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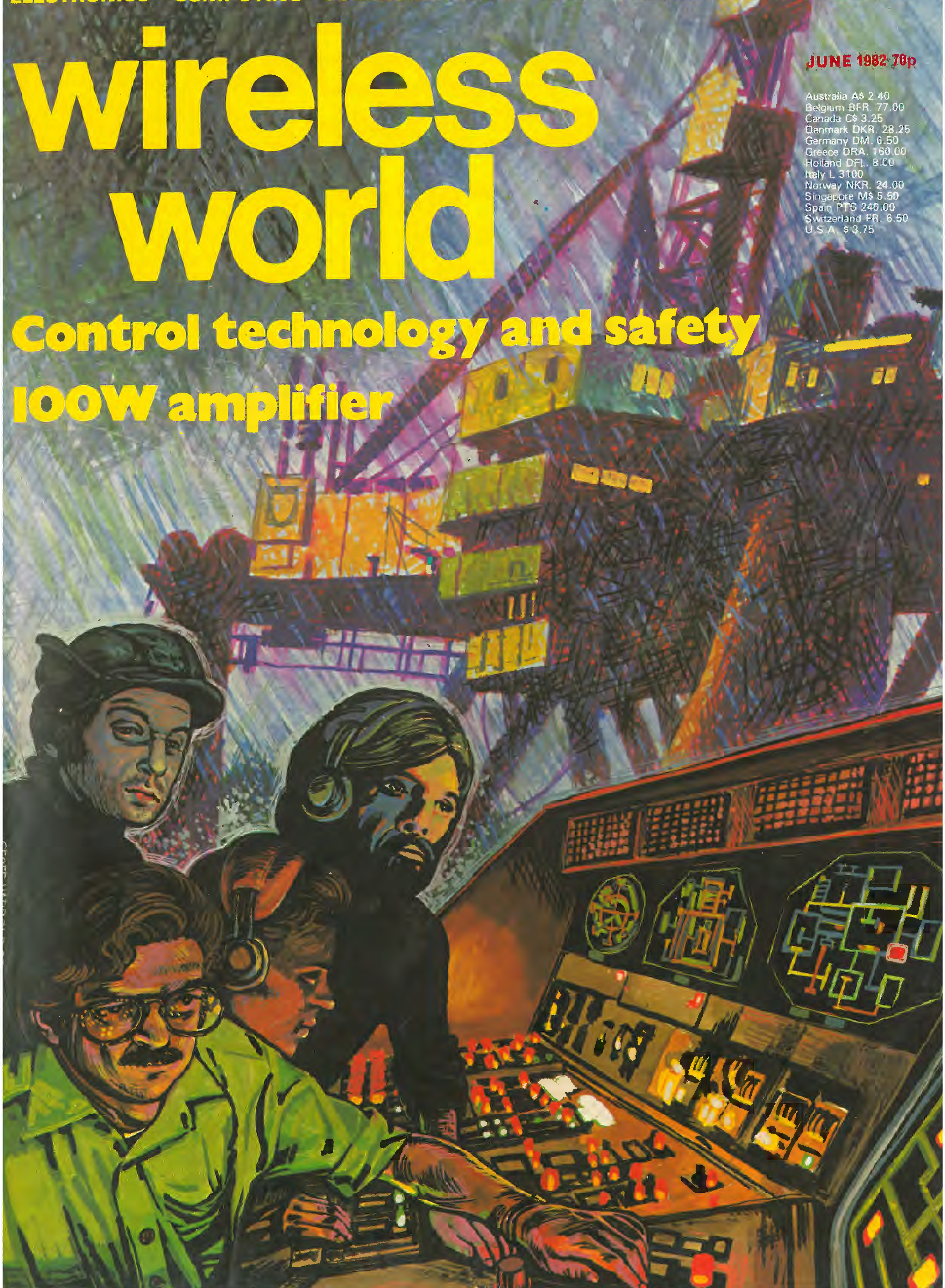

WIRELESS WORLD JUNE 1982 VOL 88 NO 1557

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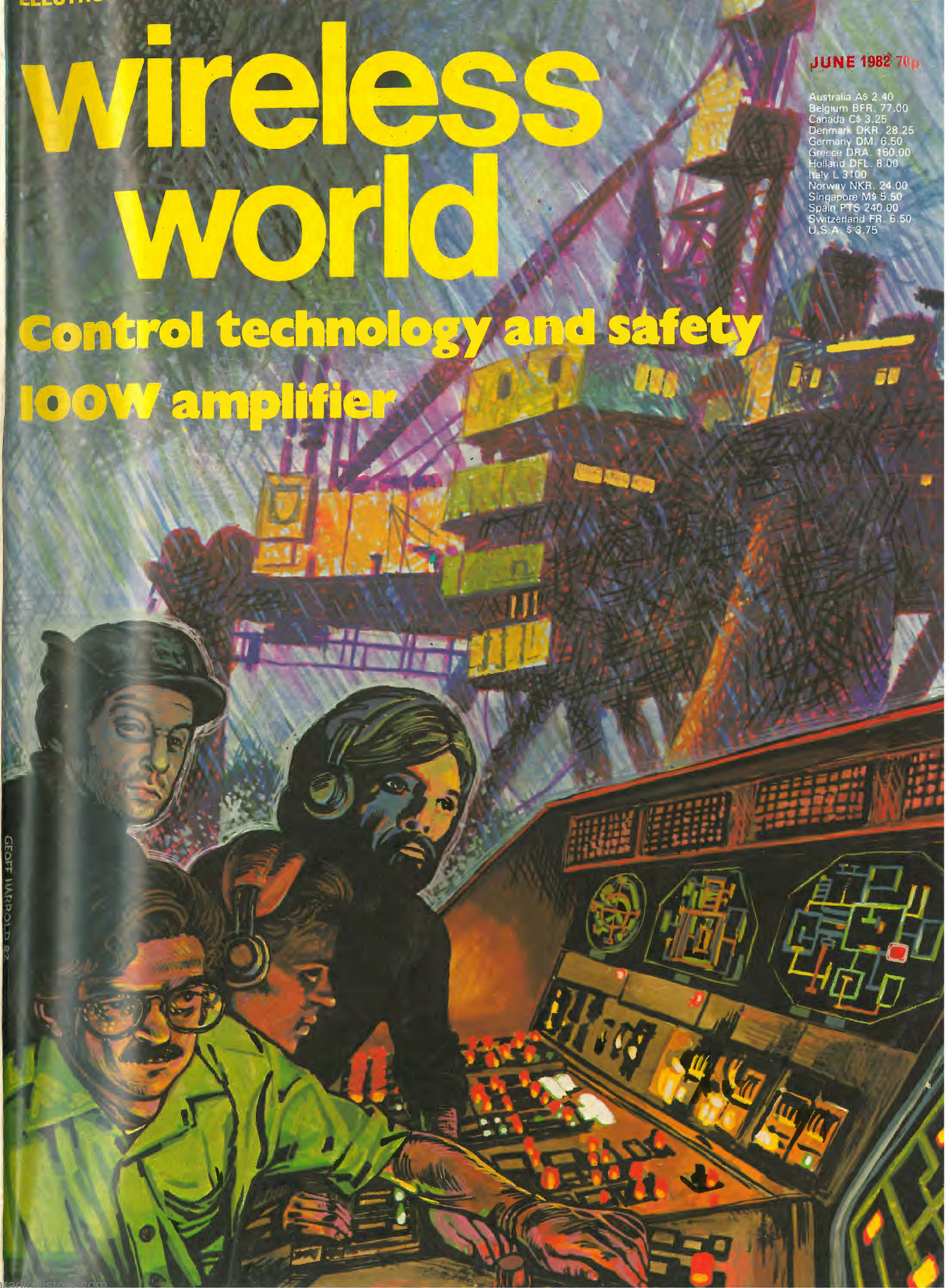

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JUNE 1982 Vol 88 No 1557

An impression of crisis control, created by Geoff Barrold, linked with R. E. Young's article on the subject in this issue.

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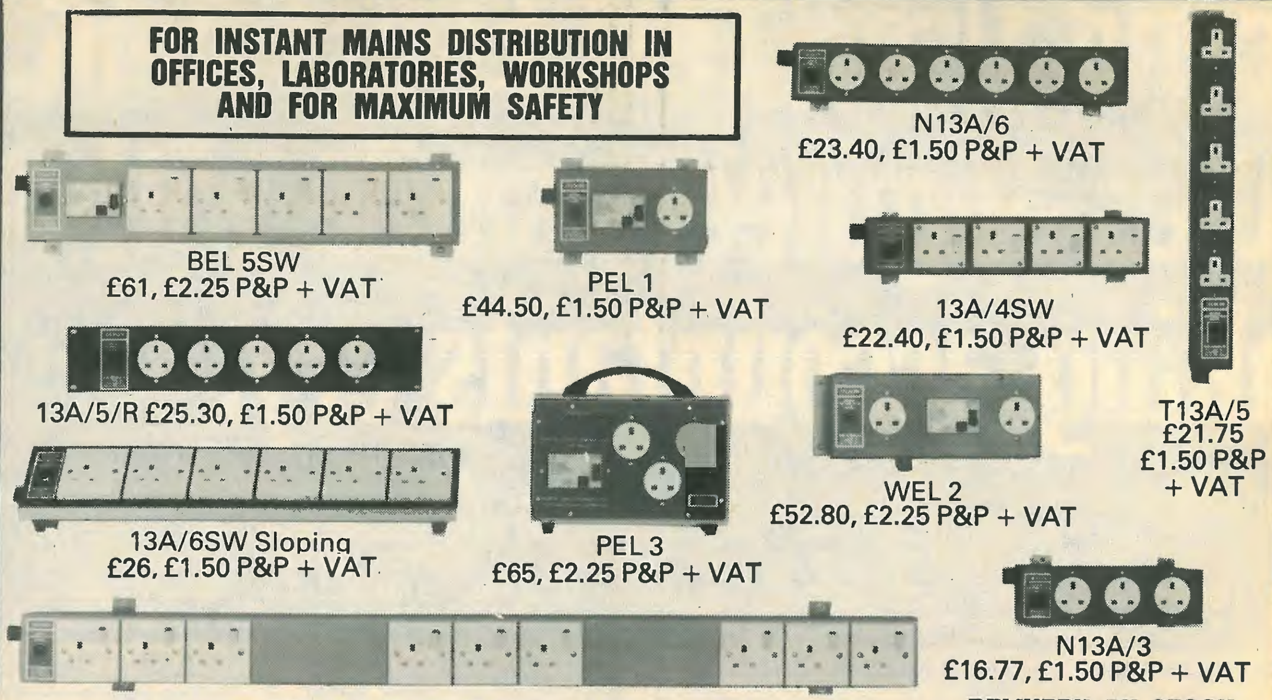

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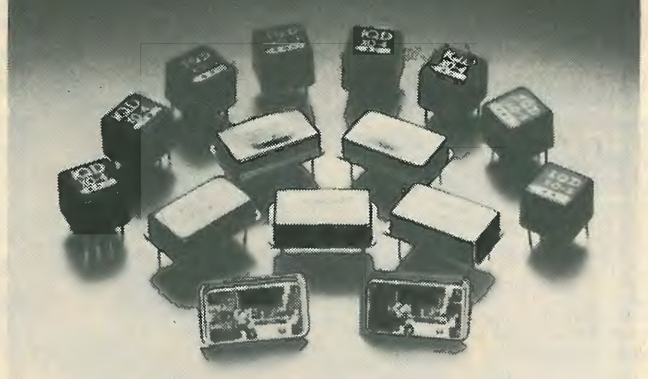

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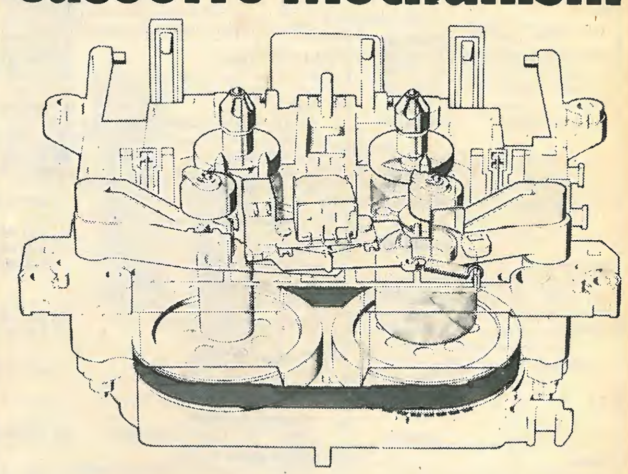

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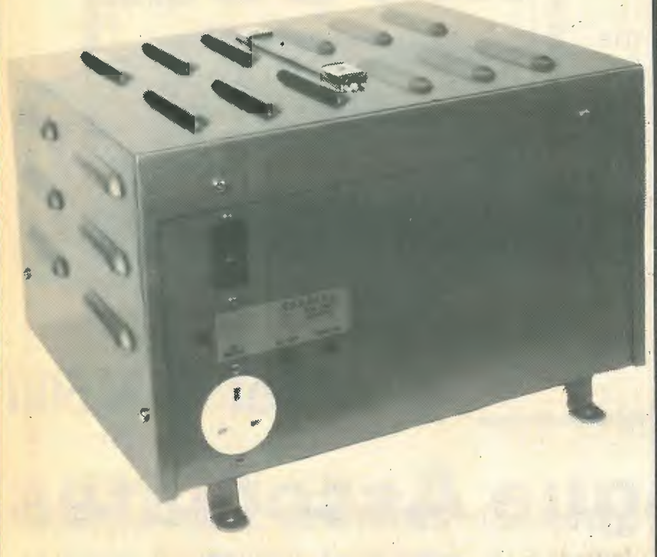

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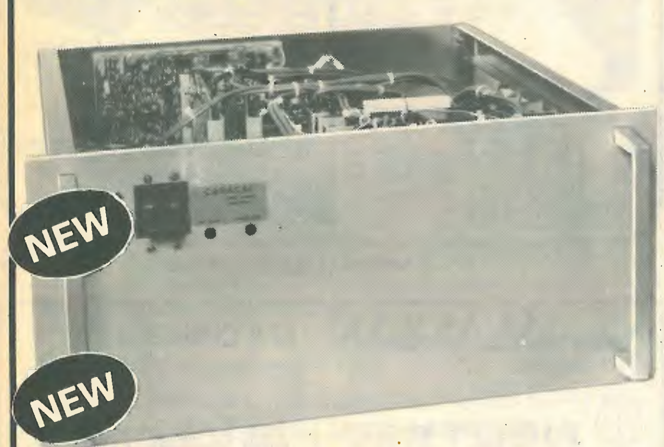

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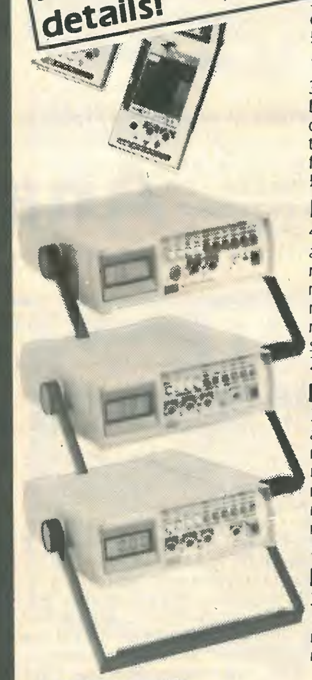

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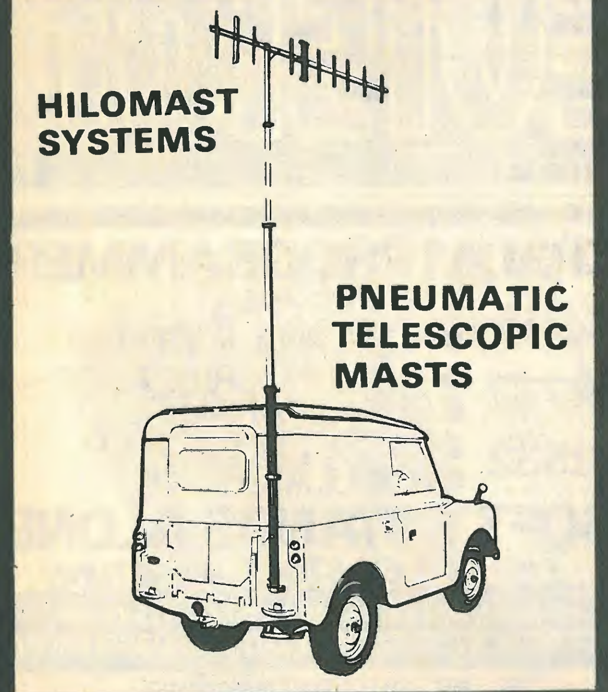

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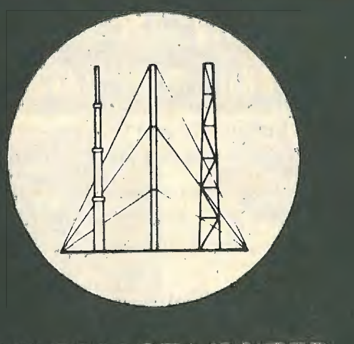

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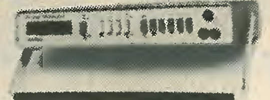

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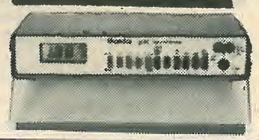

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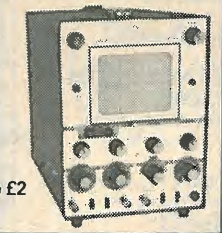

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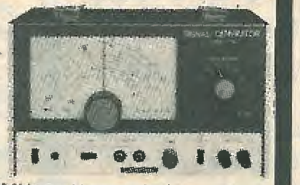

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EZ80	0.70
EZ81	0.70
GM4	5.90
GY501	1.30
GZ32	1.05
GZ33	4.20
GZ34	2.75
GZ37	3.95
KT66	6.30
	9.20*
KT88	8.95
	13.80*
MH4	2.50
ML6	2.50
MX10/01	21.50
N78	9.90
OA2	0.70
OB2	0.80
PABC80	0.60
PC85	0.75
PC86	0.95
PC88	0.95
PC97	1.50
PC900	1.15
PCC84	0.50
PCC89	0.85
PCC189	1.05
PCF80	0.80
PCF82	0.70
PCF84	0.75
PCF86	1.50
PCF87	0.50
PCF200	1.60
PCF201	1.65
PCF800	0.50
PCF801	1.75
PCF802	0.85
PCF805	2.45
PCF806	1.20
PCF808	2.75
PCH200	1.35
PCL81	0.75
PCL82	0.95
PCL84	0.90
PCL86	1.05
PCL805/85	1.25
PD500/510	4.30
PFL200	1.10
	2.80*
PL36	1.25
PL81	0.85
PL82	0.70
PL83	0.60
PL84	0.95
PL504	1.45
PL508	2.40
PL509	5.60
PL519	6.10
PL802	3.95
PY33	0.70
PY80	0.70
PY81/800	0.85
PY82	0.65
PY83	0.80
PY88	0.85
PY500A	2.10
PY809	6.45
PY801	0.80
QQV03/10	7.50
QQV03-20A	21.50
QQV03-25A	36.50
QQV06/40A	16.10
QV03-12	4.20
SC1/400	4.50
SC1/600	4 50

SP61	1.80
TT21	23.00
TT22	18.50
U25	1.15
U26	1.15
U27	1.15
U191	0.85
U281	0.70
U301	0.65
U600	11.50
U801	0.90
UBC41	1.30
UABC80	0.75
UAF42	1.20
UBF80	0.70
UBF89	0.70
UBL21	1.75
UCC84	0.85
UCC85	0.70
UCF80	1.30
UCH42	1.65
UCH81	0.75
UCL82	0.95
UF41	1.35
UF80	0.95
UF85	0.95
UL84	0.95
UM80	0.90
UM84	0.70
UY82	0.70
UY85	0.85
VR105/30	1.25
VR150/30	1.35
X66	0.95
X61M	1.70
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	82.90
Z759	19.00
Z749	0.75
Z800U	3.45
Z801U	3.75
Z803U	16.00
Z900T	2.45
1A3	0.85
1L4	0.50
1R5	0.60
1S4	0.45
1S5	0.45
1T4	0.45
1U4	0.80
1X2B	0.80
2D21	1.10
	1.85*
2K25	16.95
	24.50*
2X2	1.15
3A4	0.70
3A7	2.40
3D6	0.50
3D22	23.00
3E29	19.00
3S4	0.60
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5V4G	0.75
5Y3GT	0.95
5Z3	1.50
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5Z4GT	1.05
6/30L2	0.90
6AB7	0.70
6AC7	1.15
6AG5	0.60
6AH6	1.15
6AK5	0.65
6AK8	0.60
6AL5	0.60
6AL5W	0.85
6AM5	4.20
6AM6	1.50
6AN8A	2.50

6AQ4	3.40
6AQ5	1.00
6AQ5W	1.80
6AS6	1.15
6AT6	0.90
6AU6	0.80
6AV6	0.85
6AX4GT	1.30
6AX5GT	1.30
6BA6	0.55
6BE6	0.60
6BG6G	1.60
6BJ6	1.30
6BQ7A	0.85
6BR7	4.80
6BW6	6.20
6BW7	0.90
6C4	0.50
6C6	0.55
6CH6	8.20
6CL6	2.75
6CX8	3.80
6CY5	1.15
6D6	0.70
6F6	1.60
6F6GB	1.10
6F7	2.80
6F8G	0.85
6F12	1.50
6F14	1.15
6F15	1.30
6F17	3.20
6F23	0.75
6F24	1.75
6F33	10.50
6FH8	0.95
6GA8	4.20
6GH8A	0.95
6H6	1.60
6J4	1.35
6J4WA	2.00
6J5	2.35
6J5GT	0.90
6J6	0.65
6J6W	0.90
6JE6C	3.35
6JS6C	2.95
6K7	0.80
6KD6	3.35
6L6M	2.80
6L6G	2.50
6L6GC	2.10
6L6GT	1.25
6L7G	0.65
6L18	0.70
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6SG7	1.15
6SJ7	1.05
6SK7	0.95
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6SN7GT	0.80
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6SQ7	0.95
6V6G	1.50
6V6GT	0.95
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6X4WA	2.10
6X5GT	0.65
6Y6G	0.90
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6Z4	0.70
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9D6	2.90
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10P13	1.50
11E2	19.50
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12AT6	0.70
12AT7	0.65
12AU7	0.60

12AV6	0.95
12AX7	0.65
12BA6	0.90
12BE6	1.25
12BH7	1.95
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12C8	0.65
12E1	18.95
12J5GT	0.95
12K7GT	0.70
12K8GT	0.80
12Q7GT	0.60
12SC7	0.65
12SH7	0.65
12SJ7	0.85
12SQ7	1.45
12SQ7GT	0.85
12Y4	0.70
13D3	0.70
13D5	0.90
13D6	0.80
14S7	1.15
19AQ5	0.85
19G3	11.50
19G6	8.50
19H5	39.55
20D1	0.80
20F2	0.85
20E1	1.30
20P1	0.65
20P3	0.75
20P4	1.25
20P5	1.35
25L6GT	0.95
25Z4G	0.75
30C15	0.50
30C17	0.50
30C18	2.45
30F5	1.15
30FL2	1.40
30FL12	1.25
30FL14	2.15
30L15	1.10
30L17	1.10
30P12	1.15
30PL13	1.25
30PL14	2.45
35L6GT	1.40
35W4	0.80
35Z4GT	0.80
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50CD6G	1.35
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75C1	1.70
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78	0.95
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	2.55*
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	68.50*
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955	1.20
956	0.60
957	1.05
1625	1.80
1629	1.85
2051	2.90
5763	4.20
5842	7.50
5881	3.40
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6057	2.20
6060	1.95
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6065	3.20
6067	2.30
6080	5.30
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6550	6.60
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CV1597	
CV2116	
BR 189	
BR 179	
CV6131	
GMU 2	
TY4-500	
BK485/5552A	
MIL 5948/1754	
IC	
SN5402N	0.28
SN5410F	0.32
SN5470F	0.48
SN54196J	1.20
SN7407N	0.29
SN7408N	0.18
SN7445P	0.85
SN74453P	1.10
SN7453N	0.18
SN74L73N	0.38
SN7474N	0.30
SN7485N	0.85
SN74L85N	1.10
SN7491AN	0.32
SN74123N	0.42
DM74123N	0.38
SN15836N	0.28
cX4	0.95
SN76013N	1.80
SN76003N	1.80
SN76033N	1.35
MC6800P	5.80
MC6880OP	
	6.40
MC14511BA	
	2.20
B1702AL	3.30
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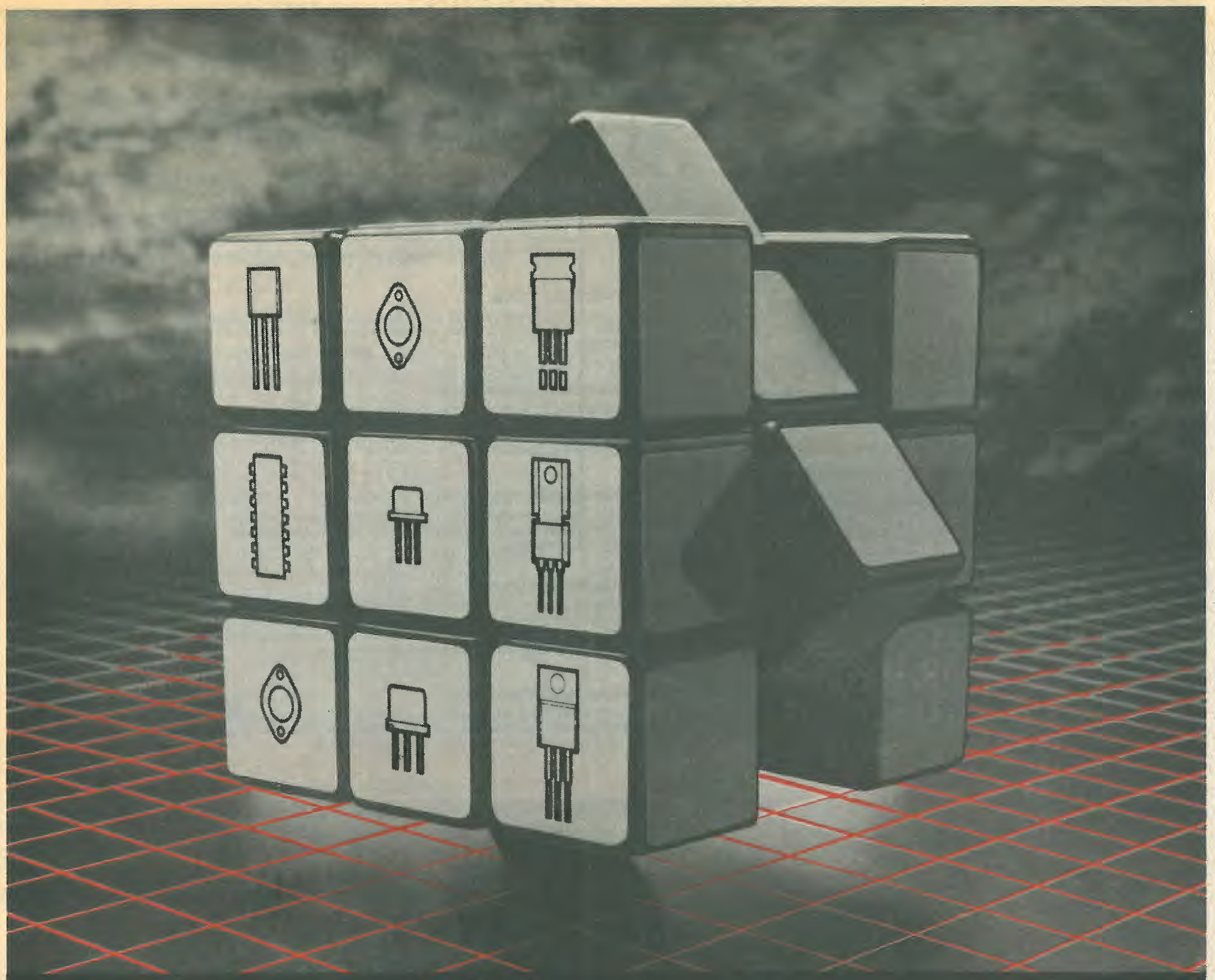

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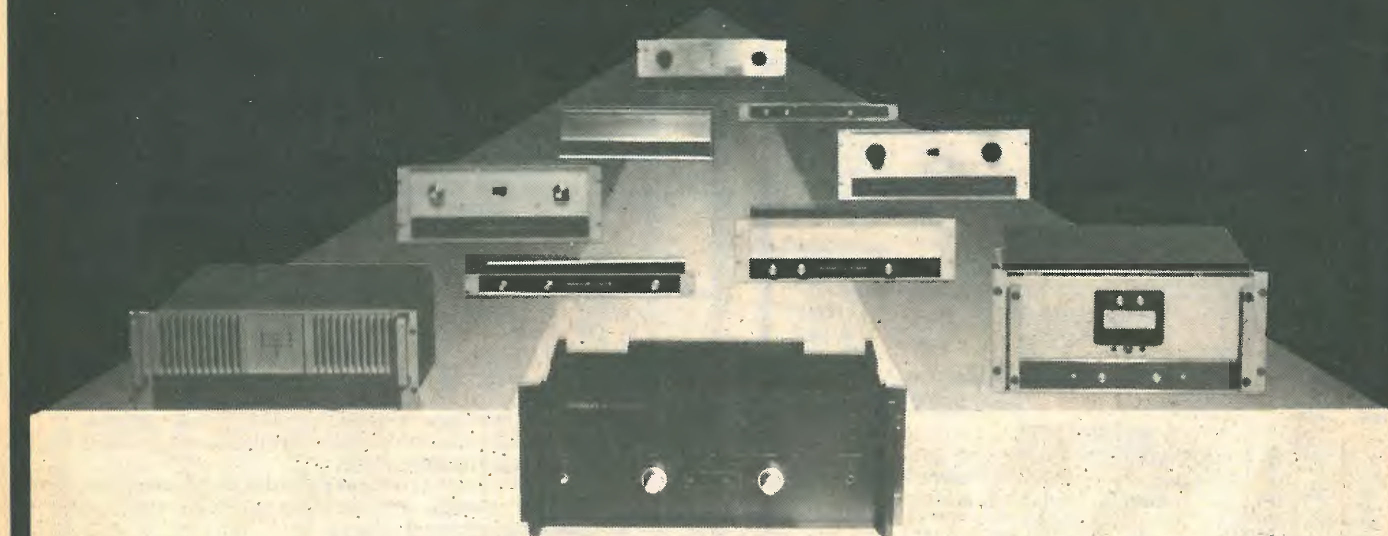

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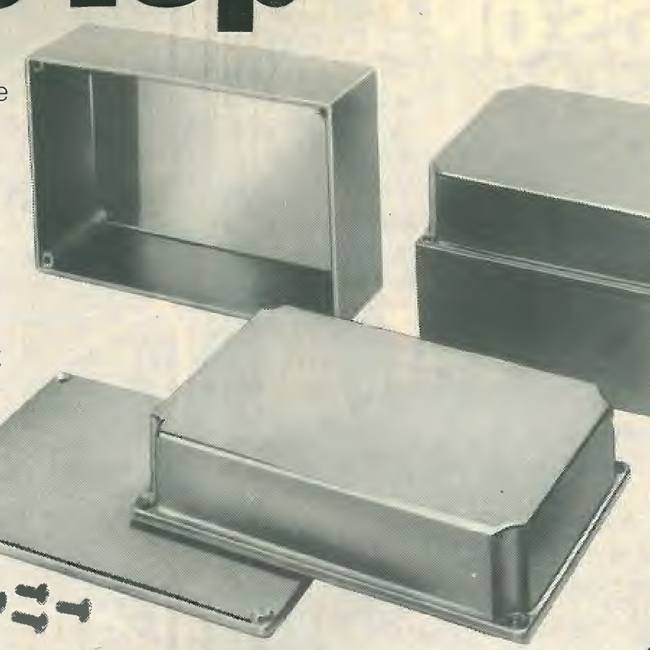

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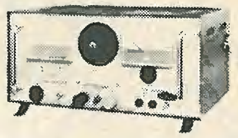

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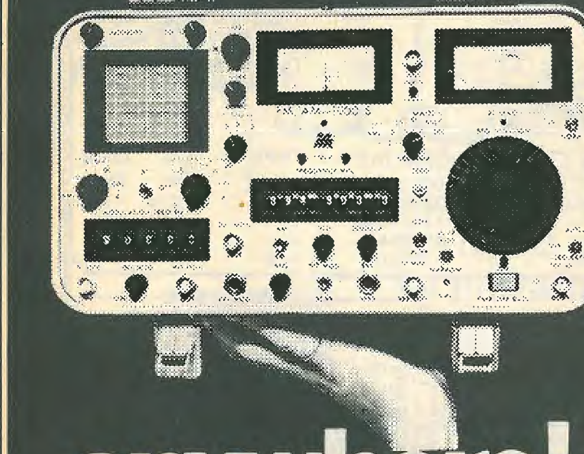

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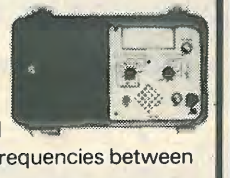

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ZN429E-8	2.10
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LM556CN	0.45
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LM741CN	0.14
LM747CN	0.64
LM748CN	0.34
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78L12	0.29
78L15	0.29
7905	0.55
7912	0.55

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79L12	0.59
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4014	0.58
4015	0.58
4016	0.25
4017	0.45
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4019	0.29
4020	0.58
4021	0.60
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4023	0.17
4024	0.35
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4034	1.55
4035	0.72
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4042	0.54
4043	0.59
4044	0.84
4045	1.65
4046	0.68
4047	0.68
4048	0.54
4049	0.26
4050	0.26
4051	0.59
4052	0.68
4053	0.59
4054	1.20
4055	1.20
4060	0.79
4063	0.95
4066	0.34
4068	0.17
4069	0.17
4070	0.17
4071	0.17
4072	0.17
4073	0.19
4075	0.17
4076	0.62
4077	0.22
4078	0.24
4081	0.14
4082	0.19
4085	0.63
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4093	0.39
4502	0.69
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4510	0.60
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4514	1.49
4515	1.49
4516	0.69
4518	0.40
4519	0.28
4520	0.69
4521	1.49
4522	1.20
4526	0.70
4527	0.69
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4532	0.85
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4556	0.44
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74LS01	0.11
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74LS03	0.12
74LS04	0.12
74LS05	0.13
74LS08	0.12
74LS09	0.12
74LS10	0.12
74LS11	0.12
74LS12	0.12
74LS13	0.22
74LS14	0.37
74LS15	0.12
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74LS27	0.12
74LS28	0.15
74LS30	0.12
74LS32	0.12
74LS33	0.16
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74LS40	0.12
74LS42	0.33
74LS47	0.39
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74LS49	0.59
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74LS138	0.33
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74LS151	0.39
74LS153	0.28
74LS155	0.38
74LS156	0.36
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74LS158	0.31
74LS160	0.39
74LS161	0.37
74LS162	0.39
74LS163	0.38
74LS164	0.45
74LS165	0.75

Device	Price
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74LS173	0.69
74LS174	0.45
74LS175	0.45
74LS181	1.28
74LS190	0.49
74LS191	0.49
74LS192	0.49
74LS193	0.45
74LS194	0.39
74LS195	0.39
74LS196	0.57
74LS197	0.59
74LS221	0.54
74LS240	0.69
74LS241	0.69
74LS242	0.78
74LS243	0.79
74LS244	0.60
74LS245	0.88
74LS247	0.80
74LS248	0.60
74LS249	0.63
74LS251	0.40
74LS253	0.39
74LS257	0.43
74LS258	0.39
74LS259	0.78
74LS261	1.95
74LS266	0.22
74LS273	0.70
74LS279	0.39
74LS283	0.44
74LS290	0.54
74LS293	0.45
74LS365	0.34
74LS366	0.36
74LS367	0.34
74LS368	0.35
74LS373	0.70
74LS374	0.70
74LS375	0.47
74LS377	0.70
74LS378	0.60
74LS379	0.64
74LS386	0.28
74LS390	0.54
74LS393	0.59
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14 pin	0.09
16 pin	0.09
18 pin	0.13
20 pin	0.14
22 pin	0.17
24 pin	0.19
28 pin	0.25
40 pin	0.29
LOW PROFILE GOLD	
8 pin	0.22
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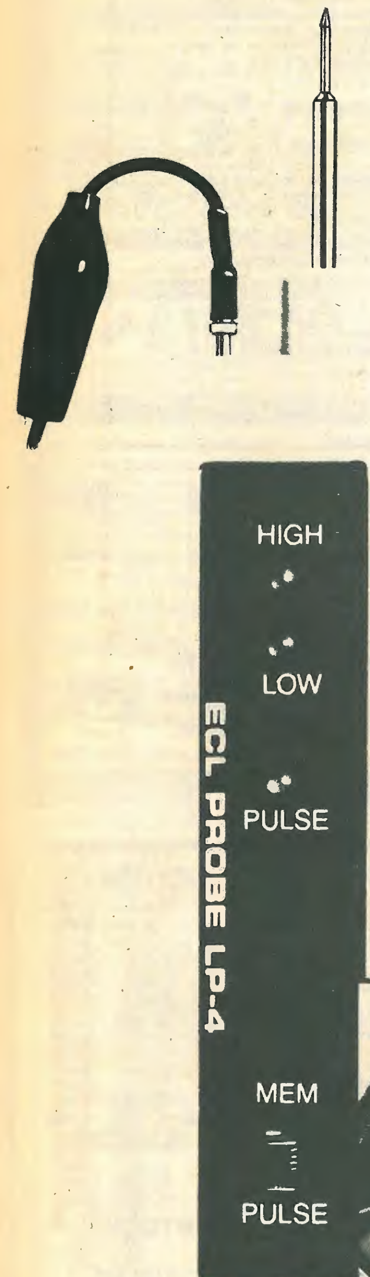

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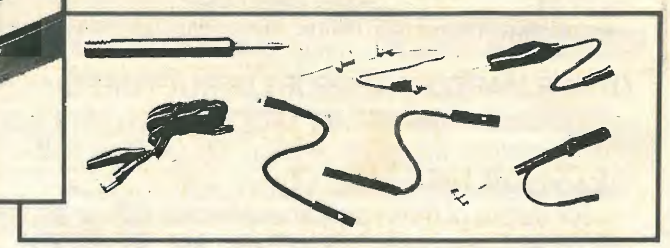

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Graven images, new style

When Lech Walesa, the Polish union leader, was asked by the military regime to appear on television and speak for them, he replied, according to a report, "You will have to cut my body into a thousand pieces first". This violently negative response speaks louder of the power of televison than all the strivings of young hopefuls in the entertainment business desperate to claw their way in front of the cameras. It also speaks loudly of the exclusive control of a whole broadcasting system by one authority. But state control only exemplifies, in a brutal way, the fact that television is a highly selective, and therefore exclusive, medium in the hands of anyone who uses it.

Consider first how reality is excluded by the physical details of the television set operating in the home. The images on the screen are produced by means of transmitted light from glowing phosphor sources, whereas we normally see most things in the world by reflected light. Sound is gratuitously present, though the real world often has to be interpreted in silence. The pictures are viewed in the banal surroundings of household objects and events: furious drama may work itself out on the screen immediately next to a plate of sandwiches and a sleeping cat. On this screen the people and objects are miniatures, and the convergence feedback of our binocular vision tells us that the images are not the same as the originals seen at a distance. Overall the effect is that of a brightly-lit toy theatre.

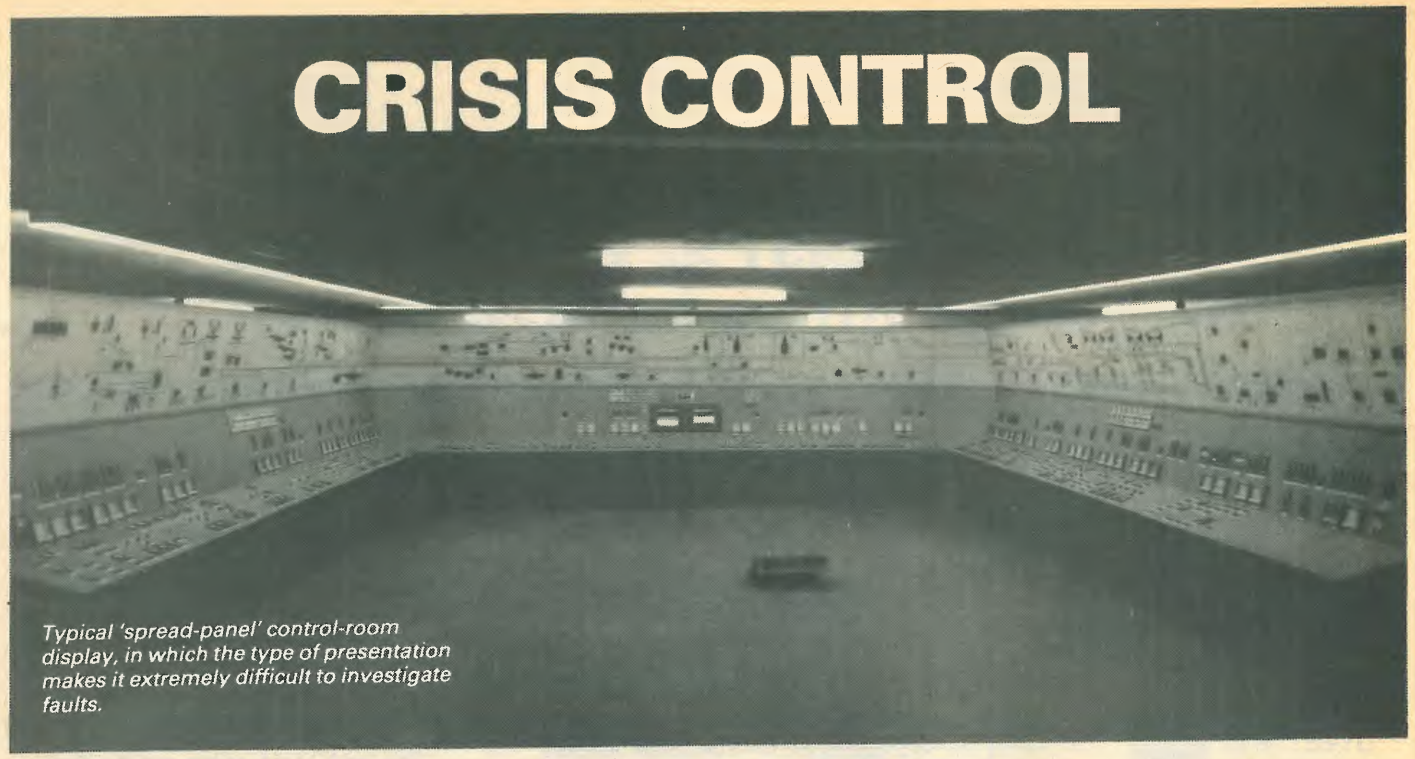
Within these physical limitations the deliberately selective processes of the broadcaster are operating. There is an old adage that newspapers can't tell you what to think but they can tell you what to think *about*. Selection and exclusion begin on television with the overall policy of the broadcaster and are implemented first in programme planning and then in the structure of individual programmes. In political discussions, for example, the agenda is set in advance: any controversy is held within previously determined limits which can be enforced by editing the recording.

It is in the "language" and conventions of the visual presentation, though, that selection and exclusion are seen most immediately. First the sequence of shots is selected. Then each shot is structured by a variety of artifices: by choice of duration, framing, lighting, the angle and distance from which it is taken and so on. The precise meaning or mood which the viewer is expected to attach to the resulting images is fixed by the use of sound. Speech and music are very specific.

One result of this concentration of selective processes is a distancing or alienation of the viewer from normal experience. Jonathan Miller, the television producer and man of many other parts, describes it well in his book on Marshall McLuhan. The images which television presents, he says, are "curiously dissociated from all other senses. The viewer sits watching them all in the drab comfort of his own home, cut off from the pain, heat and smell of what is actually going on. Even the sound is artificial . . . . All these effects serve to distance the viewer from the scenes which he is watching, and eventually he falls into the unconscious belief that the events which happen on tv are going on in some unbelievably remote theatre of human activity. The alienating effect is magnified by the fact that the tv screen reduces all images to the same visual quality. Atrocity and entertainment alternate with one another on the same rectangle of bulging glass. Comedy and politics merge into one continuous ribbon of transmission."

Perhaps what Miller describes is the result of the most powerful selective process of all – in the mind of the viewer. If we seem to go into an hypnotic trance when watching television it is because we have involuntarily agreed to suspend judgement on the reality of what is appearing on the screen. We have colluded with the modern graver of images in allowing ourselves to be transported by his art, shutting out our normal, hard-headed, sceptical selves. In this irrational state of suspended belief the contrived images and sounds become more 'real' to us than the real world. Temporarily we can be persuaded of anything. This mental state, and how to produce it, is well understood by television playwrights, propagandists and makers of commercials. The unperceived surrender of the mind is where the real power of the medium lies.

CRISIS CONTROL



Typical 'spread-panel' control-room display, in which the type of presentation makes it extremely difficult to investigate faults.

There have been occasions in large control complexes when the amount of information presented to controllers has been so vast and, possibly, suspect, that operators have found it impossible to take in. The author presents his views on methods of training for such moments, and on methods of measurement and presentation to reduce the number of malfunctions and the level of stress on operators.

by R. E. Young,
B.Sc. (Eng.), M.R.Ae.S., F.I.E.E.

The Senior Shift Engineer was more bewildered than horrified when he answered the "panic" call to the control room and saw his staff rushing backwards and forwards along the almost infinite assembly of panels, obviously not knowing in the least what was going on, and deafened by the ceaseless and ever-growing noise of the alarms. Full horror was to come almost immediately when he realised that hordes of men were surging into the room from all over the huge plant, summoned in a last desperate attempt to find out which, if any, of the meters and indicators were "telling the truth".

Fictitious? – yes. Exaggerated? – perhaps. Irrelevant? – no, for here is an extract from an account of such a catastrophic 'Incident': " . . . the operators were misled . . . because of a false indicator . . . and by signs of an apparent excess of water . . . All the events were accompanied by an unhelpful cacophony of about 100 alarms sounding off, distracting the operators as they faced panel upon panel of red, green and amber lights and dials indicating entirely unexpected combinations of conditions, and as they tried to grasp the significance of the mystifying changes that were happening . . .".

This description of a 'developed' crisis situation is taken from the masterly summary by Sir Alan Cotterell[1] of what tends to be thought of as the sole example of the large-scale industrial disaster – the Three Mile Island accident. In actual fact, this near-catastrophic type of accident, with its far-reaching consequences, has been experienced in many industrialized countries in the world, but it was not until "Three Mile Island" had occurred that public awareness of the dangers involved became much more apparent. At the same time steps to introduce improved safety precautions were being taken at both national and international levels: for example, the EEC produced a Directive in 1980 aimed at avoiding major industrial disasters, and the Atomic Energy Agency called an international conference in the same year to harmonize nuclear reactor safety standards worldwide.

Both technical and political interests now realize the need to provide additional, entirely new, safeguards in the operational design of large industrial complexes. The implied requirement for major changes in the 'total' approach to high-risk installations in particular becomes clear, and has been brought out in what amounts to public debate, largely critical in nature.

This critical element has been especially marked in relation to "Three Mile Island", where various forms of human error were blamed for the accident; and with strong attacks being made, in effect, not only on the operators themselves but on their selection and training. It is only when one builds up the background to this Incident in the light of 'crisis control' and allied considerations[2] that an entirely different picture emerges, which can be largely summarized by the statement that *within the circumstances in which they were operating*, the control engineers at Three Mile Island could not have achieved more than they did.

Before going into the way in which measurement and other information was presented to them – quite inadequately – and into the whole question of the 'goodness' (integrity) of this information, it is necessary to look at the state in which these engineers found themselves. It was, in a sense utterly fluid and had features which completely defied explanation. In the words of the Cottrell acount: " . . . (they were) faced suddenly with a totally strange combination of events . . ." preceded by " . . . (being) misled . . . *because of a false indicator* . . . and by signs of an *apparent* excess of water . . ."

This is the classic example of the final – full emergency – stage of an incident carrying high potential risk, where the control staff come to the chilling realisation that they are being given wrong information *somewhere* and that they have no possible means of finding out where. Furthermore, as far as their control of the plant is concerned, they are completely out of contact with it; and, added to this, all their sources of control information have to be treated as totally suspect.

Thus it has to be recognized that in such circumstances the stress to which the engineers are subjected can only be described as extreme in the full sense of the word. When it is remembered that the conditions producing the emergency are entirely unforeseen, it becomes clear that it is virtually impossible to predict how any individual will behave in circumstances where "there is nothing to get hold of".

Nevertheless, there are two related areas where experience can be quoted which bears very positively on this general question of behaviour under stress.

The first of these areas is in the field of 'control room' type operations carried out under wartime aerial bombing attacks. Secondly, it is not usually realized that in World War II, 'ultimate crisis control' was carried out under these hostile conditions. This was with extensive integrated broadcasting networks under central master control and with large radar command systems, which became, in effect, the equivalent of modern high-risk installations.

As will be treated in more detail in Part 2 of this article, observation has led to a number of conclusions on unexpected operational stress and methods of preparing people for it.

The most significant of these conclusions was that under such stress, one or other of two conditions – A or B – was reached; assuming, of course, that a panic state was not developed, where 'panic' is differentiated from conditions A and B by being "unreasoning" and "sudden" in nature.

Briefly, condition A is one in which – despite the severe stress – the person thinks and takes action exactly (as far as the outside world is concerned) as before the incident occurred. As will be indicated in Part 2, it is suggested that methods of training and preparation for this condition should be largely 'subliminal' in nature, i.e. without the training mechanism being really obvious.

One of the main differences between conditions A and B is the virtual disappearance in B of the ability to take action, i.e. all initiative appears to be lost. In most cases "leadership by example" can restore it, and those in this condition are "near the surface" from this point of view. In practical terms, they appear not to be in a state of shock, as it is generally understood.

However, it is in the almost invariable change in facial expression to the blank look of mental handicap, familiar to those working in that field, which is most characteristic in Condition B. The 'mentally handicapped look' and other relevant aspects of that complaint will be discussed later in the context of a reserch programme which has been in progress for a number of years. This programme, conducted essentially on a voluntary basis, has nevertheless attracted an increasing amount of interest, particularly during the last two years; and has enabled major correlation to be established between observations made in one field and 'practice' in the other. Examples of such correlation include 'thinking fatigue' which is relevant to the design of data presentation and other equipment for crisis control, and which appears as a major problem with the mentally handicapped.

Thinking fatigue may be defined as fatigue produced by 'multi-channel' mental activity, i.e. where several ideas (lines of thought) have to be carried simultaneously and coordinated. Instances are afforded by an author writing a book with a complex, interacting, plot, or by a pilot landing an aircraft under adverse conditions with all the controls having to be operated simultaneously and in relation to a number of rapidly changing events. Another instance was offered by the editor of *Wireless World* last year (1981) as "An engineer preparing for or taking part in a highly technical and diversified international conference".

It may be noted that the direction of 'spin-off flow' in this area has been largely from mental handicap to operational design in that the effects of 'thinking fatigue' can actually be seen with the former, whereas it tends to be regarded as almost hypothetical in the ordinary world. In contrast, in the other example to be quoted at this point, viz. data marshalling (to be described in the next section), the flow has been mainly from the technical side towards mental handicap. This flow tends to become two-way as time goes on; it develops to benefit both sides, and then – in many cases – extends as an interchange of information to other areas, such as geriatric care.

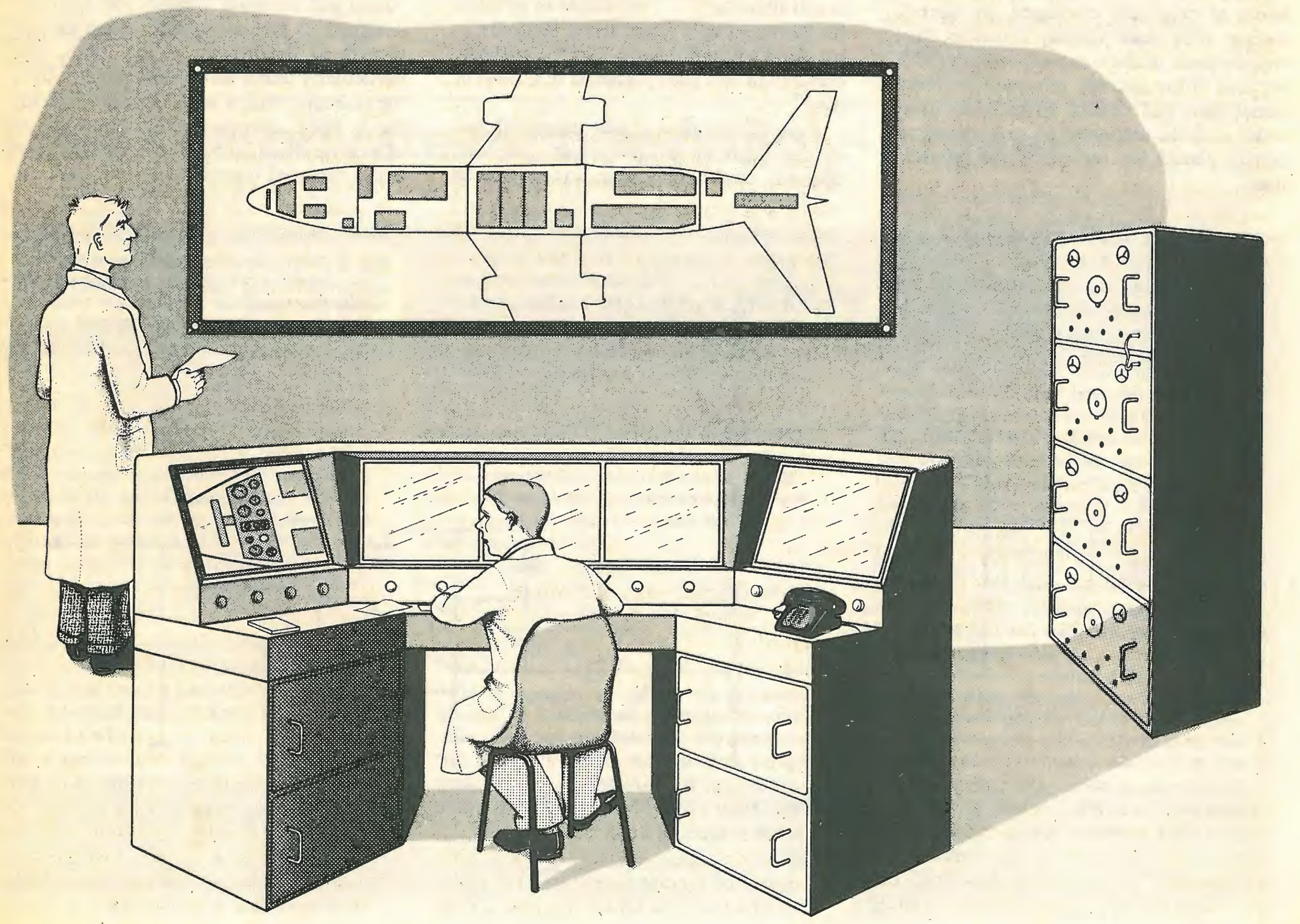
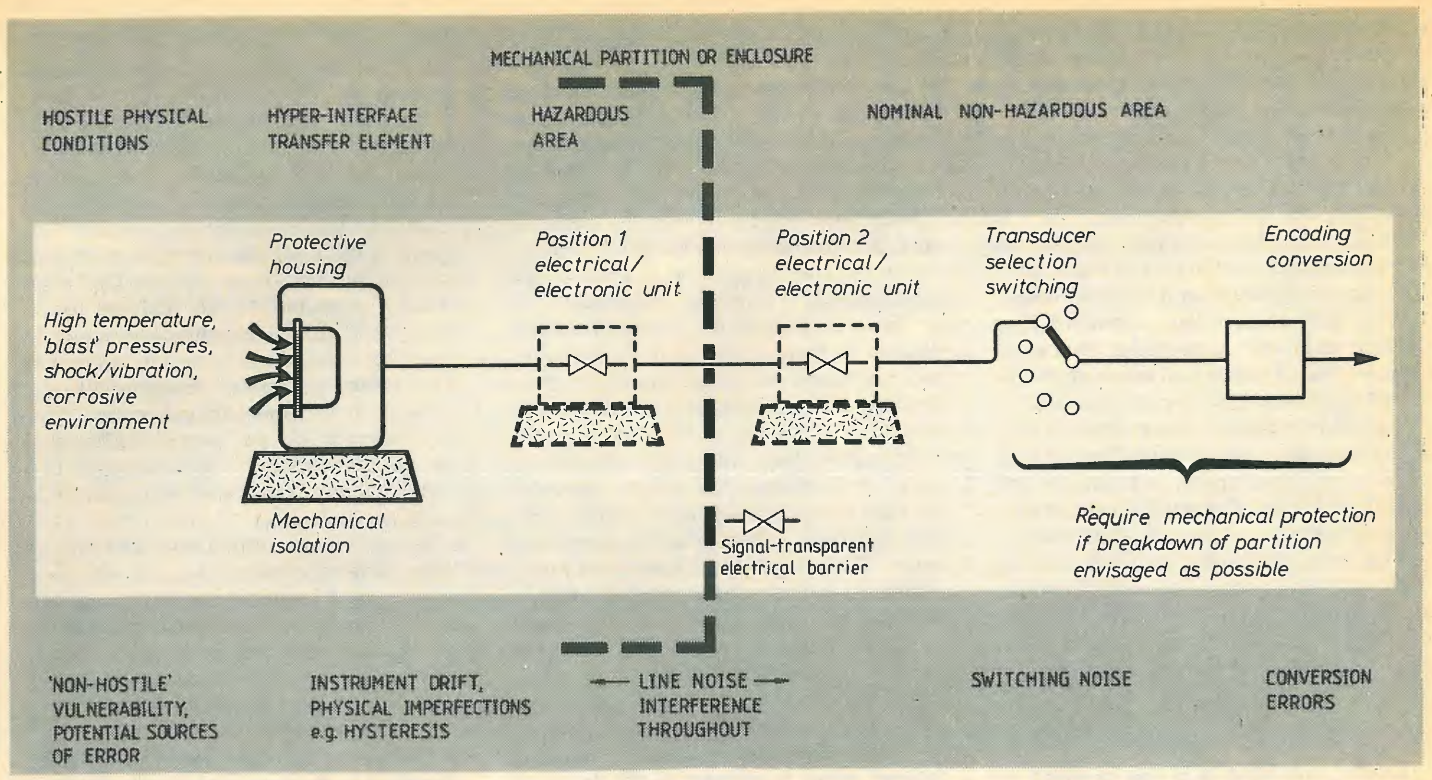


***Fig. 1.** 'Data-marshalling' type of control-room, proposed for aircraft testing. Overall situation on wall diagram is broken down on operator's screens.*



Data marshalling

This concept, first introduced by the writer in 1960[3] mainly in connexion with the testing of complex aerospace vehicles and aircraft generally, may be described as the separation, streaming and systematic presentation of "masses of data".

Figure 1 shows the control room installation as originally envisaged for aircraft testing. The main display consisted of a wall diagram of the complete aircraft, with sections of the aircraft, selected on a functional basis and broken down to the next order of detail displayed on corresponding panels placed on the operator's console itself.

Fig. 2. Transducer instrumentation chain in a totally hostile environment.

The design of the system was aimed at ensuring that "faults at once strike the attention of the operator, and their location and magnitude are clearly apparent to him". Also as a statement of an ideal which is still aimed at ". . . the action to be taken (in the event of a fault) arises naturally in his mind without the need for the analysis, correlation and interpretation of a mass of data".

Even at that time it was suggested that all the displays, and particularly the wall diagram, might be by large-screen (cinema size) television. It is not generally known that large-screen television was in operation in certain London cinemas before World War II, and in a more advanced form, using high-voltage projection tubes, was demonstrated to the Press in 1948.

In non-aerospace, such as process-plant, applications, the wall diagram becomes an 'alarm and situation' mimic-type diagram, designed to give an overall instantaneous picture of the system at any time, and particularly under alarm conditions. Thus, for example, with a large offshore oilfield, the A. and S. diagram would show a group of well-heads with their main pipeline runs and a pictorial representation of (say) the block of plant at the central platform. Working within a conventional computer-based (central processor) type of control system with visual display unit presentation, alarms shown at critical points on the A. and S. diagram give a combined 'alerting' and 'location' signal to the control engineer should a fault develop.

The resultant system is in sharp contrast with the 'spread-panel' type as shown in the heading illustration, and as described for Three Mile Island. Thus, even if the control engineers had been able to trust all their information during that incident, the presentation of this information was such as to make their task of investigating the fault quite impossible. With this in mind, it would seem pertinent to add that they would appear to have remained in operational Condition A throughout, something which should not be overlooked.

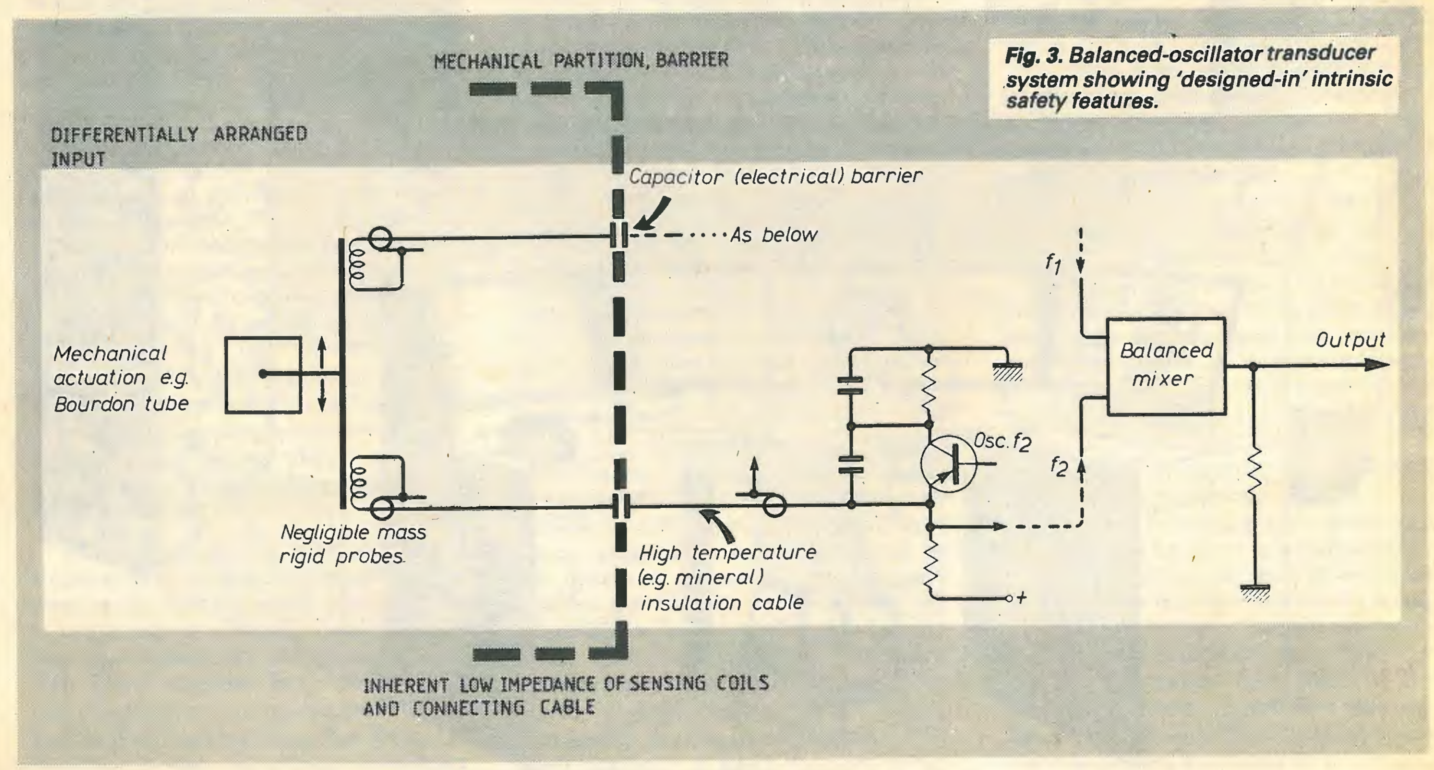


Fig. 3. Balanced-oscillator transducer system showing 'designed-in' intrinsic safety features.

'Telling the truth'

As has been stressed already, the most disturbing influence that a control engineer can experience during a dangerous break-down is the realization that instruments are not 'telling the truth'. It is manifestly impossible to ensure that information will never be false (equivalent to infinite reliability), as, for example, when a position transducer is mounted on the 'drive' side of a break in an actuator shaft. Nevertheless, design for full 'crisis management' demands that every effort must be made to give the maximum integrity to every element in the control system. The process of selecting these elements and determining their ultimate form must be carried out jointly between user and control-system designer[2] if the maximum protection is to be given to the whole installation.

The provision of crisis-level protection is obviously a big subject; as on illustration of the principles involved, consider of transducers as key elements in the overall (combined) main plant and control-system complex. They occupy the most exposed position in the instrumentation chain since they are vulnerable both to severe mechanical and electrical disturbance. Apart from various forms of mechanical protection determined by environmental engineering techniques, reducing vulnerability is fundamentally a matter of the design and construction of the transducer itself and (in the mechanical context) of its physical environment. Instrumentation transducer design is far more involved than is generally thought: this is particularly true at the 'hyper-interface' where the parameter/electrical signal transfer takes place; and it is at this interface that the actual measuring process may be said to take place.

Consequently, it is vital that, under all conditions, the transducers should maintain the integrity of this transfer (i.e. that the transducer itself should 'tell the truth'). The system diagram of Fig. 2 shows how protection against totally hostile physical conditions can be 'designed-in' to the instrumentation chain system; and brings out the extra vulnerability of the transducer in meeting the 'real' hostile conditions such as blast pressures at the hyper-interface. The balanced-oscillator (dual-chain) transducer system of Fig. 3 has been designed from the beginning to exhibit minimum error under abnormal and 'catastrophic' conditions.

This family of transducers have a 'parameter-dependent' frequency output derived as a difference (beat) frequency between two matched oscillators, themselves tuned by high (electrical) resistance moving probes. These probes move differentially within their sensing coils at the input to the dual-chain system, which is terminated with a balanced mixer. The basic symmetry of the mechanical-electrical combination and its inherent balance provide a high degree of compensation for ambient changes.

The greatest attraction of this configuration, however, arises with extreme environments, where the series-resonance operation of the Clapp oscillators which are used, makes it possible to instal the sensing coil and probe remotely from the less 'hard' electronic unit. This is by virtue of the low-impedance (coaxial cable) connexion, which can be used between the two with the series-resonant oscillators. From the diagram it will be seen how this isolation as part of the intrinsic safety precautions built-in to the system, is an advantage. Thus, with the 'barrier' protection shown, the coil can be situated in the hazardous area and can be made effectively neutral, with little risk of energy at this point to initiating a spark.

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To be concluded □

Action on private mobile radio — urgent!

Unless the Government takes action quickly to introduce repeaters, trunking and cellular systems, along with other measures to improve the effectiveness of private mobile radio, commercial enterprises, emergency services and public bodies will become much less efficient than their foreign counterparts and the UK radio industry will suffer.

This is the burden of a report, prepared by PACTEL (PA Computers and Telecommunications) on behalf of the Electronic Engineering Association, which is entitled 'Mobile radio — the case for urgent action'*. It is the latest in a series of attempts to persuade the Government, who control the use of the spectrum, to allow mobile radio users access to a share of the 30MHz-1GHz spectrum more in keeping with potential usage than the nominal 8% they now have (PACTEL say this is really 6.5% because of interference, terrain limitations and official restrictions).

Among the consequences of the failure to secure a more equitable share are high costs, delays, pollution and a poor service to customers of the haulage industry, and a lower level of safety and security.

The depressed use of m.r. in the U.K. (around 0.5% against 2.7% in the US) is, says PACTEL, no help at all to the UK radio industry, in contrast to the state of affairs in Japan and the USA, where a healthy home demand acts as a spur to development and production and results in efficient systems and lower costs.

According to the report, five actions must be taken now. Another 30-35 MHz of spectrum must be released for m.r. use — little enough, since this would bring the total to a mere 10% of the usable 30MHz-1GHz band. Secondly, the Government should take a close look at the amount of spectrum inhabited by the military and by broadcasters. PACTEL says "there is little doubt" that the MoD keeps some of its allocations on ice, unused in peacetime, simply because they might be needed some day, in spite of claims by m.r. lobbyists that, if they were released, they could be taken back in emergency in a matter of hours or even minutes. In the case of broadcasting, the report is critical of the efficiency with which the spectrum is used and says that it could be improved upon without any deterioration in service.

Thirdly, three types of system should, says PACTEL, be promoted to give wide-ranging, low-cost, flexible and spectrum-efficient services: community repeaters, trunked systems and cellular systems.

In a repeater, a common base station, located on high ground, gives a wide coverage, many rented mobiles sharing one channel: a 'channel busy' indicator helps to avoid a user hearing messages not meant for him. The current position is that not enough channels are available for this type of system.

Trunked systems provide a number of channels to each user and are therefore frequency-efficient: the operator uses a multi-channel radio which is tuned automatically to a free channel. For this purpose, blocks of channels are needed — too few are currently available for widespread adoption of this kind of system. More channels are needed.

The third type, the cellular system, relies on dividing the area into cells, each of which is treated as a trunked system. By geographically separating cells which use the same block of frequencies, interference is avoided and the usage of the channel is increased by up to 1200%. Blocks of channels are needed for trunking. Once again, not enough are available (56 instead of a required 300).

The fourth recommendation is that the Government actively promote the use of private mobile radio instead of restricting it.

Finally, says the report, the delay of between three and six months in the issue of licences should be reduced. Otherwise, illegal installations will multiply and make spectrum management virtually impossible.

As John Carlson, chairman of the EEA's Mobile Radio Committee, points out, "if even the manufacturers of private mobile radio equipment put forward the difficulties to their customers, the situation must be bad". And Ray Northcott of PACTEL sums up the current p.m.r. scene by saying "Other countries have taken a conscious decision to encourage p.m.r. — the UK hasn't". □

**Available at £3.00 from The Director, The Electronic Engineering Association, Leicester House, 8 Leicester Street, London WC2H 7BL.*

80-100 WATT MOSFET AUDIO AMPLIFIER

A three-part article on the design and construction of a modern, high-power amplifier begins with a description of the problems of amplifier design in relation to the characteristics of 'vertical' power mosfets. A matching, modular preamplifier design will follow.

by John L. Linsley Hood

The problem with designing audio amplifiers is that there are a number of design requirements which are impossible to satisfy completely: such things as freedom from harmonic and intermodulation distortion; independence, in terms of distortion or transient response, on the nature of the load reactance; freedom from spurious (amplifier-generated) signals over the whole range of signal inputs and likely load characteristics; rapid settling time and freedom from 'hang-up' on step-input or overload, particularly under reactive load conditions; and complete absence of input-signal or load-induced instability.

Not only are these requirements impossible to achieve absolutely, but the work needed to improve one of these may simultaneously bring about a worsening in other respects, so part of the task of the designer is to choose, within the appropriate limits of cost and complexity, between conflicting possibilities and requirements. No two designers (or their commercial or advertising managers) are likely to come to the same balance of compromises in these respects, and this leads to subtle differences in the tonal characteristics of the designs.

A characteristic of commercial trends in the last twenty years, which I view with regret, is an overwhelming concentration on the attainment of very low harmonic-distortion figures over the whole of the audible spectrum, to the extent that many modern commercial designs attain steady-state t.h.d. figures fifty or more times better than possible, under any conditions, from the signal sources which feed the amplifiers. A similar amount of effort is expended commercially in achieving very high signal-to-noise ratios – which would be valuable if it were matched in the handling of programme material by the programme producers.

The reason for this commercial interest is a simple one. the major emphasis in most equipment reviews is placed on t.h.d. and s/n ratio, coupled with, in the case of power amplifiers, power output in watts/pound (sterling) or, occasionally, watts/pound (avoirdupois). This trend would be wholly praiseworthy if it could be achieved without impairment in other desirable characteristics of the equipment: unfortunately, it cannot. If one wants some quality very good, one must accept some others relatively bad! If only one knew which ones were important to the listener, this choice would be easy, but one doesn't. Quite a lot of work has been done in the field of psycho-acoustics to try to characterize the effects on the listener of specific electrical defects, but this work is far from complete and impaired, from the point of view of the designer, by the omission of most of the minor performance defects practical amplifier designs are heir to.

Nevertheless, a predictable result of the accumulation of experimental findings on acoustic effects, coupled with a greater awareness on the part of designers of the existence of residual performance shortcomings, is that there is a keen interest in new developments in components and circuit techniques, as a possible route to improved performance.

Of these new component developments, one of the most interesting, in the field of active devices, has been the growing availability of rugged, reliable and reasonably priced power mosfets (metal-oxide-silicon field-effect transistors). These devices have a very much better h.f. response – almost embarrassingly so – than the normal audio power transistor, and allow a considerable extra freedom in solving h.f. loop-stability problems, where some compromise must always be reached in a feedback amplifier design between the conflicting requirements of gain (or phase) margins and the need to retain a high loop gain at the upper end of the audio spectrum to achieve a high degree of steady-state linearity. In addition, these devices are almost completely free of the charge-storage effects found in junction transistors, which tend to impair complex-signals transfer.

Unfortunately, power mosfets have electrical characteristics and circuit requirements which are very different from those of the junction power transistor, so that they cannot be used as a direct replacement for junction transistors in existing designs. One must reappraise the circuitry.

Power mosfet

Insulated-gate field-effect transistors, of the type shown in outline in Fig. 1(a), and which operate by means of a mobile layer of charge induced in an otherwise non-conducting region of a semiconductor, have been known and used in small-signal applications for many years – particularly in v.h.f. circuitry, where their very fast response times are of great value. However, the conducting path in these devices is, by the nature of their method of construction, parallel to the surface of the semiconductor element. It is difficult, although some semiconductor manufacturers have achieved this in an endeavour to avoid restricting patents, to make this conducting path sufficiently short to achieve a low enough resistance for larger signal use.

The technical breakthrough in this type of device came about when it was appreciated that a 'V' or 'U' groove etched through the junctions in a fairly conventional transistor gave the possibility of an insulated-gate, induced-charge f.e.t., in which the current flow would be 'vertical. (as in the conventional junction transistor) rather than 'lateral. (in relation to the surface of the chip) as in the normal insulated-gate component. This gave a method of manufacture of 'mosfets', as

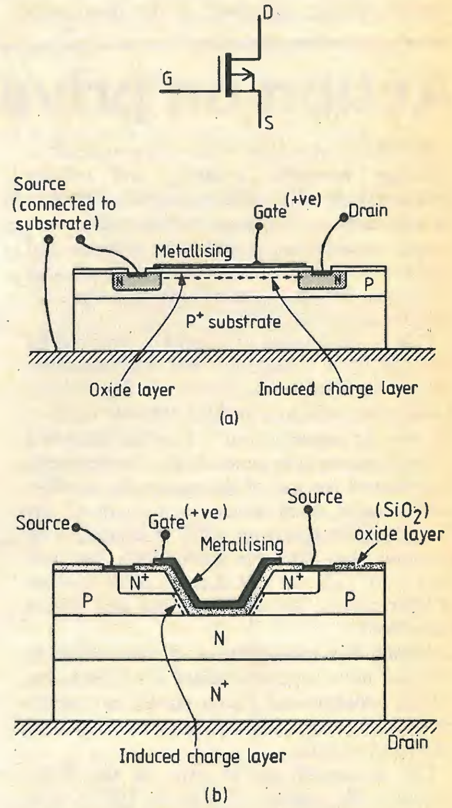


Fig. 1. *Small-signal, n-channel, insulated-gate f.e.t. of 'lateral' construction is shown at (a), while at (b) is the vertical power mosfet, in which the conducting path is a great deal shorter.*

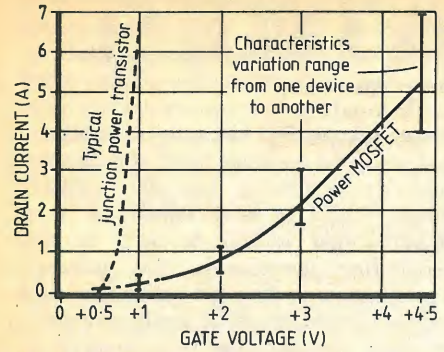


Fig. 2. *D.c. operating conditions of typical power mosfet compared with those of a junction power transistor.*

these devices are now almost universally known, which was open to any manufacturer of epitaxial planar junction transistors with the necessary skills in mask manufacture to fabricate the large number of parallel-connected igfet gates on a single chip, which are needed to lower the conducting resistance and increase the effective mutual conductance (g_m). A typical construction for such a vertical – or power – mosfet is shown in Fig. 1(b), though the proliferation of such designs within the past few years makes the concept of a 'typical' construction progressively less tenable. However, they do all have in common the parallel connexion of a large number of elements, which makes the mask design more complex, and the potential manufacturing reject rate and cost relatively high in comparison with the larger power junction transistors.

The electrical performance, under d.c. conditions, of a power mosfet, is shown in Fig. 2, with a superimposed curve from a junction power transistor added to the graph to draw attention to the differences in performance. Two features are immediately obvious from this graph – that a significantly higher forward voltage applied to the gate of the mosfet is necessary to obtain an adequate, and adequately linear, operating current, and that the mutal conductance of the power mosfet (about 2A/V in its linear region) is very much lower than that of the bipolar junction transistor (which can be in excess of 15A/V, or many hundreds of amps/volt in the case of Darlington-connected pairs).

In conventional audio-amplifier design, as it has become established over the past 20 years, the 'architecture' normally employed in the circuit is a low-power voltage amplifier, usually operated in class A, with as high an a.c. gain as is practicable without the use of an inconvenient number of gain stages, followed by an impedance-converter stage – usually a push-pull pair of compound emitter followers, forward biassed into AB operation, with an operating point chosen so that the mutual conductance of the pair of emitter followers is close to that which will be given by one half, alone, when operating in its linear region. Negative feedback is then applied from output to input to improve the overall linearity and other operating characteristics of the amplifier.

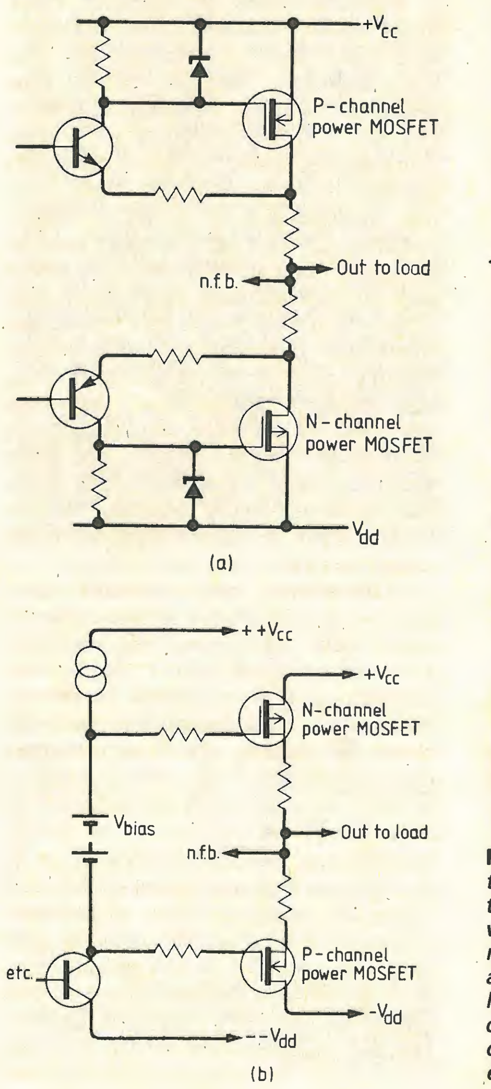
This configuration gives satisfactory bandwidth and linearity, and allows high power outputs with low quiescent thermal dissipation. The main drawback in this system is that there are invariably some low-level residues of crossover distortion, which increase in magnitude at higher frequencies, as the open-loop gain of the class A amplifier decreases – mainly as a result of the added h.f. loop-stabilizing components. This problem is worsened by the loss in loop gain through the output emitter-following stage, in which the gain is always less than unity.

To a first approximation, the output impedance of an emitter follower is $1/g_m$, which would give a gain in the output stage of $Z_L/(1/g_m+Z_L)$ at low frequencies. However, there is a further loss of gain at higher frequencies due to the limitations in turn-on and turn-off times of the transistors, so it is customary to use two or more transistors in a compound configuration in each half of the emitter follower, partly to sustain a low output impedance, and partly to allow the 100% negative feedback within the emitter-follower group to force improvements in the internal h.f. characteristics.

Inevitably, therefore, a complementary pair of output source followers using power mosfets, having a maximum g_m of some 2A/V, will perform less well in terms of linearity with a comparable class A amplifier stage than a pair of compound emitter followers with an effective (overall) g_m of, at least, some 100A/V.

Two solutions are available to this problem, of which the first is simply to incorporate the power mosfets in a compound source-follower circuit, with one or two small-signal bipolar transistors to increase the internal loop gain and reduce the dynamic output resistance and non-linearity of this stage. This provides a very simple answer to the difficulties introduced by the low g_m of the power mosfet, while allowing full advantage to be taken of the advantages of these devices (freedom from 'secondary breakdown', simpler output stage overload protection, and much better h.f. characteristics leading to lower overall t.h.d. and better transient response). I have shown a practical audio amplifier circuit using mosfets in the output stage configuration of Fig. 3(a), in another place[1].

For relatively low-power use, up to say 50 watts, this type of output stage is entirely satisfactory, and gives a good performance without calling for unconventional circuit design. However, there is a practical limitation in higher power use in that the forward voltage drop in each compound half cannot be less than the forward gate voltage (ref. Fig. 2) added



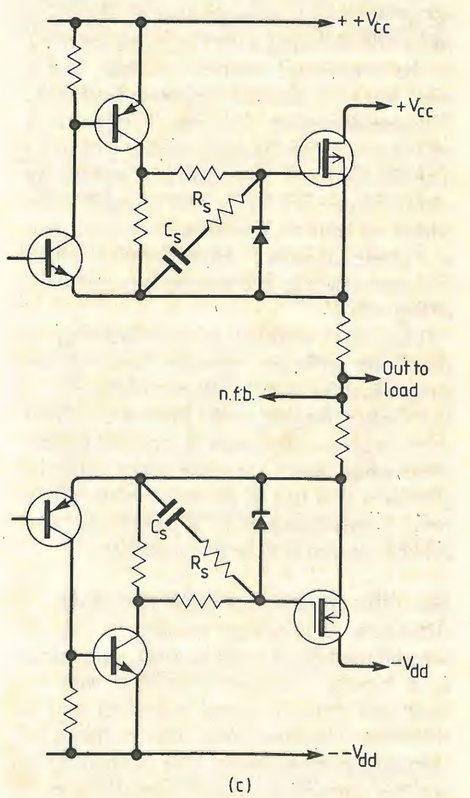


Fig. 3. *To increase the internal loop gain, the mosfet can be used with a small-signal transistor in compound source-follower, which enables low mutual-conductance mosfets to be used, while retaining their advantages, as in (a). At higher powers, larger supply voltages must be used, but circuit at (b) shows that only low-current drivers need be so supplied. Circuit at (c) is extension of approach shown in (b).*

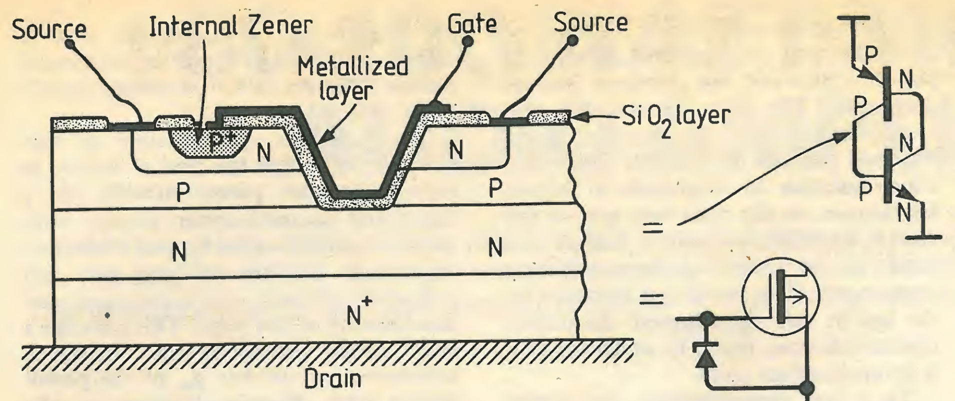


Fig. 6. *Internal protective Zener diode can cause inadvertent thyristor action, in which gate loses control.*

to the saturation voltage (V_{ce}) of the driver transistor, even though the necessary drain-source voltage of the mosfet for this output current may be less than this. This leads to the need for higher supply-line voltages, with a consequent increase in the cost of mains transformer and smoothing capacitors.

In the very simple complementary-mosfet output stage of Fig. 3(b), the driver stages can be supplied from a higher voltage line without so much of a cost penalty, since the supply currents required by the driver stages are comparatively small.

This advantage can be retained by the use of the circuit arrangement of Fig. 3(c), while still allowing a very high effective g_m of the compound emitter follower, and a high level of internal negative feedback. The problem with this circuit is that it is no longer unconditionally stable within its internal feedback loop, and h.f. stabilizing components such as C_s and R_s need to be added to achieve the desired overall gain and phase margins – an elaboration which is unnecessary in the simpler arrangement of Fig. 3(a).

The other solution to the difficulty of the lower effective transfer ratio of the simple mosfet source follower of Fig. $2_{(b)}$ is to increase the gain of the class A amplifier stage, and this approach is explored below. Meanwhile, there are some other potential pitfalls in the use of power mosfets which need consideration if a workable and reliable design is to be put together.

Specific problems with mosfets

Although the power mosfet is, in its normal method of construction, equivalent to a bipolar junction transistor with its base and emitter joined together, and is therefore immune from the problem of 'secondary breakdown' (the funnelling of emitter current through diminishing areas of the base-emitter junction and consequent localized overheating and damage) it does suffer from other problems which are unique to itself. Of these, the first and most immediate is that the gate insulation layer, an oxide film formed on the surface of the silica, is less than 0.0001in (2.5 microns) thick, and will break down if the voltage between the gate and the source exceeds some 10-20 volts – depending on the device manufacture. Since the time delay involved in this breakdown, which will destroy the device, is likely to be very short, the circuit must be designed to protect the gate against even very brief voltage excursions beyond this limit.

This difficulty can be lessened, in the construction of the device, by incorporating a Zener diode between source and gate, as shown in Fig. 4. However, this technique in its turn leads to the problem that the device must then be protected against a reverse bias – of the order of 0.6 volts – which would cause this internal Zener to conduct, since this can sometimes lead to the triggering of a thyristor-type action within the mosfet, in which the gate is irrelevant. This may not destroy the device, but may damage associated circuit elements. The simplest form of protection is the use of an external germanium diode, connected in parallel with the gate/source Zener, and arranged to conduct before the internal diode. This is not a preferred solution, however, since the reverse insulaton resistance of the Ge diode is much poorer than the unmodified input resistance of the mosfet, and is non-linear with voltage. The circuit of Fig. 3(a) is immune from this problem.

The second difficulty in the use of power mosfets arises from the very high operating frequencies possible with these components. This leads to an effective circuit element of the form shown in Fig.

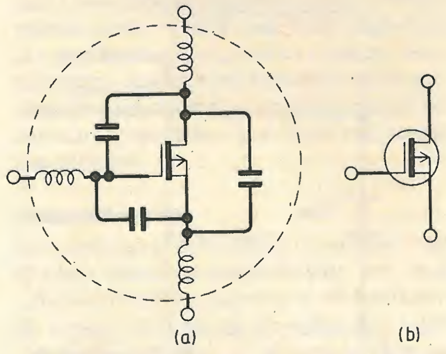


Fig. 5. *At high frequencies, stray capacitances and inductances, shown at (a), turn mosfet into an oscillator, with damaging effect.*

5(a), when the user expects the device to behave as in Fig. 5(b)! This causes immediate high-frequency oscillation, with frequently destructive effects, when such mosfets are incorporated into apparently sensible circuit configurations, and since the resultant burst of oscillation probably occurs in the 200-1000MHz range – and is brief anyway – it is unlikely that it will be seen on any monitoring instrument. The unhappy experimenter is then left contemplating a defunct device, thinking that its sensitivity to static electrification is so great as to render it unusable.

Happily, the internal gate-source capacitance is sufficiently high, typically in the range 600-1500pF, that stray static charge is unlikely to induce an electrical breakdown of the gate insulation. This internal capacitance, which must not be overlooked in circuit design considerations, also provides a convenient means for taming the h.f. behaviour of the transistor, sicne an external 'gate-stopper' resistor can then cause a predictable roll-off in h.f. response, to bring the unity gain transition frequency down to a more manageable level. An external resistor in the range 470R-4k7 is normally adequate.

Given these precautions, my experience is that power mosfets are at least as durable as normal bipolar power transistors, and allow a substantial improvement in circuit performance for a relatively small extra cost.

References

1. Linsley Hood., J. L., *Hi-Fi News and Record Review*, Vol. 25, No. 12 (Dec. 1980) pp. 83-85.

To be continued

LITERATURE RECEIVED

Power semiconductors and d.c. power supplies are detailed in the 1982 edition of the Lambda catalogue. It includes some application notes and dimensional drawings which makes it useful as a handbook. Lambda Electronics Co, Abbey Barn Road, High Wycombe, Bucks.
WW 400

Some literature that you would like to receive may be out of print. A service to help you find that elusive book is provided by **The Out of Print Book Service,** 17 Fairwater Grove East, Cardiff, CF5 2JS. They do not charge a fee but ask that all requests for the service should be accompanied with a postage stamp at the current first class letter rate and full details of the book required.
WW 401

Ceramic chip capacitors for high frequency applications are described in a bulletin from Hy-Comp Ltd, 7 Shield Road, Ashford Industrial Estate, Middlesex TW15 1AV. The bulletin lists their stability, specifications and terminations.
WW 402

DIGITAL FILTER DESIGN

Will digital filters take over from their analogue counterparts? Accuracy, versatility and falling cost suggest they will. This second article on microprocessor implementation details a procedure for recursive filter design of Butterworth and Tchebychev response, with examples and Basic programs for pole, zero calculation.

by B. M. G. Cheetham and P. M. Hughes

It has long been recognized that because of the high sensitivity of the roots of high-order polynomials to coefficient values, recursive digital filters are best implemented as cascade or parallel arrangements of second-order filter sections. Cascading several second-order filter sections as illustrated in Fig. 1(a) is equivalent to re-expressing $H(z)$ as the product of second-order transfer functions, i.e.

$$H(z)=H_1(z)\,.\,H_2(z)\,.\,.\,.\,H_m(z)$$

where each $H_i(z)$ is of the form

$$H_i(z)=\frac{1+a_1z^{-1}+a_2z^{-2}}{1+b_1z^{-1}+b_2z^{-2}} \qquad (1)$$

This is often referred to as a biquadratic transfer function and the circuit by which it is implemented a biquadratic section. The signal flow diagram of a commonly used biquadratic section is shown in Fig. 1.

The filter design problem is now that of deciding how many biquadratic sections are required, and calculating the coefficients for each of them. There are a variety of solutions to this problem, and the procedures are divided ito two broad groups: direct methods in which the poles and zeros of the digital filter are computed directly in the z-plane, and indirect methods which involve the design of an analogue prototype filter which is transformed to give a suitable digital equivalent. Details of the commonly used indirect design methods can be found in the standard texts[1]. The direct method detailed here for the design of Butterworth and Tchebychev-type digital filters is probably the simplest approach; it is non-iterative and does not require the use of a large computer. Further details and program listings for the more complex direct design methods are to be found in the standard texts.'

Recursive filters

The effect of a digital filter may be represented in z-transform notation as

$$Y(z) = H(z).X(z)$$

where $X(z)$ is the z-transform of the filter input sequence, $Y(z)$ is the z-transform of the output sequence, and $H(z)$ is the filter transfer function, as shown in a previous article (May issue). For a general filter, $H(z)$ may be expressed as

$$H(z)=\frac{\sum_{i=0}^{M} a_iz^{-i}}{1-\sum_{j=1}^{N} b_jz^{-j}}$$

where coefficients $a_0, a_1, \ldots a_M$ and $b_1, b_1, b_2, \ldots b_N$ are fixed constants. This article considers the design of a class of recursive filters i.e. filters for which $N > 0$ in the above expression. Recursive filters, generally speaking, are the most economic in both hardware and processing time requirements.

Butterworth and Tchebychev filters

Most filtering applications require an amplitude response that allows selected frequency bands to pass through the filter unaltered and eliminates as nearly as possible frequency components outside these bands. In many applications, particularly where a degree of phase distortion is acceptable in the passbands, Butterworth and Tchebychev-type filters are often used, whose responses approximate the magnitude-frequency response of an ideal filter. The difference between the two classes lies in the nature of the approximation. In the case of Butterworth-type filters, the magnitude of the frequency response is maximally flat over the passband, falling by 3 dB at the cut-off frequencies and decreasing monotonically in the stopbands. The rate of fall-off of gain in the stopbands is fixed and determined solely by the order of the filter. Tchebychev-type filters display an equi-ripple passband response with specificable ripple amplitude δ. The filter gain falls monotonically in the stopband at a rate depending on the passband ripple amplitude: the larger the ripple amplitude, the sharper is the transition from passband to stopband. Tchebychev-type filters generally show an increased rate of fall-off of gain over the equivalent Butterworth-type filter of the same order. Unlike analogue filters, a number of alternative methods exist for the design of Butterworth and Tchebychev digital filters. For a given set of design parameters, there will be a number of digital filter transfer functions which may be classed as Butterworth or Tchebychev type. The method presented here, referred to as the squared magnitude approach, is described in detail by Rader and Gold[2], and used by Ackroyd[3]. The poles and zeros of the digital filter are deducted directly from the squared magnitude of the required response with no constraints placed on the phase response, which may therefore be non-linear. Once the pole and zero positions have been determined, it is a simple matter to calculate the a_i and b_i coefficients for each of the cascaded second-order sections. As the design procedure simply consists of the evaluation of a number of given formulae, it is particularly suitable for a programmable calculator or microcomputer. The first design procedure given is that for a lowpass digital filter, which also forms the basis of the design methods for highpass, bandpass and bandstop filters.

Lowpass Butterworth filters

An analogue lowpass filter is an nth-order Butterworth type if its response $G(j\Omega)$ satisfies

$$|G(j\Omega)|^2=\frac{1}{1+\left(\frac{\Omega}{\Omega_c}\right)^{2n}}$$

where Ω signifies angular frequency. This formula gives a passband which is maximally flat, with a gain of 0dB at d.c. falling to −3dB at the cut-off frequency

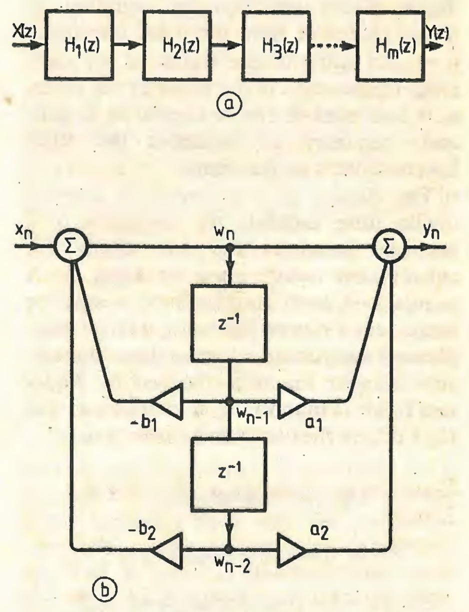


***Fig. 1.** Recursive digital filter is realised as a cascade of second-order sections with transfer functions $H_1(z)$, $H_2(z)$, etc. $X(z)$ and $Y(z)$ are the z-transformed input and output sequences respectively (a). Biquadratic section (b) has input sequence $\{x_n\}$ and output $\{y_n\}$ with $w_n=x_n-b_1w_{n-1}-b_2w_{n-2}$*
so $(1+b_1z^{-1}+b_1z^{-2})W(z)=X(z)$.
Also $y_n=w_n+a_1w_{n-1}a_2w_{n-2}$
so $Y(z)=(1+a_1z^{-1}+a_2z^{-2})W(z)$
$=\left(\frac{1+a_1z^{-1}+a_2z^{-2}}{1+b_1z^{-1}+b_2z^{-2}}\right)X(z)$, as required.

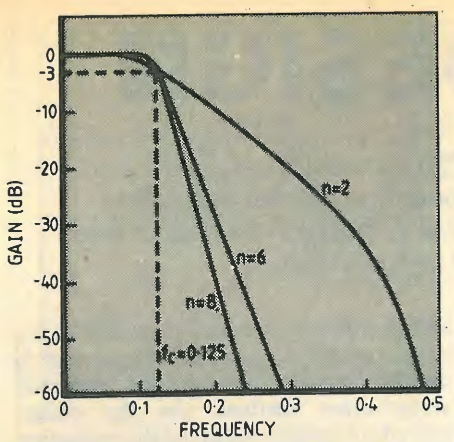


Fig. 2. Amplitude response of three Butterworth-type lowpass digital filters with cut-off equal to 0.125 effect of increasing the filter order.

Ω_c. A corresponding forumla for digital filters which produces a frequency response with similar properties to that of an analogue Butterworth filter is

$$|H(e^{j\omega})|^2=\frac{1}{1+\left(\frac{\tan\omega/2}{\tan\omega_c/2}\right)^{2n}} \quad (2)$$

where ω_c is equal to the radian cut-off frequency. In place of radian frequency ω it is often convenient to refer to relative frequency f, obtained by dividing ω by 2π. Hence f is the relative frequency of a sinusoid of frequency $F=f\times(1/T)$. Because F must be less than $1/2T$ to satisfy the sampling theorem, values of f need only be considered in the range 0 to 0.5. Fig. 2 shows the magnitude responses of three digital filters specified by equation 2. These filters all have a cut-off frequency $f_c=\omega_c/2\pi$ equal to one-eighth of the sampling frequency. As the order of the filter, n, is increased the more closely its magnitude response approximates the ideal lowpass filter characteristic.

The design of a Butterworth lowpass digital filter requires the invention of a suitable function $H(z)$ for which the substitution $z=e^{j\omega}$ gives equation 2. A number of such functions of z may be found, their derivation being a fairly complicated mathematical procedure. A suitable transfer function derived by Rader and Gold[2] is found to have poles located at U_m+jV_m in the complex z-plane, where

$$U_m=(1-\tan^2\pi f_c)/D_m$$

$$V_m=(2\tan\pi f_c\sin\theta_m)/D_m$$

$$\text{with } D_m=1-2\tan\pi f_c\cos\theta_m+\tan^2\pi f_c$$

$$\text{and } \theta_m=\frac{(2m+n+1)\pi}{2n} \quad (3)$$

where $m=0$ 1, . . . $(n-1)$ for a filter of even-order n; odd-order Butterworth filters may also be specified. Each of the n transfer function zeros is located at $z=-1$.

Once the poles and zeros of a recursive digital filter have been calculated the next stop in the design process is to determine the coefficients a_1, a_2, b_1 and b_2 of equation 1 for each cascaded second-order section. For even-order filters, the transfer function poles can always be grouped as complex conjugate pairs.

Hence the overall filter transfer function $H(z)$ expressed as a product of $n/2$ second-order (biquadratic) transfer functions is

$$H(z)=A_0\prod_{m=0}^{n/2-1}\times\frac{(1+z^{-1})(1+z^{-1})}{[1+z^{-1}(U_m+jV_m)][1+z^{-1}(U_m-jV_m)]}.$$

Expanding the above equation yields the required form

$$H(z)=A_0\prod_{m=0}^{n/2-1}\times\frac{1+2z^{-1}+z^{-2}}{1-2U_mz^{-1}+(U_m^2+V_m^2)z^{-2}} \quad (4)$$

$$\frac{1}{A_0}=H(z)|_{z=1}=H(e^{j\omega})|_{\omega=0}. \quad (5)$$

Tchebychev filters

The squared magnitude of the frequency response of a Tchebychev low-pass digital filter with cut-off ω_c is

$$|H(e^{j\omega})|^2=\frac{1}{1+\epsilon^2V_n^2\left(\frac{\tan\frac{\omega}{2}}{\tan\frac{\omega_c}{2}}\right)} \quad (6)$$

where $V_n(x)$ denotes the nth-order Tchebychev polynomial function of x and ϵ is a parameter used to set the pass band ripple amplitude. Fig. 3 shows the frequency response of a typical fourth-order Tchebychev filter. The magnitude of the pass band ripple δ is related to the parameter ϵ by

$$\epsilon=\left(\frac{1}{(1-\delta)^2}-1\right)^{1/2} \quad (7)$$

Increasing δ has the effect of sharpening the cut-off region of the filter and increases the stop band attenuation, but only at the expense of increasing the pass band ripple.

The transfer function poles of an nth-order digital filter whose frequency response satisfies equation 6 are located at U_m+jV_m in the complex z-plane where

$$U_m=\{2(1-a\tan\pi f_c\cos\theta_m)/D_m\}-1$$

$$V_m=(2b\tan\pi f_c\sin\theta_m)/D_m$$

$$\text{with } D_m=(1-a\tan\pi f_c\cos\theta_m)^2+b^2\tan^2\pi f_c\sin^2\theta_m$$

$$\text{for } m=0, 1, 2, \ldots n-1 \quad (8)$$

$$\text{and } 2a=(\sqrt{\epsilon^{-2}+1}+\epsilon^{-1})^{1/n}-(\sqrt{\epsilon^{-2}+1}+\epsilon^{-1})^{-1/n}$$

$$2b=(\sqrt{\epsilon^{-2}+1}+\epsilon^{-1})^{1/n}+(\sqrt{\epsilon^{-2}+1}+\epsilon^{-1})^{-1/n}$$

θ_m is calculated as for Butterworth filters from equation 3 and each of the n transfer function zeros is again located at $z=-1$. Once the poles and zeros of the transfer function have been calculated, substituting for U_m and V_m in equation 4 gives the $n/2$ biquadratic transfer functions used in the realization of the filter.

Example 1

Design a low pass Butterworth digital filter with the specification

sampling frequency	12 kHz
cut-off frequency	3.2 kHz
attenuation at 4.0 kHz	>20dB i.e. 10
passband gain	0dB

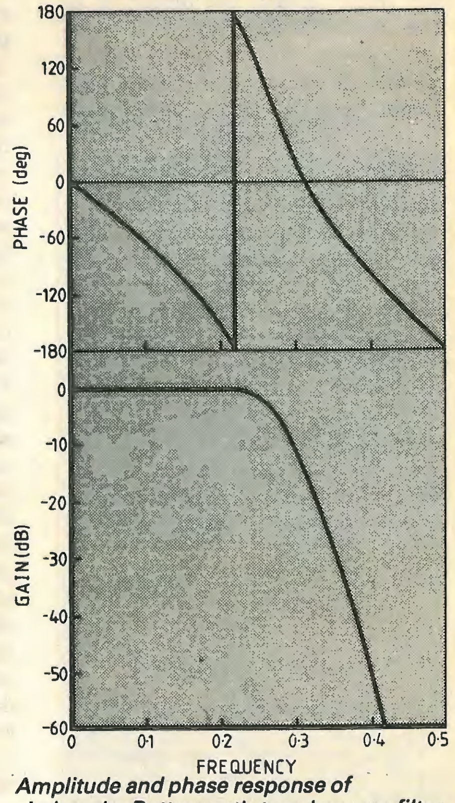
From this the relative cut-off frequency is $f_c=3.2/12=0.2667$. For the attenuation at 4 kHz to be greater than 10, it follows from equation 2 that

$$\frac{1}{10}>\left(1+\left(\frac{\tan4\pi/12}{\tan3\cdot2\pi/12}\right)^{2n}\right)^{-1/2}$$

The lowest even-order filter which will satisfy these requirements had $n=6$. The six poles for this filter are

m	U_m	V_m
2, 3	−0.0533	0.1313
1, 4	−0.0613	0.4129
0, 5	−0.0831	0.7640

Substituting these values into equation 4 gives three biquadratic transfer functions in the form of equation 1 with coefficients a_1, a_2, b_1 and b_2 as shown below. Scaling factor A_0, calculated by means of equation 5, should be set to 0.0401 to give unity passband gain. The magnitude and phase responses of the filter are shown alongside.



Amplitude and phase response of sixth-order Butterworth-type lowpass filter designed in Example.

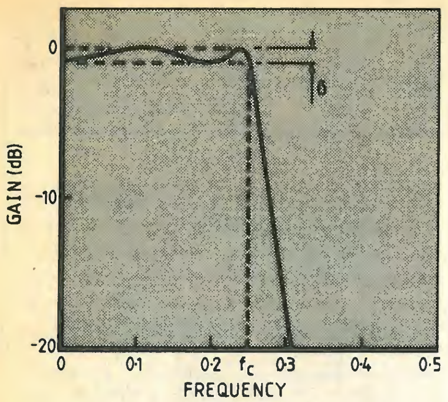


Fig. 3. Amplitude response of a typical fourth-order Tchebychev low pass filter with cut-off equal to 0.25 and specified ripple, δ.

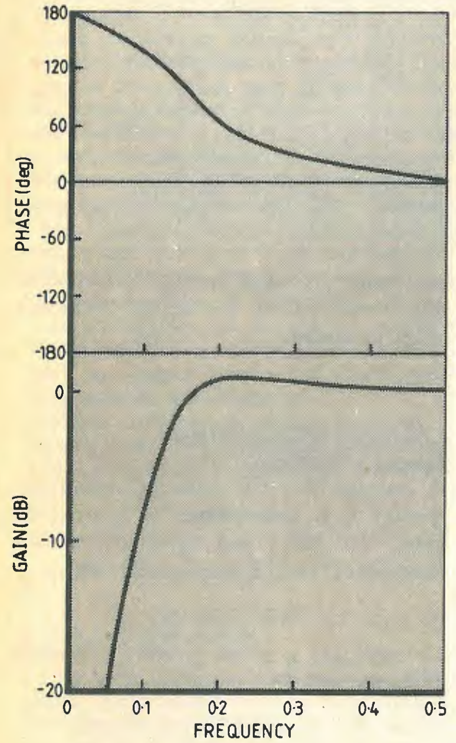


Fig. 4. Amplitude and phase response of second-order Tchebychev-type highpass digital filter designed using lowpass to highpass transformation.

Example 2

Design a second-order Tchebychev low-pass digital filter with $f_c=1/3$ and a pass-band ripple of 0.1 (0.915 dB).

By equation 7, $\epsilon=0.4843$ and a and b and thus calculated to be 0.844 and 1.2834 respectively. Substituting these values into equations 8 with θ_m obtained from equation 3 for $m=0$ and 1 shows that the poles of the filter lie at

$$z=-0.38075\pm j\,0.49030.$$

Calculating now the biquadratic transfer function for the single section required gives

$$H(z)=A_0\frac{1-2z^{-1}+z^{-2}}{1+0.7615z^{-1}+0.3854z^{-2}}$$

Substituting $z=1$ into the above shows that A_o must be set to 0.5367 for unity gain at d.c.

Example 3

Design a fourth-order bandpass Butterworth type filter with a sampling frequency of 16 kHz and cut-off frequencies of 2 and 4 kHz.

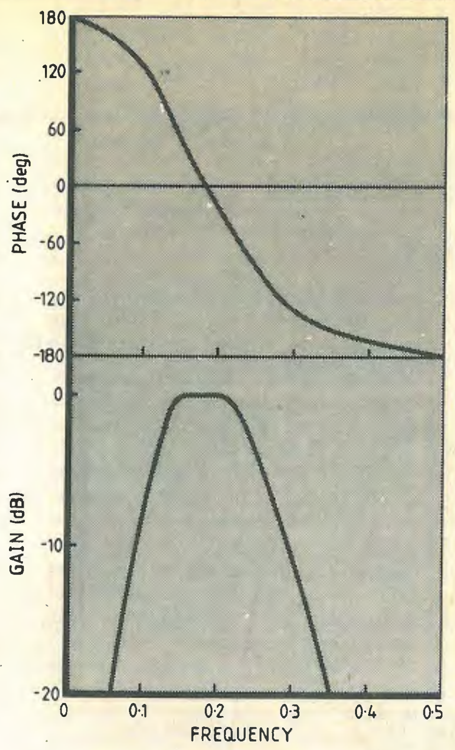
As the bandpass transformation doubles the order of the prototype filter, a second-order Butterworth lowpass prototype is required with a relative cut-off frequency $f_c=(4-2)/16=0.125$. The two poles of the prototype filter, calculated from the expressions for U_m and V_m (eqns 3), lie at $z=0.4714\pm j0.3333$. The two zeros are at $z=-1$. Substituting for $f_h=4/16$ and $fl=2/16$ in equation 9 gives $\alpha=0.4142$. By equation 10 the poles of the bandpass filter are calculated to be

$$z=0.5262\pm j0.5885$$
$$z=0.0833\pm j0.7266.$$

The zeros of the prototype filter transform to two zeros at $z=-1$ and two at $z=+1$. Pairing the zeros at $z=-1$ with the poles at $z=0.0883\pm j0.7266$, gives $H(z)$

$$=\left(\frac{1-2z^{-1}+z^{-2}}{1-1.0524z^{-1}+0.6232z^{-2}}\right)\left(\frac{1+2z^{-1}+z^{-2}}{1-0.1665z^{-1}+0.5348z^{-2}}\right)$$

The passband gain is set to 0 dB by scaling $H(z)$ by a factor of 0.0976. *Graph* shows the magnitude and phase responses of the designed filter.



Amplitude and phase response of fourth-order Butterworth-type bandpass filter designed in Example

Highpass, bandpass and bandstop filter design

The design of Butterworth and Tchebychev filters may be carried out by transformations applied to lowpass prototype filters obtained by the methods described earlier 4. The simplest of the transformations is that of low to highpass. A lowpass filter with relative cut-off frequency f_c, is transformed to a high-pass filter with relative cut-off frequency $0.5-f_c$, by replacing z^{-1} with $-z^{-1}$ in the filter transfer function $H(z)$.

Applying this transformation to the lowpass Tchebychev filter in the previous example, we obtain a high-pass Tchebychev filter with a relative cut-off frequency of 1/6. The resulting transfer function $H(z)$ is

$$0.5367\left(\frac{1-2z^{-1}+z^{-2}}{1+0.7615z^{-1}+0.3854z^{-2}}\right)$$

The magnitude and phase responses of this filter are shown in Fig. 4.

Low to band pass transformation

A low pass prototype filter with relative cut-off frequency f_c is transformed to a bandpass filter with lower and upper cut off frequencies of f_l and f_l+f_c $(=f_h)$ respectively, by the replacement of z by

$$\frac{z(z-\alpha)}{(\alpha z-1)}$$

in the prototype transfer function, where

$$\alpha=\frac{\cos\pi(f_h+f_l)}{\cos\pi(f_h-f_l)}. \quad (9)$$

This has the effect of doubling the order of the prototype filter, and consequently each second-order section in the lowpass filter is transformed to a fourth-order transfer function in the bandpass filter. It is necessary therefore to reduce each of these fourth-order transfer functions to the product of two-second order sections. The transformation may be conveniently applied by considering the poles of the prototype each of which produces two poles in the resulting bandpass filter. For each pole of the prototype filter situated at $z=p$, say, the transformed filter will have poles at values of z which satisfy

$$p=\frac{z(z-\alpha)}{(\alpha z-1)}$$

$$\text{i.e. at } z=\tfrac{1}{2}\alpha(1+p)\pm\left\{\frac{1}{4}\alpha^2(1+p)^2-p\right\}^{1/2} \quad (10)$$

As p in this equation will normally be a complex number, computation of the poles of the bandpass filter involves the calculation of a complex square root, see appendix. The $2n$ poles of the band-pass filter may then be determined by substituting each of the n prototype poles into equation 10. The band-pass poles may then be grouped into complex conjugate pairs to produce the denominators for each of the n biquadratic transfer functions which comprise the overall filter transfer function, $H(z)$. The zeros of the bandpass filter are calculated from the zeros of the prototype using the same formulae as were used for the poles. By substituting $p=-1$ into equation 10 the n zeros of the lowpass prototype located at $z=-1$, transform to $2n$ zeros in the bandpass filter, n located at

```
10 ! CALCULATE BWORTH LP. POLES
15 DISP "FILTR ORDER" @ INPUT N
20 DISP "REL. CUT-OFF FREQ.?"
25 INPUT F1@ I2=3*N/2-1 @ I1=N
30 DISP "POLES OF FILTER"
35 DISP "     REAL        IMAG"
40 FOR I=I1 TO I2 @ T=2*I+1
45 T=T/(2*N)*PI @ C1=PI*F1
50 A=TAN(C1) @ A2=A*A @ B=2*A
55 D=1-B*COS(T)+A2 @ U=(1-A2)/D
60 V=B*SIN(T)/D
65 DISP USING 70 ; U,V
70 IMAGE 2X,SD.4D,5X,SD.4D
75 V=V*-1 @ DISP USING 70 ; U,V
80 NEXT I
85 END
```

Programs to calculate poles of lowpass filters – Butterworth above, Tchebychev below – and for lowpass transformation to bandpass and bandstop, right.

```
10 ! CALCULATE TCHEB LP. POLES
15 DISP "FILTR ORDER" @ INPUT N
20 DISP "REL. CUT-OFF FREQ.?"
25 INPUT F1@ DISP "RIPPLE ?"
30 INPUT R@ W=1-R @ E=W^2
35 W=1/E-1 @ E=SQR(W)
40 M=1/(E*E)+1 @ M=SQR(M)+1/E
45 M1=LGT(M)/N @ M1=10^M1
50 M2=1/M1 @ A=.5*(M1-M2)
55 B=.5*(M1+M2) @ Y=N @ Z=3*N/2
60 Z=Z-1 @ DISP " FILTER POLES"
65 DISP "     REAL        IMAG"
70 FOR I=Y TO Z @ T=2*I+1
75 T=T/(2*N)*PI @ C1=PI*F1
80 W=TAN(C1) @ W2=W*W
85 U=1-A*W*COS(T)
90 D=U*U+B*B*W2*SIN(T)^2
95 U=2*U/D-1 @ V=2*B*W*SIN(T)/D
100 DISP USING 105 ; U,V
105 IMAGE 2X,SD.4D,5X,SD.4D
110 V=V*-1
115 DISP USING 105 ; U,V
120 NEXT I
125 END
```

```
10 ! BANDPASS TRANSFORMATION
15 DISP "PROT. ORDER" @ INPUT N
20 DISP "LWR CUTOFF" @ INPUT F1
25 DISP "UPR CUTOFF" @ INPUT F2
30 F=(F2+F1)*PI @ F4=(F2-F1)*PI
35 A=COS(F)/COS(F4)
40 FOR I=1 TO N/2
45 DISP "REAL(POLE)" @ INPUT U
50 DISP "IMAG(POLE)" @ INPUT V1
55 V2=V1*-1 @ P1=1+U
60 R7=A*P1/2 @ I1=A*V1/2
65 I2=A*V2/2 @ A2=A*A
70 V=V1 @ GOSUB 175
75 U8=R7+B0 @ U9=R7-B0
80 V8=I1+B1 @ V9=I1-B1 @ V=V2
85 GOSUB 175
90 U7=R7+B0 @ U6=R7-B0
95 V7=I2+B1 @ V6=I2-B1
100 DISP "TRANSFORMED POLES"
105 DISP "     REAL         IMAG"
110 DISP USING 115 ; U8,V8
115 IMAGE 2X,SD.4D ,5X,SD.4D
120 DISP USING 115 ; U7,V7
125 DISP USING 115 ; U6,V6
130 DISP USING 115 ; U9,V9
135 DISP @ DISP "QDRTIC COEFFS"
140 B1=-2*U8 @ B2=U8*U8+V8*V8
145 DISP USING 150 ; B1,B2
150 IMAGE X,"B1,B2",2(4X,SD.4D)
155 B1=-2*U9 @ B2=U9*U9+V9*V9
160 DISP USING 150 ; B1,B2
165 DISP @ NEXT I
170 END
175 ! CALCULATE (1+P)^2
180 R2=P1*P1-V*V @ J2=2*P1*V
185 ! CALCULATE A^2*(1+P)^2/4
190 R0=A2*R2/4 @ J0=A2*J2/4
195 ! SUBTRACT P
200 R1=R0-U @ J1=J0-V
205 ! CONVERT TO EULER FORM
210 R=R1*R1+J1*J1 @ R=SQR(R)
215 T=ATN2(J1,R1)
220 ! TAKE SQUARE ROOT
225 R=SQR(R) @ T=T/2
230 ! BACK TO CARTESIAN
235 B0=R*COS(T) @ B1=R*SIN(T)
240 RETURN
```

```
10 ! BANDSTOP TRANSFORMATION
15 DISP "PROT. ORDER" @ INPUT N
20 DISP "LWR CUTOFF" @ INPUT F1
25 DISP "UPR CUTOFF" @ INPUT F2
30 F=(F2+F1)*PI @ F4=(F2-F1)*PI
35 A=COS(F)/COS(F4)
40 FOR I=1 TO N/2
45 DISP "REAL(POLE)" @ INPUT U
50 DISP "IMAG(POLE)" @ INPUT V1
55 V2=V1*-1 @ P1=1-U
60 R7=A*P1/2 @ I1=A*V1/2
65 I2=A*V2/2 @ A2=A*A
70 V=V1 @ GOSUB 175
75 U8=R7+B0 @ U9=R7-B0
80 V8=I1+B1 @ V9=I1-B1
85 V=V2 @ GOSUB 175
90 U7=R7+B0 @ U6=R7-B0
95 V7=I2+B1 @ V6=I2-B1
100 DISP "TRANSFORMED POLES"
105 DISP "     REAL         IMAG"
110 DISP USING 115 ; U8,V8
115 IMAGE 2X,SD.4D ,5X,SD.4D
120 DISP USING 115 ; U7,V7
125 DISP USING 115 ; U6,V6
130 DISP USING 115 ; U9,V9
135 DISP @ DISP "QDRTIC COEFFS"
140 B1=-2*U8 @ B2=U8*U8+V8*V8
145 DISP USING 150 ; B1,B2
150 IMAGE X,"B1,B2",2(4X,SD.4D)
155 B1=-2*U9 @ B2=U9*U9+V9*V9
160 DISP USING 150 ; B1,B2
165 DISP @ NEXT I
170 END
175 ! CALCULATE (1-P)^2
180 R2=P1*P1-V*V @ J2=2*P1*V
185 ! CALCULATE A^2*(1-P)^2/4
190 R0=A2*R2/4 @ J0=A2*J2/4
195 ! ADD P
200 R1=R0+U @ J1=J0-V
205 ! CONVERT TO EULER FORM
210 R=R1*R1+J1*J1 @ R=SQR(R)
215 T=ATN2(J1,R1)
220 ! TAKE SQUARE ROOT
225 R=SQR(R) @ T=T/2
230 ! BACK TO CARTESIAN
235 B0=R*COS(T) @ B1=R*SIN(T)
240 RETURN
```

$z=1$, the other n at $z=-1$. When the order of the prototype filter is even, as is normally arranged, the zeros of the bandpass filter may be grouped in pairs such that of the n second-order sections which comprise the filter, half will have transfer function numerators equal to

$$1+2z^{-1}+1 \text{ (two zeros at } z=-1)$$

whilst the remaining half have numerators of

$$1-2z^{-1}+1 \text{ (two zeros at } z=+1)$$

The n biquadratic transfer functions may now be completed by pairing each of the calculated numerators with a suitable denominator. As a general rule, the denominators of the second-order sections should be parried with the numerators in such a way that the poles of each section are those nearest to the zeros it implements. The filter transfer function $H(z)$ is simply the product of the n calculated biquadratic transfer functions. The passband gain may be set to one by the introduction of a scaling factor A_0 equal to the reciprocal of $H(e^{jw})$ evaluated at the central frequency of the passband.

Example 4
Design a fourth-order bandstop Tchebychev filter with a sampling frequency of 20 kHz, lower and upper cut-off frequencies of 1 and 1.2 kHz respectively and a passband ripple of 0.1.

A second-order Tchebychev lowpass prototype is required with a relative cut-off frequency of 0.49 and a passband ripple amplitude of 0.1. The two poles of this filter are found to lie at $z=-0.9681\pm j0.0482$, with the two zeros at $z=-1$. Using the transformation of equation 11 the poles and zeros of the band-stop filter are

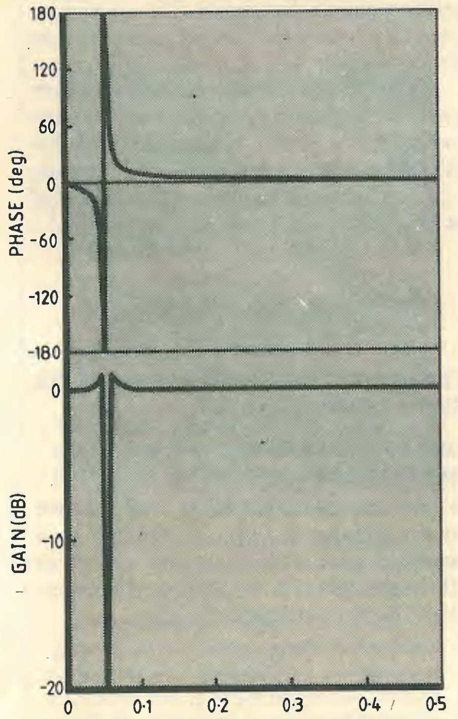
poles	zeros
0.9356±j0.3099	0.9414±j0.3374
0.9171±j0.3553	0.8414±j0.3374

The filter transfer function $H(z)$ is thus calculated to be

$$\left(\frac{1-1.8827z^{-1}+z^{-2}}{1-1.8712z^{-1}+0.9714z^{-2}}\right)$$

$$\left(\frac{1-1.8827z^{-1}+z^{-2}}{1-1.8341z^{-1}+0.9672z^{-2}}\right)$$

which must be scaled by 0.969 to set the d.c. gain to unity. The magnitude and phase responses of this filter are shown alongside.



Amplitude and phase response of Tchebychev-type bandstop or notch filter designed using lowpass to bandstop transformation in Example 4.

Lowpass to bandstop transformation

A lowpass filter with relative cut-off frequency f_c is transformed to a bandstop filter with lower and upper cut-off frequencies of f_l and f_h respectively, where

$$f_c=0.5+f_l-f_h$$

by replacing z in the prototype transfer function with

$$\frac{z(z-\alpha)}{1-\alpha z}$$

where α is defined as for the bandpass transformation. Apply the transformation to poles and zeros it is readily shown that a pole or zero of the prototype located at $z=p$ transforms to two poles or zeros of the bandstop filter located at

$$z=\tfrac{1}{2}(1-p)\alpha\pm\left\{\frac{1}{4}(1-p)^2\alpha^2+p\right\}^{1/2} \quad (11)$$

The procedure for calculating the second-order sections of a bandstop filter is similar to that for a bandpass filter. In this case, however, the zeros are no longer located at $z=\pm1$ but form complex conjugate pairs which must be expanded in the normal way to give the numerators of the second-order sections. The numerators of all the second-order sections will be identical.

Continued on page 79

LETTERS

WIRELESS?

I have just bought a copy of your magazine, for the first time in some years, and after reading it with some amusement I have the following comments to make.

Firstly, I suggest you rename the magazine *Computer World (Wireless Computer?, Computer Wireless??)* – I bought the magazine because I want to buy a communications receiver – I work with computers and I find it amusing that your magazine is now full of computer projects rather than the radio projects it used to be.

Secondly, with respect to Mr B. Reay's letter (April 1982) – I have heard of the RSGB's opposition to c.b., but not experienced it directly – I am a c.b. user, but find the RSGB has nothing to worry about. Most c.b. users would not join the RSGB (they can barely operate a 'rig' much less pass the RAE). Those who are interested in radio (myself included), find c.b. very frustrating (2 watts e.r.p. and 200 users per channel – at least in London) and this exposure to radio usually causes strong motivation to take (and pass) the RAE, thus escaping from the school-children and 'wallys" (although I hear 2 metres is not much better); and thus the serious user ends up as an amateur anyway. I have heard that the RSGB has been swamped with enquiries since c.b. legalization!

On "Microchips and Megadeaths" – if all of us who work irrelevant fields would refuse to work on weapons systems (as I have done), there would be no such systems.

Finally, what has happened to the glossy paper *WW* used to be printed on? and where can I buy a second-hand communications receiver?.

Hugh J. Davies
East Barnet
Herts

DISC DRIVES

I refer to Mr Watkinson's article on disc drives in your March issue in which he states that a 16-bit computer can only address 32K memory locations. It is well known that 16 bits can address 64K locations (2^{16} = 65,536), a feature common to most microprocessors. Could Mr Watkinson please explain his apparently erroneous mathematics?

P. C. Major
Winchester
Hants.

The author replies:
I am grateful to Mr Major for giving me the opportunity to explain a problem which has been encountered by many familiar with microprocessors when faced with more powerful hardware. I am sure, however, that Mr Major would agree that such treatment would be out of place in a series which is primarily devoted to disk drives.

The definition of a 16 bit machine which I find most satisfactory is that it possesses a data path of 16 bits in the c.p.u. Neglecting more complex machines which employ block fetching in conjunction with a cache memory, a 16 bit machine will have 16 bit wide memory locations. This allows an entire parameter to be fetched with one memory cycle. In such machines, the smallest entity of interest is still the 8 bit byte, of which there are two in each location. As one address bit is necessary to specify whether the high byte or the low byte is of interest, only 15 bits are left to specify the location. 2^{15} locations equals 32K as stated.

The address range of a processor is determined by circuitry quite distinct from the data paths in both microprocessors and minicomputers. For example, the 8085 has an 8 bit data path, but has the ability to generate 16 bit addresses by the time consuming expedient of a two-byte program counter which has to be loaded a byte at a time. This is simply the result of designing for a different price performance target. Thus on the one hand we might have a minicomputer which obtains two bytes at a time from 32K locations, and on the other hand a microprocessor which obtains one byte at a time from 64K locations. The number of available bytes is the same in each case, but all other things being equal, the minicomputer will be about twice as fast.

For further reading on the subject I would recommend the Intel 8085 Manual and the PDP 11/04 Processor Handbook.

BLUMLEIN BIOGRAPHY

The September 1973 issue of *Wireless World* carried a letter from Rex Baldock which contained the following paragraph:

"The full story of Alan Blumlein's contribution to technical history will appear in the forthcoming biography by Mr F. P. Thomson, written in conjunction with Simon Blumlein, and anyone with information likely to be of value should write to Mr Thomson at 39 Church Road, Watford WD1 3PY Herts, England."

On the 1st June 1977, a GLC plaque to Blumlein was unveiled and Mr F. P. Thomson gave a speech in which he said he had been "persuaded to write a biography" on Blumlein. I, for one, reported the prospect of a Blumlein biograophy in print (e.g. *Hi Fi News*, August, 1977) and in February 1982 wrote to Mr Thomson querying progress on his biography. In reply I have received letters which tell me that I am "impertinent" and my own articles on Blumlein are "inaccurate", but offer no information on a publication date of the mooted biography.

I would be interested to hear from anyone who answered the call for biographical information nearly ten years ago, and is now concerned. My address is Garden Flat, 5 Denning Road, London NW3 1ST.

Barry Fox
Hampstead
London NW3

MICROCHIPS AND MEGADEATHS

Steve Coleman (April letters), has taken too literally my plea for refusal to fight. This was intended to apply to opposition in its widest sense, both militarily and industrially, now and in the future.

Electronic engineers in the UK, the USA and USSR are in ideal positions to frustrate the arms race. Also well placed are their supporting teams, without whose help they cannot operate. This applies to those who type, drive lorries and serve in canteens. When the Americans say "tell that to the Russians", I have little doubt that the Russians say "tell that to the Americans". The decisions of Nuremberg tell us that obedience to orders from higher authority is no defence against charges of immoral acts.

The records of both super-powers are appalling, and give us no justification for an alliance with either power. Balance the massacre of Katyn against that of Mi Lai. Balance Afghanistan against Vietnam. Balance the rapes of the Baltic states and Central Europe against the many American invasions of South America, the destabilization of its emerging democracies and the military support of regimes as foul as that of Hitler. It was the US which dropped two atomic bombs on Japan when the latter was on the verge of surrender. The US has come close to using atomic weapons several times since then.

How can we depend upon British political parties? The government of Anthony Eden handed Russian prisoners over to Stalin and almost certain death, and later ordered the invasion of Suez. The government of Clement Attlee agreed to the use of atomic bombs on Japan. The Wilson government supported the US over Vietnam. The answer is that individuals must act according to their own consciences, or like lemmings we will follow each other into oblivion.

By the laws of chance, if weaponry be retained at its present level, sooner or later, by virtue of a technical fault or the action of a person of deranged mind, disaster must follow.

Yes, Steve Coleman, let us extend our scientific interest in technology into a scientific analysis of society.

R. Whitehead
Sutton
Surrey

POOR DEAL FOR AMATEUR RADIO

I was very interested in the letter from B. Reay in your April issue. My copy was available before the 1st of the month, otherwise I might have thought that this was a special "All Fools" feature.

May I first of all make it quite clear that, as many of your readers will know from my call-sign, although I am employed by the RSGB I am not privy to any matters of policy and I do not even work at the Society's premises. These comments are therefore made purely from the standpoint of an ordinary member of forty years standing and a reasonably active transmitting amateur since 1948.

Everyone will regret that your correspondent receives no replies to his letters, and it is up to him to take this matter up with the appropriate representatives and committee members. On the other hand however, nobody is obliged to reply to unsolicited communications which contain nothing but unconstructive criticisms and/or impossible demands, particularly from individuals claming to speak for many others.

Dealing with Mr Reay's points in roughly the same order as in his letter, first of all I do not agree that the RSGB have been anti-c.b. I have no feelings either way in this matter, but it seems to me that the Society dealt with the matter very fairly, even issuing a list so that the uninitiated could be told the essential differences between the two Services.

With regard to the 70 MHz band, this is of course not an international allocation but is made available on a "grace and favour" basis subject to the requirements of the priority users of the band. Incidentally, Mr Reay, either by accident or design, makes no mention of the new u.h.f. allocations made available, together with the three new h.f. bands, at the 1979 WARC as a result of much hard work done by national societies, including the RSGB. So far one h.f. band has been allocated to UK amateurs, with the remainder to follow. It is also worth remembering that, despite the weight of ARRL, amateurs in the USA do not yet have any of these facilities.

With regard to the abuse of the London repeaters, this is often referred to in *"Radio Communication"*, and to close the repeaters down would be about as logical as terminating the telephone system because some people "fiddle" coin-boxes or make obscene calls. As for persuading the Home Office to catch the offenders, this suggestion fails to appreciate the finance and manpower position in which the authorities find themselves at present, not to mention the limitations of the Wireless Telegraphy Act in its present form.

Concerning the RAE, I agree that this to many people seems to be more a lottery than an examination, this being all part of the general lowering of standards. However, Mr Reay has presumably joined in the joke and obtained his callsign. If the "B" licence (introduced at the insistence of RSGB, by the way) does not give him sufficient scope, he only has to pass the Morse test to obtain extended facilities and, by his operating, help to stem the undeniable decline in amateur radio.

With regard to amateur radio being a technical hobby, not many people build or repair their equipment these days, and why on earth should they, unless they are so inclined? None of my licences, not even the old "Experimental" one, has made any reference as to who makes my equipment; why is it, therefore, that from time to time those who are presumably still living in the times when it was not possible to buy ready-made gear, raise this old red herring. The acid test always used to be the quality of one's signals and the manner in which the station was operated. No-one expects Keegan to tan a piece of leather for his boots or Boycott to turn his bat on the lathe!

Finally, therefore, I would submit that as a result of having a strong and vigilant society the British amateur, far from being given a raw deal, can look to the future with confidence. An example of this is the prompt action taken after the recent Home Office gaffe over the licence schedule.

If Mr Reay, and any who think like him, would only stop and reflect, they may yet see the wood through the trees.
E. G. Allen, G3DRN
London SW20

As an RSGB member, the letter from B. Reay in the April edition was very interesting to me, and I agree with many of the points raised. It is sad that the growing numbers of concerned and disenchanted members have to air their views in a companion journal, as the Society claiming to represent all radio amateurs is totally out of touch with, and does not appear to want to listen to, its membership.

The current licensing fiasco show the Society is not in touch with the Home Office either, admitting in March *Radio Communication* "The first the Society knew about the publication of the Gazette notice was when a Member telephoned . . .". At £229, I don't subscribe to the Gazette, but surely with an income of around £700,000 the Society should. It seems the member who kindly telephoned was lucky to get an answer – it's difficult to hear the 'phone with your head in the sand!

When tuning across the bands you often hear someone complaining about the RSGB – not another band plan – yet more repeaters to be introduced – the list is endless, so I won't compile yet another one here. I'll simply say to the "Old Boys" at the RSGB, listen to the disenchanted membership, and get a much-needed decent deal for the radio amateurs you claim to represent.
Peter Thurlow, G8SUH
Dagenham
Essex

DATA STORAGE

In reply to the letter by B. Savage on the subject of "An economical Z-80 development system" (*Wireless World*, April 1982), I must say, I find his comments somewhat irrelevant. Although I agree with him on the availability and advantages of fast, reliable serial data transfer facilities, for bulk storage on cassette or for long distance communication, the article describes a technique which allows the two otherwise separate microcomputers to become integral components in a much more powerful, versatile and self-contained development system. Using this philosophy the interconnecting wires which Mr Savage refers to present absolutely no hardship.

Parallel data transfer generally requires no error checking, communication medium bandwidth presents no problem, transfer speed is optimized, more flexibility is possible in the software 'management' package and as an added bonus, the tape interface remains available on both systems for bulk storage.

On the subject of efficiency, it may be noted that two-way handshaking is involved. This was a deliberate policy to ensure universal compatibility with 'host computers' of widely varying operating speeds and with their own special data transfer facilities. Transfer speed may be improved by utilising the interrupt facilities of both computers but for absolute maximum efficiency, DMA must be used.
G. Winstanley
Bio-medical Eng. Unit
N. Staffordshire Polytechnic

AMATEURS AND C.B.

I see from recent reports that the number of applicants for amateur radio licences is continuing at a high level. I listen to the anarchy that prevails on the h.f. bands at present I wonder why anyone bothers to go to all the expense and effort of passing the requisite examinations. Quite often there seems to be more DX activity going on around 11 metres by illegal c.b. operators than in the 10 metre Amateur band. One can often hear these operators describing their 250 watt rigs over the air and exchanging names and addresses for QSL cards, indicating that they consider the chances of being prosecuted to be negligible.

It is when you start talking to the engineers of British Telecom who operate the detection service for illegal transmissions on behalf of the Home Office that the magnitude of the problem becomes apparent. Until a few years ago complaints came in from the public about interference on radio or television at such a rate that in a typical telephone area about 100 were waiting to be investigated at any time. During the past couple of years so many complaints have been received that the backlog in many areas has risen into the thousands, completely overwhelming the small detection teams that the Home Office is prepared to pay for. In fact, one engineer told me sadly that if you confined your illegal transmissions to the bands between 1.6 and 30 MHz and minimized the interference you caused to tv or the broadcast bands your chances of being caught were practically nil.

By all accounts, the Home Office has little interest in the problem and hopes that it will just go away. An example of its attitude to radio matters can be seen in the recent modification to Amateur licensing conditions announced in the London Gazette. Here we had a government ministry modifying the law of the land by a decree which was so incompetently drafted and carelessly checked that even to the most casual reader several mistakes were immediately apparent. Obviously, these matters are of so little concern to the Home Office that they can be dealt with by the office junior.

I suppose the policing of the radio waves must be a very dull job. But unless the Home Office gets a grip on the problem in the next few months it might just as well give up, abandon the licensing of radio transmitters altogether and open the whole radio spectrum to all comers. Personally, I am not sure that it hasn't started along that path already.

In the event of your using this letter, I would be grateful if you would not publish my address as many of us have found to our costs that criticism of any aspect of c.b. activity usually leads to harassment.
C. G. Howard
Oxfordshire

ALIEN INTELLIGENCE

In your March, 1982 editorial you speak of the lack of response to the Japanese challenge. If your thinking is prevalent in Britain then I can explain why. After all, if Britain were to make a large effort to develop 5th generation computers as the Japanese have announced they will, then it would be "evidently out to destroy all of Western industry". and so, being a decent nation, it doesn't!

Please, no more racist remarks in *Wireless World!*
Erich Unteregelsbacher
Kingston
Ontario
Canada

SELF-HELP TELEVISION

V. Lewis reported in your February issue that a licence had not yet been issued for the Redbrook active-deflector system (Self-Help Television, p.71). In fact, the licence was issued on September 14, 1981.
M. S. D. Granatt,
Home Office.

HIGH-LOSS POWER SUPPLY?

The low-loss power supply shown on page 42 of your February issue is potentially lethal! It is essential that the voltage rating of R_1 is not exceeded, as a failure could put mains voltage on the +12V terminal, having blown up the 470μF capacitor. This usually means that R_1 should be at least two resistors in series. Experience shows that carbon-composition resistors fail to any value – high or low. Carbon, metal-film and metal-oxide resistors almost always fail to a higher value, so should be used in this circuit.

Any mains-supplied circuit should have an appropriate fuse in the live line. A zener clamp (say 22V0) across the 0-12V output would be an additional safety factor. It should be hefty enough to carry sufficient current to blow the fuse in the case of other components failing (1.7 times rated current of the fuse).

Direct connexion of low-voltage equipment to the mains is always potentially dangerous and is best avoided.
R. Jenkins
Cheltenham,
Gloucestershire

COST-EFFECTIVE IGNITION

It is unfortunate that you have published another constructional article – Cooper's article on an electronic ignition system, March, 1982, p4 – which falls into the common trap of using device characteristics that are not specified by the manufacturer.

The devices in question are the 1N4000 series. These are low-frequency rectifiers, and the JEDEC specification to which their numerous manufacturers conform contains no information about their use at at high frequencies. While the devices used by Mr Cooper were evidently adequate, devices from another batch or from another manufacturer could be completely unusable at 15kHz. There is no shortage of devices designed and characterized for use at tv line frequency; for example, the 1N4933 would replace the 1N4001. If a 600V rating is considered sufficient for the s.c.r., D_2 and D_3 could be the BY299, an 800V, 2A fast device.
R. E. Pickvance
ITT Semiconductors
Foots Cray
Kent.

The author replies:
Mr Pickvance is quite wrong to say the article "falls into a trap", because I went to some lengths in the article to show why I used the 1N4000 series diodes and also included a graph to show their limitation, with the advice to use high-frequency diodes beyond 15kHz. Perhaps he hasn't read this part of the article?

Most of the reputable manufacturers (Motorola for instance) actually provide information on the frequency characteristics of the 1N4000 series although it isn't, strictly speaking part of the JEDEC specification. One firm devotes a whole page of its data sheet to this aspect (see enclosed photostat), so they clearly acknowledge that this diode is used at higher frequencies.

The main reason why I chose this diode is because it is readily available to the constructor. Although high-frequency diodes appear in manufacturers' catalogues, they are not available from suppliers such as Electrovalue, Maplin, Marshall's, or Semiconductor Supplies. Nowhere in these firms' catalogues are the 1N4933 or BY299 diodes to be found. To reinforce this point, I rang ITT's own main stockists, VSI (tel 0279-29666) and Nobel (tel 01-309-0500) and asked them to quote for the 1N4933 and BY299. Neither firm had any stocks of the 1N4933 or intended to stock it, and would not even quote a price. Neither firm had any stocks of the BY299, but Nobel did at least quote a delivery (6 to 8 weeks!) and a price of 17p. This latter firm, when they found out the application of the diode, went so far as to recommend the 1N4000 series as a substitute!

However, all the suppliers mentioned sell the 1N4000 series, and at a moderate cost, too – about 50p will buy all five diodes used in this circuit; and this brings me to my third point.

If the circuit were designed around specialized components instead of general-purpose components, the price would go through the roof, and the whole essence of this particular electronic ignition was that it could be paid for out of the savings on one year's motoring. If I uprated the diodes it would make even better sense to substitute the TIP3055 switching transistor, improve the transformer windings, select a military-grade capacitor for C_1 and specify a high-grade thyristor. This I would like to do, but there would be very little economic sense in it; it would no longer be cost-effective.

Finally, I suggest that Mr Pickvance reads *Wireless World* Letters to the Editor for Oct 1975 p.465, and June 1975 p.265 – and indeed anyone intent on criticism, before they put pen to company notepaper, would do well to read these letters.

HEATING-FUEL SAVER

Mr Ryder's central heating fuel saver is pretty obviously going to be cost effective but it does seem to be amenable a further cost-effective modification. This is to add a further thermistor to measure the indoor temperature and hence the temperature difference between inside and out which will better reflect the time needed to reach the desired temperature.

As it stands there is a distinct and obvious weakness in using the device with the popular time clocks providing a gap during the day, long for those at work and shorter for those staying at home. The lowest outdoor temperatures are almost always reached at night and the house has far longer to cool down, whilst during the day it is possible to have quite large solar gains to further diminish indoor temperature reduction. Therefore it obviously makes sense to respond to both temperatures.

What may be less obvious is that a linear type response is not really needed In cold weather the heating system loses more heat to the outside and thus take longer to warm inside, (in the extreme it cannot even reach the desired indoor temperature). If, as seems reasonable, non linearity can be a virtue it does seem feasible to replace R_s, the fixed resistor, with a combination of a thermistor and resistance. the actual choice of values may be made empirically if one has a fairly good idea of the characteristics of one's own heating system performance.

An allied device would be a "time extender" when switching off. The "off" time would be chosen to suit a warm day and the actual close-down would be later according to the temperature outside. One can usually tolerate a few degrees drop but this is a matter very much of personal feelings, and keenness to save fuel. With cast iron boilers and systems having a large water content there is also hot water left after the boiler stops which can be used if the pump is allowed to run longer. Thus two "time extenders" can be of value – one to let the boiler run longer in cold weather and the other a simple fixed "extender" to give about half an hour extra for the pump.

Having suggested that two thermistors are needed to measure the temperature difference it might be worth experimenting, when the time clock has no OFF period during the day, with a thermistor mounted indoors but near to a window so that it is exposed both to the outside and room temepratures. Quite obviously such a thermistor must not be exposed to the sun but the author's choice of a north facing window is unduly restrictive.

As probably 99% of domestic heating systems are just thrown together rather than designed to suit the actual house and the needs of its occupants, it is fairly safe to say that great precision in timing the heating will be uncalled for. The occupants will already suffer from wrongly sized or placed radiators and many other problems, so errors in timing of ± 15 minutes are unlikely to be noticed in terms of comfort.
L. Streatfield
Poole
Dorset

The author replies:
I am obliged to Mr Streatfield for his constructive remarks. If the thermometer facility is not required, then certainly an indoor thermistor could partly replace resistor R_s; or it might be used to modulate the 555 period, via pin 5. With a divide-by-two circuit (such as that of p. 66 *W.W.* Nov. 79) the a.m. and p.m. signals from the time-clock could be distinguished, and the operation modified to suit, for example by switching the 555 timing resistor. The difficulties lie not so much in meeting a particular set of requirements, as in defining the requirements, in the first place.

CARTRIDGE ALIGNMENT

When dealing with tone-arm geometry the tendency is to picture things as they are seen on the turntable and to always include the arc described by the stylus. If instead the stylus/cartridge assembly is imagined to be fixed and the turntable spindle itself moving about the arm pivot, the relative positions of stylus, spindle and pivot are as before but the facts are more clearly illustrated. More importantly, new facts reveal themselves.

Starting from a point representing the stylus, a perpendicular line – a datum line – from which tracking errors may be determined is drawn. Along this line the two zero tracking radii are marked. Through these points an arc with radius equal to the spindle-to-pivot distance is described from a point which, of course, represents the pivot. Any important platter radius may now be marked on the arc directly from the stylus point.

The diagram here is drawn considerably out of scale to avoid crowding. For the same reason lines have been omitted: in an endeavour to avoid confusion, points are symbolized by some letters not customarily employed.

Position of spindle when stylus is on:	
outermost groove,	A
innermost,	B
intermediate radius of high error,	C
any radius (A B etc) included,	R
inner zero tracking radius	p
outer zero tracking radius	q

D, spindle to pivot dist. (arc radius)
L, stylus to pivot
O, offset angle
L-D=overhang=$\sqrt{D^2+pq}-D$
x, angle at R.

$$\sin=\frac{R^2+L^2-D^2}{2LR}=\frac{R^2+pq}{2LR}.$$

When this is applied to p,

$$\sin=\frac{p^2+pq}{2Lp}=\frac{p+q}{2L}=\sin O$$

Similarly with q. $((p+q)/2L=\sin O$ is clear from the diagram). When applied to C,

$$\sin=\frac{C^2+pq}{2LC}.$$

Now $\sqrt{pq}=C$ therefore $pq=C^2$ and

$$\frac{C^2+pq}{2LC}=\frac{2C^2}{2LC}=\frac{C}{L}=\sin x \text{ at } C.$$

To clarify this, it can be seen from the diagram that $C^2=L^2-D^2$. Now if we join point q to the pivot point, a triangle is completed one side of which is common to another triangle whose hypotenuse is L, because of this

$$D^2-\left(\frac{q-p}{2}\right)^2=L^2-\left(\frac{p+q}{2}\right)^2,$$

therefore

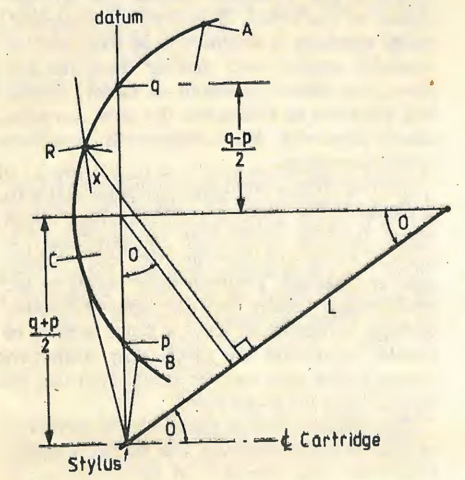
$$L^2-D^2=\left(\frac{p+q}{2}\right)^2-\left(\frac{q-p}{2}\right)^2=pq.$$

(Quickly proved by substituting figures for p and q.)

It is useful to note that while the magnitude of the tracking errors (difference between x and O) depends on the values of D and L, their proportions depend on the zero tracking radii. When p and q are 66 and 121, for instance, the errors at A and B are 1.7 and 0.7 of that found at C. When p and q are 49 and 110, at A the error is double that found at C, while at B it is two thirds. If a diagram is drawn to scale showing only the arc, datum line and points A, B and C joined by straight lines to the stylus point, tracking errors might be measured directly with a protractor. R. J. Gilson's factors would place B on the other side of the datum line.

It follows from the foregoing that $p+q=\sin O2L$. This facilitates the process of calculating the zero tracking radii in a case such as that dealt with in Gilson's final paragraphs page 64, *Wireless World* Oct. 1981. After finding O with his formula 4(b), find $p+q$ from this equation. Then p and q can be found from $pq/p=(p+q)-p$. There seems to be quite a bit of latitude for rounding off the results while ensuring negligible changes in the values of L, the intermediate radius and all angles.

P. E. Cryer,
Thornlie,
Western Australia.



THE NEW ELECTRONICS

I have every sympathy with Hugh Jacques article in your January issue – and I certainly do not find low standards in Germany an excuse for our own low standards, as C. Wehner's letter in the April issue seems to imply (in part at least).

I am now a secondary school teacher of physics and have been appalled at the philosophies built into education; standards here are definitely falling – but a whole re-shuffle of aims and objectives and a change in examination syllabuses and in the exams themselves all combine to camouflage the drop in standard. I have often wondered when this fall in standard was going to affect university standards and higher up. Mr Jacques article confirms my fears.

What with a philosophy that views the child in terms of its needs instead of in terms of its responsibility and society's expectations from it – there has developed the sort of approach which has the following characteristics:

1) educationally – the child considered in terms of its needs must be given automatic promotions to prevent any sense of inferiority, frustration or maladjustment;
2) socially – the same child must be guaranteed cradle-to-grave security lest a trauma be produced;
3) the cure for failure to learn is to devaluate learning and the cure for social failure is to devaluate success.

I trust this will give food for thought for concerned parents and then, perhaps, lead them to action.

S. Georgeoura
Ardgay
Ross-shire

WOODPECKER

Mr Martinez' letter, (April), gives an interesting and quite possibly correct explanation of the Russian "Woodpecker" transmissions. There are one or two points arising from his letter.

The suggestion that the code auto-correlation, *i.e. the "compressed" radar signals, would have virtually no sidelobes may be a little optimistic. One might expect, in a practical system, that the peak signal sidelobes would not be more than about 25 dB at best below the main lobe. One would also have to examine the ambiguity functions of these signals to determine their properties in the range-Doppler domain where their sidelobe performance might be rather different if the radar were used to detect high-velocity incoming targets.

Another point arises from the statement that the compressed signal would have "31 times the amplitude . . ." etc. The equalizer, (i.e. the matched filter), would theoretically conserve signal energy and its peak output would have 31 times the peak *power* of the uncompressed signal, not 31 times its amplitude.

Finally the statement about the radar having 31 times the "sensitivity" of a 100μs radar of the same power, should be interpreted with caution. Two radars of differing pulse durations but of the same *mean* power, and having properly matched filters in the receivers, would have the same "sensitivity". Their difference in the present context would, as Mr Martinez states, lie in their resolution capability. Pulse compression, as such, does not introduce some mysterious improvement in system sensitivity; with matched filter receivers, whatever the transmitted pulse duration, the "sensitivity" remains a function of the ratio of the received signal *energy* to noise power spectral density.

* (strictly the *cross*-correlation function of the transmitted signal with that received, taking account also of any "weighting" which might be used to improve signal sidelobe levels, albeit at some expense to resolution.)

M. G. T. Hewlett
Midhurst
West Sussex

THE FUNCTION OF FUNCTIONS

I was interested to read Thomas Roddam's remarks (*Wireless World*, December, 1981. p. 37) concerning the notion that used to be fairly prevalent, that denies the existence of sidebands in amplitude modulation. After all, with "pure" amplitude modulation the number of cycles per second of the wave remains constant whether it be modulated or not, doesn't it? Be it said that the idea is not entirely dead even yet; there are still people to be found who hanker after it. And it may be said that they are in tolerably good company, too, as anybody may see for themselves by consulting the files of **Nature** for 1930 (pp.92-3, 198-9, 271-3, 306-7, 726-7) in which Sir Ambrose Fleming, no less, categorically denies the existence of sidebands, declaring on the contrary that they are but a mathematical fiction, and stubbornly refusing to accept correction from his colleagues.

The curious thing about it all is that the sideband-deniers have never had any difficulty in accepting that a baseband signal occupies finite spectrum space, not realising, of course, that a baseband signal is but two (superimposed) sidebands, "centred" on zero frequency. A simple thought-experiment: displace the carrier frequency progressively upscale from zero and observe the two sidebands separating out.

And consider, furthermore, that proper reconstruction of a baseband signal to (say) audible form requires re-insertion of the zero-frequency carrier, e.g. in the polarizing firld of a loudspeaker or telephone receiver.

D. C. Sutherland
Wanganui
New Zealand

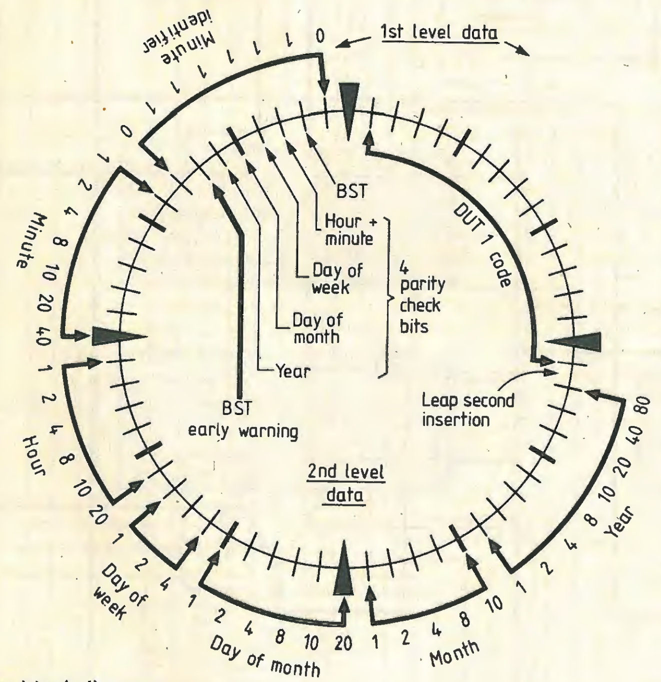
MICRO-CONTROLLED RADIO-CODE CLOCK

Several standard-frequency transmissions throughout the world provide time and date information controlled by caesium atomic clocks, with potential for automatic time and date information at reasonable cost. This design offers a compromise between economy and complexity suitable for both non-critical professional applications and domestic use.

by N. E. Sand

The 60kHz standard-frequency transmission from Rugby MSF now includes fast and slow time codes, both of which provide full time and date information once every second. The signal is transmitted 24 hours every day except for a maintenance period on the first Tuesday of each month. The transmitter power is 50kW e.r.p. which, with the long wavelength, provides propagation over a range of several hundred miles. With careful circuit design, useable reception can be achieved throughout Britain, but because there is a skywave and groundwave component, certain areas can experience cancellation or addition where mixing takes place. This problem is complicated because the areas of mixing change from daytime to nighttime as the ionosphere changes.

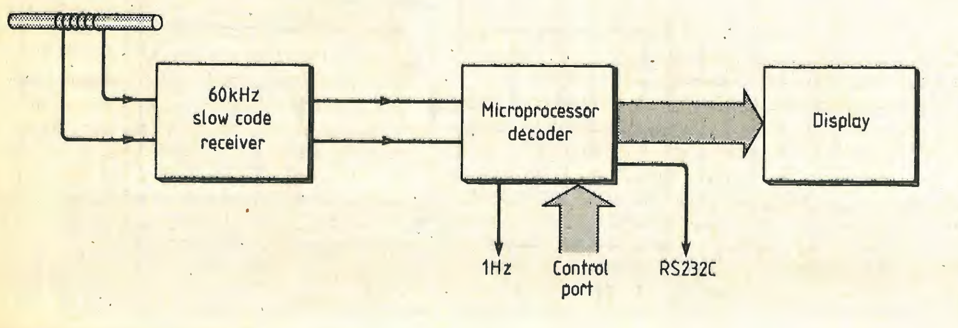
Fig. 1. Slow-code format from Rugby MSF.



Successful reception has been achieved with specialized equipment at over 3,000 km from the transmitter, but with simpler designs 750 km is a more realistic range.

This design uses the slow code, which for most applications provides better results and requires less critical timing. The slow code format shown in Fig. 1 extends from second 17 to second 59 each minute. A logic zero is represented by a carrier break of 100ms and a 1 is represented by a break of 200ms. Other information, such as parity bits, is represented by a carrier break of 100ms displaced by 200ms from the start of the second. To synchronize with the serial code it is necessary to recognize the start of each minute, so an identifier sequence from second 52 to 59 is provided.

Fig. 2. Hardware block diagram.



The most important part of the design, as shown in Fig. 2, is the receiver which must be capable of producing a consistent output in the presence of noise. Unfortunately there are several common sources of interference at 60kHz, for example the harmonics from the line output transformer of colour television receivers are powerful sources of interference for v.l.f. transmissions, as well as fluorescent tube fittings and even hand-held calculators. Carefully designed t.r.f. and phase-locked loop receivers can work satisfactorily at moderate gains, but commonly suffer from self pick up which limits their sensitivity. In the case of the p.l.l., radiation from the v.c.o. will generally weaken the signals. For best results a low-current superheterodyne receiver should be used, but it is costly and difficult to adjust for optimum noise performance at v.l.f.

An alternative receiver which is much simpler and with careful design can produce acceptable results is shown in Fig. 3. The input stage uses a cascode circuit which enables the antenna coil to be directly connected without a transformer winding and gives good stability at high gain by minimizing Miller feedback. A multiplier* is used for low-level detection of the 60kHz carrier, a technique which avoids high levels of 60kHz and therefore enhances receiver stability. The multiplier generates a double frequency component and a differential voltage proportional to the carrier level at the two load resistors. The double frequency output is ignored by the following amplifier stage which produces the demodulation carrier. To minimize power requirement, operating current in each arm of the multiplier is set to 50μA. To avoid drift at the multiplier output the potentiometer should be a ten-turn cermet type and the two load resistors metal film.

To optimize receiver performance for all signal levels at the antenna a normal gain-control loop to the cascode stage would provide satisfactory results. However, a better method is to gate the a.g.c. loop because the 60kHz carrier is 100% modulated and will cause errors in a conventional loop. In this design an undelayed signal from the output of IC_2 switches the

* Self-setting time code clock, by N. C. Helsby, *Wireless World*, August 1976, pp?

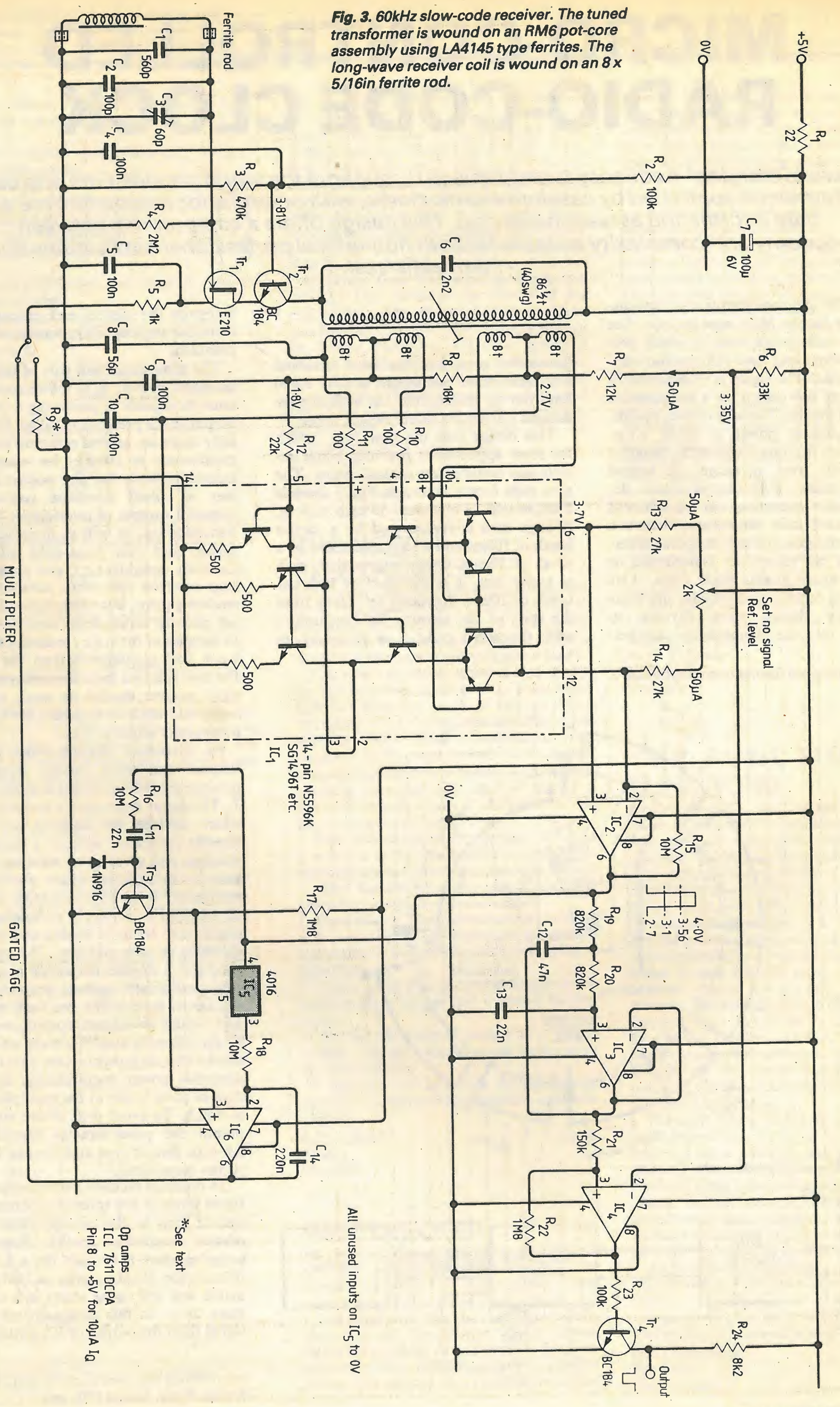
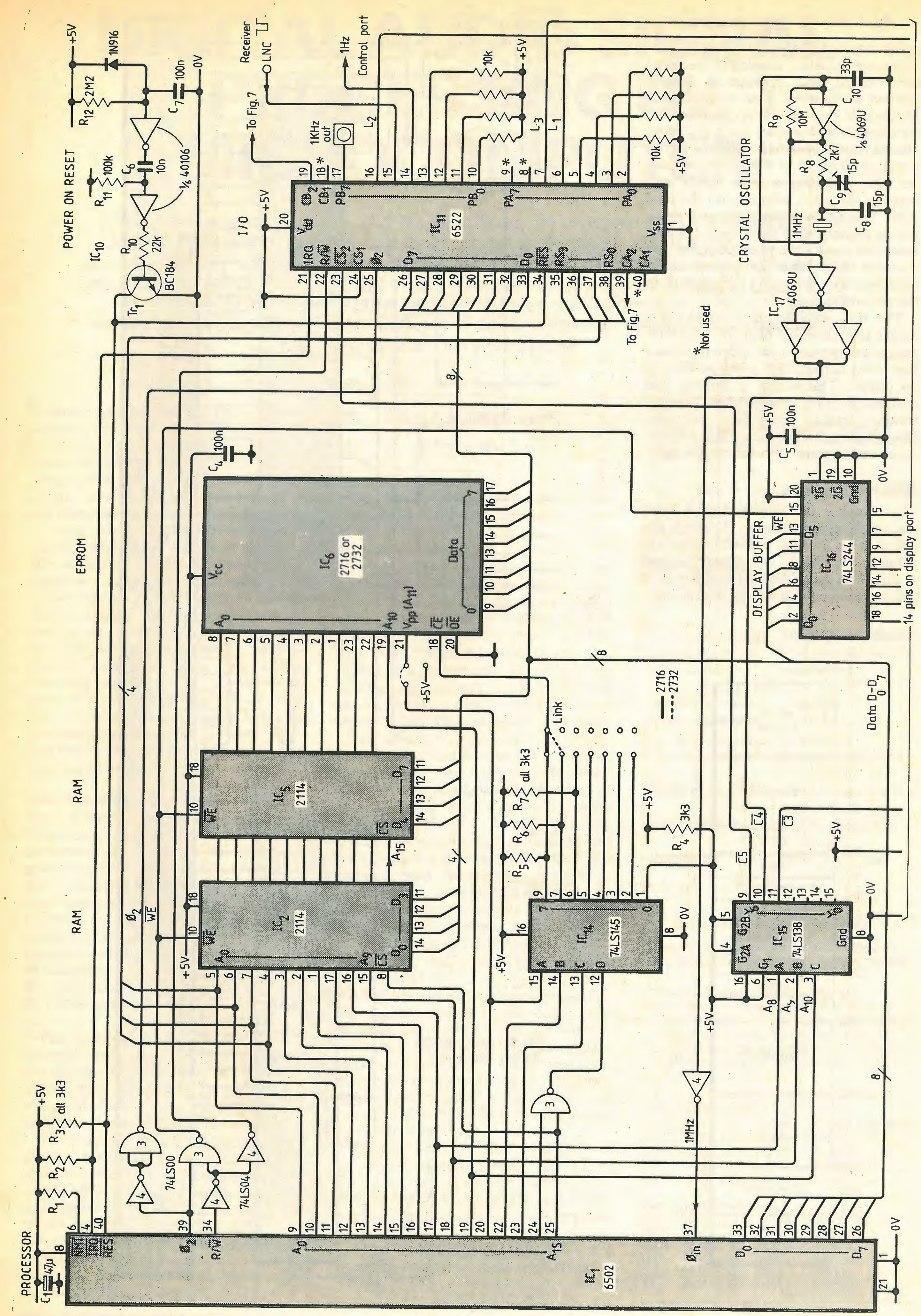


Fig. 3. *60kHz slow-code receiver. The tuned transformer is wound on an RM6 pot-core assembly using LA4145 type ferrites. The long-wave receiver coil is wound on an 8 x 5/16in ferrite rod.*



◀ *Fig. 4. Microprocessor decoder*

gating control for the a.g.c. amplifier IC_6. Transistor Tr_3 is a.c. coupled to the undelayed signal and provides a simple time-out arrangement if no modulation is detected. If this occurs Tr_3 remains off permanently and enables the a.g.c. loop. During normal operation, when the carrier is pulsed off for up to 500ms, Tr_3 remains on, the transmission gate opens and switches the a.g.c. integrator to the hold state. This action prevents the gain rising during the normal off periods which would otherwise cause noise to be amplified and excessive overshoot in the received carrier level, which in turn causes timing errors in the demodulated signal.

The d.c. amplifier is followed by a simple active low-pass filter and a Schmitt-trigger level sensing stage which provides a logic level output, high corresponding to no carrier. This output is inverted and buffered for t.t.l. compatibility. Complementary m.o.s. apmplifiers are used throughout to reduce the current requirements and dispense with dual supply rails.

Micro decoder

The complete decoding system is based on a 6502 microprocessor. Although this device is not the most powerful in terms of number crunching, its memory-mapped architecture and addressing modes make it an ideal choice for industrial control appli-

Continued on page 59

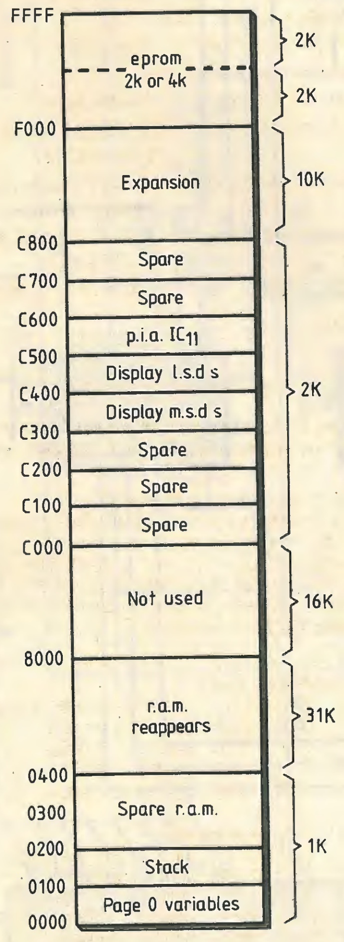


Fig. 5. Memory map.

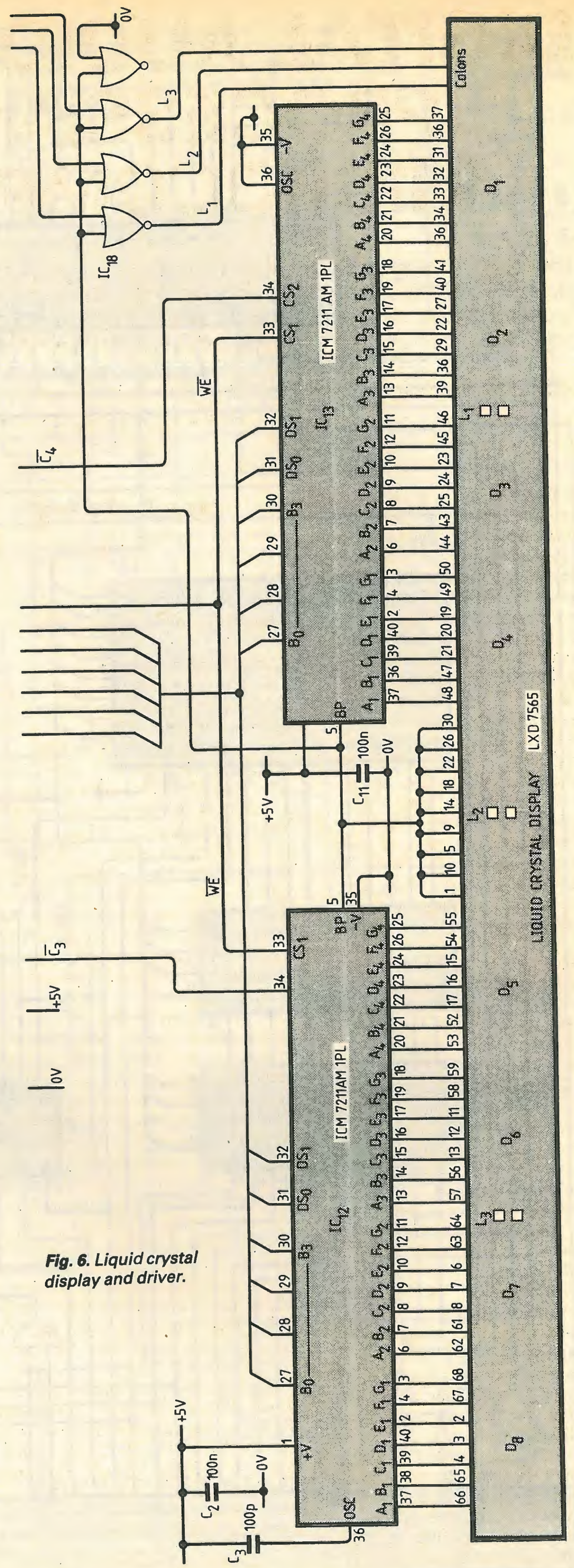


Fig. 6. Liquid crystal display and driver.

DIGITAL AUDIO SIGNAL PROCESSING BY MICROCOMPUTER

The author suggests that audio-frequency signal processing is the most recent area in broadcasting to benefit from digital technology. He lists the currently accepted digital sampling characteristics which limit the reduction of programme modulation noise, idle-channel noise, distortion and wow and flutter. The article includes a brief review of the development of microcomputers, compares analogue and digital companding and describes how companding is affected by use of a microprocessor.

by J. B. Watson*

B.Eng., M.I.E.E.

From the point of view of the broadcasting industry, there can be no doubt that digital technology carries with it numerous benefits to both listener and viewer, more than compensating for the additional complexity and expense of the signal origination and transmission plant involved.

Those who possess or have heard demonstrated audio discs cut from a digitally recorded master tape can confirm the significant improvements in clarity and fidelity now obtainable. This is in spite of the fact that the digital processing component here represents only a small part of the total recording/reproduction system. Further developments in laser optical recording may, in the future, finally resolve the contentious question of what truly constitutes 'high fidelity' audio by providing, for all practical purposes, absolute fidelity between recording studio and listening auditorium.

Digital audio characteristics

Reasons for the subjective superiority of digital audio over the traditional analogue equipment are its much reduced programme-modulation noise, idle-channel noise, distortion and wow and flutter. The magnitudes of these improvements are determined by the digital sampling characteristics, minimum standards for which are generally accepted to be, for broadcasting:

- Sampling rate: 32 000 samples/second.
- digital resolution: 14-bits/sample
- Pre/de-emphasis: CCITT characteristic.

Pre-emphasis and de-emphasis are not essential with digital audio, but give worthwhile improvements in high-frequency noise performance for signals possessing limited energy at the upper end of the audio spectrum.

These standards provide for an audio-channel bandwidth to 15 kHz (provided that great care is taken in the manufacture of the anti-aliasing filter), and a signal/-noise ratio of better than 85 dB. The needs of the recording studios, where the final output is derived from a large number of independent sources, are more stringent; hence the use of 16-bit sampling at rates between 48 000 and 64 000 per second in order to minimize signal degradation through the mixing processes.

ADC considerations

The following specification figures underline the reasons for the relative delay between the general acceptance of digital video and digital audio technology. For example, the 'aperture time' required of an 8-bit, 5 MHz video sampling circuit, is only 20% shorter than that for a 16-bit 15 kHz audio sampler; but the precision of the audio circuit needs to be better by a factor of 256. Thus, although proprietary video analogue-to-digital converters (a.d.cs) have been obtainable for the past three or four years, audio a.d.cs with adequate performance have only recently become available. Table 1 shows the specification of such a device. Although sufficiently accurate and sufficiently fast for broadcast-quality audio, this a.d.c. (in common with many other proprietary units) has the disadvantage of "offset-binary' digitally-coded outputs (see Table 2). This means that the most critical zone in its transfer characteristic occurs at the mid-point, where the digital output changes from 011 . . .111 to 100 . . .000. Since this is normally the quiescent operating region, careful circuit layout and screening are essential to minimize digital-analogue crosstalk. A more suitable coding technique for digital audio is the sign-plus-magnitude arrangement, shown in Table 2, which has the effect of moving the most critical region away from the quiescent operating point.

Digital-to-analogue converters (d.a.c.s) are usually less costly than the corresponding a.d.cs, but can introduce non-linearity into the audio channel if their output circuits are 'slew-rate limited.' This problem disappears if the d.a.c. output is re-sampled by a sample-and-hold device designed specifically for audio applications.

*Independent Broadcasting Authority

Table 1. Specification for a proprietary a.d.c. device suitable for broadcast-quality digital audio.

Analogue-to-digital converter Type MP8016, Analogic:

input voltage range:	−10V to +10V bipolar
input impedance:	5.0 kΩ
digital resolution:	16 bits
relative accuracy:	±0.0015% of full scale
absolute accuracy:	±0.003% of full scale
quantizing error:	±½ least sig. bit
monotonicity:	guaranteed
recommended calibration interval:	6 months
Warm-up time to specified accuracy:	10 minutes
conversion time:	0.6 to 2.0 µs per bit (adjustable)
digital output code:	offset binary or two's complement
power supplies:	±15V, 65 mA and +5 V, 300 mA
dimensions:	102×77×13 mm

Table 2. Comparison of offset-binary and sign-plus-magnitude a.d.c. output codes

The offset binary code is easier to implement in hardware, and is the one most commonly used in proprietary a.d.cs. The sign-plus-magnitude code may be more suitable for an audio a.d.c, but suffers the disadvantage of two equally-valid codes for zero input.

Offset-binary code:	
1111 1111 1111 1111=	+9·9997 volts input
1000 0000 0000 0001=	+0·0003 ,, ,,
1000 0000 0000 0000=	0·0000 ,, ,,
0111 1111 1111 1111=	−0·0003 ,, ,,
0000 0000 0000 0000=	+10·0000 ,, ,,
sign + magnitude code:	
0111 1111 1111 1111=	+9·9997 volts input
0000 0000 0000 0001=	+0·0003 ,, ,,
0000 0000 0000 0000=	0·0000 ,, ,,
0000 0000 0000 0000=	0·0000 ,, ,,
1000 0000 0000 0001=	−0·0003 ,, ,,
1111 1111 1111 1111=	−9·9997 ,, ,,

Table 3. Comparison of three generations of microprocessor from a single manufacturer

The 8088 is a development of the 16-bit 8086, but communicates via an 8-bit data bus.

	'First generation' (p-mos technology)			'Second generation' (n-mos technology)		'Third generation' (h-mos technology)		Future
microprocessor type	4004	8008	4040	8080	8085	8086	8088	iAPX 432
date introduced	1971	1972	1974	1974	1976	1978	1979	1981
accumulator capacity	4-bits	8-bits	4-bits	8-bits	8-bits	16-bits	16-bits	32-bits
no of instructions	46	48	60	111	113	300+	300+	very many
min. instruction cycle time	10·8 μs	20 μs	10·8 μs	2 μs	1·3 μs	0·4 μs	0·4 μs	0·1 μs
memory addressing capability	4 K×8	16 K×8	8 K×8	64 K×8	64 K×8	1 M×8	1 M×8	4000 M×8
no of general purpose registers	16×4-bit	7×8-bit	24×4-bit	7×8-bit	7×8-bit	12×16-bit	12×16-bit	?
power supplies	+5V, −10V	+5V, −9V	+5V, −10V	±5V, +12V	+5V	+5V	+5V	+5V
sub-routine nesting levels	3	7	7	unlimited	unlimited	unlimited	unlimited	unlimited
no of interrupt types	0	1	1	8	12	256	256	none
interrupt latency (approx.)	–	40 μs	30 μs	6 μs	4 μs	12 μs	12 μs	–
address, data bus width	12 addr. 4 data 4 r.a.m. sel	14 addr. 8 data	12 addr. 4 data 4 r.a.m. sel 2 r.o.m. sel	16 addr. 8 data	16 addr. 8 data	20 addr. 16 data	20 addr. 8 data	32 addr. 32 data

Signal processing by microcomputer

Before considering in detail the type of audio signal processing appropriate to microcomputers, it may be instructive to consider the history of the development of these devices since their introduction about eight or nine years ago. Table 3 summarizes the characteristics of three 'generations' of microprocessor originating from a single manufacturer, the demarcations corresponding loosely with the fabrication of the silicon chip by p-channel, n-channel and h-mos technology. Details of a prospective 32-bit device are also included.

Of particular relevance to audio processing are the accumulator capacity (16-bits being desirable), instruction cycle time and interrupt response time (latency). First generation devices (4004, 4040, 8008) were relatively slow in operation, offering instruction times of several microseconds, and possessing extremely limited – or, in the case of the 4004, non-existent – interrupt handling capability. Real-time processing of broadcast quality digital audio signals was not feasible with these processors, although the 8008 did find application in the military communications field as a speech processor.

Second generation processors, such as the still-current 8085, exhibit speed improvements of an order of magnitude over their predecessors. With an instruction cycle time of 1.3 microseconds, between 12 and 24 simple processing steps can be undertaken in the interval separating adjacent audio samples (this varies from 30 to 15 microseconds, for sampling rates of between 32 000 and 64 000 per second). Interrupt latency is of a low order, allowing rapid re-programming of peripheral devices after the treatment of blocks of data. Self-contained hardware multiply and divide features, however, are not typical of second generation microprocessors. Complex digital filtering operations requiring weighting factors related by integral powers of two. On the other hand, digital audio companding can be implemented with this type of processor, since the essential feature of a companding algorithm is a multiple-bit shifting operation.

Extremely powerful processing possibilities are now available, following the introduction of the 8086 and other similar devices such as the Z8000 and M68000. These computers operate directly on 16-bit sampled data values, have fast and comprehensive instruction repertoires including hardware multiply and divide, and can directly address memory arrays in excess of a million bytes. The very complex nature of these machines, however, necessitates some compromise in performance in the real-time environment, the most serious being an interrupt latency approaching 15 to 30 μs sampling interval of high-quality audio.

This arises from the extended memory addressing and interrupt type handling capabilities. The 8086 computer, for example, responds to an interrupt by storing the current program counter and memory segment register (16-bits each), calculating the location of the appropriate interrupt vector, and reloading new program count and segment values from the vector location. Flexibility of operation is thus achieved at the expense of speed, the complete process occupying approximately 12 μs. This trend is likely to continue with the emergence of new device types (e.g., iAP 286) capable of directly addressing thousands of megabytes of memory. Future microprocessors designed for 'mainframe' applications might dispense entirely with interrupt facilities, since these prejudice the 'number crunching' performance.

The solution to this problem, in the critical real-time signal processing area, is in the use of dedicated input/output processors. These are intended specifically for rapid peripheral device servicing duties, and communicate with the main processor system via block transfers on direct-memory-access channels. Operating speed improvements anticipated with the h-mos process are also likely to simplify the task of the digital audio engineer.

Digital companding

An example of the type of digital audio processing now possible with microcomputers occurs in signal compression and expansion (companding). Analogue companding is used extensively in the magnetic tape recording of music, where it provides an improved signal/noise ratio, particularly at high frequencies. The widespread acceptance of the Philips audio cassette as a satisfactory medium for domestic sound recording is, in fact, largely due to the adoption of Dolby or similar companding techniques.

Analogue systems of this type separate the incoming audio signals into several frequency bands, each channel during recording being compressed by amounts depending upon the peak levels present in each appropriate spectral band. The accuracy of the reciprocal expansion process during playback is seldom perfect, since great reliance is placed on carefully matching the filters, time constants and level-dependent amplifiers, in the record-replay chains.

The processes involved with digital companding are, on the other hand, accurately reversible. The degree of compression applied by the encoder is transmitted, together with the audio sample values, along a separate time-multiplexed channel to the decoder. No gain errors or time constant mismatching errors occur, but a degree of programme-modulation noise is introduced, the magnitude of which is governed by the type of digital companding used. It might seem paradoxical that digital companding apparently worsens the audio signal/noise ratio, whereas analogue companding improves it. This arises from the different areas of application in which the two techniques find justification, digital companding being used in bandwidth reduction rather than noise reduction. Alternatively, the companding process can be regarded as a 'trade-off' between idle-channel (background) noise and modulation noise in a channel of defined capacity or bandwidth. This 'trade-off' takes place in both analogue and digital companding, as sensitive ears can easily recognise, and there is, therefore, no paradox.

Table 4. 14: 10 A-law companding

14-bit audio samples are compressed before transmission to a 6-bit representation (X's in the Table), the less significant bits being truncated (shown as T's in the table). A 3-bit scale factor indicating the audio level within a range of 2: 1 is appended. Finally, the 'sign' bit (13) is added, making 10 bits altogether. A refinement of the procedure is to 'round-up' the levels by one half least significant bit to improve accuracy.

A-Law range no.	Scale factor			Bit no. 13 (m.s.b.)	12	11	10	9	8	7	6	5	4	3	2	1	0 (l.s.b.)
7	1	1	1	0	1	X	X	X	X	X	X	T	T	T	T	T	T
6	1	1	0	0	0	1	X	X	X	X	X	X	T	T	T	T	T
5	1	0	1	0	0	0	1	X	X	X	X	X	X	T	T	T	T
4	1	0	0	0	0	0	0	1	X	X	X	X	X	X	T	T	T
3	0	1	1	0	0	0	0	0	1	X	X	X	X	X	X	T	T
2	0	1	0	0	0	0	0	0	0	1	X	X	X	X	X	X	T
1	0	0	1	0	0	0	0	0	0	0	1	X	X	X	X	X	X
	0	0	0	0	0	0	0	0	0	0	0	X	X	X	X	X	X

Many different types of digital companding have been proposed, these being within two main categories of 'instantaneous' and 'near-instantaneous' (or 'quasi-instantaneous') form. Table 4 depicts the operation of a simple 'A-Law' instantaneous compander, used for reducing to 10-bits per sample a 14-bits per sample audio signal. The companding algorithm is simply expressed as follows: The position of the most significant '1' digit in the 14-bit data word is a measure of the signal magnitude within a 2:1 range. A 3-bit digital code representing the number of leading (more significant) '0' bits in each word is transmitted with the word, the leading zeros and the most significant '1' being suppressed. The remainder of the digital word is transmitted with a precision of seven bits, with the less significant bits truncated.

The resulting signal comprises ten bits per sample, three bits of which define the approximate value, within a 2: 1 range, and seven the more precise magnitude within that range. One of the seven bits indicates the signal polarity. Digital sample values with more than five leading zeros experience no compression and are therefore expanded without loss of resolution. For signals of larger magnitude, the compression takes the form of a truncation of the less significant bits, giving rise to coarser quantizing steps and associated programme modulation noise. Listeners accustomed to the very high sound quality reproduced by digital systems have observed that, with experience, their threshold of tolerance to modulation noise tends to decline. Acceptable standards of performance based on subjective tests may, therefore, become more stringent as the use of digital audio equipment increases.

'Near-instantaneous' companding makes more efficient use of the available channel capacity, and thus achieves a lower level of modulation noise. By deriving scale factors related to the peak amplitudes of groups of audio samples, rather than to individual sample values, more capacity is available for accurately resolving the signal levels. In two systems recently proposed [2,3], groups of samples representing a duration of approximately 1 ms are examined before defining the degree of companding appropriate to the group as a whole. Subjective comparison tests of various digital companders tend to the conclusion that the 'near-instantaneous' principle provides a standard of performance equivalent to that of an 'instantaneous' system employing at least one more bit per sample. Many of the differences between the various contending schemes proposed for international programme exchange arise from attempts by their proponents to make maximum use of the different hierarchical levels available on Posts, Telegraphs and Telephones (PTT) networks. Until detailed plans for digital sound channels are published by the PTT authorities, however, an optimum transmission system is unlikely to be realised. Tariff structures are likely to play a more important role than technical reasons in deciding which form of the companding, if any, should be universally adopted.

Companding by microprocessors

Since both 'instantaneous' and 'near-instantaneous' companders employ the same basic principle in deriving their respective scale factors, i.e., the determination of the position of the most significant bit in a digital code, it is possible to devise hardware configurations capable of operating in either mode.

Figure 1 illustrates a digital audio compression system used in the laboratories of the IBA to evaluate the performance of a variety of companding algorithms.

The system is designed as a peripheral interface to an 8-bit 8085 microcomputer, the 14-bit audio samples requiring two operations per processing step because of the 8-bit accumulator limitation. A priority encoder device (see Table 5) with complemented data inputs locates the position of the more significant '1' bits in the sampled data, and generates a 3-bit code related to the compression scale factor. Because the chosen a.d.c. produces an 'offset binary' output, positive values require to be complemented before reaching the priority encoder. This is effected by applying the most significant bit to a set of exclusive 'Or' gates. Samples from a 'sign-plus-magnitude' converter would not require correction in this manner.

The audio samples are 'left-justified' to remove leading zeros by means of a multiplying technique. By decoding the scale factor in the 3: 8-line decoder shown in Fig. 1, a digital number of the form 2^n is produced. This number is applied to the B inputs of a pair of 8 × 8 parallel multiplier chips, causing a left shift (by n places) of the data samples entering the multiplier A inputs. Finally, the two partial products resulting from the 8 × 88 multiplying operation are combined, and the appropriate number of less significant bits truncated. Strictly, the multiply combiner should take the form of a digital adder circuit, but a simple Or gating arrangement suffices since one of the operands is of the type 2^n. The circuit described can perform either on a sample-by-sample basis, thus providing an instantaneously companded A-law output, or it can generate scale factors derived by the computer from peak measurements of groups of samples, resulting in near-instantaneous companding.

Figure 2 shows the microprocessor configuration, based on the 8085 computer and standard memory and input/output devices. The interval between audio samples is 31.25μs (32 kHz sampling), permitting a reasonable amount of signal processing on a per-sample basis, bearing in mind the 1.3 μs instruction cycle time of the 8085. For example, the peak value of a group of audio samples can be calculated by complementing negative values, Or-ing each word with previous samples, and scanning the resultant value at the end of a cycle to determine the scale factor for the group. Insufficient processing power is available from the 8085 system to perform A-law companding by software, hence the derivation of 'instantaneous' scale factors

Table 5. Transfer function, 74148 Priority encoder device

The device generates a 3-bit output code indicating the position of the most-significant 0 bit in an 8-bit word. It can be used as the basis of a digital compander if its input signals are complemented.

Inputs 7	6	5	4	3	2	1	0	Outputs A2	A1	A0
0	X	X	X	X	X	X	X	0	0	0
1	0	X	X	X	X	X	X	0	0	1
1	1	0	X	X	X	X	X	0	1	0
1	1	1	0	X	X	X	X	0	1	1
1	1	1	1	0	X	X	X	1	0	0
1	1	1	1	1	0	X	X	1	0	1
1	1	1	1	1	1	0	X	1	1	0
1	1	1	1	1	1	1	0	1	1	1

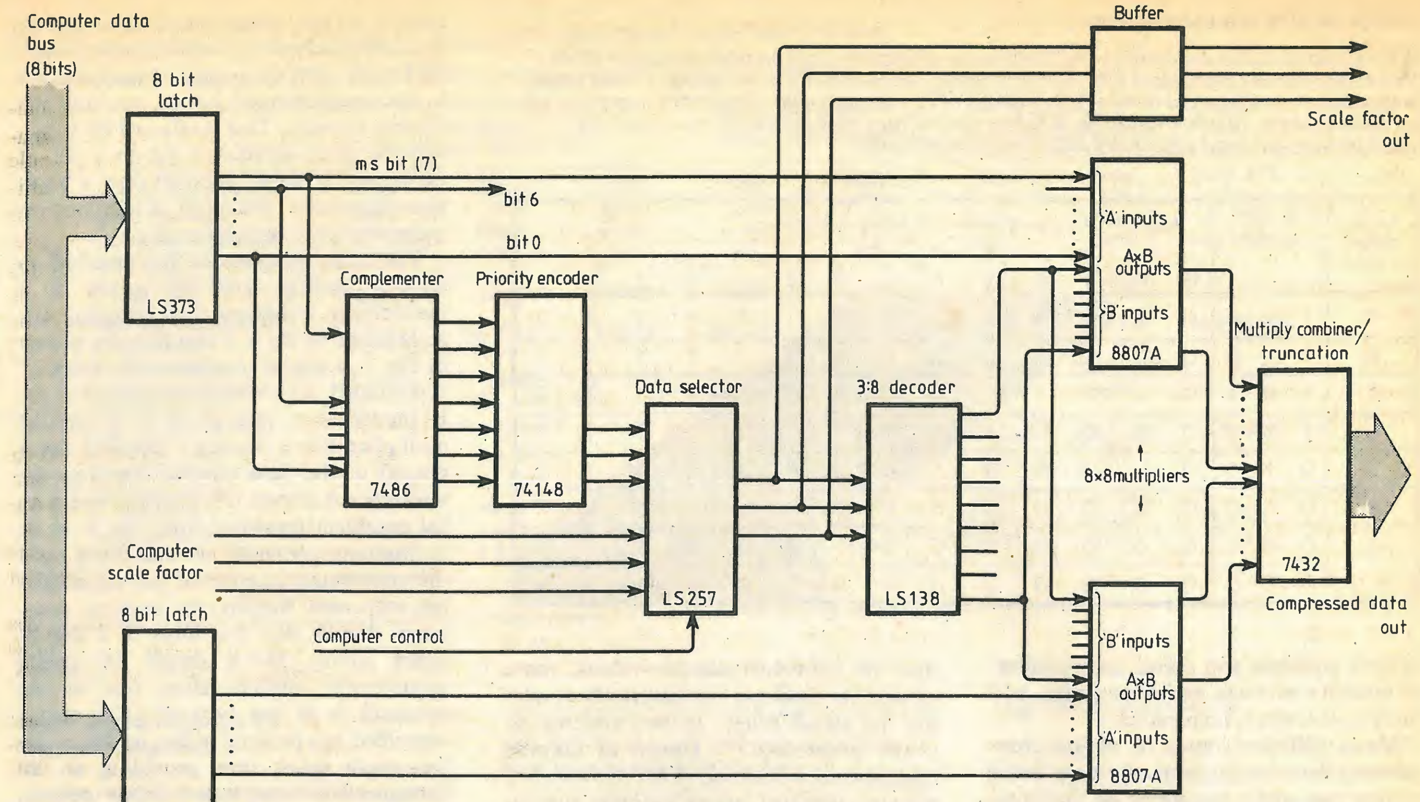
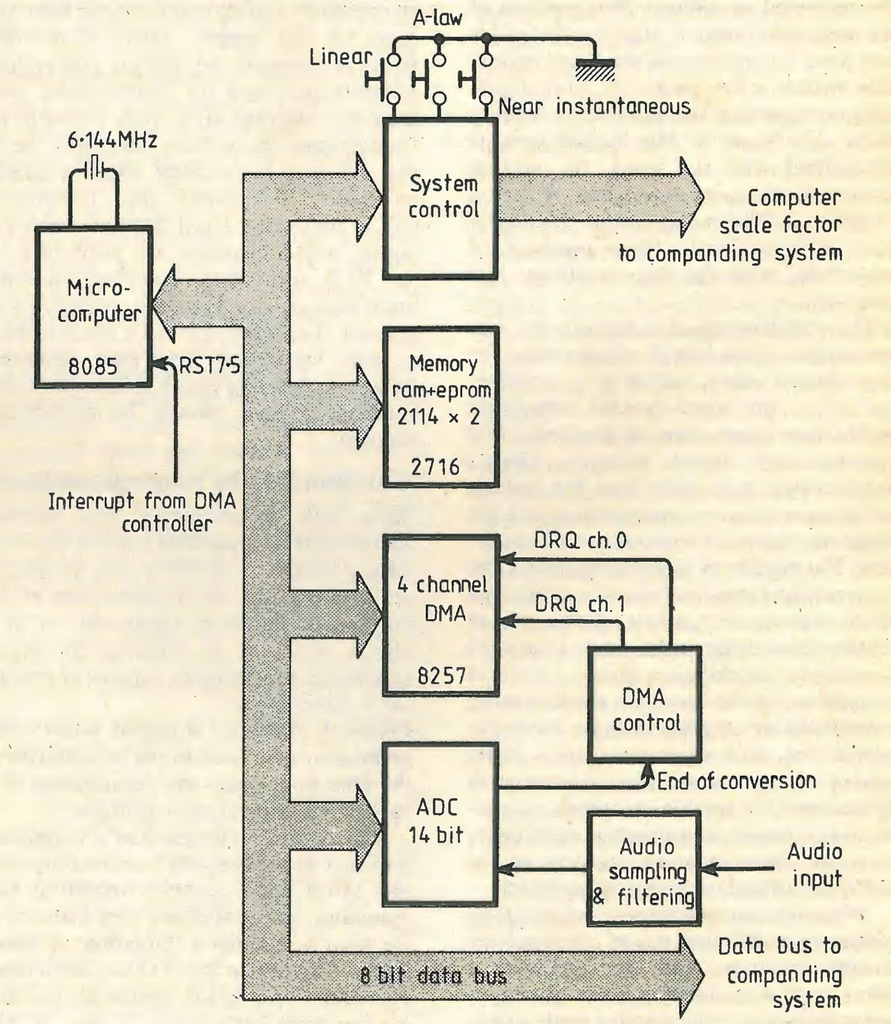


Fig. 1. Digital audio compression system. Compression is effected by shifting the digital samples left (towards the most significant bit) by an amount equal to the number of leading zeros in the data value. The 74148 priority encoder locates the position of the most significant '1' bit and produces a 3-bit code which is then decoded by the LS138 and applied to two 8 × 8 multiplier chips. These multipliers, together with the multiplier combiner, shift the input data word by the required amount.

Fig. 2. Microprocessor system for digital audio companding. Digital companding experiments can be undertaken with a relatively simple microprocessor system. The hardware configuration shown employs 1K x 8 r.a.m. and 2K x 8 e.p.r.o.m. 14-bit audio samples are processed as 8-bit word pairs.

by the 74148 priority encoder in Fig. 1. However, the usual benefits of software control, including versatility, cheapness and reliability, apply to the remainder of the data processing hardware. An interesting feature of this arrangement is that less hardware is required for 'near-instantaneous' companding than for the A-law algorithm; i.e., a reversal of the usual situation.

The system software operates on three distinct blocks of data, audio sample words originating from the a.d.c., stored blocks awaiting analysis, and output data for the expander logic arrangement (not shown, but similar in concept to the compression circuit of Fig. 1). Both input and output data streams communicate directly with memory via direct-memory-access (d.m.a.) channels. D.m.a. management is performed by the 8257 chip, which receives 'data requests' from the currently active peripheral and responds with 'data acknowledge' when the computer has disabled its bus signals. Data transfer then takes place, and the memory pointer for the channel in use is incremented. A time penalty of about one microsecond is incurred while the computer data and control busses 'freeze' during the transfer, in comparison with the several microseconds necessary for conventional input/output procedures. However, because of the automatic address incrementing performed by the d.m.a. controller, new address pointers have to be entered before the selected location exceeds the available memory space. It is convenient to manipulate the audio samples in blocks, and to re-initialise the d.m.a. controller at the end of each block. The execution time of the software routine controlling this function must, of necessity, be shorter than the audio sampling interval, otherwise samples could be lost. This segment of code is therefore designed around register manipulation instructions which are much faster than memory accessing operations. Software for the complete system, capable of rapidly switching between 14-bit linear (uncompanded), 14: 10 A-law and 14: 10 near-instantaneous companding algorithms, occupies less than 1000 words.



Subjective tests of the system described have confirmed that the effect of 14: 10 digital companding, whatever the algorithm, is inaudible with normal programme material, but discernible when pure tones are transmitted. Results from other workers in the field[4] show that 14: 10 near-instantaneous companding provides a standard of performance virtually identical with uncompanded sound. This is likely to be of great importance for satellite transmission, where significant savings in capital plant can be achieved if more channels can be accommodated within a given bandwidth.

Future techniques

Voice synthesis by microcomputer is a rapidly developing technique, especially for electronic toys and games. Most such devices currently available appear to possess American or Japanese accents, so revealing their places of origin. Economy of storage and audio bandwidth is afforded by masculine voices, but this is likely to become a minor consideration as the cost of memory chips continues to decline. Solid-state recording of high-quality musical performances is a more difficult matter, unlikely to be solved by the silicon chip for many years hence. For example, any recording of Beethoven's ninth symphony would require a digital storage array of approximately 2000 megabits. Current prices of memory chips would need to fall by a factor of 100 000 to render viable any such scheme. Meanwhile, more traditional devices such as magnetic tape, hard disc storage and the newer 'Winchester disc' continue to improve in performance, possibly rivalling the storage density achieved by laser-optical techniques.

In the digital processing area, one of the more interesting new devices to emerge is the Intel 2920 processor. This comprises an analogue-to-digital converter, a signal-processing computer and a digital-to-analogue converter, all contained on a single silicon slice. Current technology limitations restrict its operating frequency to about 14 kHz. However, speed improvements to at least five times that figure, whereby it would admirably suit the needs of the digital audio engineer, can now be expected. □

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1. Sanchey, A A, 'A new ultra-linear 16-bit digital-to-analogue conversion system for professional audio'. 66th AES Convention, Los Angeles, 6-9th May 1980.
2. Jones, A H, 'Digital Coding of Audio Signals for Point-to-Point Transmission'. Publication 166, IBC London 1978.
3. de Lamare E, 'Le Codage numerique des signaux sonores de haute qualite. . .' *Revue de Radiodiffusion-télévision*. No 44 Sept-Oct 1976, pp 12-33.
4 Gilchrist, N H C, 'Digital Sound Signals: Tests to compare the performance of five companding systems for high quality signals'. BBC Research Department Report, RD 1978/26.

Based on an article first published in *IBA Technical Review*, No 15. □

Continued from page 54

cations. And because the 6502 has been chosen by many of the microcomputer manufacturers, continued production seems assured for the forseeable future.

The hardware shown in Fig. 4 has been kept as standard as possible to reduce the overall cost, and the memory map for this arrangement is shown in Fig. 5. Circuits IC_2 and IC_5 provide 1K of r.a.m. for essential variables, the stack in pages 0 and 1, plus spare areas in pages 2 and 3. The r.a.m. is not fully decoded and appears throughout the bottom half of the 64K address space. An e.p.r.o.m. containing the firmware is assigned to the top 2 or 4K of memory with address decoding provided by IC_{14} for an expanded system. Wire-ORing of the address decoder outputs provides addressing options. Circuit 15 is enabled at C000 (hex.) to provide sub-divided outputs for display drivers and a versatile interface adapter.

The system clock is provided by a 1MHz crystal oscillator using unbuffered c.m.o.s. gates. This also provides the timing for a back-up system and is trimmed for best results in this mode. Power-on reset is provided by two Schmitt inverters which allow the power supply to stabilize before the program is initiated.

A potentially troublesome source of interference for v.l.f receivers is the conventional multiplexed display, and for this reason a low-current liquid crystal type is recommended. A suitable display and driver circuit which will plug directly into the display port is shown in Fig. 6. The eight-digit panel allows six 0.5in digits to be displayed with spaces to improve legibility. The control port of IC_{11} allows different displays to be selected. It is important to use only the AM version of the 7211 display drivers as these include display blanking and a microprocessor interface. Because the display port has been designed to drive remote displays via a short length of ribbon cable, IC_{16} is necessary to buffer the 6502 data bus.

As well as displaying time and date information, the evaluation system can be used with other equipment via an RS232C interface which transmits ASCII information. The necessary hardware additions are shown in Fig. 7. IC_7 divides the 1MHz clock to provide a 2,400 baud generator, and IC_{19} converts the serial data from the v.i.a. to an RS232 level.

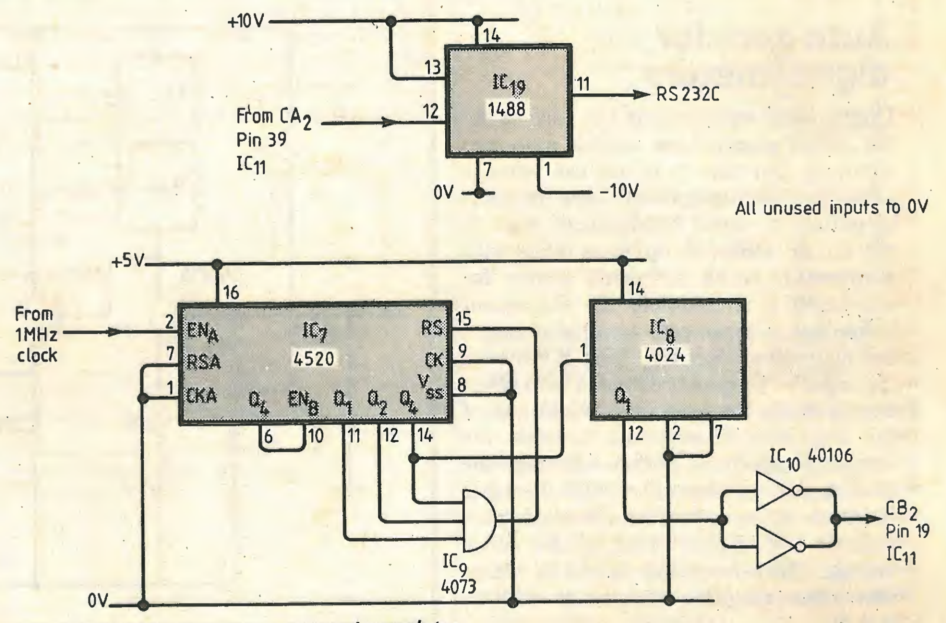
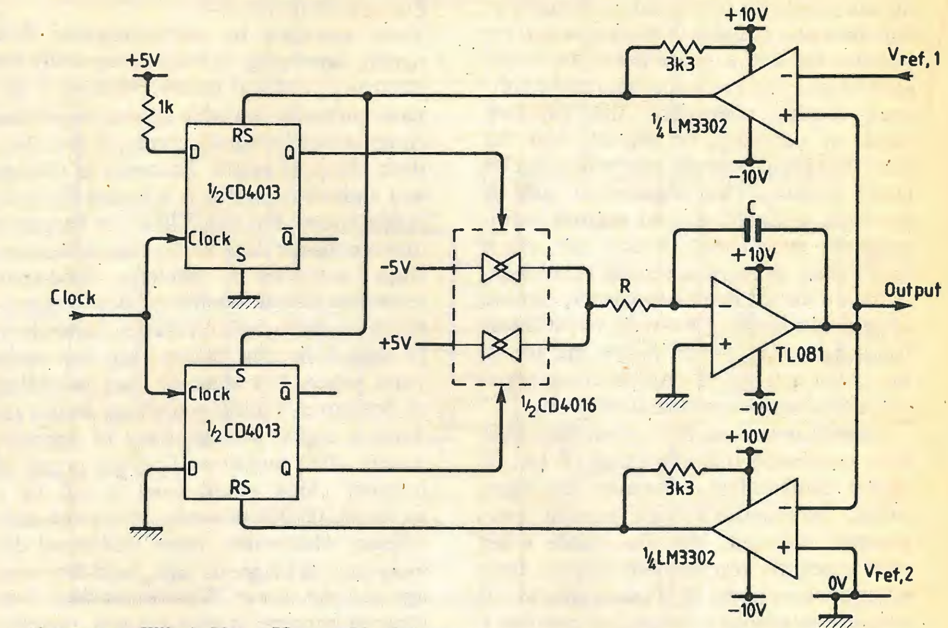


Fig. 7. RS232C generator and level translator.

Part two of this article describes firmware, construction and testing. A complete kit of components for this design will be available from Circuit Services, 6 Elmbridge Drive, Ruislip, Middx (telephone Ruislip 76962).

CIRCUIT IDEAS

Clock-triggered triangular pulse generator



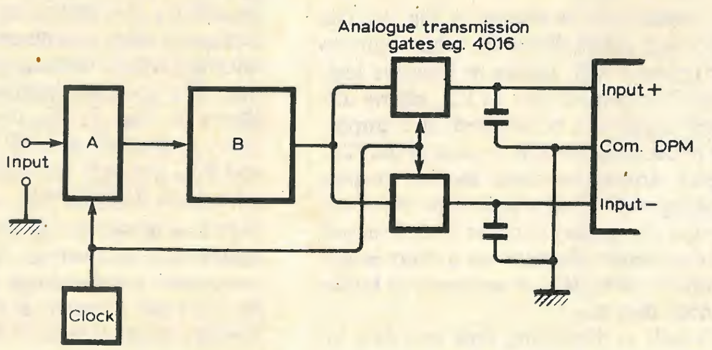
A double pulse is applied to the inverting input of a TL081 operational amplifier connected as an integrator and a triangular pulse is obtained at the output. The required double pulse is formed by two direct voltages −5V, +5V, applied to the integrator input *via* a pair of analogue switches. Two D-type flip-flops control these switches. The two flip-flops are triggered by the rising edge of the clock pulse applied to their clock inputs. When the clock-pulse triggers the two flip-flops, the First flip-flop's Q-output becomes equal to 1 and the Q-output of the second equal to 0. Consequently, one switch is enabled and the other disabled. Thus an input voltage equal to −5V is applied to the integrator. When no input voltage is applied to the integrator, V_{out}=0. Then, V_{in}=−5V and V_{out} increases; when it equals the reference voltage $V_{ref.1}$ the output of the comparator goes high, and the first flip-flop's Q-output is reset to 0, while the second's Q-output is set to 1. Thus the switches change state, so that V_{in}=+5V and V_{out} decreases. When $V_{out}=V_{ref2}$=0V, the output of the second comparator goes high resetting the Q-output of flip-flop 2 to 0. So both switches are disabled, and no input voltage is applied to the integrator. Consequently V_{out}=0 until the next rising edge of the clock pulse triggers the flip-flops. The duration of the triangular pulse is $T=2t_i$ where $t_i=1/RC$ is the time constant of the integrator.

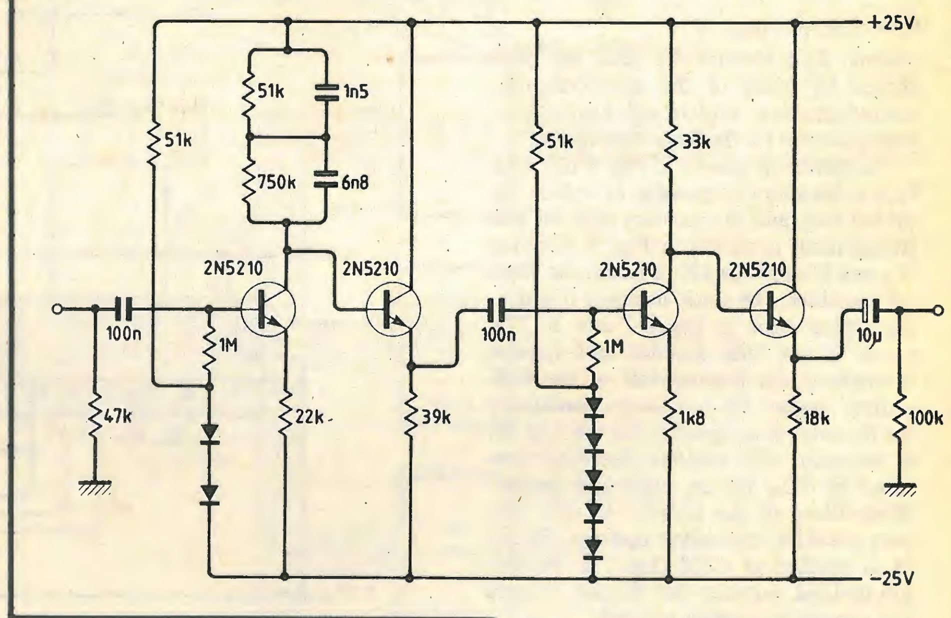
G. Tombras
Athens

Auto-zero for digital meters

Digital panel meters using i.cs such as the ICL7106/7 already have internal auto-zero circuitry, but this is of no use when a particular instrumentation case requires amplifiers or signal conditioners prior to the d.p.m. Offset in op-amps drifts with temperature so an automatic system for correcting it is desirable. In the circuit given, box A represents circuits to switch the instrumentation amplifiers B between the input to be measured and a zero reference level. At the same time, the output of the amplifiers is switched between the sampling capacitors so that one holds the amplified input plus offset, and the other holds the offset only. The differencing action of the d.p.m. cancels the offset voltage. Clock frequency should be higher than the sampling frequency of the d.p.m.

K. Wood
Ipswich
Suffolk



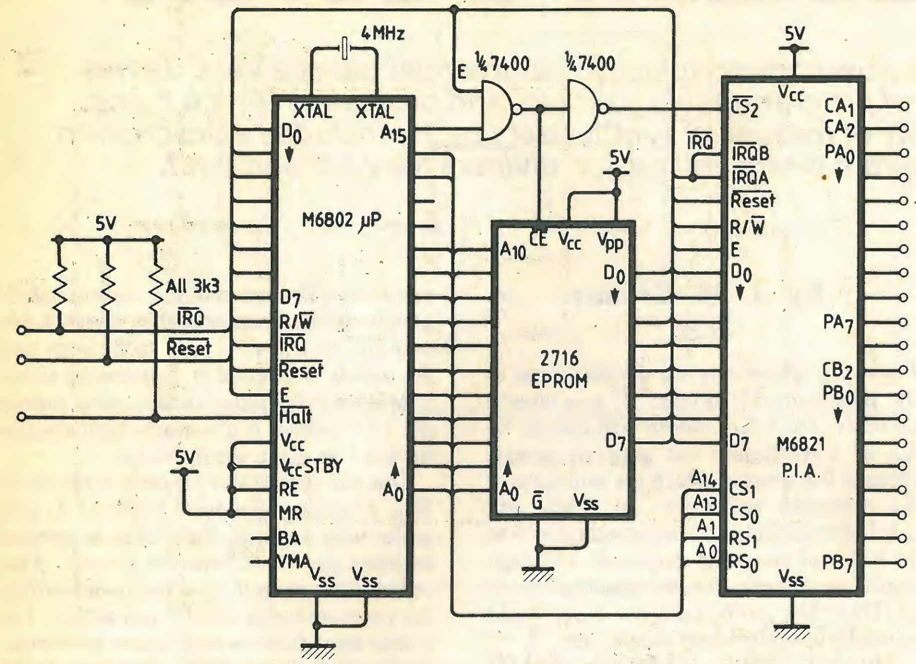


Preamp with no t.i.m.

Circuit shows a stable small-signal preamplifier with passive magnetic pick-up equalization but without overall negative feedback.

At 1kHz, the circuit has an overall gain of 50 and its input and output impedances are 47k and 1.7kΩ, respectively. Peak-to-peak maximum input and output voltages are 0.5 and 25V respectively.

Shausin Yang
National Chiao Tung University
Taiwan



Minimum component-count microprocessor

This microprocessor circuit brings the number of components required for certain control applications to a minimum and lends itself particularly to machine-control design. Address bus decoding is divided up into two 32K-byte pages and the e.p.r.o.m. is situated at 8000 or at 2K-byte images up to F800. The M6802's eight interrupt vector bytes should begin at 87F8 or its respective images. Two sets of eight i/o lines, A and B, are provided by the M6821 peripheral interface adapter; A addresses are decoded as 7000 and 7001 and B addresses as 7002 and 7003. Locations 0000 to 007F are used for the M6802's 128-byte r.a.m.

Other e.p.r.o.m.s, such as the 2758 or 2532, may be used in place of the 2716 with only minor alterations. The two spare NAND gates can be used to provide a bit-rate generator if necessary.

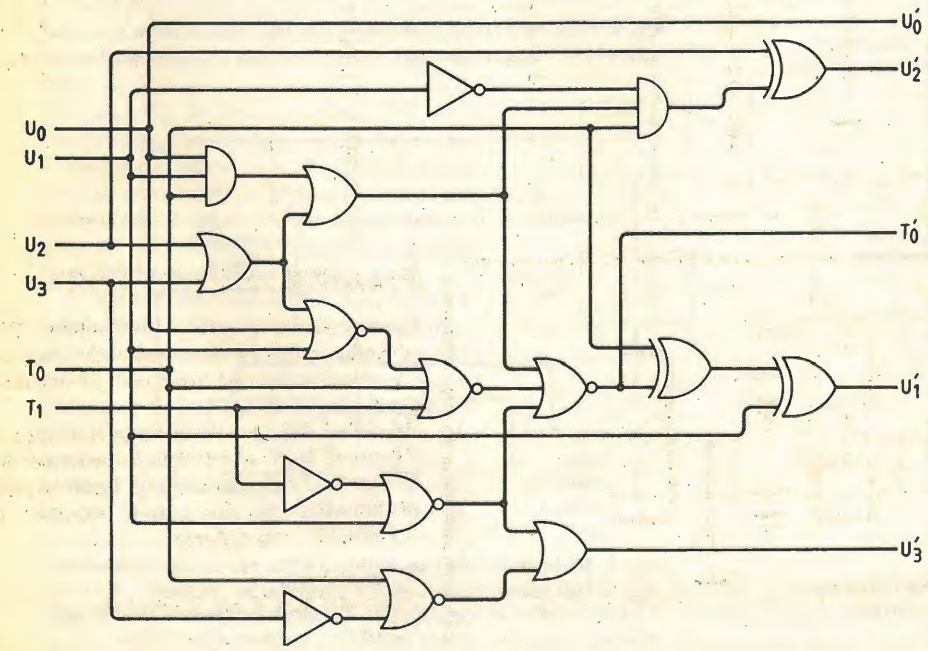
Y. C. Cheah
Wellington
New Zealand

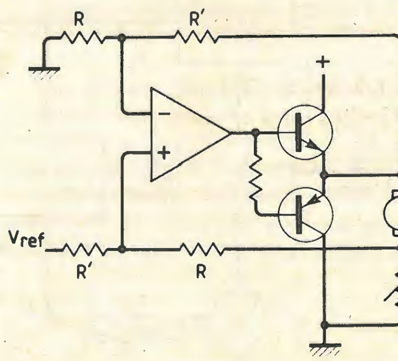
24-to-12-hour clock decoder

A digital clock may have a 24-hour display, which many people would find less preferable to the more normal 12-hour display. For example, the time-coded radio signals from Rugby work on the 24 hour clock.

The circuit shown is an economical decoder of b.c.d. 24-hour information (0 0 T_1T_0; $U_3U_2U_1U_0$) to b.c.d. 12-hour (0 0 0 T_0; $U_3'U_2'U_1'U_0'$). If c.m.o.s. i.cs are used, types 4001 (quad 2-input NOR), 4025 (triple 3-input NOR), 4030 (quad exclusive OR), 4069 (hex inverter), 4071 (quad 2-input OR) and 4073 (triple 3-input AND) are required.

W. Gough
Whitchurch
Cardiff





Accurate motor speed control with braking

Mr Malvar's 'Accurate motor speed control' (*WW* Circuit Ideas, August 1980) described a circuit in which the effect of motor armature resistance was cancelled by using the armature current to provide positive feedback to the drive amplifier. The amplifier used a booster transistor which entailed the motor stopping under open circuit conditions.

Accurate motor speed control is often required with a fast stop/start, and this can be achieved by the addition of a transistor complementary to the booster transistor.

The circuit shows this addition, with a somewhat modified bridge circuit. When $R_1=R_\alpha$ the armature back e.m.f. is equal to V_{REF}, which can be gated or switched to provide a fast stop/start (R_1 being the variable resistor). A single supply may be used.

K. G. Barr
University of the West Indies
Barbados

DIGITAL FREQUENCY SYNTHESIZER DESIGN

Digital frequency synthesis is now commonplace in commercial transceivers. James Bryant discusses the design of programmable counters and prescalers for v.h.f. and u.h.f. synthesizers using a family of frequency synthesizer i.cs. He includes a description of a basic computer program which will design dividers for v.h.f. and u.h.f.

by J. M. Bryant
B.Sc.

A basic frequency synthesizer, shown in Fig. 1, consists of a voltage-controlled oscillator, programmable divider, phase detector, low-pass filter and a stable reference frequency source. The v.c.o. and l.p.f. are the most critical parts of the design, and the v.c.o. must be isolated from the output and from the input to the divider. Operation of the synthesizer is straightforward: the v.c.o. output is fed to the programmable divider and then compared with the reference signal in the phase comparator, whose output controls the v.c.o. The system is therefore a phase-locked loop acting to maintain the divider output in phase with the reference input. The v.c.o. frequency is stabilized at *n* times the reference frequency, i.e. $F_{out}=nf_{ref}$ where *n* is the division ratio of the programmable divider. If *n* is altered by unity, the v.c.o. output will change by f_{ref}, so a synthesizer can generate several channel frequencies which are multiples of the reference frequency. In v.h.f. and u.h.f. synthesizers channel spacings of 5 to 50 kHz are normally required, although synthesizers with channel spacings down to 1Hz or less can be built but these would normally use multi-loop techniques.

Although the v.c.o., phase comparator and l.p.f. can be built using discrete components, for a complex circuit such as the programmable divider the use of i.cs is essential. Unfortunately, current integrated-circuit programmable dividers use c.m.o.s., n.m.o.s. or t.t.l. technology and are unable to operate at frequencies above 25MHz (a little higher in the case of Schottky t.t.l.) which is not nearly high enough for v.h.f. or u.h.f. synthesizers.

One solution to this problem is shown in Fig. 2 where a fixed v.h.f. or u.h.f. prescaler with a division ratio of *m* is inserted between the v.c.o. and the divider. This reduces the output to a frequency which the programmable divider can accept, but it also introduces several other problems. However because fixed dividers using e.c.l. technology are available with input frequencies up to 1.8 GHz, this system is often used in commercial equipment. Two minor problems with this approach are the high power consumption of e.c.l., and interfacing. A major and intractable problem is the effect on f_{ref} because the introduction of fixed prescaling changes the synthesizer law to $f_{out}=m\times n\times f_{ref}$. If the same channel spacing is needed, the reference frequency must be reduced by a factor of *m*, which will generally lie between 10 and 256. This complicates the design of the l.p.f., increases by a factor of *m* the time required for the synthesizer to lock and, unless extreme care is taken in the v.c.o. design, worsens the noise and reference sideband levels in the synthesizer output.

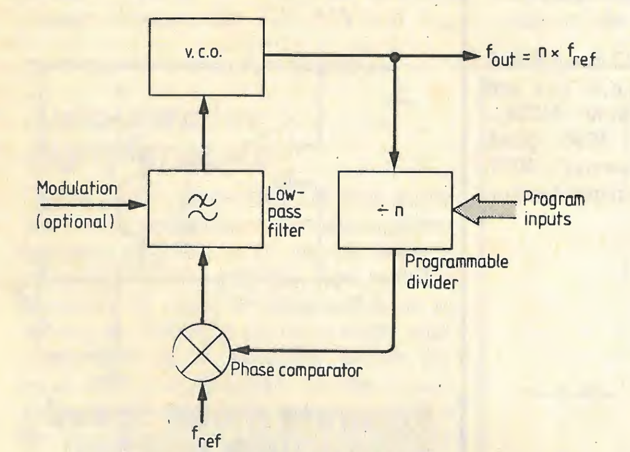


Fig. 1. *V.c.o. frequency is stabilized at n times the reference frequency in the basic synthesizer. Several frequencies can be generated by altering n.*

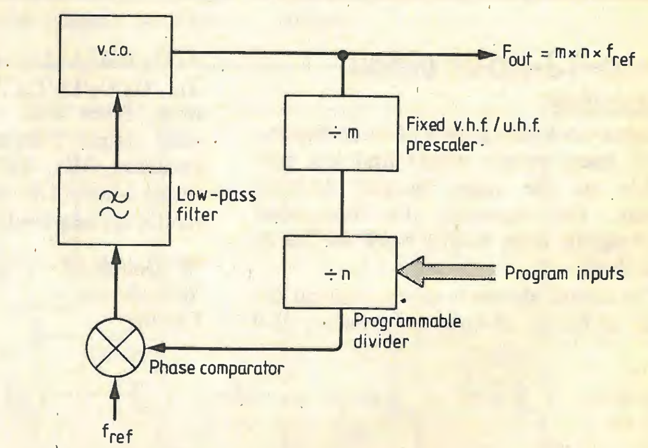


Fig. 2. *Including fixed prescaler in divider loop allows v.c.o. to operate at frequencies higher than c.m.o.s. or t-t logic allows.*

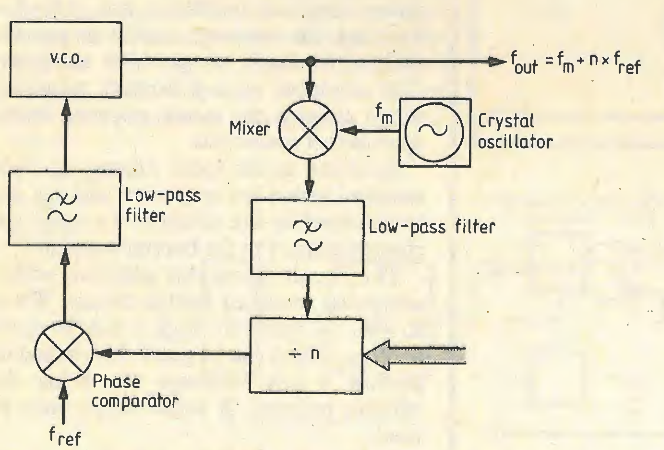


Fig. 3. *To avoid some of the problems with the fixed prescaler technique the v.c.o. is mixed with a frequency f_m and the difference applied to the divider.*

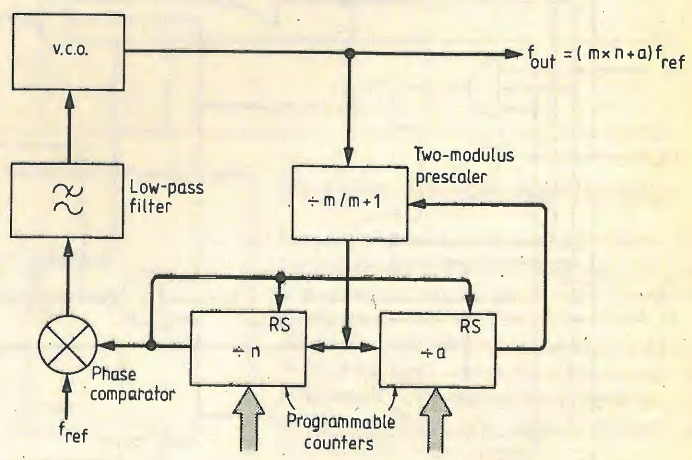


Fig. 4. *Multi-modulus prescaling avoids the noise problem of Fig. 3. Full counter cycle delivers one pulse for each a(m+1)+m(n−a) input cycles. Scheme works to u.h.f. though the two simple counters need only work to a few MHz.*

In applications where this degradation is unacceptable, the use of a mixer synthesizer as shown in Fig. 3 is often considered. The v.c.o. is mixed with a signal frequency f_m and the difference frequency ($f_{out}-f_m$) is applied to the programmable divider via a l.p.f. This system has a number of advantages, f_m can be switched to give i.f. and repeater shifts, power consumption is usually low, and it is easily understood. However, the overall system stability depends on two oscillators (f_{ref} and f_m), low noise levels in the second oscillator and the mixer are vital for good system performance (noise in the reference oscillator is less important because f_{ref} is usually divided from the reference oscillator) and the system is more complex. Nevertheless, many synthesized transceivers use this technique.

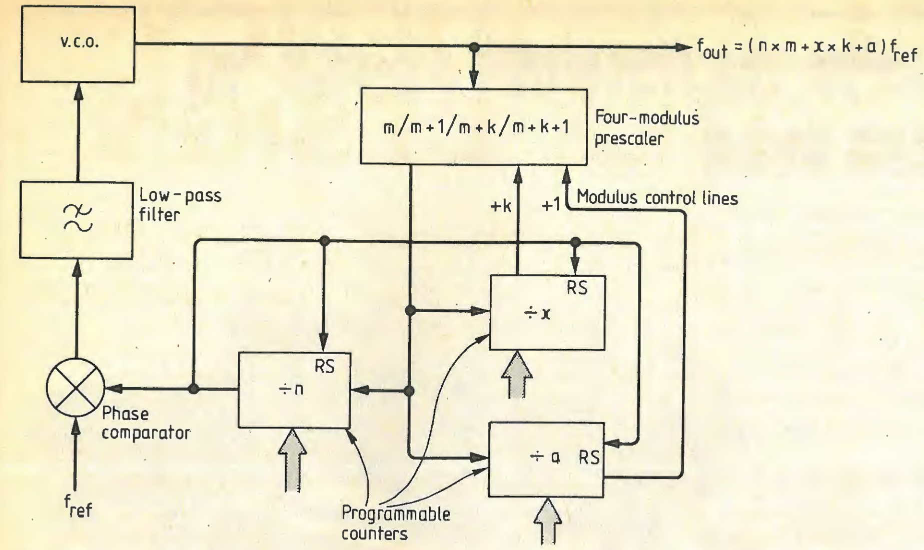


Fig. 5. *Four-modulus prescaling overcomes the limitation on lowest frequency of Fig. 4.*

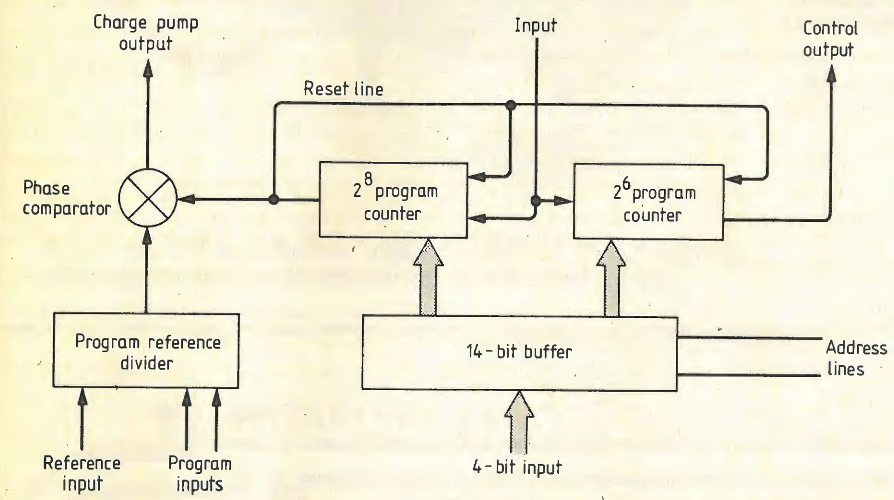


Fig. 6. *Synthesizer type NJ8811 is made from n-m.o.s. to reduce power consumption but it also reduces chip size and the number of diffusions required.*

A better system, known as a multi-modulus prescaling, is shown in Fig. 4. The simplest form uses a two-modulus prescaler (sometimes called a swallow counter) and when the system starts counting, the two programmable counters are reset to zero. The prescaler divides the v.c.o. output by $m+1$ and its output pulses increment both programmable counters. When the count reaches a, the prescaler modulus is changed to m and the a counter stops (at this point $a\times(m+1)$ cycles of the input frequency have been counted). The n counter continues to count the prescaler output until it reaches n and passes a pulse to the phase comparator. Both counters are then reset, the prescaler reverts to $(m+1)$ ratio and the cycle restarts. In the second half of the cycle, the system counts $(n-a)\times m$ cycles of the input frequency. Therefore, a full cycle of the counter delivers one output pulse for each $a\times(m+1)+m\times(n-a)$ input cycles, so the division ratio is $nm+a$:1.

The complete system forms a v.h.f. or u.h.f. programmable counter, but the programmable counter only operates at a few MHz which enables c.m.o.s., n.m.o.s. or t.t.l. devices to be used. Also, although two programmable counters are required, they are simpler than the type in Fig. 2 and the overall complexity is only slightly greater. There are, of course, drawbacks. The division ratios of a two modulus prescaler will normally be between 10/11:1 and 100/101:1, but ratios of over 20/21:1 are required at v.h.f. if the programmable counter input frequency is to be low enough for c.m.o.s. or n.m.o.s. devices.

For an $m/m+1$:1 prescaler the a counter

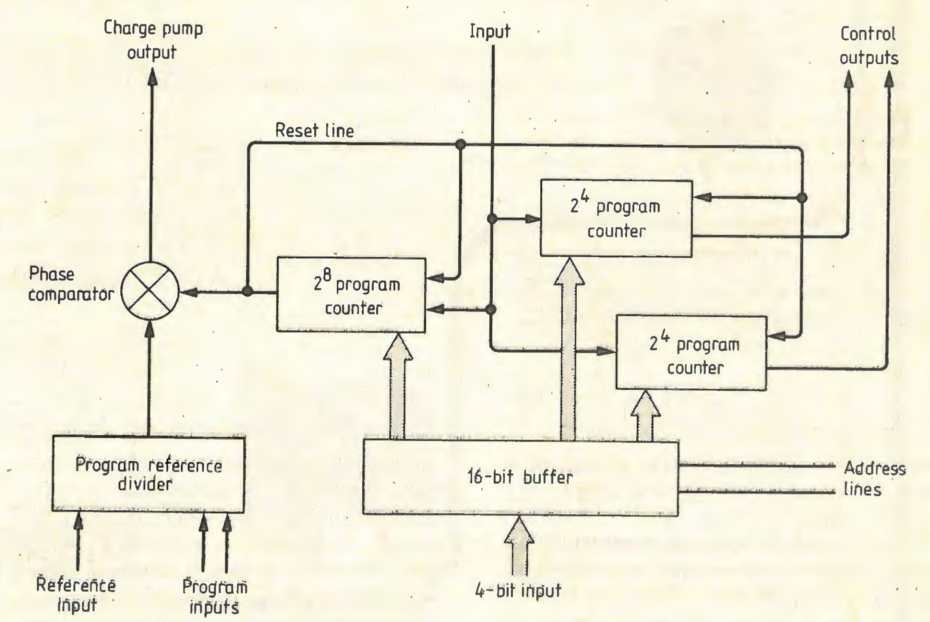


Fig. 7. *Although NJ8811 and NJ8812, above, will generally be programmed by a r.o.m. and channel switch they are compatible with microprocessor-based systems as well.*

Basic synthesizer design program

```
1 R$="REFERENCE" S$="SYNTHESIS" F$="FREQUENCY" G$=" FREQUENCIES" X$="PROGRAM
2 D$="DIVISION RATIO" E$=" DIVIDER" U$="TRANSCEIVER" P$="PRESENT" O$="PRESCALER
"
3 W$="HARMONIC" V$=" CHANNEL SPACING" C$="ENTER " Z$="INPUT ERROR - TRY AGAIN"
4 PRINT"PLESSEY SEMICONDUCTORS LTD." PRINTTAB(15)P$"" T=2 GOSUB
68:GOSUB138
5 T=3 GOSUB68 PRINT""; GOSUB138 M$="MAXIMUM " PRINT"PLESSEY SEMICONDUCTORS M
AKE A RANGE OF
6 N$="MINIMUM " O$="PIN 1 IS CONNECTED TO" PRINT"INTEGRATED CIRCUITS FOR USE I
N "F$;
7 PRINTS$"ERS.  THIS "X$" IS AN AID TOTHE DESIGN OF VHF & UHF "S$"ERS FOR";
8 PRINT"MOBILE,AERONAUTICAL,SATELLITE,MARINE ANDHAM "U$"S WITH"V$"S
9 H$=" PART " PRINT"OF BETWEEN 1 & 100 KHZ." PRINT"    THE FIRST"H$"OF THE "X$

10 PRINT" TAKES" PRINT"THE HIGHEST AND LOWEST"G$" TO BE"S$"ED, THE I.F. OFFSET
(WHICH MAY
11 PRINT"BE 0 IF THE "S$"ER IS NOT FOR A" PRINTU$"), AND THE"V$"." PRINT"IT CHE
CKS WHE";
12 PRINT"THER THE "S$"ER MAY BEMADE WITH PLESSEY INTEGRATED CIRCUITS" PRINT"AND
 WHICH ",
13 I$=" IS TOO HIGH " PRINT"ONES SHOULD BE USED." PRINT"TO CONTINUE"; GOSUB137
14 PRINT"PLEASE INPUT THE FOLLOWING"G$"  IN MEGAHERTZ:" W=4.8E6 U=W 0=0
15 PRINTM$; INPUTF IFF>1800.001THEN102
16 PRINTN$; INPUTL IFL>FTHENPRINT"ERROR! MIN>MAX" T=2 GOSUB68 GOTO14
17 O=0 PRINT"INTERMEDIATE "F$" ("C$"ZERO IF   NOT A "U$")  0"; INPUTI A=1E6
18 F=A*F L=A*L I=A*ABS(I) H=F J=L IFI>0THENPRINT"HIGH OR LOW SIDE INJECTION?",:
GOSUB66
19 IFA$="H"THENO=1 PRINT" HIGH" IFF+I>1.801E9THENF=(F+I)/1E6 GOTO102
20 IFA$="L"THENO=-1 PRINT " LOW" L=L-I
21 IFO=0ANDI<>0THENPRINT" NONE
22 T=.3 GOSUB68 PRINTC$"THE"V$" IN KILOHERTZ" INPUTP P=P*1E3 IFI/P<>INT(I/P)THE
N124
23 IFO=1THENF=F+I
24 PRINT"STANDARD "R$" INPUT IS 4.8 MHZ.    DO YOU WANT A DIFFERENT ONE?";
25 GOSUB66 IFA$<>"Y"THENPRINT" NO" T=.7 GOSUB68 GOTO27
26 PRINT" YES" PRINTC$R$" INPUT IN MHZ"; INPUTW W=W*1E6 U=W
27 IFF>1E9THENA$="SP8619 PLUS SP8906" R=4 GOTO128
28 IFF>5.12E8THENA$="SP8901" R=2 GOTO128
29 IFF>2E8ORF/P>16960THEN127
30 PRINT"THE SYNTHESISER MIGHT USE AN NJ8811 OR  AN NJ8812.  IF YOU WISH TO US
E";
31 PRINT" THE" PRINT"NJ8811 "C$"1, OTHERWISE", GOSUB137 IFA$="1"ORA$="I"THEN127
32 IFL/P<1600THEN106
33 A$="SP8793" T$="2" R=1 V=P X=1 GOSUB131
34 RESTORE F7=0 A=W/V FORB=0TO3 READC FORD=0TO3:DATA128,160,192,240
35 IFA=INT(C*2↑D+.5)THENB1=B D1=D F7=1 GOSUB71
36 NEXTD,B RESTORE IFF7=0THENA=.9*W/V GOTO107
37 PRINT""R$" "D$" IS"A PRINTR$" "F$" IS"V/1E3"KHZ." GOSUB139 PRINT"TO SET "R
$" "D$
38 PRINT"CONNECT PIN 8 TO PIN"B1 B1=3-D1 GOSUB139 PRINT"AND PIN 9 TO PIN"B1
39 IF(X=2ANDF/P>=69376ANDL/P>=36608)OR(X=1ANDF/P>11840ANDL/P>=6270)THEN132
40 IF(X=2ANDF/P<69376)OR(X=1ANDF/P<11840)THENG=0 PRINTO$" VCC" Y=1600-2240*(X=2
):GOTO47
41 PRINT""F$" RANGE IS SO LARGE THAT PIN 1
42 PRINT"(OFFSET PIN) MUST BE SWITCHED TO ALLOW  THE MAIN"E$" TO COVER THE RANG
E
43 PRINTL/1E6"TO"F/1E6"MHZ.  IN ANY CASE
44 PRINT"SUCH A WIDE RANGE OF"G$" (RATIO IS"F/L"1)",
45 PRINT"NECESSITATES SWITCHING THE" PRINT"V.C.O. RANGE. CALCULATE IN SECTIONS.
46 T=3 GOSUB68 GOTO14
47 GOSUB72 IFO=UTHENU=W 0=0 GOTO27
48 PRINT PRINT"TO ACCESS"H$"2"; GOSUB137
49 PRINT"   "H$"2 OF THIS "X$" CALCULATES
50 PRINT"THE DATA INPUTS NECESSARY TO "X$
51 PRINT"DISCRETE"G$"." PRINT
52 S=0 IFI>0THENPRINT"    ARE CHANNELS SIMPLEX OR DUPLEX?" GOTO121
53 T=3 GOSUB68 A=1E6
54 IFS<2THENPRINT""C$F$" (MHZ)", INPUTK K=K*A IFH<KORK<JTHENPRINTZ$ T=2 GOSUB6
8 GOTO54
55 IFS<2THENM=K+O*I GOTO59
56 PRINT""C$"TX "F$" (MHZ), RX "F$"  (MHZ)", INPUTK,M K=K*A M=M*A
57 IFH<KORH<MORK<JORM<JTHENPRINTZ$ T=2 GOSUB68 GOTO56
58 M=M+O*I
59 PRINT""D$, IFS>0THENPRINT" TX",
60 Z=0 PRINT" -", B=K/P PRINTB N=K/1E6 GOSUB88 IFZ=1THEN62
61 IFI<>0THENPRINT""D$" RX -", B=M/P PRINTB N=M/1E6 GOSUB88
62 IFZ=1THENZ=0 PRINT"RETURN TO"H$"1 FOR NEW VALUES." T=3 GOSUB68 GOTO14
63 IFX=1THEN134
64 PRINT"DO YOU WANT ANOTHER CHANNEL? ", GOSUB66 IFA$<>"N"THENPRINT"YES" T=1
GOSUB68 GOTO54
65 PRINT"NO" END
66 GETA$ IFA$=""THEN66
67 RETURN
68 T=TI+60*T IFT>5183990THENRETURN
69 GETA$ IFT>TITHEN69
70 RETURN
71 PRINT""R$" INPUT -",U/1E6,"MHZ." RETURN
72 A=1E6 IFH-J>4UTHENGOSUB78 RETURN
73 IFINT(H/U)-INT(J/U)=0ANDJ/U<>INT(J/U)THEN RETURN
74 PRINT"THERE IS A "W$" OF THE "R$"    OSCILLATOR AT" INT(H/U)*U/A,
75 PRINT"MHZ." PRINT"DO YOU WANT A DIFFERENT "R$" INPUT ", GOSUB66 IFA$<>"Y"THE
NRETURN
76 P=U/V FORA=0TO3 RESTORE FORB=0TO3 READC IFP<C*2↑ATHEN114
77 NEXTB,A
78 PRINT"IT IS INEVITABLE THAT ONE OR MORE
79 PRINTW$"S OF THE "R$" OSCILLATOR  WILL LIE BETWEEN"J/A"AND"H/A ,
80 PRINT"MHZ. TO LIST THEM PRESS 'Y' " PRINT"N.B.THIS ERASES "P$" SCREEN CONTENT
S
81 PRINT"OTHERWISE", GOSUB137 IFA$<>"Y"THENRETURN
82 PRINT""; GOSUB87 PRINT B=INT(J/U)+1 B=INT(J/U)+1 GOSUB86 IFB*U>HTHENRETURN
83 PRINT""; GOSUB87 PRINT,,
84 GOSUB86 IFB*U>HTHENRETURN
85 GOTO84
86 PRINTB,""B*U/1E6, B=B+1 RETURN
87 PRINTW$,"F - MHZ", RETURN
88 IFB-1E-4>INT(B)THENPRINTF$" NOT A MULTIPLE OFTHE"V$ Z=1 RETURN
89 PRINTF$" -"N"MHZ" PRINTO$; IFG=1THENPRINT" GROUND
90 IFG=0THENPRINT" VCC
91 B=B-Y PRINTX$" NUMBER IS -"B IFX=1THENB=B+192*INT(B/64)
92 C=16 B(1)=INT(B/(2↑12)) B=B-2↑12*B(1) B(2)=INT(B/(2↑8)) B=B-2↑8*B(2)
93 B(3)=INT(B/C) B=B-B(3)*C B(4)=INT(B+.1)
94 FORB=1TO4 IFB(B)<10THENB$(B)=RIGHT$(STR$(B(B)),1)
95 IFB(B)>9THENB$(B)=CHR$(B(B)+55)
96 NEXT PRINT"WHICH IS  "; FORB=1TO4 PRINTB$(B), NEXT PRINT"  IN HEXADECIMAL":P
RINT"AND  ";
97 FORB=1TO4 T=B(B) T$="0" IFT>=8THENT$="1" T=T-8
98 A$="0" IFT>=4THENA$="1" T=T-4
99 T$=T$+A$ A$="0" IFT>=2THENA$="1" T=T-2
100 T$=T$+A$ A$="0" IFT=1THENA$="1"
101 T$=T$+A$ PRINTT$;" "; NEXT PRINT"  IN BINARY" RETURN
102 PRINT""F/1E3"GHZ"I$"A "F$" FOR" PRINT"THIS TYPE OF "S$"ER AT "P$" (BUT";
103 PRINT"TRY US AGAIN NEXT YEAR!)" T=3 GOSUB68 GOTO14
104 PRINTM$D$I$ PRINT"EITHER REDUCE "M$F$" OR" PRINT"INCREASE"V$".
105 W=4.8E6 U=W PRINT"TO TRY NEW VALUES"; GOSUB137 GOTO14
106 PRINT""N$D$" IS TOO LOW" PRINT"EITHER INCREASE "N$F$" OR" PRINT"DECREASE"V
$"." GOTO105
107 FORB=0TO3 RESTORE FORC=0TO3 READE D=INT(E*2↑B+.5) IFD>ATHENB1=C D1=B B=4 C=
B NEXTC,B GOTO109
108 NEXTC,B
109 A=D U=D*V IFU>12E8THEN112
110 IFU<2.5E6THEN113
111 GOSUB71 GOTO37
112 PRINT""R$" "F$I$ GOTO37
113 GOSUB71 PRINT"(LOWER THAN OPTIMUM - SUGGEST USE OF" PRINT"EXTRA"E$")" GOTO3
7
114 D=0 IFA=3THENA=2
115 IFB=3THEN RESTORE B=0 A=A+1
116 READC B=B+1 D=D+1 R=C*2↑A O=V*R IFINT(H/O)-INT(J/O)=0ANDJ/O>INT(J/O)THEN119
117 IFD=4THENPRINT"THERE IS NO PRACTICAL VALUE OF "R$; GOTO120
118 GOTO115
119 PRINT"NEW STANDARD "R$" INPUT IS"O/1E6"MHZ." W=O U=O T=1.5 GOSUB68 RETURN
120 PRINTF$" WHICH IS FREE OF IN-BAND" PRINTW$"." RETURN
121 GOSUB66 IFA$="S"THENPRINT" SIMPLEX" S=1 GOTO53
122 IFA$="D"THENPRINT" DUPLEX" S=2 GOTO53
123 GOTO121
124 PRINT"THE I.F.IS NOT A MULTIPLE OF THE"V$; PRINT", CHOOSE EITHER A NEW I.F
 OR A
125 PRINT"NEW"V$"." PRINT"N.B.CHECK THAT CHOICE OF 1/2 OR 1/4 OF
126 PRINTP$V$" IS NOT ALL THAT IS NEEDED." T=5 GOSUB68 GOTO14
127 A$="SP8906" R=1
128 X=2 T$="1" GOSUB131 IFF/P>102143THEN104
129 IFL/P<3840THEN106
130 GOTO34
131 PRINT""O$" - "A$ PRINTX$"MABLE"E$" - NJ881"T$ PRINTR$" IS"1/R"TIMES"V$ V=P
*R RETURN
132 G=1 PRINTO$" GROUND" Y=6270-30338*(X=2) GOTO47
134 PRINT"N.B. USE OF NJ8812 MODIFIES THE BINARY  AND HEXADECIMAL NUMBERS BECA
USE",
135 PRINT" OF THE  INSERTION OF TWO EXTRA UNUSED BITS INTO THE SECOND LEAST ";
136 PRINT"SIGNIFICANT WORD." GOTO64
137 PRINT" PUSH ANY KEY." GOSUB66 RETURN
138 PRINTTAB(8)"**"F$"  "S$"**" PRINTTAB(10)"―――――――――――――――――" RETURN
139 B1=6-(B1=1)-11*(B1=2)-7*(B1=3) RETURN
```

must be programmable over a range of m and the n counter must always divide by a larger number than the a counter. Therefore, to be fully programmable, the total system division ratio must be equal to, or greater than m^2. This sets a lower frequency limit for such synthesizers, for example, with 25kHz channel spacing and a 40/41 prescaler the minimum division ratio is 40^2 so the minimum frequency is 40 MHz.

Generally, the programmable counter sets a limit which is higher than the theoretical minimum. If wider tuning is required, four-modulus prescaling can be used. A typical system is shown in Fig. 5 with four moduli, $m/m+1/m+k/m+k+1$, which are set by $+1$ and $+k$ control lines. There are three programmable counters and the conditions which limit the ratios are, a must count over a range of k, x must count over a range of $(m+k+1)/k$, and n must count at least the minimum value of a or k. For a 55/56/63/64 prescaler, the division ratio limit is 512, which allows 25kHz channel synthesis above 12.8 MHz. Again, in a practical system, the programmable counter will generally set a higher minimum. The overall division ratio of this system is $m \times n + k \times x + a$ where n, x and a are the counts in their respective counters.

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NEWS

Sinclair versus BBC

Clive Sinclair with his new colour graphics computer the ZX Spectrum

When the BBC announced that they had selected Acorn to produce their microcomputer, Clive Sinclair was furious. He said that he could offer "any facilities that the Corporation might require at a lower price than any competitor". The Sinclair ZX Spectrum is his attempt to prove it and he recently launched it as "not the BBC Computer". In his promotional literature he lists features included on the Spectrum and compares them with rival colour-display computers in an attempt to prove that the Sinclair can out-perform the others, especially the BBC model 'A', at less than half the price.

As he has invited the comparison, it is worth taking a closer look at both the BBC model A and the ZX Spectrum. The first obvious difference is the keyboard. The BBC has a conventional typewriter layout with 73 keys including ten user-definable function keys. The Spectrum has 40 keys and some have six functions which involves extra shift-keys to get the required function. This is offset by the single keystroke entry of all the Basic keywords available. Unlike the miniature membrane of the ZX81, the Spectrum has moving keys at typewriter pitch, but instead of concave full-sized keys they are in the flat, calculator style.

The Spectrum offers eight-colour graphics with 190 x 256 pixels and 32 x 24 text: other modes are user definable. The BBC 'A' has a choice of modes with four-colour graphics, 160 x 256 pixels and 20 x 32 characters. Higher definition is possible on the BBC 'A' at the expense of using fewer colours.

The Sinclair has overcome one of the major bugbears of the ZX81, that of loading and saving programs and data onto a cassette recorder. The Spectrum has a cassette interface that records a tone onto the tape: when loading, the computer automatically adjusts to the tone so that the correct input level is set. This overcomes the automatic recording level fluctuations on some recorders. The BBC 'A' uses a test tape to tell the operator when the recording level is correct after it has been adjusted manually. Spectrum can load a program faster at 1500 baud compared with 1200 baud.

Using a BEEP command, the Spectrum can generate sound which may be controlled in pitch and duration, but the model A has three-voice music synthesis with full envelope control.

The ZX printer plugs directly into the Spectrum and offers the same high resolution and graphics capabilities, without the colours, as the screen, and can reproduce anything displayed. Model A has no printer interface – it is only available on the enhanced model B. An input/output port is available for the Spectrum and not for model A.

The BBC computer uses a version of Basic with a very large number of keywords. The ZX Basic is an extended version of the ZX81 but has fewer keywords than the BBC. Some of the more useful keywords available to the BBC computer are AUTO for automatic program line numbering and RENUMBER. OLD restores a program that has been released from memory by NEW. There are about 30 other keywords which are not available on the Spectrum. However, there are some on the Spectrum which do not occur in model A, especially VERIFY which can compare a program which has been saved with the original; and MERGE which can combine a program being loaded with another in the memory. The BBC seems to have more powerful file-handling facilities. In practice, both computers will be able to perform similar functions but the Sinclair may need some additional steps in the more complex programs.

The memory (r.a.m.) capacity of both machines is the same at 16K and, while the BBC may be expanded to 32K, the Sinclair can have 48K. The Sinclair claims to have more efficient memory 'packing' so that more r.a.m. is left available when using high resolution graphics.

BBC model A may be enhanced to model B and then has a very wide choice of extra facilities; analogue inputs, serial and parallel ports. An 8-bit Centronics printer port, viewdata and teletext buffered extension bus, Red, Green, Blue, and sync. outputs and a disc memory filing system. So far the Spectrum will operate the ZX printer. An RS232 serial interface board is to become available soon, as are the Microdrives – miniature microfloppy disc drives which will hold up to 100Kbytes of data. Eight of them may be linked together. Each Microdrive will cost about £50. The prototype Microdrive demonstrated by Sinclair recently was only about 70mm wide, so the discs are very small.

When we come to compare the prices, the Sinclair has the distinct advantage. The basic model costs £125 and the 48K memory model, £175. BBC model A cost just under £300 and model B, £400. The Sinclair ZX81 attracted a wide variety of peripherals from other manufacturers: at a recent computer fair there were demonstrations of high resolution graphics, interface for all kinds of equipment including full-size parallel printers, music synthesizers and even colour graphics. If the ZX Spectrum attracts similar support from these manufacturers, or others, it seems there could be few limitations to its abilities while remaining within the price ceiling set by its rival.

Meanwhile, Sinclair Research will continue to produce and sell the ZX81. 400,000 have been sold in a year and Mr Sinclair believes that it is still the best introduction to computing for those unwilling to undertake a higher financial commitment. The price of the add-on 16K r.a.m. has been reduced to £30 but the ZX printer increased to £60. Despite the high profits of Sinclair Research over the last year, Sinclair is looking for more capital to finance some of their other activities including Clive Sinclair's pet project, an electric car. For this he is investigating the possibility of selling some shares in the company.

Sinclair might not lead the low-cost colour computer market for long. It is rumoured that Acorn, the makers of the BBC computer, are to produce their own colour device, the Electron, for about the same price as the Spectrum.

Flat-screen 'scope

It is often forgotten that liquid crystals are a British invention and that their development at the Royal Signals and Radar Establishment earns the UK royalties from all over the world. The earlier 'twisted nematic' l.c.d used in digital watches and calculators need polarized light to make them visible, but the newer 'dye-phase-change' l.c.d use the optical properties of dyes, dissolved in the liquid crystal, which make them brighter and not subject to the restricted viewing angle of the earlier type. The development of these displays has enabled them to be made much larger. This, however has led to problems of addressing and driving the display elements. Time division multiplexing may be used but the limit has been achieved in a message display with four lines of text.

Oscilloscope displays are fractionally easier to produce as the information displayed is simpler in form than that presented on message displays. It is usually necessary to distinguish only one element in each column for a waveform to be displayed. The method invented by Dr Ian Shanks of the RSRE uses row and column drive waveforms which are divided into discrete time periods. In each period the drive waveform is either on or off associated with the logic states '1' or '0' respectively. The drive waveform is therefore binary and the sequence of logic states may be repeated every 30ms. The waveforms can be supplied by standard c.m.o.s. logic circuits, unlike message displays which need special decoder/drivers. Unfortunately direct voltages cannot be used to drive liquid crystals. They would cause chemical decomposition. This problem is overcome in the 'scope display by generating pseudorandom binary sequences in the drive waveforms. A different waveform may be applied to each row of the display. Any of these or a different set of waveforms are also applied to the columns. Only when the same waveform is applied to both row and column in the matrix will there be no voltage difference and so that element will be 'off'. As there will be a voltage difference on all the other elements in the column, they will be 'on' and therefore distinguishable. The vertical height of the 'off' element is determined by the choice of waveform for that column and this is in turn determined by the value of the sample taken from the input signal waveform. As the 'off' element depends on zero voltage difference, the device is not sensitive to voltage changes caused by temperature and full performance is maintained under a variety of conditions.

A display using this system was incorporated into a prototype oscilloscope at the RSRE, Malvern, with a 100 × 100 element matrix. In 1977, the NRDC invited Scopex to see the work at Malvern and this has eventually led to the development and production of the first commercially available flat screen oscilloscope. The Scopex Voyager has a 128 × 256 element display with a graticule of 60 × 100mm. The bandwidth is d.c. to 150kHz sampled eight times for each cycle. Analogue signals are converted to digital using a successive approximation method. The 7-bit digital words so produced are written into a r.a.m. in a location which corresponds to its position in relation to the timebase or X-axis of the instrument. The contents of the r.a.m. at a particular address are used to define the pseudo-random binary sequence waveform for the column corresponding to that address and thus the vertical level of the display element. Waveforms can be up-dated or held in the r.a.m. This gives a stable image to the display which does not flicker or fade. A 'save memory' function allows a waveform to be stored after the instrument is switched off. Waveforms from different sources may be compared using the dual-trace function. There is also a pre-trigger function. The use of an external timebase or X input converts the oscilloscope to an XY plotter.

At the analogue input end of the 'scope, Scopex have retained their 'easy to use' philosophy, using as few external controls as possible. To this end, the vertical amplifier, trigger and timebase controls are similar to those on Scopex analogue oscilloscopes. There are two Y inputs to provide a dual-trace and in order to cope with the 7-bit resolution and the 1.25MHz conversion rate, a switching speed of 60ns is achieved using a Schottky diode ring gate. The 'scope is powered by rechargeable internal batteries through a switch-mode supply. All this is housed in a plastics case and weighs only 2.5kg. Along with probes and a battery charger it is presented in a leather briefcase for £2,500.

The Scopex Voyager oscilloscope is the first commercially available to include a flat l.c.d., making it smaller and with lower power requirements than conventional instruments.

Memory key

Plastic credit cards can be damaged or stolen, the magnetic strip information can be copied and is limited in the amount of information that can be stored. An alternative has been produced by a manufacturer of plastic cards, Data Card International, which is a plastic key incorporating an electrically alterable read only memory (e.a.r.o.m.). The memory has a capacity for 300 characters and the key is used in conjunction with a keyhole or 'Keyceptacle', (sic), with the appropriate electrical contacts and a 'Keytroller' – a micro-controlled interface unit which allows the contents of the key to be communicated to a host computer. The Keytroller also 'manages' the data so that the most efficient use is made of the memory.

The system is inherently secure as the details are entered at random. Sections of the data may be protected by access codes and the memory is sufficient to contain a variety of details or personal data known only to the keyholder. If anyone were able to copy the key, it would be no good without the knowledge of the appropriate codes.

The Datakey, as it is called, can be used for access to areas or machines that are secure against unauthorised users in much the same way as a plastic card. It has many additional uses as it can contain a programme for almost any computer or controller. Thus it may be used for work and time logging, for monitoring patients or medical staff, for vehicle records etc.

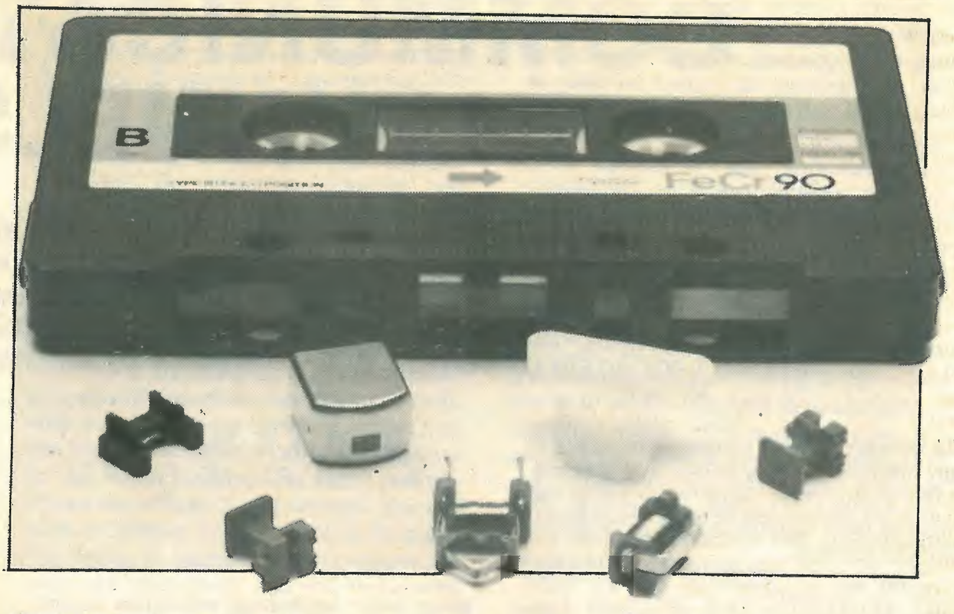
A particular advantage of the system is the ability to alter the contents of the memory. It can hold details of an account, and money or other token units may be added to or subtracted from a total held in the memory. This enables it to be used as a credit card to be used with a variety of vending or dispensing machines. An example is the dispensing of petrol at a pump controlled by the key.

A further example of the use of the key is the storage of instructions for a machine tool. A change of key instantly changes the instruction set without the need for re-programming or loading punched tape. Keys may be used in combination with one key specifying a product type while another may hold details which may vary within the type.

A Datakey development system, including a number of the keys and a keyboard/display programming unit, is available to OEM customers.

The Datakey contains an e.a.r.o.m. It fits on a keyring, is virtually indestructable and the contents of the memory are not affected by electromagnetic fields.

Sir Kenneth Corfield (right) receives the Royal Charter of the Engineering Council from John Wakeham, Under-Secretary of State for Industry. The Council was set up as a result of the Finniston Report to act 'as an engine for change in shifting national attitudes and priorities'. The new Council will take over most if not all of the executive functions of the Council of Engineering Institutions, especially the supervision of the training and qualifications of engineers. The CEI has agreed that 'when the time is right' it will seek agreement of its members to transfer the Engineers Registration Board. The right time depends on the Engineering Council and is likely to be in about two years. If the new Council assumes all the functions of the CEI, then the CEI will 'undoubtedly wind itself up', according to Bryan Hildrew, the retiring Chairman of the CEI in his foreword to the CEI Annual Report.



These miniscule plastics bobbins are used in cassette tape recording heads. The sort of thing that one associates with far Eastern manufacturers, these are made in Alcester, Warwickshire and exported to the Far East by Dyncast International, who have won a contract for eight million of them.

Intelsat VI – a new series

Five new Intelsat telecommunications satellites will offer three times the capacity of the current series V, and will be in operation by 1987.

The satellites are to be built by the Hughes Aircraft Company who will be joined by an international team of subcontractors. British Aerospace has a major role with about £50 million worth of orders over a seven-year period. After the initial five there is an option for eleven additional spacecraft.

Each satellite will be nearly 12m tall and 4m in diameter with a weight of 1,800kg. The solar panels will generate 2,200W to power up to 33,000 two-way telephone channels and four tv channels.

The spacecraft has been designed for launching by both the ESA Ariane and the NASA space shuttle. As the shuttle does not reach the required altitude for geosynchronous orbit, there needs to be an additional booster stage and a system for launching the satellite from the shuttle. BAe will design and build the cradle for carrying the spacecraft in the shuttle bay. This will include electronics units and the power and signal interface which connects the shuttle and spacecraft. They will also design and build the C and K band dish reflectors, other structures and wiring harnesses.

British Telecom has the second largest share in the International Telecommunications Satellite Organisations of which there are 106 members. The new satellites will help BT meet the demand for international telephone calls which grows at a rate of more than 20% each year.

In brief

80% useful energy conversion is claimed for NASA's latest fuel cells. Forty-five fuel-cell plants each housed in cabinets 2.7×1.5×2m and capable of generating 40kW, are to be installed experimentally in various sites around the U.S.A. The fuel cell power generators produce electricity from hydrocarbons by first converting to a hydrogen-rich fuel and then feeding it to one electrode in a cell of phosphoric acid electrolyte, while the other electrode is fed with oxygen. The cells produce heat as well as electricity and if the heat is used for heating homes or commercial buildings, the combined output of electricity and heat represents an energy conversion factor of about 80%. This compares with about 30% energy conversion from conventional electricity generators.

Home Radio (Components) have moved to 169 London Road, Mitcham, Surrey; their address for postal enquiries remains at PO Box 92, 215 London Road, Mitcham. They may be contacted by telephone at 01-648 3077.

Thomson-CSF claim the **world's fastest integrated circuit** operating at room temperature, an 11-stage ring oscillator with a gate delay of only 22 picoseconds. Structured from gallium aluminium arsenide/gallium arsenide junctions, the molecular-beam expitaxial process used is capable of controlling the crystal growth of the various layers down to the thickness of a single atomic layer.

On 5th April, a 'magazine' called Electronic Insight appeared on Prestel. This publication is intended to provide news, comment, product information and advertizing concerning electronics and telesoftware for microcomputer users. The potential exists to update news by the minute – 24 hours a day – and readers can pass from product news to product feature to comparison chart to stockists and retailers quickly. The magazine's editorial and commentary "maintains a fully independent view", says the publishers.

LEAKY FEEDER COMMUNICATION IN TUNNELS

Since the earliest days of radio – and certainly before the advent of broadcasting – attempts have been made to apply it in mines and tunnels using conventional apparatus of the day and relying on natural propagation of the waves. All these efforts were doomed to failure, and the reasons are now well understood: radio waves cannot propagate usefully in such conditions by any natural means.

by D. J. R. Martin,
B.Sc., Ph.D., F.Inst.P., F.I.E.R.E.

Radio waves cannot propagate naturally to any useful extent in mines and tunnels. Since about 1920, many attempts to use radio in such conditions have been made, without success.

There are two important exceptions to this generalization. At very low frequencies, certainly below any used for broadcasting, there is a limited capability of waves to penetrate rock or other strata. This property has recently been thoroughly investigated in the USA and exploited in the development of equipment intended for possible communication with trapped miners from the surface above. Through-the-earth ranges of up to 300 m have been demonstrated, but speech modulation is not practicable at the very low frequencies necessary and so the system uses baseband audio-frequency induction (for the 'down-link') and a coded c.w. carrier (for the 'up-link'). The possibilities have been studied but discounted for British mines, where average depths are greater and mining techniques less suitable, though equipment operating at slightly higher frequencies and using speech modulation has been successfully used by cave explorers. Useful ranges with speech modulation have also been achieved in South African mines, where geological conditions seem more favourable for propagation, and special equipment developed[1].

At the other end of the practical radio spectrum, waves in the u.h.f. range and above can often propagate usefully through a tunnel in what amounts to a waveguide mode, as first demonstrated by such investigators at Foot[2]. This form of communication is now being considered seriously by British Rail, whose operational train communications are normally in the standard mobile radio u.h.f. band following agreed European practice, and is also being investigated for the National Coal board by the University of Surrey. Negotiation of obstacles is the obvious difficulty, and in mining applications some practical means would need to be devised for re-directing the waves round a corner.

In 1955 Wyke and Gill[3] reviewed the situation as they then saw it with regard to mine communications and their paper gives many interesting references to the futile earlier experiments. One such investigator, in sheer negation of the traditional scientific attitude, made the infamous remark 'as successful means of wireless communication have not been discovered, details of the apparatus used and experiments carried out have not been included in this paper'.

Wyke and Gill themselves drew attention to the possibilities of inductive-type communication in coal mines, using frequencies in the range 15 to 150 kHz and relying on 'guidance' by any conductors present, such as power cables and telephone lines. Suitable equipment was, in fact, developed over the next decade or so and became a standard attachment to underground locomotives and cable-hauled man-riding trains. Generally, it was found worth while to install special well-positioned conductors or 'guide wires' for the purpose. By this means, reliable communication over distances of a kilometre or so could be obtained, especially if the conductor wire were galvanically connected to the base station instead of relying on inductive coupling there.

While fulfilling an important need in mine vehicular communications, the inductive equipment never achieved any success as a two-way personal system. The reasons for this were the fairly high transmitter powers required (about 5 W) and the resulting heavy batteries, the cumbersome loop or frame aerials involved, and the need for fairly close coupling to the conductor wire. In all applications, trouble was experienced with 'blind spots' or standing waves on the line, due to a lack of appreciation of the need for correct impedance transmission or periodic phasing, and this often was the limiting factor on range. Such inductive systems have now been completely superseded in UK coalmines by the later developments to be described, though they are still used widely in some overseas countries, notably the USA and West Germany.

The key to the revolution in underground communications came in 1956 with the publication by Monk and Winbigler[4] of a paper describing how v.h.f. radio communication had been successfully maintained with a moving train in a long railway tunnel. Following a logical idea, they first installed a standard coaxial cable (RG-8/U) through a section of the tunnel, connected to a normal base station at one end and having dipole aerials bridged across it at frequent intervals. This worked extremely well, and so the spacing between the aerials was then progressively increased until they had all been removed; good communication was still maintained throughout the length of the section, although it had not been possible before the installation of the cable. It became clear then, that the communication was through the stray fields of the cable, and there was born what is now known as the 'leaky feeder' or 'leaky cable' principle.

The author

Since graduating as a physicist at King's College London in 1945, D. J. R. Martin has run the gamut of the spectrum from d.c. to microwaves. At BSIRA (now the Sira Institute) he devised new techniques in low-level d.c. amplification and a.c. voltage stabilization. On joining the National Coal Board in 1958 he developed remote-control systems for roof supports in the automatic-coalface projects of the time, and introduced loudspeaking telephones to the coalface.

At the same time he became fascinated by the propagation problem which had for so long defeated all attempts at radio communication underground, and eventually obtained approval to embark on a full-time investigation.

He studied the scientific basis and developed the technology extensively for mining use; for this work he received his doctorate in 1973 – and was married in the same year.

He has continued to specialize in the subject, working closely with colleagues in Europe and North America. In 1981 he took early retirement from the NCB to work privately in the field. At present he is helping to supervise a study of microwave propagation in mine tunnels, being undertaken by the University of Surrey on behalf of the NCB for possible application in underground rescue operations.

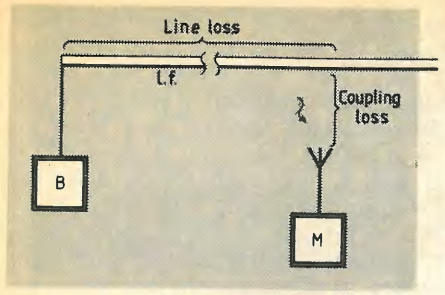


Fig. 1. Diagram of the basic leaky-feeder principle, depicting how the base station B communications both ways with the mobile station M through the leakage fields of the feeder LF. The total path loss is made up of two components, the line loss and the coupling loss.

With the realization that the cable itself was providing the necessary fields, the obvious next step was to substitute one known to have a higher leakage, and so a change was made from the coaxial to an unscreened two-wire line, a rather heavier version of the 'ribbon feeder' or 'twin-lead' then in common use for television downleads. This was also a more economical type of cable to use in terms of longitudinal attenuation for weight of copper.

Following this pioneering work of Monk and Winbigler, the leaky feeder principle was applied for communication in several American underground railway systems, notably the New York Subway, substantially unchanged in detail. But it was ignored by mining interests worldwide, and also in the UK generally, until the present author[6] commenced an investigation into its possibilities for mining use in 1966 at the same time that similar interest was being shown in Belgium[5]. Since then, the subject has been well served with expert theoretical and practical investigation in several countries, leading to a deeper understanding of the basic processes involved and improved designs of feeder cable, as will be described in this part of the article. At the same time, it has been appropriate to develop new and dedicated systems techniques to complement the new basic principle, and these will be described in Part 2.

Leaky-feeder principle

The basic leaky-feeder principle is illustrated in Fig. 1: B is a conventional two-way radio base station; LF is the leaky feeder, installed throughout the tunnel or environment where communication is required and connected to the base station in lieu of a normal aerial; M is a conventional two-way mobile or personal radio set, communicating with the base station through the leakage fields of the feeder.

The total path loss between the base station and the mobile is made up of two components: (a) the transmission loss within the feeder itself, between the base station and the region of the feeder in the vicinity of the mobile; and (b) the 'coupling loss', which is measured between the same region of the feeder and the aerial terminal of the mobile set.

Note that no assumptions are made here about the direction of transmission; the overall path is truly reciprocal, and so the principle is valid equally for mobile-to-base as for base-to-mobile communication. However, the processes involved are perhaps easier to visualize as operating in the base-to-mobile direction, and so consideration of coupling loss especially is usually from that point of view. Much of the theoretical work on the subject is similarly oriented; experimental observations, on the other hand, are often more conveniently based on the reciprocal path, with a mobile source inducing signals into the feeder. Reciprocity, of course, applies to the signal transmission only, and excludes the effects of any external interference or internal noise sources in the path; this factor is important when system aspects are considered, and will be covered in Part 2.

It may also be noted in passing that a leaky feeder may be used to allow direct communication between mobiles, without the intermediary of a base station. In this case the feeder operates in a purely passive or 'parasitic' mode, and two coupling losses are involved in the path. This form of operation was the basis of the pioneering Belgian work[5].

Consider now the separate loss components shown in Fig. 1. The feeder is basically a transmission line, in spite of its leakage, and so the signal within it will decay exponentially along its length – that is, the loss in dB will be directly proportional to distance. Generally, this attenuation will still be largely determined by the normal copper and dielectric losses, the leakage contributing little, and so a heavier and thus more expensive cable will give correspondingly better performance in this respect. It also follows that the attenuation rate is a stable characteristic which can be closely predicted or specified in the design of the cable.

The coupling loss, on the other hand, is a vague and variable quantity, being a function not simply of cable leakage (however that may be assessed) but also of the cable mounting, the environment, the characteristics and polarization of the mobile aerial and its distance from the cable. In a tunnel or any enclosed space pronounced multipath effects will inevitably occur, causing the 'instantaneous' value of the coupling loss to vary by 20 dB or more over short distances. Figures for coupling loss attributed by manufacturers to their cables, usually in unrealistic conditions, should be taken only as a very rough guide; an experienced system designer will prefer to work with more measurable fundamental characteristics of the cable together with a knowledge of its construction and a consideration of the environment and application concerned; in this particular respect the subject is best regarded as an art rather than a science.

The longitudinal attenuation of the feeder, again, is a well-established increasing function of frequency, which on this account should be set as low as possible. The leakage fields are generally considered to be substantially independent of frequency; however, other factors such as the size and efficiency of mobile aerials, the availability of suitable equipment or the need for compatibility with surface systems have discouraged the use of frequencies below 30 MHz. One therefore finds most applications of the leaky feeder principle operating in the standard v.h.f. mobile radio bands, with a minority in the h.f. and u.h.f. ranges.

Bifilar lines

For a decade or so following the original Monk and Winbigler publication, unscreened bifilar or 'balanced' types of line were used exclusively as leaky feeders, and were shown capable of giving very satisfactory performance. It became clear, in fact, that the high field strengths being

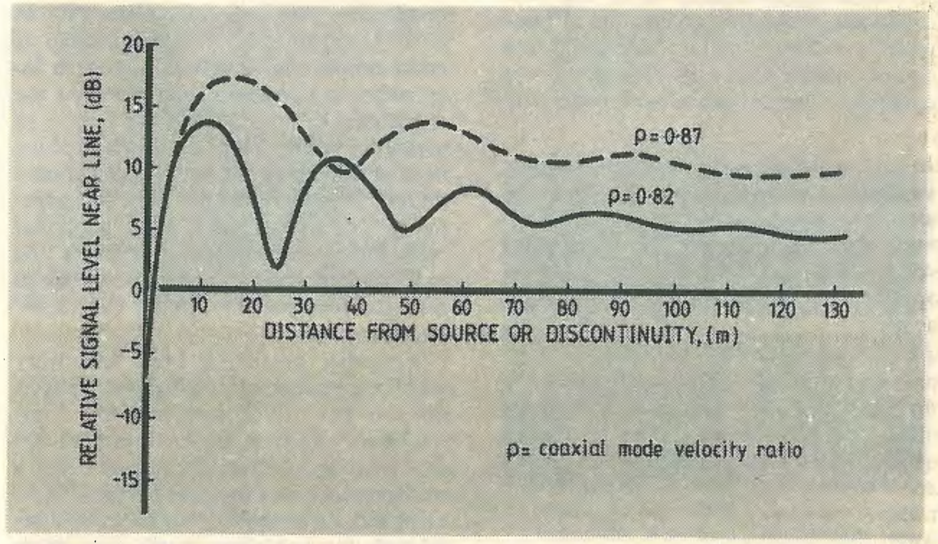


Fig. 2. Typical long standing wave as predicted to occur on a leaky feeder near the base station or any discontinuity as a result of waves travelling with differing velocities and attenuation rates (based on a computer analysis by A. M. Schmidt, Technische Hogeschol, Delft, Netherlands).

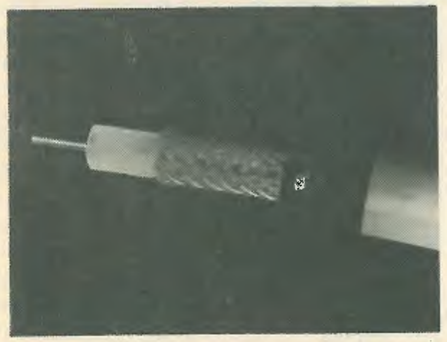
encountered in practice were far higher than would be expected theoretically from balanced cables[10], and this led to the conclusion that an imperfection or imbalance was the key factor in the success of the schemes. Such imbalance would result from inevitable assymmetries and irregularites in the mounting arrangements and in proximity effects, and led to the postulation of a continuous or continual interchange of energy between the low-field balanced mode and the high-field unbalanced or 'monofilar' mode; the balanced or 'bifilar' mode provided the longitudinal transmission, while the monofilar mode provided the coupling to the mobile set. It was shown experimentally that improving the balance by twisting the feeder, as might be expected, improved the longitudinal transmission at the expense of the leakage field, while 'careless' installation of an untwisted line close to metal structures or other cables would enhance the field locally to the detriment of the longitudinal transmission.

Deryck[7] has extensively studied the use of bifilar lines as leaky feeders, and has devised discrete 'mode converters' for introducing a controlled interchange of energy at specific points in a feeder[11].

Bifilar lines were intially considered for the first UK coal-mine system, commissioned at Longannet in 1970; the requirement there was for a radio system to serve a single 9 km tunnel linking four mines underground. But conditions there are extremely wet, and precautionary test showed that in such an environment the longitudinal attenuation of the simple 'ribbon feeder' proposed for use rose drastically and became extremely unstable. In the following years, evidence also came to light that bifilar feeders installed in the early railway stems were suffering from the effects of build-up of grime on their surfaces, and some were having to be cleaned regularly.

Coaxial feeders

For the Longannet system, further tests were made using a standard low-loss coaxial television downlead in which the outer conductor braid was applied in a 'loose weave' for cheapness and which could thus be presumed to have a high leakage. It was confirmed that this cable had broadly suitable characteristics for a leaky feeder: the longitudinal and coupling losses were both less favourable than those of the ribbon type in clean and dry conditions, but they were completely stable against surface moisture and grime, and the cable could even be installed close to a wall or alongside other cables without detriment to the attenuation. This experience established the open-braided coaxial type of feeder as the standard for use in UK mines, though the actual cables used are specially designed heavier versions than television downlead in the interests of robustness and a lower attenuation. This move back towards coaxial feeders has also been followed generally in railway and other transportation systems.



Typical leaky cable of the open-braided type. The dielectric is of semi-airspaced thread-and-tube construction, for maximum velocity ratio (about 0.87). The inner sheath is of polythene, the outer of pvc.

With the general adoption of coaxial types of leaky feeder, the need arose for a better understanding of their mode of operation, rather than the vague notion of leakage of energy through the interstices of the braid. At that time, screening efficiency of coaxial cables was assessed in terms of a 'surface transfer impedance', a measurable characteristic of the braid alone[8]. From this starting point, the present author developed a theory of continuous 'mode conversion', analogous to that of Deryck for bifilar lines, with the surface transfer impedance replacing the casual assymmetries of the bifilar line[9]. This led directly to some useful practical improvements and also predicted effects which have since been confirmed in experience, though the subject has now been dealt with more rigorously and elegantly by such writers as Wait[12] and Delogne[13].

The first important conclusion from the theoretical work – and in truth a prior hunch which had prompted it – was that the internal phase velocity of the cable in its normal coaxial mode was a key factor in determining the leakage fields, along with the more obvious surface transfer impedance of the braid. In fact, changing the dielectric from solid polythene to a semi-airspaced construction was shown to improve the external coupling in typical conditions by 20 dB – a change that would otherwise correspond to a reduction in braid cover, for example, from 93% to 39%. Unfortunately, the relationship is not a simple one, involving also such intangibles as the attenuation rate and the characteristic impedance of the line in its monofilar mode, parameters which must depend heavily on the environmental conditions. It is therefore not possible to formulate a 'figure of merit' for a particular cable for direct comparison purposes; the individual parameters and a knowledge of the environment have to be considered together.

A further prediction from the theory was the existence of standing-wave effects on the line, quite separate from any multipath effects in the environment and any reflections which may occur from a mis-termination of the line. Potentially the most serious of these disturbances occurs towards the beginning of the line where the base station is connected, and arises from the inadvertent launching of a faster wave, travelling at near free-space velocity, in monofilar mode on the outside of the feeder. Of near-identical initial amplitude to that of the 'true' leakage field, and in initial phase opposition, it results in a modal-interference standing wave being set up on the line – and thus in the near field – with a wavelength several times the free-space wavelength. Fortunately, it decays at the rate of the monofilar mode attenuation, which is generally quite high; further, its effects near the base station are mitigated by the lower system loss there, reducing the seriousness of any 'dropouts' which might result. In theory, it would be possible to suppress the launching of the interfering wave, or to launch another in phase opposition; however, the wave can be regenerated subsequently by any discontinuity, such as in mounting arrangements, which might occur further down the line. The effects have not proved serious or even noticeable in the operational systems now being installed by London Transport.

A typical long standing wave effect at the beginning of a line is illustrated in Fig. 2, which also shows the advantage in coupling obtainable through using a cable with a semi-airspaced dielectric (ρ=0.87) rather than foam (ρ=0.82). Both curves are for a frequency of 72 MHz, and assume coaxial-mode and monofilar-mode attenuation rates of 36 dB/km and 0.3 dB/m respectively. The monofilar-mode velocity ratio is taken as 0.95.

The theory has also served to discount early fears expressed that any attempt to introduce in-line amplification into a leaky feeder would risk instability through feedback between the outgoing and incoming sections of feeder unless non-leaky 'tails' were introduced. In fact, it can be confidently shown that with a typical feeder, such as the standard NCB open-braided type, repeater gains of well over 100 dB would be necessary to incur such risk.

Field characteristics

Confusion has occasionally arisen over the use of the terms 'monofilar mode' and 'mode conversion' in the operation of coaxial leaky feeders. In his early coupled-lines treatment the present author looked upon the monofilar mode as being the whole of the field external to the cable, and upon mode conversion as a continuous process which maintained it. In this simple view, the standing wave at the source as seen as a natural process of establishment and 'synchronization' of the mode. Later workers[12, 13] have taken the analysis further and resolved the external field into two major components. One of these is identified as the true leakage or 'spilling out' of the inner coaxial mode, travelling at coaxial-mode velocity and reaching its full amplitude immediately at the source. The other is the 'inadvertent' wave, launched at the source and at every discontinuity, and now accounting by interference for the long standing waves; this wave alone may be truly designated the 'monofilar mode', travelling strictly at monofilar velocity and also decaying at the higher monofilar rate. 'Mode conversion' becomes a discrete process, occurring only at the launching

points of the monofilar mode.

Resolving the wave on the line into two major components in this way simplifies a prediction of the nature of the resulting fields away from the line. It has been pointed out that the true monofilar mode will have a larger effective radius from the line than the continuous leakage mode, by virtue of its higher phase velocity. The relevant relationship approximates to

$$r_e=\frac{\rho\lambda}{2\pi(1-\rho)}$$

where r_e is the effective radius, λ the free-space wavelength and ρ the velocity ratio (in comparison with free space).

Thus, the leakage field of a typical solid-dielectric cable (ρ = 0.67) at 85 MHz would have a radius of 1.14 m. Use of a foam dielectric cable (e.g. 'Radiax', ρ = 0.82) would increase the radius to 2.6 m. Changing to a semi-airspaced type (e.g. NCB standard, ρ = 0.87) would increase it to 3.8 m. Against these figures, the velocity ratio of the monofilar mode is very close to unity and its effective radius will be several times greater again.

In fact, experiments in the Mersey (No. 1) tunnel have shown the field at 170 MHz to be reliably maintained across the full width of the 12 m carriageway from a feeder of semi-airspaced construction installed along one side. This suggests that in such practical situations of larger tunnels (or higher frequencies) the monofilar mode resulting from 'inadvertent' mode conversions must contribute substantially and usefully to the observed fields, though in smaller tunnels the effects are probably detrimental.

Whichever mode is considered, the physical picture of the external field is of a TEM wave with the outer conductor of the cable forming the inner conductor of a larger coaxial structure having the tunnel wall as its outer conductor. At low propagation velocities or higher frequencies, or in larger tunnels, the electric field lines will tend to curve and eventually will break away from the tunnel walls and return to the cable, in the manner of a Goubau wave supported entirely on the cable.

In this picture, the communication is entirely through induction fields, and so the use of the term 'radiating cables' in respect of leaky feeders in general is incorrect. It is true that any discontinuity which causes mode conversion may in the same process provoke 'inadvertent' radiation; this may be useful or even necessary in larger tunnels or at higher frequencies by extending the field in the same manner as the monofilar mode, but otherwise the effects are more likely to be detrimental by setting up long standing waves.

The simple picture becomes complicated by severe distortions in the fields, caused by irregularities and obstacles in the tunnel and by the induction-field equivalent of radiation 'multipath' effects. Polarization away from the feeder in such conditions is generally found to be random, while signal amplitudes can vary locally by the 20 dB or so that is typical for conventional mobile radio schemes in heavily built-up areas. But, as will be seen later in discussions on systems techniques, there need be no difficulty in accommodating such variations.

Practical feeders and their installation

Bifilar feeders have the advantage of being comparatively cheap and lightweight, but their use should be considered only in clean and dry conditions where they can also be installed at least 20 cm clear of walls, structures and other conductors; even then, their variability makes them unsuitable for use in long repeatered systems. The one-time favourite type, RG-86/U, has in fact not been manufactured for several years.

A bifilar type could, however, meet a need for an extemporized system in a dry and clean underground environment such as the worked-out stone quarries that have been used as store depots in Wiltshire, or for a temporary system for use during maintenance of a dry tunnel. A good choice in such a situation would be 300Ω

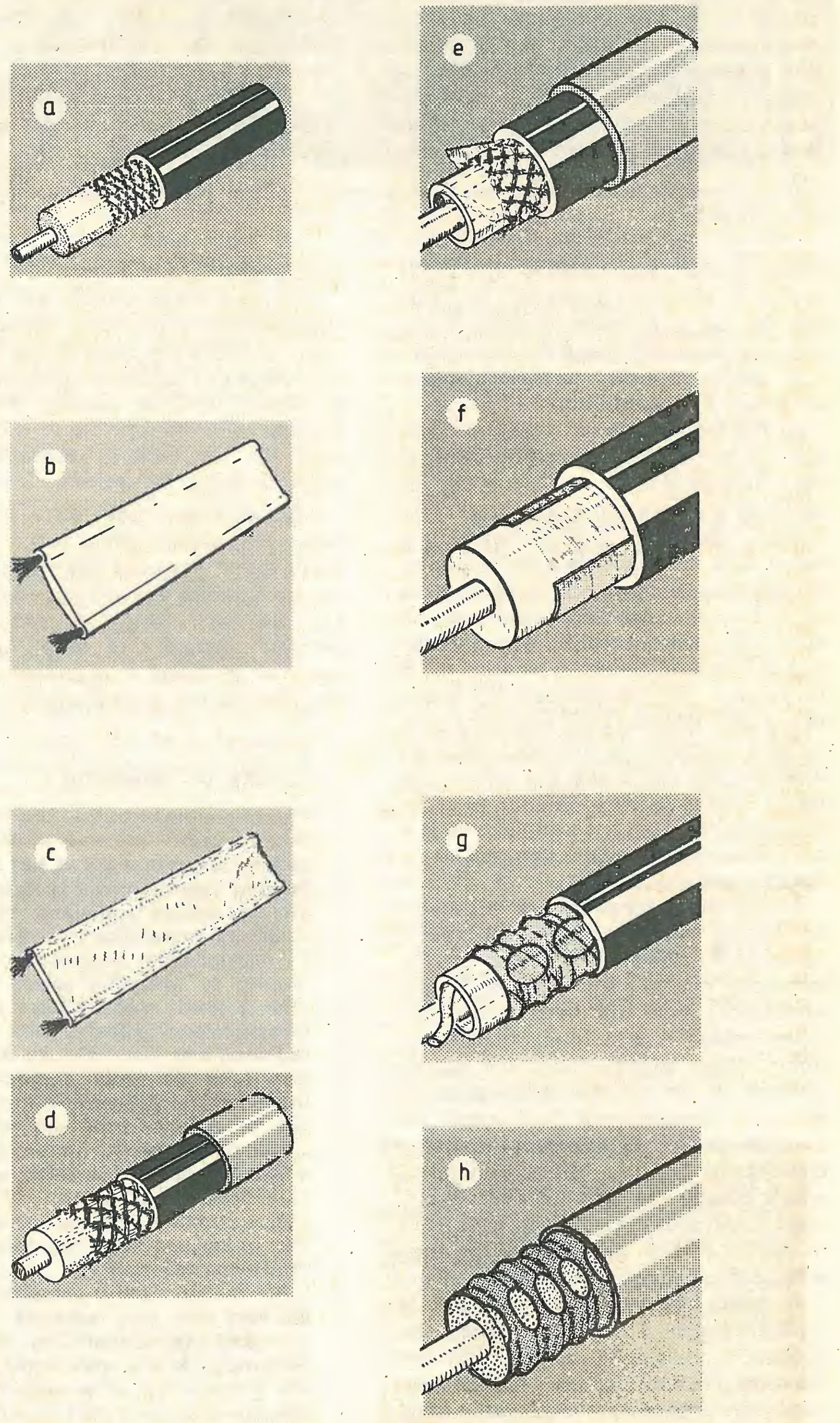
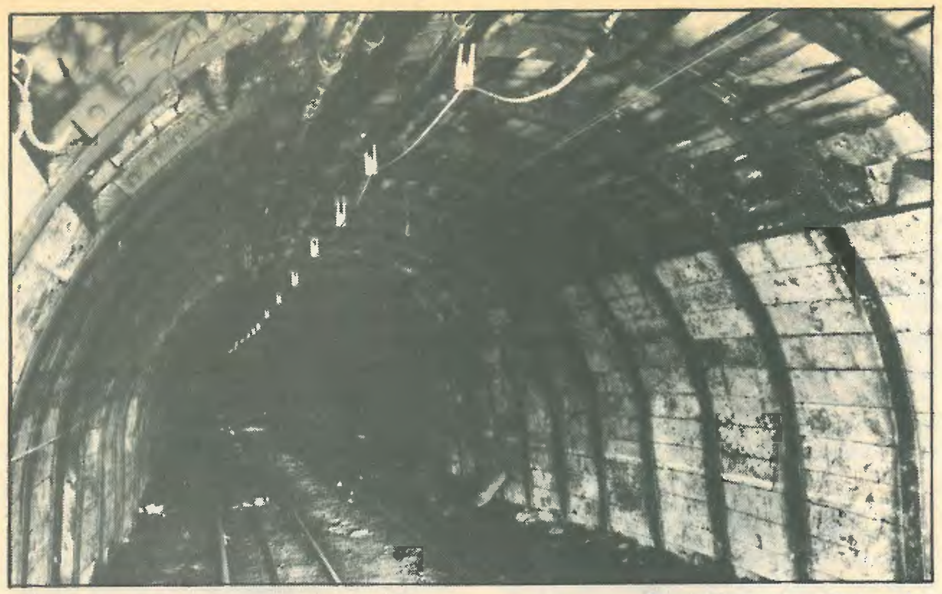


Fig. 3. Types of feeder used in underground communications. At (a) is a foam-dielectric, open-braid coaxial type, and at (b) and (c) two kinds of bifilar feeder for use in clean conditions. A solid-dielectric cable with an open braid is seen at (d). The open-braid coax. at (e) has a 'thread and tube' semi-airspaced construction and (f) shows an early design with a continuous slot in the outer. Two kinds of cable with discrete holes are (g) with punched apertures and thread-and-tube and (h) Radiax, which has milled holes in the screen and foam dielectric.



Leaky feeder installed in a coal mine. A line-powered repeater of early design is visible at the top of the picture.

'ribbon feeder', such as BICC type T3101 or T3129. However, it may well be found that this has to be made to order; in any case, for reasonably long unbroken lengths it is usually necessary to place a special order with the manufacturer.

In the type of use envisaged, a bifilar line can be suitably installed by suspending it in loops of nylon cord below existing cables or from any convenient brackets or fixings. Suspension from a tunnel roof, rather than attachment to a wall, will normally ensure the necessary clearances more easily but otherwise will not substantially affect the performance. Such an installation can be used simply in parasitic mode to enhance the communication between two or more personal radio sets over ranges up to a kilometre; or the range might be increased by connecting one end of the feeder directly to one of the sets – ideally through a balun but not essentially so. It is not necessary to terminate free ends of the feeder in any way.

In any more permanent or vulnerable application, the use of a coaxial type of feeder is advisable. Of the various constructions that are now available, many experienced users still prefer the original 'open-braided' type for its mechanical flexibility, comparative cheapness and insensitiveness of its internal attenuation to mounting arrangements. Such cables are available with a semi-airspaced dielectric construction (such as 'thread and tube') which ensures the high internal velocity ratio (about 0.87) that is so advantageous in respect of both general coupling level and the effective radius of the field.

A braid coverage factor of 67% is typical, giving a surface transfer impedance of perhaps 700 mΩ/m at 30 MHz, the usual reference frequency for that measurement. Generally, a braid with a fairly long 'lay' (smaller braid angle) gives the best compromise in attenuation and coupling for a given weight of copper. But decreasing the braid angle below about 30° may unduly weaken the cable against bending, and the situation is also influenced by systems aspects such as the span loss between repeaters.

Braided leaky cables are usually supplied with a polythene sheath, either alone or with a pvc outer sheath for flame retardance and mechanical protection. The polythene itself is usually considered necessary for waterproofing, though the need for that with a thread-and-tube dielectric might be queried.

Like all coaxial leaky feeders, braided cables will usually give a better coupling performance if spaced away from continuous conducting surfaces and other cables. On the other hand, an occasional or periodic proximity of metal (as might occur in the fixing arrangements) will often enhance the general coupling by setting up 'inadvertent' mode conversions or even radiation. Mounting arrangements have no influence on internal attenuation except at the higher frequencies (e.g. in the u.h.f. range) or with excessively open braids. Surface contamination affects neither coupling nor attenuation.

Another early type of leaky coaxial cable had a solid outer conductor with a continuous longitudinal slot, giving an aperture of typically 25% of the circumference, and having a solid or foam dielectric. Such a cable has sometimes been considered intermediate in its mode of operation between a bifilar and a coaxial type. Indeed, it has proved to be very dependent on mounting position, in the manner of a bifilar line, yet with no compensating advantage of cheapness.

A type popular in the USA and Canada has discrete holes milled in an otherwise solid corrugated outer conductor ('Radiax'). In mode of operation it is closely similar to the open-braided type. However, there is some evidence, with theoretical support, that 'hole size' is the key factor in determining the susceptibility of the longitudinal attenuation of a cable ot the mounting arrangements and surface contamination. For a given surface transfer impedance – and thus coupling efficiency – immunity to such external effects will be better with a larger number of small holes than the converse, and in this respect the braided type must be supreme (it should be noted that total optical cover is not a reliable indicator of coupling efficiency, and is not in question here).

The milled-hole type has a foam dielectric, which is slightly less favourable than the thread-and-tube construction in respect of the all-important phase velocity.

'Leaky' or 'radiating'?

The terminology to be used in referring to these 'assisted-propagation' techniques and systems has been a bone of contention for several years. The committee organizing the 1974 Guildford conference (see refs. 10 to 14) decided upon 'leaky feeder' in preference to 'radiating cable' on the main grounds that the basic process involved was probably not one of radiation but induction, an assumption now generally confirmed by theorists. The distinction is not simply one of distance from the cable, as would be the case for a normal aerial: a leaky feeder in perfectly uniform conditions cannot radiate at all except at its extremities, whereas the induction or leakage fields can and do extend along its length.

But the situation is complicated by the fact that true radiation can be provoked inadvertently, or even intentionally, at any point in the line by the introduction of irregularities or discontinuities into the line or its environment, and these will exist naturally to some extent in any practical system. In a tunnel the additional effect may be useful, but is more likely to be detrimental. Certainly, in a surface system (where the techniques can also be used) any radiation will run counter to the main purpose in such circumstances, that of spectrum conservation. The term is clearly best avoided.

On the other hand, the epithet 'leaky' has outraged some manufacturers who seem more concerned about the image of their product than with scientific accuracy, and 'leaky feeder' has come in for ribald comment. There have even been some qualms expressed in academic circles about the strict meaning of 'leaky' as applied in transmission-line theory. As for 'feeder' or 'cable', the only reason for preferring the former is that conceivably an open-wire line could be used (and in fact was considered at one time for NCB use) and this could hardly be described as a 'cable'.

The present writer stands by 'leaky feeder' in alluding to the techniques and systems; the term is certainly preferred by most serious investigators of the subject, and in particular is now general in the USA and mainland Europe. 'Leaky cable' remains a useful variant in reference to the product itself.

It is also worth noting that the international working group advising the XVth Plenary Assembly of the CCIR in such matters recently came down in favour of 'leaky cable', and this recommendation is likely to prevail.

A somewhat similar type manufactured in the UK employs a 'punched tape' construction, with an outer conductor made of copper foil in which discrete holes have been previously punched. The holes here are even larger than those of the milled-hole type, and so such cables should be used with discretion where close-mounting is inevitable. The dielectric construction is the favoured thread-and-tube.

Leaky coaxial cables are available in both 50Ω and 75Ω characteristic impedances, and it may well be asked which is the better choice. So far as the leakage is concerned, the considerations are more complex than might appear, but whatever difference may exist in practice is not likely to be significant. When it comes to longitudinal transmission efficiency, standard textbook treatment applies as for conventional coaxial cables, and shows that 75Ω is close to the theoretical optimum.

For a fuller treatment of the various types of practical leaky feeder, and an assessment of their relative performance, reference is recommended to the paper by Cree and Giles[14].

INIEX systems

It has been shown earlier that in order to account for the strong fields that are consistently maintained at distances of many wavelengths from a leaky feeder it is necessary to postulate the occurrence of 'inadvertent' mode conversions at irregularities in the feeder or tunnel. The resulting monofilar modes generally have a larger effective radius than the 'true' leakage field, though they are attenuated comparatively rapidly along the line and so require continual regeneration.

Delogne[15] has developed a system in which inadvertent mode conversions are replaced by deliberate conversions, effected through special 'mode converters' which may be inserted into the cable run at carefully determined points. Furthermore, the cable itself is a conventional non-leaky type; the true leakage field is thus abandoned and eliminated, and communication is entirely through the closely regulated monfilar mode. The system has been sponsored and promoted by INIEX, the Belgian 'Institut National des Industries Extractives'.

The main advantage of this 'INIEX-Delogne' system is that the interchange of energy between coaxial and monofilar modes is entirely under the control of the system designer and thus can be optimized for best use of the energy available by careful design and spacing of the mode converters. At their simplest, these devices could comprise a single large hole in the outer conductor; in practice, both conductors are usually interrupted and the gaps bridged by reactive elements. The spacing is typically 100 m, but the system may be 'graded' by varying the spacing according to the residual energy in the line and so maintaining a consistent performance regardless of distance form the base station. The devices are normally installed in pairs, spaced a quarter-wavelength apart, to give a forward directivity to the launch of the monfilar mode.

Since only one mode is present, there is no risk of potentially troublesome long standing waves. The system can be used over a wide range of frequencies, typically down to 2 MHz, by designing the mode converters to suit rather than by changing the design of a leaky cable. Against this, it could well be disadvantageous to have to interrupt the cable run at such short intervals. More important, perhaps, is the need for the cable to be kept well clear of walls or metal structures for the whole of the run, for if the monofilar mode becomes dissipated inadvertently by such a proximity effect it cannot be regenerated until the next converter, and an extended 'blind spot' could result.

At frequencies above the tunnel cut-off frequency – say, in the u.h.f. band for a road or twin-track rail tunnel – the INIEX-Delogne system operates differently. Here, the monofilar mode is likely to be attenuated too rapidly to be useful. Instead, the radiation which the mode-converter also provokes now launches a wave in the tunnel itself. In this case, the cable mounting arrangement between mode converters is immaterial since the cable is not called upon to support a monofilar mode. At even higher frequencies, of course, the tunnel propagation improves beyond that of the coaxial mode of the cable itself, and so the system then becomes superfluous except, perhaps, in the negotiation of corners and obstacles.

In a later development [16] the discrete mode-conversion devices are replaced by lengths of leaky feeder inserted into the otherwise non-leaky cable run. This 'leaky sections' technique operates broadly on the same principles as before, but offers the advantage that the alternating leaky and non-leaky sections could be arranged in the manufacture of a continuous cable, so eliminating the need for frequent interruptions and connections of the cable in installation.

The length of a leaky section in such a system is fairly critical, and the optimum can be shown to be equal to

$$\frac{\rho\lambda}{2(1-\rho)}$$

where λ is the free-space wavelength and ρ is the velocity ratio of the leaky cable section. Not surprisingly, this is one-half the length of the long standing wave which the original coupled-line theory shows would be set up at the beginning of a continuous leaky feeder of the same type.

It also follows simply that the monofilar mode set up by such a leaky section will have an initial amplitude 6 dB greater than the true leakage field of the corresponding continuous feeder. Furthermore, of course, it will have the advantage of the increased effective radius over the leakage field. Against this, the amplitude will decay at the rapid monofilar rate, and so the 6 dB advantage will usually be lost before the next leaky section.

Again, the leaky section will operate more as a radiator than as a mode converter at frequencies above tunnel cut-off.

A prime objective in the development of these INIEX systems has been to achieve the maximum possible longitudinal range from a system without recourse to in-line repeaters or other active techniques. Once active devices become necessary, as they normally do in any case beyond a few kilometres, the use of these systems becomes less attractive. However, there is no fundamental reason why they should not be used, in the same way as 'conventional' leaky feeders, with the active systems techniques to be described in Part 2.

To be concluded

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Further reading

Other authors who have contributed significantly to a basic understanding of leaky feeders include J. C. Beal, Q. V. Davis, P. Degauque, B. Demoulin, A. S. C. Fernandes, J. Fontaine, R. Gabillard, D. A. Hill, R. Johannessen, R. J. Slaughter and A. M. Schmidt. □

MICROPROCESSOR-CONTROLLED LIGHTING SYSTEM

This final article describes the overall operation and performance of the prototype lighting system. Details are given of the operating system, equalization table, and the hardware required to set-up the control desk's processing and recording modes.

by John D. H. White and Nigel M. Allinson*

Before discussing the operating software used in the lighting system, its relevance is best understood by considering the layout of a typical control desk, as shown in Fig. 1, and how such a desk is operated. The desired lighting pattern is set on the channel faders (presets), and this pattern is stored in the processor-system memory by pressing the 'record' button associated with a particular master fader, or 'master preset'. This pattern will be recalled and sent to the dimmer modules whenever its associated master preset is not at zero. Assuming for the moment that only one master preset is at a non-zero setting at any one time, any other master preset may now be used and another lighting pattern set in the same manner. Hence, a complete lighting pattern may be stored for each master preset.

There are two ways in which these stored patterns may be controlled using the master presets.

– Scaling – the equivalent of analogue control-desk processing – in which each preset level is multiplied by the master preset level and the resulting signals sent to the dimmer channels. Relative levels of the channels are maintained at all times.

– And stepping, where the master preset level is compared with the stored preset levels and the lesser of the two levels used for output. This type of processing is used to build up a lighting pattern, i.e., all dimmer outputs rise according to the level of the master preset and then stop at their predetermined levels. In an analogue control desk, this type of processing would require very complex circuits.

By using more than one master preset at a time, lighting patterns can be gradually changed from one stored pattern to another. As the operating program endlessly polls all the faders and record buttons, any lighting pattern produced by a combination of master and channel presets may be recorded by simply pressing the appropriate master-preset record button.

Operating Software

The operating program and the 'look-up', or equalization table, are contained in just over 1/2Kbyte of p.r.o.m. The requirements for r.a.m. depend on the overall size of the control desk. Around 256 bytes are required for the operating program and $N(M+1)$ bytes for lighting pattern storage, where N is the number of channel presets and M the number of master presets. Except for the largest of systems, 2Kbytes of r.a.m. will suffice. Organization of the data structures is shown in Fig. 2. The present memory stores the lighting-pattern preset levels associated with each master preset. The output-buffer memory is used to store the required lighting pattern temporarily, before the levels are converted to output signals for the dimmer modules, using the equalization table.

The equalization table performs two important functions. Firstly, the scaling process entails the multiplication of numerous channel and master preset levels. Without an external multiplier unit, most microprocessors carry out multiplication relatively slowly (some recent microprocessors, such as the 6809 and 9995 have such a multiplier internally). The multiplication problem could have been solved by using a logarithmic a-to-d converter, but in this case, logarithmic-law faders were used together with a look-up table containing base 2 antilogarithms for the 256 possible levels – hence, multiplications become simple additions.

Secondly, the table provides compensation for the non-linear relationship between the fader position and the subjective brightness of the lamps, mentioned in the first article. This code transformation is fairly difficult to formulate, and will be of more general interest than the code transformation used in the prototype system which combines both this subjective brightness compensation and the antilogarithm conversion, so this coding is given in Table 1.

The operating program is not listed because it is specific to the processor used and consists of only eight short sub-routines and three core-routines for lighting pattern recording and processing. However, using the flow-chart of Fig. 3, it should be possible to program most microcomputer systems to provide the facilities described. The program tests data present on the data bus to decide whether scaling/stepping processing, or recording mode is required. The hardware needed for this is described in the next section. Note that, to reduce processing time to a minimum, there are a number of conditional branches dependant on channel or master levels being zero.

* University of Keele

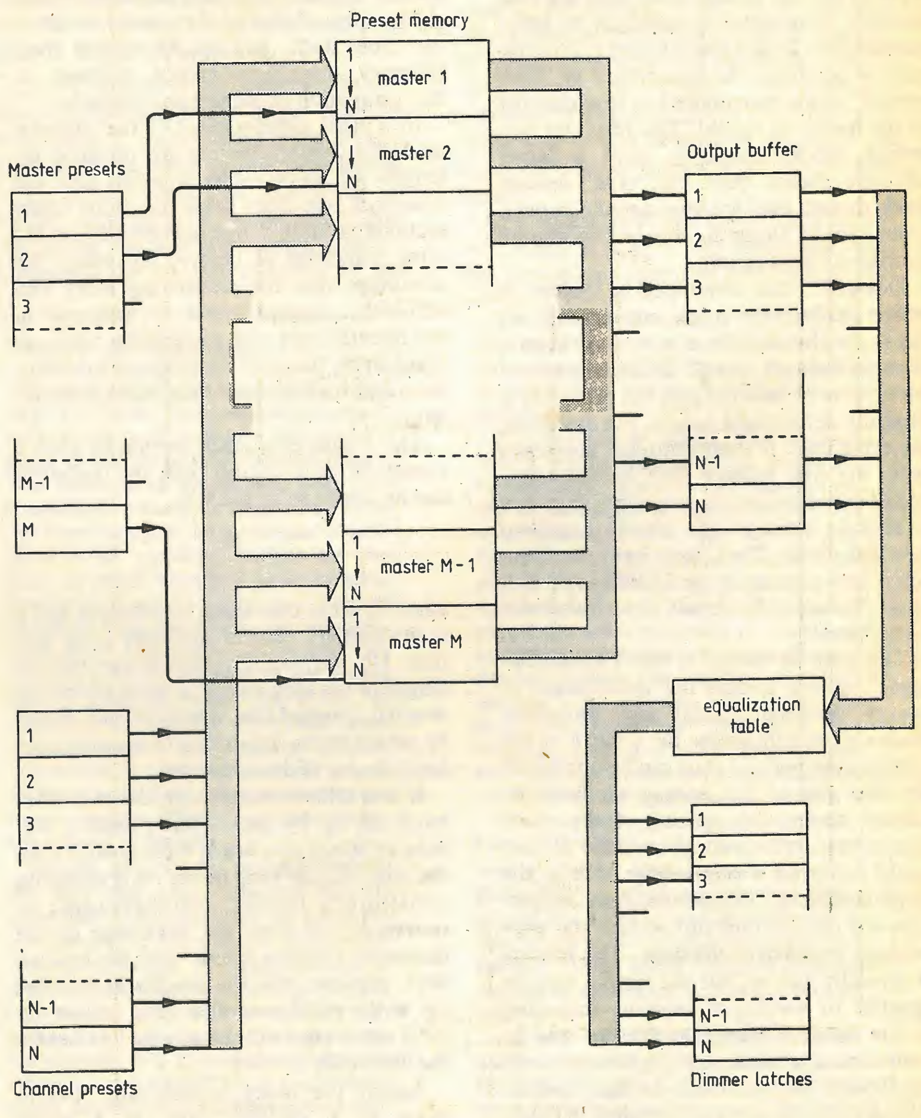


Fig. 2. Data structures used in the control-desk software. M is the number of master presets, and N is the number of channel presets.

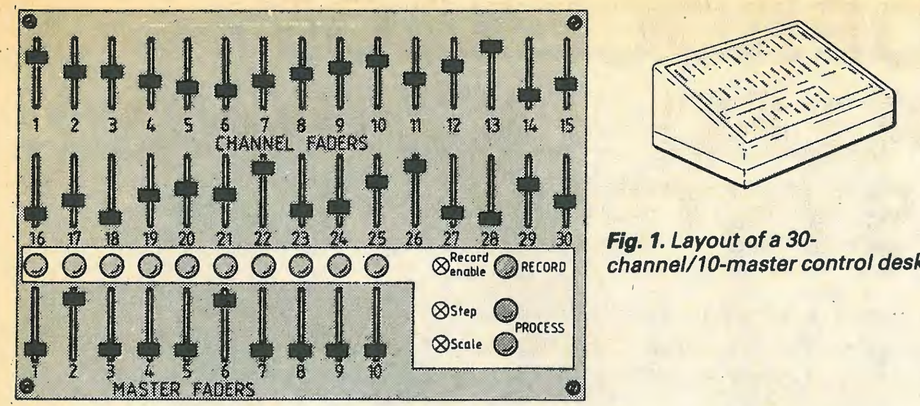


Fig. 1. Layout of a 30-channel/10-master control desk.

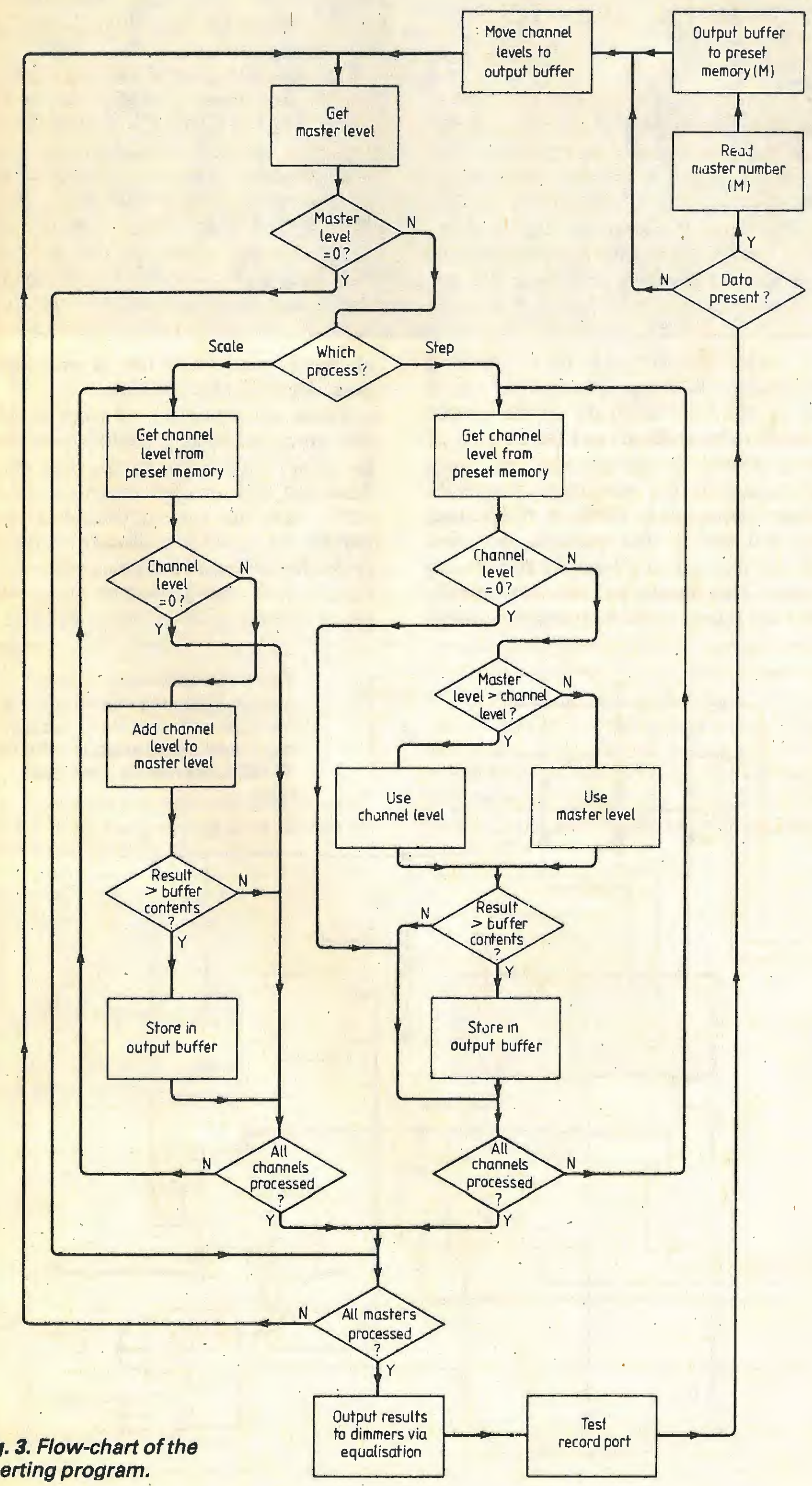


Fig. 3. Flow-chart of the operting program.

Process/record select circuits

The operating program must test whether stepping/scaling processing, or pattern recording is required. This could be achieved by connecting the control desk's record and process keys, through some form of keyboard encoding, to a programmable i/o device (such as the 8155/6). However, since mapped-memory techniques are used for all other data input and output, a single i/o port can be connected directly to the data bus which is enabled when the IO/$\overline{M}$ status line goes high. Figure. 4 shows the process/record-select circuit. When the 'record enable' key is pressed, the octal encoder (74148) is enabled and its output will stay high until a master-preset record key is pressed. The three RS flip-flops connected to the octal encoder are reset, and hence the 4-bit binary counter (74163) is enabled. The counter outputs are connected, through a 4-to-16-line demultiplexer (74154), to sixteen cross-lines in the master-preset 'record' key-matrix. When a key is pressed, at least one of the encoder's outputs goes low and disables the counter. The three-state buffer is enabled when either $\overline{E}$, W/$\overline{R}$ or M/$\overline{IO}$ is low, and the input data is transferred to the processor data bus. Also, the four inputs to the NAND gate (1/2 7420) are high, and on the next rising edge of the system-enable, $\overline{E}$, a '0' is clocked out of the D-type flip-flop and the four RS flip-flops are reset. The next $\overline{E}$ pulse will enable the system again. The 0 input of the octal encoder is not used, as a low level on this input will cause all three outputs to be high (i.e., equivalent to no key being pressed).

Table 1: Code conversion for subjective brightness correction: I is input, in hexadecimal form, O is output, also in hexadecimal form and L is relative luminous intensity.

I	O	L	1F	3D	224	I	O	L	5F	60	160	I	O	L	9F	7C	97	I	O	L	DF	A1	32
			20	3E	223				60	60	160				A0	7D	95				E0	A1	32
00	00	255	21	3F	222	40	51	191	61	61	158	80	6E	129	A1	7D	95	C0	8D	63	E1	A2	30
01	12	254	22	3F	222	41	51	191	62	61	158	81	6F	126	A2	7E	93	C1	8D	63	E2	A3	29
02	17	253	23	40	220	42	52	189	63	62	156	82	6F	126	A3	7E	93	C2	8E	61	E3	A4	28
03	1B	252	24	41	219	43	52	189	64	62	156	83	70	124	A4	7F	91	C3	8E	61	E4	A5	27
04	1E	251	25	41	219	44	53	187	65	63	154	84	70	124	A5	7F	91	C4	8F	59	E5	A6	26
05	20	250	26	42	217	45	54	186	66	63	154	85	71	122	A6	80	89	C5	8F	59	E6	A7	25
06	22	249	27	43	216	46	54	186	67	63	154	86	71	122	A7	80	89	C6	90	57	E7	A8	24
07	24	248	28	43	216	47	55	184	68	64	151	87	71	122	A8	80	89	C7	91	56	E8	A9	23
08	26	247	29	44	214	48	55	184	69	64	151	88	72	119	A9	81	86	C8	91	56	E9	AA	22
09	27	246	2A	45	213	49	55	184	6A	65	149	89	72	119	AA	81	86	C9	92	54	EA	AB	21
0A	29	245	2B	45	213	4A	56	181	6B	65	149	8A	73	117	AB	82	84	CA	92	54	EB	AC	20
0B	2A	244	2C	46	211	4B	56	181	6C	66	147	8B	73	117	AC	82	84	CB	93	52	EC	AD	19
0C	2B	243	2D	46	211	4C	57	179	6D	66	147	8C	74	115	AD	83	82	CC	94	51	ED	AE	18
0D	2D	242	2E	47	209	4D	57	179	6E	67	145	8D	74	115	AE	83	82	CD	94	51	EE	AF	17
0E	2E	241	2F	48	208	4E	58	177	6F	67	145	8E	75	113	AF	84	80	CE	95	49	EF	B0	16
0F	2F	240	30	48	208	4F	58	177	70	67	145	8F	75	113	B0	84	80	CF	95	49	F0	B1	15
10	30	239	31	49	206	50	59	176	71	68	142	90	75	113	B1	85	78	D0	96	47	F1	B3	14
11	31	238	32	49	206	51	59	176	72	68	142	91	76	110	B2	85	78	D1	97	46	F2	B4	13
12	32	237	33	4A	204	52	5A	173	73	69	140	92	76	110	B3	86	76	D2	97	46	F3	B6	12
13	33	236	34	4B	203	53	5A	173	74	69	140	93	77	108	B4	86	76	D3	98	44	F4	B7	11
14	34	235	35	4B	203	54	5B	171	75	6A	138	94	77	108	B5	87	74	D4	99	43	F5	B9	10
15	35	234	36	4C	201	55	5B	171	76	6A	138	95	78	106	B6	87	74	D5	99	43	F6	BB	9
16	36	233	37	4C	201	56	5C	169	77	6A	138	96	78	106	B7	88	72	D6	9A	41	F7	BD	8
17	37	232	38	4D	199	57	5C	169	78	6B	135	97	79	104	B8	88	72	D7	9B	40	F8	BF	7
18	38	231	39	4D	199	58	5D	167	79	6B	135	98	79	104	B9	89	70	D8	9B	40	F9	C1	6
19	38	231	3A	4E	197	59	5D	167	7A	6C	133	99	7A	102	BA	89	70	D9	9C	38	FA	C4	5
1A	39	229	3B	4E	197	5A	5E	165	7B	6C	133	9A	7A	102	BB	8A	68	DA	9D	37	FB	C7	4
1B	3A	228	3C	4F	195	5B	5E	165	7C	6D	131	9B	7A	102	BC	8B	67	DB	9E	36	FC	CA	3
1C	3B	227	3D	4F	195	5C	5E	165	7D	6D	131	9C	7B	99	BD	8B	67	DC	9E	36	FD	CF	2
1D	3C	226	3E	50	193	5D	5F	162	7E	6E	129	9D	7B	99	BE	8C	65	DD	9F	34	FE	D6	1
1E	3C	226	3F	50	193	5E	5F	162	7F	6E	129	9E	7C	97	BF	8C	65	DE	A0	33	FF	FF	0

Hence up to seven modules may be used, each with 16 master presets.

Processing-mode keys are simply latched by an RS flip-flop and connected to D_7 of the data bus. Unlike the record-key data, they do not form a destructive read circuit.

Conclusion

For the 8085A microprocessor used and with a system clock of 3MHz, Fig. 5 shows the experimentally determined relationship between processing speed (that is, the time taken for all the output channels to be updated) and the number of master presets in use at any one time. Assuming that the minimum acceptable update frequency is 20Hz, a 60-channel desk will operate fast enough, provided that less than about 20 master presets are in use. This limitation will not usually effect operation as the maximum number of master presets in use at one time is normally from four to eight.

There are a number of ways in which this prototype system could be extended. In most stage applications, the control desk and dimmers are remote from each other, and the system described would require an expensive 40-way connecting cable. Some form of high-speed serial interface, with high noise immunity, would be of greater practical use. Because the major cost of most installations is in the dimmer modules, the replacement of an analogue control desk with a digital one would most effectively be achieved by providing a low-cost interface between the digital output data and the existing 0-10V-direct voltage-controlled dimmers. More permanent lighting pattern storage could be easily achieved by providing either battery back-up for some of the r.a.m., or a tape interface. In exacting situations where the colour, or hue, of lights has to be maintained as their luminosity changes, the software could be extended, and lighting sets with three primary coloured gels used.

Estimates of the cost of a digital control desk compared with a conventional analogue desk suggest that the former solution is the cheaper alternative for systems with more than 40 channels and 20 master presets. The fixed cost element of the microprocessor system is offset by the absence of a diode-matrix board and the large reduction in the number of faders.

The authors are grateful for the use of the microprocessor development facilities provided in the final-year Electronics laboratory at the University of Keele, and the technical help provided by B. W. Cornes and E. J. T. Greasley. □

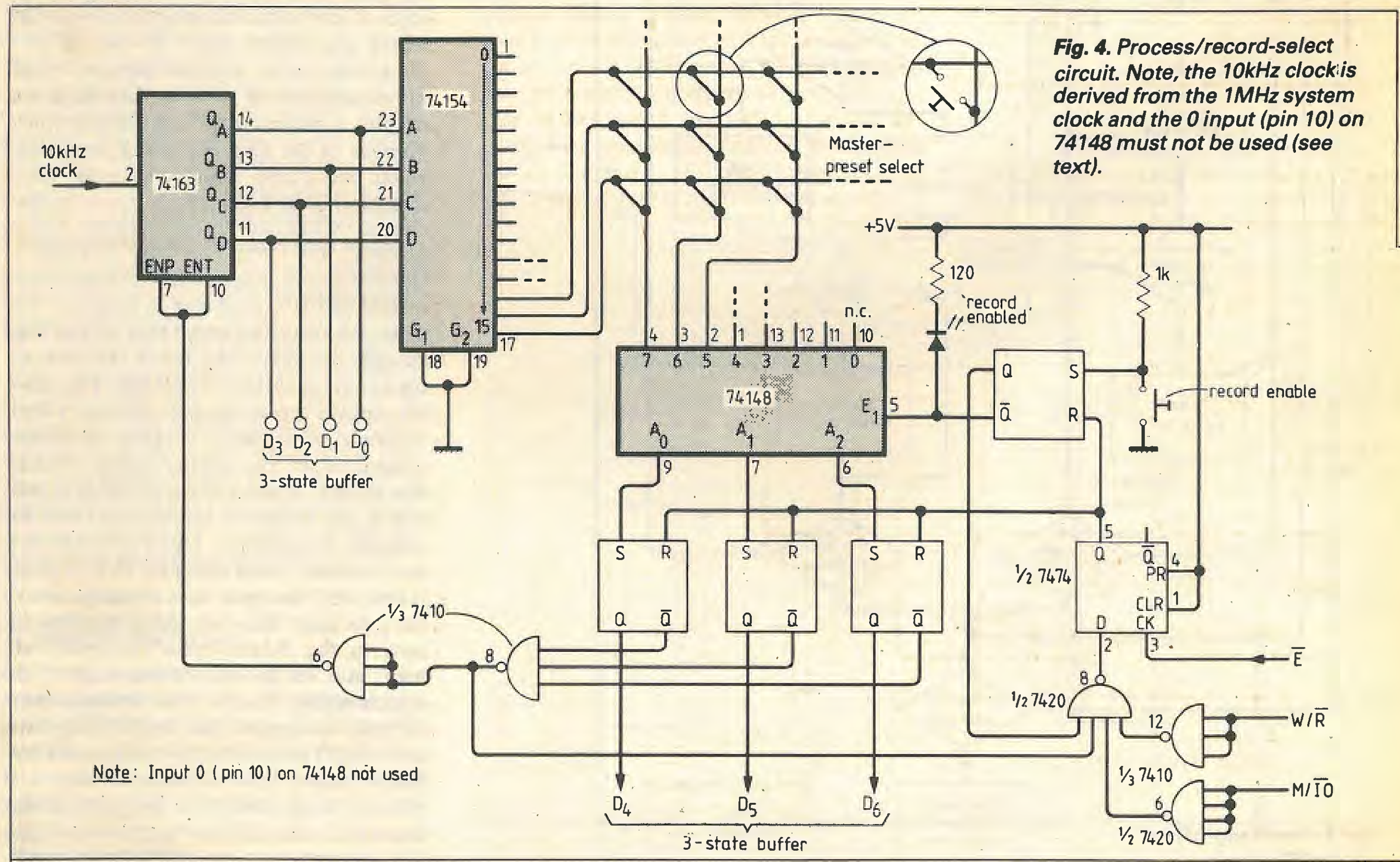
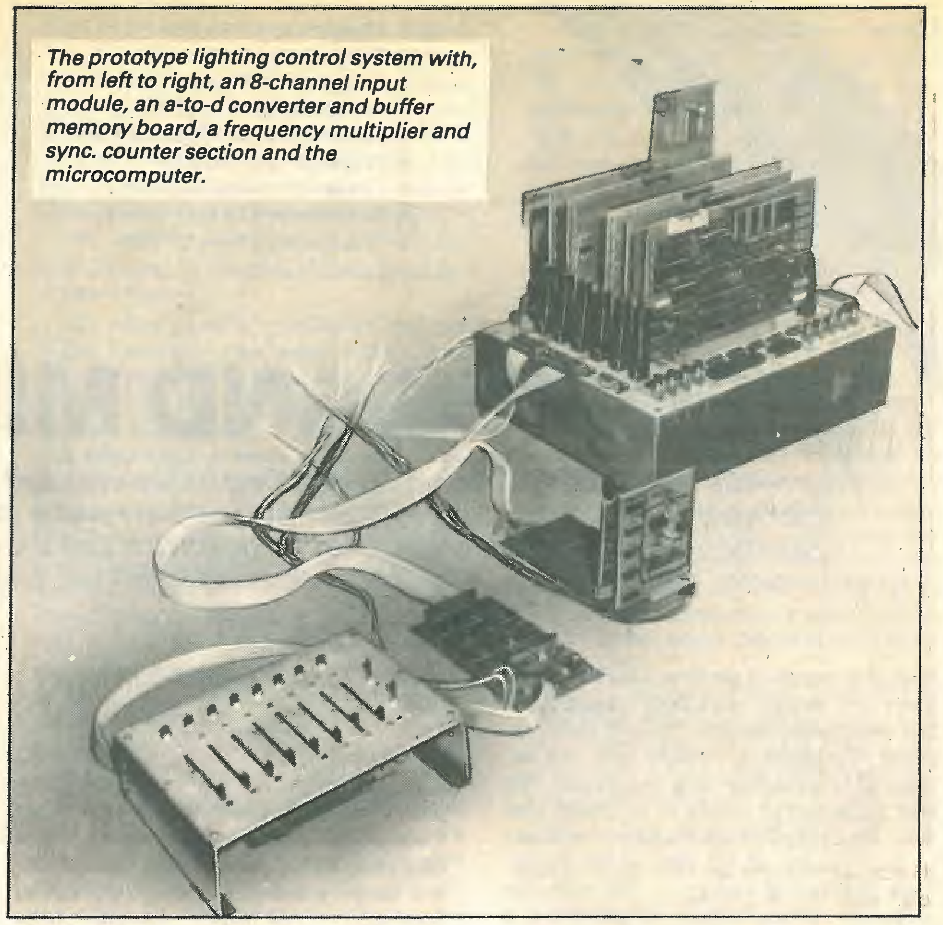


Fig. 4. Process/record-select circuit. Note, the 10kHz clock is derived from the 1MHz system clock and the 0 input (pin 10) on 74148 must not be used (see text).



The prototype lighting control system with, from left to right, an 8-channel input module, an a-to-d converter and buffer memory board, a frequency multiplier and sync. counter section and the microcomputer.

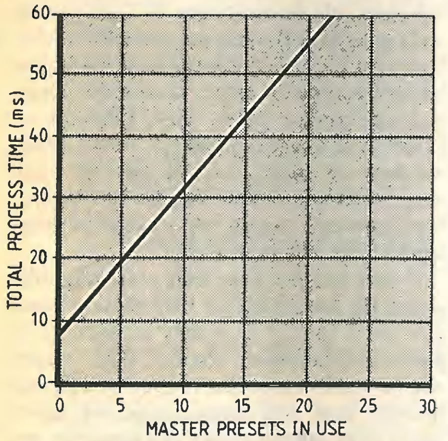
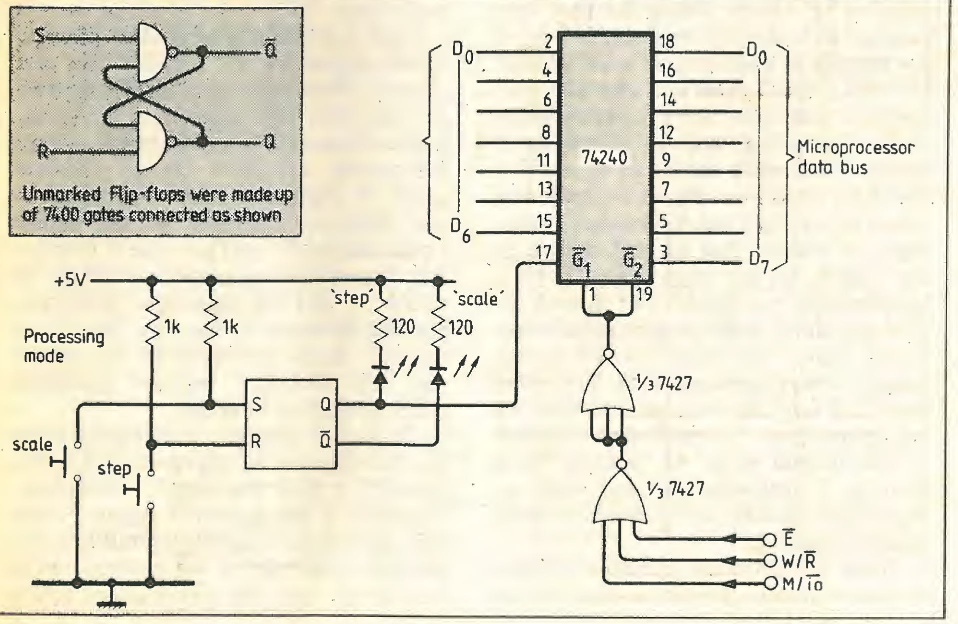


Fig. 5. Experimentally determined process time, i.e., time for updating all dimmers, for a 60-channel control desk.



From page 46

Computation

The calculations illustrated may be carried out with the aid of a pocket calculator. Access to a computer is clearly advantageous and Fortran programs have been previously published [3] for the procedures required. For convenience, four simple Basic programs are presented on page 46 for the pole and zero calculations constituting the most difficult part of the computation. These programs may be readily incorporated into a complete digital filter design package for a desk-top computer. The programs have been developed and listed on a Hewlett Packard HP 85 microcomputer which allows several statements per line, separated by @. The command DISP may have to be replaced by PRINT in may Basic implementations.

Bibliographical references have had to be held over and will be included in a third part of this series.

A Heretic's Guide to Modern Physics

THEORIES AND MIRACLES

Enormous gaps exist in our understanding of Nature, and many of our fundamental theories are not very credible. In a controversial review of current doctrine in nine instalments, Dr Murray investigates the electromagnetic theory, photons, duality, quantization, matter-waves, indeterminacy and haziness, and reviews the state of physics today.

by W. A. Scott Murray,
B.Sc., Ph.D.

Many thousands of professioanl radio engineers can design television transmitters, and almost anyone can build a radio receiver, but there is nobody who can explain in a plausible and watertight way how radio energy comes to be transferred from the Crystal Palace transmitting tower to the H-aerial on the roof of my house. This transfer of energy – the radiation process – is miraculous, if we define a "miracle" as a physical occurrence for which we can offer no physical explanation. (I'll just say that again: a miracle is *a physical occurrence for which we can offer no physical explanation*). It is just over 100 years since James Clerk Maxwell gave us a good working description of what happens – the equivalent of saying that if you lie in hot sunshine you will get sunburned – but he did not *explain* the radiation phenomenon; and nobody has explained it since.

Here, then, is a fine example of modern technology in action. We know how to build a radio transmitter and we can calculate very accurately what will happen when we switch it on. Something will travel from transmitter to receiver at the speed of light, and we shall be able to detect its arrival and make whatever use of it we please for our convenience and entertainment. But except that it may consist of physical energy, or at least that it may carry physical energy with it, *we have no idea* what it is that does the travelling.

Confronted with this true statement of our human ignorance, ninety-nine people out of every hundred will probably say they do not care. The radio is for listening to, not wondering about; wondering about such things is a job for scientists. But now we come to the crunch, for I have to make a similar report to you about the attitudes of the scientists themselves. Nine out of every ten physicists today would also say they didn't care – they are far too busy to be bothered with such abstract, impractical matters. On the other hand, the one physicist in ten who does care about such things is likely to be seriously worried.

If one were to identify and question this minority, their concensus view would almost certainly be that vast gaps exist in our knowledge of physical phenomena that take place not only in complex laboratories and remote galaxies, but also "right on our doorstep" – of which domestic radio radiation and sunlight are commonplace examples. From a purist point of view it is a pity that our progress in understanding such things should have come to a grinding halt in about 1920. (The fundamental basis for atomic energy was laid by Einstein in 1907, and that for the laser in 1917.) Of the new concepts which have arisen in physics since that time very few, if any, have dealt credibly with fundamental matters. I include in this category the major speculative adventure of the 1930s, which failed amid general confusion and is one of the main topics to be examined here.

There would seem to be little doubt that progress in fundamental physics, as opposed to technology, has not kept pace with contemporary progress in other branches of science during the past fifty years or so. It should have done, in view of the number of physicists at work all over the world, but it hasn't. Every now and then, it is true, some new hypothesis seems locally promising and is hailed as a triumph; but when one seeks to apply it elsewhere it does not fit, and it leads one sooner or later to a logical impasse. Nowadays, for reasons that we will explore in due course, we no longer reject a failed hypothesis as we should, but instead we tend to retain it on the pragmatic basis that it may prove more useful to have wrong concepts than no concepts at all. From that point it is very easy to forget that they *are* wrong concepts – scientifically disproved – and instead to go on building upon them as if they were true and valid: an elementary mistake, surely, but one which we go on making.

There are countless examples of this trouble in modern physics, so that it is the rule rather than the exception. The cumulative effect of such errors has been confusion on a majestic scale. We are left with a tangle of separate, uncoordinated, and very often mutually-exclusive concepts. "Sometimes light behaves as waves, sometimes as particles", it is said, yet the concepts of electromagnetic light-waves and particles (photons) are mutually exclusive. Our picture of the physical world has become less clear, rather than more clear, with the passing years. This, I submit, is evidence of a lack of progress. In the 1980s we have to admit that we have not yet found answers to some simple but important questions which were asked as long ago as 1920, and even earlier.

Now when you have been searching diligently for something for fifty or sixty years and failed to find it, it may be sensible to pause and consider whether there might not be some reason for the failure. In our present case two possibilities are more likely than others: either the thing we are looking for doesn't exist, so that we are mistaken in looking for it, or we are looking for it with the wrong kind of spectacles. Let us examine these two possibilities in turn.

There is a doctrine of modern physics, whose origins we will identify later and criticise, which says that scientific theories are limited in their application to providing descriptions of physical events, and are intrinsically incapable (in an absolute sense) of explaining them. According to this doctrine, questions of the nature "what happens?" may give rise to descriptive answers – in numerical detail, of course – and are legitimate questions, whereas questions of the type "how?" or "why?" cannot be answered by science and are therefore *improper* questions which should not be asked.

To take an example, experiments show convincingly that all negative electrons are identical in their behaviour – "indistinguishable" in the approved jargon – and that short of its complete annihilation the physical properties of an electron never vary in any way; one never comes across bigger or smaller electrons, or parts of an electron. Now: to the question "Why is the structure of an electron so phenomenally stable?", current doctrine returns the answer that the mass of the electron is so small that its structure must be quantum-indeterminate, which means that the question of its mechanical stability does not arise. That question is a non-question, an irrelevance that does not *require* an answer.

For convenience of reference I propose to call this the Doctrine of Haziness: "Microphysical entities are hazy, and one should not ask old-fashioned questions about them". Personally I am very suspicious indeed of this doctrine. It seems to be just a little too flexible in its application to be intellectually honest. For instance, in another example,

Question: Why are the wavelengths of the spectrum lines from a gas in a discharge tube so precisely defined?
Answer: Because the permitted energies that electrons can assume within the atoms are precisely quantized.
Question: Oh – I thought it was the electron's angular momentum that was quantized?
Answer: That is also true. Both energy *and* angular momentum are precisely quantized.
Question: If that is so, then the position of an atomic electron must be precisely determined. How far is it from the nucleus?
Answer: We cannot tell you that, because of the Uncertainty Principle of Professor Heisenberg. We can only tell you where you are most *likely* to find it.
Question: So its energy and momentum are in fact *not* precisely determined?
Answer: That is so; they may take on any values within Heisenberg's limits.
Question: Then why are the spectral wavelengths, which you now say are dependent on indeterminate energy and momentum, themselves precisely defined?
Answer: Your questions pre-suppose that the atom has a mechanical structure. Our modern theory is a *mathematical* theory, not a mechanical theory. Hence the questions you ask are meaningless.
Question: But I thought you said the mathematical theory dealt with energy and angular momentum. Are these not ordinary mechanical quantities?
Answer: You are wasting my time. It is a matter of statistics. Look up the theory in any textbook.

You will have noticed the testiness of tone which arises characteristically at that point in the discussion. We shall look into that little "matter of statistics" and form conclusions about it which are not entirely conventional. As I said earlier, the doctrine of haziness seems a shade too convenient to be true. It enables its adherents to wriggle out of logical impasses by sheltering in mysticism, a particular mysticism which as we shall see is linked directly to an unexpected and, as I shall assert, erroneous and quite unjustified denial of the Law of Causation. These are deep waters which can bear being looked into. The doctrine of haziness also offers comfort to the lazy physicist (or shall we say, the too-busy physicist?). Current theories suggest that Nature may be stranger than our forbears thought, for human understanding. If so, we should not be surprised that we have made so little progress recently. (I need hardly emphasize that if this defeatist attitude should become held generally – and it seems to be gaining ground – it must spell the end of the philosphical road for physical science.)

The other possible explanation for our failure to achieve that steadily-improving understanding of the working of the physical world which human instinct (*and* previous experience in physics, *and* current experience in other disciplines) suggests we ought to be achieving, is that there is something there to see but that we have been looking for it with the wrong spectacles. We cannot see radio waves or electrons with the naked eye, of course, but we infer their existence from the readings of our instruments. Our "electron spectacles" are not the instruments we use, but *the scientific theories* with and against which we interpret our observations. A current theory is an expression of a contemporary attitude of mind.

We can be, and historically often have been badly misled by our theories. To take a classically familiar example, in times past the motion of the planets across the night sky could be *described* to any desired degree of accuracy on the basis of the Earth being the dynamic centre of the universe. It could be *explained* – that is, accounted for rationally with a minimum of underlying assumption – much more readily by means of a Sun-centred theory. From experience we have come to believe that the more closely a scientific theory reflects the mechanism of the physical world, the simpler will its concepts appear and the wider will be its field of application. In this example, planetary astronomy had been bogged down for a thousand years under the geocentric theory, and progress had virtually stopped. Further advance depended on the rejection or overthrow of the geocentric theory and its replacement by the alternative which is still in use today. And what an advance that proved to be! One of its earliest consequences was Newton's law of universal gravitation.

We may perhaps read that experience across into the area of fundamental physics where our recent progress seems to have been surprisingly, and disappointingly, slow. Slow progress does not prove that anything is wrong with our current theories and doctrines, but it raises that possibility. It is possible that some of our fundamental thinking may have been on the wrong lines (and by wrong lines I mean lines which do not accord with those of physical Nature). If so, then much of the elaborate, self-generating and *untested* structure of mathematico-physical theory that has been built up during the past fifty years may turn out in the end to have been irrelevant, if not actually misleading. I am suggesting that the time is now ripe for a critical review of modern physical theory, much of which has not been of a type to inspire confidence.

There was for many years a powerful body of opinion which in the teeth of all the evidence for the heliocentric theory maintained that the Earth, as the abode of Man, *must be* the centre of the physical universe. To such opinion no factual proof was convincing: one can neither prove nor disprove an Article of Faith. Thus the ancient polarisation between churchman and scientist tended to continue. Yet it is a feature of modern physics, unexpected but explainable, that in its philosophy it is more akin to a religion than to a classical science. Mysticism has returned in a big way. It seems that in the fundamentals area we are dealing with matters of faith and doctrine, dogma and heresy, so that formal experimental proofs are no more to be expected in fundamental physics nowadays than in a theology. There may even be resentment against anyone who presumes to question the One True Faith; but this time the conservative Establishment is likely to be found within the ranks of science itself.

The significance of that remark will become clear when I declare my main thesis, which is that physical science made a sequence of errors during the 1930s from which it has never recovered. I am in good company in this, since that view was to a greater or lesser extent shared from the early days of Quantum Theory by Einstein, Planck, von Laue, and Schrödinger, all of whom were central in the original arguments. Theirs was a "realistic" view, which in the climate of the times did not prevail against the novel, mystical doctrines of Bohr, Heisenberg, Dirac, and others. The last-mentioned became established and remain formally accepted today. But attitudes may now be changing after fifty years: at any rate I hope so. I propose to identify some of the errors in the 1930's doctrines, show that they were indeed errors, and show how they came about. To my physicist colleagues I say, If your faith is not strong enough to withstand such criticism you should read no further, for I have no wish to cause you offence. To the layman I say, Here for you entertainment is a real-life, up-to-date version of Hans Andersen's famous story of the King's New Clothes.

To sum this up: every scientific theory is somebody's particular pet. Rather than attack the established theories of physics – which would force their doting owners to rush to their defence, and lead to quite unnecessary altercations – I propose to examine a selection of miracles. A miracle, you will remember, is a physical occurrence for which we can offer no physical explanation. There are plenty of miracles to choose from, so we can afford to be selective. We shall find that our miracles have a certain hallmark about them, from which we can deduce not understanding, perhaps, but clues towards understanding. The nature of current theories will become clearer, so that we shall discover when it is safe – philosophically safe – to use these theories, and when dangerous. When fully developed this technique should enable us

Continued on page 87

ERROR CORRECTING SOFTWARE

Software solution avoids complexity of convolutional coding; program computes error signal which would have been the output of a convolution decoder.

by N. Darwood

In digital equipment a single parity bit is often added to a word to increase reliability by detecting an erroneous read. For words which are recorded in blocks as on paper-tape, floppy discs, cassette or magnetic tape we can time-share the single parity bit over a number of words. Fig. 1 shows the idea. Here the parity bit is formed from the data bits of the current word and one data bit per channel staggered to produce a staircase parity-checking area as shown.

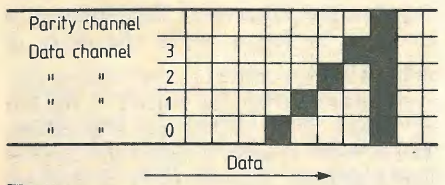


Fig. 1

Using a technique described in *Wireless World*, "Improved parity checker" Jan 1981, p. 81/2, multiple errors can be detected, although only a single parity bit is employed. Fig. 2 illustrates how eight errors would be detected.

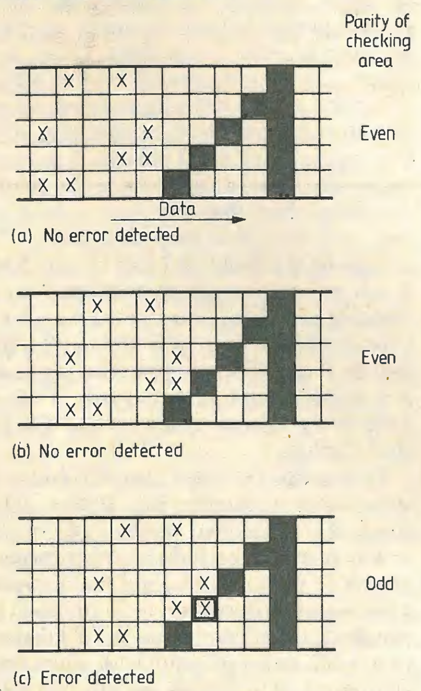


Fig. 2

If the parity bit at the transmitter was coded for say even parity then on reception any single error in the checking area will result in an odd parity for this area of bits. This fact signals an error has occurred. Single error correction with regard to the error signatures produced will be as shown in Fig. 3.

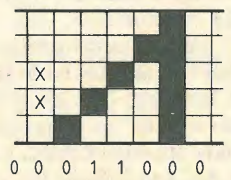
However Fig. 4 shows a double error which although detected gives an error

Parity channel	0	0	0	0	0	0	1	0	0
Data channel 3	0	0	0	0	0	1	1	0	0
" " 2	0	0	0	0	1	0	1	0	0
" " 1	0	0	0	1	0	0	1	0	0
" " 0	0	0	1	0	0	0	1	0	0

Fig. 3

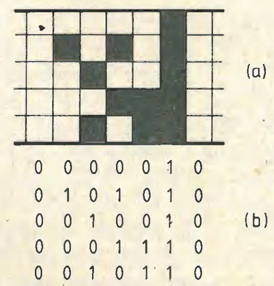
signature which corresponds to the single error signature of channel 0 in Fig. 3.

Fig. 4



If error correction is required then this problem can be overcome still employing only one redundant bit, i.e. the parity bit, by applying more complexity to the time sharing of the data bits. An example of a more complex template is given in Fig. 5(a) together with the single error signatures. This template was formed almost at random – any multiple errors are detected and it appears single errors give unique signatures. Armed with this fact we can detect multiple errors and correct single errors.

Fig. 5



The error detecting software described is useful to

- assess the merits of a particular template
- as a program in its own right used in place of hardware
- to test (2) or for the following reason.

The procedure described here is a form of convolution code. This can be seen by converting the parallel characters to a serial bit stream using a serial shift register. This is given in Fig. 6 for the template of Fig. 1.

Now the convolution approach is far more complicated than the technique described here, but the theory of convolution coding does show that the defining polynomial of Fig. 1 is

$$X^1+X^7+X^{13}+X^{19}+X^{25}+X^{26}+X^{27}+X^{28}+X^{29}+X^{30}.$$

Hence the software approach described can be useful in investigating convoluting codes where here the template, i.e. the polynomial, can easily be changed.

The prompts given by the program and the inputs required in response are shown in Fig. 7 while Fig. 8 shows the notation used for the template data.

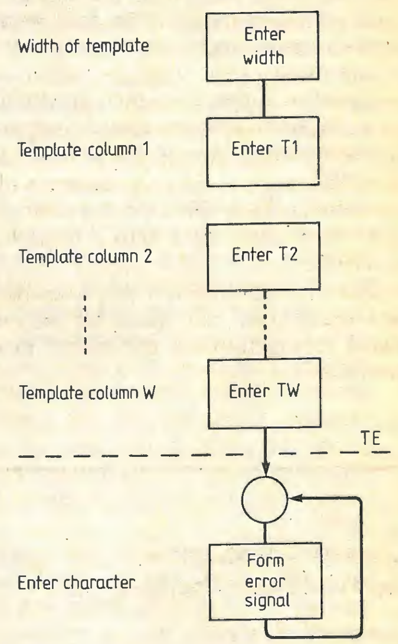
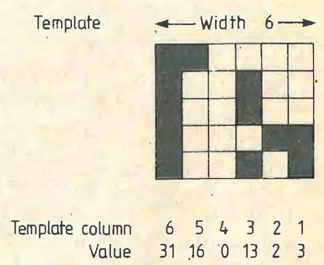


Fig. 7. *Prompts given by the program and inputs required in response.*

Fig. 8



A column is indicated by the decimal equivalent of the binary code where a 1 indicates that the bit is part of the template, that is, an input to the parity-generating hardware.

Having completed entering the template data a continuous series of decimal values are entered as shown in Fig. 9. This series is in response to the prompt *N* where *N* is the *N*th error pattern in the series. An error pattern is simply a character with or without errors, Fig. 9.

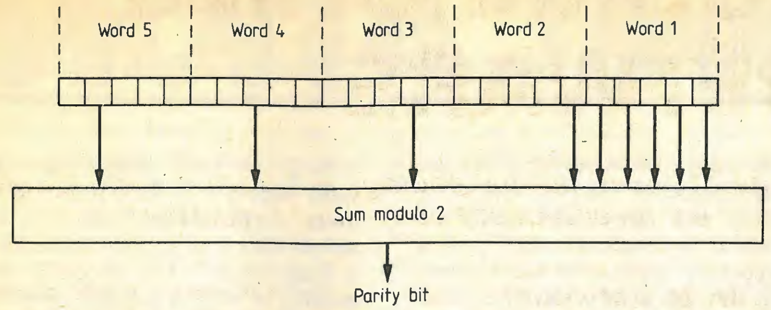
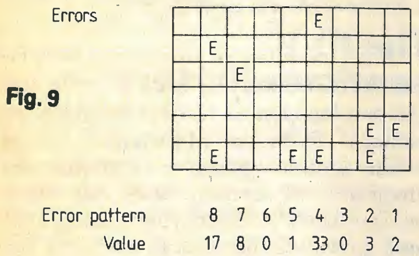


Fig. 6. *Software approach described is much simpler than convolutional approach to Fig. 1 template in which parallel characters are converted to a serial bit stream by shift register.*

Fig. 9



For each error pattern input the program computes the error signal which would have been the hardware output of the convolution decoder, that is it computes the parity of the error pattern within the template area. The output is a one if an error is detected and a zero if no error is detected.

The program assumes that even-parity was transmitted and thereafter an error signal results from an odd parity in the template.

Error detecting program

```
10 DIM T(15), P(15)
20 FOR I=1 TO 15
30 LET T(I)=0
33 LET P(I)=0
35 NEXT I

38 PRINT "TEMPLATE WIDTH IS";
40 INPUT W

50 PRINT "INPUT TEMPLATE CO-
   LUMN VALUES 1TO" W
60 FOR I=1 TO W
70 PRINT I
80 INPUT
90 NEXT I

92 PRINT "ENTER CHARACTERS"
100 LET N=1

104 PRINT N
107 INPUT P(W+1)
108 LET N=N+1
110 IF P(W+1) <0 THEN 990

120 FOR I=1 TO W
130 LET P(I)=P(I+1)
140 NEXT I
150 LET E=0
160 FOR I=1 TO W
170 LET A=T(I) (AND) P(I)
175 LET E=A (EOR) E
180 NEXT I

187 LET A=0
190 FOR I=1 TO 15
192 LET A=E (EOR) A
194 LET E=E(SHFT,-1)
200 NEXT I

210 PRINT A (AND)#1
220 GOTO 104
990 END
```

A block diagram of the program would show that for each error pattern input value, Fig. 9, the previous W values, Fig. 8, are right-shifted one place and the last one drops off the end to retain W values. The corresponding columns in the sense of Figs 8 & 9 are first ANDed with the template. This results in one word. The bit pattern of this one word is parity checked, i.e. the bits counted modulo 2, down to one bit by the exclusive-OR instructions. This bit indicates an even number of bits in the checking area if zero, and an odd number of bits in the checking area (the template) if one, which indicates an error. To change the template the program is rerun. To stop the program a negative integer is entered as data. □

Digital frequency synthesiser design

Continued from page 64

Integrated circuit two-modulus and four-modulus v.h.f. and u.h.f. counters have been available for some time, as have general purpose c.m.o.s. and t.t.l. programmable counters. However, dedicated l.s.i. circuits which can interface directly with two and four-modulus prescalers have only recently been introduced. The diagram in Fig. 4 shows a two-modulus prescaler with i/o ports and a control line which is used to alter the division ratio. The input may be balanced or unbalanced, but the output and control normally operate at c.m.o.s. or t.t.l. levels. A similar four-modulus prescaler which uses two control lines instead of one is shown in Fig. 5.

At frequencies over 1GHz, separate prescalers or mixing techniques must be used. The SP8619 divide by 4 prescaler will operate up to 1.8 GHz and can drive an SP8906 four-modulus divider. Therefore, synthesizers using this combination will have a channel spacing of four times the reference frequency. The highest output frequency available from an SP8906 or SP8901 is 512/239 MHz, i.e. 2.142 MHz. An n.m.o.s. synthesizer circuit, the NJ8811, which has been designed for use with the 8906 and 8901 is shown in Fig. 6. This device contains the programmable dividers, reference divider, phase comparator and a data buffer which enables it to be used with a 4-bit data bus for programming by a r.o.m. and channel switch or a microprocessor system.

Using the NJ8811 and SP8906, in the type of design shown in Fig. 5, provides a synthesizer which operates from 40 to 512 MHz. The programmable reference divider allows the choice of sixteen channel spacings, via two program control lines, from a single standard 4.8 MHz reference input. If the NJ8811 is used with a SP8901, channel spacings and frequency ranges are doubled. The NJ8812 shown in Fig. 7 is almost identical to the 8811, but is designed for use with the SP8793 prescaler in low-power synthesizers at frequencies up to 200 MHz. This combination has an average power consumption of around 55mW and is therefore suitable for portable equipment.

Page 64 is a Basic program for designing v.h.f. and u.h.f. synthesizers using the i.cs described earlier. The first part of the program requests frequency limits, channel spacing, injection mode and other details of the required synthesizer. The circuits to be used and interconnections are then determined together with the reference frequency and reference division ratio for minimum reference harmonics in the working band. The second half of the program requests details of the individual channels required and provides r.o.m. programming information for them. The program was written for a PET 2001 computer, although it will run with other Basic computers with minor dialect modifications. The compact form of program and lack of REM statements is due to memory limitations of the computer involved, but in its present form it works well and is a useful aid for designing frequency synthesizers in the v.h.f. and u.h.f. range. □

COMMUNICATIONS

Radio-frequency or data processing?

Modern radio communications, it is often stressed, represent the marriage of data processing with traditional radio-frequency technology. But sometimes the problem is to determine which is the dominant partner. The r.f. engineers often feel that they are in danger of being smothered by the embrace of the systems-approach of the computer people. In an analogue world, careful alignment and circuit optimization tend to be admired more than the "go/no-go" attitude of the digital designers. Digital technology has advanced spectacularly, tending to obscure the steady improvements in crystal filters, mixers of great dynamic range and the opening up of a whole new world by low-noise microwave amplifiers.

Glancing through the technical-papers programme for Communications 1982 (still at the time of writing in the future though already over when these notes appear) one notes – a little regretfully – that relatively few of the 80 or so papers cover the more traditional r.f. subjects of receivers, transmitters, aerials and r.f. propagation. The themes are mostly allied to what was once called "telecommunications" with networks, switching, multiplexing, electronic (telephone) exchanges and fibre-optics technology prominent among them.

Nevertheless, work continues in the r.f. field. New techniques for land-mobile (admittedly including digitized speech), radio pagers with monolithic v.h.f. receivers (STL work on direct-conversion f.m. techniques combined with Plessey work on the chips); more progress on the s.s.b. polar-loop transmitters developed at the University of Bath and showing one effective way in which v.m.o.s.-type devices can provide good linearity at v.h.f.); a University of Leeds paper on envelope-detectable s.s.b. (also with the land-mobile service in view); the intensive work on developing gallium-arsenide integrated circuits for microwave receivers (in this case by Plessey, Research though intensive work is also going on elsewhere).

The Communications 82 programme may not promise many entirely new developments, but does reflect continued progress over a broad spectrum. The computer/communications marriage is clearly well past the honeymoon period and seems to have settled down to quiet domesticity. But r.f. must be careful not to be dominated by the intrusive digits.

F.e.t. or bipolar r.f. power?

Recent years have seen the appearance of practical r.f. mosfet power devices with the basio V-groove or alternatively "hex" or "T" vertical channel structures where the rather difficult V-groove is replaced by a gate within a straight service. This has given the designer of m.f., h.f. and v.h.f. transmitters new scope, and much has been made of the freedom of such devices from secondary breakdown and thermal runaway, although there is still a vital need to avoid even momentary over-voltages that could cause gate punchthrough. The older bipolar devices can often (but not always) provide greater linearity at v.h.f. (this can be overcome by such techniques as the polar-loop arrangement and for some stringent applications including amplitude-compandored s.s.b. the powerfet may prove superior). It is easier to bias a bipolar device for Class B operation. Input impedance of the mosfet is more constant with varying drive levels, and intermodulation distortion can be lower. A larger die is required for comparable power with the mosfet. Each form of device has advantages and disadvantages and there is seldom a clear-cut, no-choice situation. Motorola however has recently shown an interesting form of Class D (switching) amplifier based on power mosfets that can provide 1 kW output at up to 10 MHz with an efficiency of 85 per cent and a power gain of 30 dB.

Thought-provoking antennas

Most of the books published by the amateur radio organizations are stronger on surveying accepted theory than on taking a searching and thought-provoking look at ideas that have been accepted by amateurs (and professionals) over a number of years. Once an idea has been widely accepted it tends to survive without further critical examination. But a new book (which deserves to be read by all concerned with h.f. communications and h.f. broadcast reception) is a notable exception to this rule.

"HF antennas for all locations" is by L. A. Moxon, G6XN – a name familiar to readers of *Wireless World* – who has a highly professional communications background and combines theory with practice to a rare degree, with specialist knowledge of aerial theory and h.f. radio propagation (one of the first persons in the U.K. to recognise the important implications of Albrecht's work on chordal hop). But he also possesses to an unusual degree an uncompromising ability to spot and explain why "sure-fire" designs so frequently fail to live up to the designer's expectations while at the same time often prepared himself to introduce novel ideas.

The current scene – professional as well as amateur – includes, he considers, "serious errors by some of the experts" leading to "consequent waste of much time, money and effort in the pursuit of better results by methods which either have no chance of success or lead to undesirable compromises of one sort or another; in addition it has ensured that virtually all such designs have been in some degree, or in some respect, sub-optimum". Experimental findings, he stresses, should always be subject to the test "does it make sense?". His advice is, however, positive as well as negative, with recommendations firmly based on theory *and* practice. Equally remarkable is to find in a 1982 publication 260 well-illustrated pages in hard covers selling at £5.

The III-V semiconductors

The development of 12 GHz receivers for domestic reception of satellite broadcasting is clearly going to emphasise the importance of current work on III-V semiconductor technology, now being ever more intensively pursued to overcome the limitations of silicon technology, particularly at microwave frequencies. Relatively unfamiliar terms such as "mesfet" (metal semiconductor field effect transistor) and "digfet" (diamond gate structure field effect transistor) are emerging from the laboratories with GaAs mesfets already available for u.h.f. television receivers (3SK97 etc).

The "Annual Review 1981" of Plessey's Allen Clark Research Centre at Caswell shows clearly the many applications foreseen in the communications field for both discrete and integrated circuit devices based on semi-conductors using materials made from combinations of Group III elements of the periodic table (boron, aluminium, gallium and arsenide) and Group V elements (nitrogen, phosphorus, arsenic and antimony). Indium phosphide InP, for example, is increasingly yielding millimetric active devices including advanced field-effect transistors. Gallium arsenide is being used for power f.e.t.s and for microwave integrated circuits. Apart from Plessey, Texas Instruments, Hewlett-Packard and Hughes have all reported progress in this field.

AMATEUR RADIO

dBW carrier power

Although many of the technical anomalies and actual errors of the now notorious February 12 licence schedule have been corrected – and the whole question of imposing "equivalent isotropic radiated power" limitations above 1 GHz has been deferred pending further consideration, the revised schedule still retains the controversial "dBW carrier power supplied to the antenna" definition. It replaces the traditional "d.c. input power", i.e. total direct current power inputy to (i) the anode circuit of the valve (s) or (ii) any other device energizing the antenna, as used for some 60 years as the basis of the licence for modes other than s.s.b.

Admittedly, 20 dBW output power is roughly equivalent to 150 W d.c. input. So it may seem only a quibble to worry about this change of definition. Presumably it has been made as part of a general move on the part of the licensing authority to standardize all licences issued under the Wireless Telegraphy Acts. But there are surely valid objections to the change and it seems a pity that the R.S.G.B. seems to have conceded the point.

In the first place the syllabus for the Radio Amateur's Examination has never included any requirement for candidates to understand or use decibel notation; but even more to the point is that relatively few amateurs have test equipment capable of measuring acurately the carrier power supplied to the aerial (and is this the power supplied to the feeder or to the actual radiating element – often two very different things?). Nor is there any real reason why amateurs should be forced into acquiring such a measurement capability. A d.c. input limitation encourages high-efficiency amplifiers and sets a clear, and easily measured, limit to the power.

The delays and problems in respect of the new "schedule" were reflected in the time it took to issue new licences to those who passed the R.A.E. last year. The situation in the U.K. compares unfavourably to those in many other countries. In Sweden, for example, where the multiple choice technique is also used, it is possible to take the examination on any working day throughout the year and to be told whether you have passed or failed at the time. The Swedish Morse examinations (no test for Technician licence for v.h.f.; 8 w.p.m. for 10-watt Class C novice c.w. licence; 12 w.p.m. for 75-watt Class B licence; 16 w.p.m. for 500 W Class A licence) are machine generated and the candidates sending is recorded. Sweden is even prepared to issue "guest" licences to foreign amateurs on merit without insisting on an official reciprocal licence agreement. One result is that there are some 8500 licences in a population of about 7.5-million, and Swedish amateurs are a pleasure to contact, reflecting the sensible regulations.

Cable tv problems

In North America leakage of amateur signals into and out of wideband cable television systems is becoming a major problem to radio amateurs. It has even been termed "a new strain of the radio-frequency-interference virus" in a strongly worded *QST* editorial by Richard Palm, K1CE. He points out that while, on paper, cable systems are non-broadcast facilities close to the outside environment this is often far from being the case in practice. But it does mean that the frequencies used for distributing television signals along relatively-leaky co-axial cables are chosen by the industry itself on economic grounds alone. The result is that the distribution frequencies are more and more often within the v.h.f. bands allocated internationally and by the F.C.C. to amateur radio. Instead of being true "closed-circuit" the systems leak signals out into highly-used bands, including 144MHz, in violation of F.C.C standards and so ruin weak-signal reception by amateurs, but are also themselves frequently susceptible to strong local transmissions from amateur stations operating fully in accordance with their licences. When viewers complain to the cable companies, A.R.R.L. suggest, "many companies promulgate the myth that the amateur is at fault." Poorly shielded components and installation, poor maintenance of the cable systems, the choice of frequencies with amateur bands, are all adding to the problem, while the current cut-backs in the American government-funding of F.C.C. mean that the regulatory organization is now in the throes of a financial crisis and "willing but unable to enforce its rules" on the cable companies.

An h.f. convention

V.h.f. conventions in the UK have a long and extremely successful record stretching back over many years, attracting attendences of over 1000 enthusiasts. But attempts to organise equivalent conventions for h.f. enthusiasts have a more chequered history with some having to be cancelled for lack of support. However this has not deterred the R.S.G.B.'s h.f. committee from organising an ambitious one-day event at the Belfry Hotel and Conference Centre, Milton Common, some 9 miles along the M40 (Exit 7) from Oxford, on Saturday, June 19. The programme includes a trade exhibition, lectures, forum, films, a special display and demonstration of low-power (QRP) operation with the co-operation of the G-QRP-Club using the call-sign GB2HF. A talk on h.f. aerials, including those for 10, 18 and 24MHz, is to be given by Louis Varney, G5RV while the writer of these notes is trying to work out what to say about h.f communications receivers.

Hazard of PCBs

The recent disclosure that there was a large spillage of polychlorinated biphenyls (PCBs) in 1981 due to bomb damage in Northern Ireland to a large power transformer shows once again that there is still a lot of this highly-dangerous substance around. For many years PCBs were widely used as a coolant in oil-filled transformers and capacitors, including some used with fluorescent lamps. Manufacture did not stop in the U.K. until 1977, following the discovery of its dangerous effects.

Any oil-filled component that shows signs of leakage should be treated with care. A useful test recently suggested by Brian Castle, G4DYF is to take a piece of copper wire. Put it in a gas flame and burn off the dirt until the flame is clear. Allow the wire to cool, then dip it in the suspect oil and return it to the flame. If it now burns yellow, it is ordinary oil. If it burns bright green, PCBs are *probably* present. Note that this is not a positive test but it does provide a useful guide as to whether the oil is harmless.

Here and there

A new 28MHz beacon is expected in operation shortly on Gough Island in the South Atlantic. With callsign ZD9GI it will operate on 28.2125MHz.

A balloon carrying a 146/432MHz translator plus beacon transmitters and expected to reach heights of 15 to 20km is due to be flown several times this year in South Africa. After reaching its maximum height the balloon is expected to burst and the equipment come down on parachute at distances up to about 250km from the launching point. As it comes down a 144MHz transmitter will be activated to allow the package to be tracked by d/f and recovered.

Membership of the British Amateur Television Club is expected to reach 1400 this year. The club notes that the BBC microcomputer is well suited to the display of slow-scan television pictures on domestic tv sets. The club has recently published a special issue of *CQ-TV* (no 117) largely devoted to amateur television equipment for the 24cm band. It also proposes a plan for this amateur band including a main tv repeater channel (output 1242.25MHz vision, input 1274.25 MHz vision) an alternative repeater channel (output 1250.25MHz, input 1282.25MHz) and a simplex amateur television channel (1258.25 MHz). In each case the sound channel would be 6MHz higher.

Forthcoming mobile rallies: June 20 Denby Dale at Shelley High School, Skelmanthorpe, near Huddersfield. June 27 Longleat mobile rally organized as the Bristol RSGB group's 25th event; Rolls Royce Sports & Social Club, Barnoldswick (6 miles south of Skipton). July 11 Worcester rally at High School, Ombersley Road, Droitwich. July 18 Pembroke's "Bucket and Spade Party", The Regency Hall, Saundersfoot; Sussex rally at Brighton Racecourse; and Cornish rally at Technical College, Cambourne.

PAT HAWKER, G3VA

NRZ RECORDING FOR SMALL COMPUTERS

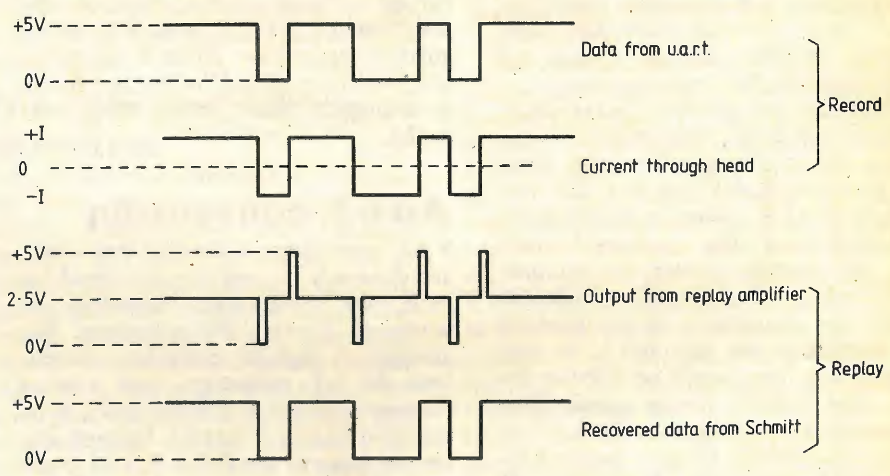
The majority of small computer recording systems use the Kansas City cassette recording format, with a data rate limited to a few hundred baud. L. Hayward proposes a non-return to zero recording system for the Nascom 1 and 2 – the circuit should be adaptable to others – and compares performance with that of the Kansas City interface.

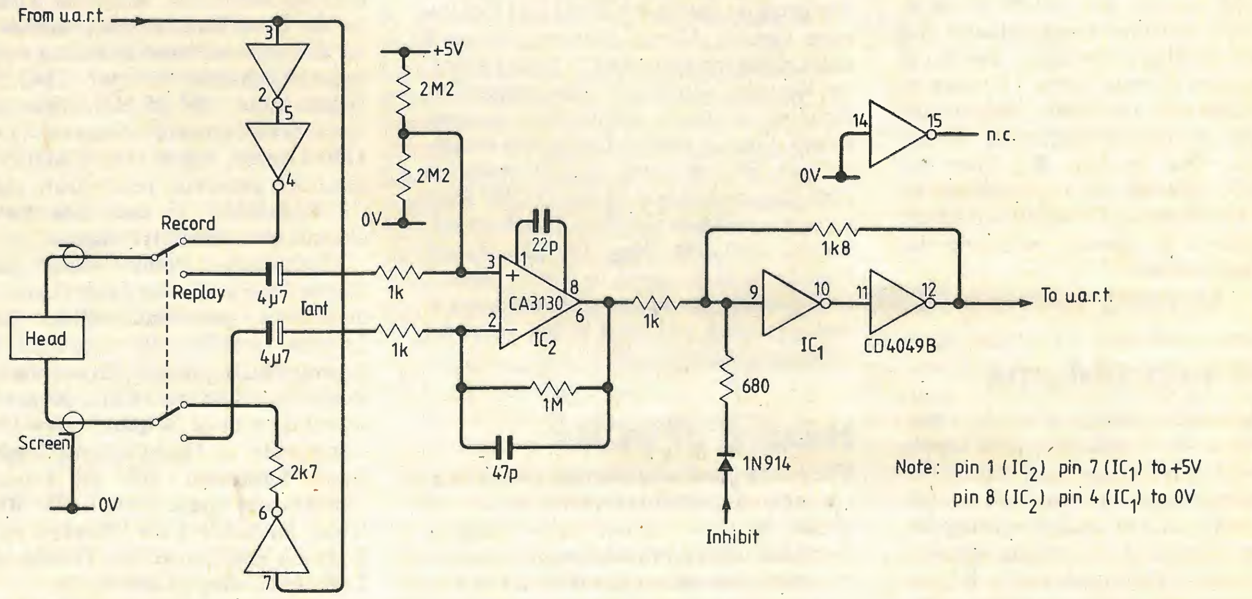
By L. Hayward

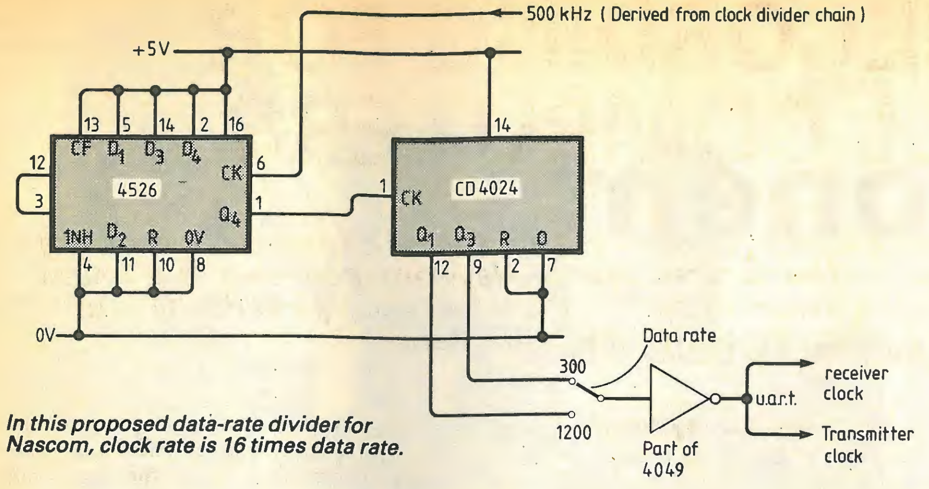
Most small computer systems, in particular those machines offered to the amateur user, have adopted the audio cassette as a convenient method of data storage. Cassette players are readily available at a low price, the only reasonable alternatives being open-reel recording or expensive disc drives. The Kansas City recording standard, developed to use the audio cassette, works well and has become popular due to its tolerance of tape-speed variation, typically 30%. This allows users to exchange recordings between machines of almost any type. But this speed tolerance is the only significant advantage of the system – its disadvantages are susceptibility to tape dropout and slow data speed. Some systems optimistically offer a data rate of 1200 baud, but using cheap cassette decks the best that can normally be attained is 300 baud. The encoding and decoding circuitry involved is fairly complex, using as many as seven or eight large-scale integrated circuits, and enables the user to adapt either the existing audio cassette recorder, or use a "bare-bones" deck, with mechanism and record-playback head only.

Using one of the worst cassette decks I have encountered reliable recording was achieved to a rate of 1200 baud, with fair operation to 2400 baud 1⅞ in/s. The only disadvantage of this is its in tolerance of speed variation: 5% instead of the 30% offered by Kansas City, assuming that the usual uart (universal asychronous receiver-transmitter) is used in the computer. This speed restriction will normally only be a disadvantage if tape transfers from one recorder to another are to be made. Any recorder having a cyclic speed variation of greater than 0.5% would be totally useless for musical reproduction, and most cassette recorder mechanisms can achieve better speed regulation.

The n.r.z. system is well known, and has been used in computing systems for years. No h.f. head bias is used, and the tape is magnetically saturated in a negative or positive sense, depending on whether a zero or a one is written. There is no condition of zero flux, hence the name: non-return to zero. As the tape is saturated no erase head is required, and the system is less sensitive to tape drop-out or variation between various types of tape. Ideally, the system should use heads and tape designed specifically for this type of operation; practical results have shown however, that ordinary heads and tape are quite suitable. The use of certified digital cassettes such as the Scotch type 834A is recommended for the most reliable operation. The circuit shown was specifically designed for use with Nascom 1 and 2 computers, but should be suitable for most other computer systems with little change.







In this proposed data-rate divider for Nascom, clock rate is 16 times data rate.

Principles of operation

Record mode. The data output from the u.a.r.t. is applied to a 4049 buffer. Sections of this buffer are connected so that the current through the head and current-limiting resistor is reversed in direction as the input changes from logic high or low. The choice of resistor shown is suitable for a typical cassette recorder audio head. It may be reduced if more current is required, up to a maximum of 10 mA for other heads.

Replay mode. A 3130 op-amp is used as a differential amplifier. This mode of operation permits considerable hum to be picked up in the head connecting wires without interference. The gain of 1000 was found to be sufficient to cause the output to clip when replaying data. The input coupling capacitors were made much larger than necessary, considering the frequencies involved, to permit the head to short-out the amplifier input at low frequencies and prevent hum pick-up by capacitive stray coupling. The high frequency response of the amplifier is rolled off to avoid possible pick-up from the nearby clock generator and dividers in the computer.

As the voltage output from the head is proportional to the rate of change of flux the amplifier output will consist of narrow pulses coincident with the timing and direction of the data. In between these pulses, the amplifier output falls to 2.5V. A Schmitt trigger circuit using part of the 4049 is used to hold the state of the previous positive or negative excursion, and thus output restored data to the u.a.r.t. Hysteresis is used to make the output insensitive to spurious small outputs from IC2. The u.a.r.t. requires that the receiver input terminal remains high until the data transmission begins. An inhibit input is provided, which when high prevents IC2 from changing the Schmitt trigger output. This point is conveniently connected to the drive l.e.d. transistor collector in the Nascom, thus making the computer ignore all data until the 'c Load' or 'R' command is executed. The suggested divider circuit is useful if the standard data rates of 300 and 1200 baud are required from the Nascom 1. Power supply required is a single +5V supply; current drain is so small that an existing computer supply should easily accommodate it.

I suggest that circuits such as this be included in small computer systems as an alternative or addition to Kansas City. It shouldn't be too difficult for manufacturers of ready-built systems to offer a completed cassette system as part of the package. If such devices are made available with accurate speed control, thus giving interchangeability, it is likely that the more logical n.r.z. will be adopted universally.

It should be fairly easy to produce a machine with a speed correct to within 5% for reasonable cost. A normal diesel engine with a crude mechanical governor can meet 5% regulation of speed, so why not a simple cassette drive? □

Heretics guide to modern physics

Continued from page 81

to judge the *physical credibility* of any new hypothesis, providing us with a critical faculty which in recent times has been woefully lacking.

The first miracle we shall examine will be the one I mentioned at the opening, namely the mechanism of the transmission of light energy through empty space. Our first philosophical milestone will be consequential and closely related to it: an understanding of the true function of "waves" in modern physics. We shall have to go back some 200 years in scientific history to find a suitable starting point. Our route will take us from Newton to Heisenberg: via electromagnetic theory and the acute distress it suffered when denied an aether; via practicable photons, quantization, non-existent matter-waves, and a *restricted* Principle of Indeterminacy; and ultimately to an affirmation that the Law of Causation is obeyed in physics not only statistically but in all circumstances. In each of these areas I will present ideas for your consideration which although far removed from conventional scientific doctrine are yet strictly in accord with the findings of experiment. These ideas will add up eventually to a self-consistent whole, but not yet, I regret, to a fully-developed Theory.

All that I have to say is very simple, and indeed I hope to show how simple Nature really is when the dust of man-made confusion has been swept away. William of Occam said that fundamental assumptions should not be multiplied unnecessarily, and I am a follower of William of Occam. □

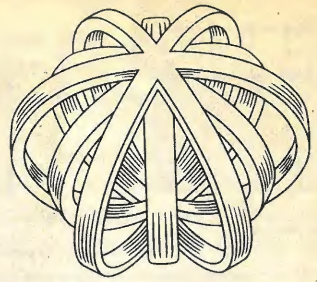
Summary

The phenomenon of the radiation of light and radio energy is a "miracle" – a well-established physical occurence for which science can offer no physical eplanation. Modern technology uses these radiations and others every day without understanding them. Progress toward understanding such things effectively came to a halt in about 1920, after which fundamental concepts in physics began to become confused and mutually contradictory. This lack of progress may have been due to one of two possible factors: either Nature is too mysterious for us to understand, so that it is not worth the bother of trying, or our fundamental thinking may have taken a wrong turning 50 years ago. There are historical precedents both for such errors and for conservative pressures against correcting them. Nevertheless, enough material now exists to warrant a major re-think, based on a return to the earlier philosophy of realism in physical science which reflects the underlying simplicity of Nature.

C.b. frequency synthesis

In Fig. 4 of the article on 40 channel c.b. frequency synthesis, which appeared in the November 1981 issue of *Wireless World*, there should be a 1nF capacitor in the line between the bottom end of L_1 and the MV2110 variable-capacitance diode. Without this capacitor the a.f.c. is inoperative.

Paris components show

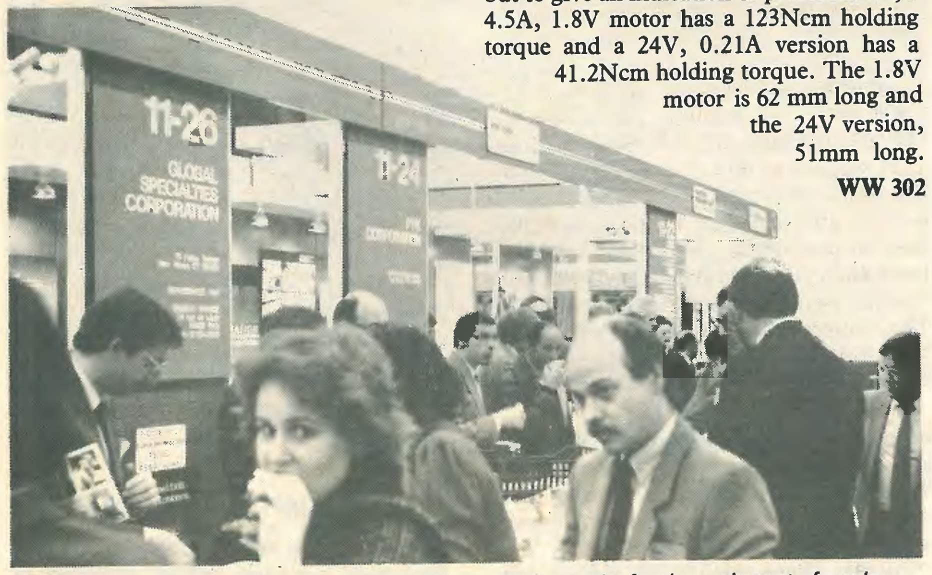


from Martin Eccles in Paris

France's foremost electronics exhibition – Salon International des Composants Electroniques – this year attracted over 1700 exhibitors representing 31 different countries. Held for the last time at Paris's Parc des Expositions, the 25th annual Paris Components Show, despite slight increases in the number of visitors from outside France and the total number of exhibitors, saw a fall in attendance. According to the French Trade Exhibitions office in London, there were just over 85700 visitors, as opposed to last year, when 95124 permanent passes were issued. But considering current economic restraints, the figures are still quite impressive.

In 1983, the show is to be held in November, instead of early April as has been the tradition, at the North Paris Exhibition Grounds, and after that become biennial to alternate with the Munich exhibitions. A more specialized exhibition will be held at the North Paris site each even year.

Surprisingly, perhaps, the current 'world's fastest' **16K c.m.o.s. static r.a.m.** has been developed in France and is, initially at least, to be manufactured there. Access time of the HM65161 2K by 8-bit memory is 55ns (maximum) and its power consumption is 500µW in standby mode or 30mW while enabled. This device is the outcome of joint efforts by Harris and the exhibitors, Matra Harris, who were also showing the HD6409, a c.m.o.s. Manchester II encoder/decoder for full duplex operation up to 1Mbit/s, the MA1200 30 MHz gate array with 1200 gates each with a propagation-delay time of between 2 and 4ns, and the HM8048 microcontroller together with its c.m.o.s. counterpart.
WW 301



By far the majority of contemporary exhibition stands consist for the main part of nooks, often formed by zig-zagged partitions, the most effective of which are erected diagonally across the podium. Often, traders without visitors will step away from their stand waiting for an unsuspecting fly to be attracted into one of these embryonic niches and, when fairly certain of their potential prey's species, move in from behind and weave their web. This show is no exception, but as it is so large, one can still find many of the more modest exhibitors with stands where one can browse without being pounced upon.

Oriental Motors are contemplating distributing their products in the UK, possibly by the end of the year. When asked why their motors might be bought in the UK in preference to similar products from other manufacturers, the representative replied predictably that they would be competitively priced. The company's **stepping motors,** part of a large range of induction, reversible, synchronous, geared and fan motors, have a step angle of 1.8° and are designed for a −10 to +50°C operating-temperature range. Direct-voltage ratings of the 19 motors range from 1.8 to 24V and current-per-phase ratings range from 0.2 to 4.5A. Maximum and minimum voltage/current ratings don't always coincide, but to give an indication of performance, a 4.5A, 1.8V motor has a 123Ncm holding torque and a 24V, 0.21A version has a 41.2Ncm holding torque. The 1.8V motor is 62 mm long and the 24V version, 51mm long.
WW 302

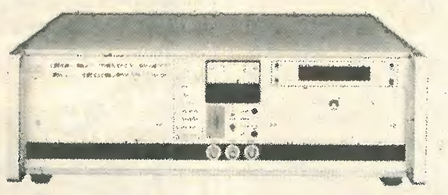
A **computer-aided design system** shown by EIE, a Swiss company specializing in the manufacture and distribution of printed-circuit board design equipment since 1974, has its limitations. Firstly, boards greater than 232dm^2 and containing more than 360 i.cs cannot be designed on it. Further, only 15 colours from a choice of 4096 may be used, restricting the system's use to designing boards with no more than 15 layers. But as 25ft^2 boards and printed circuits with more than 15 layers are hard to find, further comment is justified. The computer can resolve layout tolerances down to 0.0025mm and, when combined with the company's drum photo-plotter, be used to produce art work accurate to within ±0.01mm. Dimensions may be displayed on the screen in two ways, either from the board's pole to the cursor or between two points on the 'drawing', and commonly used component forms may be selected from a permanent memory. To the operator, System 81 looks like a large desk with a keyboard, swivelling v.d.u. and joystick. A 32-bit, bit-slice processor coordinates the graphics, supplemented by 1.5M byte of semiconductor memory, an LSI11 for handling i/o and arranging data files and two 8in disc drives. Options other than the drum photo-plotter already mentioned include 20M byte hard-disc drives, a printer and GPIB pen plotter interface – so one can imagine the further limitations.
WW 303

Still on the same subject, but at the other end of the price scale, Colvern had a microcomputer with colour graphics adapted for computer-aided design on their stand. The company isn't moving into this area though – they only manufacture the 3-axis controller used. Bitstik-Apple graphics, as the system is called, is a product of Robocom and can be supplied as a basic conversion kit for existing Apples, with or without colour facilities, or alternatively as a complete system with various options for hard-copy, etc. The kit, priced at around £187, comprises the previously mentioned controller (which is actually a joystick, but not to be confused with a games paddle) design-aid software on disc and a manual. As this is a **general-purpose design aid** intended for compiling anything from artistic to architectural drawings, component 'library' software for printed-circuit board design is to be available as an option. One feature of this

system is that a single picture element may be zoomed in on to fill the whole screen, i.e., a single master page may be broken down into 16000 pages. This means that information, such as an op-amp's parameters, may be stored in a small area which is invisible on the overall view.
WW 304

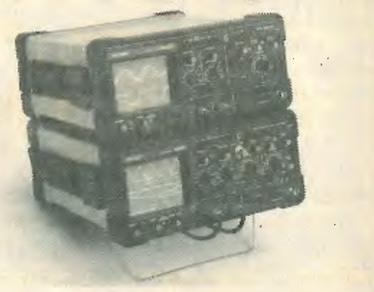
Numerous photographs of disc-drive heads initially attracted us to this company's stand. On glancing at the brochures there, we found that Paris's Samson Data, and their Belgian counterpart, Samson Computer Supplies, represent Information Magnetics Corporation, an American company offering a hard-disc head refurbishing service. They also supply numerous **professional computer and computer-related appliances,** such as disc packs, read/write heads, magnetic tapes, alignment discs and anti-glare filters for v.d.us. Occasionally at such exhibitions we meet people who hear the name *Wireless World* for the first time and respond cautiously, and probably in their eyes tactfully, with "I don't think we have anything that will be of interest to your readers." To explain the evolution and current scope of the magazine is time consuming so we usually take one or two issues along to keep initiation discussions brief. A genial M. Samson hadn't heard of us and on seeing our only two issues decided to keep them, despite our insisting that we wanted them back. We thought at first that he had simply become accustomed to people handing out free magazines but after a brief rapport realized that he was serious. Finally, M. Samson won, after fruitlessly offering a cheque or cash and the promise of a subscription in exchange for the two magazines, by tendering a slightly imperfect disc-drive carriage fitted with 13 heads – an offer we couldn't refuse.
WW 305

Caesium frequency standards are not uncommon nowadays. Besides their more obvious uses in national time services, metrological applications and power stations, they are also used for a number of scientific and industrial purposes and in data-transmission systems, satellite ground stations and observatories. In outline, the International Telegraph and Telephone Consultative Committee (CCITT) recommends that international digital links should be synchronized with a frequency error of less than 1×10^{-11}: the only practical way of achieving this accuracy is by using a caesium frequency standard. Oscilloquartz was showing part of its range of caesium standards and systems alongside its more conventional quartz crystals, oscillator units and frequency references. The 3000 is an uncased caesium oscillator with frequency and long-term stability errors of $\pm 7 \times 10^{-12}$ and $\pm 3 \times 10^{-12}$ respectively, intended as a module for use by



equipment manufacturers. It gives a 1V r.m.s. sine-wave output at 5MHz. Model 3120 incorporates the 300 oscillator and gives sine-waves of 1, 5 and 10MHz with the same accuracy. This cased instrument is fitted with a 6-digit clock, control and monitoring facilities, output buffers, batteries and p.s.u. It measures 131 by 428 by 546mm and is suitable for rack mounting. The company can also provide complete systems for all the applications mentioned above and a number of accessories are available for the standards mentioned.
WW 306

Of course, the more familiar faces were also at the exhibition. Philips were exhibiting quite a number of recent and new products, among them a dual-clock analyser/35MHz oscilloscope. The PM3543, **10MHz logic-state analyser** has disassembly facilities for 16-bit microprocessors, including the Z8001/2 and 8086, and for 8-bit devices from a number of manufacturers. Because of the dual clock, multiplexed-bus data and address lines may be separated so up to 16 data bits and 21 address bits may be displayed, together with status information, using only 24 inputs. The analyser section has a 255-word memory, and threshold levels may be set for t.t.l. or varied between −3 and +12V. Once logic has been analysed, the instru-



ment can be used as a 35MHz oscilloscope to aid fault location. An IEEE-bus interface is available as an option. Among other new products shown by Philips were two 75MHz lightweight service oscilloscopes, one with a single timebase and the other dual, and an audio-distortion meter for measurements to DIN standards.
WW 307

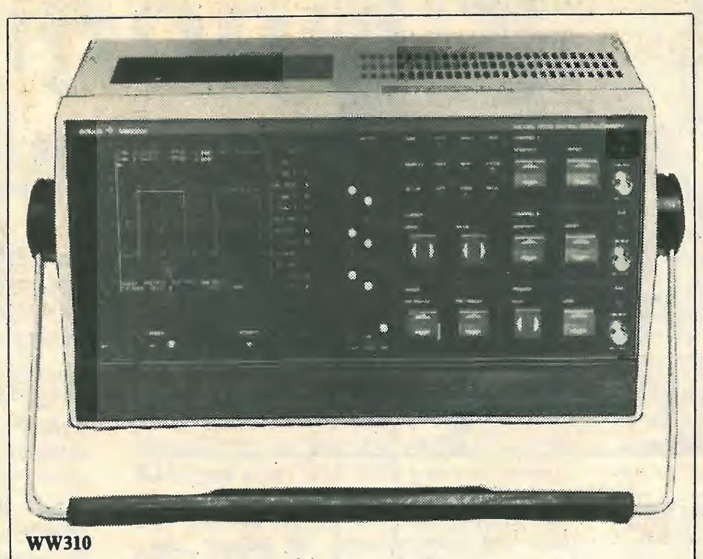
Two types of **liquid-crystal display** with **129mm-high characters,** one with seven segments and the other with a 6-by-7 dot matrix, have been produced by Fairchild's optoelectronics division. Measuring 165 by 110mm, these displays require typically 5V r.m.s. at 5µA with all segments on and take either 45 and 130ms or 80 and 150ms to turn on and off respectively, depending on the type-number suffix. Seven-segment types are disignated LTR1341 and dot-matrix types LTR1401.
WW 308

In addition to a number of new semiconductors, RCA were illustrating the advantages of their **colour-enhanced development system** for 1802 c.m.o.s. microprocessor products. "Colour", say the company, "not only enhances the display, but also simplifies and speeds up screen editing." Fairly obvious, we thought, but that doesn't detract from the usefulness of the system. Floating-point Basic is held in part of a 30K r.o.m., as is assembling, editing and monitoring software. A further 5K r.a.m. is included, leaving 29K free memory area for expansion, and permanent storage is possible using one of two cassettes. Any CDP18S600 series Microboard may be used with the system. The semiconductors mentioned earlier are firstly; two 4MHz microprocessors, similar in architecture to the 1802 but with 113 instructions as opposed to the 02's 91, one with 64 bytes of r.a.m., the 1805, and one without, the 1806. Secondly, an 8-bit flash a-to-d converter for sampling speeds up to 15MHz with 150mW power consumption, and a range of N-channel power m.o.s.f.e.ts, dubbed 'low cost', with current ratings from 1 to 18A and voltage ratings from 80 to 450V were mentioned. The converter i.c. is the CA3308, and the power f.e.ts are RCA91XX and 92XX devices.
WW 309

NEW PRODUCTS

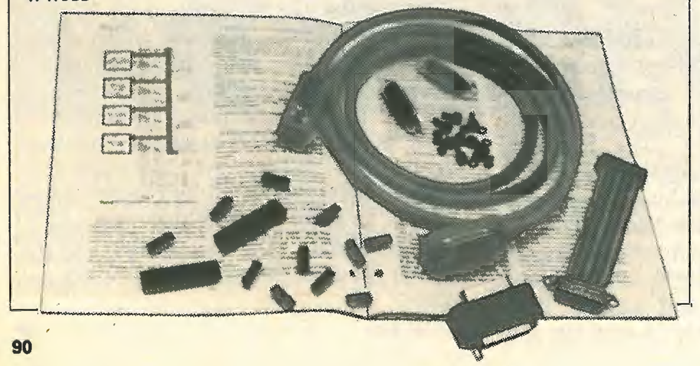
DIGITAL STORAGE OSCILLOSCOPE

Two 100MHz a-to-d converters, one for each 35MHz bandwidth channel, are used in Gould's 4500 storage oscilloscope. Shown for the first time in the UK at the All Electronics Show, this instrument can be used to store and display 'single-shot' or repetitive waveforms and is suitable for bench use, when it will be operated by means of its front panel controls, or as part of a test system under GPIB control. It can resolve 5.1 bits at 35MHz, introduces a maximum absolute voltage error of ±6.6% f.s. over the recorded range and, after 40ns, responds to transients with a relative error of ±0.4% Setting up of the front panel is aided by software-generated menus displayed on the screen; once a setting has been made, it may be stored for later use or comparison. With these menus, the operator can select control functions for signal averaging, cursor positioning, trigger source and filtering options and plotter digital interface operation. Mathematical comparisons of reference and acquired waveforms are possible. For waveform comparisons, the 4500 has a 4Kbyte reference memory and for acquired waveforms, a 1Kbyte per channel (or 2Kbyte in single-channel mode) memory. A floppy disc will be available for storing up to 30 waveforms for later use either with the oscilloscope or with an external computer/controller. The 4500's price is around £11,500, and it can be obtained from Gould Instruments Ltd, Roebuck Rd, Hainault, Ilford, Essex IG6 3UE.
WW310


WW310

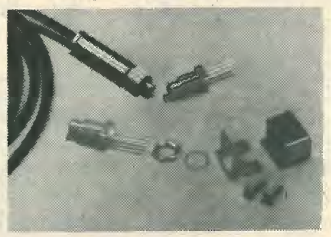
GPIB COMPONENTS SERVICE

Accessories and components specifically for general-purpose interface buses are available from one source. With prior knowledge of the problems involved in designing GPIB's, Wasec have formed a service offering a range of bus transceivers, control i.cs, connectors, studs, cables, adapters, specifications and other data pertaining to these buses for enthusiasts, manufacturers and designers alike. A leaflet describing the service is available on request. Wasec, 45 Hurstcourt Rd, Sutton, Surrey, SM1 3JF.
WW311


WW311

SMALL FIBRE-OPTIC LINK

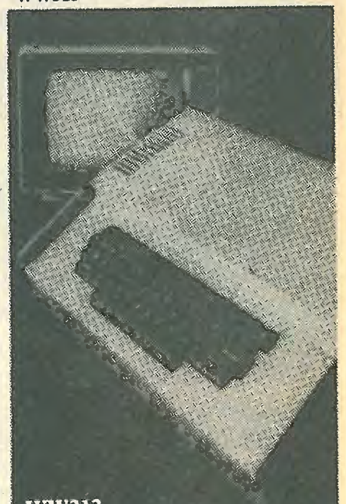
This 0 to 5Mbaud fibre-optic link with very small emitter and detector elements can be used to transmit data over distances of up to 1200m. Transmitter and receiver



units of Hewlett-Packard's HFBR-0200 measure 7.8mm high, including shielding and connectors, and function at temperatures within the range −20 to +85°C. The receiver i.c. has an optically transparent, yet electrically conductive, shield to improve its immunity to electromagnetic interference and electrostatic discharge; the receiver also has a logic-compatible output. Suspended sapphire-ball lenses optically connect emitter and detector i.cs with the cable, without mechanical contact. These lenses also make connector alignment problems less critical. A kit, including transmitter, receiver, mounting hardware and 10 metres of cable costs £104.60 if less than ten are bought. Hewlett-Packard Ltd, King Street Lane, Winnersh, Wokingham, Berks RG11 5AR.
WW312

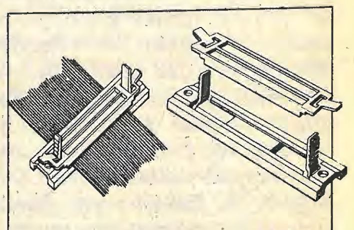
MICROCOMPUTER FOR EDUCATION AND CONTROL

Some 12Kbytes of Basic tailored for control applications is a feature that Midwich Computer Co. hope will bring their MC system into educational establishments and laboratories. If a separate board, called an experimental unit, is connected to the Z80-based computer, combinations of an analogue-input board, analogue-output board and digital-input/output board may be used with the system to perform many complex control applications simply. The computer has 16Kbytes of r.a.m., a 300 or 1200 baud RS232 interface and a 300 or 1200 baud cassette-recorder interface with motor switch. Binary, octal, decimal and hexadecimal numbers can be handled by the interpreter, which also has facilities for simplifying communications between Basic and machine-code routines, and real-time interrupt handling and nested scheduling facilities. A machine-code monitor is included. Prices are, £375 for the Microcontroller computer, £145 for the experimental unit, £55 and £35 respectively for analogue input and output boards and £55 for a digital i/o board. Discounts on the computer are available for educational establishments. Midwich Computer Company Limited, Hewitt House, Northgate Street, Bury St Edmunds, Suffolk IP33 1HQ.
WW313


WW313

FLAT-CABLE CLAMP

A two-part plastic assembly for clamping flat cables is available from Winslow. The Ribbon Wrap can hold up to eight, 0.89mm-thick flat cables and is obtainable in three widths, for 26, 40 or 64-core ribbons. Rivets, bolts, screws or self adhesive pads may be used to secure the base-plate section. Two ratchet-type lugs on the base plate hold the clamping section in place. Winslow International, 71 Tunnel Road, Tunbridge Wells, Kent TN1 2BX.
WW314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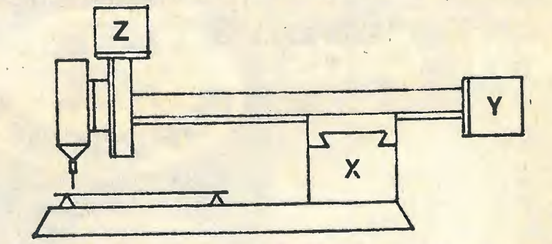

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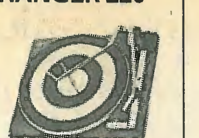


Plays 12", 10" or 7" records, Auto or Manual. A high quality unit backed by BSR reliability. Stereo Ceramic Cartridge. AC 200/250V. Size 13½×11¼in. 3 speeds. Above motor board 3¾in. Below motor board 2½in.
Post £2 Cut Board **£1** extra

HEAVY METAL PLINTHS Post £2

Cut out for most BSR or Garrard decks. Silver grey finish, black trim. Size 16x13¾in. **£4**

DECCA TEAK VENEERED PLINTH. Post £1.50 Superior finish with space and panel for small amplifier. Board is cut for B.S.R. 18¾in.×14¼in×4in. Black/chrome facia trim. Also with boards cut out for Garrard £3. **Plastic cover £6** **£5**

TINTED PLASTIC COVERS Post £2

17⅞×13⅛×3¼in.	£6	18¼×12½×3in.	£6
17¼×9⅜×3½in.	£3	14⅜×12½×2⅞in.	£5
13¾×12×2¼in.	£5	16⅝×13×4in.	£6
15¼×13½×4in.	£6	14½×13⅛×2¾in.	£5
17×12⅞×3½in.	£6	17¼×13¾×4⅛in.	£6

Callers Only (not suitable for post)

21½×14¼×2½in.	£6	21×13⅞×4⅛in.	£6
23¾×14×3⅞in.	£6	30¾×13⅜×3¼in.	£6

BSR SINGLE PLAYER DECKS

BSR P232 BELT DRIVE QUALITY DECK
Manual or automatic play. Precision ultra slim arm. Cueing device. Bargain price With stereo ceramic cartridge **£24** Post £2

BSR P204 SINGLE PLAYERS SPECIAL OFFERS
Two speed 33/45 r.p.m. hi-fi decks with stereo cartridges, cueing device and snake arm. Ceramic – 240V AC **£15** or 9V DC **£19** Magnetic – 240V AC **£20** or 12V DC **£24**

GARRARD 6-200 SINGLE PLAYER DECK £22 Post £2
Brushed Aluminium Arm with stereo ceramic cartridge and Diamond Stylus, 3-speeds. Manual and Auto Stop/Start. Large Metal Turntable. Cueing Device. Ready cut mounting board £1 extra.
GARRARD 730SP. 240V. **£27.50.** Magnetic Cartridge Snake arm, 3-speed single player, p.p. £2

BATTERY ELIMINATOR MAINS to 9 VOLT D.C.
Stabilised output, 9 volt 400 m.a. U.K. made in plastic case with screw terminals. Safety overload cut out. Size 5 x 3¼ x 2½in. Transformer Rectifier Unit. Suitable Radios, Cassettes, models, **£4.50.** Post 65p.

DE LUXE SWITCHED MODEL STABILISED. £7.50. Post £1. 3-6-7½-9 volt 400ma DC max. Universal output plug and lead. Pilot light, mains switch, polarity switch.

DRILL SPEED CONTROLLER/LIGHT DIMMER KIT. Easy to build kit. Controls up to 480 watts AC mains, **£3.** Post 65p.
DE LUXE MODEL READY-BUILT 800 watts. Front plate fits standard box, £5. Post 65p.

EMI 13½x8in. LOUDSPEAKERS

Model 450, 10 watss R.M.S. with moving coil tweeter and two-way crossover; 3 ohm or 8 ohm.
£9.50 post £1.50. "Final Clearance".
SUITABLE BOOKSHELF CABINET
£6.50. Size 18×11×6in. Post £1.50.

RELAYS. 6V DC **95p.** 12V DC **£1.25.** 18V **£1.25.**
BLANK ALUMINIUM CHASSIS. 6×4–**£1.45**; 8×6–**£1.80**; 10×7–**£2.30**; 12×8–**£2.60**; 14×9–**£3**; 16×6–**£2.90**; 16×10–**£3.20.** All 2½in. deep. 18 swg.
ANGLE ALI. 6×¾×¾in. 18 swg. **30p.**
ALUMINIUM PANELS, 18swg. 6×4–**45p**; 8×6–**75p**; 14×3–**75p**; 10×7–**95p**; 12×8–**£1.10**; 12×5–**75p**; 16×6–**£1.10**; 14×9–**£1.45**; 12×12–**£1.50**; 16×10–**£1.75.**
PLASTIC AND ALI BOXES IN STOCK. MANY SIZES
ALUMINIUM BOXES. 4×4×1½ **£1.** 4×2½×2 **£1.** 3×2×1 **£1.** 6×4×2 **£1.60.** 7×5×3 **£2.40.** 8×6×3 **£2.50.** 10×7×3 **£3.** 12×5×3 **£2.75.** 12×8×3 **£3.60.** All with lids.
BRIDGE RECTIFIER 200V PIV 2a **£1.** 4a **£1.50.** 8a **£2.50.**
TOGGLE SWITCHES SP **30p.** DPST **40p.** DPDT **50p.**
RESISTORS. 10Ω to 10M. ¼W, ½W, 1W, **2p**: 2W **10p.**
HIGH STABILITY. ½w 2% 10 ohms to 1 meg. **10p.**
Ditto 5%. Preferred values, 10 ohms to 10 meg, **3p.**
WIRE-WOUND RESISTORS 5 watt, 10 watt, 15 watt **20p.**
PICK-UP CARTRIDGES SONOTONE 9TA **£2.50.** 9TAHC **£3.80.**
BSR Stereo Ceramic SC7 Medium Output **£2.** SC12 **£3.**
PHILIPS PLUG-IN HEAD. Stereo Ceramic. AU1020 (G306 - GP310 - GP233 -AG3306 - AG3310) **£2.**
LOCKTITE SEALING KIT DECCA 118. Complete **£1.**
ANTEX SOLDERING IRON 240V 15W 3mm bit **£5.25.**
JACK PLUGS Mono Plastic **25p**; Metal **30p.**
JACK PLUGS Stereo Plastic **30p**; Metal **35p.**
JACK SOCKETS Mono Open **20p**; Closed **25p.**
JACK SOCKETS Stereo Open **25p**; Closed **30p.**
FREE SOCKETS – Cable end **30p.** Metal **45p.**
2.5mm and 3.5mm **JACK SOCKETS 20p. Plugs 20p.**
DIN TYPE CONNECTORS
Sockets 3-pin, 5-pin **10p.** Free Sockets 3-pin, 5-pin **25p.**
Plugs 3-pin **20p**; 5-pin **25p**; Speaker plugs **20p**; Sockets **15p.**
PHONO PLUGS and SOCKETS ea. 15p.
Free Socket for cable end **20p.** Screened Phono Plugs **25p.**
300 ohm TWIN RIBBON FEEDER 10p yd.
U.H.F. COAXIAL CABLE SUPER LOW LOSS, 25p yd.
COAX PLUGS 30p. COAX SOCKETS 20p.
NEON PANEL INDICATORS 250V 30p.
SOLID DIELECTRIC 100pf, 500pf £1.50.

POTENTIOMETERS Carbon Track

5kΩ to 2MΩ. LOG or LIN. L/S **50p.** DP **90p.** Stereo L/S **£1.10.** DP **£1.30.** Edge Pot 5K. SP **45p.**

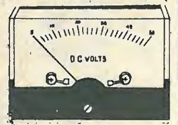
MINI-MULTI TESTER NEW

De luxe pocket size precision moving coil instrument. Impedance + Capacity – 4000 o.p.v. Battery included.
11 instant ranges measure:
DC volts 5.25, 250, 500.
AC volts 10, 50, 500, 1000.
DC amps 0-250μa, 0-250mA.
Continuity and resistance 0 to 600K ohms.

£6.50 Post 65p

De Luxe Range Doubler Model, 50,000 o.p.v. £18.50. 7×5×2in. Post £1

NEW PANEL METERS £4.50



50μa 100μa, 500μa, 1ma, 5ma, 50ma, 100ma, 500ma, 1 amp, 2 amp 25 volt, 50 volt, VU Meter. 2¼x2x1¼in. Post 65p

RCS SOUND TO LIGHT CONTROL KIT

Kit of parts to build a 3 channel sound to light unit. 1,000 watts per channel. Suitable for home or disco. Easy to build. Full instructions supplied. **£15** Post 95p
Cabinet £4.50 extra. Operates from 200MV to 100W.
200 Watt Rear Reflecting White Light Bulbs. Ideal for Disco Lights, Edison Screw. 6 for **£4**, or 12 for **£7.50.** Post 65p. Suitable panel mounting holders **85p.**

RCS "MINOR" 10 watt AMPLIFIER KIT £14

This kit is suitable for record players, guitars, tape playback, electronic instruments or small PA systems. Two versions available: Mono, **£14**; Stereo, **£20.** Specification 10W per channel; size 9½×3×2in. SAE details. Full instructions supplied. 240V AC mains. Post £1.

RCS STEREO PRE-AMP KIT. All parts to build this pre-amp. Inputs for high, medium or low imp per channel, with volume control and PC Board Can be ganged to make multi-way stereo mixers **£2.95** Post 65p

MAINS TRANSFORMERS

		Post
250-0-250V 70mA, 6.5V, 2A	£4.50	£2
250-0-250V 80mA, 6.3V 3.5A, 6.3V 1A	£5.00	£2
350-0-350V 250mA, 6.3V 6A CT	£12.00	£2
300-0-300V 120mA, 2×6.3V 2A C.T.; 5V 2A	£12.00	£2
220V 45mA, 6.3V 2A	£2.50	£1

AUTO 115V to 240V 150W **£9.** 250W **£10.** 400W **£11.** 500W **£12.** £2.

GENERAL PURPOSE LOW VOLTAGE

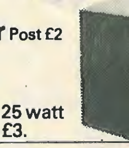
Tapped outputs available

	Price	Post
2 amp. 3, 4, 5, 6, 8, 9, 10, 12, 15, 18, 25 and 30V	£6.00	£2
1 amp. 6, 8, 10, 12, 16, 18, 20, 24, 30, 36, 40, 48, 60	£6.00	£2
2 amp. 6, 8, 10, 12, 16, 18, 20, 24, 30, 36, 40, 48, 60	£10.50	£2
3 amp. 6, 8, 10, 12, 16, 18, 20, 24, 30, 36, 40, 48, 60	£12.50	£2
5 amp. 6, 8, 10, 12, 16, 18, 20, 24, 30, 36, 40, 48, 60	£16.00	£2

5-8-10-16V. ½ amp.	£2.50	80p	15-0-15V. 2 amps	£3.75	£1
6V. ½ amp.	£2.00	£1	20V 1 amp	£3.00	£1
6-0-6V. 1½ amp.	£3.50	£1	20-0-20V 1 amp	£3.50	£1
9V. 250ma.	£1.50	80p	20-40-60V 1 amp	£4.00	£2
9V. 3 amp	£3.50	£1	25-0-25V 2 amps	£4.50	£1
9-0-9V. 50ma	£1.50	80p	28V 1 amp Twice	£5.00	£2
10-0-10V. 2 amps	£3.00	£1	30V 1½ amp	£3.50	£1
10-30-40V. 2 amps	£3.50	£1	30V 5 amp and 17-0-17 2a	£4.50	£2
12V. 100ma	£1.50	80p			
12V. 750 ma	£2.00	80p	35V 2 amps	£4.00	£1
12V 3 amps	£3.50	£1	TOROIDAL 30-0-30V 4a and 20-0-20V ½a £10 £2		
12-0-12V. 2 amps	£3.50	£1			

CHARGER TRANS	Post		RECTIFIERS	Post
6-12 volt 3a	£4.00+£2		6-12 volt 2a	£1.10+80p
6-12 volt 4a	£6.50+£2		6-12 volt 4a	£2.00+80p

OPUS COMPACT SPEAKERS £22 pair Post £2



TEAK VENEERED CABINET
11×8½×7in, 15 watts
50 to 14,000 cps. 4 ohm or 8 ohm
OPUS TWO 15×10½×7¾in 25 watt 2-way system **£39 pair.** Post £3.

LOW VOLTAGE ELECTROLYTICS **10p**
1 mf, 2 mf, 4 mf, 8 mf, 10 mf, 16 mf, 25 mf, 30 mf, 50 mf, 100 mf, 250 mf. All 15 volts. 22 mf/6v/10v; 25 mf/6v/10v; 47 mf/10v; 50 mf/6v; 68 mf/6v/10v/16v/25v; 100 mf/10v; 150 mf/6v/10v; 200 mf/10v/16v; 220 mf/4v/10v/16v; 330 mf/4v/10v; 500 mf/6v; 680 mf/6v/10v/16v; 1000 mf/2.5v/4v/10v; 1500 mf/6v/10v/16v; 2200 mf/6v/10v; 3300 mf/6v; 4700 mf/4v.
500mF 12V **15p**; 25V **20p**; 50V **30p.** 1200mF 76V **86p.**
1000mF 12V **20p**; 25V **35p**; 50V **50p**; 100V **70p.**
2000mF 6V **25p**; 25V **42p**; 40V **60p**; 1200mF 76V **80p.**
2200mF 63V **90p.** 2500mF 50V **70p**; 3000mF 50V **65p**;
4500mF 64V **£2.** 4700mF 63V **£1.20.** 2700mF/76V **£1.**

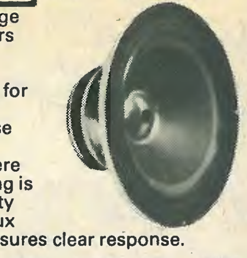
HIGH VOLTAGE ELECTROLYTICS

2/500V	45p	8+8/450V	75p	32+32+16/350V	90p
8/450V	45p	8+8/500V	£1	100+100/275V	65p
16/350V	45p	8+16/450V	75p	150+200/275V	70p
32/500V	75p	32+32/350V	50p	220/450V	95p
32/350V	50p	32+32/500V	£1.80	32+32+32/325V	75p
50/450V	95p	50+50/300V	50p	50+50+50/350V	95p

CAPACITORS WIRE END High Voltage
.001, .002, .003, .005, .01, .02, .03, .05 mfd 400V **5p.**
.1MF 200V **5p.** 400V **10p.** 600V **15p.** 1000V **25p.**
.22MF 350V **12p.** 600V **20p.** 1000V **30p.** 1750V **50p.**
.47MF 1500V **10p.** 400V **20p.** 630V **30p.** 1000V **60p.**
VALVE OUTPUT Transformers (small) **90p.**
TRIMMERS 30pF, 50pF, **10p.** 100pF, 150pF, 500pF **30p.**
MICROSWITCH SINGLE POLE CHANGEOVER 40p.
SUB-MIN MICROSWITCH, 50p, Single pole changeover.
TWIN GANG, 120pF **50p.** 500 plus 200pF **£1.**
GEARED TWIN GANGS 25pF **95p.**
GEARED 365+365+25+25pF **£1.**
TRANSISTOR TWIN GANG. Japanese Replacement **£1**

HEATING ELEMENTS, WAFER THIN
Size 11×9×⅛in. Operating voltage 240V, 250W approx. Suitable for Heating Pads, Food Warmers, Convector Heaters, Propagation, etc. Must be clamped between two sheets of metal or ceramic, etc.
ONLY **60p** EACH (FOUR FOR **£2**) ALL POST PAID.

NEW baker Star sound



high power full range quality loudspeakers produced to give exceptional reproduction. Ideal for Hi-Fi, music P.A. or discotheques. These loudspeakers are recommended where high power handling is required with quality results. The high flux ceramic magnet ensures clear response.

MODEL	INCHES	OHMS	WATTS	TYPE	PRICE	POST
MAJOR	12	4-8-16	30	HI-FI	£14	£2
DELUXE MK II	12	8	15	HI-FI	£14	£2
SUPERB	12	8-16	30	HI-FI	£24	£2
AUDITORIUM	12	8-16	45	HI-FI	£22	£2
AUDITORIUM	15	8-16	60	HI-FI	£34	£2
GROUP 45	12	4-8-16	45	PA	£14	£2
GROUP 75	12	4-8-16	75	PA	£18	£2
GROUP 100	12	8-16	100	Guitar	£24	£2
DISCO 100	12	8-16	100	Disco	£24	£2
GROUP 100	15	8-16	100	Guitar	£32	£2
DISCO 100	15	8-16	100	Disco	£32	£2

BAKER 150 WATT MIXER/POWER AMPLIFIER £89 Post £2

SLAVE VERSION £75

For Organs, Discotheque, Vocal, Public Address. Three loudspeaker outlets for 4, 8 or 16 ohms. Four high gain inputs, each 20 mv, 50K ohm. Individual volume controls "Four channel" mixing. 150 watts into 8 ohms R.M.S. Music Power. Distortion less than 1%. Slave output 500 M.V. 25K.ohm. Frequency Response 25 Hz – 20kHz ± 3dB. Integral Hi-Fi preamp separate Bass & Treble. Compact – 16" x 8" x 5½". Lightweight – 14lb: Master volume control. Made in England. 12 months' guarantee. 200/250v A.C. mains or 120V to order. All transistor and solid state devices. 100 Volt Line £15 extra.
New Stereo Slave Model 150 + 150 watt £125. Post £4.
BAKER'S NEW PA150 MICROPHONE PA AMPLIFIER £129. PP £3
4 channel 8 inputs, dual impedance, 50K-600 ohm 4 channel mixing, volume, treble, bass. Presence controls, Master volume control, echo/send/return socket. Slave input/output sockets.

BAKER £69 Post £2 50 WATT AMPLIFIER

Ideal for PA systems, Discos and Groups. Two inputs, Mixer, Volume, Controls, Master Bass, Treble Gain.

RCS offers MOBILE PA AMPLIFIERS. Outputs 4-8-16 ohms
20-watt RMS 12v DC, AC 240v, 3 inputs. 50K **£46** PP £2.
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Mic 1; Mic 2; Phono; aux. outputs 4 or 8 or 16 and 100v line
60-watt RMS, Mobile 24 volt DC & 240-volt AC mains. inputs 50K. 3 mics + 1 music. Outputs 4-8-16 ohm + 100 volts line **£95** PP £2

FAMOUS LOUDSPEAKERS "SPECIAL PRICES"

MAKE	MODEL	SIZE	WATTS	OHMS	PRICE	POST
SEAS	TWEETER	4in	50	8	£9.50	£1
GOODMANS	TWEETER	3½in	25	8	£4.00	£1
AUDAX	TWEETER	4in	30	8	£6.50	£1
SEAS	MID-RANGE	4in	50	8	£7.50	£1
SEAS	MID-RANGE	5in	80	8	£12.00	£1
SEAS	MID-RANGE	4½in	100	8	£12.50	£1
GOODMANS	HIFAX	7½×4¼	100	4/8/16	£22	£2
GOODMANS	WOOFER	8in	25	4/8	£6.50	£1
GOODMANS	HB	8in	60	8	£12.50	£1
RIGONDA	GENERAL	10in	15	8	£5	£2
AUDAX	WOOFER	10in	50	8	£16.00	£2
GOODMANS	HPG	12in	120	8/15	£29.50	£2
GOODMANS	GR12	12in	90	8/15	£27.50	£2
GOODMANS	HPD	12in	120	8/15	£29.50	£2

SPEAKER COVERING MATERIALS. Samples Large S.A.E.
B.A.F. LOUDSPEAKER CABINET WADDING 18in wide **35p** ft.
CASSETTE MONO REPLAY. Complete working **£12.50**
CASSETTE MECHANISM. 12V Stereo Heads **£5**
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LOUOSPEAKER BARGAINS
3 ohm, 4in, 5in, 7×4in, **£1.50**; 6½in, 8×5in, **£3**; 8in, **£3.50.** 10in, **£5.**
8 ohm, 2⅝in, 3in, 5in, **£1.50**; 6½in, **£3**; 8in, **£4.50**; 12in, **£6.**
15 ohm, 3½in, 5×3in, 6×4in, **£2.50.**
25 ohm, 3in, 5×3in, 7×4in, **£2.50. 120 ohm,** 3¼in dia. **£1.**

MOTOROLA PIEZO ELECTRIC HORN TWEETER, 3⅜in. square **£5**
100 watts. No crossover required. 4-8-16 ohm, 7⅜x3⅛in. **£10.50**

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Suitable for cassettes and all sizes of tape reels. AC mains 200/250V. Hand held size with switch and lead (120 volt to order).
Will also demagnetise small tools.
Head Demagnetiser only £5.

R.C.S. LOW VOLTAGE STABILISED POWER PACK KITS **£3.95.** Post 65p

All parts and instructions with Zener diode printed circuit, mains transformer 240V a.c. Output 6 or 7½ or 9 or 12V d.c. up to 100mA or less. Please state voltage required.

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HM307.4 £138

Y: Bandwidth DC-10MHz (-3dB) – Sensitivity 5mV-20V/cm (±5%)
X: Timebase 0.2s-0.5µs/cm (±5%) – Triggering 2Hz-30MHz (3mm) – Built in component tester – Calibrator – Screen 6 x 7 – 1kV.

HM203 £220

Y: Bandwidth DC-20MHz (-3dB) – Sensitivity 5mV-20V/cm (±3%) – Dual trace
X: Timebase 0.2s-40ns/cm incl.x5 Magn. – Trigger 3Hz-30MHz (4mm) – X-Y operation – Calibrator – Screen 8 x 10cm – 2kV.

HM412 £350

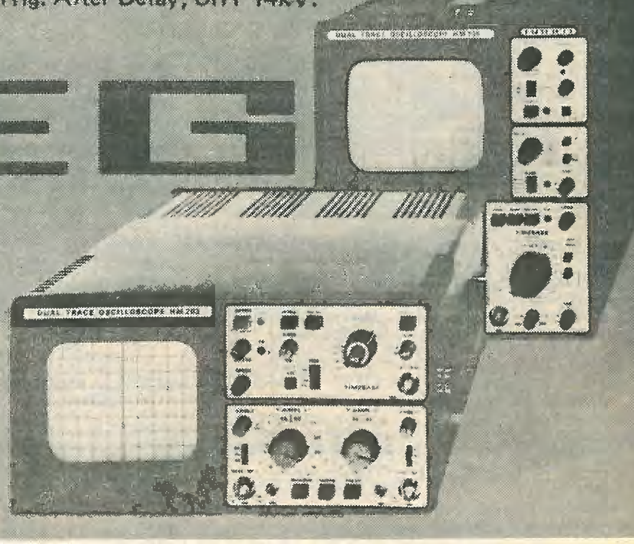
Bandwidth DC-20MHz (-3dB) Sensitivity 2mV/cm - 20V/cm(±3%) Timebase 40ns/cm Triggering DC - 40MHz (5mm) Algebraic Add., Sweep Delay, x5 Mag., Overscan Ind., Var. Holdoff, Single Sweep.

HM705 £580

Bandwidth DC-70MHz (-3dB) Sensitivity 2mV/cm -20V/cm(±3%) Timebase 5ns/cm - 2.5s/cm -Triggering DC - 100MHz (5mm), Algebraic Add , Sweep Delay, x10 Mag.|Alt. Trigger, Trig. After Delay, CRT 14kV.

For free data sheets of the full range contact:

HAMEG



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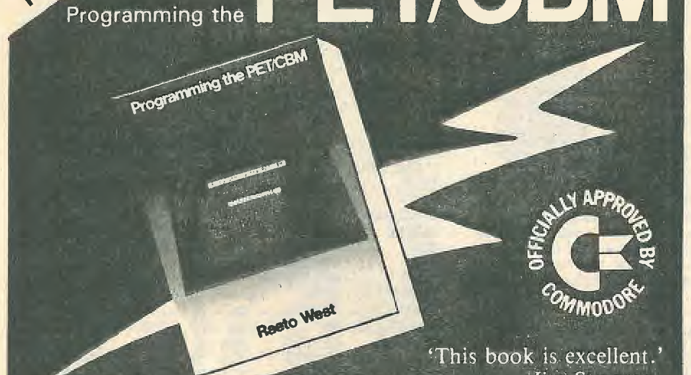
Programming the **PET/CBM**



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THE CINTEC

SINUSOIDAL FREQUENCY AND VOLTAGE STABILIZER

APPLICATIONS

* SOUND RECORDING
* VIDEO RECORDING
* MEDICAL
* MARINE
* COMPUTERS
* NAVIGATIONAL SYSTEMS

Applications for the use of CINTEC FREQUENCY & VOLTAGE STABILIZER are more numerous than can be listed. Therefore, if you have a supply problem, contact CINTEC LIMITED whose engineers will be only too pleased to assist.

★ 500VA OR 250VA
★ SOLID STATE
★ HIGH STABILITY
★ ROBUST
★ VERSATILITY
★ RELIABILITY
★ SINUSOIDAL

Reliable Frequency & Voltage Stabilization

The efficient operation of sophisticated electrical and electronic equipment is, in many cases, dependent upon an electrical supply which is stable in both frequency and voltage.

In many countries and even in the United Kingdom during periods of heavy demand, the variation in the frequency and voltage is sufficient to introduce errors and the malfunction of such items as Recording equipment etc. Likewise, in certain areas, the only source of supply is from a Generator, the output of which can vary considerably when different loads are imposed. This has precluded the use of a wide range of equipment in many countries. Voltage Stabilizers are readily available, but these do not stabilize the frequency of the supply which, in many instances, is essential.

The CINTEC FREQUENCY & VOLTAGE STABILIZER provides the answer to both these problems

When the supply frequency is fluctuating wildly, between 45Hz and 65Hz and the voltage by more than 10% the output from the Stabilizer will not vary more than .01% from 50Hz or 1% in voltage, even when different loads are imposed.

Used by Government establishments, oil rigs, hospitals, police, video and electronic industry, shipbuilders etc, for a wide range of applications including video systems, medical, frequency conversion, navigational aids and sound recording systems.

The CINTEC FREQUENCY & VOLTAGE STABILIZER is also available for supplies of 100-125 volts, 45-65Hz with an alternative output of 50Hz or 60Hz at 115 volts or 230 volts and as a dual frequency model with a switchable output of 50Hz or 60Hz.

The Stabilizer may also be used as a frequency converter. For example, the supply to it can be any frequency between 45-65Hz and the output can be switched to either 50Hz or 60Hz.

SPECIFICATION

INPUT	105-125 volts or 210-250 volts at 45-65Hz.
OUTPUT	115 volts at 230 volts
RATING	500VA or 250VA
STABILITY Voltage	± 1% No load to full load –
Frequency	± 0.01% No load to full load
FREQUENCY	50Hz or 60Hz. Single or dual versions
WAVEFORM	SINUSOIDAL
DISTORTION	Less than 2%
AMB TEMP	−20°C to +40°C
COOLING	Fan cooled
DUTY	Continuous
DIMENSIONS	432 (W) x 196 (H) x 508mm (D) (17″ x 7¼″ x 20″)
WEIGHT	45 or 30Kg unpacked
CONSTRUCTION	Cabinet or rack mounting
TERMINATION	Cannon Connections at rear of case
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24V DC Inverter

In addition to the A.C. operated models, a 24v D.C. INVERTER Stabilizer is available which operates from a heavy duty 24 volt battery and has output ratings similar to the A.C. models. This type of Stabilizer is particularly suitable for mobile operation.

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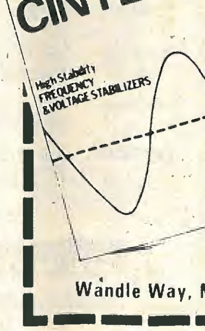

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AD161/2	0.90	BC212	0.09	BD238	0.40	BFX29	0.30	TIP3055	0.66
AF124	0.34	BC212L	0.09	BD410	0.55	BFX84	0.26	TIS91	0.20
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AF126	0.32	BC213L	0.09	BD437	0.50	BFX86	0.30	2N2219	0.28
AF127	0.32	BC237	0.09	BD517	0.60	BFX88	0.25	2N2905	0.40
AF139	0.42	BC238	0.09	BD520	0.65	BFY50	0.21	2N3054	0.59
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AF239	0.42	BC252A	0.15	BF115	0.35	BFY52	0.25	2N3702	0.12
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BC107A	0.10	BC327	0.10	BF177	0.38	BT108	1.22	2N5294	0.38
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BC116A	0.13	BC550	0.08	BF185	0.26	BU208	1.39	2SC1172Y	2.20
BC117	0.19	BC557	0.07	BF194	0.11	BU208A	1.52	2SC1173	1.15
BC119	0.24	BC557B	0.08	BF195	0.11	BU208/02	1.89	2SC1306	1.25
BC125	0.12	BC558	0.08	BF196	0.11	BU326A	1.42	2SC1307	1.85
BC140	0.31	BD115	0.30	BF197	0.11	BU407	1.24	2SC1449	0.80
BC141	0.25	BD116	0.52	BF198	0.10	BU500	1.75	2SC1678	1.25
BC142	0.21	BD131	0.32	BF199	0.14	MJ3000	1.98	2SC1945	2.10
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BT5B	43.50	EC91	7.00	EY86/87	0.50
BT17A	95.00	EC93	0.80	EY88	0.60
BT19	32.50	EC95	7.00	EY91	1.00
BT79	12.50	EC8010	6.00	EY500A	1.50
CIK	14.00	ECC32	3.00	EY802	0.70
C3JA	16.00	ECC33	4.00	EZ35	0.60
C143	18.50	ECC34	4.00	EZ40	2.00
C180	18.50	ECC35	4.00	EZ41	1.25
C1108	55.00	ECC40	2.30	EZ80	0.75
C1111	15.50	ECC80	1.50	EZ81	0.56
C1112	52.15	ECC81	0.55	EZ90	0.96
C1134	15.50	ECC82	0.55	F6059	3.20
C1148A	60.00	ECC83	0.60	F6060	2.50
C1534	32.00	ECC82 Philips	1.10	F6158	2.75
CBL31	2.00	ECC83 Philips	1.10	F6870	6.00
CCA	2.60	ECC84	0.80	FW4/500	2.00
CL30	1.05	ECC85	0.60	G1/371K	30.00
CMG25	9.00	ECC86	1.65	G55/1K	8.00
CMG29	10.85	ECC88	0.65	G120/1B	8.00
CMV25	10.00	ECC88 Mullard	1.65	G150/2D	4.00
CV Nos prices on request		ECC91	0.65	G180/2M	9.00
D3a	35.00	ECC180	0.72	G400/1K	14.00
D63	0.60	ECC189	0.78	G550	11.50
DA107	6.00	ECC230	4.50	GC10B	11.50
DAC32	1.20	ECC801S	2.50	GC10D	11.50
DAF91	0.45	ECC803S	2.50	GC10/4B	11.50
DAF96	0.65	ECC804	0.40	GC10/4E	11.50
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DF91	0.45	ECF812	1.20	GR10G	4.00
DF92	0.60	ECH3	2.50	GR10J	4.00
DF93	1.20	ECH4	3.00	GS10C	12.00
DF96	0.65	ECH35	1.50	GT1C	17.00
DF97	1.50	ECH42	1.00	GT1C S/S	13.50
DH76	1.00	ECH81	0.58	GTE175M	8.00
DH77	0.90	ECH83	0.78	GTR150W	1.00
DH79	0.56	ECH84	0.93	GU50	11.50
DH81	2.00	ECL80	0.66	GXU1	13.50
DH149	2.00	ECL82	0.58	GXU3	24.00
DK91	0.90	ECL83	1.13	GXU50	12.50
DK92	1.20	ECL84	0.74	GY501	1.20
DK96	2.50	ECL85	0.74	GZ30	1.00
DL35	1.00	ECL86	0.74	GZ32	1.00
DL63	1.00	ECLL800	13.50	GZ33	4.50
DL69	1.20	ED510	3.50	GZ34	2.00
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DL94	1.20	EF40	3.00	HBC90	0.75
DL96	1.00	EF41	3.50	HBC91	0.80
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DLS15	11.00	EF50	1.50	HF94	0.60
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DY80	1.20	EF86	0.70	HL92	1.50
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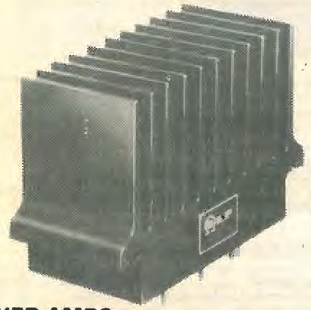

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TYPE	SERIES No.	SECONDARY Volts	RMS Current	PRICE
30 VA 70×30mm 0.45Kg Regulation 18%	1x010	6+6	2.50	**£5.12** +p/p £1.04 +VAT £0.92 TOTAL £7 08
	1x011	9+9	1.66	
	1x012	12+12	1.25	
	1x013	15+15	1.00	
	1x014	18+18	0.83	
	1x015	22+22	0.68	
	1x016	25+25	0.60	
	1x017	30+30	0.50	
50 VA 80×35mm 0.9 Kg Regulation 13%	2x010	6+6	4.16	**£5.70** +p/p £1.30 +VAT £1 05 TOTAL £8 05
	2x011	9+9	2.77	
	2x012	12+12	2.08	
	2x013	15+15	1.66	
	2x014	18+18	1.38	
	2x015	22+22	1.13	
	2x016	25+25	1.00	
	2x017	30+30	0.83	
	2x028	110	0.45	
	2x029	220	0.22	
	2x030	240	0.20	
80 VA 90×30mm 1 Kg Regulation 12%	3x010	6+6	6.64	**£6.08** +p/p £1 67 +VAT £1 16 TOTAL £8.91
	3x011	9+9	4.44	
	3x012	12+12	3.33	
	3x013	15+15	2.66	
	3x014	18+18	2.22	
	3x015	22+22	1.81	
	3x016	25+25	1.60	
	3x017	30+30	1.33	
	3x028	110	0.72	
	3x029	220	0.36	
	3x030	240	0.33	
120 VA 90×40mm 1.2 Kg Regulation 11%	4x010	6+6	10.00	**£6.90** +p/p £1 67 +VAT £1 29 TOTAL £9.86
	4x011	9+9	6.66	
	4x012	12+12	5.00	
	4x013	15+15	4.00	
	4x014	18+18	3.33	
	4x015	22+22	2.72	
	4x016	25+25	2.40	
	4x017	30+30	2.00	
	4x018	35+35	1.71	
	4x028	110	1.09	
	4x029	220	0.54	
	4x030	240	0.50	
160 VA 110×40mm 1.8 Kg Regulation 8%	5x011	9+9	8.89	**£7.91** +p/p £1.67 +VAT £1 44 TOTAL £11 02
	5x012	12+12	6.66	
	5x013	15+15	5.33	
	5x014	18+18	4.44	
	5x015	22+22	3.63	
	5x016	25+25	3.20	
	5x017	30+30	2.66	
	5x018	35+35	2.28	
	5x026	40+40	2.00	
	5x028	110	1.45	
	5x029	220	0.72	
	5x030	240	0.66	

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TYPE	SERIES No.	SECONDARY Volts	RMS Current	PRICE
225 VA 110×45mm 2.2 Kg Regulation 7%	6x012	12+12	9.38	**£9.20** +p/p £2.00 +VAT £1 68 TOTAL £12 88
	6x013	15+15	7.50	
	6x014	18+18	6.25	
	6x015	22+22	5.11	
	6x016	25+25	4.50	
	6x017	30+30	3.75	
	6x018	35+35	3.21	
	6x026	40+40	2.81	
	6x025	45+45	2.50	
	6x033	50+50	2.25	
	6x028	110	2.04	
	6x029	220	1.02	
	6x030	240	0.93	
300 VA 110×50mm 2.6 Kg Regulation 6%	7x013	15+15	10.00	**£10.17** +p/p £2 00 +VAT £1 83 TOTAL £14 00
	7x014	18+18	8.33	
	7x015	22+22	6.82	
	7x016	25+25	6.00	
	7x017	30+30	5.00	
	7x018	35+35	4.28	
	7x026	40+40	3.75	
	7x025	45+45	3.33	
	7x033	50+50	3.00	
	7x028	110	2.72	
	7x029	220	1.36	
	7x030	240	1.25	
500 VA 140×60mm 4 Kg Regulation 4%	8x016	25+25	10.00	**£13.53** +p/p £2 35 +VAT £2 38 TOTAL £18 26
	8x017	30+30	8.33	
	8x018	35+35	7.14	
	8x026	40+40	6.25	
	8x025	45+45	5.55	
	8x033	50+50	5.00	
	8x042	55+55	4.54	
	8x028	110	4.54	
	8x029	220	2.27	
	8x030	240	2.08	
625 VA 140×75mm 5 Kg Regulation 4%	9x017	30+30	10.41	**£16.13** +p/p £2 50 +VAT £2 79 TOTAL £21 42
	9x018	35+35	8.92	
	9x026	40+40	7.81	
	9x025	45+45	6.94	
	9x033	50+50	6.25	
	9x042	55+55	5.68	
	9x028	110	5.68	
	9x029	220	2.84	
	9x030	240	2.60	

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4164 200ns	6.15	5.25	4.65
2114 200ns Low power	1.25	1.15	.95
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4118 250ns	3.45	3.15	2.65
6116 150ns CMOS	4.95	4.45	3.65
2708 450ns	1.95	1.90	1.80
2716 450ns 5 volt	2.25	2.15	1.95
2716 450ns three rail	6.40	6.00	4.95
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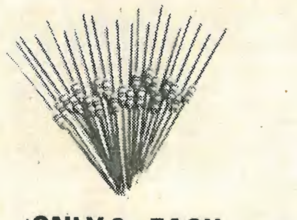

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200R	2k	20k	200k
220R	2k2	22k	220k
240R	2k4	24k	240k
270R	2k7	27k	270k
300R	3k	30k	–
330R	3k3	33k	330k
360R	3k6	36k	–
390R	3k9	39k	–
430R	4k3	43k	–
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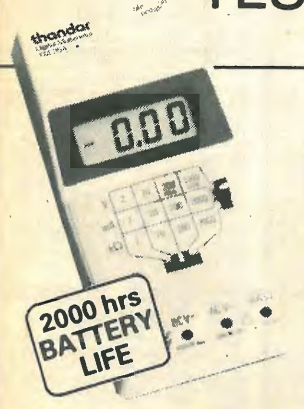

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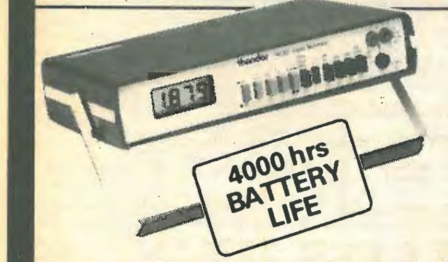

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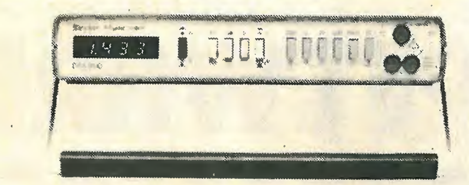

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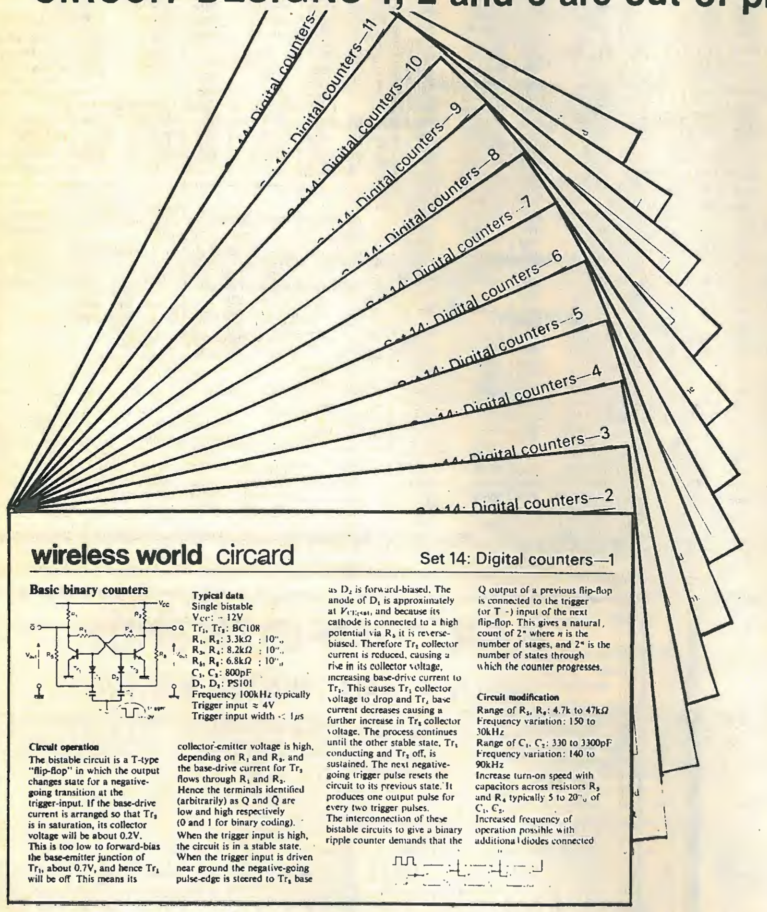

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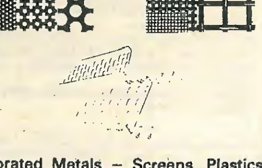

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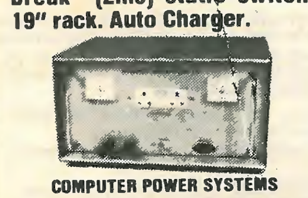

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