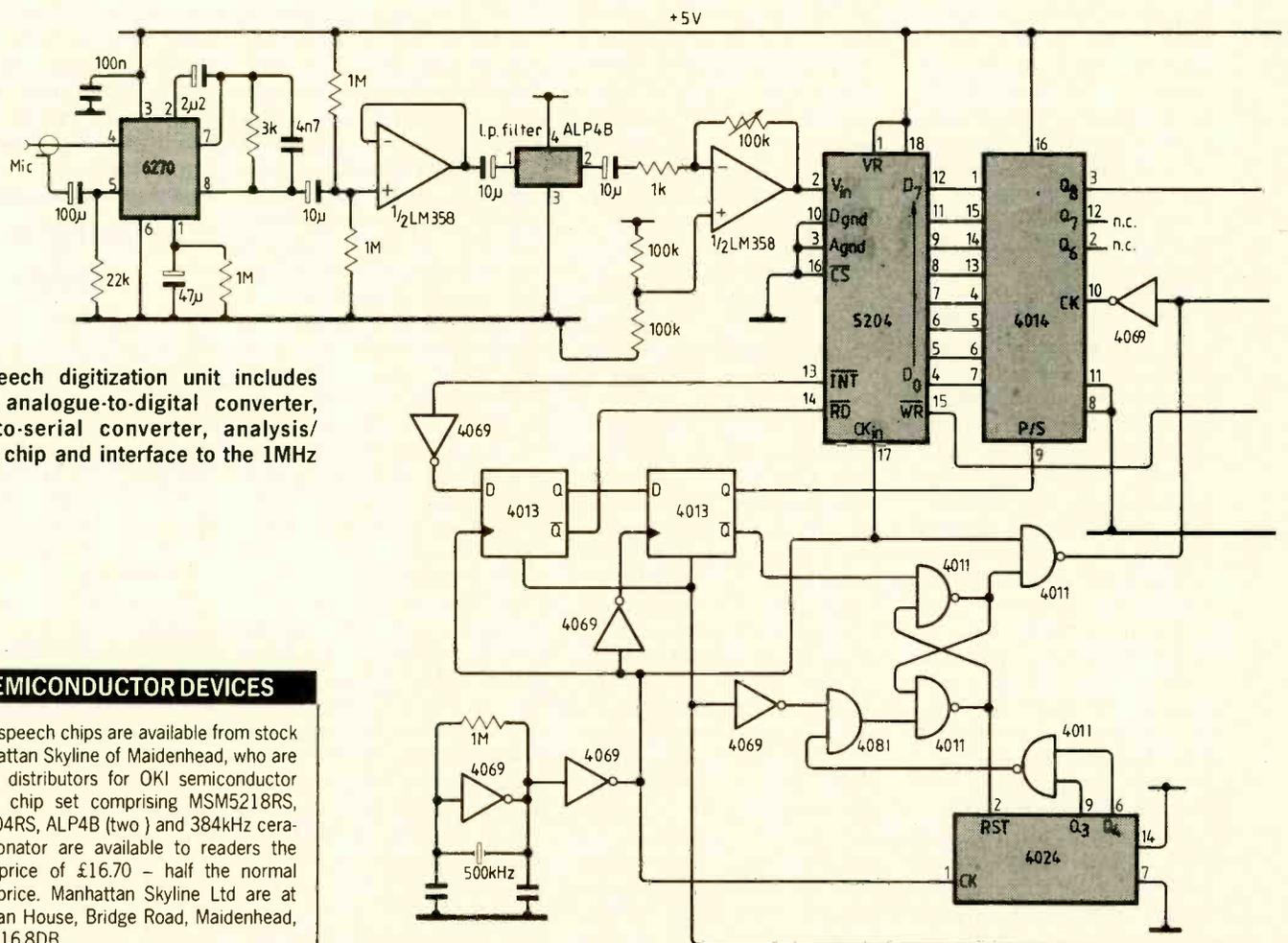


Fig.3. Speech digitization unit includes eight-bit analogue-to-digital converter, parallel-to-serial converter, analysis/synthesis chip and interface to the 1MHz bus.

OKI SEMICONDUCTOR DEVICES

ADPCM speech chips are available from stock at Manhattan Skyline of Maidenhead, who are sole UK distributors for OKI semiconductor parts. A chip set comprising MSM5218RS, MSM5204RS, ALP4B (two) and 384kHz ceramic resonator are available to readers the special price of £16.70 - half the normal one-off price. Manhattan Skyline Ltd are at Manhattan House, Bridge Road, Maidenhead, Berks SL16 8DB.

back mode – a prompt appears on the screen so that the user is aware of the currently selected mode. Writing the data file to disc takes one to two seconds and if the selected work or phrase is not currently in memory, there is a similar delay while data is read from the disc before speech starts.

SPEECH DIGITIZATION UNIT

Direct digitization of speech waveforms at 8kHz only allows some three seconds' worth of data to be acquired in the 32K memory of the standard BBC B microcomputer (allowing a modest 8K for program storage). To gain more recording time, the adaptive differential p.c.m. technique implemented by the OKI company in their MSM5218 chip is used to give a data compression factor of two. In addition, the system allows software selection of sampling frequency, implementation of analysis and synthesis on the same chip, easy interfacing to an input/output port, and low-power c-mos circuitry.

In the digitization unit, Fig.2, the audio input from the microphone is buffered, filtered and amplified before entering the eight-bit analogue-to-digital converter (MSM5204). The data converter has a built-in sample and hold function, and is controlled by the start-conversion signal of the analysis/synthesis chip. Data is transferred to a parallel-to-serial converter (4014) before being presented to the 5218 as eight-bit serial data. Since the 5218 expects a 12-bit

data input, the least significant four bits are padded out as additional zeros using hard-wired circuitry (4024 and 4011 on the full circuit diagram in Fig.3). The analysed data is transferred to memory from the data pins D₀-D₃ of the 5218. Figure 4 shows the timing diagrams for the data transfer to and from the microcomputer via the 6821 peripheral interface adapter. Port A of the 6821 p.i.a. is reserved for bidirectional data transfer while port B dedicated to control the 5218 chip functions. The master timing signal V_{CK} from the 5218 is detected via CB₁ pin of the p.i.a. The p.i.a. was addressed by the computer at &FCEX using minimal decoding logic (Fig.3).

Data returns from the computer again via

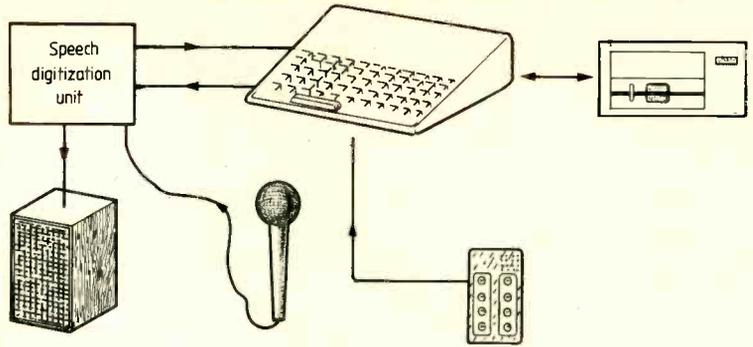
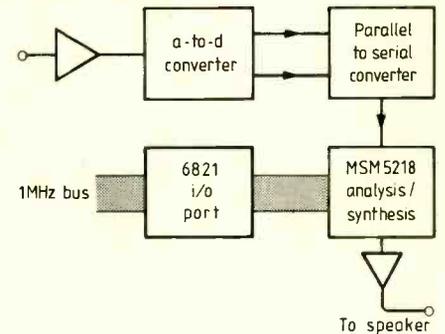


Fig.1. In the random access phrase recorder, all units are standard except the speech digitization unit which uses a.d.p.c.m. to compress the speech data.

Fig.2. Speech digitization unit is based on the OKI MSM5218 a.d.p.c.m. analysis/synthesis chip which compresses the digitised speech waveform before sending it to the computer. The same chip also decodes the data and gives a synthesized waveform as an analogue output.



the 6821 p.i.a. The 5218 is placed into synthesis mode by taking pin 6 low (port B bit 3). The V_{CK} signal from the 5218 still controls the data transfer (Fig.4). An analogue output is produced at pin 18 and after low-pass filtering the signal enters an audio amplifier before the speech is generated by an external speaker.

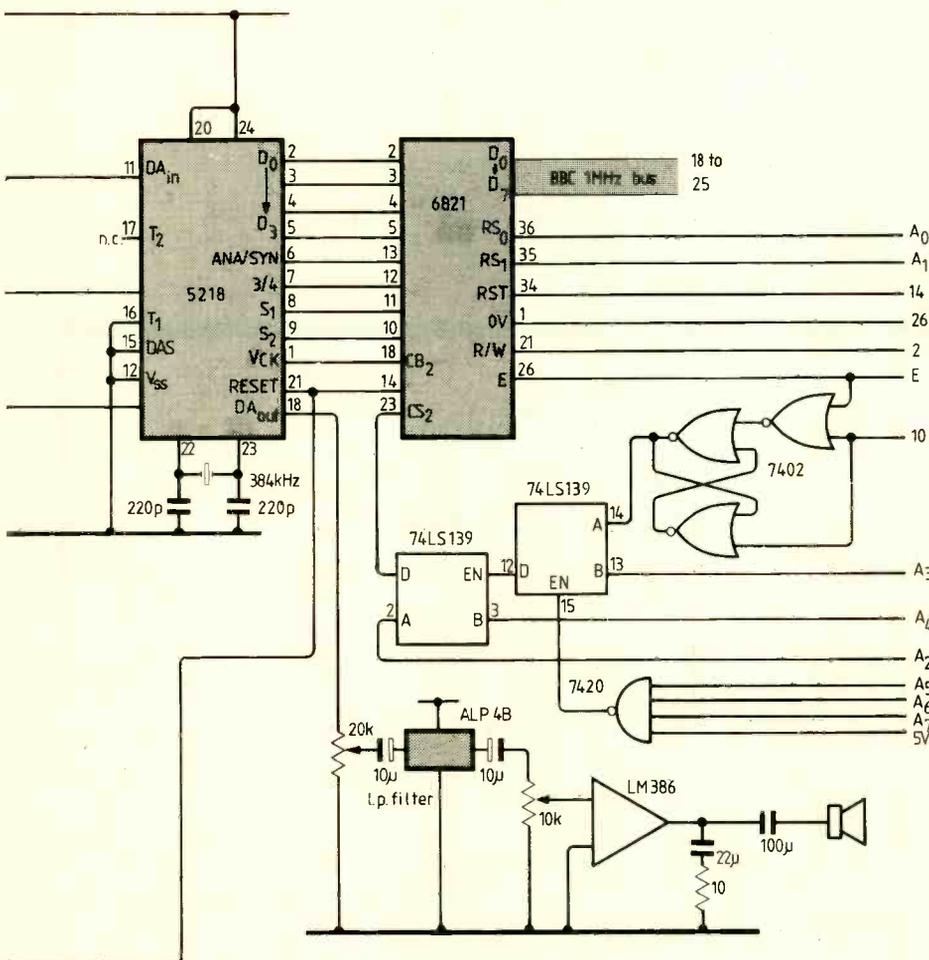
The components used in the speech digitization unit are readily available (OKI chips, active filters and resonator from Manhattan Skyline, Bridge Road, Maidenhead, Berks SL6 8DP). Component costs, including interconnecting plugs, cable and the enclosure, are below £55.

The quality of the speech produced is good, its intelligibility being close to high quality recorded speech and significantly better than l.p.c. speech in a similar environment².

SOFTWARE

The programs are written in Basic with machine code routines controlling the acquisition and playback of data. The main program is described by the flow diagram of Fig.5. The system waits for a touch-pad key to be selected and then branches to the routines controlling the data transfer.

If recording of speech data is selected the acquired routine is entered. All interrupts are disabled so that data acquisition occurs at a predictable rate. After the p.i.a. has been initialized, the system is synchronized with the V_{CK} signal from the digitization unit and data acquired until the recording is terminated by the touch pad. When the allocated



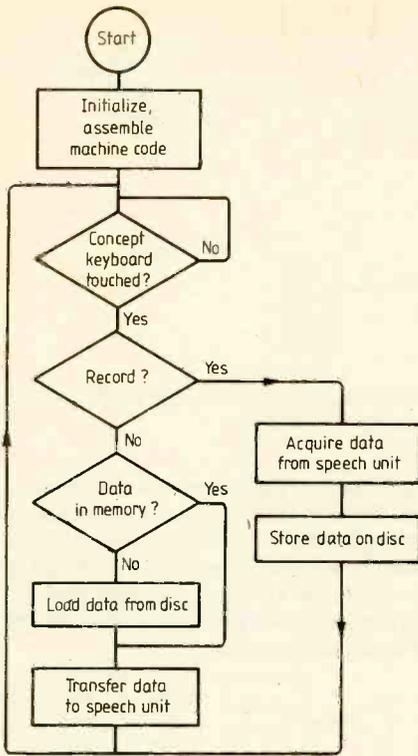


Fig.5. Flow diagram of Basic program controls the recording and playback of speech data.

section of memory is full, the data starts overwriting the same section so that using a pointer allows the most recent six seconds of speech to be retained. When acquisition is terminated, the program adds a stop code, re-enables interrupts and returns to Basic so that the data may be transferred to disc.

In a similar way, when playback is selected, the program initially decides whether the desired speech is already in memory and if necessary reads the data in from a disc file. Again, all interrupts are disabled before the system initializes the p.i.a. for transfer of data to the digitization unit. The system detects the V_{CK} signal from the 5218 and transfers the data according to the timing diagram of Fig.4. When the stop code is detected, interrupts are re-enabled and the system returns to the main program.

The system's versatility has been extended by providing an overlay editor which allows for example, therapist to design an overlay on the touch pad (see below). Thus the user has control of the number and size of sensitive elements on the touch pad. The overlay editor can be entered at the beginning of the main program but the program defaults to using the current overlay after a suitable time delay.

Software. Listings of the program controlling the acquisition and replay of the speech data, and for the overlay editor, together with the p.c. prototype layout are available from the editorial office in return for a self-addressed A4 envelope, marked 'phrase recorder'.

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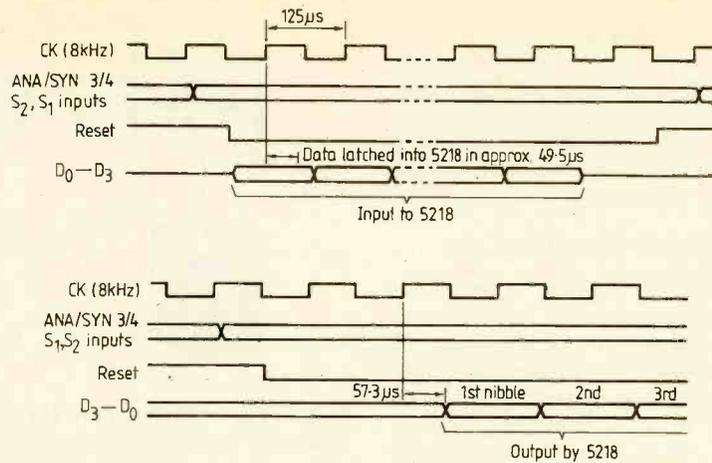


Fig.4. Timing diagrams for the data transfer between the MSM5218 and microcomputer, (a) synthesis (b) analysis. The MSM5218 provides an 8kHz clock to control the data transfer.

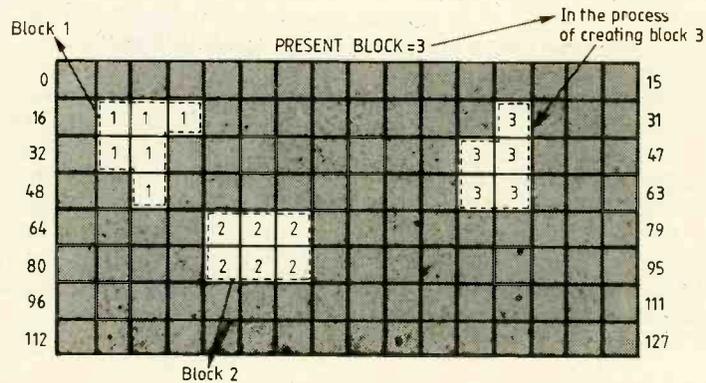


Fig.6. System has been made 'user-friendly' by using a touch pad to control the recording and playback. A program is provided to edit layout of the paper overlay.

References

- OKI Semiconductor Application Note. Simultaneous speech analysis and synthesis with MSM5218, 1982.
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- 2. Keating, D., Evans, A.L., Wyper, D.J. and Cunningham, E. Comparison of the intelligibility of some low-cost speech synthesis devices. *British Journal of Disorders of Communication* (vol.21), 1986, pp167-172.

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There exist many applications in which outputs exceeding 100 volts and not infrequently output current considerably in excess of 25 milliamperes are required, with a slew rate of several hundred volts per microsecond and a large-signal gain-bandwidth product of several megahertz. It is often possible to obtain commercial amplifiers employing thermionic valves to satisfy such requirements, but these are usually expensive.

To design a single amplifier capable of meeting such a wide range of requirements would be a formidable task and, instead of this, a design is suggested in which each stage is unilateral and essentially independent of the following one. This method renders the design of an amplifier to meet a specific requirement systematic and relatively simple. The suggested designs are entirely solid-state though discrete, and offer as economical a solution as is possible within the current state of semiconductor device development.

DESIGN PRINCIPLES

The block diagram of the amplifier is shown in Fig.1. Each stage is designed to be unilateral for all practical purposes, and the output of each of the first two is a high impedance current source, so that each of these stages is characterized by a transconductance and an input admittance. The circuit branch consisting of capacitance C and resistance R in series shown shunting the input of the intermediate stage is a compensating network which will be discussed later. The input and intermediate stages normally require only the 15 volt power supplies which operational amplifiers commonly need.

The output stage is designed to drive the load and may take several forms, dependent upon the performance requirements. It is only this stage which requires high-voltage

supplies which are chosen to suit the application, but at the time of writing must satisfy the condition $V_1 + V_2 < 1\text{kV}$.

Input and intermediate stages. The circuit employed for the input and intermediate stages is shown in essence in Fig.2. In a balanced quiescent state, the achievement of which is assisted by resistors r, each of the six transistors will have an intrinsic transconductance $g_m = 20\text{I}$. The effective transconductance of Tr_1 and Tr_2 will be modified by the presence of resistance r to $g_m' = g_m / (1 + g_m r)$, so that the greatest voltage gain from an input base to the corresponding collector is $-1 / (1 + g_m r)$, whose magnitude is less than unity. Thus the Miller effect at the bases of Tr_1 and Tr_2 is reduced.

The admittance seen at the bases of Tr_1 and Tr_2 is also decreased by the presence of r. Since $y_{\pi}' = y_{\pi} / (1 + g_m r)$ the effect is to increase r_{π} and to decrease C_{π} in the same ratio, so that the input resistance is increased. The intermediate stage is current driven and its bandwidth will be determined by the time constant $(C_{\mu}' + C_{\pi}') / (G + g_m')$, which has been decreased by the presence of r. Thus another effect of r is to increase the bandwidth of the intermediate stage. In designing the overall amplifier circuit, the values of r in the input and intermediate stages are chosen to give the required value of open-loop, low-frequency gain.

A further advantage of the use of current mirrors in the input stage is that it enables a large common mode signal to be accommodated. If the current mirrors employ discrete single, rather than dual transistors, their current ratio may be improved by adding

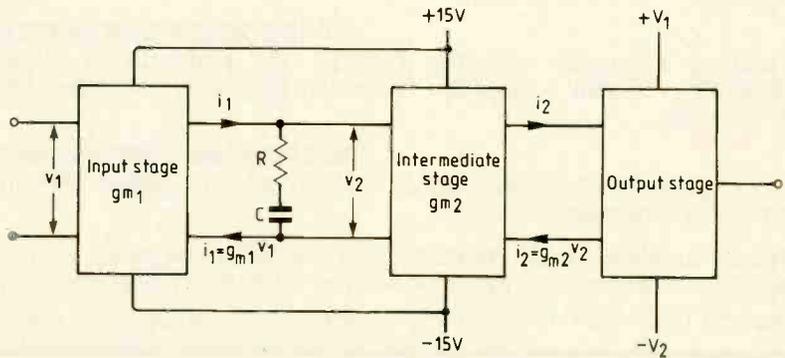


Fig.1. Block diagram of the amplifier. Each stage is considered separately, the output stage being varied to suit the application.

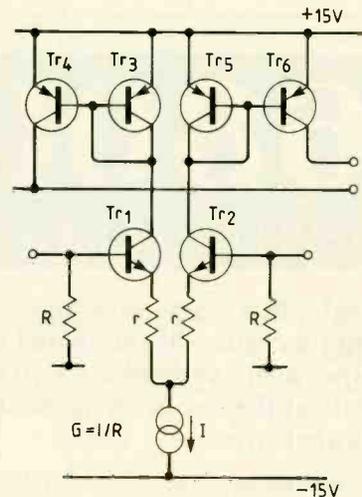


Fig.2. Circuit of input and intermediate stages. Current mirrors allow large common-mode signals.

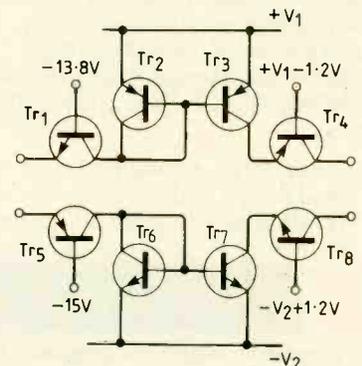
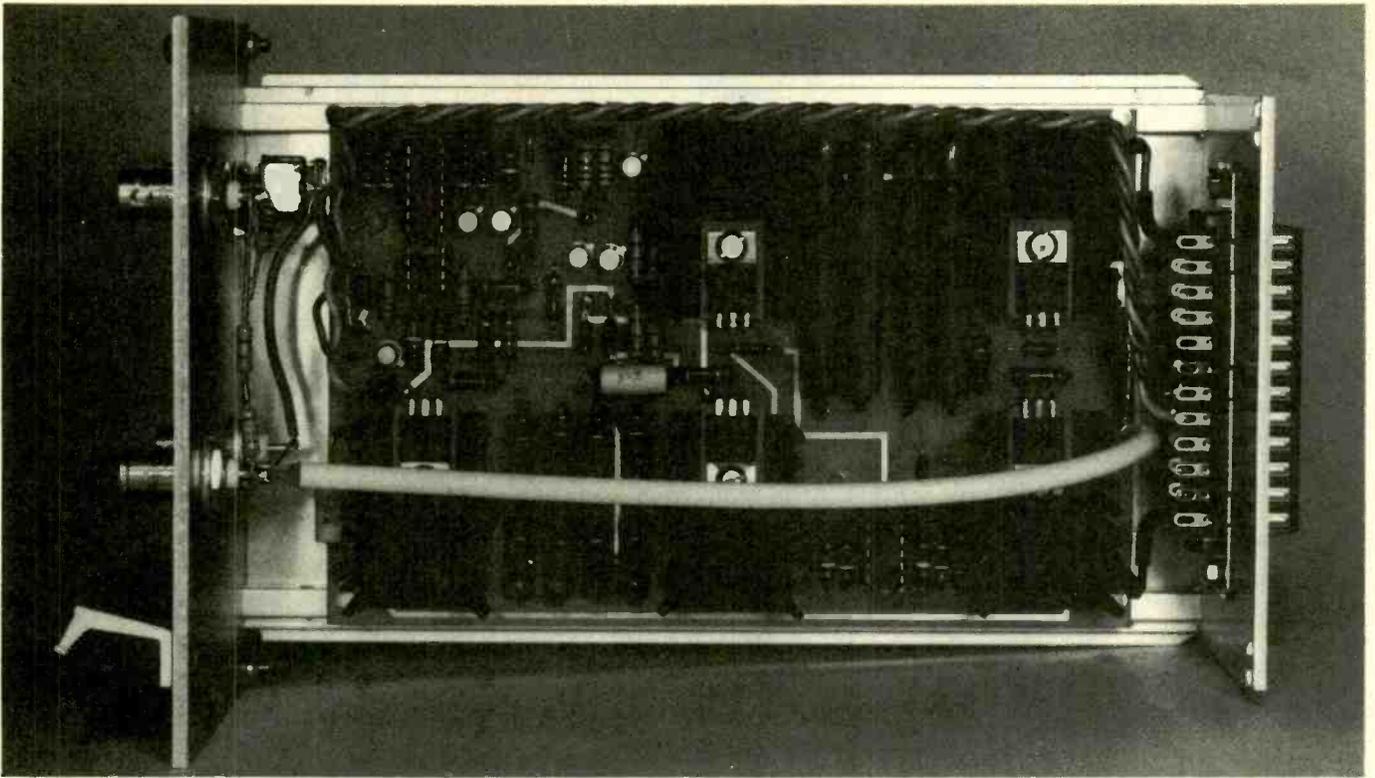


Fig.3. One version of input to final stage.



identical resistors in series with each emitter. These resistors should not be of too large a value since they will increase the Miller effect in the emitter coupled pair.

Output stage. One of a variety of circuits may be used for the output stage, depending upon the particular application of the amplifier. The type of device used in this stage is determined by the output voltage and current required by the load. For voltages up to 350 and currents up to several tens of milliamperes, high-voltage video-amplifier bipolar transistors may be employed in the common-base mode, simply to withstand the required voltage, whilst passing current from the intermediate stage to the load.

Where the required voltage exceeds 350 or the current is greater than several tens of milliamperes, power fets are more suitable because of low current gain in bipolar power transistors, or the lack of p-n-p devices, or both. These may be used either in the same way as the video transistors, or as source followers in order to reduce the standing power dissipation in the amplifier. The output stage may be either single-ended or push-pull, again depending upon the performance required. At the highest voltage rating currently of 1kV, only n-channel fets are available and, if a push-pull system is needed, a type of totem-pole amplifier becomes necessary.

Whatever arrangement of devices is used in directly driving the load, the complete output stage is designed to be current driven by the intermediate stage. The input section of the output stage will usually contain transistors in the common base configuration, at least one of which will serve to translate the d.c. level of an input current to that of one of the high-voltage power-supply lines.

Figure 3 shows a typical version of the

input section of the output stage. In this example it is assumed that the intermediate stage uses p-n-p transistors so that its output current may be at the $-15V$ level. One of the intermediate stage collectors connects directly to Tr_5 , a high-voltage transistor whose function is simply to transmit the current from one output of the intermediate stage of the current mirror formed by Tr_6 and Tr_7 and thence to Tr_8 , which is again a high-voltage transistor passing the current to the final driver stage, or in some cases directly to the load. Transistors Tr_1 to Tr_4 perform a similar function for the other output from the intermediate stage. The bias voltage of $1.2V$ which occurs in three places in Fig.3 is provided to ensure that the output transistor of a current mirror is not saturated.

The simplest form of output stage has the collectors of Tr_4 and Tr_8 connected together to form the output terminal, the load being connected from this point to ground. This form is suitable for applications in which the current from the intermediate stage is sufficient to drive the load.

A complementary source-follower circuit is shown in Fig.4. In this case the current from the intermediate stage drives the resistor chain, and the source followers provide large currents on demand, enabling heavy loads to be driven.

Currently the highest rated p-channel power fet which is available will withstand 500V, so that for outputs of more than 500V the complementary source follower cannot be used. A circuit having similar properties but using only n-channel devices is shown in Fig.5.

Stability and transient response. The design of a feedback amplifier normally commences from a knowledge of the required gain and of the stability of this gain against circuit

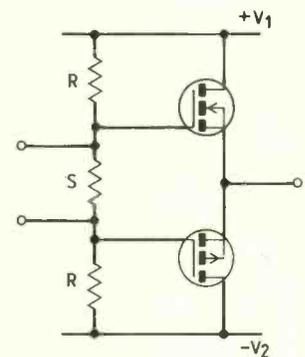


Fig.4. Complementary source follower driven by intermediate stage.

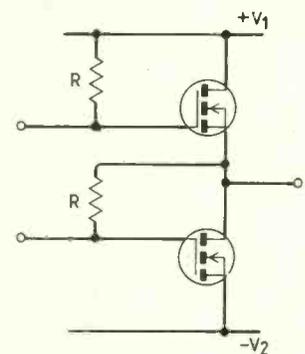


Fig.5. Circuit has similar characteristics to that of Fig.4, but at higher voltages.

parameter changes. These two data immediately determine the feedback fraction β and the low frequency loop gain $A_o\beta$, so that the required open loop gain A_o is known. A major advantage of the present method of design is that A_o can be adjusted to the required value, and this has a significant effect upon the problem of achieving a.c. stability and the desired transient response.

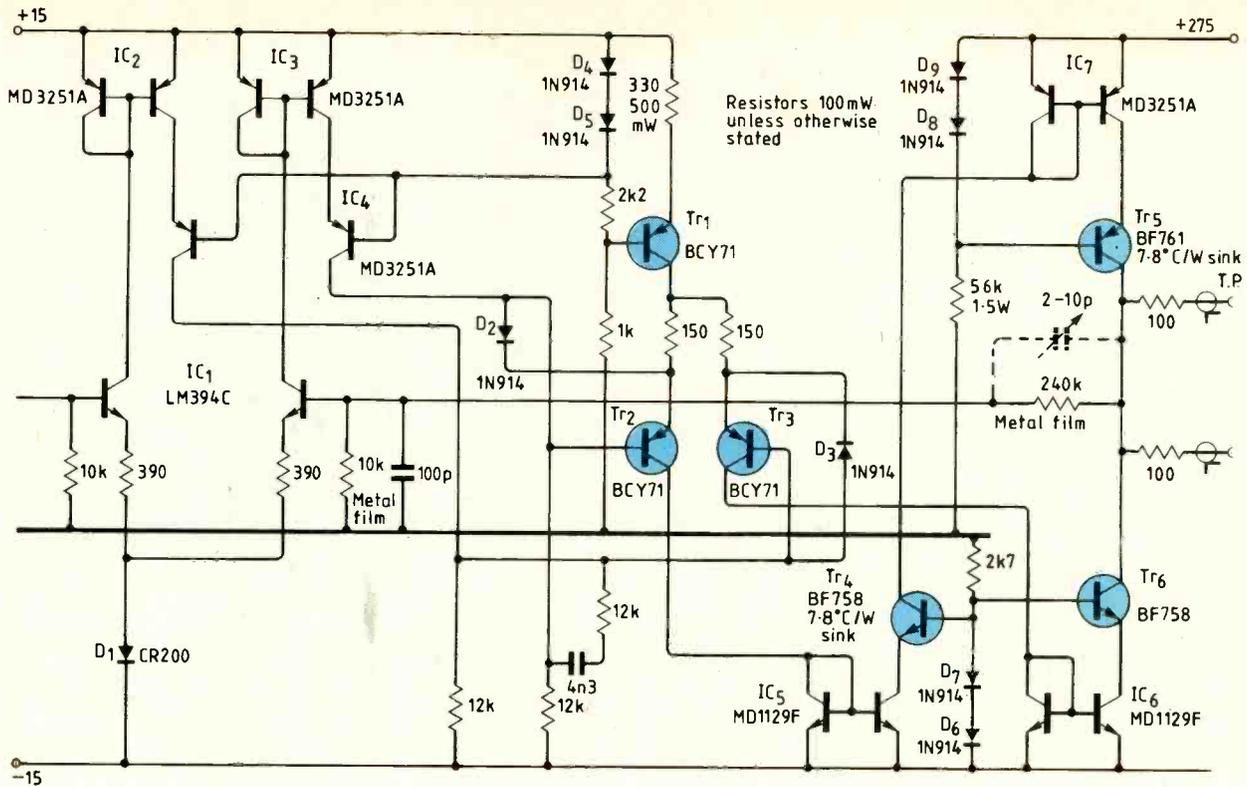


Fig.6. Complete circuit diagram of one version of amplifier.

To reduce phase shift in this circuit at high frequencies, transistors should be selected whose f_T is at least one and preferably more orders of magnitude higher than the desired 3 dB frequency of the amplifier. Transistors in the common-base configuration have a current gain which falls by 3 dB at approximately f_T and current mirrors behave in the same way as common-base transistors having transition frequencies of $f_T/2$.

The input stage will have a small input capacitance at each base reduced by the emitter resistors, and one of these capacitances will form a time constant with the feedback potentiometer. This time constant may be compensated in the usual manner in order to make β independent of frequency. Often in order to bring the compensating capacitor to a practical value it will be necessary to add capacitance at the base of the input stage.

Two situations now arise. If the load is current driven, effectively by the output of the intermediate stage, the feedback circuit will form part of the load and the overall time constant will be large compared to $1/\omega_T$. If on the other hand the load is driven by a source follower or a totem pole stage, the load time constant will become insignificant compared to that at the input of that stage. In either case one significant time constant will exist in the output stage. For example in the circuits of Fig.4 and Fig.5, the dominant time constant is $(C_{oss} - C_{rss})R$. A second significant time constant will usually arise at the input of the intermediate stage.

If the significant time constants are one or two orders of magnitude greater than the remainder, as is often the case, then these latter may be ignored at least to a first approximation, and the amplifier is essen-

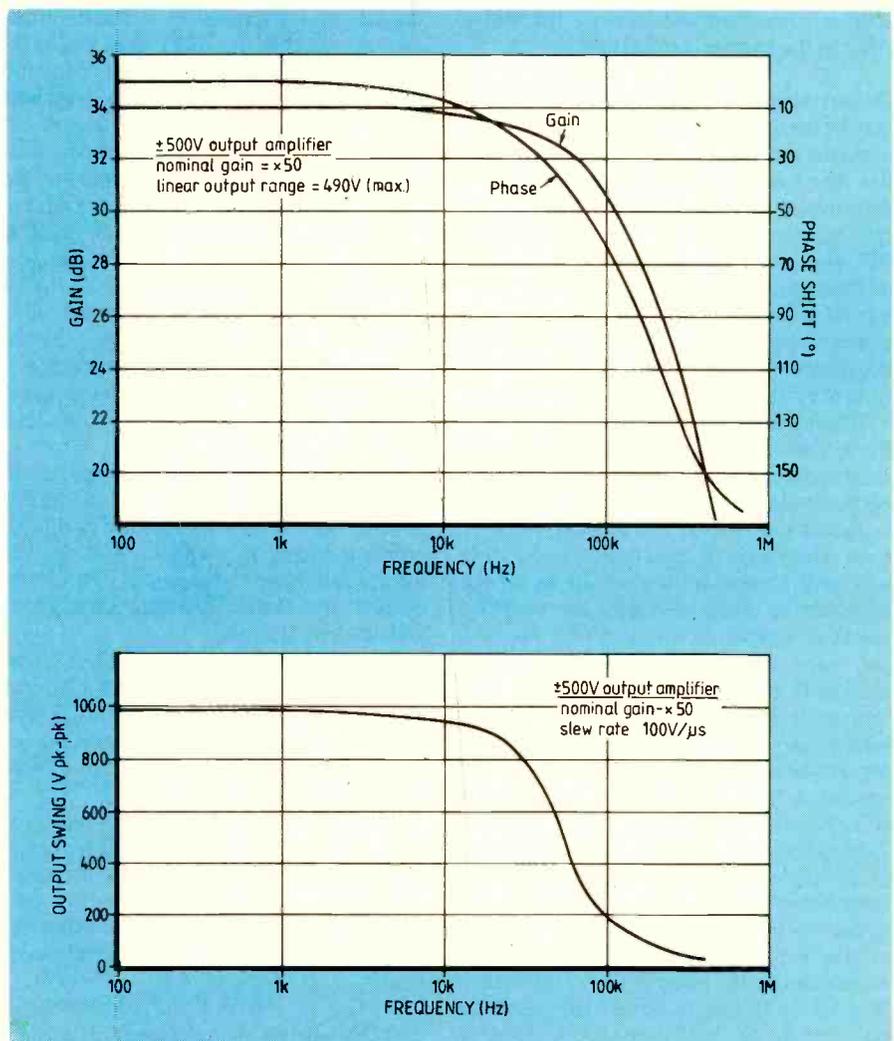


Fig.7. Characteristics of the design arranged to drive a capacitive load at 20kHz, with a gain of 50 and 1000Vpk-pk swing. This design shown in the photographs.

tially one with two lagging time constants. If $A_o\beta \gg 1$, then for a non-oscillatory step response it is necessary that the ratio of these time constants be at least $4A_o\beta$. The compensation circuit consisting of C and R at the input to the intermediate stage enables this ratio to be adjusted as shown in the appendix, so that an acceptable transient response may be achieved.

EXAMPLE OF METHOD

As an example of the design method suggested above, an amplifier to meet the following requirements is considered here. The load is a capacitance of 200 pF in which a non-oscillatory response of 250V is required to an input step of 10V. The output is to reach 99.9% of the final value within 2 μ s, and the gain stability required is 0.1%.

The suggested circuit is shown in Fig.6 and is of the type in which the load is driven by the current from the intermediate stage. Evidently we need $\beta = 1/25$ and $A_o\beta = 1000$, so that $A_o = 25000$. The total current in the intermediate stage must yield a slew rate of at least 125V/ μ s and thus must be at least $2 \times 10^{-10} \times 250 \times 1000 / 2 \times 10^{-6} = 25$ mA. The total current as designed is 28.5 mA.

The transistors BF758 and BF761 are video devices chosen to withstand 300V, and, with suitable heat sinking, the greatest mean dissipation of 4 W: their f_T is 45MHz and minimum h_{FE} is 40. The current mirrors have minimum values of $h_{FE} = 100$ and $f_T = 200$ MHz. The BCY71 devices chosen for the intermediate stage have values h_{FE} typical 200 and f_T minimum 200MHz, $C_{\mu} = 5$ pF.

Hence for the intermediate stage $g_m = 1/240$, $S = 4.17$ ms, $g_{\pi} = g_m/h_{FE} = 20.8 \mu$ s, $C_{\pi} = g_m/2\pi f_T = 3.32$ pF, and thence $C + g_{\pi} = 104 \mu$ s. The first significant time constant is hence $T_1 = (5 + 3.32) \times 10^{-12} / 104 \times 10^{-6} = 8 \times 10^{-8}$ s.

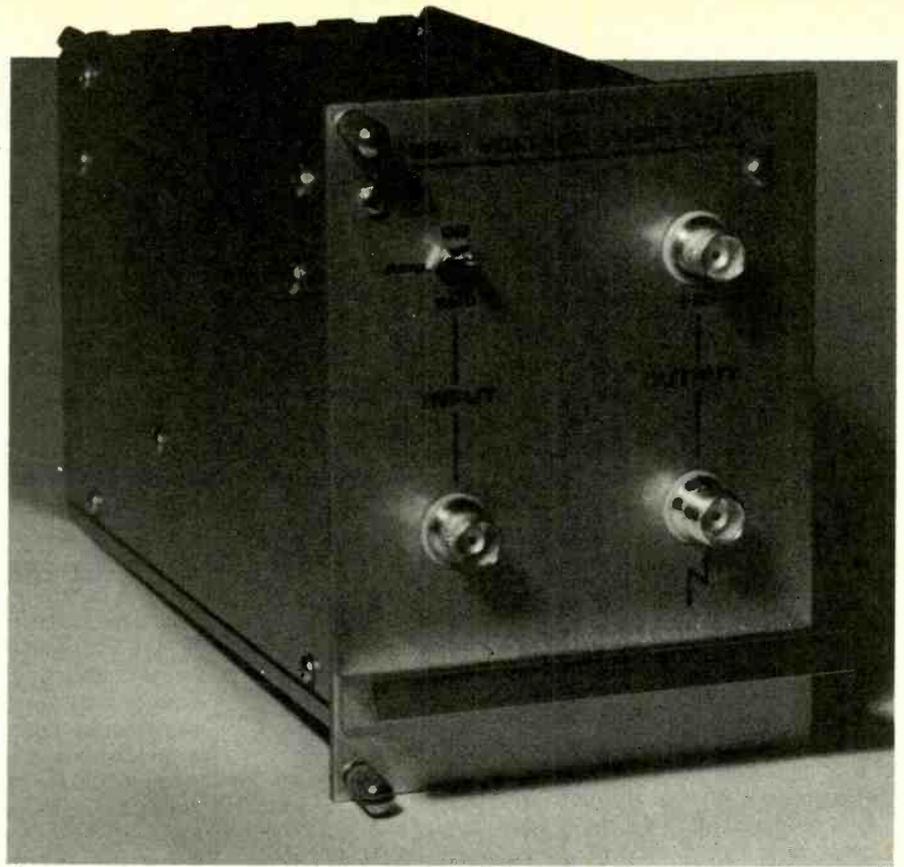
For the LM394C used for accurate balance, low drift and low noise in the input stage, $h_{FE} = 500$, $C_{\mu} = 15$ pF and $f_T = 60$ MHz. Hence $g_m = 1/390$, $S = 2.56$ ms, $g_{\pi} = 5.13 \mu$ s and $C_{\pi} = 6.8$ pF.

The input stage will have a differential mode signal only 1/25 of the common mode signal, so that the contributions of g_{π} and C_{π} to the input admittance may be ignored in comparison with that of C_{μ} . Thus the capacitance loading the feedback potentiometer is 15 pF, requiring about 0.6 pF compensating capacitance. This value is impracticably small, and to make it more reasonable 100pF is added at the feedback base of the input stage, bringing the compensating capacitance to about 4.6 pF.

In this circuit the time constant T_2 is that of the load and feedback network: $T_2 = (200 + 115/25) \times 10^{-12} / (10^{-4}/25) = 51.15 \times 10^{-6}$ s. Also $A_o = 4.17 \times 10^{-3} \times 2.56 \times 10^{-3} / (10^{-4} \times 1.04 \times 10^{-4}) = 25035$, $A_o\beta = 1001$, $T_2/T_1 = 639.4$. The ratio of the time constants is thus too small and correction is needed.

We require $CR = 51.15 \times 10^{-6}$, $T_3T_4 = T_1T_2 = 4.092 \times 10^{-12}$ with $T_3 = 4004T_4$, whence $C = 3.99$ nF for which we use the larger preferred value 4.3 nF, leading to $R = 12$ k. The equal time constants in the closed loop are then each $2T_4 = 6.4 \times 10^{-8}$ s.

With this choice of values the rise time to



99.9% of final value is $9.23T = 591$ ns for small signals. For signals greater than about 10% of maximum the output rise time is determined by the slew rate of 135V/ μ s, giving 1.85 μ s for 250V output, which is within the specification.

In this example some typical and some minimum parameter values have been used. Using minimum values of gain parameters will give assurance of adequate open-loop gain and hence of closed-loop gain stability. In practice, to deal with the question of transient response either a worst case calculation must be performed, or the values of compensation components may be found by trial in a give case, using the mean values as a starting point.

As a further example, Fig.7 shows some characteristics of an amplifier designed to drive a capacitance load of several hundred pF at frequencies up to 20kHz. A gain of 50 was required stable to 0.1% and an output swing of 500V peak was desired. It can be seen that a swing of 450V peak was achieved at 20kHz. The total noise output was 700 μ V r.m.s., which could be improved if necessary by attention to the input stage. This amplifier uses an output stage like that shown in Fig.5. Figure 8 is a photograph of this amplifier.

APPENDIX-STEP-RESPONSE CORRECTION

Suppose the amplifier has d.c. gain A_o and two lagging time constants T_1 and T_2 . The differential equation relating output V_o and input V_i when the loop is closed with a feedback fraction β is then $[A_o\beta + (1+T_1D)(1+T_2D)]V_o = A_oV_i$. The transient response is determined by the complementary function of the equation, whose nature is in turn determined by that of the roots of the auxiliary equation $A_o\beta + (1+T_1x)(1+T_2x) =$

0. For a non-oscillatory solution it is necessary that the roots shall be real and this requires that $(T_1+T_2)^2 > 4(1+A_o\beta)T_1T_2$ or $(T_1-T_2)^2 > 4A_o\beta T_1T_2$. Since in practice $A_o\beta \gg 1$, it follows that one time constant, say T_1 , must be much greater than the other, whence $T_1 > 4A_o\beta T_1T_2$.

Now suppose that the time constant T_1 is at the input of the intermediate stage, and is formed by capacitance C_1 and resistance R_1 in parallel. With the compensating series CR branch connected, the impedance at this point is $R_1(1+j\omega CR)/[(1+j\omega T_1)(1+j\omega CR)+j\omega CR_1]$ which takes the place in the gain expression of the impedance $R_1/(1+j\omega T_1)$ which obtains in the absence of the CR branch. The additional branch will thus lead to the gain expression $A_o(1+j\omega CR)/[(1+j\omega T_1)(1+j\omega CR)+j\omega CR_1](1+j\omega T_2)$.

If we now choose $CR = T_2$, this becomes $A_o/[(1+j\omega T_1)(1+j\omega T_2)+j\omega CR_1]$ which may be written $A_o/[(1+j\omega T_3)(1+j\omega T_4)]$ where $T_3T_4 = T_1T_2$ and $T_3T_4 = T_1T_2 + CR_1$. Thus the two time constants T_1 and T_2 have been replaced by T_3 and T_4 whose ratio may be adjusted by choice of C whilst keeping $CR = T_2$.

ACKNOWLEDGEMENT

The author would like to acknowledge the contribution of R.R. Thorp, M.A., senior design engineer in the Cambridge University Engineering Department, in the form of constructive criticism, for the design of the second amplifier mentioned and for Figs 7 and 8.

J. Barron, M.A., graduated from the University of Cambridge, subsequently researching in nuclear physics and electronics for eight years before entering industry. In 1967 he returned to Cambridge as a lecturer in the Department of Engineering.

FEEDBACK

Or not, as the case may be

I recently bought a hairdryer. As an integral part of the hairdryer a cable was supplied which is to be connected to a mains plug by the user. The instructions read as follows:-

IMPORTANT

The wires in this mains lead are coloured in accordance with the following code: BLUE:NEUTRAL BROWN:LIVE.

The wire which is coloured blue must be connected to the terminal marked with the letter N or coloured black. The wire which is coloured brown must be connected to the terminal marked with the letter L or coloured red.

Can any of your readers help me convince a member of the opposite sex that engineers really do think logically or are we all quite mad?

W. Scott,
Highgate,
London N6.

Heat transfer

I refer to a letter by Mr G. Nalty in the October issue in the Feedback column. I am afraid that Mr Nalty, like so many others, has been confused by Specmanship in his interpretation of the thermal performance of semiconductors and especially transistors. Mr Nalty claims to have heard the difference in distortion in an amplifier when changing from a power transistor rated at 12.5 watts to one rated at 50 watts due to lower junction temperature. I would suggest that the truth of the matter is that the difference was due to the actual transistor characteristics rather than junction temperature. Let us examine the facts.

Mr Nalty's circuit conditions are such that the device is dissipating 60 milliwatts. Presumably the transistor is mounted in free air for such a small dissipation. What is important in this condition is not the R_{th} junction to case (R_{thj-c}) but the R_{thj-a} (junction to ambient) which is considerably different and very much larger. Furthermore, this figure is in the main independent

of package for a given die size, although not necessarily true for all die sizes. It is easy to find devices in TO220, SOT82 and SOT32 (TO126) in the SGS Power Transistor Databook which use the same die and exhibit very different R_{thj-c} but the same R_{thj-a} . But let us be generous and use some figures which do vary, for example an R_{thj-a} of 62.5 degC/W and 83.4 degC/W which can be found for certain devices in TO220/SOT82 and TO126 respectively. We have here a difference of about 20 degC/W which for the quoted dissipation of 60 milliwatts will yield a junction temperature difference of a mere 1.2 degrees centigrade between the two devices. Taking this to its logical conclusion and using figures for more representative types we would have negligible temperature difference, certainly much less than the normal measurement tolerance capability, tending to prove my previous statement that the distortion difference is due to transistor (die) type alone.

Mr Nalty might like to note that it is not thermal resistance which decreases with increase in power dissipation, but rather the other way round. The R_{th} defines the maximum allowable dissipation with reference to the maximum allowable junction temperature. It is the first link in the chain from the die to the outside world, so to speak.

Let us look at this phenomenon a little further and take as an example the devices TIP110, TIP120, TIP130, SGS110, SGS120 and SGS130. The three TIP--- types use increasingly larger die assembled in TO220 while the SGS--- types are the equivalent die in SOT32 package, which is physically the same size as TO126 (or SOT32) but without the fixing hole, which thus enables a larger die to be accommodated. The R_{thj-a} is identical for all six types (regardless of die size or package type), but the R_{thj-c} is another matter. Die for die (e.g. TIP110 to SGS110) the R_{thj-c} is the same but it decreases from 2.5 through to 1.78 degC/W as the die size increases from 110 to 130. As a parallel to this the figure for 7800 regulators in TO220, which use very much smaller die, shows a R_{thj-c} of 4

degC/W.

I hope this sheds some light on a rather ill-understood topic. Nigel Pritchard, Field Applications Engineer, SGS Semiconductor Ltd, Aylesbury.

Filters and pulses

I was interested to read at the end of Mr C.F. Coleman's letter (July issue), the question "Why, if the Fourier transform of a single square pulse can represent the zero level before the pulse arrives, should the calculated output which it generates when applied to a low pass filter begin before the leading edge of the pulse arrives?"

I remember as a student (many years ago) being puzzled by precisely this anomaly. The answer which emerged after a great deal of effort, mathematics not being my strong suit, was that the ideal pulse and network postulated were not physically realisable. If one utilised in the calculation the characteristics of a pulse and a network which were physically realisable, then no output occurred before the input pulse was applied, and the effect was delay the rising edge of the output pulse with respect to the input pulse.

No doubt your more mathematically inclined readers could furnish a convincing proof.

J. Hollingworth,
Grand Caymen,
B.W. Indies.

S5/8

I regret that I am still unconvinced about the proposed S5/8 system, despite the enthusiastic flag waving by its proponents.

The worst point about this system is that it is not a properly designed transmission system; it takes no account of the properties of the transmission line, and it is not terminated. It is thus very prone to interference pick-up and ringing; a problem that must increase as line length is increased. It is obvious that lengths in excess of 1 metre will be required. Mr Tomlinson's

point regarding the t.t.l. parallel interface to disc drives is very valid. These are used only on short runs, but the cable is at least terminated!

I am not impressed with arguments about an extra pound or two for components, nor even an extra 100 mA or so. I am sure that anyone who has lost valuable programmes or text from a computer system would gladly pay the price for a reliable system. Such a properly designed interface need not tie up software time with CRC checks etc. These need only be used when unstable media such as radio link or magnetic recording is involved. A wired system should be as reliable as possible.

I would be very interested to learn what system is employed for consultation between BSI and the engineering community before standards are proposed. Clearly this proposal would have benefited by input from engineers in the television, radar or audio field where the points I have listed have been well known and understood for many years. Perhaps wider publication of such proposals in *Wireless World* might be advantageous.

Les Hayward,
Eastpoint Ltd,
Corfe Castle,
Dorset.

German v.l.f.

I am engaged on a programme of research into a very low frequency transmitting radio station established in Germany, late autumn 1941.

The station called "Goliath", was near the village of Kalbe (Saxony, Prussia) and partially straddled the River Milde.

I seek any information, no matter how small or seemingly trivial. Any drawings, photographs, technical details or personal memories by German or Allied personnel would be especially welcome. Any documents loaned to me would be treated with care and postage refunded.

T.F. Bernascone,
Goliath Research Project,
Teeside Polytechnic,
Borough Road,
Middlesbrough,
Cleveland TS1 3BA.

FEEDBACK

Relativity

In the October issue of *E&WV*, Alan Watson of Mallorca challenges sceptics to provide physical evidence against the second postulate on which Albert Einstein based his relativity doctrine. Perhaps I can help.

The postulate was presented in the second paragraph of Einstein's famous 1905 paper in which he introduced his Special Relativity. He wrote in German. Translated into English, the postulate reads:

"...light is always propagated in empty space with a definite velocity c which is independent of the state of motion of the emitting body."

Statements about motion and velocity are meaningless unless they specify (or unambiguously imply) a frame of reference. In later years, Einstein and his followers interpreted the postulate to mean: The velocity of light measures as c under all conditions.

Consider the radar equipment used by traffic police to catch people driving their automobiles too fast. Typical is the case in which a police vehicle sits at the side of a highway, aiming a radar beam at approaching traffic.

One person is driving toward the police vehicle at a speed (a velocity) of 100 km/h (relative to the road surface) in violation of the law. The radar beam bounces off the offending vehicle, bounces back to the police vehicle, and records the illegal land speed in the police radar equipment. The police officer signs a citation, and the offender pays a fine.

How does the radar work? The radar beam emitted by the police vehicle travels at c (nearly 300,000 km/s) relative to its source. It is reflected back to the waiting police vehicle. How fast does the reflected beam travel? At c , relative to the vehicle from which it is reflected.

So far, Einsteinians and anti-Einsteinians agree. But then we ask the critical question: What is the velocity of the incoming reflected beam relative to the police vehicle? It is only c , the Einsteinians reply. But the anti-Einsteinians say no, insisting that the returning beam comes in at a velocity of $c+100\text{km/}$

$h=c+0.02777\text{..km/s.}$

The radar equipment in the police vehicle shows the Doppler effect, a slight shift in the frequency of the incoming beam, and calculates the oncoming vehicle's illegal land speed of 100 km/h, which is equal to 0.02777...km/s. If the incoming beam's velocity were only c (as Einsteinians contend), there would be no shift in frequency. But there is a shift, and Einstein was mistaken.

Now a bit of direct evidence.

Scientific American, in its issue of November 1961, published an article describing the two-mile linear accelerator to be built at Stanford University in California (it was built a few years later, and has been in operation for many years). On page 50, the article said the velocity of particles would be 0.999999997c relative to the accelerator. Then, a few words later, on the same page, came this:

"...the accelerating (electromagnetic) field must travel at a velocity close to that of the particles. Some slippage between the two is allowable, so long as the particles stay near the wave crests (the points of maximum accelerating force). To bring the wave velocity close enough to that of the particles, the inside of the pipe is designed with a series of ridges. These have the effect of slowing the electromagnetic waves travelling through the pipe. If the pipe were perfectly smooth, the waves would travel faster than c and would be unable to transfer energy to the particles... A proper choice of the dimensions of the ridged structure slows the wave to the desired velocity."

Reference

1. A Einstein, "On the Electrodynamics of Moving Bodies," in *The Principle of Relativity*, translators (1923) W. Perrett, G.B. Jeffery (Dover, New York, 1952), p.38. (Also, General Publishing, Toronto, and Constable & Co., London.)
Lee Coe, Berkeley, California.

I was very interested in H. Aspden's letter concerning relatively and the Sagnac effect.

STC in London are developing a optical-fibre gyroscope which depends upon this effect for its operation. However, what is detected is a rotation of an optical-fibre coil, not linear motion, thus the general theory and not the special theory is involved. In E.W. Silvertooth's experiment as described, surely what is being detected is a rotational effect as the earth rotates around the sun. D. Marquis, Westerham Hill, Kent.

As every schoolboy surely has been taught, it is perfectly possible to "prove" (mathematically, of course) that $2=1$. Permit a reminder:

Let $a=b$ and multiply both sides by a , so $a^2=ab$.

Subtract b^2 from both sides -
 $a^2 - b^2 = ab - b^2$

Factorize
 $(a+b)(a-b) = b(a-b)$

Cancel
 $(a+b) = b$

But $a=b$
Therefore $2b=b$
 $2=1$

It is quite like removing every vestige of energy from a system so that a transform may work in the static state of mathematics, simultaneity having broken down, so that the assumption may be made that the speed of light is infinite to correct the situation, so that the interval in space-time between two events on the world line of a photon may be made zero, in order that the said transform may work as intended! Of course, it is just another Ezekiel's wheel revolving round an axle of presumption.

One is then left with the infinite speed of light when the system is devoid of energy: very neat! This is further explained by the strange statement that mass and energy are the same thing, whence energy must have inertial mass because it is non-existent.

Unfortunately for the mathematicians, both these are interrupted by catastrophe which makes both the numbers and their symbols totally valueless.

What this demonstrates is the lack of value of mathematics unless they are fully backed by an active dynamic logic which may

be expressed in words: that is to say, the mathematics must be done without first freezing the system solid by removing every vestige of energy, and existing thought is sheer poppycock. Sorry Mr Coleman, but at least Mr Snoswell should be pleased, along with Mr Aspden and Mr Abdullahi when they get down to it, all in November Feedback.

Incidentally, Mr Editor, and presumably for the benefit of your (what used to be called) "comps", the word is "couch" as recorded in the Book of Job: to "couch" implies servility but "couch" implies a sleepy state of static boredom!
James A. MacHarg, Wooler, Northumberland.

I was interested in the comment in the letter from M.D. Abdullahi of Zaria, Nigeria in your issue of November, where he obtained the energy of a charge as $(mc^2/2)$, rather than mc^2 .

$mc^2/2$ is the finite limiting value of the translational kinetic energy of the electron, and we can obtain this as equal in magnitude to the finite limiting value of the spin angular kinetic energy of the electron as the translational velocity approaches zero, which is the rest energy.

As the translational velocity of the electron approaches its finite limiting value, the fluctuations in the magnitude of the spin angular velocity increase indefinitely, and the electron liberates an infinite sequence of photons of indefinitely increasing energy. Thus the electron can take as much energy from the accelerating field as the field can give it, and most of this energy is radiated as an infinite sequence of photons of indefinitely increasing energy.

During the fluctuations in the spin angular velocity of the electron, the resultant angular momentum is conserved by fluctuations in the lateral component of the translational velocity of the centre of mass, this representing the commutation relations of translational and angular velocity of quantum mechanics. At the minima of the total kinetic energy as the translational velocity approaches its finite limiting value, the magnitude of the spin angular velocity

FEEDBACK

approaches zero, and the translational kinetic energy, which is approaching its finite limiting value, has almost entirely replaced the finite limiting value of the spin angular kinetic energy when the translational velocity approached zero, which is the rest energy, and the magnitudes of the two finite limiting energies are equal.

R. Fricker,
Surbiton,
Surrey.

In the November issue, M.J. Snoswell asks "what happens to a top"? The answer to that must surely depend upon how one treats the top. One result is shown in the picture. The 24 pound top is spinning at 2000 r.p.m. and is rising quite rapidly up a helical path.

The downward pressure on the finger is no more than two pounds and so it seems that some 22 pounds of gravitational mass has mysteriously vanished.

A further aspect of the affair is that the calculated centrifugal force in the horizontal plane from the centre of the helix was to have been 17 pounds but in the event there was no c.f. at all. Again it seems that *all* of the inertial mass vanished.

Perhaps some dedicated relativist who fervently believes that to increase the energy of a body is to increase its mass will tell us where the mass got hid.

It may be that if such an explanation is forthcoming we shall know how the Cheshire cat's grin got out of the bag in order to join MacHarg's menag-

ery or, is it perhaps that we have a new science and Albert was wrong?

Alex Jones,
Alderney.

Alan Watson (April Letters) rightly emphasizes that the current experimental evidence for the time dilation predicted by Special Relativity establishes the effect beyond question. However the work which he actually cites was presumably based on the use of atomic clocks in fast flying aircraft or in satellites. Even for the latter the predicted effect is only one part in 10^{10} , so that the experiments are open to the criticism that some small systematic effect may have been neglected, and the gravitational effects on clock rates are of comparable magnitude.

Much more dramatic are the measurements on fast-moving unstable subatomic particles such as the muon and pion. The measured lifetimes of these two particles at rest are 2.20 μ s and 26.0 ns respectively, and when they are in motion the values should, according to Special Relativity, be increased by the time dilation factor. The earliest measurements of this kind involved comparing the muon component of the cosmic rays at the top of the atmosphere and at ground level, and were made in the late thirties. The muons travel downwards at almost the speed of light, and to account for the large fraction that survive the journey it was necessary to assume a lifetime several times larger than the estimated value for muons at rest.

Precise quantitative measurements became possible with the advent of high-energy particle accelerators fifteen to twenty years ago. Measurements on the fraction of positive pions which survived transit along a twelve metre flight path gave a time dilation of 2.40, within $\frac{1}{2}\%$ of the Special Relativity prediction for pions of the selected energy¹. In principle, this measurement corresponds closely to the earlier cosmic ray measurements. Relativistic muons, however, travel about 700 metres in 2.2 μ s, so that to observe time dilation in the lab. it was necessary to force them to travel repeatedly round a circular path, and to measure the survival fraction by sending off bunches of muons at suitable intervals, and observing the number that decayed near a fixed position on the path as a function of time after starting. Such measurements have been made for muons of two different energies. The more recent set gave an increase in lifetime by a factor of 29.3, within 0.1% of the Special Relativity prediction^{2,3}. Because the muons are forced to return time after time to the same position in the lab., these last experiments actually measure differential ageing, i.e. the phenomenon involved in the so-called 'twin paradox'.

The meson experiments test aspects of Special Relativity which are completely inaccessible to measurements made with atomic clocks, and confirm its predictions with unparalleled precision. Moreover they are virtually immune to the effect of the earth's rotation, of the earth's

orbital motion round the sun, and of its gravitational field. May we now forget the Haefele-Keating experiment?

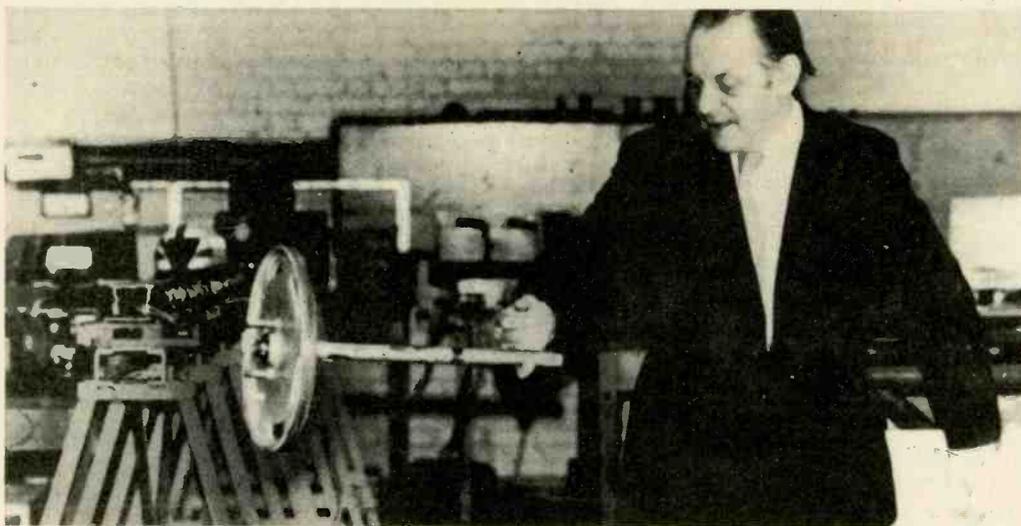
C.F. Coleman,
Grove,
Oxfordshire

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1. D.S. Ayres, A.M. Cormack, A.J. Greenberg, R.W. Kenney, E.F. McLaughlin, R.V. Schafer, D.O. Caldwell, V.B. Elings, W.P. Hesse, and R.J. Morrison, *Nucl. Instrum. Methods* 70(1969)13
2. J. Bailey and E. Picasso, *Prog. Nucl. Phys.* 12(1970)43
3. J. Bailey, K. Borer, F. Combley, H. Drumm, F. Krienen, F. Lange, E. Picasso, E. Picasso, W. von Ruden, F.J.M. Farley, J.H. Field, W. Flegel, and P.M. Hattersley, *Nature* 268(1977)301.

Apart from the problem he poses (Feedback, November, 1986) Mr Abdullahi should expect to encounter discrepancies involving factors of $\frac{1}{2}$ from general considerations. The reason is that our perception of the physical world is based on sampling of our external environment by our sensory apparatus. Even with the very large number (10 to 15 thousand million) nerve cells in the brain, we cannot sample at twice the rate at which some external events occur in order to comply with the requirements of Nyquist sampling frequency and thus avoid 'aliasing' as a necessary component of some aspects of mental activity. In addition, the electrical and chemical activity of the central nervous system is quantized as trains of voltage spikes (called 'action potentials' by neurophysiologists). From this, it would be expected that, at some (very low) threshold of perception and cognition (called 'limen' by psychologists), events of the physical universe would only be detected with probability 0.5, constituting a possible explanation of, for example, Bell's inequality.

B.E.P. Clement,
Crickhowell,
Powys,
Wales.



Electrometer amplifiers for sub-picoamp currents

Switched-range and logarithmic amplifiers for currents down to 20 femtoamps

D.F. CAUDREY

Electrometer amplifiers are widely used in the nuclear industry for amplification and measurement of direct currents from ionization chambers used, for example, in reactor control and safety instrumentation. In these applications they process currents ranging from femtoamps to nanoamps (10^{-15} A to 10^{-9} A).

This article presents the basic design principles for a switched-range linear amplifier and an examination of those concerned with its single-range logarithmic counterpart. To quantify the design parameters, it is assumed that either type is to measure currents over the four and a half decades from 20 femtoamps to 0.6 nanoamps; in the linear case, by ten switched ranges – 20fA f.s.d. multiplied by 1,3,10,30 and so on to 0.6nA f.s.d.

DESIGN PRINCIPLES

Figure 1 shows a block diagram of a direct-coupled measuring amplifier. The current flowing through the milliammeter, I_{out} , is essentially equal to

$$\frac{I_{in}R_f}{R_m}$$

provided that no current is diverted to or fed from the amplifier input terminal and the loop gain $AR_m/(R_m+R_1)$ is much greater than unity.

The amplifier must, therefore, have very high input resistance, its gain must be large and R_m must not be much less than R_1 . From expression (1) it can be seen that I_{out} is directly proportional to I_{in} and is equal to I_{in} multiplied by the factor $-R_f/R_m$, the current gain of the amplifier. It can also be seen from (1) that I_{out} is independent of the resistance of the milliammeter and within reason any milliammeter may be used to indicate the magnitude of the input current.

The range voltage V_m is given by $I_{out}R_m$ and inserting this into (1) gives $V_m = -I_{in}R_f$. Therefore V_m is equal to, but opposite in polarity to V_f , the voltage drop across R_f due to I_{in} . Of course, V_f and V_m differ by the voltage needed to drive the amplifier, which is $V_f/\text{loop gain}$, but if loop gain is very high the drive voltage will be virtually zero. As will emerge later, this *virtual-earth input* is an essential requirement for current-measuring electrometers.

Choice of range voltage. Consider the most sensitive range of 20fA f.s.d. and assume that this is to be displayed on a 1mA f.s.d. meter.

From (1) the magnitude
$$\frac{I_{out}}{I_{in}} = \frac{R_f}{R_m} \quad (2)$$

Therefore
$$\frac{R_f}{R_m} = \frac{10^{-3}}{2 \times 10^{-14}} = 5 \times 10^{10}$$

or
$$R_f = 5 \times 10^{10} R_m \quad (3)$$

Thus it would appear that any combination of resistors having this relationship could be used to achieve the desired result. However, the amplifier shortcomings must be taken into account. All directly-coupled amplifiers are subject to drift, both as a random function of time or as a result of temperature change. Fortunately, time-dependent drift is small in solid-state amplifiers, but the effects of temperature have considerable bearing on the choice of V_m . In common with most solid-state, directly-coupled amplifiers, the electrometer amplifier is a balanced, differential type having one input terminal driven by the input signal and the other held at a reference potential. This, of course, minimizes the direct effects of temperature and supply voltage changes by rendering them common-mode, and the degree of temperature independence achieved depends on the matching of temperature coefficients in the balanced input stage. In all practical electrometer circuits, there will be appreciable temperature dependence.

The temperature dependence of operational amplifiers is usually expressed in terms of the equivalent difference voltage between the input terminals and, for modern monolithic integrated circuits, figures of 1 to $5\mu\text{V}/^\circ\text{C}$ are commonly specified. Unfortunately, due to the need for very high input resistance, it is essential to employ mosfet transistors in the input stage of electrometer amplifiers; devices which can normally be matched to better than $100\mu\text{V}/^\circ\text{C}$ in monolithic pairs. Consequently, the virtual-earth driving point can drift by up to 2mV for 20°C temperature excursions. This effect, together with other sources of drift, can be represented by an additional voltage generator V_t , as shown in Fig.2.

As V_t is within the feedback loop, it must be small in comparison with $V_m (-V_f)$ if the dominant temperature-dependent drift is not to be a nuisance. However, the choice of V_m is constrained by the maximum permissible value for the feedback resistor R_f and its relationship to R_m defined in expression (3).

High-value resistors are available in values up to $100\text{T}\Omega (10^{14}\Omega)$ but about $3\text{T}\Omega$ is the maximum value which gives accept-

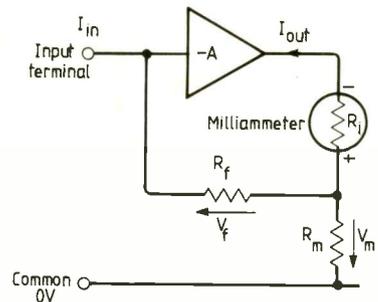


Fig.1. D.c. amplifier block diagram, in which I_{in} is the input current; R_f the input resistor; R_m a range-multiplier resistor; R_1 the internal resistance of the meter; and V_m the range voltage. The meter can be replaced by a resistor terminating the input to a d.p.m., simplifying range switching.

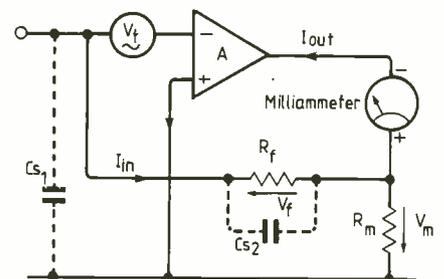


Fig.2. Voltage generator V_t represents temperature and other sources of drive.

able selection tolerance (1%) and stability in respect of temperature and voltage coefficients and ageing. This constraint defines V_m as 60mV for the 20fA f.s.d. range. For the higher ranges of input current, it is possible to reduce R_f or to increase R_m and concomitantly V_m . Of these options, the former gives the benefit of reduced response time, as is discussed later, and the latter is less costly, because fewer expensive, high-value resistors are required to accommodate the projected ten ranges of current measurement. The latter option also simplifies the range switch at the input to the amplifier, with a consequent reduction in static-charge overloads during range selection. In fact, it is possible to employ only two values for R_f ($3\text{T}\Omega$ and $10\text{G}\Omega$) in conjunction with seven values of R_m to cover the assumed range of current measurement 20fA to 0.6nA.

Response time. Also shown in Fig.2 are two stray capacitances C_{s1} and C_{s2} , which can

influence the response time of the electrometer amplifier. C_{s1} represents the input capacitance made up mainly by the capacitance of the signal source – ionisation chamber etc. – and the gate capacitance of the input mosfet. The magnitude of C_{s1} might be as much as 100pF and, by virtue of its position, it must be charged by the input current before the electrometer can achieve a correct indication. Remembering that V_m for the 20fA range is 60mV, it is worthwhile to examine the time required for the input current to charge 100pF to 60mV

$$t = \frac{CV}{I} = \frac{100 \times 10^{-12} \times 60 \times 10^{-3}}{2 \times 10^{-14}} \text{ seconds.}$$

Therefore, without virtual-earth operation, f.s.d. would be reached after five minutes. However, for the most critical 20fA range, the loop gain $AR_m/(R_m + R_f)$ for the projected design is likely to exceed 40,000 if a typical operational amplifier is included in the loop, and thus C_{s1} will be charged to less than $60/40,000$ mV, i.e. 1.5µV. This requires approximately 0.008 seconds and C_{s1} can be rendered insignificant with respect to its effect on response time.

C_{s2} represents the stray capacitance across the high-value resistor and, until it is charged by the input current, the full range voltage cannot be developed across R_f . For increasing input current, C_{s2} and R_f form an integrating pole with a settling time constant $T = C_{s2} R_f$ seconds and the indication may be considered to have settled after 3T. C_{s2} is typically 2.5pF and therefore T is approximately 7.5 seconds for ranges employing the 3TΩ resistor. For these current ranges, the output of the amplifier will reach 95% of its final value in approximately 22 seconds from a step increase of input current. For ranges employing the 10GΩ resistor, the response is much more rapid and it is sometimes desirable to create a similar pole by shunting the resistor with a polystyrene capacitor.

Input circuit. As mentioned previously, a mosfet input stage is essential to prevent the signal current from being diverted into the input terminal of the amplifier and to prevent it from being augmented by current from within the amplifier.

Mosfets are highly susceptible to breakdown of their gate insulation, due either to voltage transients or to static charges received through handling. Breakdown due to voltage transients is a major problem in instruments employing ionization chambers, since these often require polarizing potentials of several hundred volts. It is therefore essential to protect the input gate with devices which break down or avalanche non-destructively at a voltage lower than that which causes destructive breakdown of the gate insulation, and for less stringent applications zener or avalanche diodes are used for this purpose. However, for electrometers, the conductance of normal zener diodes would be unacceptable and special low leakage "picoamp" diodes are used. The simplified input stage of a typical electrometer amplifier is shown in Fig.3.

The gate-protection diodes are connected

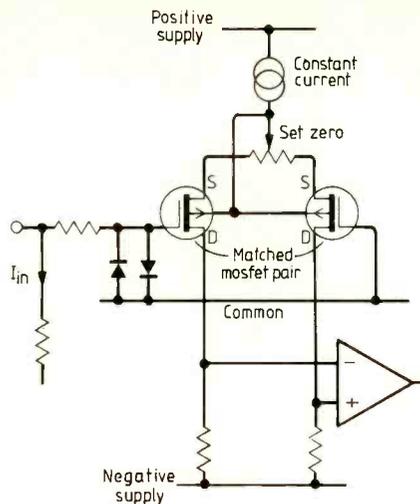


Fig.3. Typical input stage using mosfets and protection diodes.

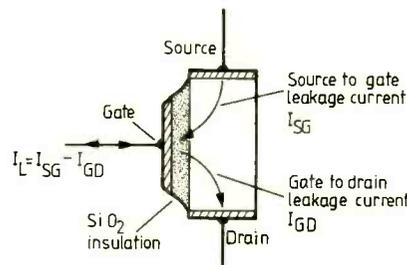


Fig.4. Internal leakage paths in a mosfet, which tend to cancel when balanced.

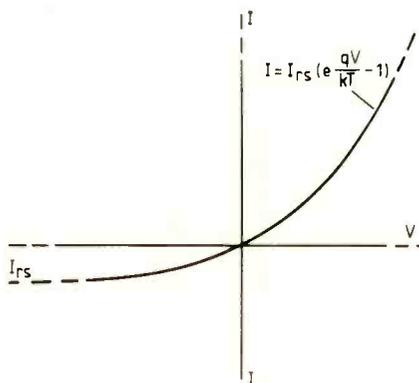


Fig.5. Diode characteristic.

between the vulnerable input gate and the common 0V line. In view of the requirement for negligible leakage, it might seem surprising that the diodes are connected in parallel and apparently short circuit the input terminal to zero. However, the picamp diodes are of very small area and are characterized by very low reverse leakage in comparison with ordinary signal diodes. This property, in conjunction with virtual earth operation, ensures that leakage current to common 0V is very small in comparison to the input current for the most sensitive range, irrespective of whether the diodes are forward or reversed biased; provided, of course, that there is negligible voltage offset at the input terminal. However, the diode leakage approximately doubles for every 7°C temperature rise and consequently a 20°C rise might result in an eight-fold increase. With

the input terminal at zero the magnitude of this increased conductance is not excessive at the elevated temperatures but, as discussed in connection with the choice of range voltage V_m for the most sensitive range, it can be accompanied by a voltage offset of 2mV.

Taken together, these effects of increased temperature can result in unacceptable offset current and it is therefore necessary to make provision for precise adjustment of the 'Set Zero' control to minimize voltage offset, particularly at high ambient temperature. The Set Zero control, shown in Fig.3, compensates for differences between the gate-source voltages for the mosfet pair when $I_{out} = 0$: it is adjusted in a Set Zero position of the range-selection switch, which short-circuits the high-value resistor R_f and selects a lower value V_m . The control is normally a ten-turn potentiometer which, in conjunction with a V_m of the order of 10mV, permits the voltage offset to be easily maintained within $\pm 100\mu V$ of 0V.

In view of the fact that leakage through the protection diodes is rendered insignificant by virtual earth operation, why is it not practical to use bipolar transistors or junction-gate field-effect transistors (jugfets) in the input stage? The answer to this question is that the internal leakage current, that is current to and from the input terminal from the other electrodes, is too great in these devices. In mosfets, internal leakage is minimized by the gate insulation to a level many orders below the base current for a bipolar transistor and several orders below the leakage across the reversed biased gate junction of a typical jugfet. Even when mosfets are used, it is advantageous to minimize offset current with equal gate-to-source and gate-to-drain voltage, under which condition the internal leakage currents depicted in Fig.4 tend to cancel. This balanced operation also cancels external inter-electrode currents due to leakage across the header, which can be minimized by installing the electrometer amplifier within a desiccated, sealed compartment and using devices having headers of low-leakage glass.

Some years ago GEC-Marconi marketed an electrometer dual mosfet (MBH1) with a header assembly having inter-electrode guarding by means of an earth-plane. Unfortunately this device is no longer available and matched dual mosfets are now difficult to obtain, semiconductor manufacturers finding the market too small.

However, Intersil market a complete electrometer operational amplifier (ICH8500A) which is suitable for all but the lowest-current applications without selection and which embodies the essential features discussed in the foregoing text.

LOGARITHMIC ELECTROMETERS

For some applications of electrometer amplifiers, a single-range logarithmic display is an ideal way of presenting measured data, particularly for process-monitoring applications, where a wide dynamic range

continued on page 112

Two novel oscillators

A voltage-controlled active-R oscillator and a minimum-component design using a single-op-amp, resistor and capacitor.

M.T. ABUELMA'ATTI AND W.A. ALMANSOURY

Voltage-controlled oscillators find numerous applications in instrumentation, measurement and communication systems. The literature contains a large number of oscillator designs and analyses based on active-RC and active-R principles. The active-R design technique is, however, more attractive for monolithic i.c. fabrication. Our first circuit is a new voltage-controlled active-R oscillator using only two general purpose operational amplifiers and two resistors.

ACTIVE-R OSCILLATOR

The circuit is shown in Fig.1. Let the open-loop gain of the operational amplifiers be represented by the two-pole model given by

$$A(s) = \frac{A_0 \omega_a \omega_b}{(s + \omega_a)(s + \omega_b)} \quad (1)$$

where A_0 is the d.c. gain, ω_a is the first corner frequency, ω_b is the second corner frequency and $B = A_0 \omega_a$ is the gain-bandwidth product of the operational amplifier. The transfer function of the circuit can be expressed as

$$\frac{x}{y} = \frac{(1+n)A_1A_2}{(A_1-n-1)(1+A_2)} \quad (2)$$

where $n = R_2/R_1$.

For the circuit of Fig.1. to produce and sustain oscillations, the transfer function of (2) must be equal to unity, giving

$$[(\omega_a + s)(\omega_b + s) + B\omega_b](\omega_b + s)(\omega_a + s)(n+1) = [(\omega_a + s)(\omega_b + s) - nB\omega_b]b\omega_b \quad (4)$$

By rearranging (4) and equating the imaginary part to zero, the frequency of oscillation can be shown to be

$$\omega_0 = \left[\omega_a \omega_b \left(1 + \frac{nA_0}{2(n+1)} \right) \right]^{1/2} \quad (5)$$

Further simplification of (5) can be obtained if $nA_0 \gg 2(n+1)$, giving

$$\omega_0 = nB\omega_b/2(n+1) \quad (6)$$

By equating the real part of (4) to zero and using (5) and noting that for most practical cases $\omega_b \gg \omega_a$ the condition of oscillation can be shown to be approximately

$$A_0 \geq \frac{2(n+1)}{3n+4} \frac{\omega_b}{\omega_a} \quad (7)$$

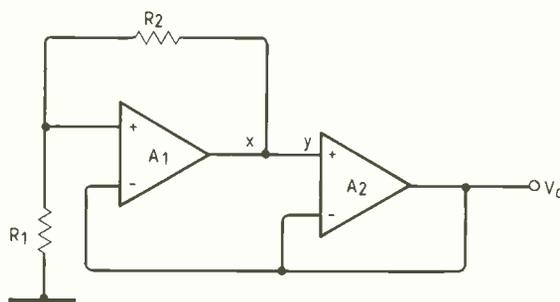


Fig.1. Active-R oscillator circuit. Frequency can be adjusted conveniently by replacing R_1 with a j-fet and varying its gate-to-source voltage.

This condition is easily satisfied in practical operational amplifiers.

Equation 5 shows that the frequency of oscillation is a simple function of the resistance ratio n . This feature of the proposed circuit is of practical significance because the resistance ratio is effectively insensitive to temperature changes over a considerable range. Moreover, by keeping R_2 constant, it is possible to adjust the frequency of oscillation simply by changing R_1 . By replacing R_1 with a j-fet it is, therefore, possible to adjust the frequency of oscillation by adjusting the gate-to-source voltage of the j-fet.

The circuit was built and tested using the operational amplifier TL064CN and the j-fet 3N187 and fairly good sinusoidal oscillations were obtained: Fig.2 shows a typical output waveform at 250kHz.

Figure 3 shows how the frequency of oscillation altered with the gate-to-source voltage of the j-fet. When the gate-to-source voltage was varied from 0 to $-2.5V$, the corresponding change in the frequency of oscillation was from 500kHz to 128kHz.

Changing the d.c. supply voltage of the operational amplifiers from $\pm 9V$ to $\pm 15V$ raised the frequency of oscillation from 100kHz to 133kHz (Fig.4).

MINIMUM-COMPONENT OSCILLATOR

Because of their numerous advantages, there has been considerable interest in the design of active-RC oscillator circuits with grounded capacitors. Besides a single grounded capacitor, our second circuit uses only a single resistor and a single operational amplifier. The circuit, therefore, uses the minimum number of passive and active elements (Fig.5).

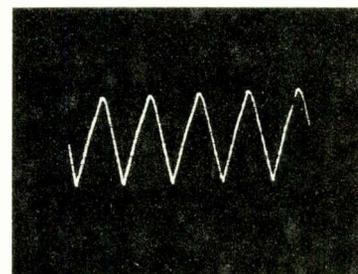


Fig.2. Output waveform at 250kHz of the circuit in Fig.1 with a j-fet replacing R_1 ; V_{GS} of the j-fet is $-1.5V$, R_2 is $3.3k\Omega$ and the op-amp's supply is $\pm 10V$.

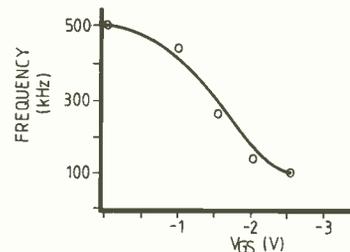


Fig.3. Frequency of oscillation varies with the gate-to-source voltage of the j-fet. Op-amp supply is $\pm 12V$, R_2 is $3.3k\Omega$.

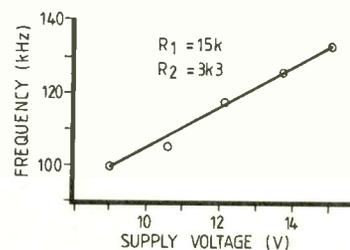


Fig.4. Frequency can also be varied by altering the supply voltage of the op-amps.

Let the open-loop gain of the operational amplifier be represented by the two-pole model given by equation 1. The transfer function of the circuit can be expressed as

$$\frac{y}{x} = \frac{-A(s)}{(1+sCR)} \quad (8)$$

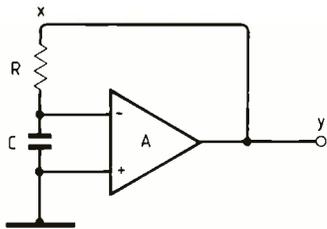


Fig.5. Op-amp oscillator for the minimalist: for practical reasons the grounded capacitor arrangement is very desirable.

For the circuit of Fig.5 to generate and sustain oscillations, equation 8 must equal unity. Substituting (1) into (8) and rearranging, then

$$(1+sCR)(\omega_a\omega_b+s^2+s(\omega_a+\omega_b)) = -A_0\omega_a\omega_b \quad (9)$$

The frequency of oscillation can be obtained from (9) by equating the imaginary part to zero, giving

$$\omega_0^2 = \omega_a\omega_b + \frac{\omega_a + \omega_b}{RC} \quad (10)$$

By equating the real part of (9) to zero and using (10) and noting that for most practical cases $\omega_b \gg \omega_a$, we can show that the condition of oscillation will be given, approximately, by

$$A_0 > \left[\frac{\omega_b}{\omega_a} + \frac{1}{\omega_a CR} + \omega_b CR \right] \quad (11)$$

This condition can easily be satisfied in practical operational amplifiers.

From equation (4) it follows that the frequency of oscillation is a simple function of the product RC. Therefore, by changing R or C, or both, it is possible to adjust the frequency of oscillation.

The circuit of Fig.5 was built and tested with a 741 op-amp, polystyrene capacitors and metal film resistors. A typical waveform is shown in Fig.6.

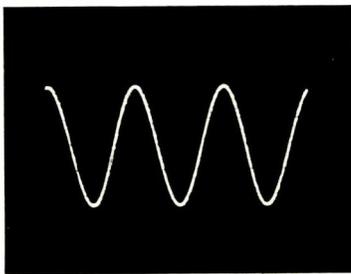


Fig.6. Output waveform at 3kHz of the circuit in Fig.5 built with a 741 op-amp. The resistor is 19k Ω , the capacitor 1 μ F and the supply voltage \pm 15V.

When the RC product was adjusted from 57ms to 1 μ s, the frequency of oscillation was found to vary over more than two decades, from 1.4kHz to 200kHz (Fig.7).

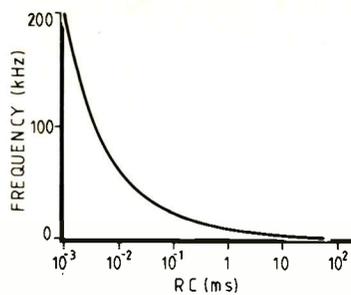


Fig.7. Altering the RC product of the circuit in Fig.5 shifts the output frequency over more than two decades.

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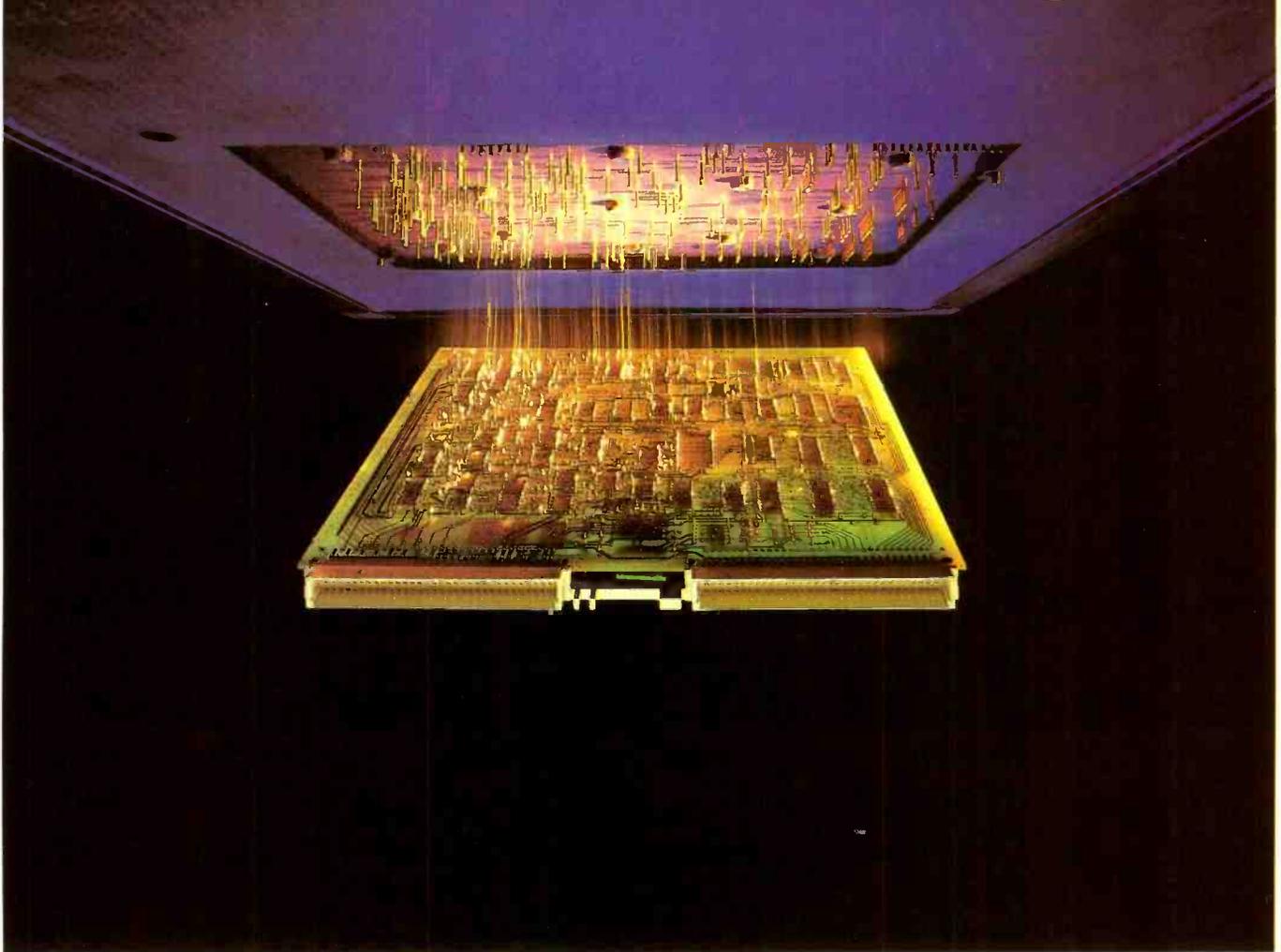
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Automatic testing



An overview of techniques used in the automatic testing of printed-circuit board assemblies

GRAHAM ELLIOTT

Automatic test equipment, a.t.e., is widely used to test electronic components and printed-circuit board assemblies. It comprises a set of test instruments whose functions are controlled by a computer via a test program. A test fixture provides a hardware interface through which access to the unit under test is achieved. The a.t.e. market breaks down into three major groups; p.c.b. testers, component testers and bare board testers. All a.t.e. systems can have a great effect on the commercial viability and attainment of quality and specification for a variety of electronics products.

The accelerating growth of technology means that improved and more complex devices are coming to market faster. Product lifetimes are reduced and the problems of developing and introducing new lines amplified. Customers now familiar with high-technology products demand a very high level of reliability. The general shortage of skilled electronics engineers makes the ap-

plication of manual test stages impractical and existing stages of automation have grown to provide the solution. Efficiency, productivity and quality are all improved, while costs are reduced. The use of automated systems in design, manufacture and test has increased dramatically over the last few years, and a.t.e., in particular, provides important information relative to scheduling, costs and product quality.

WHY TEST AT ALL?

Testing electronic devices and assemblies is an expensive process and one might well ask why test at all? If designs were error-free, all components purchased functioned to specification, all sub-assemblies were correctly connected and no calibration or adjustments were needed, the product could go directly to the customer. Unfortunately it is not a perfect world and should a manufacturer choose to omit the test stage it would soon go out of business; field service costs would be astronomical and a reputation for poor

quality and reliability would quickly put off prospective customers. Testing is an essential part of the manufacturing process, providing a high degree of confidence but, because of the costs involved, must be performed only at key points in the process to be economical.

It is possible for testing to take place at every stage in the production line shown in Fig.1. But to do so would waste resources and result in less than optimum performance. Usually a fault coverage of around 95% can be achieved at a reasonable cost, the remaining 5% being uneconomical for most companies to achieve and requiring a disproportionate investment in additional equipment. The a.t.e. systems described below may be implemented at any stage, each having been designed to match the requirements of a particular point in the production process.

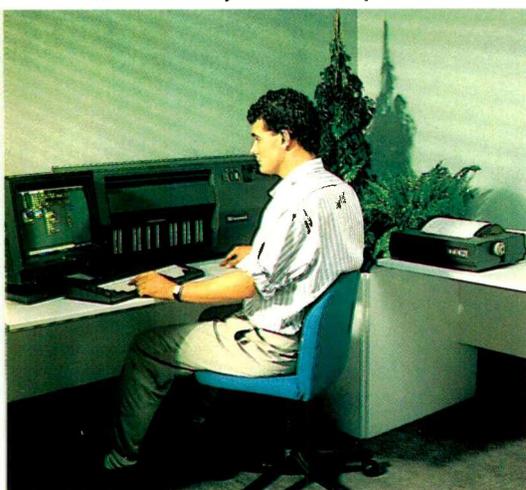
Component testers. These range from small bench-top bridges to large parametric test-

ers. The latter perform full parametric tests on integrated circuits prior to their insertion into printed circuit boards. Costs range from a few thousand pounds for the smaller systems to several hundred thousand pounds for the parametric testers. The installation of parametric testers is not usual for most applications.

Bare-board testers. Blank p.c.b.s may be tested on bare-board testers prior to their population with components, each track checked for breaks and spurious short circuits. Test speeds and board throughputs are high, making their application popular where expensive or very high volumes of printed-circuit boards are used. Prices begin at approximately twenty-five thousand pounds and fixturing costs (fixtures providing a hardware interface between tester and a p.c.b. under test) are minimal. Bare-board testers are usually used prior to later testing on in-circuit or functional a.t.e..

Manufacturing-defect analysers. M.d.as detect manufacturing process faults by testing for track continuity, spurious short circuits, component presence, correct value and orientation, without applying power to the printed circuit board or assembly. They are often used with other testers, pre-screening p.c.b.s prior to in-circuit or functional testing. Costs for m.d.as commence at approximately thirty-five thousand pounds.

In-circuit testers. If a printed-circuit board is correctly manufactured and each component correctly fitted and operational, then



Bench-top automatic test equipment

the p.c.b. will function correctly. That is the principle on which in-circuit testers are based. Track continuity tests and checks for spurious short circuits are performed, although not as quickly as on the bare board tester. Passive component measurements establish presence, value and orientation of discrete components. Controlled powering up of the board prevents component failures occurring at this stage, which is followed by active analogue and digital tests. In-circuit testers are available in the range £25,000 to £200,000.

Functional testers. A different test strategy is used by functional testers, that of testing the p.c.b. as a single functioning unit. This is

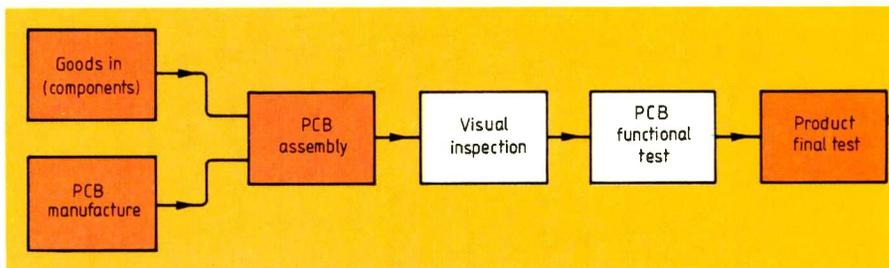


Fig.1. Manually testing at each production stage is wasteful of resources, while still only providing a 95% fault detection at reasonable cost.

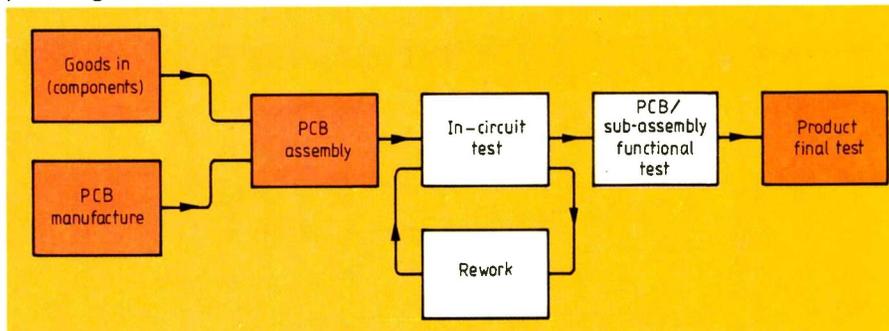


Fig.2. Replacing one manual inspection by in-circuit testing reduces cost, increases production and improves records.

done by injecting a series of test patterns into the p.c.b.'s inputs and monitoring for expected patterns at the outputs. When errors occur, diagnostic algorithms, generated by simulator software, identify the failing component. This method is particularly effective for locating digital device errors occurring at high speed, functional testers operating at speeds of up to 30MHz, as opposed to the 1MHz digital test rate generally available from in-circuit testers. Prices for functional test systems range between £150,000 and £250,000 (bench-top testers start at around £50,000).

Traditional a.t.e. tends to be rather large and expensive, but smaller and cheaper bench-top systems are also available, as indicated above. The performance of these systems, in terms of test speed, p.c.b. throughput and test-point capacity (the maximum number of circuit nodes with which the system can make contact) is generally lower than the larger systems. But bench-top systems are attractive, since they provide an entry path into automatic testing for small organizations unable to afford the traditional types. Larger companies may choose to purchase several system, providing virtual immunity to failure and allowing a more flexible test strategy.

The option therefore exists to test at every stage, and the best place or places to test will depend upon the requirements of the particular installation. Identifying the optimum point at which to site test equipment is achieved by analysing the types and proportions of faults occurring. A typical production line may suffer the following failures:

- 30% short circuit tracks
- 25% faulty components
- 12% unsoldered or dry joints
- 9% wrongly fitted components
- 8% wrong components
- 2% broken components
- 2% open circuit tracks

To determine the optimum position in the production line to place test equipment, one

must first consider how and where the faults occur. Short circuits and soldering problems are likely to occur at the soldering stage, while most wrongly inserted, wrong and broken components are introduced at the component insertion stage prior to soldering. Broken components and open circuit tracks can be introduced at any point, but are more likely at the earlier stages, or during rework. Any areas of manual assembly are often a weak link in the chain and are susceptible to any of these types of error. One must now assess the most suitable tester test option.

Continuity testing will only find the track-related faults introduced prior to board assembly, a small proportion of those identified. Component testers find a similarly small proportion of p.c.b. faults. A manufacturing-defect analyser will find most of the faults in the list, but not those requiring active tests (a possible 25%). The functional tester positioned at the end of the assembly line is better placed, but is not best used to detect track and analogue failures, being very effective at locating digital faults. This is also the most expensive system to buy and operate.

In-circuit testing provides the best solution in this case, being capable of detecting all of the fault types in the analysis. All p.c.b. failures are detected during a single test of the board, enabling a high throughput to be maintained, and fault reporting to component level simplifies rework.

The in-circuit tester. The production process diagram of Fig.1 now becomes that shown in Fig.2. If the in-circuit tester replaces a manual test stage, several advantages immediately become apparent; skill level at the test stage and cost per p.c.b. are reduced, documentation is improved and production capacity is increased.

Producing the test program and fixture. A block diagram of system architecture is shown in Fig.3. The central computer con-

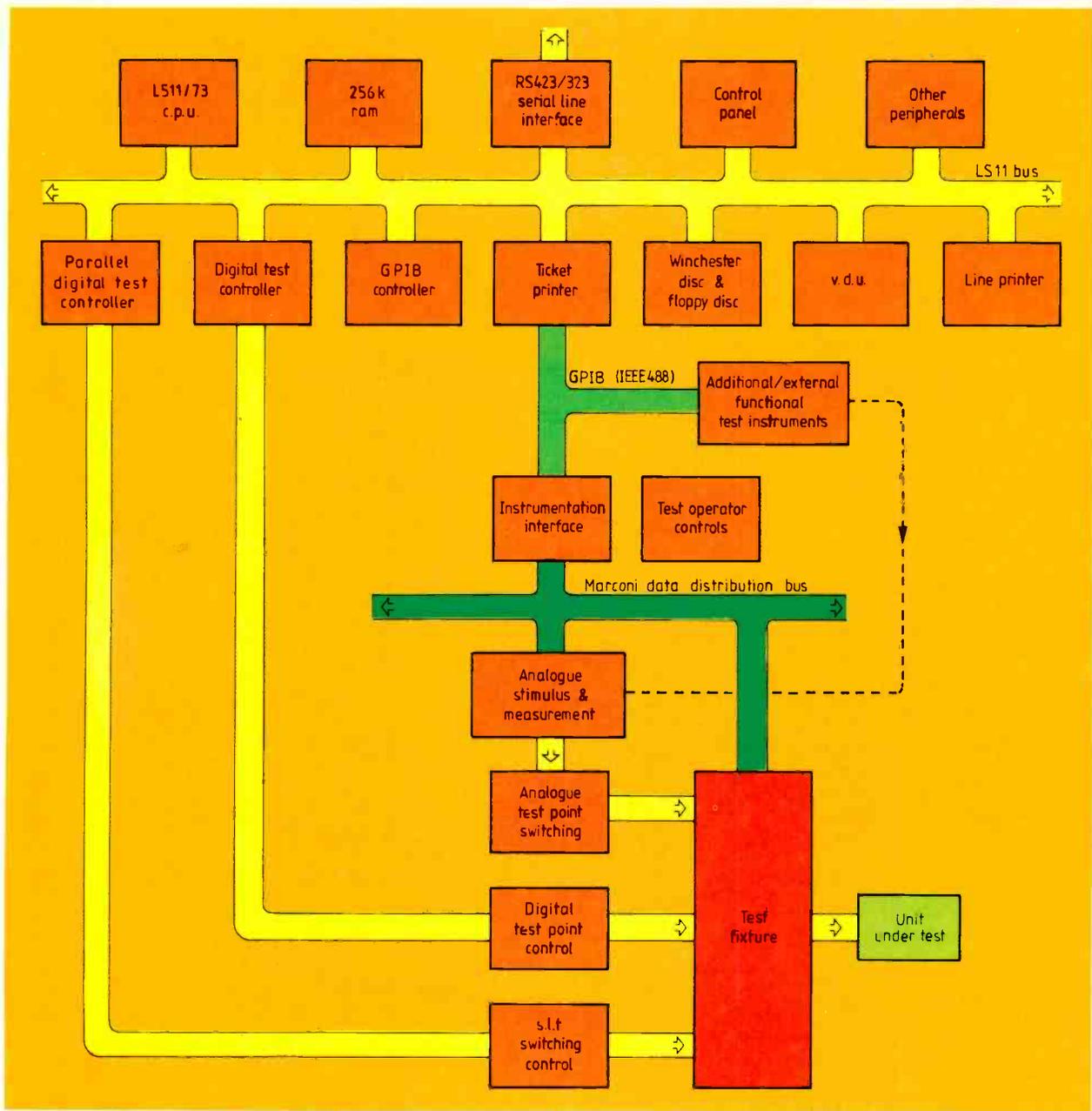


Fig.3. System architecture of an in-circuit tester.

trols the operation of the analogue and digital test electronics. In some applications one may wish to connect external GPIB-controlled measuring instruments to the p.c.b. under test, and this may be done through the in-circuit testers' own hardware interface. System software will control the operation of these instruments via a GPIB port.

The operation of the hardware is invisible to the programme, the high-level programming language automatically configuring the a.t.e.'s measuring instruments to perform the necessary measurements. There are three basic steps in the generation of a working test program and fixture.

The first step is to allocate test points (which connect the a.t.e. hardware to the p.c.b.) to a circuit diagram of the board to be tested. This allocation will use the analogue and digital test points available on the test system, and is very straightforward. Once complete, it is transferred to a 1:1 artwork of the p.c.b. The artwork is used by the fixture manufacturer to ensure accuracy: in-circuit testers use a hardware interface known as

the 'bed of nails' fixture to provide easy and accurate connection to the board under test. Figure 4 and the accompanying photograph show a section through a bed of nails fixture and a picture of it in use. The fixture relies on accurately positioned spring probes (or test points) and the marked up artwork gives these positions. Wired back to an interface block which connects to the tester, they make contact with every electrical node (i.e. component interconnection) on the board. This requires very accurate positioning and registration, target areas for the probes typically being component leadouts or special dedicated pads. Location pins on the fixture are made and positioned to match artwork-referenced fixing holes in the p.c.b., the holes being chosen to ensure that the board is correctly oriented. Operation of the fixture is quite simple: the p.c.b. is positioned on the fixture's carriage plate and the 'operate' button pressed. A vacuum source, external to the test system but piped through it, evacuates the air between the carriage and bed plates, sucking the board onto the probes.

The second activity, occurring in parallel with the manufacture of the fixture, is the generation of the test program. Software provided on the a.t.e. prompts for parts-list information, such as component identifier, nominal value, tolerance or device type and the assigned test point numbers for each component. Active tests for digital devices are stored in a disc-resident 'library'. Special software locates the active test for the specified device and converts it to the high-level program language, appending the relevant test point information as part of this process. When all parts list information and the library-based tests have been added to the program, a second programming stage takes place. The program is analysed and partially debugged by a special software package which searches for device configurations which affect the accuracy of measurements performed by the a.t.e. 'Guard' points and other special test routines, for example bus tristate and open-collector test, are automatically added and the circuit checked for potentially dangerous testing and circuit configurations.

The final activity is to prove the test fixture and program together on the a.t.e.. This final program debugging resolves all remaining test problems and the p.c.b. can now be production tested on the a.t.e..

Order of test execution. In-circuit test execution is carried out in the following order: track tests for continuity and spurious short circuits, tests for the presence of wire links, measurements of discrete components, a controlled powering up of the p.c.b. and active analogue and digital tests. All faults are usually located on a single run of the test program, although the program run will be aborted if the track tests fail. This avoids the possibility of damaging the p.c.b. by applying power to it while there are track shorts on it. The test program is also aborted if an error occurs while performing the initial p.c.b. power up prior to the active analogue and digital tests.

Analogue test pin electronics. The analogue test-point switching system connects specified test points to the measurement hardware and is fundamental to accurate and flexible analogue measurement. It must be capable of performing an efficient scan for random short circuits between two or more points, provide high insulation between test points and system ground, enabling resistances of at least 10 milliohms to be measured without significant error and allowing leakage measurements to be performed. For capacitors in the order of 10 picofarads to be measured, inter-test point capacitance must be limited.

A number of design features allow these requirements to be met. Mechanical switching elements meet the isolation and series impedance requirements. A three-tier switching 'tree' reduces the selection time for specified test points and reduces stray capacitance in the switching system. A technique to solve the problem of single-pole switch losses is to employ p-n-p transistor switch-control stages. This avoids driving the coils direct from high-power, open-collector integrated circuits and all switches have their coils at zero voltage, minimizing coil-to-switch leakage. 'Break-before-make' operations protect the system from possible damage from stored charge in the board under test. The reed relays used are especially designed to minimize the effect of thermal e.m.f..

Analogue measurement techniques. The



Marconi MIDATA 540 in-circuit tester.

analogue measurement system is shown diagrammatically in Fig.5. It features two reference power supplies, a programmable resistor, a 12 bit analogue-to-digital converter and a measurement zero-volt line. These may be connected to any analogue test point via the four-pole switching matrix. The poles (or wires) of the matrix are designated A, B, C0, and C1.

An a-to-d converter is used in place of a digital voltmeter for greater speed. The system also features additional fast, floating reference power supplies to provide power to

the p.c.b. under test and stimulus for digital device drivers.

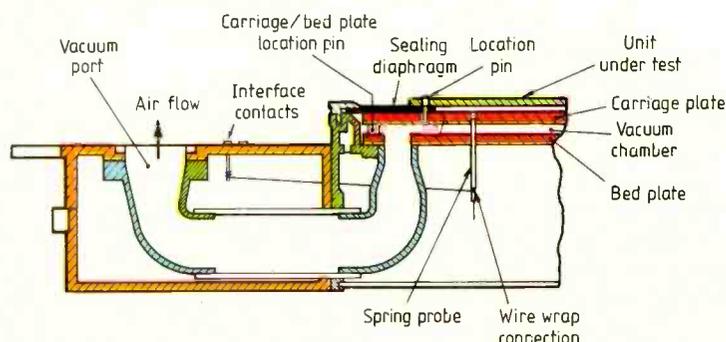
Under normal operating conditions, the system configuration for all standard measurements is arranged automatically by the high-level test language. However, experienced programmers have the option of full control of all parameters through the test program, enabling special non-standard tests to be performed if required.

Resistance measurement. A simple example of how tests are modified to cope with certain measurement conditions is the resistance test. The theoretical measurement circuit is shown in Fig.6. R_x is a resistor in the range 400Ω to $20M\Omega$. The high gain and high input impedance of the operational amplifier results in points X and Y being equipotential. Provided that system switch, lead and contact resistance are negligible compared with R_x , the gain equation of the operational amplifier is

$$\frac{V_{out}}{R_{ref}} = \frac{-V_{ref}}{R_x}$$

$$\text{or } R_x \propto \frac{1}{V_{out}}$$

Fig.4. Section through a bed-of-nails test fixture.



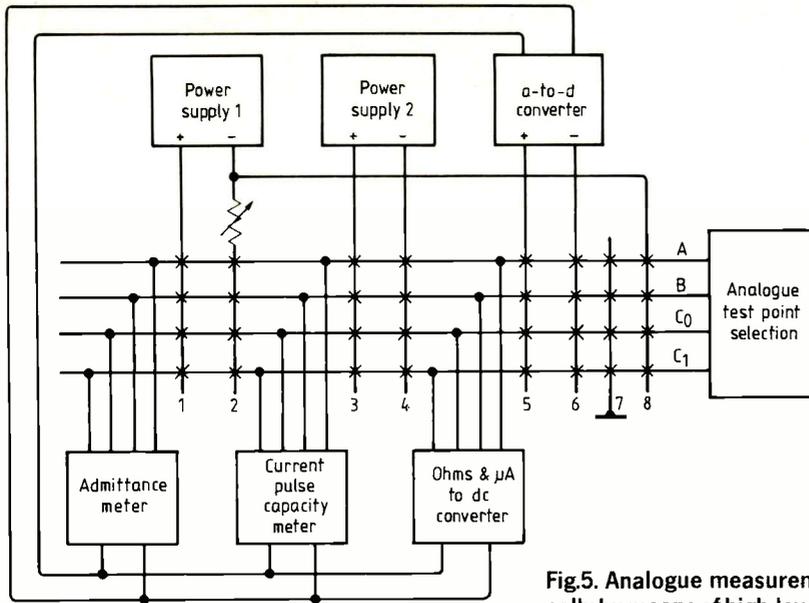


Fig. 5. Analogue measurement system, which is normally configured automatically by means of high-level test language.

This configuration is often sufficient to provide a reliable test for a resistor, but the component configuration shown in Fig. 7 is more often encountered. When testing R_5 there is a low-impedance path through resistors R_4 , R_9 and R_{10} . Referring to Fig. 8, the combined shunt resistance of these components will significantly effect the measured value of R_5 .

Test point 5 is selected as a 'guard' point and connected to pole C. R_4 shunts the voltage source, so provided it does not pass too much current it will not effect the voltage at Z. Since X and Y are equipotential, the series combination of R_9 and R_{10} does not pass any current and the original gain equation remains true.

The system will also perform a special low-value resistance test (using Kelvin techniques) and an a.c. resistance measurement where a resistor is shunted by an inductor. Other measurements include those on capacitors, inductors, leakage, diodes and transistors. Space does not permit the discussion of examples of these types of measurement.

Digital test execution. A prime requirement of all automatic test equipment is the location of faults to component level. Most in-circuit test systems meet this requirement when testing digital devices by using a test method known as node forcing. Digital pin drivers force logic levels at the device's inputs or control pins and override the normal quiescent state of the node due to preceding stages. The resulting logic states at the device outputs are monitored to establish the correct functioning of the device without needing to propagate logic states through complex digital circuitry. Much discussion has taken place as to the long-term effect of this method on other components on the p.c.b. Research has shown that it is the output stages connected to a force node which are stressed by the currents used in node forcing and that junction temperature rise within the device is the main hazard. Tests have shown that for node forcing to cause a t.t.l. device to reach a junction temperature of 70°C under normal

ambient conditions of 25°C, a forcing period of at least 30ms is required. The tester therefore applies a hardware timeout of 400µs on all node forcing stimuli, ensuring total device safety. The biggest advantage of node forcing is that it is easy and quick to program and provides a high test speed.

When the test program is run, each test routine is loaded from computer memory into a local high-speed ram under system software control. Up to 2048 test instructions may be stored in this ram at any time and a high test speed is achieved by making the speed of test execution independent of the LSI11 computer. The ram outputs at 1MHz with the test instructions being executed one per microsecond: tests for s.s.i. and m.s.i. devices typically require only a few test steps, so the hardware timeout is not exceeded. A truth-table system drives input and control pins and monitors for outputs as expected from the devices' truth table.

Failure of any part of the test results in device and p.c.b. failure. More complex digital devices are tested using the same basic method, although the tests do not follow the same truth-table approach. This serial test method may be supplemented

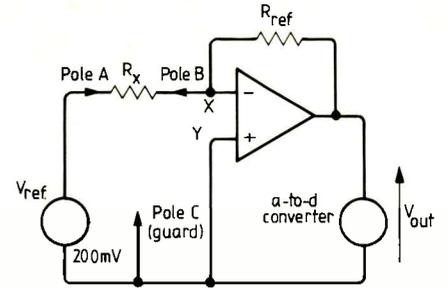


Fig. 6. Resistance (R_x) measurement in basic form.

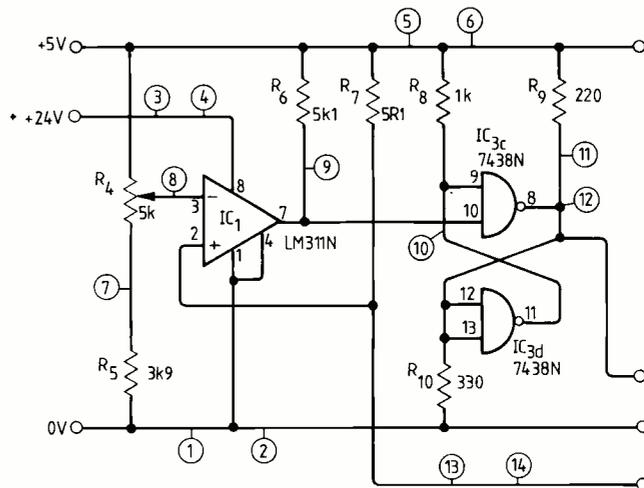
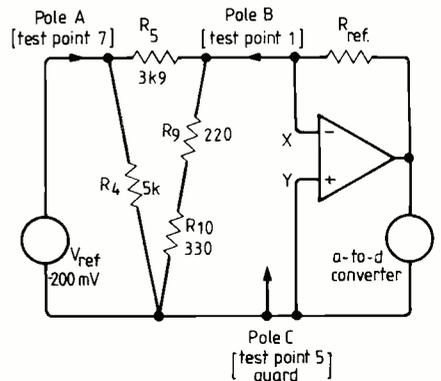


Fig. 7. More usual resistance-measurement circuit, in which shunt resistance affects measurement.

Fig. 8. Circuit arrangement in which shunt resistors of Fig. 7. are either shunting the source or across the equipotential points and do not affect reading.



with a parallel test method, setting up a series of pin states simultaneously and achieving a much faster test rate.

FUNCTIONAL TESTING

The main advantages of in-circuit testing are; inherent fault diagnosis to component level, low programming cost and skill requirements and short programming times. The disadvantages of this method are relatively high recurring fixture costs and the inability to detect faults which only become apparent at or around the p.c.b.'s normal operating speed, for example timing problems. These are especially important in some applications, for example military products, where these faults must be found at the board test stage, rather than final test. For example, the electronic systems in a rocket only have one chance.

This latter type of fault is located by another a.t.e., the functional tester. The function test program injects a pattern of stimuli into the inputs of the p.c.b. and monitors for expected results on the outputs. If the p.c.b. fails, diagnostic software, using a technique known as reverse trace, guides the operator back through the p.c.b. network from the output, prompting the operator to probe specific nodes with hand-held probe, wired back to the a.t.e., until the failing component is located.

Generating the test program and software is quite a complex task and is usually carried out by a software package known as a simulator. One widely used simulator will run on a variety of functional a.t.e. systems or on a mainframe computer. To generate the necessary software for functional testing, the programmer must take the following steps.

A database or image file, containing details of all devices on the p.c.b. and their interconnections, must be written. To do this, the programmer simply specifies each component connection (e.g. IC₁ pin 1) and the track or signal name to which it is connected. When processed by the simulator the signal names reference particular component pins on the same node to one another, producing a network description of the p.c.b.

The database is now run through the first part of the simulator where mathematical models of each device fitted are located and combined, using the network description of the p.c.b., to produce a mathematical model of the whole board. This is in Boolean algebraic form and it is using this model that the sequence of input stimuli and expected output results are determined.

The simulator now produces series of input stimuli and applies them to the mathematical model of the p.c.b. When this process is complete the following files are produced:

- the test program;
- a file indicating the percentage of possible faults that will be detected by the test program;
- a list of untested nodes;
- wiring lists for fixture manufacture;
- the diagnostic files to be invoked should a p.c.b. test fail.

An important requirement of functional

testing is good node visibility: i.e. should any single node fail that failure must be flushed through to and be visible at the p.c.b. output. If good node visibility is not achievable a partial bed of nails fixture may be required to supplement the standard edge connector 'fixturing' and provide the necessary node visibility.

When creating and debugging files the programmer has some additional software tools available: a database creator aids the programmer at the p.c.b. description-generation stage: an emulator, part of the simulator, creates mathematical models for complex digital devices; and a logic analyser shows the logic state of any node at any stage during the test and is particularly useful when creating and debugging new mathematical device models.

Functional testing has several advantages over in-circuit testing. Fixturing costs are low, and the p.c.b. is tested at or near its actual operating speed (up to 30MHz on some functional testers). Further, timing problems in digital circuits can be detected and their source identified. P.c.bs which have been conformally coated can be tested, since the edge connectors are clear and the diagnostic probe is able to pierce the coating. In-circuit testers must test p.c.bs prior to the application of the coating. Additionally, surface mounted devices do not pose any special problems, whereas true in-circuit testing sometimes requires the use of special (and expensive) fixturing.

However, in-circuit testers have advantages over functional testers, in that functional hardware is much more expensive than in-circuit testers, and in-circuit testing requires a much lower skill level than functional testing, typically taking one fifth of the time to write, for the same p.c.b. A functional tester can only find one fault on each test run and these often take some time to locate using the reverse trace method. In-circuit testers find all faults on a single pass in a much shorter time.

PRODUCTION LINE USING ATE

Once installed in the production line, a.t.e. provides a number of benefits. Throughput is increased: for example an in-circuit test time around 10 seconds would not be unusual for a p.c.b. comprising 60 discrete and 60 digital components (including track testing). Quality is improved as the level of inspection or test does not vary. For example, manual visual inspection of a p.c.b., in addition to being slower, is dependent on the degree of concentration of the inspector, which varies during the day, and also throughout the week.

When the a.t.e. is fully utilized in testing p.c.bs, it may become difficult to find sufficient free time to write new test programs. Off-line programming on a multi-user workstation reduces the time required on the a.t.e. to that needed for fixture and program debug.

Efficiency may also be improved in the rework area. When faults are found by an in-circuit a.t.e. they are printed out on a paper ticket. This contains details of the component identity and its failing value if analogue, or device type if digital. But as

component packing densities increase, the p.c.b. legend is usually omitted, making component location difficult. If a track short is found, the a.t.e. can only report the test point numbers between which the short circuit occurred. On a large double-sided or multi-layer board it can take the repair technician many minutes to simply locate the fault. Computer aided repair (car) provides a solution: the paper ticket is replaced by a software fault file, sent to a multi-user workstation by a a.t.e. via a wire link. When the board goes to the rework area for repair, the technician enters its serial number, possibly using a bar-code reader, and the fault file belonging to that p.c.b. is read. Faults located on the test run are displayed on a full colour graphics display which shows the location of all components or tracks on that board type. Fault location time is greatly reduced and productivity in the repair area increased.

Information about the types and numbers of faults from both the a.t.e. and car can be put together by process-management software to give a 'real-time', up-to-the-minute picture of production. The most frequently occurring failures, identified either by component identity, or failure type, for a particular p.c.b. production run is displayed. Analogue failures may also be checked for the spread of failures around the nominal value. Using this software, earlier process errors, e.g. a wrongly set flow solder bath, may be identified and the line stopped until they have been corrected.

THE FUTURE

A.t.e. capabilities have improved, some of today's low-cost bench-top a.t.es having a superior performance and specification to the larger testers available a decade ago. Many manufacturers are attempting to combine the capabilities of in-circuit and functional testers into a single test system. These combinational testers will overcome the limitations of the two system types, although no company has yet succeeded in bringing a true combination tester to market at a realistic price.

But perhaps the most important development in the industry is that of data sharing. Real-time fault analysis is one example of how data gathered at the a.t.e. stage can be put to good use elsewhere. Another example is taking the p.c.b. design information available in a computer-aided design system files and automatically producing an a.t.e. test program. Such a system is already available and also generates car and fixture design files, greatly reducing development costs and times.

The future of the industry lies in integration. Automation reduces manufacturing costs, but it is integration which reduces the applications costs associated with the introduction of new products (e.g. test programs, manufacturing information). When these are overcome the 'factory of the future', much heralded within the industry, will be much nearer. A.t.e. is playing a key role.

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Television standards conversion

Thirty years development – from image transfer to digital conversion and onwards

S.M. EDWARDSON and C.K.P. CLARKE

The conversion of television signals that have originated on one scanning standard to provide signals on a different scanning standard presents problems that have taxed the ingenuity of television engineers for well over thirty years. In that time, the methods used for conversion have undergone several major changes: first from optical methods to analogue signal processing and then to digital signal processing. These changes have come about partly as the result of changes in technology and partly through a progressively improving understanding of the underlying theory.

The conversion methods have developed from the relatively simple techniques used to alter just the number of lines per picture, while keeping the same field rate, to those in which the field rate is altered as well. As the conversion methods have developed, there have been corresponding changes in the ways of visualizing the action of the conversion process. Some have been based in the time domain, synthesizing intermediate lines or fields and correcting their timings, while others have used a frequency-domain representation to view the process in terms of low-pass filtering and resampling.

Transatlantic conversions, changing between the 625/50 (625 lines, 50 fields per second) and 525/60 standards, (which grew out of the different mains electricity frequencies used in Europe and North America) have become particularly important to allow the exchange of programmes between these two major programme producing areas. A further complication that has arisen over the same period is the inclusion of colour. This has prompted the development of special techniques for handling colour at each stage in the evolution of converter technology.

IMAGE-TRANSFER CONVERTERS

The image-transfer or optical method of standards conversion¹ consisted of pointing a television camera, working on the output standard, at a television picture on a cathode-ray tube display (c.r.t.) working on the input standard. In principle, therefore, the intention was to reproduce an image of the original scene in front of the camera so that it could be rescanned on the new standard. Clearly, in such a circumstance, the geometrical matching and the stability of the two scans were very important to the

effectiveness of the technique. The need for synchronizing the two signals, so that the camera scan followed appropriately after the c.r.t. scan, was also important.

Flicker and movement portrayal. The combination of the c.r.t. and the camera differs somewhat from the case in which a television camera views a normal scene. The light from a normal scene falls continuously on to the camera tube, to be integrated as an accumulating charge, whereas the light from the converter c.r.t. is not continuously present. Instead, a bright 'writing' spot traces the raster on the c.r.t. screen where it remains only during the afterglow period of the phosphor, which has an approximately exponential decay. With normal phosphors the persistence is very short, being only a few line periods, so that the image dies away rapidly. The 'reading' beam of the camera then discharges the integrated signal from the previous field period, but the signal level varies with the relative field phasing of the c.r.t. 'writing' beam and the 'reading' beam in the camera tube.

This problem is particularly noticeable when converting between standards with different field rates^{2,3}. Then there is either a surplus or deficit of fields or parts of a field within the scanning period of the converter camera tube. The display brightness at any point in the field is renewed as the scanning spot passes and then rapidly dies away. If the field period of the output standard is of shorter duration than that of the incoming signal, then, at one particular phase relationship, the camera receives no exposures at all. Similarly, an output field longer than that of the incoming signal can embrace two full exposures from the short-persistence display tube. Either of these conditions creates a brightness variation in the converted signal at the difference frequency between the field repetition rates and results in a bright or dark bar travelling vertically through the picture. However, it was found that this problem, and the related one of field phasing, could be reduced substantially by the use of a c.r.t. display phosphor with much longer persistence than normal.

The advantage of using longer persistence c.r.t. phosphor was lessened somewhat by the fact that it exacerbated the problems of movement portrayal. Moreover, the three

camera tubes in general use at that time (image-orthicons, cathode potential stabilized Emitrons and vidicons) all tended to leave residual images. These resulted primarily from the incomplete erasure of the stored charge by a single scan of the camera tube target and an image could still be discernible as many as ten fields after it was originally scanned. This effect, in combination with the use of long-persistence phosphors in the c.r.t. produced rather poor portrayal of movement, the main impairment being due to multiple imaging.

Line structure and vertical resolution. 'Line beating' was a problem that arose when the number of lines per picture was changed. This was again a form of brightness modulation of the output picture, at the difference between the two spatial line-frequencies. It resulted from the discrete nature of lines on the c.r.t. display which thus did not form a continuous image. The scanning lines of the camera tube could then fall between or on top of the lines of the original scan as the phasing of the two line structures varied down the picture. This effect, coupled with slight geometrical differences between the rasters in the c.r.t. and the camera, gave rise to moiré patterns on the converted picture. These patterns could be quite severe and had to be reduced by broadening the display spot, for example by introducing sinusoidal vertical 'spot-wobble', although this unavoidably had an adverse effect on vertical resolution. Because of these effects, all image-transfer converters were operated on a field-by-field basis, thereby losing any extra vertical information carried in the interlaced fields. An additional benefit of this field-by-field approach was that it tended to improve the portrayal of movement, by preventing double imaging (resulting from displaying two fields at once) from adding to the multiple imaging effects caused by the dynamic characteristics of cameras, as described above.

In the circumstance where the input and output field frequencies could be made exactly synchronous, it was possible to select a particular phase of the input and output signals in which the vertical resolution was significantly improved. This was achieved by phasing the output scan to take equal contributions from two interlaced input fields.

Unfortunately, this phase was found to be the most disadvantageous for movement because the use of two fields gave a double edge to any moving object, in addition to the normal multiple-imaging effects of the display tube and the camera.

Standards conversion of colour pictures.

An extra problem that was addressed in later field-rate image-transfer converters was that of converting colour signals⁴. This was approached by adding a second c.r.t. and camera, amounting to an additional standards converter. The first then carried the luminance component of the picture signal whilst the second carried the colour difference signals in the form of a phase modulated carrier. This added a new dimension to the problems of matching, requiring accurate correspondence between both the c.r.t. and camera scans in both the luminance and colour channels. Colour transcoding was achieved by, first, transcoding from the original colour system, either PAL, NTSC or SECAM, into the special form of coding used only during conversion and then transcoding again to provide the required output colour signal.

Limitations of image-transfer methods.

Throughout their development, only limited facilities were available for controlling the performances of image-transfer converters, such as choosing a display phosphor time-constant or adjusting optical or beam focus. In essence these techniques were used to filter out the unwanted products of the original scanning standard, but did not give much freedom of control of the low-pass filter characteristic. In practice, the c.r.t. could not give out 'negative' light, so limiting the effective scanning aperture to positive-only values. Because of this, only a relatively slow roll-off frequency characteristic could be achieved, which resulted in a noticeable loss of picture detail. Similarly, flare in the various optical components, especially the glass face-plate of the c.r.t., also caused a loss of detail, particularly in bright areas of picture.

For moving pictures, the temporal filtering effect was determined by the exponential decay properties of the c.r.t. display and the camera target discharge process. This resulted in an asymmetrical impulse response, akin to that of a recursive filter, and was less suitable than a symmetrical response such as might be produced by a transversal filter.

In converters used to change both the number of lines in the picture and the number of fields per second, the problems of maintaining accurate matching between the scan geometries of the c.r.t. and the camera were increased. In addition, because the timing of the scans relative to one another was changing systematically, flicker, particularly that associated with line beating, became more noticeable. The use of long-persistence phosphors in the display did not

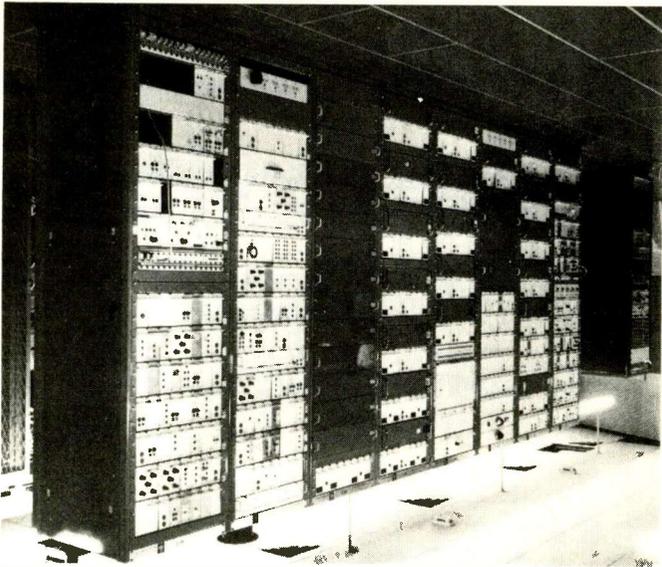


Fig.1. BBC analogue converter at Television Centre.

provide a complete answer to this problem. It ensured that the amount of light integrated by the camera tube did not vary appreciably as the phasing of the two scans altered, but had the disadvantage of making movement very blurred, as stated earlier.

Over and above these problems and shortcomings, there existed the need for careful maintenance and skilled operation. The cleanliness of the optical system and the adjustments of focus and raster sizes are just a few examples of the requirements to produce consistent and acceptable results.

ELECTRONIC ANALOGUE CONVERTERS

The fundamental nature of the problems with image-transfer converters made it very desirable to seek a wholly electronic solution in which the signals did not have to go through an intermediate optical form. Electronic conversion had the potential to eliminate many of the problems already outlined and to give much greater freedom to control the filtering effect of the transfer. However, whereas the processes involved in an image-transfer conversion are reasonably easy to visualize, those in the non-optical, electronic case are more difficult.

In electronic standards conversion, the picture signal occupying each active-line period of the input television signal was seen as a discrete element to be reformatted to match the synchronizing pulses of the output standard. A means of changing the number of lines in the picture was needed so that some lines could be discarded or new lines produced, and processes such as time stretching or compression were used to make the new lines fit the active period of the output standard. The filtering action implicit in this transfer of signal information was determined by controlling which input lines were used in the formation of each output line. These processes required the development of methods of storing portions of the input signal until required for the output, of altering the line duration and of combining lines together.

Changing the active-line duration.

An electronic store for stretching or compressing input lines to match the active-line period of the output standard was formed by a series of capacitors⁵. Each capacitor was used to hold a charge proportional to the picture brightness at one point in the line, so as many capacitors were needed as there were discrete picture elements in the active-line period. The charge was stored by using electronic switches to connect the input signal voltage to each capacitor in turn as the line scan progressed. Then the change of line duration could be achieved by 'reading out' the stored values in the same order, but at a different rate.

The use of discrete storage elements in this way caused the input signal to be sampled and may well have constituted the first application of 'horizontal' sampling in a television context. Because of this, some pioneering work into the filtering requirements and the impairments resulting from video sampling was carried out at this stage. This showed that, for the video signal bandwidths then in use, about 550 elements (capacitors) were needed for a time-redistributing line store⁶.

With the analogue devices then available, there was a certain amount of variability in the efficiency of the switching and storage circuits. As the same switching and storage unit was used for corresponding picture elements on consecutive lines, this variation caused a modulation of the output signal which could appear on the converted picture as a fixed pattern of vertical striations. Very careful matching was required to reduce this pattern to a sufficiently low level to avoid a visible pattern⁷.

Line-store conversion.

With standards converters operating between asynchronous input and output field periods, any interval up to a complete picture period may elapse between the moments at which a given picture element is 'written' at one standard and 'read' at the other. In such converters, therefore, provision must be made to store an entire picture period of video information in the converter. In terms of the capacitive storage cells described in the previous section, this would have required about 200,000 separate capacitors. In an image-transfer converter, this storage capacity was provided by the persistence and the camera charge integration process. If, however, two standards having the same nominal field period are rigidly synchronized, so that their field periods are coincident, it can be arranged that each picture element is read out of the converter within one output line period of being written in. In that case, therefore, it is necessary to provide storage for only one line of video information. Such a device, known as a line-store converter, was then able to replace the image-transfer method in application such as 625/50 to 405/50 conversion^{8,9}. This avoided the careful adjustment

and constant attention from a skilled operator required with the image-transfer equipment. Because of this, it was possible to install a line-store converter at each 405-line transmitter, thus avoiding the need to provide duplicate distribution for the 405 and 625-line broadcasts.

Nevertheless, the construction of suitably matched storage and switching devices operating at, for the mid-1960s, very high switching speeds represented a considerable problem. The main difficulties were those of distributing the video and switching control pulses to the successive stages of the store without allowing crosstalk from the input to the output. This was accomplished with manageable complexity by using diode-bridge switches and low-loss transmission lines for the video and pulse signal distribution¹⁰.

Ultrasonic delay lines. In circumstances in which the input and output field rates are different, or where the input and output fields cannot be rigidly synchronized, much larger amounts of storage are required. This can be provided by storing the signal in quartz ultrasonic delay lines¹¹. In analogue field store standards converts the basic conversion operation was achieved by a network of switched delay units and by-pass routes that determined the delay to which each television line was subjected. In this 'main store' process, lines could be repeated or omitted so that the average line frequency of the output signal from the main store was that of the final output standard¹².

Ultrasonic propagation through a solid medium can be up to 100,000 times slower than that of electromagnetic waves in free space. In addition, certain substances exhibit fairly low absorption of acoustic or ultrasonic energy and therefore make suitable transmission media for delay devices. Glass can be used for fairly short delays, although natural or synthetic fused quartz is preferred for longer delays in which lower absorption is needed. A long transmission path is obtained in a small volume by adopting a form of construction in which the acoustic wave is internally reflected between the facets of a polygon. The acoustic wave is launched into the quartz and collected from it by means of piezoelectric transducers, such as quartz crystals.

In the development field store standards converters, the main factors found to limit the performance of such delays were their insertion loss, their group-delay and amplitude-frequency characteristics, and their susceptibility to producing spurious signals. Many of these factors were improved by amplitude and group-delay equalization¹³ and by operating the longer delays at a high temperature (75°C) and at a common centre frequency of about 30MHz. Because the value of delay was temperature dependent, the quartz delay lines were housed in accurately controlled ovens to maintain stable temperature during operation.

Several factors contribute to the insertion loss of a delay line. The transducers, which use lead backing to damp their frequency and group-delay characteristics and to reduce reflections from the transducers, pro-

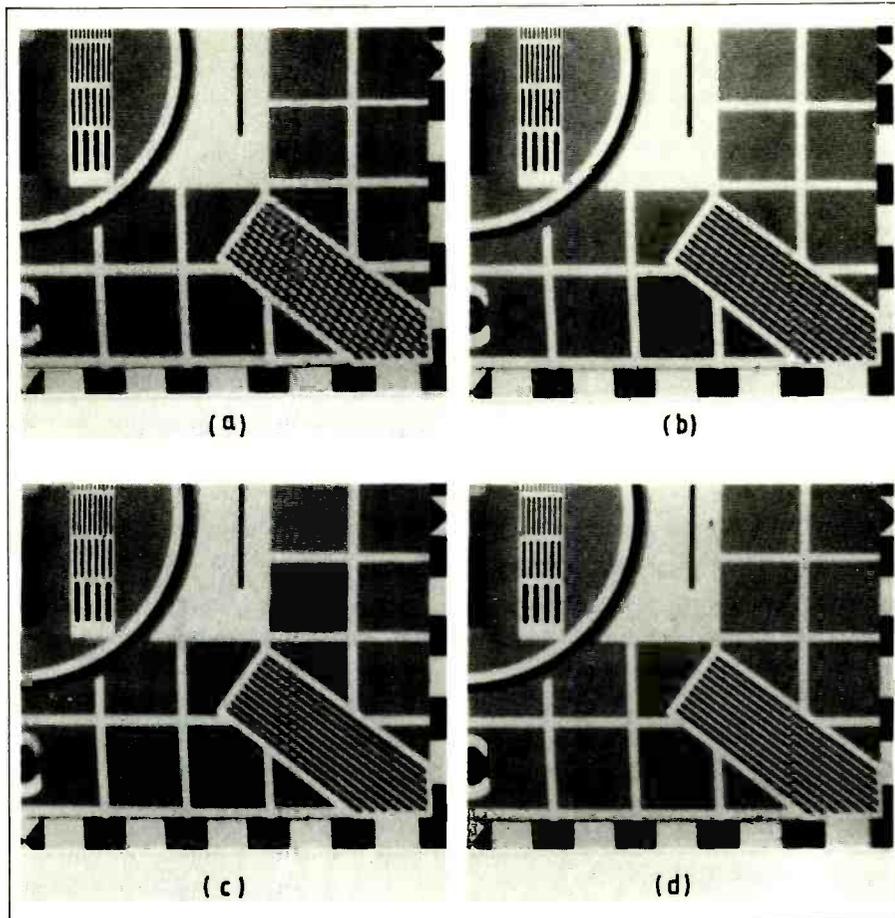


Fig.2. Line-store conversion: zero-order interpolation(a); 2-ordinate linear interpolation (b); and 4-ordinate optimized.



Fig.3. Quartz delay line.

duce about 36dB loss. Beam spreading and energy absorption losses in the quartz depend on the length of the delay, but can each contribute another 10dB, making a substantial loss overall.

Spurious signals could reach the output transducer either before or after the main signal. These could arise partly through the presence of faster modes of propagation along the main path and through the existence of alternative paths in the quartz block. Although very numerous, most, but not all,

were of very low level. In the case of longer delays, the most serious spurious signal was the so-called 'third-time round' component. This resulted from successive reflections at the output and input transducers, the unwanted signal finally emerging after having traversed the prescribed path exactly three times. In addition, direct electrical coupling between input and output circuits was found to cause an unwanted signal with zero delay. These effects were covered by careful specification¹⁴ and minimized by using spe-

cial design and manufacturing techniques.

Frequency modulation. In a field-store standards converter, the relative phasing of the input and output fields can vary rapidly from zero to the maximum value. Delay changes were effected by the systematic switching of delay units in and out of circuit. Even a small change of delay could involve the switching in or out of several delay lines at once and any consequent variation in gain could appear on the output picture as flicker or moving horizontal banding, or both. Because of the large number and the wide range of switched delays used in field-store standards converters, care was necessary in equalizing the gains and responses of the many signal paths.

This problem was greatly reduced by using frequency modulation of the ultrasonic carrier used in the delay lines, instead of amplitude modulation. This not only suppressed the effects of gain inequalities, but it also substantially reduced the picture impairment resulting from the spurious signals produced in the delay lines.

Interpolation. The earliest experimental standards converters simply omitted or repeated the lines and fields of the input standard as necessary to produce the output standard signal. This use of the nearest line from the nearest field amounted to a zero-order interpolation process and led to two important forms of picture distortion: geometric distortion and impairments to movement portrayal.

Geometric distortion is caused by the disturbance of the relationship between picture information and the line structure. This appears as serrations on moving edges and, due to the effects of interlace, spurious movement within the picture. For moving objects, the omission or repetition of fields results in smooth movement being reproduced as a series of jumps, known as judder. Higher-order interpolation processes were, therefore, investigated as a means of producing improved performance.

Higher-order interpolation requires the weighted addition of delayed and undelayed versions of the signal. Having chosen the use of frequency modulation, it was then desirable to use frequency modulated signals in the interpolation process; this calls for a frequency mixing process, such that the input frequencies (and hence their modulations) are added. Otherwise, each weighted addition would require the demodulation of each of the contributions followed by subsequent remodulation. Conventional methods of mixing use, for example, a bridge-ring modulator to multiply the two carriers together to generate a carrier with a frequency equal to the sum of the two input frequencies. This is unsuitable here because the input frequencies are close to one another, so that second-harmonic components produced by the modulator would interfere with the wanted sum frequency. Instead, a novel method of frequency averaging was developed from the principle that addition or subtraction of equal-amplitude carriers produces half-sum and half-difference frequencies¹⁵.

The frequency averaging process limited the complexity of the interpolation methods so that only relatively simple binary ratios could be used. Nevertheless, this produced useful improvements in picture quality. One of the early analogue field-store colour standards converters^{16,17} had three interpolation methods, known as 'field', 'line' and 'mixed' interpolation. The converter was normally operated in the mixed mode.

With field interpolation, the two lines from adjacent fields nearest to the required output line position were averaged. This avoided flickering impairments on vertical detail, but produced a double-image effect on moving vertical edges. Line interpolation, on the other hand, combined weighted proportions of the two adjacent lines from the input field nearest to the required output line position. Using proportions of three-quarters of the nearer line of the two and one-quarter of the other line gave a good compromise between complexity and the suppression of serrations, but introduced flicker on vertical detail and did nothing to reduce judder. The mixed interpolation method consisted of a combination of line and field interpolation in which one or other method was used depending on the relative field phasing of the input and output signals. When the input and output field timing was nearly coincident, the line mode was used, while when the output field position fell roughly mid-way between two input fields, the field mode was therefore to reduce judder on moving objects by producing a representation of the image closer to its actual position. The performance on vertical detail was somewhat poorer than that of the field mode alone, due to the introduction of flicker, but this was not as serious as that in the line mode and was offset by the considerably improved movement performance.

For analogue implementation, the mixed interpolation method, with its relatively large number of possible contributions and the switching required to change between the different modes, must have been close to the limit of practicable complexity. Also, the interpolation methods were generally thought of as time-domain processes, producing lines appropriately positioned in time and space. With this approach, therefore, it becomes increasingly difficult to visualize how the interpolation methods should be developed to improve their performance.

Electronic standards conversion of colour signals. With normal PAL and NTSC colour signals the instantaneous phase of the colour subcarrier changes from line to line. If these signals were used for transmission through the interpolation processes of a standards converter, it would not be possible to average successive lines for interpolation. To avoid this difficulty, the input signal was transcoded to an alternative 'intermediate' form.

The intermediate signal used a colour subcarrier frequency which was an integral multiple of the input line frequency. In addition, the signal was similar to that of NTSC and did not include the alternate-line phase switching of PAL. Moreover, since it

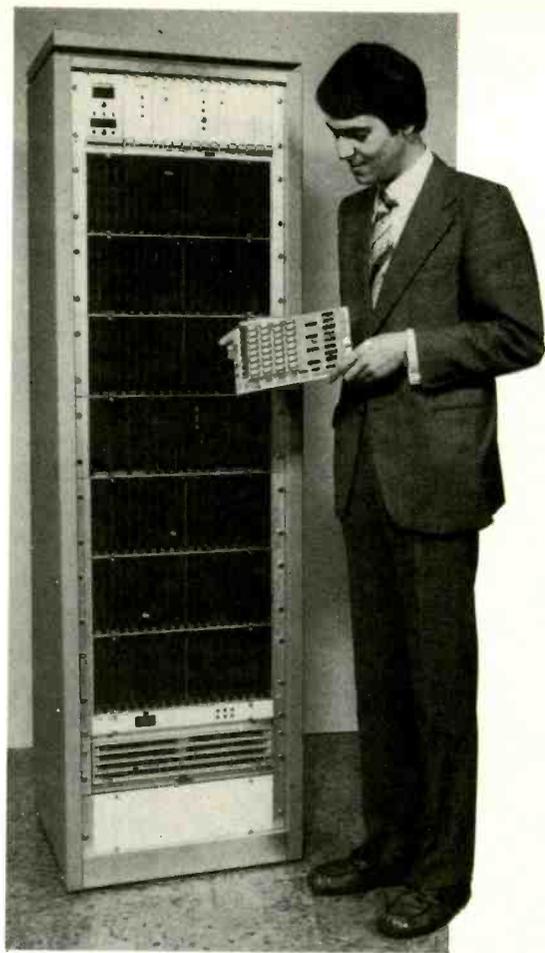


Fig. 4. A 1978 digital television field store on one board. The rack is a 1973 field store.

was necessary to separate the luminance and chrominance components of the intermediate signal after standards conversion, the chosen frequency of 4.5MHz placed the chrominance signal outside the rather limited luminance range. This particular frequency had the advantage of being an integer multiple of the output line frequency as well. This minimized the possible effect of beats, because it ensured a half- or quarter-line offset with the NTSC and PAL subcarriers, respectively.

DIGITAL STANDARDS CONVERTERS

Replacement line-store converters. The development of new digital integrated circuits with sufficient speed for video signal processing opened the door to the development of digital line-store converters¹⁸. At this stage it was clear that the UK 405-line service was going to be maintained for a considerable further period. In view of this, the replacement of the analogue line-store converters by more stable and reliable digital equivalents was easy to justify. Moreover, the very high signal quality and immunity to distortion of digital implementation brought with it the possibility of using more contributions and a wider range of coefficients in the interpolation process. This held out the possibility of significantly improved interpolation.

The two fundamental numerical properties of a binary pulse code modulation

system are the sampling or clock frequency and the number of binary digits (bits) used to describe the value of each sample. For 625/50 PAL colour signals, a sampling frequency of about three-times the subcarrier frequency was conventionally used (about 13.3MHz). However, for 625-to-405-line conversion it was possible to reduce the sampling frequency because of the relatively restricted upper frequency limit of the 405-line system. This, coupled with the necessity of removing the colour subcarrier and most of its significant sidebands, allowed the use of 729-times line frequency sampling (approximately 11.4MHz).

The number of bits used in a digital system must be sufficiently large to ensure that no visible 'contours' occur when an area of almost uniform brightness is being scanned. In common with other broadcast quality television signals, 8-bit digital signals were used for the conversions between analogue and digital signals at the input and output. Also, somewhat greater precision was used internally at critical stages to avoid a build-up of quantizing distortion.

Apart from the input and output conversions, the main processes in a digital standards converter are digital signal storage and digital arithmetic to allow the summation of weighted contributions from different lines. The store units of the digital converters were based on digital shift registers. Reasonably low power consumption was achieved by using mos technology, in which the individual transistors only draw appreciable current when changing state. Even so, the maximum operating frequency of these devices (2MHz at that time) was insufficient to carry the input samples directly and it was necessary to demultiplex the signals into six parallel paths to achieve the necessary operating rate. Thus six 100-element registers were a convenient means of providing the storage necessary for one television active-line period.

New lines are formed by combining weighted proportions of lines from adjacent positions in the field. In digital arithmetic, the multiplication process consists of repeatedly shifting the significance of the input sample value and adding the result to a total under the control of the individual bits of the multiplier coefficient. In early line-store converters, each multiplier was assembled as an array of binary adders and gates. The complexity of this array grew in proportion to both the number of bits per sample and the number of bits per coefficient. Thus, at first, the precision of the coefficient was limited to three bits, allowing coefficient values of one-eighth, two-eighths, etc. to be used. Another means of limiting the complexity of the arithmetic in the early experiments was to use linear interpolation. Because the coefficient values were always C and $(1-C)$, a single multiplier could be used operating on the difference between two sample values. In later versions, as the level of integration in the digital circuits was increased, the limitations on both the number of coefficients and their precision were progressively removed.

Interpolation in the frequency domain. Un-

til this point, the design of interpolation methods had been restricted by the small number of coefficient values available. The particular set of coefficients to be used was in general determined by approximating the geometrical position of the required line relative to its neighbours. A more theoretical approach was therefore required in order to make full use of the much larger numbers of coefficients available in a digital interpolator. This was provided by using a frequency-domain interpretation of interpolation^{19,20}.

The interpolation process can be considered as equivalent to the processes of low-pass filtering, to return the signal to the form of a continuous image, followed by rescanning on the new standard. This concept is very similar to the mode of operation of the optical, image-transfer converters. In the frequency domain, the scanned signal consists of a series of spectral components, which repeat the baseband spectrum of the image at harmonics of the sampling frequency (the line and field frequencies). The interpolator, in its low-pass filtering role, then suppresses the spectral harmonics to leave the baseband spectrum. Resampling at the scanning rates of the output standard then introduces new harmonic spectra in different positions from those of the original standard.

With this spectrum-based interpretation of the standards conversion process, the signal impairments can be explained in terms of imperfect operation of the low-pass filter. If the filter fails to suppress the harmonics adequately, then the resampling process will repeat the spectra with some components falling into the low-frequency region. As these residual components will be incorrectly related to the original baseband spectrum of the output scanning standard, they will appear in the picture as patterns of the wrong frequency, an effect usually known as 'aliasing'. Alternatively, if the filter has adequate passband response, then some of the wanted baseband components will be lost, and the picture will have a soft or blurred appearance.

Field-rate conversions. Whereas digital implementation and the freedom to use more coefficients improved both the basic signal quality and the interpolation performance of the line-rate converters, the quality of the conversion was always limited by the lack of access to the lines of the adjacent fields. In an interlaced scan, the high-frequency vertical information is shared between adjacent fields. So, by restricting access to only one field, some wanted component frequencies were replaced by unwanted alias patterns. Although, in principle, field storage could have been used, it was always judged that the benefit in performance was unlikely to justify the high cost involved. However, in field-store conversion, digital storage was used almost as soon as it became practicable, despite its high cost.

Cost was still a considerable constraint in the design of early digital field-store standards converters^{21,22}. As in the line-storage converters, mos shift registers were used as the storage elements, but in this case the larger 1024-element devices were suitable.

Even so, the enormous number of devices involved resulted in power dissipation difficulties. Since, in these devices, most of the power dissipation occurs as the charge is transferred from one stage to the next, it was advantageous to clock the registers more slowly when access to the stored lines was not required. Unfortunately, the registers could not be stopped altogether, because the charge would leak away over a prolonged period and this led to the added complication of correct phasing of the data in the registers when the time came for them to be read-out. This involved marking the start of the stored data with a code.

As well as complicating the store operation of early digital field-store converters, the high storage cost resulted in completely different modes of operation in the two directions of conversion. This also resulted, in part, from the converter initially being designed only for one direction of conversion, from 525/60 to 625/50. In this direction, two fields of the incoming signal were stored directly, still in composite NTSC form. This approach reduces the storage requirement in two ways: first, the 525-line standard has fewer lines and, secondly, storing the NTSC colour signal avoids the extra capacity necessary for separate Y, U and V signals. However, separate decoders were needed at the outputs of the two field stores to convert the stored NTSC signals to multiplexed YIQ form for interpolation. This was followed first by movement interpolation, to produce a 525/50 signal, and then line interpolation, to produce a 626/50 signal, with a conventional analogue PAL coder at the output.

Conversion from 625/50 to 525/60, which was added later, was constrained to use the same two field-stores, again in 525-line NTSC form. The PAL or SECAM 625-line input signal was first decoded and digitized to form a multiplexed YIQ signal. Line interpolation was then used to produce 525/50 signals which were digitally coded in NTSC form for storage. After leaving the field stores, the signals were decoded again to multiplexed YIQ form for conversion to 525/60 by the movement interpolator, before finally being encoded once again as NTSC.

In addition to the improved signal quality through using digital storage, the interpolation performance was significantly better than for either the optical or electronic analogue converters. Line interpolation was very similar to the methods used in the earlier line-rate converters although, because the movement and line interpolation were carried out as separate processes, it was only possible to use vertical detail from one field, thus limiting the resolution. The field-rate conversion process was visualized in terms of the displacement of the input fields from their correct positions in time. To counteract this displacement, the movement interpolator used an approximation to linear interpolation, so that proportions of the adjacent fields were used, weighted according to their proximity to the required position. The complication of the multipliers was still limited by cost considerations.

Subsequently, the costs of digital storage and arithmetic fell dramatically and a number of relatively low-cost digital field-store synchronizers were developed. As storage was still the main cost, however, it was possible to add extra circuitry for features such as noise reduction, based on recursive temporal filtering, and special effects. The combination of these extra circuits provided the basic requirements for field-rate standards conversion, although the recursive temporal filtering tended to limit the performance of the movement interpolation. This was primarily the result of the asymmetrical impulse response of the recursive filter which tended to leave a trail of multiple images following moving objects, reminiscent of optical conversion methods. In some cases, movement detectors were added to change the filtering between 'still' and 'movement' modes. However, such converters were always prone to choosing the wrong mode because of the difficulty of detecting movement reliably over a wide range of signal conditions.

Eventually, the constraints on interpolation were removed by the falling cost of random-access memories (rams) and the reduced complexity of making multipliers of full accuracy. In

particular, the use of rams greatly simplified the provision of multiple outputs from the field stores, thereby removing the need for separate line stores for line interpolation^{23,24}. Simultaneous access to four lines from each field was provided by arranging the field stores as four quarter-field blocks, each with its own output. The writing of successive items into different quarters then ensured that four consecutive lines were always available at the outputs. A consequence of this store organization was that it placed a limitation on the maximum size of ram device that could be used. Thus, 4096-by-1 rams could be used, whereas the later 16384-by-1 devices could not, because the minimum store block became too large and could not provide sufficient access points.

It also became possible to combine movement interpolation and line interpolation into a single process, so that lines of the output standard were synthesized directly from an array of stored input lines. This had the advantage that more use could be made of vertical detail carried in the two fields of an interlaced picture, without degrading movement performance.

Experiments showed that simultaneous access to four lines from each of four input fields gave useful improvements in interpolation performance over the use of fewer contributions. However, with such a large number of contributions, it was no longer feasible to visualize the movement inter-

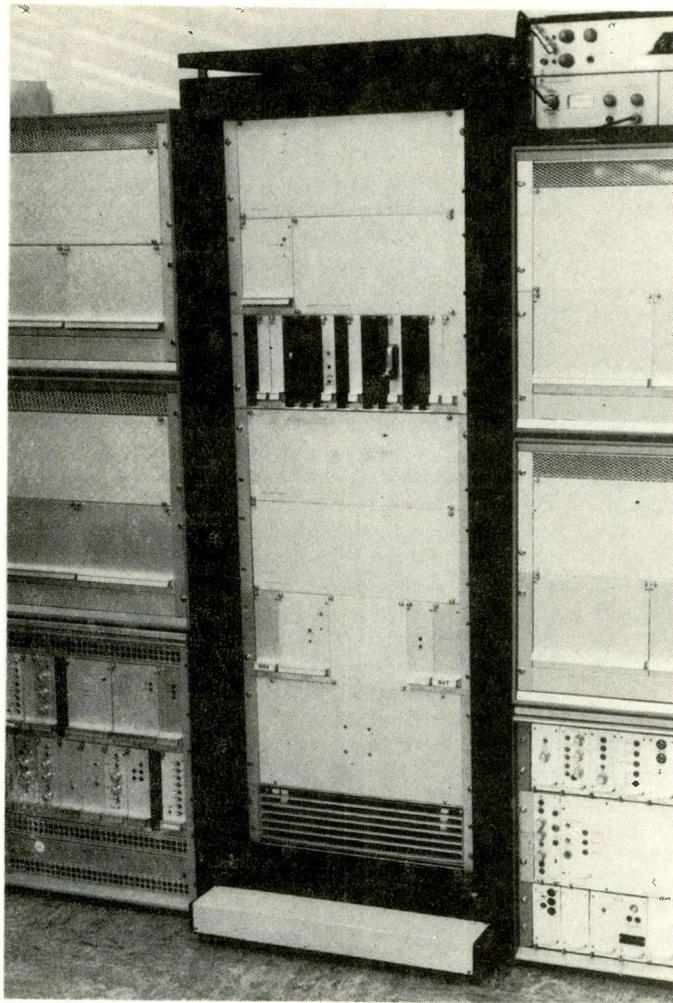


Fig.5. Rack of instrumentation for studying interpolation apertures.

polation process in terms of positional displacement. Because of this, the frequency-domain understanding of interpolation, applied successfully to the line interpolation process in the line-rate converters, was broadened to cover the movement interpolation process as well²⁵.

In this case, because of the two-dimensional (vertical-temporal) nature of television scanning, it was necessary to define a two-dimensional interpolation aperture function to control the choice of the coefficient values used to weight the input lines. This function then defined the vertical-temporal low-pass filtering action of the interpolator, used to suppress the unwanted scanning products of the original scanning process. The size of the aperture determined the spacing at which the frequency characteristic could be defined. By setting the required frequency response values at this predetermined spacing, the associated aperture function could be calculated by two-dimensional Fourier transformation. Initially, the relative importance of retaining or suppressing parts of the spectrum was not known, but the apertures were optimized by comparing the performances of different apertures using a versatile experimental standards converter. With these techniques, further improvements in the conversion quality were made, virtually eliminating impairments to still pictures and substantially reducing those on movement. Nevertheless, some residual move-

ment judder effects remained, which became particularly noticeable on moderately fast camera panning.

In practical terms, this later generation of digital field-store converters benefitted from a rationalization of the organization of earlier converters. Thus, the converters were truly reversible, having the same organization in both directions of conversion. Cheaper storage allowed the signals to remain as multiplexed digital YUV components throughout, so that the storage and interpolation processes were common to luminance and colour interpolation. Because of the relative bandwidth requirements of the luminance and colour components, the input standard video was stored as an interleaved sequence of five words YUYVY. At the multiplexed data rate of 16MHz, this provided sufficient samples for bandwidths of 4.2MHz for luminance and 1.3MHz for each colour difference signal. However, a further feature of this converter was the use of different interpolation apertures for the luminance and colour difference signals. This required the multiplier coefficients to change at sample rate rather than line rate, but

to reduce the vertical and temporal bandwidths' of the converted chrominance signals, thus suppressing unwanted noise and cross-colour components left by the input standard colour decoder²⁶.

FUTURE DEVELOPMENTS

The full potential of the frequency-domain approach has been exploited by the present generation of converters, but their performance is limited by aliasing built into the input signal. This is present because the camera integration process fails to filter temporal frequencies in the signal to within the Nyquist bandwidth. Because of the resulting overlap of true and aliased frequencies, it is impossible to separate the two.

Experimental work on high-definition television (h.d.tv) has produced a promising new approach to the standards conversion problem²⁷. This uses a technique known as movement compensation to overcome most of the residual judder effects of standards conversion. When the motion of an object from field to field is reasonably predictable, such as during camera panning, the probable position of the object during an arbitrarily-timed output field can be calculated. Whereas a conventional interpolator would just combine the input fields in weighted proportions, the movement compensator would alter the positions of the moving object to match the required output position and then interpolate. Thus, provided that the boundary of the moving object could be identified accurately, judder would



Fig.6. Split-screen picture with two apertures. Apertures 10 (left half) gives vertical aliasing visible in the diagonals, while aperture 27 virtually eliminates the aliasing.

could be identified accurately, judder would be eliminated. Less predictable movements, however, such as the leg movements of a galloping horse, could not be accommodated.

The development of standards converters over a considerable period of time has been the result of successive improvements, both in the technologies used and in the supporting theories. Each change of technology, first from optical to electronic analogue and then from analogue to digital, has been accompanied by the need for new ways of visualizing the conversion methods to obtain a better understanding of the processes involved. In particular, the alternative approaches of viewing conversion either in the time domain or the frequency domain have now almost come full-circle, with the time-domain method of movement compensation perhaps about to take over from frequency-domain methods.

ACKNOWLEDGEMENTS

Many individual engineers and many enterprises deserve credit for the evolution of television standards converters. However, it is felt that it would be invidious to name individual engineers or enterprises in this paper. Nevertheless, the list of authors in the reference does give a fair indication of the key innovators in the field.

The permission of the BBC Director of Engineering to publish this paper is gratefully acknowledged.

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NEW PRODUCTS

COMPONENTS

Pressure transducers

Two new pressure transducers are available from HBM; the P12 gauge-pressure device for level measurements and P9V absolute-pressure transducer with integral amplifier. HBM's P12 is designed specially for hydrostatic pressure measurements with long-term stability such as reservoir water levels. Models are available with ranges from 0 to 0.1bar and 0 to 20bars. The inductive measuring system can be connected to any HBM 5kHz carrier frequency amplifier.

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217 on reply card.



Easy p.c.b. design

Protel-PCB is a software package which enables p.c.b.s up to 32in by 19in and with up to six track layers to be designed on an IBM personal computer (PC, XT or AT and compatibles with colour graphics board). The program is claimed to be the first of its kind at under £1000.

Users can select from nine grid sizes, with grid lines 1, 5, 10, 25, 50, 100, 125, 156 or 200 mils apart. Resolution is just 1 mil, which allows precise placing of components such as edge connectors with odd spacings. Five zoom levels are available, from 1.6x0.95in. up to the full workspace capacity.

A special feature of Protel-PCB is its track-stacking capability, by which it can memorize and repeat a structure – a useful time-saver in laying out memory i.c.s. Auto-

routing is not provided, but rubber-banding makes it possible to drag any component to a new position complete with all its connections – though any tracks which tangle as a result must be edited afterwards.

Finished artwork can be produced on a plotter as a multi-colour check-plot or as separate track layers, with component overlays, solder resist overlays and drilling plots. The system also compiles a list of all component used.

Protel-PCB, which comes from Australia, is available in the UK at £799 complete from Engineering Solutions Ltd (telephone 0628-36052). An evaluation disc, which lacks only the ability to save designs, costs £25 (refundable against purchase).
224 on reply card.

Rom, i.c. tester and programmer

Available from Petrtec is the Triple Crown 700, an i.c. tester, eeprom programmer and memory tester. The instrument can be run on the IBM PC and Apple II series and can be used to transfer private or commercial programs onto eeproms, test digital i.c.s for both d.i.y. and commercial circuitry or be turned into a memory tester for customized devices.

Manufactured by Computer Service Technology, the device can test some 700 pre-programmed t.t.l. and c-mos i.c.s. It also allows the user to devise a logic table for testing non-standard or custom digital i.c.s. As an eeprom programmer the 700 will read, programme and edit the eeproms, load and save the program onto disk. In the memory tester mode it will exercise ram and read and verify the roms and prompts to permit the mass testing of the memory devices.
214 on reply card.

Schottky logic devices have high speed

The Signetics N74F00 series of 'Fast' advanced Schottky t.t.l. s.s.i. and m.s.i. devices is available through BA Electronics. Features of the series include an s.s.i. gate propagation of 3ns at 4mW, a flip-flop toggle rate of 100MHz, and an m.s.i. a.i.u. 4-bit add time of 10ns. The N74F00 series also has a supply voltage of 7V and a continuous input voltage of $-30mA$ to $+5mA$.

The N74F00 series now comprises 119 parts and 57 types. Devices new to the range include octal buffers with parity generator/checker; an octal bus transceiver and register; and an octal latched transceiver.
221 on reply card.

VLSI test system

The GR125 VLSI Test System from GenRad tests very-large scale integration (v.l.s.i.) devices. It is a general-purpose system for incoming test of digital i.c.s as well as production tests of v.l.s.i. devices.

Notable features of the GR125 include menu-driven software for test operations and test program development, availability of clocks up to 35MHz, a 25MHz multiplexed data rate and total overall system accuracy of $\pm 2.5ns$. Versions are available which can test 40 and 64-pin devices but clever 'split' architecture enables the pin capacity to be doubled to 80 and 128 pins. Future expansion will cope with 96 or 128-pin devices; again doubled by the split system.

In addition to its menu-driven programming environment, the GR125 offers vector pattern translation for all the Gen-Rad's testers. The 64K pattern memory available on the 64-pin GR125 eliminates the time-consuming reloading of test vectors.

The GR125 runs under a Unix operating system and is based on two Motorola 68000 microprocessors. One processor functions as the system c.p.u. and the second MC68000 runs as the dedicated test c.p.u. Additional local microprocessors are imbedded throughout the system, which control local functions, increasing throughput.
209 on reply card.



NEW PRODUCTS

STE BUS



High-performance STEbus disc controller

An intelligent floppy disc controller is able to control up to four drives over an STEbus.

Based on Western Digital's control i/c, the FDC-1110 incorporates an 8085A-2 c.p.u. and an 8237A-5 memory access controller for local processing and high-speed data transfer. The board communicates with bus masters over the STEbus through a simple two-byte slave interface.

In addition to the 2797 controller, there are software-controllable input/outputs to the disc interface. These make it a simple matter, the makers say, to control the 5.25in drives directly without having to modify the interface cable.

Data rates supported are 125, 250 and 500Kbits/s.

The FCD-1110 has provision for an eprom up to 32Kbyte and static ram upto 32Kbyte forming a 64Kbyte local memory. The STE interface, FDC interface and the d.m.a.

controller all lie within the i/o space of the processor, whilst the memory controller configured for memory-to-memory, STE-to-memory and FDC-to-memory transfers.

Provision of these features allows management algorithms to be executed locally, while the STE master processor boards can be executing other code. This reduces bus overhead since only simple command sequences are required between the FDC-1110 and the STE bus masters.

The STE interface consists of a data byte and a status byte within the STE i/o address space. Data transfers over STEbus may be software-pollled by monitoring the status register. With the local d.m.a controller running from a 4MHz clock it can transfer data between local memory and the STE interface at 1.3Mbytes/s. Datapulse Ltd, Tel: 0491 571955.

228 on reply card

Stepper motor controller for STEbus

Stepper motor control facilities, the first for the STEbus, are provided by Arcom Control Systems' new STMC module. This single-axis controller imposes no processor overhead on an STEbus system c.p.u. all that is required is initialization with a command string to set-up the type of motor and its start and finish speeds. STMC accepts a variety of simple software commands or string for actions including single step, a number of steps control acceleration, move to a reference point and emergency stop.

The module includes an on-board four-phase bipolar driver chip which can output up to 2A per phase. Also on-board are 16 i/o lines, providing general-purpose facilities. Arcom Control System Ltd, Tel: (0223) 242224. 231 on reply card

Stepped packaging for STE

Latest item in Dean Microsystems STEP system is the STEP-1-SUB, a 3U by 84E self-contained 10-slot subrack in which the backplane containing mains input and i/o connectors is an intrinsic part of the assembly. This means that the rack can be removed from the front of an enclosure without the need to disconnect the wiring.

The STEP (STE Packaging) system has been developed to make life easy for STE users. It is a standard packaging system, that calls up all the options essential to the STE user, from power supply to disc drives.

It's unique feature is the stepped front in which the extrusions have been stepped by 60mm to allow board to be set back with room for front access i/o cabling to be routed to the rear, whilst disc drives can be located forward of the boards.

230 on reply card

Kemitron adopt STEbus

Kemitron have chosen STE for their data acquisition and control systems because it is the only international standard bus that specifies a double eurocard option. The format has two advantages: the second connector may be used for i/o connections, and the larger area allows more facilities per board.

With single eurocards, the front of the board is generally used for signal termination and the front panel holds the connectors. If i/o is routed to the rear of the system, ribbon cables are laid across the front of other boards in the rack making them difficult to withdraw.

In the Kemitron design, input or output signals are taken via the second connector. From here ribbon cable runs directly to panels on the rear of the system. This allows the front panel to be used for led indicators and the whole board can be unplugged without affecting any connections.

A secondary advantage of double eurocard is its size. This is greater than that available on two single eurocards since one STE interface is saved and there is a free area in the middle of the board. Kemitron say that one double eurocard can replace at least three, usually more, single eurocards.

STE-based computer uses 64180

Unlike most STE computers, the STE/RIO system accommodates five horizontally-mounted double eurocards, which approach facilitates the routing of i/o signals to the rear of the case. Board front panels are then free for led indicators and each board may be unplugged from the front of the rack without disturbing plant connections.

The front of the integral 19in by 3U rack holds the five slot backplane and a disc-drive module. Several disc options are available including 3.5, 5.25in floppy and hard discs, all supported by the CP/M 3.0 operating system.

The 64180 processor is used on the processor card. Its double eurocard



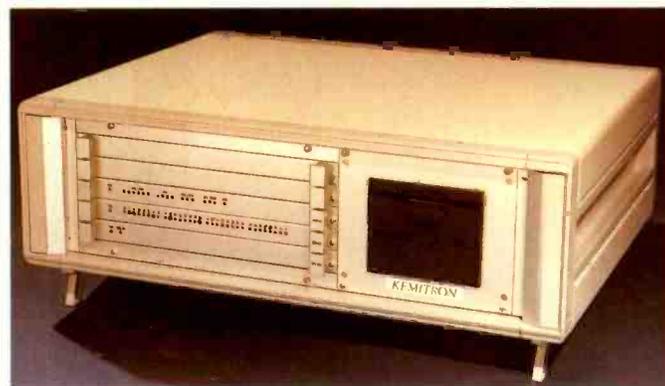
One of Kemitron's first boards which takes advantage of the double eurocard design is a processor card with Hitachi's 6MHz 64180, 768ram, dma, four serial channels, floppy and hard-disc interface, maths processor, counter/timer and r.t.c. with battery. Two interface cards are also available which are a 64/32 channel analogue input with eight analogue outputs and a 32-channel opto-isolated input/output board with 10-channel c.t.c.

Two CP/M 3.0 systems are also available. Dubbed STE/RIO, the name reflects the STE bus, Rear I/O and the double eurocard format. 226 on reply card

design allows the card to contain all the essential system-level features such as 768Kb memory with d.m.a. four serial channels, maths processor, counter timer, watchdog, real-time clock, floppy-disc controller hard-disc interface and the STE bus.

A selection of i/o boards is also available which includes a 32-channel digital input/output/counter/timer card with feedback sensing and opto-isolation, as well as a 64/32 channel single-ended or differential analogue input/output card with sample and hold amplifier. Kemitron Limited, Tel: 0244 536123

227 on reply card



The Archer Z80 SBC

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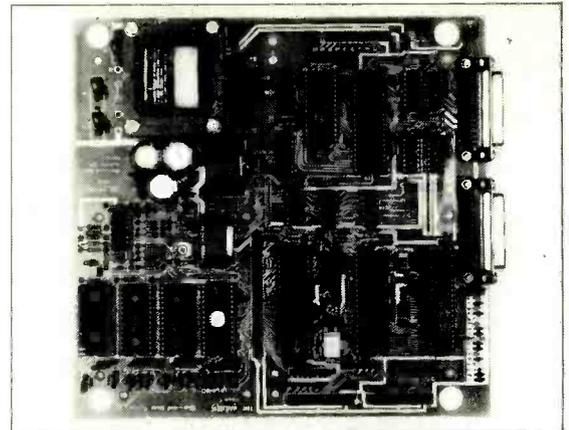
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ENTER 9 ON REPLY CARD

NEW PRODUCTS

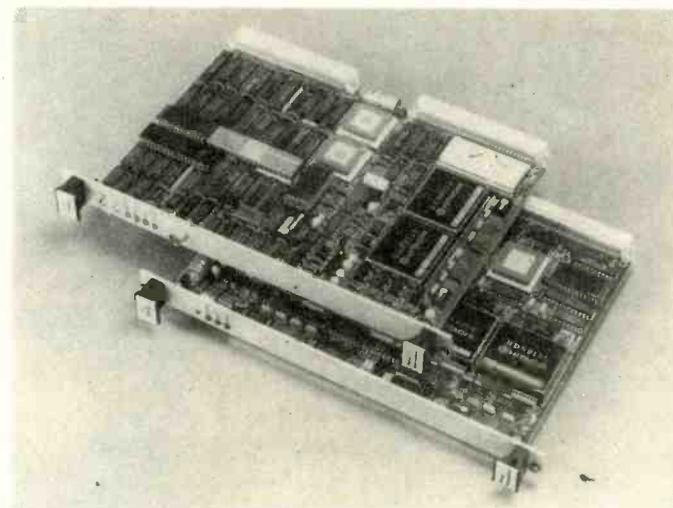
COMPUTING

High-speed eprom programmer

Newly launched is the G-Stack an addition to the Elan Universe 1000 Universal Programmer range. It incorporates 32-pin zif sockets, enables up to eight devices to be programmed concurrently either in sets or in gang (all identical) arrangements. The set mode for a single G-Stack allows expandable data set sizes of two to eight concurrent devices from 8 bit through 16, 24, 32, 48 to 64 bits wide. For higher production requirements, two G-Stacks may be linked together in a Universe 1000 to program up to 16 devices at a time in either sets or gangs. This would allow one set of say eight eproms to be programmed whilst another set is being loaded/unloaded and labelled. Label printing output to a printer is included. The control of the Universe 1000 operation is carried out by its Master Control Unit (m.u.c.) which may be connected by the system's parallel bus to any selection of "stacks" including those needed to program pals, IFLs, bipolar proms etc. Only the top stack is fixed; the lower stacks are all motorized with a simple reliable mechanism to move forward or retract under m.c.u. control so that the operator is clearly directed to use the correct stack and socket for the device selection. G-Stack can handle 28-pin devices up to the latest 1MB Intel 27011 EPROMS as well as the 32 pin 1MB versions from Intel, Fujitsu and Hitachi. New algorithms may be easily added by cartridge replacements. 206 on reply card.

68020 emulator

An in-circuit emulator for Motorola's 32-bit processor has been produced by Applied Microsystems, the first to do so apart from Motorola themselves. The EP68020 module allows full emulation at up to 12.5MHz, and is used in conjunction with an ES 1800 stand-alone emulator. Together they provide real-time analysis of all bus and control signals, with full-width address and data buses, by using trace modes. The trace memory allows the capture of 2046 words of 72-bit processor cycles to reveal what is actually happening. The patented events monitoring systems enables the user to control the emulation by setting the break vector on any combination of address, data, status, pass counter and logic state. An event or combination of events can be defined by logic statements and then used to break emulation, trace software sequences, count events or trigger outputs. 232 on reply card.



Bubble memory for VME bus

An intelligent bubble memory has been developed by Plessey for the VMEbus. The memory is totally non-volatile, without the need for battery back-up. It has no moving parts, is insensitive to adverse environments, and provides a high level of reliability. This results in a longer operating life and completely maintenance-free operation. The system comprises a master card (PME BB-1M) containing 1Mbyte of memory plus control logic, which may be configured with up to 8 Slave cards (PME BB-1S), each providing an additional 2Mbytes. In total, the

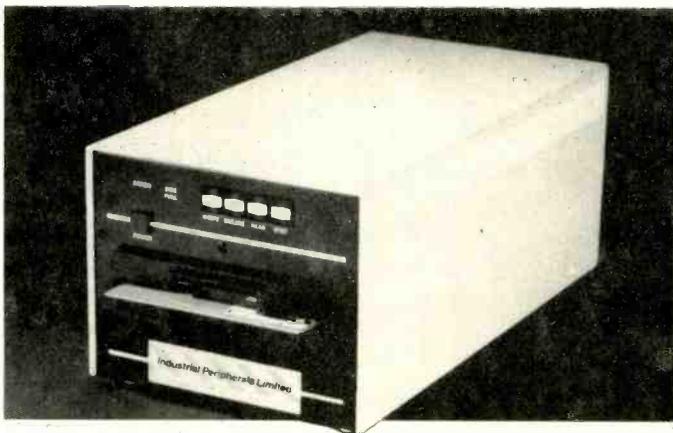
sub-system offers up to 17Mbyte of storage. BB-1 implements standard SCSI-type commands and supports VME transfers at up to 2.6Mbytes/s on read cycle and 3.6Mbyte/s on write operations. Bubble access takes, on average, 11ms; 8 and 16-bit transfers are supported and the memory base address is user-selectable within a 4Gbyte address space. Applications include process control and automation, offshore equipment, mobile systems and embedded applications. 211 on reply card.

Disc drive for IBM compatibility

An intelligent disc drive has been developed which will allow data loggers, instruments and any other device with RS232c interface, including other computers, to generate discs compatible with the IBM-PC. The T14, as it is called, receives data on one of its RS232 ports and stores it on a PC-DOS disc. The disc may then be used to transfer the data to an IBM-PC. Disks written by the PC can be read by the device connected. The internal software enables the T14 to operate in 'remote' mode; commands can be sent from the host device to give a

directory, write files, erase files, read files, rename files as well as controlling the unit. In addition to converting data into IBM format, the unit can be used as a complete disc operating system for devices without one and can be used, for example as a printer replacement for data loggers and remote test and instrumentation sites.

Other models in the range include a PC-AT compatible version, twin discs and an intelligent tape drive with 3.8 or 15Mbytes of storage. Industrial Peripherals Ltd 215 on reply card.



8-bit one-chip Microcomputer

Expanding its range of 8048 and 8049 single chip microprocessors, Toshiba introduces a 4Kbyte rom c-mos 80C50A device in a 40 pin dil or 44 pin flat package.

The 8 bit TMP 80C50A is made in Silicon gate c-mos which enables low power consumption during normal operations of 10mA at its fixed supply of +5V and frequency of 6MHz. A stand-by mode reduces current consumption 10µA maximum. On the chip is an 8bit c.p.u., 256byte ram data memory, 4Kbyte rom program memory, 27 I/O lines and an 8 bit timer/event counter. Another version, the TMP 90C40A may be supplied without the rom program on chip. By using this chip with an external eprom or ram software debugging is made easy. Instruction cycle time for both devices is 1.36µs for temperatures between 0 and 70°C. Software is upwardly compatible with Toshiba's TMP80C49AP/TMP80C49AP-6 or Intel's 8049. 213 on reply card.

Hard drive with s.c.s.i. interface

Including an s.c.s.i. controller in the circuit of a hard disc drive simplifies interfacing and reduces space requirements. The Mini Scribe 8425S 3.5in drive includes a standard s.c.s.i. controller but drives with a custom interface are also available.

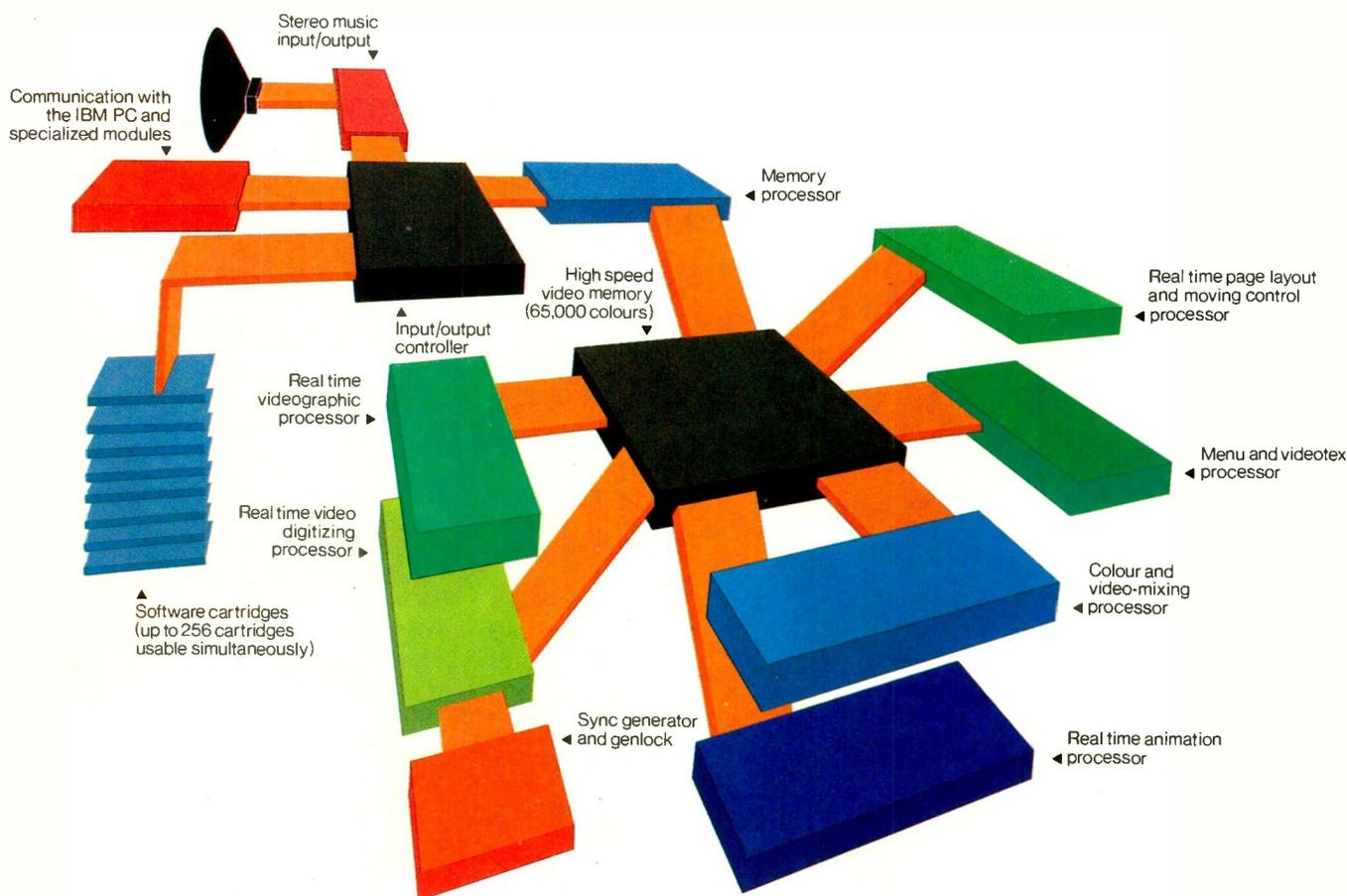
When formatted, data capacity of the half-height 8425 is 21.3M-byte. A single p.c.b. with surface-mounted components is used so the drive consumes only 12.5W. Average access time is 68ms and data is transferred to the host at up to 1M-byte/s. 223 on reply card.

68000 emulators

Plug-in pods for the emulation of 68000/68010 and 68008 processors are available for the Position SDT (symbolic debugging tool). The pods offer full emulation at the maximum rated clock speed (16MHz). The STD consists of a 8/16-bit universal control unit built around a Z80B processor with its associated memory. Additional ram boards fit inside the controller for real-time trace overlay operations. The emulation is performed by a second dedicated processor, housed at the target-board end of the pod cable. The system can be used with most development systems, many computers and terminals and Position's own Technical Development Computer. 208 on reply card.

NEW PRODUCTS

COMPUTING



Video workstation

This compact all-in-one video workstation brings to small studios a package of image creation and manipulation tools which until now have been restricted to much more lavishly-equipped production centres.

Once digitized by the system, an image can be retouched using textures and colours selected from within itself; it can be mirrored,

deformed, dissolved, multiplied, or turned into a mosaic. Moving pictures can be captioned, colourized artificially, inlaid and manipulated with a wipe effects generator, all in real time. Up to four video sources can be handled; Pal, Secam or NTSC, with automatic genlock on each. The makers plan to offer still more effects later by means of plug-in cartridges.

Besides video effects, the unit also

has a facility for page layout and for creating and editing business graphics: with the addition of an extra module, results can be turned into a photographic slide with a resolution of 4,000 by 4,000 pixels.

The system is based on seven dedicated processors, with an open architecture which allows the user to add further applications modules or to control it from existing equipment

— for example, an IBM PC.

The De Grafe Silver Workstation is made in France and distributed in the UK by Bell & Howell (telephone 01-902 8812). Prices for a complete system are in the range of £10,000-£15,000, which includes all updates or new versions of the integrated software for five years.

225 on reply card.

Pcb design on a PC (including Amstrad)

The major new features of the Quickpad 1.1 p.c.b. design package from Conguin Software, which runs on an IBM PC, include Block Move, Block Copy and Block Save commands. It has 126 pad and track sizes, ranging from 0.002 to 0.254in, a maximum board size of 32in. square and mass change commands for track and pad sizes. Other improvements over its predecessor are faster screen refresh times and the ability to generate silk screen images directly. Camera-ready artwork can be produced at up to eight times full size on a dot-matrix printer and a wide range of pen plotters. The system can be used on an IBM XT or AT computer and a version is being developed to run on the GEM-based Amstrad 1512 PC. The company are offering complete

low-cost workstations centred around the Amstrad PC. An IBM evaluation disc offers the full facilities but will not plot the result (Conguin will produce one free double-sided plot as a gesture of good will). This comes with the full user manual and costs £36.50 while the complete package, including a mouse, comes to £995. 207 on reply card.

Graphics controllers

A family of high performance graphic display controllers is manufactured by Control Systems Inc. and distributed by KGB of Windsor. Designed to work with IBM AT and high resolution monitors the 'Artist' board is expected to become a standard graphic enhancement for the operation of computer design software, such as AutoCAD, on the IBM micro.

The Artist series comes in a range of six models depending on the application and supports 42 software products, including all the major graphic and c.a.d. vendors using personal computers, including Autodesk with AutoCAD, Calcomp with Cadvance, and Computervision with Personal Designer.

The Artist range starts with the Monochrome 1024 by 768 Resolution Board at £695, through to the Artist 10 1024 by 1024 Resolution in 256 colours at £2095. 218 on reply card.

8-bit video-speed a-to-d

A bipolar monolithic, video-speed 8-bit flash a-to-d converter comes from Datel, the ADX-300. It is capable of digitizing an analogue signal at conversion rates up to

20MHz, yet the device's maximum power dissipation is 700mW. A serial/parallel technique is employed to obtain high conversion speeds using a single -5V power supply. The analogue input range is 0 to -2V, with -2V ref, and digital inputs and outputs are e.c.l. compatible. The outputs are buffered and provide an open emitter. The ADC-300 is designed for use with an external sample-and-hold amplifier together with external clock and reference source. The device features maximum non-linearity of $\pm 1/2$ l.s.b. and maximum differential non-linearity of $\pm 1/2$ l.s.b. The ADC-300 is ideally suited to applications requiring a combination of high-speed digitization and low power including high-speed data acquisition, radar pulse analysis and optical character recognition.

Datel UK, Tel: 0256 573675. 216 on reply card.



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ENTER 49 ON REPLY CARD

RADIO BROADCAST

Spotty outlook

There seems little doubt that we are now in the early months of Sunspot Cycle 22, the transition from Cycle 21 having occurred in the summer of 1986. There are already signs of a significant rise in the maximum usable frequencies on h.f. with a tearing up of the ionospheric predictions that forecast the continuation of Cycle 21 into 1987 or even 1988.

This promise of higher frequencies being usable over longer daily periods will come as a major relief to all users of the extremely crowded h.f. spectrum for communications or broadcasting.

For those concerned with h.f. broadcasting, the arrival of Cycle 22 has come at a particularly opportune time, just ahead of the important second session of the World Administrative Radio Conference for the planning of the h.f. bands allocated to the broadcasting service and due to be held in Geneva from January 27 to March 13, 1987.

The very low evening and night-time usable broadcast bands, including the 3.9 and 6 MHz bands, have exacerbated the problems brought about by the ever increasing powers and schedules of external broadcasting, not only those attempting to provide virtually 24-hour world coverage by using ever more relay bases capable of providing strong "one-hop" signals to audiences that tend to be loosely defined (the BBC has recently upped its claimed "regular" audience from 100 to 120 million, but the basis for these figures remains rather obscure and must include a substantial contribution from its domestic and overseas m.f. outlets).

The WARC may or may not come to grips with the pressing problem of jamming, much of it stemming from the network of Soviet jammers intended to blot out foreign Russian-language broadcasts but inevitably affecting h.f. broadcasting in all parts of the globe. If this jamming could be dramatically reduced, the use of multiple frequencies serving the same target area could be minimized, and some sort of order restored to the broadcasting bands.

The difficulty of achieving order out of chaos is underlined

by the current BBC use of 87 high-power h.f. transmitters at 10 sites throughout the world, with a further three sites under construction.

The equipment industry is clearly hoping for a phased introduction of compatible single-sideband transmissions with partially reduced carrier. Engineers from Marconi and Brown, Boveri presented papers on this subject at IBC86, based on the use of transmitters rated up to about 100kW peak envelope power. The Marconi work is based on the technique of 'envelope elimination and restoration' (e.e.r.) proposed by Leonard Kahn in the 1950s for m.f. and l.f. broadcasting, with possible enhancement by the polar-loop feedback technique developed at Bath University by Gosling and Petrovic.

Rejected in the 1960s, it would seem that compatible single-sideband is a technique whose time is coming – both for broadcasting and for mobile radio communications.

Noise-induced hearing problems

To what extent does listening on the once-again-popular headsets pose a hazard to the user? This question has come up recently in connection with amateur radio, where morse enthusiasts have always largely remained faithful to listening on 'cans'. But nowadays the pavements, trains and buses are witness to the enormous number of people with walk-about audio who listen at a level of volume that produces audible chatter at distances of many feet. Then, again, many people with hearing problems use headphones to follow television sound.

Nigel Neame, formerly G2AUB, is one of several correspondents, who have warned of the need for those with normal and partial hearing to take care not to listen on headphones at excessive volume, stressing that some degree of automatic peak limiting is highly advisable to prevent the impairment or even destruction of the remaining hearing ability. Reg Taylor, G3AVQ, has similarly drawn attention to the insidious prob-

lem of tinnitus (ringing in the ears) that can result from excessive noise yet may not manifest itself until years after exposure. As a headphone addict, I have always used a pair of back-to-back diodes across my ancient, high-impedance headphones to eliminate loud switching and similar clicks, when operating morse.

In 1984, the BBC Designs Department introduced headphone protectors (type EP5/25A stereo and EP5/25B) that similarly protect the wearer against potentially harmful sound levels when using high-quality, low-impedance headphones (8 to 600-ohms) such as Pioneer SE550, Beyer DT 220 etc. These were, apparently, rather more sophisticated passive limiters including averaging and weighting networks to prevent limiting-type distortion on short-term peaks and low-frequency signals, but including fast-acting voltage clippers to limit all signals well below a level at which instantaneous hearing loss could occur (although possibly, one must assume, not entirely ruling out later development of tinnitus). Sound limiting level can be set within the range 95 to 110 dBA and the protectors were intended for use in broadcast production. One gathers, however they are not widely used, either in the corporation or elsewhere.

Brian Davies, G30YU, was born profoundly deaf but by means of a series of operations recovered a good deal of hearing and has had a lifelong interest, including at times a professional interest, in high-fidelity sound reproduction. He is convinced that there is a good deal of emotional feeling against high-powered music environments, such as discos, that cannot be substantiated. He believes that, for the most people with good hearing, tinnitus is usually a temporary effect which may disappear overnight. The loud noises that, he agrees, do induce tinnitus or hearing impairment are impulsive sounds with steep wavefronts. These can be induced in the music output from audio amplifiers by using passive back-to-back diode limiters; he advocates instead the use of attenuator-type i.c. devices which can reduce volume rapidly but without distorting the wave-

fronts. Even better, he believes, is to use a low-distortion amplifier with plenty of headroom.

He notes that a paper by J.J. Knight, a senior ear, nose and throat consultant, reported an investigation of the hearing of a number of recording studio engineers who had worked for long periods in control rooms with very high sound-pressure levels. Out of 20 such engineers, only one showed signs of clinical deafness, and this seemed likely to have been age-induced deafness. This again suggests that while distorted or impulsive loud sounds can and do cause deafness, or tinnitus, undistorted sound is unlikely to do.

Varying the dynamic range

It is well recognized that achieving a universally acceptable balance between music and speech, programmes and adverts is virtually impossible, due to the large age and environmental differences of the listeners. Various systems have been proposed during the past decade or two for providing control signals that automatically adjust the dynamic range of the receiver output to suit the environment.

The latest process, this time demonstrated and tested, comes from the German Institut fuer Rundfunktechnik (IRT) and is described in *EBU Review – Technical* (August 1986, pp. 230-240).

This allows the dynamic range of a programme signal, whether speech or music, to be varied to match it to various radio, television or satellite circuits and, more especially, to match it to quiet or noisy surroundings, high or low listening levels, use of headphones or loudspeakers, etc.

An inaudible control signal is inserted in the stereo audio signal. This, according to the listener's choice, either matches the output of the receiver to the dynamic range of the transmission (30dB), or allows the listener to set the range to his choice, over the range 20 to 55dB, or after home taping.

In practice the system serves to enhance the signal-to-noise ratio of the channel by a maximum of 25dB.

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Anodyne history

The many events – and programmes – that almost swamped the 50th anniversary of the start of television broadcasting from Alexandra Palace tended at times to reduce the development of television to a personal struggle between Baird and Shoenberg and between 'mechanical' and 'electronic' systems. There was little recognition of the many contributions, some of fundamental importance, that stemmed from Germany and the USA, including the vital requirement to move away from m.f. to v.h.f., or that much of the Baird 240-line system (for which the technical director was not J.L.B. but Captain West) was derived from German work on 180 lines. At least the IEE's three-day conference (which will be over before these comments appear) promises us a number of international papers that will help to put the record straight, at least for the delegates, though the public will inevitably, retain the simplistic and rather chauvinistic myths that were sedulously propagated by many commentators.

Of lasting interest and value, however, are the two new television galleries at the National Museum of Photography, Film and Television at Bradford which had their gala opening and preview on 31 October. The upper gallery, *The Story of British Television*, has been sponsored mainly by the BBC and Thorn-EMI; a lower gallery, *Television Behind the Screen*, is sponsored by Yorkshire Television, Philips, Lee Cortran, Central ITV, LWT and Microvitec including an initial donation of £100,000 by Yorkshire Television. This gallery departs from normal museum display by combining broadcast technology with audio-visual exhibits. It shows how programmes are made, taking the risk of rapid dating. All programme material is stored on Philips Laservision discs, considered the only recording system rugged enough to withstand constant playing!

Very well worth a visit – though it was unkind to show at the gala opening a stereoscopic film on the enormous (52ft 4in high by 64ft 8in wide) Imax

screen. Someone was obviously keen to make the point that even with modern, high-definition tv, there is still a long, long way to go to catch up with this sort of film display!

Somehow, the three-day "International Conference on the History of Television – from Early Days to the Present" failed to spark the nostalgic vein to the same degree as the 1985 radar conference. But perhaps this is a jaundiced view of a reporter who had hoped for surprising new revelations, but found most presenters sticking firmly to the world of engineering which, though important, is only part of television, often largely ignored by those who have always shaped broadcasting policy. While Alan Blumlein was amply confirmed as the engineering star, little was said of the political background not only to the launch of Alexandra Palace tv in 1936 but also to the German services on 180 lines and then 441 lines under pressure from the Nazi party as a propaganda coup, though with no receivers available to the public.

But then, few of those who spoke were in the original Baird or Marconi-EMI camps, with the notable exception of Tony Bridgewater, Joshua Sieger (Scophony) and H.G. Lubzyski. If all the speakers had been as lively and as forthright as the Berlin-born, one-time employee of Siemens and Telefunken, before joining (Sir) Isaac Shoenberg's remarkable team, then indeed it would have been a conference to remember!

Lubzyski was one of the few speakers who attempted to correct some of the myths that surround early developments. He firmly ascribes the invention of the vitally important storage principle in camera tubes not to Campbell Swinton (as often suggested) but to Zworykin. He also confirmed that the 240-line "Baird" system was cobbled together by Captain West, including the purchase of the high-speed scanning disc and intermediate-film system from Germany.

Dr Maurice was brave enough to maintain that the UK should have stuck to 405-lines rather than change to 625. Pat Leggatt told Peter Mothersole that it was "nonsense" to suggest the pub-

lic were not interested in the better black-and-white pictures that would have been obtained with d.c. restoration rather than a.c. coupling. Otherwise few contentious issues were raised – few myths overturned in an all-too-genial atmosphere of mutual backslapping.

Only H.G. Lubzyski ended on a sad note: "We never thought in 1933 that our efforts would be abused to pump out crime, violence and murder every night to an audience of millions. Instead of increasing the broadcast hours, I believe they should be reduced and the quality of the contents be improved." Whether you agree or not, it was a change from the frequent harping on British television being "the best in the world".

As Professor R.W. Burns pointed out in his closing remarks, few of the firms that pioneered television have remained fully committed to this field. He regretted that nobody spoke on the French work or represented the original Pye company. With 128 registrations, Savoy Place was far from full, with few younger engineers seemingly interested in what Henry Ford is supposed to have regarded as "bunk", but is in fact all our yesterdays, and from which lessons can still be learned.

"Live-Net" grows

It is about 20 years since large educational closed-circuit television cable networks were set up in Glasgow and London, only to disappear later when they became enmeshed in local politics. However a significant revival of interest seems to have arisen within the widely-spread colleges of London University. A specialized modern cable network is now emerging in collaboration with British Telecom.

The original requirement was based on the desire to relay medical lectures and seminars from the new Charing Cross teaching hospital to its seven satellite hospitals south of the Thames, to reduce travel costs and the time wasted in students travelling across London.

The resulting two-way in-

teractive system has encouraged the university to plan optical-fibre links between the Senate House complex and Kings College, Queen Mary College, Imperial College, Bedford College and the university's audio-visual and computer centres. Each link provides four video circuits and data transmission facilities.

Installation of all these links should be completed next summer before the start of the new academic year.

Tv at 2Mbit/s

For several years, British Telecom has been offering video-conferencing facilities, with a data rate of 2Mbit/s, based on a GEC-McMichael codec. This bit-reduction system was also offered for satellite links but was more suited to talking-heads than for pictures with fast-moving content.

It has recently been announced that Siemens AG in Munich have developed a rather 2Mbit/s similar system, based on the mathematical technique of discrete cosine transformation (d.c.t.) and is studying further bit reduction to 384 Kbit/s and eventually 64 Kbit/s.

For d.c.t., the tv picture is first divided into blocks of, for example, 16 by 16 pixels to which are assigned coefficients that describe the picture content of that block in terms of a real number identifying a signal frequency. Bit reduction of the coefficients is held to be more effective than working on the individual pixel intensities. The signal is processed to eliminate the need to retransmit information on stationary images. Where there are abrupt movements only the most significant variations are transmitted immediately, with minor variations and details sent in subsequent frames. Viewers see a slight, temporary loss of resolution during abrupt movement.

It is claimed that Siemens has gone beyond existing d.c.t. and differential-p.c.m. systems in the detection of groups of coefficients in the transform domain, adaptive Huffman coding and postbuffer control. A 2Mbit/s, the Siemens techniques are claimed to provide excellent tv images, even when picture details are changing rapidly.



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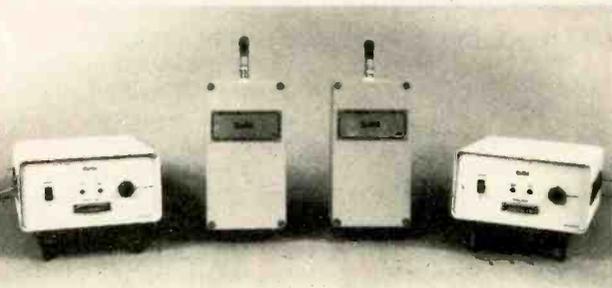
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RADIO COMMUNICATIONS

Mobiles in 5kHz

During the past decade there have been a series of research projects – including the Wolfson project – aimed at achieving practical mobile-radio systems that would permit 5kHz channelling without significant degradation of the performance achieved by current f.m. or a.m. systems using 12.5, 15, 25 or 30 kHz channelling. It has long been recognized that, if analogue transmission is retained, this requires some form of single-sideband transmission, either with pilot or full carrier, to overcome Doppler frequency shifts, and with an effective a.g.c. system in receivers capable of minimizing the deep fading caused by the severe multipath conditions experienced on moving vehicles.

In the UK, Professor J.P. McGeehan at Bristol University continues to advocate feed-forward with pilot-tone s.s.b. for mobile operation. In the USA amplitude companded single sideband (a.c.s.s.b.), in which the signal is heavily compressed during transmission, has been FCC approved for mobile communications, but the systems so far marketed are significantly more costly than n.b.f.m.

Japanese engineers of NTT have proposed the use of real-zero single sideband (r.z.s.s.b.) with full carrier. This system (*IEEE Trans. on Vehicular Technology* vol VT-35, No 1, February, 1986) can provide compatibility with f.m. receivers and can use 5kHz channelling, yet incorporates an f.m.-type amplitude limiter in the receiver. This, it is claimed, can largely overcome the amplitude degradation caused by fading. This system, like compatible h.f. s.s.b. broadcasting, can draw on the techniques proposed in 1952 by Leonard Kahn, enhanced by the use of the high-efficiency polar-loop transmitter configuration

developed by Petrovic and Gosling at Bath University.

The r.z.s.s.b. system offers the attraction that there is no requirement to regenerate in the receiver the clean and stable carrier necessary for product detection. In an experimental system the Japanese engineers used phasing-type s.s.b. generation, with full-carrier lower sideband signals developed at 455 kHz and up-converted to 70 MHz.

Reverse Danegeld

Amid all the fervid commemoration of 50 years of television, members of the Special Forces Club in Knightsbridge met to receive from the Danish firm of Bang & Olufsen a modern tv set with v.c.r. in connection with events now more than 40 years ago.

For from 1941-1945, B & O and, in particular, their chief engineer, the late L.A. Duus Hansen (OZ7DU), his surviving widow, and his former secretary Edith Bonnesen, were at the very centre of possibly the most successful of all the wartime clandestine radio networks set up in enemy-occupied countries.

Duus Hansen, born in 1901, while still nominally working for B & O pursued single-mindedly his efforts to provide manual and high-speed machine telegraphy and v.h.f. telephony links with the UK and/or Allied representatives in Sweden, working first through British and Danish ("The Princes") Intelligence but later through SOE which, unusually, became responsible for obtaining intelligence from that country, as well as organizing sabotage. The German security police reacted by destroying not only his home but also the B & O factory where, with the help of Svend Bagge and Steen Hasselbach, some 60 of his ingenious

"telephone-directory" (Telefonbogen) lightweight h.f. transmitter-receivers had been secretly built, using standard domestic-receiver components that came largely from Germany. Much of Denmark was then using d.c. mains unsuitable for the British-built SIS/SOE "suitcase" sets which also had to be relatively heavily built to withstand being dropped in parachute containers. Duus Hansen rejected such equipments and the hastily-trained operators sent from the UK. He recruited or trained his own 'amateur' operators. His seven-valve Telefonbogen unit, the size of the Copenhagen telephone-directory, had an output of 10 watts from parallel UBL21 valves and a three-valve superhet receiver using three triode-heptode UCH21 valves. The "transformerless" a.c./d.c. series-heater techniques kept the weight to under 1.5kg. He persuaded SOE to send out crystals and signal plans. His enthusiastic and dedicated group was one of the very few to use a high-speed auto-sender (GNT) on the short-range links with Sweden.

The enormous professionalism shown in Duus Hansen's approach to covert radio had one important outcome not mentioned at the recent presentation: unlike the tragic experience in many other occupied countries where the vulnerable radio-operators had an average operational life of only about six weeks before being overtaken by capture or death or being "played-back" by the enemy, the Danish radio group had an outstanding record of survival.

Ole Lippmann, one of the leading members of the Danish SOE group, and (Colonel) J.D. Parker recalled the events in which Duus Hansen and B & O were involved. (Major) John Brown and (Flt Lt) Charles Bovill of SOE Signals were there, as was the remarkable Mrs Yvonne Cormeau, MBE, who, as a young WAAF radio operator, was parachuted into France in August 1943. As "Annette", working for the Section F organizer, George Starr, in the following twelve months she transmitted and received more than 4000 messages before the village where she had set up her station was overrun by Allied forces, so becoming one of

the most successful radio-agents to survive in France. And this was despite breaking the convert-radio rule book in operating from the same house for about six months. Usually, in practice, the greatest danger was enemy penetration of the Resistance movements or "amateur" informers rather than the German d/f teams.

Among SOE and SIS veterans one finds a growing irritation with the books being produced by young military historians and journalists whose research may be diligent but often fails to recapture the wartime atmosphere in which Resistance developed and blossomed despite its many tragic shortcomings.

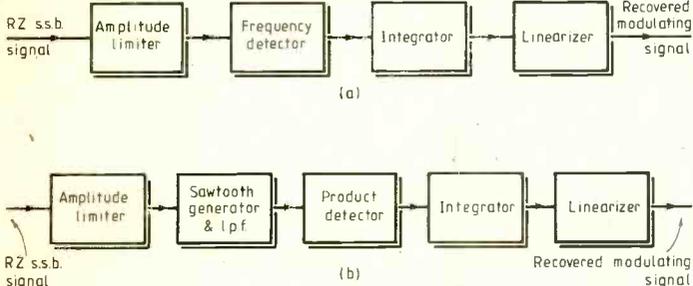
Space logic

Dr Karl Meinzer, DJ4ZC, one of the leading members of AMSAT-DL, interviewed for the New Zealand amateur radio society, NZART, has expressed some forthright views on future satellites for the amateur service.

He shows little support for the use of packet radio transmissions, which he feels may be a craze that will pass like other crazes, becoming just one activity in amateur radio, at a relatively low level. He notes that the general trend in computing, at least in Europe, is starting to wane. This, he feels, is because people have found that doing really useful, complex things with computers is still hard work. He advocates linear transponders on the satellites, since they place few constraints on what you may wish to do.

AMSAT-DL is also not in favour of attempting to have a geostationary satellite, because the project cost would be an order of magnitude greater than present amateur satellites and require much more mechanically complex satellites with despun antennas and attitude-control jets.

Dr Meinzer also believes that geo-orbital satellites would diminish the educational value of amateur satellites since they would remove the need for tracking and knowledge of the space environment. A geostationary satellite would be, he suggests, no different in essence from an f.m. repeater or telephone.





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JA96	0.65	EC980	1.15		
XC70	1.75	EC982	1.15		
XC90	1.20	EC986	1.70		
XC94-5000		EC9200	1.85		
DET16	28.50	EC9202	1.85		
DET18	28.50	EC9201	0.85		
DET22	35.00	EC9205	1.85		
DET23	35.00	EC9210	1.85		
DET24	39.00	EC9215	1.85		
DET25	22.00	EC9220	1.85		
DET29	32.00	EC9225	1.85		
F91	1.00	EC981	1.00		
F92	0.60	EC983	1.00		
F96	0.65	EC984	1.00		
H77	0.90	EC985	1.00		
H79	0.56	EC986	1.00		
H149	2.00	EC987	1.00		
K92	1.20	EC988	0.95		
LD35	2.50	EC989	0.95		
LD63	1.00	EC990	0.95		
LD70	2.50	EC991	0.95		
LD73	1.50	EC992	0.95		
LD92	0.95	EC993	0.95		
LD93	1.10	EC994	0.95		
LD94	2.50	EC995	0.95		
LD96	2.50	EC996	0.95		
LD97	1.50	EC997	0.95		
LD98	1.50	EC998	0.95		
LD99	1.50	EC999	0.95		
LD100	1.50	EC1000	0.95		
LD101	1.50	EC1001	0.95		
LD102	1.50	EC1002	0.95		
LD103	1.50	EC1003	0.95		
LD104	1.50	EC1004	0.95		
LD105	1.50	EC1005	0.95		
LD106	1.50	EC1006	0.95		
LD107	1.50	EC1007	0.95		
LD108	1.50	EC1008	0.95		
LD109	1.50	EC1009	0.95		
LD110	1.50	EC1010	0.95		
LD111	1.50	EC1011	0.95		
LD112	1.50	EC1012	0.95		
LD113	1.50	EC1013	0.95		
LD114	1.50	EC1014	0.95		
LD115	1.50	EC1015	0.95		
LD116	1.50	EC1016	0.95		
LD117	1.50	EC1017	0.95		
LD118	1.50	EC1018	0.95		
LD119	1.50	EC1019	0.95		
LD120	1.50	EC1020	0.95		
LD121	1.50	EC1021	0.95		
LD122	1.50	EC1022	0.95		
LD123	1.50	EC1023	0.95		
LD124	1.50	EC1024	0.95		
LD125	1.50	EC1025	0.95		
LD126	1.50	EC1026	0.95		
LD127	1.50	EC1027	0.95		
LD128	1.50	EC1028	0.95		
LD129	1.50	EC1029	0.95		
LD130	1.50	EC1030	0.95		
LD131	1.50	EC1031	0.95		
LD132	1.50	EC1032	0.95		
LD133	1.50	EC1033	0.95		
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LD144	1.50	EC1044	0.95		
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LD148	1.50	EC1048	0.95		
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LD150	1.50	EC1050	0.95		
LD151	1.50	EC1051	0.95		
LD152	1.50	EC1052	0.95		
LD153	1.50	EC1053	0.95		
LD154	1.50	EC1054	0.95		
LD155	1.50	EC1055	0.95		
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LD158	1.50	EC1058	0.95		
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LD165	1.50	EC1065	0.95		
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LD189	1.50	EC1089	0.95		
LD190	1.50	EC1090	0.95		
LD191	1.50	EC1091	0.95		
LD192	1.50	EC1092	0.95		
LD193	1.50	EC1093	0.95		
LD194	1.50	EC1094	0.95		
LD195	1.50	EC1095	0.95		
LD196	1.50	EC1096	0.95		
LD197	1.50	EC1097	0.95		
LD198	1.50	EC1098	0.95		
LD199	1.50	EC1099	0.95		
LD200	1.50	EC1100	0.95		

M8195	6.50	Q51203	4.15	VU39	1.50
M8196	5.50	Q51205	3.95	W21	4.50
M8204	5.50	Q51206	1.05	W77	5.00
M8223	4.50	Q51207	0.90	W29	1.00
M8224	2.00	Q51208	0.90	W39	1.50
M8225	3.95	Q51209	3.15	X24	4.50
ME1401	29.50	Q51210	1.50	X61X66	4.95
ME1402	29.50	Q51211	1.50	X76M	1.95
ME1501	14.00	Q51212	3.20	XG24	1.50
MH4	3.50	Q51213	5.00	XC25	0.50
MHL6	4.00	Q51215	2.10	XF747	1.50
ML4	4.50	Q51218	5.00	XF750	1.50
EL509	5.25	Q51219	9.50	XG1-2500	75.00
EL51	6.95	Q51220	1.50	XG2-6400	135.00
EL802	3.65	Q51221	3.20	W729	1.00
EL821	6.50	Q51222	5.00	W739	1.50
EL822	12.95	Q51223	2.10	X24	4.50
EM1	9.00	Q51224	5.00	X61X66	4.95
EM4	9.00	Q51225	15.00	X76M	1.95
EM80	0.70	Q51226	15.00	XG24	1.50
EM81	0.70	Q51227	15.00	XC25	0.50
EM84	1.65	Q51228	15.00	XF747	1.50
EM85	3.95	Q51229	15.00	XF750	1.50
EM87	2.50	Q51230	15.00	XG1-2500	75.00
EN32	15.00	Q51231	15.00	XG2-6400	135.00
EN91	1.95	Q51232	15.00	W729	1.00
EN92	4.50	Q51233	15.00	W739	1.50
EX35	235.00	Q51234	15.00	X24	4.50
EY1	0.80	Q51235	15.00	X61X66	4.95
EY81	2.35	Q51236	15.00	X76M	1.95
EY83	1.50	Q51237	15.00	XG24	1.50
EY84	5.50	Q51238	15.00	XC25	0.50
EY85/87	0.50	Q51239	15.00	XF747	1.50
EY88	0.55	Q51240	15.00	XF750	1.50
EY91	5.50	Q51241	15.00	XG1-2500	75.00
EY92	0.70	Q51242	15.00	XG2-6400	135.00
EY93	0.70	Q51243	15.00	W729	1.00
EY94	0.70	Q51244	15.00	W739	1.50
EY95	0.70	Q51245	15.00	X24	4.50
EY96	0.70	Q51246	15.00	X61X66	4.95
EY97	0.70	Q51247	15.00	X76M	1.95
EY98	0.70	Q51248	15.00	XG24	1.50
EY99	0.70	Q51249	15.00	XC25	0.50
EY00	0.70	Q51250	15.00	XF747	1.50
F606	2.95	Q51251	15.00	XF750	1.50
F5WA/800	2.95	Q51252	15.00	XG1-2500	75.00
G591K	9.00	Q51253	15.00	XG2-6400	135.00
G1802M	6.95	Q51254	15.00	W	

Mobile radio for Lancashire's police

Lancashire Constabulary's radio engineers pioneered many techniques now used in v.h.f./u.h.f. networks everywhere. This article traces their system's development.

J. DAVIES

In 1924, police in Lancashire used four-valve broadcast receivers for the reception of news and weather forecasts. But in 1925, a 200-watt transmitter was installed at the Constabulary headquarters at Preston and six divisional headquarters were equipped with four-valve receivers. A mobile set — which may have consisted of a receiver only — was fitted in a van for use in connection with a royal visit.

The year after, a Dennis 30cwt van was equipped with a transmitter capable of 200W output and which could transmit telephony. It was later discovered that a lower power would suffice and a 15W transmitter using Marconi type LS5 valves was installed.

Also in 1926, all police divisions in the country were equipped with three-valve receivers and seven divisional headquarters with 10 transmitters. During the general strike, temporary base stations were installed in the vicinity of collieries to assist in maintaining police strengths to protect safety workers against hostile crowds.

The original equipment was replaced in 1930 with, according to the records, a Hartley-type transmitter fitted with an Osram transmitting valve and a modulator. Crystal control of transmitter frequency was also introduced about this time.

Daily contact was made with Scotland Yard for important traffic but the system was abandoned in 1931 as a result of economy measures.

THE MOVE INTO VHF

In 1930, the Radio Branch, or Wireless Workshops as it was known until recently, was formed by the appointment, as a constable, of the late Frank Gee (G60G). Frank retired with the rank of superintendent. His responsibility was to investigate and, if feasible, establish a radio system to communicate with cars.

I believe that the enthusiasm of Frank Gee and his staff of the early years, particularly Keith Eve, who succeeded as chief engineer, should be given credit for a large amount of the pioneering work which laid the foundation of v.h.f. mobile radio communications on this side of the Atlantic.

After the fixed station network became inoperative in 1931, some experimental



1963: Lancashire's home-grown personal v.h.f. radio. Production models were made by GEC; some were sold even to the USSR.

work was done on medium frequencies with two-way communication with cars; but it was soon discontinued, the conclusion being that the frequencies were unsuitable for short-distance mobile work.

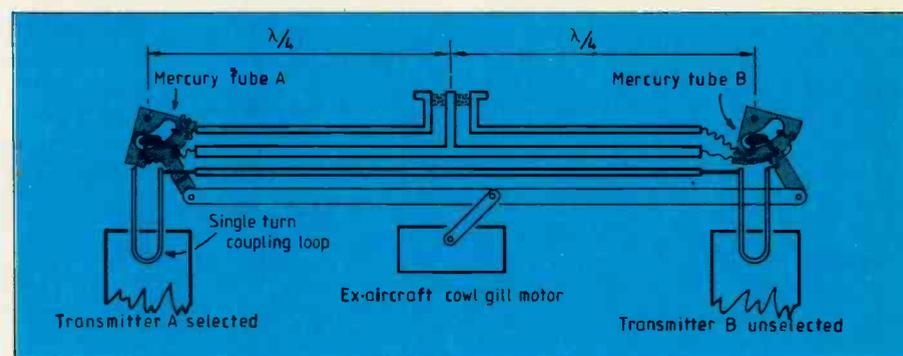
Later experiments were made on 60MHz over a small area of the county, under an experimental licence. They used a base station installed at force headquarters at Preston and a car worked within an eight mile radius. It was possible to use telephony with a loudspeaker in the car and the results were so promising that the work continued.

No records are available of the type of equipment used. Certainly it was home-made, none being available commercially, and staff of that time tell of de-capping valves to improve their performance. At the time, the view was generally held that very high frequencies were not suitable for this type of communication because of their limited range — which then was believed to be about 1½ miles, but the preliminary work disproved this.

With hindsight, it was probably the lack of suitable components which restricted the performance of the early equipment, giving rise to the belief that v.h.f. was unsuitable. It worries me today that technicians are so used to dealing with much higher frequencies that v.h.f. tends to be treated almost like audio. Consequently, sufficient attention is not always paid to such items as lengths of leads and the materials used.

The first v.h.f. scheme expanded until twelve cars were operating in the Fylde area

Fig.1. 1950: Lancon high-power antenna relay. The unit was actuated by an aircraft cowl motor, with contacts to prevent operation of the transmitter during switching.



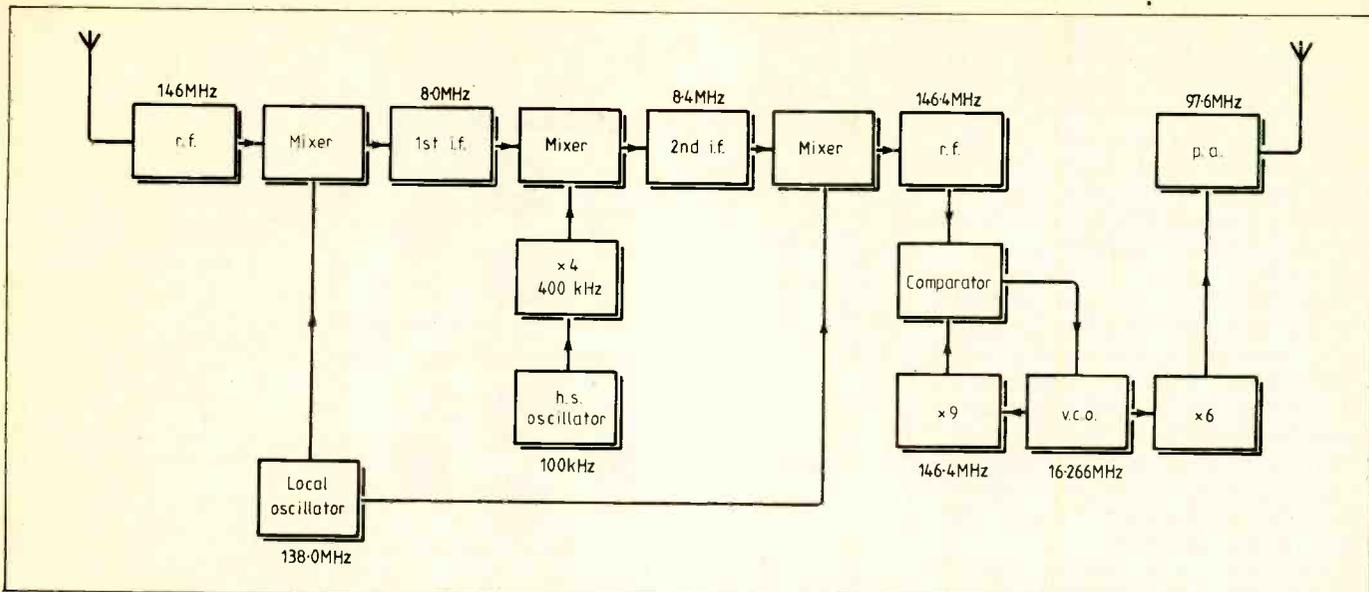


Fig.2. 1953: quasi-synchronous non-demodulating transmitter, for extending system coverage.

from a 100W transmitter at Preston. Additional fixed stations at Garstang and Swinton were introduced in 1936 and 1937. More cars were equipped to work in these areas and the earlier successes were repeated.

What seems to have been the first use of radio for aerial traffic control came in 1937, for the Grand National race meeting at Aintree.

By 1939, four new fixed stations were providing radio coverage over a large part of the county. Each transmitted on a different frequency but received on a common one, both frequencies being in the 70-80MHz band. The mobiles used National type 1-10 tunable receivers with home-built transmitters. The receivers were super-regenerative and used types 954 and 955 acorn valves.

As war became imminent, the divisional headquarters were equipped with transmitter-receivers operating in the 128-131MHz band as a back-up to the telephone system.

The wisdom of this was demonstrated when numerous telephone exchanges in the Manchester area were put out of action. During the prolonged bombing of Liverpool in 1941, Lancashire Constabulary's radio system, through the fixed station at Billinge near Wigan, was the sole means of communication between the city and the regional headquarters and all civil defence services relied upon it for a time.

In 1941, a committee was appointed by the Secretary of State to advise upon the future developments of radio for the police and fire services. To enable this committee to reach its conclusions, v.h.f. trials were carried out by Home Office engineers assisted by technical officers from Lancashire, in Buckinghamshire, Oxfordshire and Berkshire. The results were favourable and as a result the services were supplied with v.h.f. equipment on hire from the Home Office.

Approval was given for three forces, Lancashire, Liverpool City and Birmingham City, which had developed their own systems, to remain independent from the Home Office for a time but Liverpool and Birming-

ham ultimately joined the Home Office scheme.

The v.h.f. schemes used amplitude modulation and the equipment used was largely the type P40 receiver and S4350 transmitter produced by Eddystone Radio, now part of Marconi Communications Systems Ltd. Eddystone Radio also produced a separate a.m. transmitter from a Lancashire prototype.

FREQUENCY MODULATION

At the time of the Home Office trials, Lancashire technicians were considering the possible superiority of frequency modulation and began their own trials with it. As experience was gained, some base transmitters and P40 mobile receivers were converted to f.m.

At this stage, f.m. was used only in the base-to-mobile direction. The receivers were single superheterodynes which were converted to double superhets suitable for f.m. reception by the addition of a progressively limiting second i.f. of 1.2MHz, a Foster-Seeley discriminator and a squelch circuit. The original detector was made into a mixer circuit and the additional circuits were housed in a Mu-metal case mounted atop the receiver with the discriminator audio output and the squelch voltage fed back into the original audio stages.

The radio scheme at this time used four manned, hill-top base stations. But at first, only the one serving the north-west of the county was converted to f.m. in the outgoing direction. During the change-over period, both a.m. and f.m. were successfully applied to the one transmitter.

Commercially-produced f.m. mobile equipments were first delivered in 1948 and were supplied by a company called Electronic Transmission Equipment (E.T.E.) Ltd., which was a subsidiary of Mullard. By 1950, all main radio schemes in the country had been converted to f.m. and the only remaining a.m. equipment was that used for local portable networks.

With the success of f.m. a radio system controlled over point-to-point radio links

from a central location was envisaged. It was desirable to achieve coverage of a wide area by simultaneous transmission from a number of base stations on a single frequency. An initial step towards this was taken with the engagement of Mullard Ltd, who supplied three radio-link controlled, non-demodulating, synchronous f.m. base station transmitters.

Two of these were installed at the Billinge and Barnacre base stations and controlled from Billinge. They operated by dividing the 146MHz radio link frequency down to 100kHz and then multiplying the result to a final frequency of 98MHz.

Local oscillator injection in the receiver was obtained from an oscillator locked to the 100kHz oscillator which was itself locked to the receiver i.f. Another oscillator, also locked to 100kHz, generated the transmitted frequency. All locks were achieved by phase-locked loops.

No attempt was made to equalize the audio phasing in the transmitter, although the audio to the directly-controlled Billinge transmitter may have been delayed so that audio from both transmitters would be transmitted in phase.

REMOTE CONTROL OF BASE STATIONS

Work was also progressing on the development of the county-wide remotely-controlled system and by 1950 a base station at Winter Hill was nearing completion. Today this is a well-used site shared by many services; but then, even the road had to be made and this was done by policemen.

1950 was an eventful year. No only was the work of installing the remotely-controlled system under way, but the design and construction of several 600W base transmitters for the scheme was undertaken.

The remotely-controlled system at first used four base stations which were equipped with pairs of the 600W transmitters for main and standby purposes. In the transmitters were Mullard QY4/250 valves with blown-air cooling. The antenna switching relay —

again home made — was novel in that two concentric quarter-wavelength lines were used (Fig.1). Connection to the transmitter at each end of the line was made through a mercury tube. The tubes were coupled mechanically and the one connected to the unselected transmitter provided a short circuit across its section of the line and thus an element of filtering in addition to the switching function.

High power components were not available and such items as power amplifier tuning capacitors were made in the wireless workshops. Two of the base stations were equipped with the Mullard synchronous transmitters which drove the 600W amplifiers through in-house-built 45W stages. By 1952, the new radio system was operational from police headquarters at Hutton using high-band v.h.f. radio links.

Once the system was proved, a set of control and base station equipment to replace everything except the synchronous transmitters and the high power amplifiers was purchased from GEC. This equipment controlled three radio channels at four base station sites.

The Mullard transmitters were prototypes; no production models were made. They worked well, but servicing was something of a nightmare and in 1953 they were replaced and the area coverage scheme extended by quasi-synchronous, non-demodulating transmitters. These were designed and built by ourselves, largely from GEC components (Fig.2).

To the link receiver i.f. was added the output from a high stability oscillator, the frequency being the difference between the radio link frequency and $3/2$ times the transmitting frequency. The resulting second i.f. was again mixed with the receiver local oscillator output to produce a frequency exactly $3/2$ times the transmitting frequency. A voltage controlled oscillator, locked to this frequency by means of a phase-locked loop, produced a sub-harmonic of the transmitting frequency and carried the modulation.

The high stability oscillator could be adjusted over a small range to set the transmitters to a desired frequency offset. Experiments showed that an offset of approximately 10Hz between adjacent frequency transmitters was optimum on this system. In practice, the high stability oscillators drifted about ± 3 Hz over the period between re-adjustments. This gave offsets of up to about 70Hz between transmitters at the extremes of the frequency range which was acceptable. Offsets of less than 2Hz produced very undesirable effects in transmitter overlap areas because of the slowly moving interference patterns.

These transmitters were used until 1968 when the GEC system was replaced by a six-channel scheme supplied by Pye Telecommunications, which used quasi-synchronous, demodulating transmitters. Audio phase equalization was incorporated into these by the constabulary radio branch. At the same time, a fifth station in what is now Cumbria was commissioned to serve the Furness area and to provide a better signal in the Lune valley.

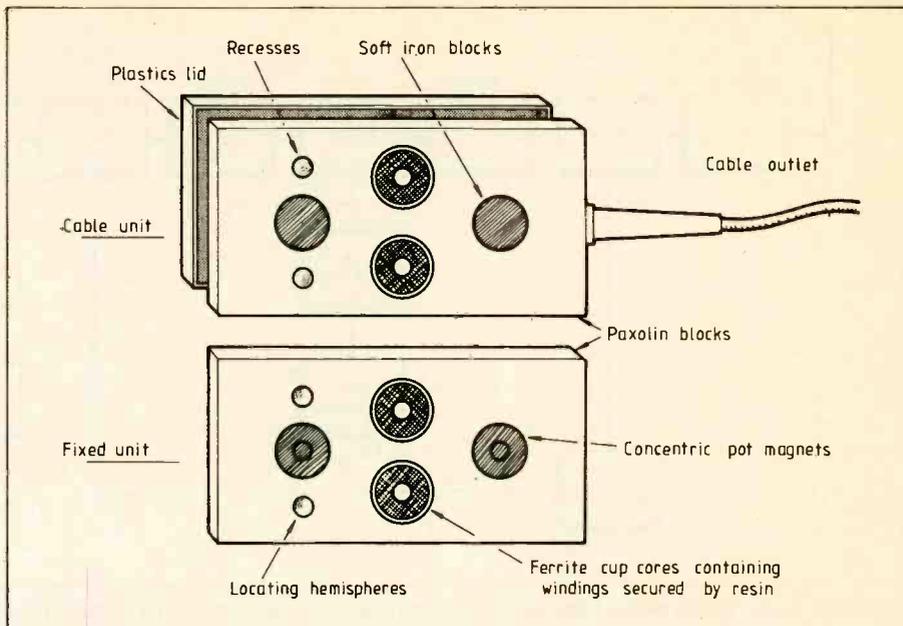


Fig.3. Magnetic connector for connecting body-worn microphone and earpiece to the motorcycle radio. Vibration will not dislodge it, but it will safely disconnect with a slight pull. The design is still in use.

Whilst all the base station work had been going on, there had been three generations of mobiles: Mullard type GME 501, Hudson types FM 105 and 106 and GEC RC660. However, apart from an experimental Plessey unit, no motorcycle equipment had been used.

MOTORCYCLE TRANSMITTER-RECEIVER

Development of a purpose-built motorcycle radio was begun in 1957. The requirements were that it should use, as far as possible, solid-state components and be safely usable whilst moving.

The main problems were the type of microphone and earphone to be used and the safe connection between man and machine; the control of transistor temperature; the design of transistorized circuitry for the crystal oscillator and the limiting i.f. amplifier; vibration from the motorcycle; and the effect of the radio equipment on the stability of the motorcycle at speed.

After trials with throat and cheek microphones, a small diaphragm-type microphone on a boom in front of the rider's mouth and a rubber-mounted earpiece were adopted on the early models. Later, a change was made to the modern practice of using acoustic coupling between the operator and the transducers.

The problem of connecting these body-worn components to the radio was overcome by the use of a magnetic connector (Fig.3). This design is still used today and commercially produced radio sets have been modified to accept it. It has an insertion loss of 1.5dB over the audio frequency range and will safely disconnect with a slight pull from any direction, yet is not dislodged by vibration. It is completely weatherproof and has no contacts to give trouble; but it is expensive to produce.

The only solid-state components available to us were germanium transistors with an

upper frequency limit of about 20MHz. Valves therefore had to be used for the r.f. circuits.

The transistors also had a temperature restriction of 55°C; and the presence of valves worsened the temperature problem, but this was overcome in the design of the case. The temperature inside an empty black box standing in the open air in sunlight, was found to rise to 48°C, leaving a safety margin of only 7°C for operation of the transistors. The heat from the valves could well raise the ambient temperature to above the limit.

A white glass fibre shield fitted over the box but spaced from it and with a slot in the upper surface to permit an air flow reduced the internal temperature under the same conditions to 36°C. The final design of the case and chassis segregated the valves and transistors with some heat insulation between them and a matt black box dissipated the internally generated heat into the space beneath the shield. No defects due to excessive temperature are known to have occurred during the life of the equipment.

Circuits for the series-resonant crystal oscillator and the progressively-limiting i.f. amplifier were straightforward. Nevertheless, provisional patents were issued for these, the housing and the magnetic connector.

An interesting point arose with regard to vibration. After the sets — about 100 were made to our prototype by GEC — had been in use for some time, the lead-out wires of a transformer in the power supply unit repeatedly fractured. The power supply was mounted in a recess in the petrol tank and it looked as if the trouble emanated from engine vibrations. This was confirmed when a stroboscope showed that with the engine running and the vehicle on its stand, the transformer windings were rotating back and forth around the core to the extent of the lead out-wires although the windings were absolutely immovable by hand. Potting the

transmitter in silicone rubber eliminated the problem.

PERSONAL RADIO

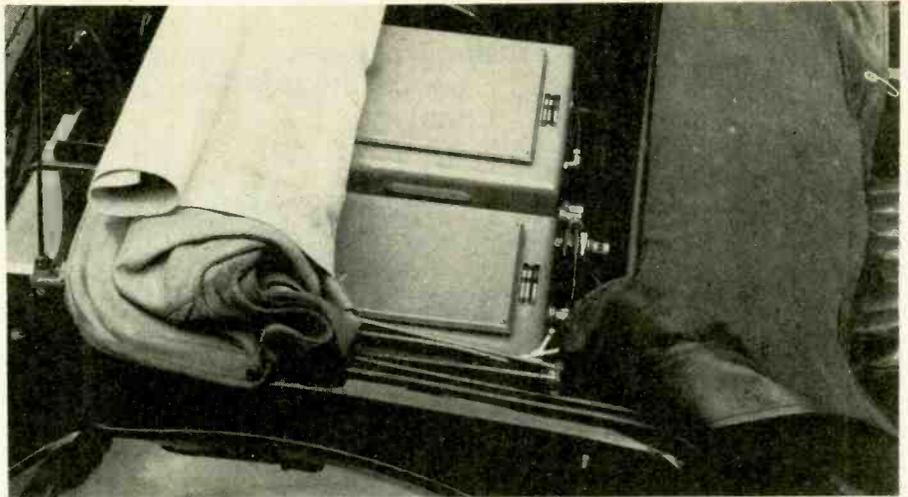
After a successful demonstration at Stretford in 1959, ten high-band v.h.f. personal radio sets were imported from Motorola in America. The receivers were all solid state and beautiful pieces of equipment; but the separate transmitter used miniature valves and was very expensive on batteries. A personal radio scheme for foot patrolmen using these sets was installed at Chorley in 1961; they worked well and undoubtedly proved the value of radio to a policeman.

Consequently, British manufacturers were approached for a similar type of equipment but the response was poor. Only existing hand-portables, too large for the purpose, were offered. The general opinion seemed to be that the potential market was too small to justify development. There was no alternative but for the Constabulary again to go it alone.

By 1963 it had produced prototypes of the Lancon v.h.f. personal transmitter-receiver. The set was intended to be worn on a belt with leads to a microphone, transmit switch and antenna and, separately, to an audio transducer. The circuitry was built on small printed boards which were accommodated in a plastic honeycomb, the complete assembly being housed in a plastic case.

Radio manufacturers were invited to tender for the production of the equipment and GEC were successful. They produced 800 sets for Lancashire plus a quantity for the Home Office and others were sold to Russia. It was a matter for pride that the sets entered production without a major modification.

The main problem in service was the failure of the leads where they entered the



1948: Lancashire police's first commercially-produced mobile f.m. set, shown here in an MG car. Equipment was made by E.T.E. Ltd, a subsidiary of Mullard.

case. This was overcome in Lancashire by redesigning the grommeting; but I shudder to think how the Russian sets fared at their temperatures. The Lancon was used for several years with an average fault rate of 4.5% per week until replaced by commercial u.h.f. equipment.

UHF AREA COVERAGE

Personal radio was first introduced into urban areas only but it was not long before coverage of wider areas was required. More than one base station was then needed to cover the service area. Initially, the base transmitters were selected manually although receiver voting had been introduced.

Now, the base station operator's dream is a free-running talk-through system without any manual switching; and the nearest prac-

tical approach is a quasi-synchronous transmission system with receiver voting. The manually-selected transmitter schemes had already been provided with interconnect facilities to the county-wide, headquarters-based v.h.f. system and worked started on the development of a u.h.f. quasi-synchronous system controlled over land-lines. A successful system was produced and is officially known as LANPACCS — Lancon personal radio area coverage and control system — but is often mis-pronounced or given other names by the technicians.

The system almost achieves synchronization. Where transmitters are geographically closely spaced with overlap areas, the output from a high stability oscillator at a central base station is divided down to approximately 3kHz and transmitted over a land line to all other base stations to provide a reference for their carriers. Modulation is transmitted to the base stations over a separate land line. Because two lines are needed for the quasi-synchronous stations, the more remote stations with little overlap are not synchronized, but use high-stability oscillators directly for their carrier frequencies.

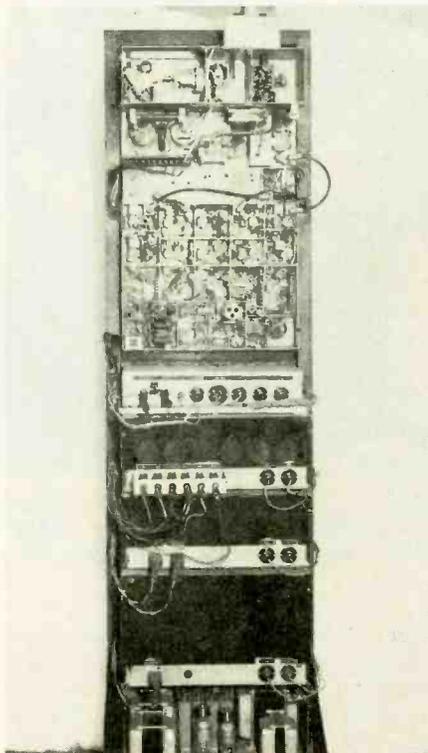
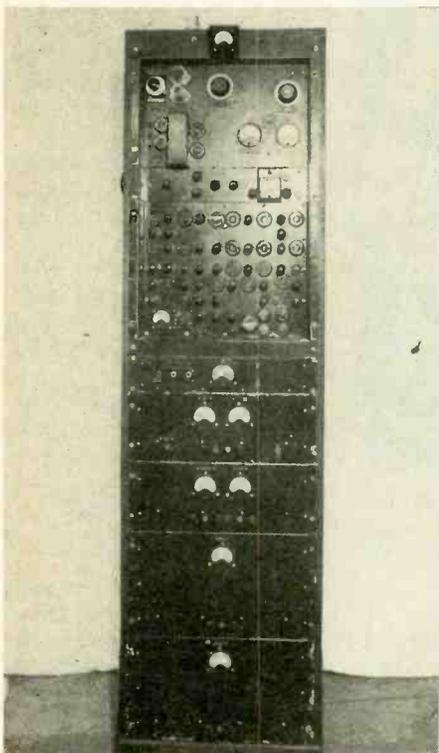
At the quasi-synchronous sites, the 3kHz pilot frequency is multiplied to approximately 42MHz and compared with the third harmonic of the transmitter oscillator. The oscillator is phase-locked to the multiplied 3kHz pilot frequency and itself multiplied to the final transmitter frequency of approximately 450MHz.

Because of the high multiplication factor, the biggest problem was that of noise and jitter on the 3kHz pilot tone transmitted over the line. This was overcome by the use of a low frequency crystal bar as a filter.

In use, the system is more synchronous than quasi, the frequencies of the transmitters remaining locked absolutely for several seconds before slipping out of synchronization and quickly re-synchronizing. This avoids the annoying repetitive beats apparent on some quasi-synchronous systems.

John Davies was with the radio branch of the Lancashire Constabulary for 36 years, latterly as chief engineer, and was personally involved with many of the developments described in his article.

1949: One of the three remote-controlled synchronous f.m. base-station transmitters supplied by Mullard. Finer details of its operation seem to have been forgotten.



In this new series, Tony Atherton recounts some turning points in the development of electrical communications and sketches the personalities who brought them about.

Pioneers

1. Stephen Gray (c.1666-1736): discoverer of electrical conduction

W.A. ATHERTON

The discovery of electrical conduction and insulation must rank as one of the most important technical discoveries in history. The names of later pioneers such as Morse, Bell and Marconi are well known to the public, yet the name of the man who took the very first step is unknown even to most electrical engineers.

Scientific breakthroughs are not usually made by gentlemen enjoying a quiet retirement, but the conduction of electricity was indeed discovered by a pensioner: Stephen Gray, by then ten years into his retirement.

Knowledge of electricity was primitive in the early 18th century. From ancient times it had been known that if amber was rubbed with fur, for example, it would attract small, light objects (such as a feather) to itself. This discovery of electrostatic attraction, the first discovery in electrical science, is still remade daily, usually by young children wielding plastic combs. William Gilbert (the founder of magnetism as a science) in 1600 announced other materials possessing the same property and gave them all the collective name of electrics, from the Greek name for amber. We now call them insulators. Anything else was a "non-electric".

By 1700 progress had occurred. Otto von Guericke in Germany had made a spinning ball of sulphur, which if allowed to rub against the hand would build up a sizeable static electric charge. It is now acknowledged as the first electrical machine. Magnetic and electrical phenomena were known to work in vacuum as well as in air, and electrical glow discharges in vacuum — the first electric light — were being studied.

Such was the state of what would eventually be called electrostatics with which Stephen Gray became familiar as he took to the fascinating pastime of electricity. Whilst living in Cambridge he had assisted the first

Plumian professor of astronomy and experimental philosophy and he had given a few electrical demonstrations. He had also learned of what for 40 years or so was to become the standard method of producing static electricity — vigorously rubbing a long glass tube.

began "that justly celebrated series of experiments which...only ended with his last breath in 1736." He even dictated an account of his final experiments to the secretary of the Royal Society of London the day before his death.

The principle of electrical conduction was discovered in February 1729, a discovery which would lead in the following century to world-wide electric telegraphy.

Gray knew that amber, glass, resin and so on (all insulators) could be electrically excited (charged) by the standard techniques of rubbing, hammering and heating; and he had tried, but failed, to do the same thing to metals. It occurred to him that if a glass tube was charged it might be able to pass on the charge (electric virtue, in Gray's words) to the metal. So he tried it.

His glass tube was 3ft 5in long, had a bore of one inch, but was larger at the ends than in the middle. (I often wonder if it was a yard of ale!) He capped the ends with corks to keep out the dust, but the corks as it turned out did more than that.

The first experiment was to check whether the corks affected the power of attraction of the tube. They did not, but...

"I then held the feather over-against the flat end of the cork, which attracted and repelled many times together: at which I was much surprised."

The corks were electrified. The glass tube had passed on the "electric virtue" to the corks!



Gray also found that the human body is a conductor. In this demonstration, a boy suspended from the ceiling has been electrified by the glass tube and is showing his powers of attraction.

In 1719 Gray became a pensioner at the Charterhouse, a London institute for retired gentlemen, and soon after he started the work for which he is remembered more than 250 years later. In one writer's words, he

begin "that justly celebrated series of experiments which...only ended with his last breath in 1736." He even dictated an account of his final experiments to the secretary of the Royal Society of London the day before his death.

GRAY PURSUES HIS DISCOVERY

A four-inch long fir stick was next stuck into a hole in an ivory ball and the other end Gray fixed into the cork. Both the cork and the stick conducted the electricity and the ivory became charged. The same thing happened with an eight-inch stick and with a 24-inch one, and with brass

wire and iron wire. Finally he tied the ball to a length of packthread (strong string) and hung it directly from the glass rod. It worked!

As the string got longer Gray ran out of height in which to experiment and so was forced to lay the string horizontally. To prevent it from touching the floor he supported it with another piece of packthread tied to a nail driven into a beam. Alas when the glass tube was charged nothing happened to the ivory ball. The experiment had failed and he correctly deduced that the charge had escaped up the vertical string, through the nail and into the wooden beam.

Further experiments were then postponed. Working horizontally had failed and he determined to pursue the matter with even greater lengths of string held vertically, which he felt sure would work. He even dreamed of trying the experiment

"...from the top of the cupola of St Paul's, not doubting but the electric attraction would be carried perpendicularly down from thence to the ground."

By late June, Gray was demonstrating his experiments to Granville Wheler (1701-1770) at Wheler's country house; and Wheler, an accomplished "electrician", was soon accompanying Gray in extending the work. The greatest height they could achieve was a 34 feet drop from the clock turret of the house, a distance through which the electricity was conducted with ease.

Gray then tried again to lay the string horizontally, this time using silk supports at Wheler's suggestion. Gray thought it a good suggestion since the silk was so much thinner than the packthread

"...there would be less virtue carried from the line of communication."

The concept of two classes of materials, conductors and insulators, was then unknown (though not for much longer) and there was therefore no idea that silk is an insulator.

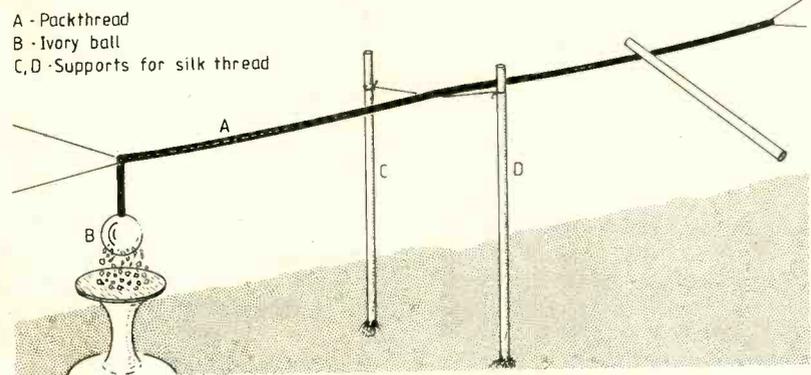
The experiment was tried at 10a.m. on July 2, 1729 and was a great success. The line was 80 feet long and was soon extended to 147 feet by doubling it back on itself. The charge was transmitted from a glass rod through the packthread to the ivory ball which promptly displayed the property of attraction, the then one sure test for the presence of electricity.

Next morning the line was extended as far as 293 feet before the extra weight broke the silk supports. Stronger supports were needed. Fine brass wire was substituted for the silk. It was stronger but very thin and, to their thinking, this thinness meant it would be as good as the silk at stopping the electricity from escaping.

The experiment of course failed. No electricity was conducted to the ivory ball, which gained not the slightest power of attraction. All the charge escaped via the brass wire supports, despite their fineness.

The friends now realised that the critical consideration was not the thickness of the supports but the materials from which they were made. Silk prevented the electricity from escaping, packthread and brass wire did not. The phenomena of conduction and insulation were now apparent.

At ten o'clock in the morning of July 2, 1729 two men in England began an experiment which was to change the course of history. Using string and silk thread they discovered that electricity can be transmitted over great distances.



Gray's crucial experiment: electricity was conducted from the charged glass rod (right) to the ivory ball. Though neglected in our own time, Gray and his achievement were recognized by some, at least of his contemporaries: in 1732 he was elected a fellow of the Royal Society.

42

E S S A Y S.

ON THE DEATH OF

STEPHEN GREY, F. R. S.

THE AUTHOR OF

THE PRESENT DOCTRINE OF ELECTRICITY*.

LONG haft thou born the burthen of the day,
Thy task is ended, venerable GREY!
No more shall Art thy dextrous hand require
To break the sleep of elemental fire;
To rouse the powers that actuate Nature's frame,
The momentaneous shock, th' electric flame,
The flame which first, weak pupil of thy lore,
I saw, condemn'd, alas! to see no more.

Now, hoary Sage, pursue thy happy flight,
With swifter motion haste to purer light,

* The Publisher of this Miscellany, as he was assisting Mr. Grey in his experiments, was the first that observed and notified the emission of the electrical spark from a human body.

E S S A Y S.

43

Where BACON waits with NEWTON and with BOYLE
To hail thy genius, and applaud thy toil;
Where intuition breaks through time and space,
And mocks experiment's successive race;
Sees tardy Science toil at Nature's laws,
And wonders how th' effect obscures the cause.

YET not to deep research or happy guests
Is ow'd the life of hope, the death of peace.
Unblest the man whom philosophick rage
Shall tempt to lose the Christian in the Sage;
Not Art but Goodness pour'd the sacred ray
That cheer'd the parting hour of humble GRAY.



This eulogy, published in 1766 in *Miscellanies in Prose and Verse*, is by Anna Williams, who knew Gray during his time at the Charterhouse and whose father was also a pensioner there. The copy is now in the British Library.

A 54-page illustrated article, *The Enigma of Stephen Gray* by David H. Clark and Lesley Murdin of the Royal Greenwich Observatory, appeared in *Vistas in Astronomy*, 1979, vol.23 (Pergamon Press).

It was soon clear that all materials (or so it seemed) could be classified as behaving either like silk (insulators) or brass (conductors).

Gray prophetically termed his conducting thread a "line of communication". A little later J.T. Desaguliers, with whom Gray had once lived, suggested the terms insulator (from the Latin for island) and conductor. Previously the only words had been Gilbert's "electric" and "non-electric".

Gray and Wheler used their new-found knowledge to extend their maximum distance to 765 feet in July and to 886 feet in August. The basis for electrical communication had been laid — but not yet recognised.

In France, Charles du Fay (1698-1739) confirmed and extended Gray's discoveries and found that metal wire and moist string worked even better than packthread. He went on to realise that there are two fundamental types of electricity which he called vitreous and resinous (amber is a resin). Benjamin Franklin eventually gave them the easier but less imaginative names of positive and negative.

Gray meanwhile had found that the human body makes a nice conductor, as anyone who has touched the live mains will confirm. In doing this he suspended from the ceiling a young (and probably rather frightened) boy. He was strung up by insulating cords and laid horizontally, face down. Gray electrified him by holding a charged glass tube near his feet and noted that metal leaf was then attracted to the boy's face. Du Fay repeated the experiment and even suspended and charged himself.

The experiment was repeated in Germany in about 1743, the same year in which Georg Boze in Leipzig amused himself and others by charging pretty girls and daring young men to kiss them. The shock, it was said, broke their teeth.

Gray had also found in 1732 that the effect could be passed from one boy to another if they held hands or were linked by a conductor. After the invention of the Leyden jar in 1745 (the first storage capacitor) the Abbé Nollet in France persuaded 180 Royal Guards to link up and act as a conducting path through which he discharged a fully-charged Leyden jar. Not surprisingly all 180



guards leapt into the air simultaneously. He next formed a circle of 5800 feet consisting of Carthusian monks linked together with iron wire. Again their vigorous reactions were simultaneous.

The object of these strange experiments was to determine the speed of transmission of the electricity. It was apparently instantaneous.

Two years previously J.H. Winkler, a German physicist, concluded that the speed was comparable with that of lightning and it could be transmitted to the ends of the Earth.

In the following century it was.

Tony Atherton works at the Independent Broadcasting Authority's Engineering Training College in Devon. His book, "From Compass to Computer, A History of Electrical and Electronics Engineering" was published by Macmillan Press in 1984.

Hans Christian Oersted, discoverer of electromagnetism, will be the subject of the next article.



IEE CONFERENCE ON TELEVISION HISTORY

On 13-15 November the Institution of Electrical Engineers paid its own tribute to the pioneers who gave us television by staging a comprehensive conference on its history, its present state, and — to a limited extent — its future. Around 100 distinguished members of the profession, mostly from the UK but with an American and European presence also, took part in the three-day meeting (which was organised jointly by the IEE, the RTS and other professional bodies). Over 40 papers were presented and discussed, and this brief report cannot cover them all but highlights a few that readers may find of particular interest.

The first two sessions looked in some detail at the early pre-war history, and presented an intriguing picture of the closely parallel course of development in the three main centres — Germany, Britain and the United States, with papers on each national history. As might be expected, there was some gentle jockeying for position and rival claims on who was first with what, where. The picture remained far from clear (out of sync?) even when the session chairman was led to suggest that the arguments be continued elsewhere over a glass of wine — but the conclusion reached by this observer at least, is that here was a genuine case of the almost simultaneous development of a technology — electronic television — whose time had arrived.

In the early stages at least, the exchange of information on a commercially sensitive subject — which television undoubtedly was — very limited, and confined to the broadcast principles. Zworykin himself published the basic patent on the iconoscope on 31st March 1932 and described it in the IEE's own proceedings in 1933 (*JIEE* 73, 437); but the essential development of a working, reliable camera tube from those basic principles seems to have been accomplished largely independently in the three countries.

As to the argument "who first transmitted a public service of high-definition pictures?", well, it all depends on what you mean by high-definition. From Dr J. Knies-tedt's paper, it was clear that the German postal authorities, with their trial transmissions in 1932-1934, and their opening of regular public transmissions in March 1935, antedated the BBC's service from the Alexandra Palace by over a year. But the Berlin transmissions used electromechanical scanning on a 180-line standard — arguably medium rather than high definition — and in addition, receivers were not available for purchase: the programmes were watched in public viewing rooms. By August 1936 however, for the Berlin Olympic Games, a German-built iconoscope camera was in use. And it was, of course, on August 26th, 1935 that the BBC's electronic services really started, with the special transmissions of the show "Here's Looking At You" to the Radiolympia exhibition.

The story of British television development was succinctly put by Pat Leggett of the BBC, and readers can follow this in more detail from his excellent article on p.17 of last month's issue. There was also a very revealing paper on the early days at HMV and EMI, presented by J.A. Lodge.

While in Europe the race was neck-and-neck, in the United States development proceeded at a rather slower pace as far as public broadcast services were concerned. The paper by M.J. Sherlock on the NBC's role in the development of television, and that by Leslie and Robert Flory on the early work at RCA, left no doubt that the efforts put in, and the achievements, were considerable. Low-definition experiments started around 1930, not long after the NBC was created as a subsidiary of RCA, and test transmissions on a 343-line standard were taking place by 1936; but a full public service did not commence until the opening of the New York World's Fair in 1939.

The final two sessions ranged over the present state of the art, with digital techniques coming to the fore, and prospects for the future — satellite technology, digital standards conversion and the progress towards fully compatible higher-definition systems. To illustrate this theme, a demonstration of h.d.t.v. was mounted during the conference by the IBA research department. Delegates were also able to see a small but fascinating exhibition of historical documents and equipment. Overall, the conference was a stimulating event, and its record will be of undoubted value in the history of this twentieth-century technology.

P.B.

* A 200-page volume of papers, Conference Publication 271, is available at £30 from IEE Publications, PO Box 26, Hitchin, Herts.

Within the 68020

Besides new instructions, the 68020 has additional hardware features – cache memory, dynamic bus sizing and pipelining.

DAVID BURNS AND DAVID JONES

Adding a coprocessor enhances the main general purpose processor by incorporating new instructions, registers and data types into the system without overloading the main processor.

Interfacing between the main processor and coprocessor is not noticed by the user i.e. the programmer need not be aware that a separate piece of hardware is executing some of the program-code sequence. In the 68020 microcode within the device takes care of coprocessor interfacing so that any coprocessors appear as a natural extension to the main processor architecture.

Using devices without a coprocessor interface such as the 68008, 68000, 68010 and 68012, communication between the main c.p.u. and coprocessor is possible by observing the correct sequence of coprocessor primitives necessary for the interface. These primitives are a method of passing commands and data between the main processor and coprocessor.

Accessing coprocessors over the coprocessor interface is straightforward since the interface is implemented using standard M68000 asynchronous bus structure without the need for any special signals. This not only makes the interface simple; because of the asynchronous nature, the main processor and coprocessor can be operating at different clock frequencies. Designers can therefore optimize a system to make best use of the speed options available.

The coprocessor need not be architecturally similar to the main processor but can be designed so that it best suits its required application. The only requirement is that it adheres to the coprocessor interface protocol. A coprocessor can indeed be implemented as a v.l.s.i. device, as a separate board or even as a separate computer.

When communicating with a coprocessor the MC68020 executes bus cycles in c.p.u. space to access a set of interface registers (CIR). The 68020 indicates that it is accessing c.p.u. space by encoding the function-code lines as all high ($FC_{0-2} = 111_2$). Chip selection of the coprocessor and the relevant register is then performed by the address bus.

Encoding of the address bus during coprocessor communication is shown in Fig. 1a. This illustration shows that by using the 'Cp-ID' field on the address bus up to eight separate coprocessors can be interfaced concurrently to the MC68020. Figure 1b shows how simply this can be done.

You can see that, if required, there could be several 68881 floating point coprocessors operating concurrently in your system to facilitate very fast and complex number crunching. Interfacing to these separate

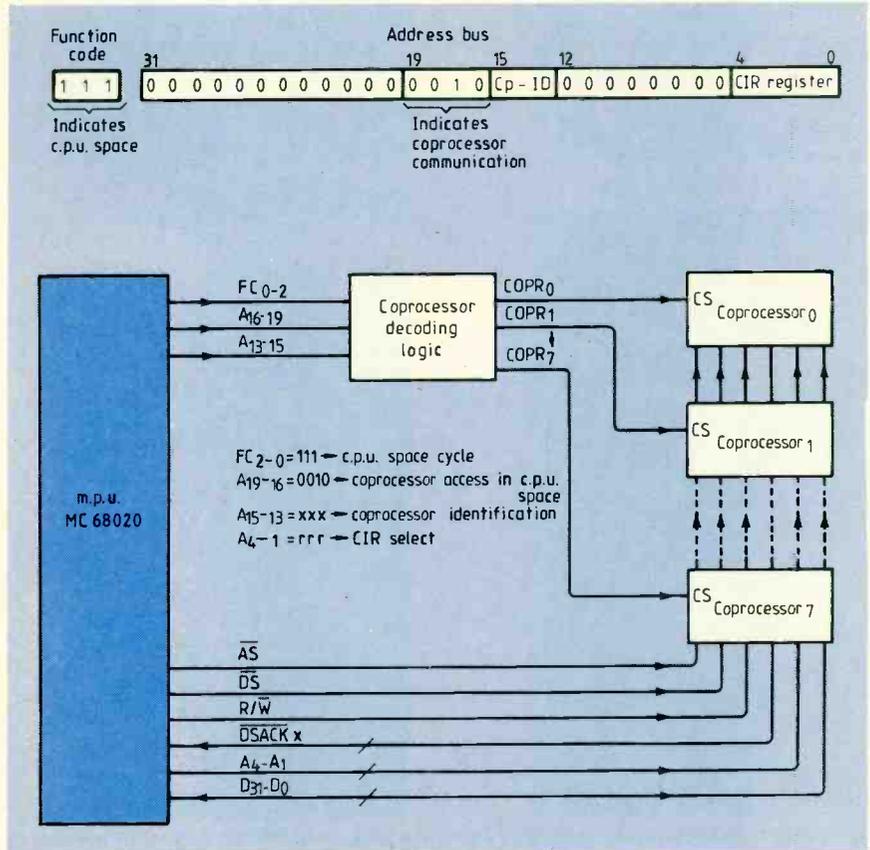


Fig.1. Address bus coding during coprocessor communication (a) and an illustration of how up to eight coprocessors can be connected (b) using the Cp ID field.

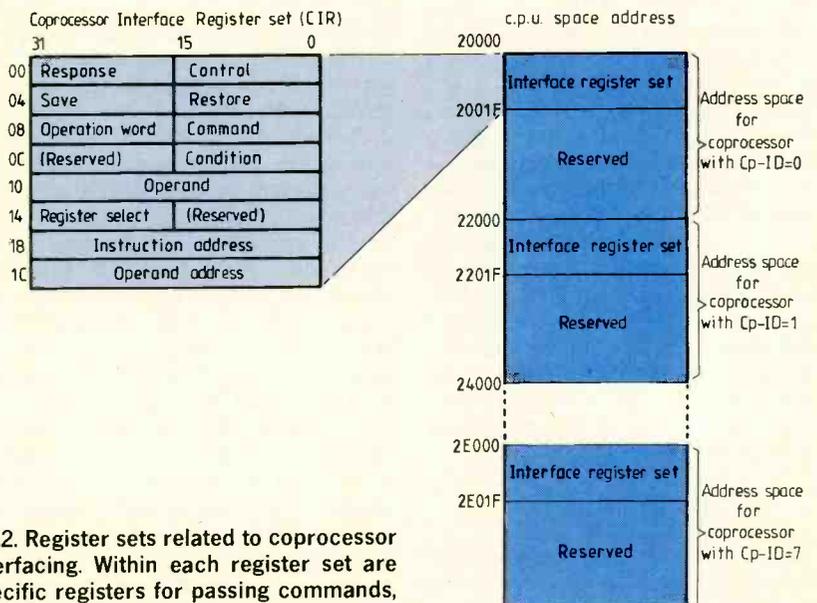


Fig.2. Register sets related to coprocessor interfacing. Within each register set are specific registers for passing commands, operand data and effective addresses.

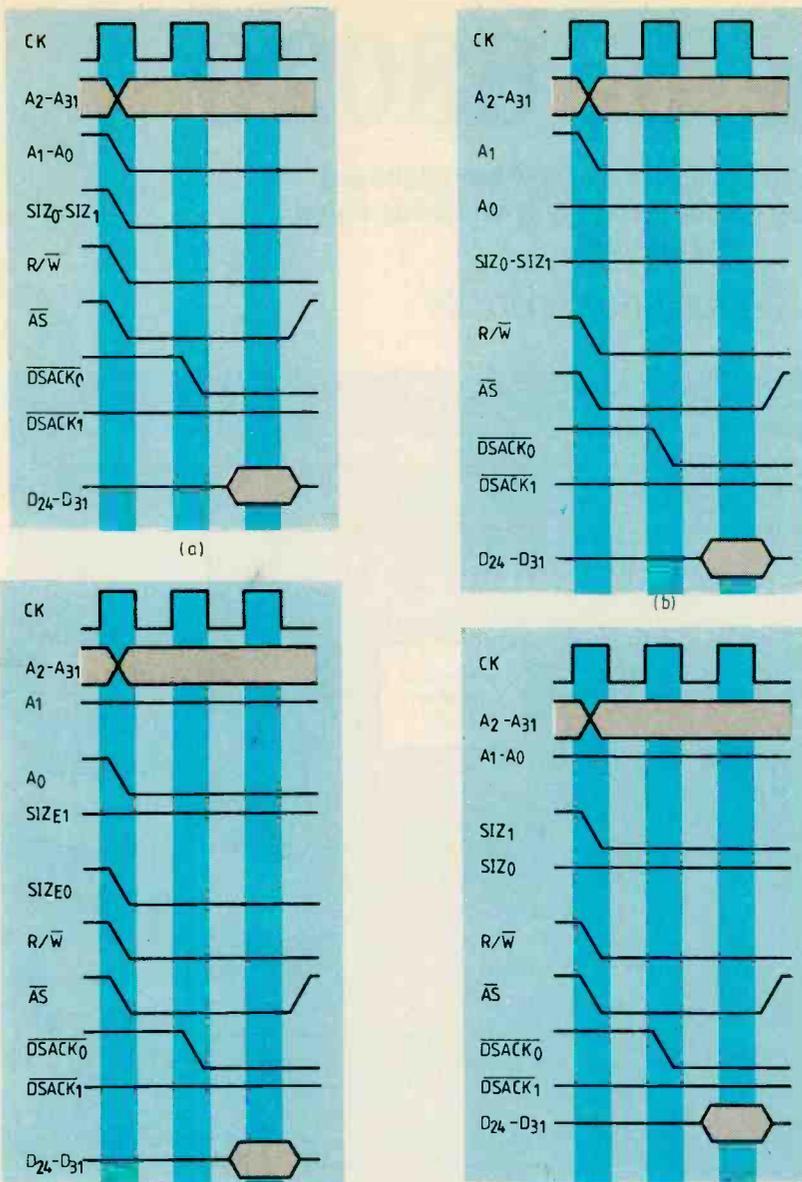


Fig.3. Timing of the four cycles required to write a long word into an eight-bit memory. In all four diagrams, \overline{DSACK}_0 is zero and \overline{DSACK}_1 is one, indicating that the memory is eight bits wide. At (a), signals $SIZ_{0,1}$ are both low to indicate that the processor has four bytes to transfer and the states of these two signals change from 1,1 to 0,1 to 1,0 in (b), (c) and (d) to indicate three, two and one bytes to transfer respectively.

coprocessors is simply a matter of encoding the relevant Cp-ID in the coprocessor instruction, and hence on the c.p.u. space address-bus encoding, so that the MC68020 will communicate with the relevant register set in c.p.u. space.

Figure 2 shows how the separate interface register sets are located in c.p.u. space. Within this interface register set the various registers are allocated to specific functions required for operating the coprocessor interface. There are registers specifically for passing information such as commands, operand data and calculated effective addresses (effective address calculations, and associated operand fetches, are performed by the main processor). Other registers are allocated for use during a context switch when the internal state of the coprocessor needs to be saved and then restored.

DYNAMIC BUS SIZING

The 68020 can dynamically change the size of the data bus on each bus cycle. This feature has been included so that the proces-

sor can communicate with peripheral devices intended for 32, 16 or 8-bit buses. Dynamic sizing can also be used to retrofit a 68020 in a 16-bit system and although the full performance increase of a 32-bit processor is not gained, performance improvement can be considerable.

Four signals have been added to support dynamic bus sizing, namely $\overline{DSACK}_{0,1}$ and $SIZ_{0,1}$. Data transfer and device-size acknowledge signals $\overline{DSACK}_{0,1}$ replace the \overline{DTACK} (data transfer and acknowledge) asynchronous-bus handshaking on the 68000. As with the 68000, these signals are used to terminate the bus cycle but they also indicate the external size of the data bus. Signals $SIZ_{0,1}$ are outputs indicating how many bytes are still to be transferred during a given bus cycle.

To illustrate this principle, consider the 68020 writing a long data word (32 bits) to an eight-bit memory device. (This would have a serious effect on system performance as four write cycles would be needed). Bus timing diagram Fig.3 shows the four cycles required to write the information into memory.

Using dynamic bus sizing, during a write cycle the 68020 always drives the entire 32-bit data bus even though all 32 bits may not be used. If data is transferred as a byte, it is placed on D_{24-31} , if it is transferred as a word it is placed on D_{16-31} and if it is transferred as a long word D_{0-31} are used.

A multiplexer within the 68020 routes data to various sections of the data bus depending on bus size. Address lines $A_{0,1}$ are linked to this multiplexer. Their encoding indicates a one, two or three-byte offset for a long word to be read from or written to memory.

Unlike other 68000-family processors, the 68020 allows you to place data misaligned in memory, including the user and supervisor stacks. As far as the programmer is concerned this misalignment goes unnoticed in the hardware but it affects performance by increasing the number of data transfers. In fact the only limitation on data storage in the 68020 is that the instruction word, or opcode, must lie on a word or long-word boundary. This is to retain upward software compatibility with 68000-family software.

Consider this example illustrating the principle of misaligned data transfers. Data is transferred to memory by the 68020 over a 32-bit data bus, however the memory address has been offset by one byte from a long-word location.

Figure 4, in which the timing diagrams are simplified to show just the signals used to control data flow, clarifies the situation. Tables summarize decoding of the SIZ and \overline{DSACK} signals. Also shown is a representation of how the data is organized within the processor.

Since the data is misaligned by one byte the processor needs to make two memory accesses to transfer the long word. During the first cycle, three bytes of data are transferred and the second cycle transfers the last byte.

During the first transfer the processor sets the SIZ pins low to tell the memory that the processor has four data bytes to transfer (a long word). The memory address is odd and offset by one byte from a long-word address so address lines $A_{0,1}$ are at logic one and zero respectively.

Using $\overline{DSACK}_{0,1}$, the memory controller indicates that it is 32 bits wide. Information placed on the data bus, in long-word form displaced by eight bits, is carried on D_{23-0} . The lower eight bits are not transferred during this cycle. Information placed on D_{31-24} is just a mirror image of data on D_{23-16} and should be ignored by the controller during a write cycle.

During the second transfer, the long-word transfer is completed. Again, $\overline{DSACK}_{0,1}$ indicate a 32-bit port but the SIZ pins indicate to the memory controller that only one byte remains to be transferred. Data is transferred as one byte on address/data lines D_{24-31} . Inside the processor, data is transferred within the registers on the opposite end of the data bus. The remainder of the data bus carries a mirror image of this data and should be ignored during write cycles to memory.

Figure 5 shows data transfers over a 16-bit port misaligned by one byte. In this example,

it takes three data transfers to transfer a long word. These examples illustrate that the 68020 can be designed into systems with 32, 16 or indeed 8-bit data buses and with a large possibility of data being placed in aligned and misaligned memory.

CACHE MEMORY

Increasing the speed of the 68020 from 16.7MHz to 20 and 25MHz has been accompanied by increases in the cost and difficulty of interfacing the device to external memory without using wait states. The device's internal 256-byte instruction cache relieves this problem.

A minimum of three clock cycles is required when the processor accesses external memory. However if the information is held in the cache, which can be thought of as very fast on-chip local memory, then only two clock cycles are required.

Computer simulation tests were carried out for the 68020 based on the 68000 architecture to find out what type and size of cache would be most beneficial; 256 bytes was found to be the best compromise between efficiency and cost. The cache theory rightly assumes that modern computer programming involves the program repeatedly executing small sections of code as opposed to randomly jumping over large linear address spaces.

When the processor fetches an instruction from memory, the processor is redundant since no processing can be performed until the instruction has been decoded (this is not strictly true for the 68020). If this instruction fetch can be performed from the cache then the processor spends less time waiting for information from external memory. This has an even greater effect if there is a memory-management unit (m.m.u.), external bus or magnetic backup storage in the system. In such cases it may not always be possible to access external memory with no wait states.

The cache, Fig.6, is 'hit' when address field $A_{8,31}$ and function-code FC_2 (indicating instruction accesses) match the international cache tag field. Cache hit is a term used to describe the condition where the address and any other control information presented on the bus matches information previously placed in the cache tag field.

Some 64 long words are available for storage of cache information. Since the cache is always updated on a long word, maximum throughput is achieved when two instructions in memory are held in the cache. Address lines $A_{2,7}$ select one of 64 entries. Upon reset, the cache is disabled and all entries are made invalid; the v bit in the tag field is also cleared.

Two registers are used with the cache – the cache-control register, CACR, and the cache address register, CAAR. Enabling, disabling and clearing of the cache is carried out by the control register. This register can also be used to freeze individual entries in the cache so critical code sequences can be run within the cache.

It is not possible for the programmer to access cache entries directly. Programming the cache registers is performed using the MOVEC instruction and so can only be done in

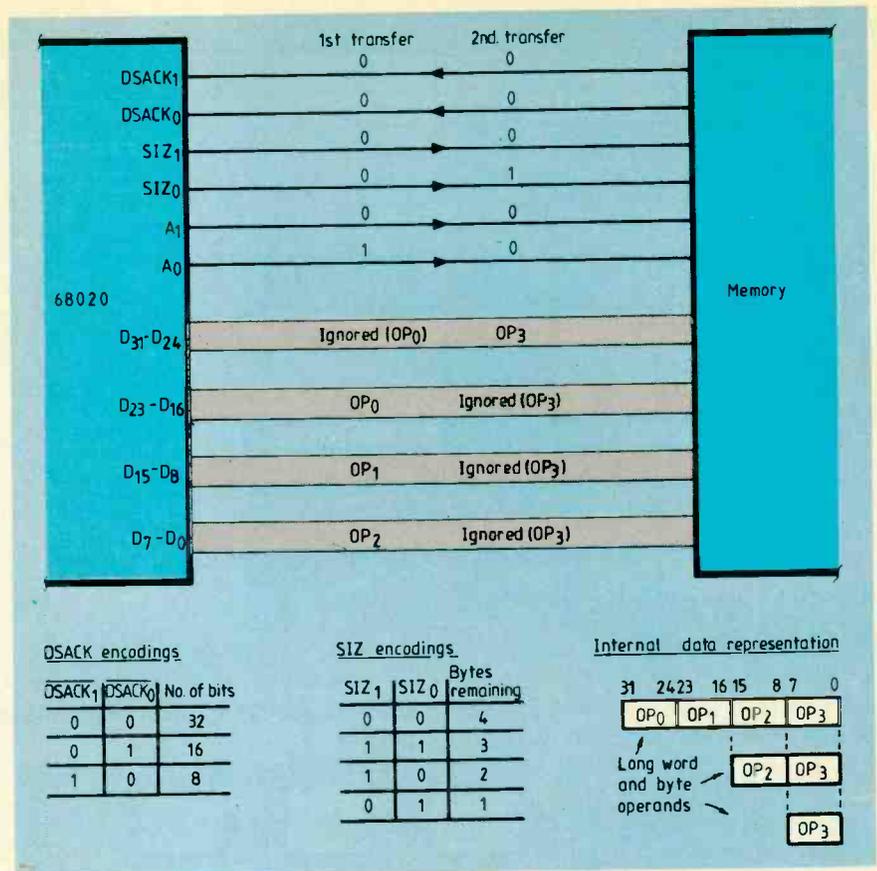


Fig. 4. To send data to memory over a 32-bit address bus when the memory address is offset by one byte from a long-word location requires two transfers. The operand is 32 bits and the boundary is one byte.

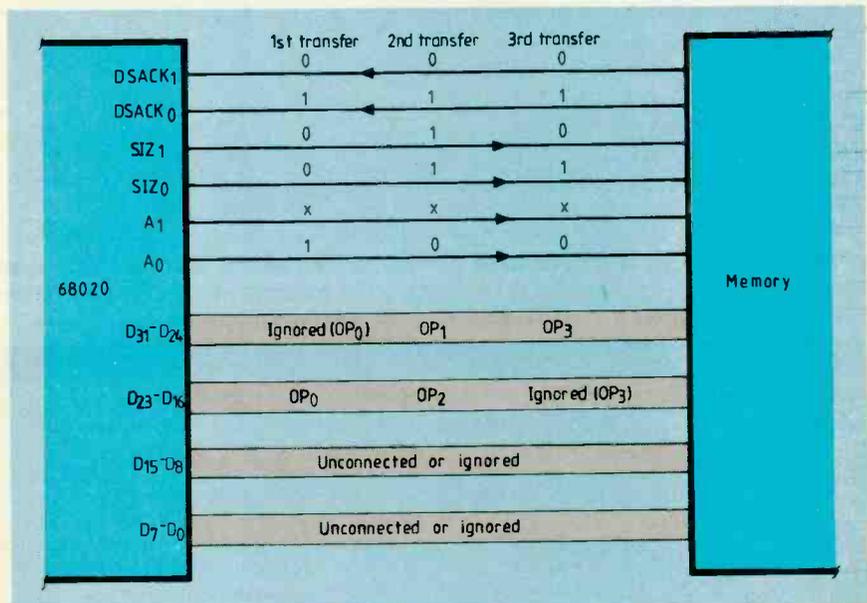


Fig.5. To transfer a long word over a 16-bit port misaligned by one byte takes three transactions. The operand is 32 bits and the boundary odd. Data lines $D_{0,15}$ are either unconnected or ignored by both the processor and memory on 16-bit port-sized transfers.

supervisor mode. This ensures that the user cannot accidentally effect the cache operation. Using cache clearing, the operating system can perform a fast context switch in just one instruction.

In addition to the software cache-enable facility, an external hardware cache disable pin, CDIS, can be used to dynamically disable the cache on the next internal cache-access boundary.

THE PIPELINE

Within the 68020 is a three-stage pipeline used for instruction execution, Fig.7. Instructions enter the pipe either from the external bus or from the instruction cache, also within the processor. These instructions are not the ones currently being executed but are 'prefetched' instructions obtained by the bus-interface unit.

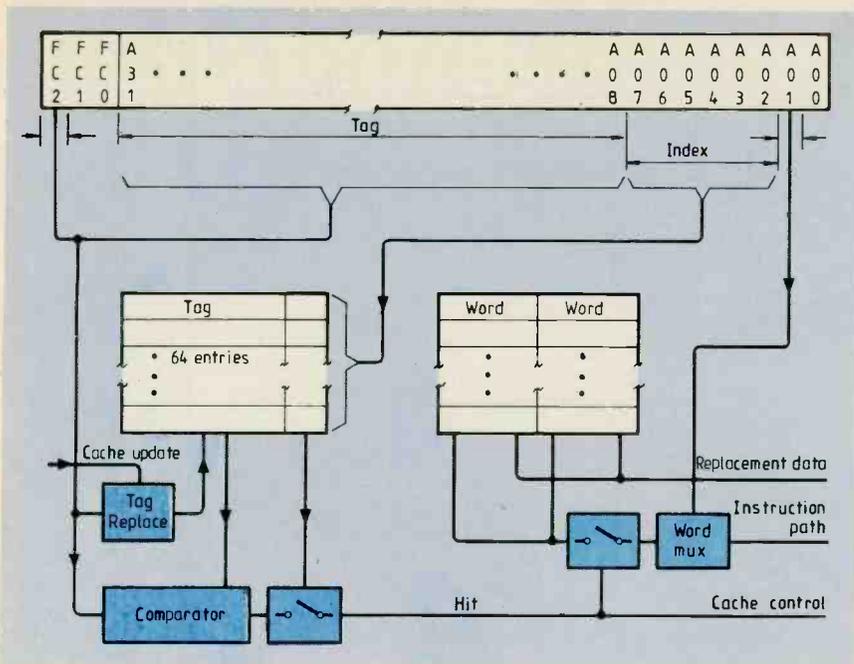


Fig.6. Accessing external memory slows down processing so the 68020 processor has a 256-byte internal cache memory to reduce this problem.

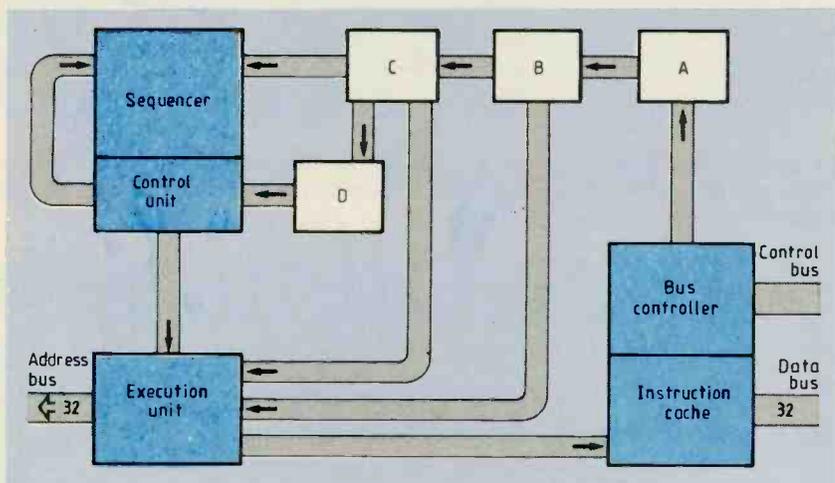


Fig.7. Within the 68020 is a three-stage pipeline which reduces execution time since external bus cycles are not needed to fetch instruction extension words, etc. Instructions enter the pipe either from the external bus or from the instruction cache.

MASTER/INTERRUPT STACKS

Most high-performance microprocessors have two stack pointers. One is usually to the system stack, reserved for interrupts, etc., and the other is the user stack for temporary data storage and parameter passing. In the 68020, these stacks are A_7 and A_7' .

During normal operation most code will be executed in user space and programs will use the A_7 stack for temporary data storage and parameter passing between software routines. Interrupt stack A_7' will only come into use when an exception occurs, such as an external interrupt when control is passed to supervisor mode and the relevant exception processing performed.

In many microprocessors this supervisor stack pointer is the only one accessible during exception processing and all data storage and context switching has to be performed on only one stack. With complex multi-tasking maintaining the main system stack costs processing time; interrupt information for the program counter and status register is interleaved with process-control blocks for various software tasks.

This problem is alleviated in the 68020 by a third stack – the master stack – specifically for holding process-related information for the various tasks. When the master stack is enabled, through bit M in the status register, all non-interrupting exceptions like divide-by-zero, software traps and privilege violation are placed in the user's process-control block on the master stack.

When the first interruption occurs, typically a timer interrupt from a preemptive scheduler, the processor places the program counter, PC , the status register, SR , and the vector offset on the master stack A_7' . It then duplicates this information on the interrupt stack A_7' . The processor is now free to manipulate the processor control block without any further interrupt information being placed on the master stack.

All subsequent interrupts received while performing exception processing are placed only on the interrupt stack A_7' . An effective context switch can now be performing by simply reloading the master stack pointer and mapping in another task's process-control block. This allows context switching to be performed without any master-stack modification by higher-priority interrupts which may occur during the exception processing for the preemptive scheduler's timer interrupt.

With prefetching, long words are always fetched and the cache is arranged as long words. After a prefetch has been requested, the long-word obtained is placed in a 32-bit holding register (stage A) called the cache holding register. This register is used to hold the prefetched long word in the case of a cache miss or to hold the prefetched long word from the cache if a hit occurs.

When a cache hit occurs, the external bus performs an aborted cycle with the \overline{ECS} pin asserted followed by a second assertion of \overline{ECS} . Address-strobe pin AS is not asserted during a cache hit. If data is prefetched from the external data bus, it is routed directly to the 32-bit arithmetic logic unit, or a.l.u. This a.l.u., with which is associated a 32-bit barrel shift register, performs arithmetic and logical operations on data registers.

From the cache holding register, instructions pass to stages B, C, and D, where the instruction is executed. Movement through the pipe is governed by the execution time of the instruction at stage D. On reaching stage D of the pipe, the executing instruction corresponds to the program counter. However the processor needs to know where the extension words, etc, for the instruction are so temporary pointers are set up for each stage of the pipe and holding register. These pointers are used to obtain data from each stage of the pipe to allow completion of the instruction.

Using pipeline architecture, the processor operates much faster since no external bus cycles are needed to fetch extension words, etc. Flow of the temporary pointers is tracked by a 32-bit arithmetic unit. There is a third arithmetic unit in the 68020 for arithmetic operations between address registers such as calculating effective addresses. To produce effective addresses needed by the instruction when it is in stage D of the pipe, stage C can be used for effective address calculation. Stages C and D have inputs for allowing them to control the 32-bit a.l.u. and a.us.

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David's hobbies include orienteering, tennis and badminton.

David Jones graduated with an honours degree in Electronic Engineering at Heriot-Watt university in 1982. Since then he has worked for Motorola Semiconductors in East Kilbride. After working as an equipment engineer David transferred to the microprocessor applications group, progressing from eight to 16 to 32 bit processor design. He is currently working on a 25MHz high performance MC68020 computer board.

David's hobbies include white-water canoeing (both recreational and competitive), swimming and car maintenance. He enjoys outdoor life.

Digital computer modelling of linear systems

By one of those odd coincidences, whilst preparing this article, I came across a war-time bombsight computer in a junk sale. It was, in essence, a mechanical analogue computer and must have been extremely difficult to design, manufacture and adjust, particularly as it was mass produced. It served a timely reminder of the traditional difficulty in providing rapid solutions to complex physically related mathematical problems. Consider a mechanical system we all take for granted – the suspension system on cars, consisting of a spring and damper. To describe in detail how this system works under all circumstances is not easy since it requires the solution of a second-order linear differential equation. In terms of automotive engineering, getting the solution wrong can be disastrous because examination of the equation and its solutions reveals that certain combinations of spring and damper characteristics can render the whole system unstable, with potentially lethal results. This is, of course, one of the classic systems studied by generations of engineering students who are taught how to solve the system equation longhand (see example over) but are not expected to do so as a matter of routine, since they are also taught how to simulate the system by means of an analogue computer, and thus provide much more useful continuous solutions.

This article describes how small personal computers can be used to solve these ordinary engineering problems. With the correct approach digital simulations of linear systems are relatively easy, with results that can be shown to be mathematically valid on even the humblest of micros.

ANALOGUE SIMULATION

Most linear differential equations can be simulated by an arrangement of integrators, scalars and summers, and various methods produce the final circuit and correct scaling factors. It must be said however that some of the complexity of the final circuit derived will be due to the sign change introduced by each op-amp and also of course due to the practical limitations of the devices themselves. For instance, the output voltage of an integrator cannot exceed the power supply voltage.

First, let us conveniently assume that the system components are perfect devices. To illustrate this take the equation used in the example on page 108, transpose and divide by M:

$$\ddot{x} = \frac{f(t)}{M} - \frac{D\dot{x}}{M} - \frac{Kx}{M}$$

Ignoring scaling factors and sign changes, this results in the analogue circuit shown to the right.

How to use analogue techniques and a digital computer to solve second-order differential equations with a few simple algorithms.

RICHARD OXLEY

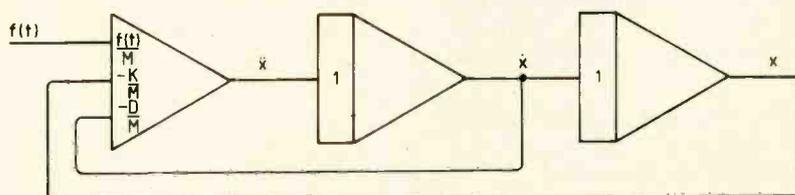
The analogue simulation of a differential equation leads directly to a very useful and simple method of digital system modelling, since it is possible to write down simple algorithms which model all the components of an analogue computer, that is, scalar, summer, integrator, multiplier and even differentiator, although this component rarely needs to be used. Further, just as in the analogue computer where components are connected together, algorithms can be connected together to model systems of any complexity.

Before proceeding, it is as well to review how continuous signals as are used in analogue systems are represented in digital modelling.

DIGITAL MODELLING OF SIGNALS

Unit steps. In an analogue computer, the unit step is simulated by the instantaneous application of a scaled voltage by a switch. The digital modelling equivalent of this is merely the statement currently being executed in the program:

LET A=X where X is the scaled value.



Continuous signals. The normal interface approach is to use sampling. A sinusoidal signal can be sampled by an a-to-d converter to give a series of numbers in real time for input to a computer at a rate determined by a clock. However, for modelling and simulation purposes, the converter can be replaced by a simple software routine which will produce exactly the same set of numbers, but at a rate determined by the execution speed of the computer. This has no significance, just as the absolute scale of the axes of a graph drawn on paper in no way alters the meaning of the results depicted.

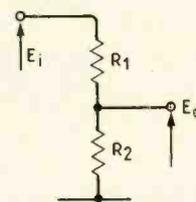
For instance, try the routine:

```
10 Input "AMPLITUDE": A
20 Input "FREQUENCY": F
30 Input "SAMPLING FREQUENCY": S
40 Q = 0
50 X = A * SIN(Q * 360 * F * S)
60 Q = Q + 1
70 PRINT X
80 GOTO 50
```

This will model a sinewave of frequency F, amplitude A at a regular sampling interval of S by producing an endless stream of numbers to either printer or screen.

DIGITAL MODELLING OF LINEAR COMPONENTS

Scaler. A potentiometer divides a signal by a constant which is determined by the position of its wiper on a track, viz:

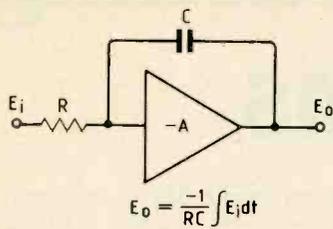


Provided the wiper does not change its position this is the equivalent of division by a constant. Obviously, simple multiplication by the constant shown will produce the desired result:

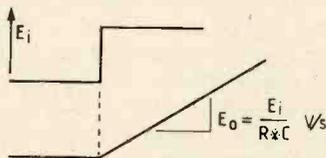
$$Y = X_1 (R_2 / [R_1 + R_2])$$

Y is the instantaneous value of the output signal after being processed by a potential divider with an instantaneous input signal X.

Integrator. The integrating element of an analogue computer normally comprises an op-amp with capacitive feedback as shown.



In practical terms, if a voltage is suddenly applied to the input, the output will change at a linear rate according to the time constant RC and the magnitude of the input voltage.



The digital form of this mechanism is a series of numbers of ascending value separated by a time relationship. The difference in voltage between one sample and the other is

$$X = E_i \star S / R \star C$$

where S is the sampling interval. To produce a ramp the simple expedient of continuously adding the increment to the previously derived value can be adopted:

$$X = X + E_i \star S / (R \star C)$$

This is known as a recursive function and is the digital equivalent of a linear integrator.

Summer. The other important component used is the op-amp summer. From what has already been given it is pretty obvious that the digital equivalent of that is merely adding two variables, viz:

$$A = B + C$$

Multiplier. Similarly, multiplications of two variables – horrendously difficult in its analogue form – is likewise easy:

$$A = B \star C$$

Multipliers, incidentally, are used in the solution of *non-linear* differential equations, often met in systems designed to solve navigation and guidance problems.

DIGITAL MODEL

Armed with all the algorithms needed to implement a full linear system and returning to the original example, the schematic of an analogue computer connected to solve the linear second-order differential equation this time uses the symbols to represent algorithms – see top of page 109.

Using the derived algorithms and working from left to right, now write down the program steps required to simulate/model the circuit:

- 10 $X = (F/M) - (D/M \star Y) - (K/M \star Z)$ summer
- 20 $Y = (X \star S / (R \star C)) + Y$ integrator
- 30 $Z = (Y \star S / (R \star C)) + Z$ integrator
- 40 Print Z

CLASSICAL ANALYSIS OF SECOND-ORDER SYSTEM

If a spring is either compressed or extended a distance x by a force:

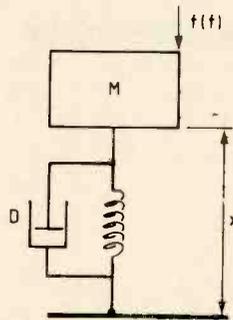
$$x \propto F \therefore F = Kx$$

where K is the spring constant ($K = Mg/x$).

A damper only has an effect when the system moves, in particular the resistive force to movement:

$$\frac{dx}{dt} \propto F \therefore F = Dx$$

where D is the damping factor.



If a mass is at rest it can only move by an application of a force. That is, the application of a force imparts an acceleration

$$\frac{d^2x}{dt^2} \propto F \therefore F = M\ddot{x}$$

In the case of the spring-mass system, the force producing the acceleration is equal to the sum of the forces acting on the mass. In equilibrium

$$M\ddot{x} = -Kx - D\dot{x}$$

$$M\ddot{x} + D\dot{x} + Kx = 0$$

When the system is subjected to an applied force, which may be a function of time,

$$M\ddot{x} + D\dot{x} + Kx = f(t)$$

The classical method for solving practical examples of this mechanical system involves the use of the Laplace operator. Unlike the analogue or digital simulator, when differing values for $f(t)$, M , D and K can be used directly, the two examples given next illustrate how in the classical method different values lead to the use of different mathematical tricks to implement a solution.

EXAMPLE 1

$M = 5\text{kg}$, $D = 40\text{N/m/s}$, $K = 80\text{N/m}$, $f(t) = F = \text{stepforce of } 60\text{N}$.

$$M\ddot{x} + D\dot{x} + Kx = f(t)$$

$$5\ddot{x} + 40\dot{x} + 80x = 60$$

$$\ddot{x} + 8\dot{x} + 16x = 12$$

Applying Laplace:

$$s^2x^2 + 8xs + 16x = \frac{12}{s}$$

$$x(s^2 + 8s + 16) = \frac{12}{s}$$

$$\therefore x = \frac{12}{s(s+4)^2}$$

To proceed further, the equation should be split into partial fractions such that:

$$\frac{12}{s(s+4)^2} = \frac{A}{s} + \frac{B}{s+4} + \frac{C}{(s+4)^2}$$

$$\therefore 12 = A(s+4)^2 + Bs(s+4) + Cs$$

Put $s=0$, so $A=0.75$. Put $s=-4$, so $C=-3$.

To find the value of B requires the application of another devious mathematical trick. Equating highest powers: $0 = 0.75 + B$, $\therefore B = -0.75$

$$\therefore f(t) = \frac{12}{s(s+4)^2} = \frac{0.75}{s} + \frac{-0.75}{s+4} + \frac{-3}{(s+4)^2}$$

Applying inverse Laplace transforms gives the final general solution:

$$x(t) = 0.75 - 0.75e^{-4t} - 3te^{-4t}$$

EXAMPLE 2

If the value of D is reduced to five the general solution is radically different. Even the method to find it although basically the same is tediously different. For the intrepid:

$$5\ddot{x} + 5\dot{x} + 80x = 60$$

$$\ddot{x} + \dot{x} + 16x = 12$$

By Laplace, $xs^2 + xs + 16x = \frac{12}{s}$

$$\text{and } x = \frac{12}{s(s^2 + s + 16)}$$

Proceeding by partial fractions:

$$\frac{12}{s(s^2 + s + 16)} = \frac{A}{s} + \frac{Bs + C}{s^2 + s + 16}$$

$\therefore 12 = 0.75(s^2 + s + 16) + s(Bs + C)$. Dating $s=0$ given $A=0.75$, while equating coefficients of powers gives $B=-0.75$, $C=-0.75$.

$$\therefore x(t) = \frac{0.75}{s} - 0.75 \left(\frac{s+1}{s^2 + s + 16} \right)$$

The quadratic in the denominator of the second term is factorized by completion of squares:

$$x(t) = \frac{0.75}{s} - 0.75 \left(\frac{s+1}{(s+0.5)^2 + 15.75} \right)$$

$$= \frac{0.75}{s} - 0.75 \left(\frac{s+1}{(s+0.5)^2 + 3.97^2} \right)$$

The s -term in the numerator of the second term requires some trickery to enable the 'first shift' theory to apply

$$x(t) = \frac{0.75}{s} - 0.75 \left(\frac{(s+0.5)}{s+0.5^2 + 3.97^2} + \frac{0.5}{(s+0.5)^2 + 3.97^2} \right)$$

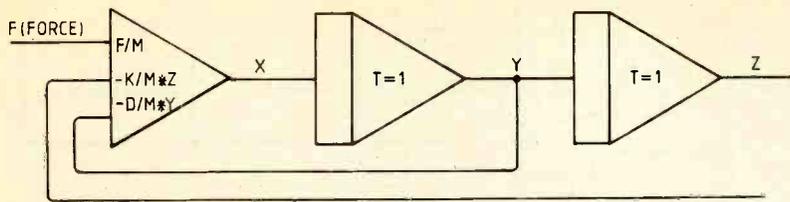
Applying the inverse transform $x(t) =$

$$0.75 - 0.75(1 - e^{-0.5t} \cos 3.97t + \frac{0.5}{3.97} e^{-0.5t} \sin 3.97t)$$

$$= 0.75[1 - e^{-0.5t}(\cos 3.97t + 0.126 \sin 3.97t)]$$

This result is a damped oscillation with a final settling value of 0.75 , shown diagrammatically on page 110. The oscillation has a frequency $3.97/2\pi = 0.632\text{Hz}$, a period of 1.58s .

For more about the solution of differential equations using Laplace transforms consult: Laplace Transforms by K.A. Stroud, Stanley Thornes (Publishers) Ltd.



where S is sample time, RC is nominal integrator time constant.

It must be remembered that these three equations form a set since they are recursive functions. That is, one equation without the others is meaningless. This is highlighted by the way in which the equations must be incorporated in a program loop. The program incorporating these equations is given in the listing. On a microcomputer it will run until escape is keyed and will output to screen a stream of numbers that represent the action of the real system.

PARALLEL PROGRAMMING

It must be said that the spring/damper used for our example is a very simple second-order system allowing a direct approach to analogues simulation and thence to a digital model. The example solutions show that even with such a simple system obtaining a solution relies on the use of such operator methods as Laplace.

Turning again to the example, it was shown that the displacement of the mass, x, when a step force was applied had the form.

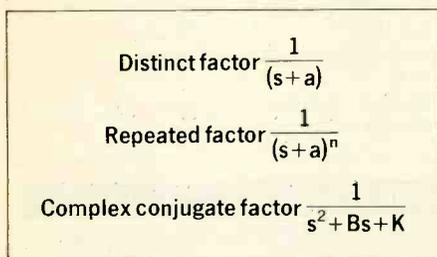
$$x = R \frac{1}{s} \frac{1}{(s^2 + Bs + K)}$$

The denominator has two parts: 1/s as the Laplace transform of the step function, and

$$\frac{1}{s^2 + Bs + K}$$

the Laplace system transfer function or impulse function. Therefore the algorithm previously derived through the direct method forms a digital model of the system transfer function.

When attempting to model complicated systems, the direct approach can become involved. If the system transfer function (or indeed the compound transfer function comprising the system transfer function operated on by the transform of the forcing function) is expanded by partial fractions only three kinds of terms arise, viz: distinct factors, repeated factors and complex conjugate factors, as shown in the panel.



Distinct factor. Remembering that the linear integrator is the equivalent of 1/s, the factor 1/(s+a) results in an exponential rather than a linear output. In fact, it is the equivalent of a simple RC circuit.

Producing such an exponential by algor-

ithms is quite simple since it is a modification of that derived for the linear integrator, viz:

$$X = X + E_i \star S / (R \star C) \quad \text{linear}$$

$$X = X \star (1 - |S / (R \star C)|) + E_i \star S / (R \star C) \quad \text{exponential}$$

remembering $R \star C = 1/a$.

Repeated factors. The algorithm for the repeated factor is the algorithm for the distinct factor repeated n times, viz

$$10X = X \star (1 - S \star a) + E_i \star S \star a$$

$$20Y = Y \star (1 - S \star a) + X \star S \star a$$

$$30 \text{ etc.}$$

where E_i is input stimulus.

Complex conjugate factor. An algorithm that models complex conjugate factors has already been described in the spring/mass example.

Having expanded a complex system equation by means of partial fractions into distinct, repeated and/or complex conjugate factors, a software model would run the algorithms for these in parallel; the full solution would be the simple sum of the algorithms. An example which illustrates this method and models a complex second-order system is given as programming example 2.

DIGITAL FILTERING

I have stuck with the example of the damped spring mass because it leads to a further application of the algorithms that were derived earlier. This system can be considered as a filter. (The suspension system of a car filters out a high proportion of the undulations presented to the wheels by the road surface.) It comes as no surprise then that the analogue computer circuit can be used as a general purpose filter which, depending on the particular output connection, acts as a high-pass, low-pass or even bandpass filter.

By outputting either X, Y or Z, a high-pass, band-pass or low-pass filter can be modelled whose a.c. characteristics can be

predicted. To illustrate this consider the equation for the centre frequency of the circuit.

$$f_0 = \frac{1}{2\pi} \sqrt{\frac{R_6}{R_5}} \cdot \sqrt{\frac{1}{R_1 C_1 R_2 C_2}}$$

If the gain of the two integrators is chosen to be unity: $1/R_1 C_1 = 1/R_2 C_2 = 1/f_0$ becomes

$$\frac{1}{2\pi} \sqrt{\frac{R_6}{R_5}}$$

Now this is equivalent to the well-known formula relating to the resonant frequency of the original mechanical system:

$$f_R = \frac{1}{2\pi} \sqrt{\frac{K}{M}}$$

where K is the spring constant, and M is mass. Comparing diagrams it can be seen that K/M is equivalent to R_6/R_5 , which sets the gain about that particular loop.

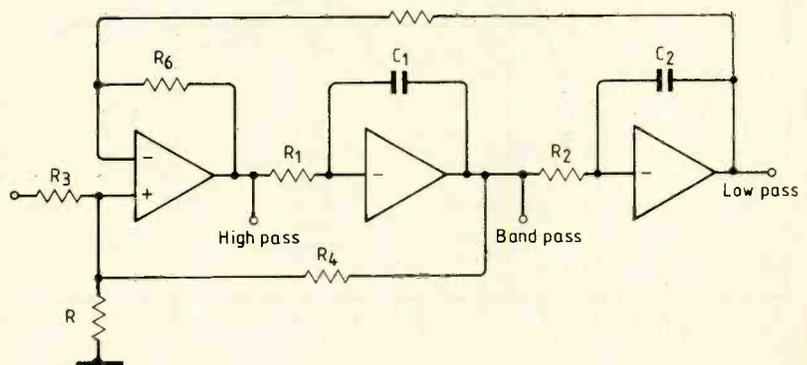
This shows incidentally that to speed up or slow down the response of an analogue simulation the integrator gain (1/RC) can be changed from unity. For digital modelling the gain of the integrator has less significance and can normally be kept at unity, thus simplifying the algorithms. Conversely, to model such an analogue filter to examine its characteristics, then the appropriate time constants for the integrators can be entered into the algorithms directly.

All this provides a useful way of checking the validity of the algorithms. In the original example $F = 60$, $M = 5$, $D = 40$, $K = 80$.

$$f_0 = \frac{1}{2\pi} \sqrt{\frac{K}{M}} = 0.6366 \text{ Hz}, T = 1.57 \text{ s.}$$

The first programming example (page 110) shows that these values result in no oscillation because the system is so damped. However, reducing D to 5 is a different story, and the second trace exhibits damped oscillations with a period of between 1.5 and 1.6s. In fact, the trace exactly models a step function being processed by a low-pass filter with a centre frequency of 0.6366Hz.

This form of recursive digital filter is widely used in the digital implementation of linear control systems. Other forms of digital filter designed to extract continuous signals from random broadband noise can be very complex, relying on statistical techniques. But for most purposes the recursive digital filter works very well provided that due attention is paid to sampling theory to avoid



problems such as aliasing, that is at least two samples per period.

The filtering algorithms can be very useful in some practical applications of micro-computers, for instance, control of heating systems by remote sensors. The obvious system to use is shown in the diagram to the right.

Old hands will know that such a system is liable to all kinds of interference pickup problems from the electricity mains. The sensors will be measuring temperatures that will be varying at a very slow rate but picking up interference at 50Hz and above. Instead of adopting the usual method of hanging big capacitors on everything the problem is avoided by incorporating the above algorithms in the control program to realise a low-pass digital filter.

SOFTWARE OSCILLATION

From the examples given, it will be apparent that under certain circumstances the second-order system as shown will oscillate. In general, if the damping factor (for the mechanical example) or circuit resistance (for the electrical equivalent) is reduced to zero then the general solution will be a simple sinewave.

Now an algorithm for generating the sine function for input purposes to digital models and filters was given earlier, but of course, it was written in Basic and relies on the host computer containing an algorithm for determining the sine of an angle. In some circumstances this facility may not be available and so the algorithms derived for the spring-mass (L,C,R) system may be used instead. By choosing appropriate values any frequency may be generated and may, indeed, be faster than the internal routine. The oscillator algorithm then can be used to generate forcing functions for other algo-

ithms which model systems (see second programming example, page 111).

This method is very convenient because it is obviously possible by simple software to switch the algorithm on and off at any point in the cycle by loop counting.

Engineers who have struggled with the Heaviside unit step function and the Dirac delta impulse function to show how a complex control system behaves when the input forcing function is an arbitrarily starting sinewave or a pulse will know what an emancipation this is!

It is possible also, by resort perhaps to simple arithmetic, to derive algorithms for forcing functions which model other waveforms too - like square waves or sawtooth waves. The output of a bridge rectifier, for example can be modelled by merely ignoring the sign of the sinewave generator algorithm.

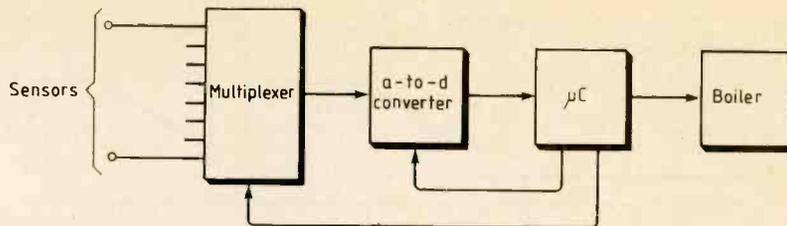
In the spring-mass system, if the damping factor B is reduced to zero, the equation becomes

$$\frac{1}{s^2 + K}$$

This is the Laplace transform of

$$\frac{1}{\sqrt{K}} \sin \sqrt{K} t.$$

where of course, $f = \sqrt{K}/2\pi$.



USING THE ALGORITHMS

Care is required when modelling systems to choose a suitable sampling interval. Often surprisingly fast transients are generated by systems, particularly when second or higher order. However, a two low a sampling rate can usually be recognised in the nature of the output from the algorithm since it will show variations in magnitude at the sampling rate. In extreme cases, the output will merely oscillate between rapidly increasing positive and negative values. Trial and error usually determines a suitable value.

However, when the algorithm has been derived from a partial fraction expansion of the system transfer function it may be useful to continue and solve the resultant equation into its cyclic and exponential parts. Since the Laplace transform of the delta Dirac impulse function is 1, the solution gives the impulse response of the system and reveals its frequency components.

It must also be remembered that complex forcing functions will contain frequency components higher than its fundamental. No matter how slow the system is the output will contain harmonics, admittedly attenuated, of that fundamental. The sampling rate then becomes a function of the forcing function.

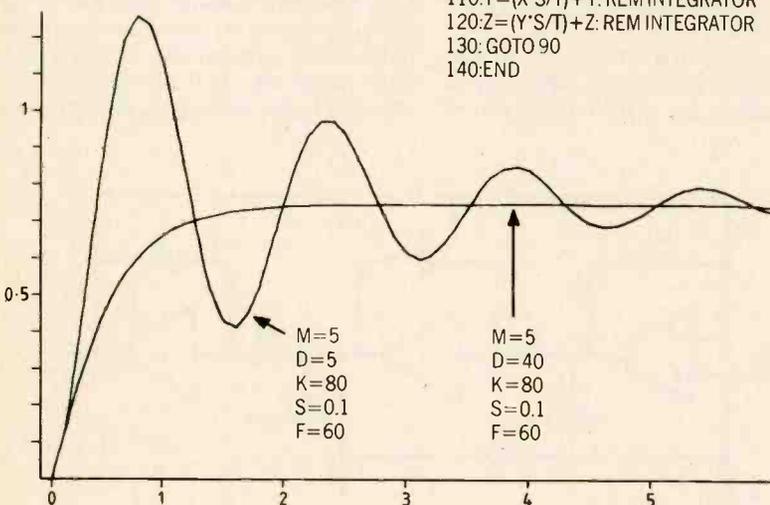
Dick Oxley is a senior systems engineer with British Aerospace Air Weapons Division at Lostock, near Bolton, currently working on infra-red reconnaissance equipment.

EXAMPLE 1: DAMPED SPRUNG MASS

Program shown prints output as a number. Output value Z in line 90 can be linked to a graphics routine to produce traces shown.

```
10:REM SECOND ORDER SYSTEM
20:X=0:Y=0:Z=0
30:INPUT "MASS=";"M:PRINT" M=";"M
40:INPUT "DAMPING=";"D:PRINT" D=";"D
```

```
50:"SPRING=";"K:PRINT" K=";"K
60:INPUT "SAMPLE TIME=";"S;
PRINT" S=";"S
70:INPUT "R'C=";"T:REM NORMALLY 1
80:INPUT "FORCE=";"F:PRINT" F=";"F
90:PRINT USING"#####"Z
100:X=(F/M)-(D/M*Y)-(K/M*Z);
REM SUMMER
110:Y=(X*S/T)+Y:REM INTEGRATOR
120:Z=(Y*S/T)+Z:REM INTEGRATOR
130:GOTO 90
140:END
```



FURTHER READING

Automatic Control Engineering, F.H. Raven, McGraw Hill.
 Random Data Analysis and Measurement Procedures, J.S. Bendat & A.G. Piersol, Wiley.
 Digital Processing of Signals, Gold & Radar, McGraw Hill.
 Frequency Analysis, R.B. Randall, Brüel & Kjaer.
 Digital Control of Dynamic Systems, G.F. Franklin & J.D. Powell, Addison-Wesley.
 Digital filter design, W.J. Rees, *Wireless World* Oct. 1976.
 Simple digital filters, Palham, *Wireless World* July 1979.
 Analogue computing techniques, D.F. Dawe, *Wireless World*, June/July 1980.

GPIB INTERFACE

T. Segaran's December article on low-cost automated response measurement showed two devices marked 74160 and 74161; these two should be types 75160 and 75161.

EXAMPLE 2: PARALLEL PROGRAMMING

Consider a system with the following transfer function:

$$T_f = G \left\{ \frac{1 + T_{11}s + T_{12}s^2}{(1 + T_{21}s)(1 + T_{22}s)} \right\}$$

When b is an arbitrary gain constant in its present form this cannot be expanded by partial fractions as the numerator and denominator are of the same order; some manipulation is therefore necessary. Multiplying out the denominator:

$$\begin{aligned} T_f &= G \left\{ \frac{1 + T_{11}s + T_{12}s^2}{1 + (T_{21} + T_{22})s + T_{21}T_{22}s^2} \right\} \\ &= G \left\{ 1 - \left[1 - \frac{1 + T_{11}s + T_{12}s^2}{1 + (T_{21} + T_{22})s + T_{21}T_{22}s^2} \right] \right\} \\ &= G \left\{ 1 - \left[\frac{1 + (T_{21} + T_{22})s + T_{21}T_{22}s^2 - (1 + T_{11}s + T_{12}s^2)}{1 + (T_{21} + T_{22})s + T_{21}T_{22}s^2} \right] \right\} \\ &= G \left\{ 1 - \left[\frac{(T_{21} + T_{22} - T_{11})s + (T_{21}T_{22} - T_{12})s^2}{1 + (T_{21} + T_{22})s + T_{21}T_{22}s^2} \right] \right\} \end{aligned}$$

Assume that we can choose the following conditions – the reason will soon be apparent: $T_{21} = T_{11}$ and $T_{21}T_{22} = T_{12}$. Substituting:

$$T_f = G \left\{ 1 - \left[\frac{T_{22}s}{1 + (T_{21} + T_{22})s + T_{21}T_{22}s^2} \right] \right\}$$

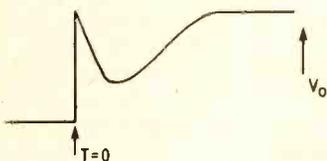
Factorizing the denominator gives

$$T_f = G \left\{ 1 - \frac{T_{22}s}{(1 + T_{21}s)(1 + T_{22}s)} \right\}$$

which can now be expanded by partial fractions to the form:

$$T_f = G \left\{ 1 - \left[\frac{A}{1 + T_{21}s} + \frac{B}{1 + T_{22}s} \right] \right\}$$

The general response of this form of transfer function when excited by a step function is



which is recognised as that of a notch filter. Assume that this needs to be tuned to 50Hz, which gives $T_{21} = 0.00072$, and $T_{22} = 0.0138$. Substituting these values in the above equation and expanding by partial fractions, the expansion is

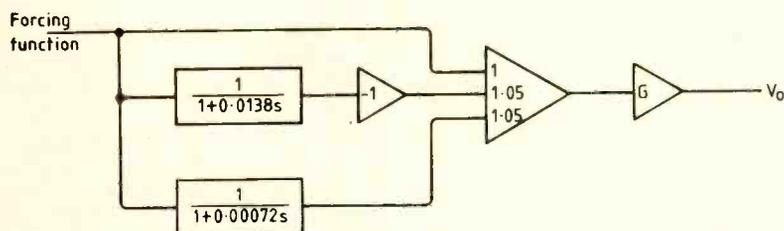
```

10: "A": GRAPH :
    TEXT
20: LPRINT "NOTCH
    FILTER"
30: CLEAR
40: DIM C(50)
50: INPUT "S="; S
60: LPRINT "S="; S;
    "secs"
70: INPUT "INPUT F
    REQUENCY"; F;
    LPRINT "f="; F;
    "Hz"; AA=2*PI*F
80: INPUT "T21="; A
90: LPRINT USING "
    ##.#####"; "T2
    1="; A; "secs"
100: INPUT "T22="; B
110: LPRINT USING "
    ##.#####"; "T2
    2="; B; "secs"
120: INPUT "G="; G;
    LPRINT USING "
    ###.###"; "GAI
    N="; G
130: Q=1.05
140: LPRINT "Q/P fo
    r 1mS interval"
150: PAUSE "CALCULA
    TING"
160: "CALC": XX=1-(A
    A*AA*ZZ): REM
    
```

$$\begin{aligned} T_f = \frac{V_o}{V_{in}} &= G \left\{ 1 - \left[-\frac{1.05}{1 + T_{22}s} + \frac{1.05}{1 + T_{21}s} \right] \right\} \\ &= G \left\{ 1 + \frac{1.05}{1 + 0.0138s} - \frac{1.05}{1 + 0.00072s} \right\} \end{aligned}$$

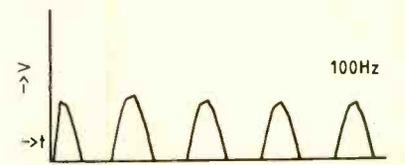
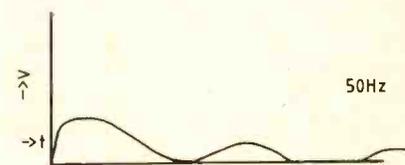
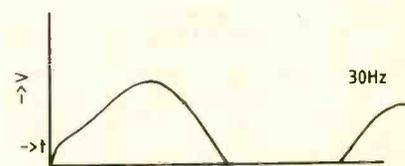
The listing gives a complete program that models the analogue simulation. Included is the algorithm to generate sinusoidal forcing functions. Numerical printouts and their graphic equivalent demonstrate the action of the digital filter.

In this way it is possible to model practically any analogue filter. Although some mathematical dexterity is required to expand the transfer function by partial fractions the resultant algorithms will be very simple and will consist of only the three factors described, viz: distinct, repeated and complex conjugate.



```

LOOP ENTRY POINT
170: YY=(S*XX)+YY;
    REM 1/AA SINA
    AT
180: ZZ=(S*YY)+ZZ;
    REM END OF SI
    NE ALGORITHM
190: X=(AA*YY*S/A)+
    ((1-(S/A))*X);
    REM 1/s+A
200: Y=(AA*YY*S/B)+
    ((1-(S/B))*Y);
    REM 1/s+B
210: W=(Y*AA)-(Q*X
    )+(Q*Y); REM A
    DD TERMS
220: U=(G*W); T=T+1;
    REM U=Q/P
230: IF T=10 THEN
    LET Z=Z+1;
    LPRINT USING "
    ##.#####"; U; "
    V"; T=0; C(Z)=U*
    300
240: REM PRINT AFT
    ER 10 ITERATIO
    NS & STORE IN
    ARRAY FOR GRAP
    HICS
250: IF Z=50 THEN
    LPRINT "END";
    GOTO "GRAPH"
260: GOTO "CALC"
    
```



By entering appropriate constants and setting appropriate input frequencies the behaviour of the filter can be readily observed (dip at 50Hz for instance). Note response to sudden application of forcing function.

Electrometer amplifier

continued from page 65

may be displayed on a strip-chart for 'at a glance' interpretation. Freedom from the disturbances attendant upon range switching operations are added advantage of this mode of operation.

It is well known that, if the feedback resistor of the conventional operational amplifier circuit is replaced by a junction diode, the amplifier output approximates to the logarithm of the input signal and it may be of interest to see how this principle is applied to our projected electrometer amplifier.

Figure 5 depicts the typical semiconductor diode characteristic; a simplified expression for this relationship between applied voltage and the resulting current is

$$I = I_{rs}(e^{\frac{qV}{kT}} - 1)$$

I_{rs} = reverse saturation current

where V = applied voltage

q = electron charge

k = Boltzmann's constant

From this expression it can be seen that when $V = 0$, $I = 0$ and when $V \geq 60\text{mV}$, it may conveniently be further simplified to

$$I \approx I_{rs}e^{\frac{qV}{kT}}$$

at normal ambient temperatures.

Taking logs of the revised expression,

$$\log_e I = \log_e I_{rs} + \frac{qV}{kT}$$

and hence

$$V = \frac{kT}{q} \log I - \frac{kT}{q} \log I_{rs}$$

As the last term is constant at constant temperature,

$$dV/d\log I = kT/q$$

At 300°K this gives a figure of approximately 60mV per decade: a relationship which holds for perhaps ten decades determined by the physical dimensions of the diode.

Normal signal diodes are unsuitable at the minute currents measured by electrometer amplifiers and it is common practice to use a high H_{FE} small-signal transistor as the feedback element as shown in Fig.6. In this arrangement $I_{in} = I_c$, which closely approximates to I_e if H_{FE} is high and constant. As with the linear amplifier, external leakage is

minimized by the use of a transistor of small area and, for this reason, the transistor is often reversed to make its emitter the input terminal, it being the smaller electrode in epitaxial devices.

The lower level of input current to which the logarithmic relationship extends is determined by the external leakage and by recombination current within the transistor, and few discrete devices are suitable at femtoamp levels. However, some of the very small geometries used for l.s.i. microcircuits, for example the Ferranti CDI process, are well suited to the application.

The upper limit of the input current range is determined by ohmic resistance, which only becomes significant at current levels well above our projected maximum, even for devices of such small size.

One interesting outcome of the use of a three-terminal logarithmic feedback element is that the output voltage is essentially unaffected by input offset-voltage variations of the operational amplifier: that is, there is no voltage drift due to this cause. However, the logarithmic response can be distorted at low current resulting from the voltage offset, which must therefore be as small as possible.

Referring to the preceding paragraphs in which the parameters of the logarithmic response were discussed, it will be seen that the output voltage (V_{be}) will range between approximately -60mV and -400mV over the five-and-a-half decades of input current. This is insufficient to drive a strip-chart

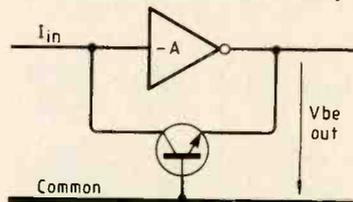
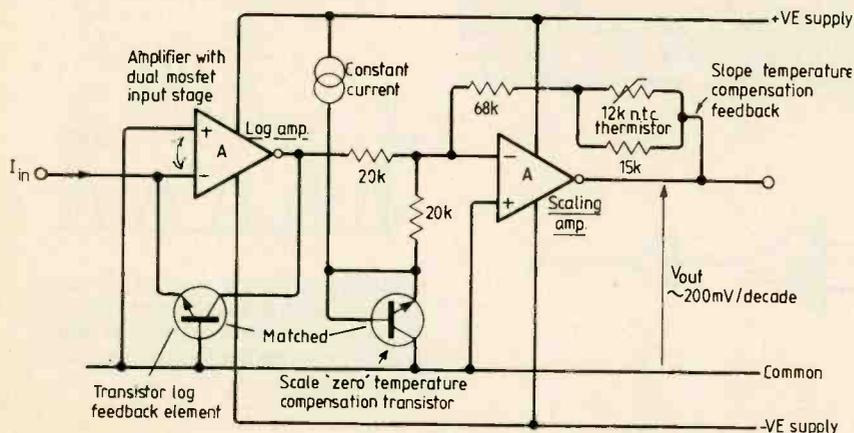


Fig.6. Replacing the feedback resistor by a junction diode provides a log response.

Fig.7. Circuit diagram of complete log amplifier with linear stage to drive a recorder. Second transistor compensates for temperature characteristic of feedback transistor.



recorder efficiently and further amplification is necessary. The following linear scaling amplifier also facilitates back-biasing to 'zero' the recorder for the lowest current to be measured and the means for compensating the temperature dependence of the feedback element.

Temperature effects. If V_{out} is plotted against $\log_{10} I_{in}$, the curve produced will of course be a straight line over the range of input current for which the logarithmic relationship holds; the slope of the line being $2.3026 kT/q$ volts per decade, as previously derived. Thus, the slope varies directly in accordance with absolute temperature. To compensate for this temperature dependence, the gain of the scaling amplifier is made to vary inversely with temperature.

EXHIBITIONS AND CONFERENCES

3-4 February 1987

Cellular Radio and mobile communications Conference at the London Press Centre, Shoe Lane, London EC1. Organized by IBC Technical Services Ltd. Tel: 01-236 4080.

12-13 February 1987

Convergence of computing and telecommunications. Conference at Royal Lancaster Hotel, London. Organized by IBC Technical Services Ltd. Tel: 01-236 4080.

17-20 February 1987

Which Computer? Show. National Exhibition Centre, Birmingham. Cahners Exhibitions. Tel: 01-891 5051.

23-27 February 1987

Fiarex 87 international electronics trade fair. Rai Exhibition Centre, Amsterdam, Rai Gebouw, Europaplein, Amsterdam.

3-6 March 1987

International Open Systems Conference (and a MAP seminar, 4 March). Barbican Centre, London Online Exhibitions. Tel: 01-868 4466. Semicon Europa 87 exhibition of semiconductor production equipment and materials. Zuspa Convention Centre, Zurich, Enquiries to Cochrane Communications, Tel: 01-353 8807.

24-26 March 1987

Cadcam 87 exhibition and conference. NEC Birmingham. EMAP Int. Exhibitions, Tel: 01-608 1161. Internecon Production Show and conference - "from CAD to testing" NEC Birmingham. Cahners Exhibitions. Tel: 01-891 5051.

25-26 March 1987

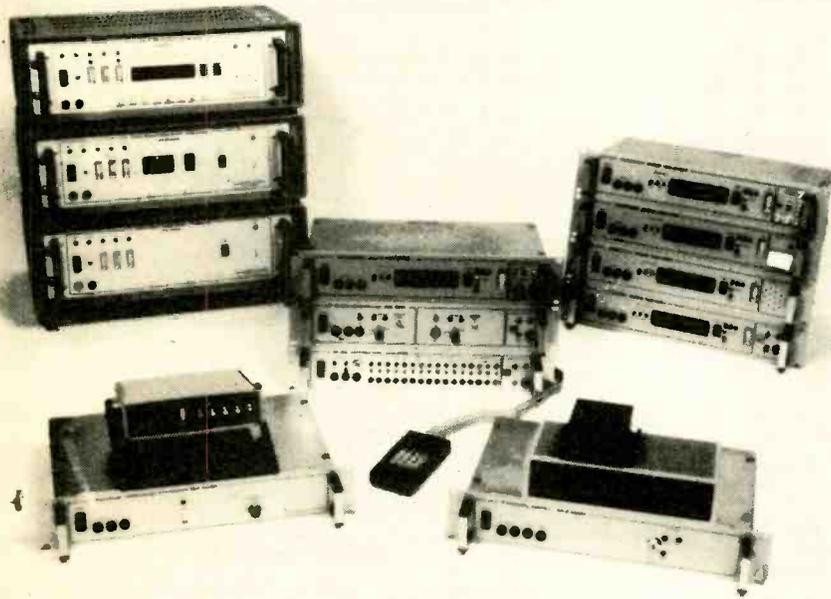
Instrumentation Bristol 87 Exhibition. Bristol Crest Hotel. Trident Int. Exhibitions. Tel: 0822 4671.

31 March-2 April

Scottish Computer Show and conference. Scottish Exhibition Centre, Glasgow. Cahners Exhibitions. Tel: 01-891 5051.

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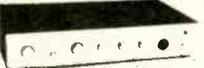
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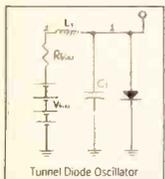
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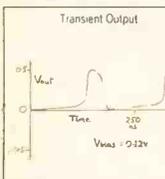
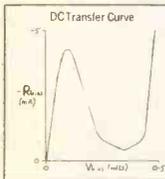
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This month's piece comes from Brendan Rooney, technical director of Exchange Resources.

It comes as a surprise to many people in the electronics and computing fields to find that less than 25% of the industrial jobs market is in the defence sector. The reality of the market is heavily distorted by the recruitment strategies of the large multinationals involved in defence work. For example, most undergraduates are only aware of the larger companies who appear at the 'milk rounds': in reality, there is a vast number of small to medium sized firms working in non-defence electronics and computing who are never seen at recruitment fairs. They can generally offer secure work in non-military employment.

Large numbers of advertisements appearing in technical publications are also for military jobs: again, this distorts the image of the job markets. Job turnover in the defence industry is high and there are certain vacancies that can never be filled, hence the need for intense marketing campaigns. As a result of this state of affairs the defence industry has to employ contractors to fill some of the vacancies, which tend to be very highly paid positions compared with those of the staff working alongside who, in turn, can be underpaid compared with their commercial equivalents. To attract new graduates, reasonable rates of pay are offered, but this can lead to a small pay differential between the new people and long standing employees. Long-standing employees feel that their skills are not portable across to the commercial sector, a belief which is not necessarily true.

Exchange Resources is a new venture in the electronics and computing recruitment field. It aims primarily to encourage employment outside the defence sector. In the belief that this country has far too much of its resources invested in an unnecessarily large and provocative system that has little to do with actual defence. However the ethos of the agency is not simply anti-defence. The agency has been set up to encourage those companies that provide a good working environment for their employees and that take into account each employee's personal views. Hence, Exchange Resources deal with some of the large electronics corporations who are obviously involved in defence work but who have other, non-defence departments. These companies are willing to take on people from Exchange Resources, with the explicit understanding that the employee has specific moral objections to working on certain applications of technology. Exchange Resources therefore wishes to encourage the use of our resources — human and material — in ways which actually strengthen industry and liberalize the working environment. We are not affiliated to any political party or any other organization or their ideologies.

The ideas behind our company originated in

1984. While Tony Wilson was co-ordinator of Electronics for Peace (EFP) in 1984/5, several people had asked him to find them non-military work. At this stage he was in no position to do so, but as a result of this and discussions within the EFP steering group, there was a vague idea of providing a jobs agency and skills register (since at least one such agency already exists in the USA). Then, in the summer of 1984, James Plummer of Prospect International — a successful London-based recruitment agency in the computer field — contacted Tony and asked, "Why don't you set up an agency for people who don't want to work in defence?". A couple of meetings with James gave EFP the impetus to act and the Barrow Cadbury Trust provided an initial grant.

As a result of this backing Tony Wilson began work at the start of 1986 as m.d. of Exchange Resources, initially working three days a week on the project. There are now three directors and we are presently looking for a fourth person with drive and commitment to fill the vacant fourth place. All the directors' shares are held in trust.

MILSOM STREET

Exchange Resources have been based at 28, Milsom Street, Bath since the beginning of July 1986. At present there are three full time workers Tony Wilson — managing director, Jackie Greenwood and Sally Aust — who both work in the Milsom Street offices and three part time workers Hugh Deynem, Brendan Rooney and Steve Bryant. Although the organisation is a limited company it is intended to keep a simple two-tier hierarchy, so that we remain fairly close to the idea of a co-operative. It was decided fairly early on in the process of formation to set up as a limited company since there was no experience of co-operatives within those of us interested in Exchange Resources.

The aims of Exchange Resources are to provide staff training and development for our own staff and pay the market rate for our skills.

OPERATION

We offer a recruitment-agency service at very competitive rates.

Since the formation of the agency we have received over 140 c.v.s from very high calibre people in the electronics and computing fields. Presently, we are dealing with around 100 companies and have placed three candidates in the first few months.

The agency receives c.v.s from candidates attracted by our service, which are summarized and held on our Amstrad PCW8256. The summarised c.v.s are then passed to companies having specific vacancies for which the candidate is suited. The summaries are deliberately short so that employers

won't have to trudge through mountains of text to find what they are looking for. When they show interest in a candidate, we can then provide more detailed information. Since these c.v.s are on a database, they can be searched at will, to identify specific skills for example. As part of our business strategy we believe in communicating with all the candidates and identifying the type of work they are looking for; this is not always obvious from a c.v.

Since we are in the electronics and computing fields we feel that we should make full use of electronic mail. Hence the Amstrad is now linked into GEONET mailbox EFP and One-to-One mailbox 25020001. We are currently taking up Telecom Gold and Prestel. These networks will provide candidates and companies with an immediate link to us as well as another media for advertising.

PROJECTS

As well as the recruitment agency business Exchange Resources aims to support companies working on socially useful products and projects. This work includes supporting a number of projects which in the fullness of time can be taken to market. At present we have the following projects on our books.

Eritrean radio. At the present time, we are talking with Warwick University about this project. It is currently seen as a combination of an a.m. (s.w. and m.w.) radio for receiving infrequent national broadcasts and an f.m. radio for receiving low-cost local transmissions. There will be two versions, one for small travelling family groups, using a small loudspeaker and rechargeable batteries, and the other for village centres with a larger loudspeaker. This radio will be assembled and maintained in Eritrea in existing workshops which employ people with disabilities.

Radiation hazard monitoring. This project is aimed at monitoring the effects of non-ionizing radiation specifically the physiological hazards.

Waveform synthesizer. We are presently looking at the feasibility of bringing to market a low-frequency transient-waveform synthesizer. This will be based around a standard PC with corresponding hardware. The system will allow a waveform to be designed on the computer screen.

To conclude then, Exchange Resources are showing industry that there is a market for the novel type of service we provide. We are gradually building up contacts in industry and have met with some very positive responses. Perhaps the least expected came when we were approached by a large electronics company, whose non-military division had been told to deal with us.

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INDEX TO ADVERTISERS

Appointments Vacant Advertisements appear on pages 116-119

PAGE	PAGE	PAGE	PAGE
Advance Bryans Instruments.....18	Ellmax Electronics.....7	Langrex Supplies Ltd.....83	Sherwood Data Systems85
AEL Crystals.....29	EMS Mfg. Ltd.85	Levell Electronics Ltd.....29	Silicon Glen92
Air Link Transformers6	E&WW Editorial Features.....114	Lloyd Research Ltd18	Soft Option39
Cambridge Kits.....12	Feedback Instruments29	Micro Concepts.....57	Stewart of Reading.....39
Carston Electronics.....2,18	Field Electric Ltd57	Microkit Ltd12	Strumech Engineering46
CA Special Products.....6	Flight Electronics Ltd.....18,40	Monolith Electronics Co7	Surrey Electronics Ltd.....22
CIL MicrosystemsIBC	Fluke (GB) Ltd Advert Card	Micromake Electronics92	Taylor Bros (Oldham) Ltd.....8
Cirkit Holdings Ltd.....50	Happy Memories22	MicroProcessor Engineering28	Technomatic Ltd4/5
Clark Masts Ltd8	Harris Electronics (London).....46	Network Satellite Systems.....46	Thanet Electronics Ltd85
Cocking and Drury Ltd7	Hart Electronic Kits50	Number One Systems39	Those Engineers.....114
Computer Appreciation120	Henry's Audio Electronics6	Pineapple Software114	Thurlby Electronics88
Conguin Software Ltd.....46	Henson, R. Ltd28	PM Components Ltd94/95	TIC Semiconductors.....28
Crotech Instruments Ltd22	Hypertac Ltd12	Radiocode Clocks Ltd50	Triangle Digital Services Ltd114
Crouzet Ltd.....28	John's Radio22	Raedek Electronics92	Valradio Power Ltd12
Dean Microsystems18	Kestral Electronic Components.....92	Research Communications Ltd 57	Webster Electronics6
Display Electronics.....112/113			
Electronic Brokers Ltd IFC, Advert Card, 90, OBC			

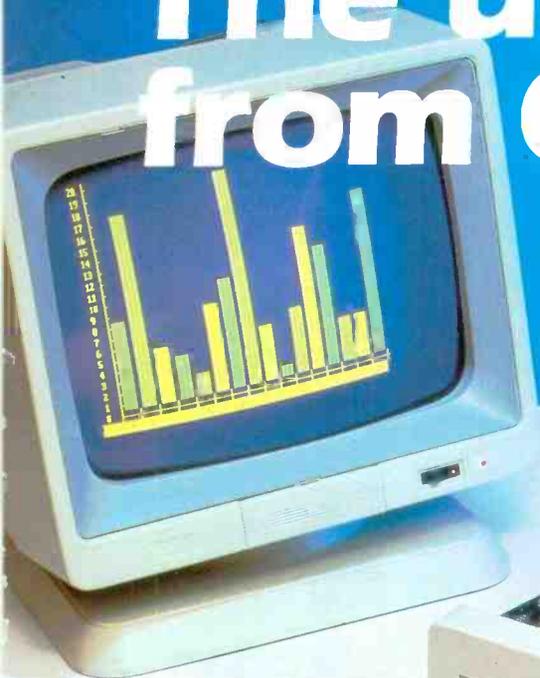
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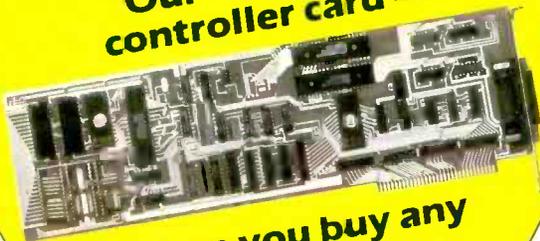
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