TELEVISION

SERVICING · CONSTRUCTION · COLOUR · DEVELOPMENTS

25p

NOVEMBER 1974

IMPROVING PICTURE QUALITY with our BLACK-LEVEL CLAMP

Also:-

RASTER CORRECTION IN 110° COLOUR SETS ■

CHANNEL IDENTIFICATION WITH VARICAP TUNERS ■

TRANSISTORS, ETC.

Type	Price (£)
AC107	0.35
AC117	0.24
AC126	0.25
AC127	0.25
AC128	0.25
AC141	0.26
AC141K	0.27
AC142K	0.19
AC151	0.24
AC152	0.25
AC153K	0.28
AC154	0.20
AC176	0.25
AC187	0.25
AC187K	0.25
AC188	0.25
AC188K	0.26
AC193K	0.30
AC194K	0.32
ACY28	0.25
ACY39	0.68
AD140	0.50
AD142	0.52
AD143	0.51
AD149	0.50
AD161	0.48
AD162	0.48
AF114	0.25
AF115	0.25
AF116	0.25
AF117	0.20
AF118	0.50
AF121	0.30
AF124	0.25
AF125	0.20
AF126	0.20
AF127	0.20
AF139	0.35
AF147	0.35
AF149	0.45
AF178	0.55
AF179	0.60
AF180	0.55
AF181	0.50
AF186	0.40
AF239	0.40
AF279	0.69
AL100	1.10
AL102	1.10
AL103	1.10
AL113	0.95
AU103	1.40
AU110	1.10
AU113	1.70
BC107	0.12
BC107A	0.13
BC107B	0.14
BC108	0.12
BC108B	0.13
BC109	0.13
BC109C	0.14
BC113	0.13
BC114	0.20
BC115	0.20
BC116	0.20
BC117	0.20
BC119	0.29
BC125	0.22
BC125B	0.25
BC126	0.20
BC132	0.15
BC134	0.20
BC135	0.15
BC136	0.20
BC137	0.20
BC138	0.20
BC142	0.30
BC143	0.35
BC147B	0.13
BC148	0.12
BC149	0.14
BC152	0.25
BC153	0.20
BC154	0.20
BC157	0.15
BC158	0.13
BC159	0.15
BC161	0.48
BC167B	0.15
BC168B	0.13
BC169C	0.13
BC170	0.15
BC171	0.15
BC172	0.14
BC173	0.20
BC176	0.22

Type	Price (£)
BC177	0.20
BC178	0.20
BC178B	0.22
BC179	0.20
BC179B	0.21
BC182L	0.11
BC183	0.11
BC183K	0.12
BC183L	0.11
BC184L	0.13
BC186	0.25
BC187	0.25
BC208	0.12
BC212L	0.12
BC213L	0.12
BC214L	0.15
BC238	0.10
BC261A	0.28
BC262A	0.18
BC263B	0.25
BC267	0.16
BC268C	0.14
BC294	0.27
BC300	0.58
BC301	0.35
BC303	0.60
BC307B	0.12
BC308A	0.10
BC309	0.15
BC323	0.38
BC377	0.22
BC441	1.10
BC461	1.58
BCY33	0.36
BCY42	0.16
BCY71	0.22
BCY88	2.42
BD115	0.65
BD123	0.98
BD124	0.80
BD130Y	1.42
BD131	0.45
BD132	0.50
BD135	0.40
BD136	0.46
BD137	0.48
BD138	0.50
BD139	0.55
BD140	0.62
BD144	2.19
BD145	0.75
BD163	0.67
BD183	0.56
BD234	0.75
BD519	0.76
BD520	0.76
BDX18	1,45
BDX32	2.55
BDY16A	0.38
BDY18	1.78
BDY20	0.99
BF115	0.20
BF117	0.45
BF120	0.55
BF121	0.25
BF123	0.28
BF125	0.25
BF127	0.30
BF158	0.25
BF159	0.27
BF160	0.22
BF161	0.45
BF162	0.45
BF163	0.45
BF167	0.25
BF173	0.25
BF177	0.30
BF178	0.33
BF179	0.33
BF180	0.35
BF181	0.33
BF182	0.44
BF183	0.44
BF184	0.26
BF185	0.26
BF194	0.15
BF195	0.15
BF196	0.15
BF197	0.17
BF198	0.20
BF199	0.25
BF200	0.35
BF218	0.35
BF222	1.08
BF224J	0.15
BF240	0.20

Type	Price (£)
BF241	0.20
BF244	0.18
BF254	0.45
BF255	0.45
BF256	0.45
BF257	0.49
BF258	0.66
BF259	0.93
BF262	0.70
BF263	0.70
BF273	0.16
BF336	0.35
BF337	0.35
BF458	0.46
BF459	0.57
BF596	0.70
BF597	0.15
BFR41	0.30
BFR61	0.30
BFT43	0.55
BFW10	0.55
BFW11	0.55
BFW16A	1.70
BFW30	1.38
BFW59	0.19
BFW60	0.20
BFW90	0.28
BFX16	2.25
BFX29	0.30
BFX30	0.35
BFX84	0.25
BFX85	0.26
BFX86	0.26
BFX87	0.28
BFX88	0.24
BFY18	0.53
BFY40	0.40
BFY41	0.43
BFY50	0.25
BFY51	0.23
BFY52	0.23
BFY57	0.32
BFY64	0.42
BFY72	0.31
BFY90	0.70
BLY15A	0.79
BPX25	1.65
BPX29	1.60
BPX52	1.90
BRC4443	0.60
BRY39	0.40
BR101	0.35
BSW64	0.38
BSX19	0.13
BSX20	0.19
BSX76	0.15
BSX82	0.52
BSY19	0.52
BSY41	0.22
BSY54	0.50
BSY56	0.80
BSY65	0.15
BSY78	0.40
BSY91	0.28
BT101/ 300	1.05
/500	1.15
BT102/ 300	1.02
/500	1.12
BT106	0.99
BU105/02	1.95
BU108	3.25
BU126	1.93
BU204	1.98
BU205	1.98
BU207	3.00
BU208	3.15
BU209	2.55
BUY77	2.50
BUY78	2.55
BUY79	2.85
C106F	0.43
C111E	0.56
CRS1/40	0.45
CRS3/40	0.55
D40W1	0.45
E1222	0.55
E5024	0.20
ME6001	0.16
ME6002	0.17
ME8001	0.18
MJE340	0.68
MJE341	0.72
MJE370	0.65
MJE520	0.85
MJE521	0.95
MJE2955	1.20

Type	Price (£)
MJE3055	0.74
MM721	0.70
MPF102	0.40
MPS6566	0.21
MPSA05	0.47
MPSA55	0.50
MPSU05	0.66
MPSU06	0.76
MPSU55	1.26
OC26	0.38
OC28	0.65
OC35	0.50
OC36	0.55
OC42	0.35
OC44	0.15
OC45	0.15
OC70	0.15
OC71	0.15
OC72	0.15
OC73	0.51
OC75	0.25
OC81	0.25
OC81D	0.30
OC139	0.28
OC140	0.30
OC170	0.25
OC171	0.30
OCP71	0.43
ON188	2.19
ON236A	0.65
ORP12	0.55
R2008B	2.05
R2010B	2.10
TAG3/ 400	1.54
TIC44	0.29
TIC46	0.44
TIC47	0.58
TIP29A	0.49
TIP30A	0.58
TIP31A	0.65
TIP32A	0.67
TIP33A	0.99
TIP34A	1.73
TIP41A	0.80
TIP42A	0.91
TIS43	0.30
TIS73	1.36
ZTX109	0.12
ZTX300	0.16
ZTX304	0.22
ZTX310	0.10
ZTX313	0.12
ZTX500	0.17
ZTX502	0.17
ZTX504	0.42
ZTX602	0.24
2N525	0.86
2N696	0.23
2N697	0.15
2N706	0.12
2N706A	0.15
2N708	0.15
2N744	0.30
2N914	0.19
2N916	0.20
2N918	0.42
2N930	0.35
2N1304	0.21
2N1305	0.21
2N1306	0.31
2N1307	0.22
2N1308	0.26
2N1309	0.36
2N1613	0.34
2N1711	0.45
2N1890	0.45
2N1893	0.48
2N2102	0.31
2N2217	0.36
2N2218	0.60
2N2219	0.50
2N2221A	0.41
2N2222A	0.50
2N2369A	0.42
2N2401	0.60
2N2484	0.41
2N2570	0.18
2N2646	0.53
2N2712	0.12
2N2904	0.22
2N2904A	0.26
2N2905	0.72
2N2905A	0.73
2N2926G	0.13
2N2926Y	0.12
2N2926O	0.12
2N3019	0.75

Type	Price (£)
2N3053	0.21
2N3054	0.55
2N3055	0.60
2N3133	0.54
2N3134	0.60
2N3232	1.32
2N3235	1.10
2N3254	0.28
2N3323	0.48
2N3391A	0.23
2N3702	0.13
2N3703	0.15
2N3704	0.15
2N3705	0.11
2N3706	0.10
2N3707	0.13
2N3715	2.30
2N3724	0.72
2N3739	1.18
2N3771	1.70
2N3772	1.90
2N3773	2.90
2N3790	4.15
2N3794	0.20
2N3819	0.35
2N3820	0.49
2N3823	1.45
2N3866	1.70
2N3877	0.25
2N3904	0.16
2N3905	0.18
2N3906	0.15
2N4032	0.31
2N4036	0.52
2N4046	0.35
2N4058	0.17
2N4123	0.13
2N4124	0.15
2N4126	0.20
2N4236	1.90
2N4248	0.12
2N4284	0.19
2N4286	0.19
2N4288	0.13
2N4289	0.20
2N4290	0.14
2N4291	0.18
2N4292	0.20
2N4871	0.24
2N4902	1.30
2N5042	1·05
2N5060	0.32
2N5061	0.35
2N5064	0.45
2N5087	0.32
2M5294	0.35
2N5296	0.37
2N5298	0.38
2N5322	0.85
2N5449	1.90
2N5457	0.30
2N5458	0.35
2N5494	1.85
2N5496	2.05
2N6027	0.65
2N6178	0.71
2N6180	0.78
2SC643A	1.36
2SC1172Y	2.80
3N140	1.21
40250	0.60
40327	0.67
40361	0.48
40362	0.50
40429	0.80
40439	2.67

All items advertised ex-stock on magazine copy date. All prices subject to availability.

Tel: Stoke Climsland (05797) 439
Telex: 45457 (A/B Mercury Calgton)

MATCHED PAIRS

Type	Price (£)
AC128/ AC176	0.52
AC141K/ AC142K	0.56
AC187/ AC188	0.60
AC187/ AC188K	0.61
AC193K/ AC194K	0.71
AD161/ AD162	0.95
BC142/ BC143	0.70

Any other transistors can be matched at a supplement of 20p per pair

DIODES

Type	Price (£)
AA113	0.15
AA119	0.09
AA129	0.20
AA143	0.10
AA213	0.30
BA100	0.15
BA102	0.25
BA110U	0.30
BA115	0.12
BA141	0.17
BA145	0.17
BA148	0.17
BA154	0.13
BA155	0.16
BA156	0.15
BA157	0.25
BAX13	0.06
BAX16	0.07
BB103	0.25
BB104	0.45
BB105B	0.45
BB105G	0.35
BR100	0.50
BY100	0.15
BY103	0.22
BY126	0.16
BY127	0.17
BY133	0.23
BY140	1.40
BY164	0.55
BY176	1.00
BY179	0.70
BY206	0.31
BYX10	0.15
BYZ12	0.30
FSY41A	0.30
OA10	0.20
OA47	0.07
OA81	0.12
OA90	0.08
OA91	0.07
OA95	0.07
OA200	0.10
OA202	0.10
OA210	0.29
OAZ237	0.78
S2M1	0.22
TV20	1.20
IN914	0.07
IN914E	0.06
IN916	0.10
IN4001	0.05
IN4002	0.06
IN4003	0.07
IN4004	0.08
IN4005	0.09
IN4006	0.11
IN4007	0.14
IN4148	0.05
IN4448	0.10
IN5400	0.15
IN5401	0.17
IN5402	0.20
IN5403	0.22
IN5404	0.25
IN5405	0.27
IN5406	0.30
IN5407	0.34
IS44	0.07
IS310	0.45
IS920	0.07
IS923	0.12

Diodes can be supplied balanced at a supplement of 5p per device—e.g. four balanced OA91 would be £0.48 per set.

Variable capacitance diodes can be supplied matched at a supplement of 3p per device.

LINEAR INTEGRATED CIRCUITS

Type	Price (£)
CA3046	0.70
CA3065	1.90
MC1307P	1.19
MC1310P	2.94
MC 1327PQ	1.01
MC1330P	0.76
MC1351P	0.75
MC1352P	0.72
MC 1358PQ	1.85
MC1496L	0.87
MC3051P	0.58
MFC 4000B	0.43
MFC 4060A	0.70
MFC6040	0.91
PA263	1.90
SL414A	1.91
SL901A	2.60
SL917B	3.80
SN 76003N	2.92
SN 76013N	1.95
SN76013 ND	1.72
SN 76023N	1.95
SN76023 ND	1.72
SN 76033N	2.92
TAA300	1.46
TAA320	0.94
TAA350	1.54
TAA435	0.85
TAA450	1.85
TAA550	0.49
TAA570	1.39
TAA611	0.73
TAA630Q	3.29
TAA630S	3.29
TAA700	3.30
TAA840	1.64
TAA861A	0.49
TAA100	1.42
TBA120	0.68
TBA120S	0.99
TBA240A	1.10
TBA480Q	1.24
TBA500	1.99
TBA500Q	2.00
TBA510	1.99
TBA520Q	2.72
TBA530	1.98
TBA530Q	1.99
TBA540	2.20
TBA540Q	2.21
TBA550Q	3.29
TBA560C	2.71
TBA 560CQ	2.72
TBA570	1.17
TBA641	0.76
TBA673	1.80
TBA700	1.90
TBA720Q	2.20
TBA750Q	1.54
TBA800	1.75
TBA 810AS	1.75
TBA920Q	3.29
TBA990	3.29
TBA990Q	3.29
TCA270Q	3.30
ZN414	1.25
μ6A995- 159	2.25
709:	
TO99	0.45
DIL8	0.45
DIL14	0.47
710:	
DIL14	0.50
711:	
DIL14	0.62
723:	
TO100	0.95
DIL14	0.95
741:	
TO99	0.48
DIL8	0.50
DIL14	0.55
748:	
DIL8	0.52

DIGITAL INTEGRATED CIRCUITS

Type	Price (£)
7400	0.20
7401	0.20
7402	0.20
7404	0.24
7406	0.45
7408	0.25
7410	0.20
7411	0.25
7412	0.28
7413	0.50
7416	0.45
7417	0.30
7420	0.20
7430	0.20
7440	0.20
7441	0.85
7447	1.30
7450	0.20
7451	0.20
7454	0.20
7460	0.20
7470	0.33
7472	0.38
7473	0.44
7474	0.48
7475	0.59
7490	0.65
7491	1.10
7492	0.75
7493	0.65
7494	0.85
8495	0.85
7496	1.00
74100	2.16
74121	0.60
74122	0.80
74150	1.44
74151	1.15
74154	1.66
74164	2.01
74192	2.05
74193	2.30

HARDWARE

BASES

Type	Price (£)
DIL8	0.16
DIL14	0.16
DIL16	0.18

PADS

T0-5	½p
T0-18	½p

MOUNTING KITS

T0-3	0.06
T0-66	0.06

VALVES

Type	Price (£)
DY87	0.39
EB91	0.30
ECC82	0.41
EF80	0.41
EF183	0.53
EF184	0.53
EH90	0.55
PC86	0.67
PC88	0.76
PCC89	0.58
PCF80	0.47
PCF86	0.58
PCF801	0.58
PCF802	0.63
PCL82	0.50
PCL84	0.54
PCL805/ 85	0.58
PCL86	0.58
PFL200	0.74
PL36	0.80
PL84	0.61
PL504	0.80
PL508	0.95
PL509	1.44
PY81/800	0.45
PY88	0.52

ZENER DIODES

400mW 3.0-33V 12p each
1W 3.3-68V 18p each

MINIATURE BRIDGES

2A Plastic Encapsulated

N6001	50V PIV	35p each
N6002	100V PIV	38p each
N6003	200V PIV	41p each
N6004	400V PIV	45p each
N6005	600V PIV	50p each
N6006	700V PIV	59p each
N6007	800V PIV	68p each

VDR'S, PTC & NTC RESISTORS

Type	Price (p)	Type	Price (p)
E295ZZ /01	12	E299DD/P116- P354	all 7
E295ZZ /02	12	VA1015	20
E298CD /A258	6	VA1026	17
E298ED /A258	5	VA1033	7
/A260	5	VA1034	7
/A262	5	VA1040	7
/A265	5	VA1053	7
/P268	5	VA1055S	9
E298ZZ /05	6	VA1077	10
/06	5	VA1104	27
		VA8650	85

RESISTORS

Carbon Film (5%)
¼W 5.6 Ω-330k Ω (E12) 1p ea
½W 10 Ω-10M Ω (E24) 1p ea
1W 10 Ω-10M Ω (E12) 2p ea
2W 10 Ω-10M Ω (E12) 3p ea

WIREWOUND (5%)
2½W 0.22 Ω-270 Ω 15p ea
5W 10 Ω-8.2k Ω 12p ea
10W 10 Ω-25k Ω 15p ea

CAPACITORS

Full range of C280, C296, tubular ceramic, pin-up ceramic, miniature electrolytics, mica, mixed dielectric and TV electrolytics stocked.—Please see catalogue.

MASTHEAD AMPLIFIERS
Labgear type CM6000 complete with mains power unit CM6001/PU.
Groups A, B, or C/D £9.87

PATTERN GENERATORS
Labgear CM6004/PG giving crosshatch dots, greyscale and blank raster on 625-lines. Tuning can be preset for anywhere in Bands IV and V as well as Band III (for relays). £45.39

COLOUR-BAR GENERATORS
Labgear CM6037/DB: Dual standard bar generator gives standard 8 band colour bars +greyscale step wedge+red raster+centre cross+centre dot+crosshatch+dot pattern+blank raster. Sync output also provided £101.31

VHF/UHF CONVERTERS
Labgear "Televertas" for DX-ing or single-standard receiver use on relay systems.
Type 6018/RC £10.00
Type 6022/RA (with pre-amp gain control) £12.27

AERIALS, ETC.
Please see catalogue.

OUR NEW CATALOGUE IS NOW AVAILABLE AT 30p (refundable).

P. & p.: UK £0.08 per order
Overseas: At cost
Please Add 8% for V.A.T.

EAST CORNWALL COMPONENTS

CALLINGTON - CORNWALL

TELEVISION

VOL 25 No 1
ISSUE 289

SERVICING·CONSTRUCTION·COLOUR·DEVELOPMENTS

NOVEMBER 1974

A BENEFIT TO THE PUBLIC?

The abolition of retail price maintenance some years ago was at the time hailed as a major breakthrough in giving the consumer a better bargain. Some of us always wondered whether in fact it would turn out to be something of a mixed blessing. True the discounters and cut-price stores have flourished, bringing benefits to many who keep an eye out for a bargain and, sometimes a necessity, have their own transport. Others may feel however that they don't know quite where they stand in a totally free—or free-for-all—marketplace. If you've no standard price you've no yardstick with which to assess what's on offer. There are still manufacturers' recommended prices of course, but how realistic are these? It has often been suggested and seems likely that they are set unrealistically high so that each and every retailer can dress up his window with "price-slashed" offers. It certainly seems that many imported monochrome sets have recommended price tags that never anywhere appear to apply. There is another snag. Just what does a price include? There's not the scope in the domestic electronics trade that there is in some others to play about with "extras", but it's still possible to feel that you've come off worse than you expected.

Liberal economists seem to have a rather unreal picture of the shopper as someone with all the time in the world to enquire, get quotes and leaflets, ask around and "shop around". This ideal has never seemed to bear much relation to the average busy person's need to fit his shopping in amongst the many other claims on his limited available time. More recently we have swung in the opposite direction, towards extensive consumer protection. Some may seriously wonder whether the net result has been to make confusion total. All good stuff for the legal boys who set up the test cases to decide what legislation does or doesn't mean!

It is difficult to sum up the pros and cons of free pricing. Prices have probably been kept down by competition, to the benefit of the consumer. Whether standards of dealing have been kept up is another matter. Then there is the problem now confronting us rather strikingly: how does a high turnover system of retailing manage when the market collapses? Profits of many of the more adventurous concerns have plummeted. Could we be left with a drastically reduced number of retail outlets and a consequent contraction in consumer choice?

These problems are likely to arise in a new form before long since it is now proposed that servicing and services be brought within the scope of restrictive trade practices legislation. This means for a start that recommended minimum service charges could be outlawed. The public is not generally aware of these anyway, but they do nevertheless give some sort of guidance as to what a fair charge is—always assuming the use of competent staff backed by adequate equipment.

The whole subject of recommended prices and charges can generate a great deal of emotional heat. It is a crucial matter however, determining the nature of the interface between a trade and its public. The proposal to bring servicing within the legislation on restrictive practices is an opportunity for the whole matter of what is fair practice in this area to be brought up for public scrutiny. We hope that the opportunity to do so will be taken, rather than that some doctrinaire decision will be hurriedly implemented. It is not very reassuring that only two months have been allowed for comment, and that an order is proposed within six months.

L. E. HOWES—*Editor*

THIS MONTH

THE NEXT ISSUE DATED DECEMBER WILL BE PUBLISHED ON NOVEMBER 18

TELETOPICS

MARKET UPTURN?

Deliveries by UK setmakers of colour and monochrome TV sets showed a welcome improvement in July, with colour set deliveries the highest for five months. The fall in colour set deliveries during the first seven months of the year compared to the same period in 1973 was 14%, not nearly as bad as was once feared, though monochrome set production really seems to be going through the floor, down by a massive 44% compared to the equivalent period in 1973 which was itself the worst year for UK monochrome receiver production for twenty years. This need not concern us too much however with the public still avid for colour TV. These figures are exclusive of imports—the Customs Statistical Office computer still seems to be hors de combat—but at a guess these must be down quite substantially. How continental setmakers are faring at present is not known but in Japan where colour set sales have fallen by 32.4% the scene is one of production cutbacks all round. That massive 85% colour set market penetration we mentioned in this column in September must be having an effect, along with the slowing down of export markets. At the end of May Japanese colour setmakers were understood to have over a million colour sets in stock. Present plans by Japanese colour setmakers seem to be concentrated on efforts to establish overseas plants to serve their export markets. Most Japanese setmakers are now established in the US, and Matsushita (National Panasonic) Japan's largest setmaker has announced plans to follow Sony in setting up a plant in South Wales. The site chosen is on the outskirts of Cardiff and Matsushita expect to be producing five thousand sets a month there by the beginning of 1976. Let's hope the market is forging ahead by then.

COLOUR RECEIVER FAULTS

Enough time has now gone by to be able to summarise the faults experienced with colour receivers. So here goes. As with monochrome sets the majority of faults are in the power supplies or timebases, so the usual monochrome fault diagnoses apply. The added problems are the more complex power supplies, especially in all solid-state chassis, and the more elaborate line output stage circuitry, with c.r.t. first anode and focus supplies giving rather more trouble than in monochrome sets. Colour drifting is commonly due to the first anode supply networks and the plugs and sockets associated with the preset controls.

Convergence circuitry does not seem to have presented any insuperable problems. Faults seem to consist of controls burning out, faulty electrolytics, dry-joints and AC128 clamp transistors playing up—sometimes causing convergence drifting. There is the occasional unfortunate tube/scan coil/convergence yoke combination that simply will not converge nicely, but this is evident upon initial delivery. Experiences with 110° chassis do not seem to be too happy.

At the start of colour reception and servicing in the UK we were inclined to worry about the terrors the complex decoder might have in store—especially when reading about experiences in the USA. But whilst US colour decoders have been primarily valved the vast majority of PAL decoders have been transistorised. The resultant low voltage/low current operation has meant that this is one of the most reliable sections of a colour receiver, with never a suggestion of a charred resistor. Faults seem to be mainly defective transistors, the occasional faulty crystal, electrolytics and diodes and the odd defective capacitors which cause some of the more difficult faults. Faulty power supply lines can also occasionally produce odd results. The transistors and diodes presumably fail as a result of transients, and with improved flashover protection this seems to be less of a problem than in earlier sets. Colour faults are sometimes caused by lack of or low-amplitude burst gating pulses but the trouble here is generally back in the timebase section.

With colour-difference drive the colour-difference output stages and clamps and the high-power luminance output stage bring us back to valve defects and burnt resistors. RGB drive seems to give rather less trouble: the occasional faulty transistor, defective coupling electrolytics, and sometimes a tricky clamp fault.

Have we missed anything significant? Let us know if your experiences have been different.

RECEPTION PROBLEMS

We get queries from time to time about difficulties in receiving one of the u.h.f. channels in a local group while the others are received perfectly satisfactorily. The BBC have commented recently on this problem. It seems that there can be a very complex pattern of reflected signals over irregular terrain and especially over rooftops in urban areas, and that this is the basic cause of the trouble. If careful aerial positioning does not give any improvement a more directional aerial should be tried. You guessed it, a log-periodic type. The BBC report that in a few locations they have found more than 10dB difference (more than three to one ratio) in signal strength between the strongest and weakest channel in a local group. In such extreme cases the BBC suggest trying the use of separate aerials in different positions.

A recent PO survey found that in 87% of TV aerial installations the coaxial feeder was not soldered to the receiver aerial input plug. Whilst this is not likely to cause problems initially, oxidation over a period of time can result in a high-resistance connection with impaired reception in consequence. An unsoldered high-resistance connection can also act as a diode, increasing the possibility of interference from nearby transmitters through acting as a mixer.

CEEFAX/ORACLE DECODER

The first decoder for the Ceefax/Oracle news transmissions in data form has been announced by Jasmin Electronics of Leicester. This is expected to be available early next spring at a price of about £150 plus tax. It is envisaged mainly for sale by dealers to business customers who may find the "instant news" service of particular benefit. This is on the expensive side but specialised i.c.s for this application are still in the development stage and the use in the meantime of i.c.s from currently available ranges means that quite a number are required—and not cheap ones at that. RRI's development head Bernard Rogers has suggested that it will be 1977 before sets incorporating Ceefax/Oracle decoders are available in the shops. This is because of the time it will take—some 18 months—to develop and put into production the specialialised l.s.i. i.c.s for this application that will bring the price of a decoder down to the level where it becomes a domestic proposition. By then the use of as few as three specialised i.c.s should bring the price of an adaptor down to the region of £60.

TRANSMITTER OPENINGS

The high-power (100kW) transmitters at the **Chatton** (Northumberland) u.h.f. station are now in operation. BBC-1 is radiated on channel 39, BBC-2 on channel 45 and ITV (Tyne Tees Television programmes) on channel 49. Horizontally polarised group B receiving aerials are required. In addition the following relay transmitters have now been brought into operation:

Blaina (South Wales) ITV (HTV Wales programmes) channel 43. Receiving aerial group B.

Cwmafan (South Wales) ITV (HTV Wales programmes) channel 24. Receiving aerial group A.

Ogmore Vale (South Wales) ITV (HTV Wales programmes) channel 60. Receiving aerial group C/D.

Peterhead (Aberdeenshire) BBC-1 channel 55, BBC-2 channel 62. Receiving aerial group C/D.

Sedbergh (Cumbria) BBC-1 channel 40, BBC-2 channel 46. Receiving aerial group B.

Stanton Moor (Peak Park area) BBC-1 Midlands channel 55, BBC-2 channel 62. Receiving aerial group C/D.

Taff's Well (South Wales) ITV (HTV Wales programmes) channel 59. Receiving aerial group C/D.

These relay transmissions are all vertically polarised.

SALARIES IN TV SERVICING

SERT has issued details recently of the salaries paid in 1973 to qualified staff in the radio and television servicing industry. For the under 20s salaries started at about £1,300, averaged £1,800 for the 20-25 year old group, rose to just under £2,100 by the age of 30 and stabilised at around £2,200 above this age. SERT found that although salaries improved in 1973 they still lagged behind those paid to similarly qualified staff in broadcasting, industrial electronics, education, the civil service and the forces.

TV GAMES KIT

A TV games kit—the Videomaster—is to be made available "in time for the Christmas trade" by The Sales Team, 119/120 Chancery Lane, London WC2A 1QU. The recommended price is £57.75 including VAT. The kit has joystick controls for each player and plugs into the set's aerial socket. It is said to "simulate a surprising number of the features that normally make up a game of football, tennis and squash", a cautious way of describing what appears to be quite a versatile system. The unit—which is understood to be UK made—employs 17 i.c.s and 20 transistors and is powered by a PP7 battery. It will carry a one year guarantee.

ECONOMICS OF THE DOMESTIC TELLY

According to stockbrokers Greene and Co. the fifteen year period of constant TV set prices is coming to an end. The reasons given are that no further economies in set manufacture are possible, while competition from imported sets will cease to keep prices down since the rates of inflation in the countries of origin are now similar if not worse than in the UK. Not only set prices but rental charges are likely to rise. Annual maintenance costs work out at £16 a set whilst rental overheads work out at £13. The cost of labour, spares and overheads is expected to rise at the rate of 15% a year while the cost of the sets themselves is likely to rise at the rate of 5% a year. A recommendation for improved profitability made recently to rental organisations is to prolong set life (something our readers certainly seem to know how to do): it is suggested that some sets can continue to be profitable after twelve years while others may have to be scrapped after nine years, the main determining factor being the availability of spares (most setmakers seem to think five years a long enough life for a set, though the problems of keeping stocks of parts for large numbers of obsolete chassis are not to be belittled). A warning given is that colour sets may be cheaper relative to monochrome ones by 1978 so that the rental life of monochrone sets could now be decidedly limited.

DIGITAL MONITORING SIGNALS

The use of unmanned transmitters has led to the development of automatic monitoring techniques. The BBC recently announced the conclusion of successful field trials of a new technique of transmitting digital monitoring information: this uses differential phase modulation of an existing low-level pilot subcarrier signal which is at present radiated by TV sound transmitters for continuity monitoring. The system makes use of a very narrow band of frequencies above the useful audio range. The BBC comment that the mode of transmission, coupled with error protection in the transmitted message, results in a very rugged system which can operate even when the signals are degraded to the point where the vision and sound channels themselves are unusable. The pilot subcarrier is presumably the signal detected by some Sony KV1800UB colour sets (see *Some Foreigners* last month) as intereference, leading to shut down of the sound channel in the set if it is fitted with a sound interference suppressor board.

RASTER Correction for 110° COLOUR SETS

HAROLD PETERS

WITH six years of colour behind us we take purity, grey-scale tracking, static and dynamic convergence in our stride. The introduction by several setmakers of 110° colour sets brings our complacent selves back to earth with a bump however by adding a new dimension in the form of raster correction circuits for us to adjust. Enthusiasts will rub their hands at the prospect of more lovely knobs to twiddle; but to those who don't look upon innovations with such relish we bring the glad tidings "don't panic, it's easy".

First, why should raster correction circuits be needed? Think back to the 110° monochrome set and to the way in which the sawtooth line scan current has to be modified by means of a series scan-correction capacitor which slows down the scan at both ends of the line to compensate for the fact that the scanning light spot would otherwise travel faster at the edges of the picture than towards the centre of the screen where it is closer to the tube cathode. Even with s-correction the raster shape is like a pincushion and needs little fixed bar magnets carefully positioned around the scan coils to provide correction. Bar magnets cannot be used to provide this sort of correction in a colour set however since if they were able to penetrate the tube's internal shield (the degaussing shield is internal with 110° tubes) they would influence each beam by a different amount. So, as with picture centring, it is necessary to pass a current through the scan coils in order to get the effect we require.

Controls and Their Effects

Six raster correction controls are usually provided in the new sets: E–W pincushion and E–W keystone which affect the sides of the raster; N–S phase, N–S amplitude and N–S symmetry which affect the top and bottom of the screen; and a second harmonic transformer which irons out a wiggly droop predominant in the upper central picture area. The East–West adjustors are usually resistors—the width control is also included in this part of the circuit—while the North–South adjustors are mainly inductances. The correction actions, simplified since in practice a badly adjusted set exhibits a combination of all five distortions, are illustrated in Fig. 1.

Since the correction waveforms are fed into the scan coils all three guns are equally affected and no terrible misconvergence results. Correction is carried out after convergence therefore, on a white crosshatch. The conventions "N–S" and "E–W" have no significance other than to associate control functions with the appropriate knobs: they merely refer to the top, bottom, right and left raster edges respectively.

If you are familiar already with the way in which a transductor works in 90° colour sets—providing pincushion distortion correction by feeding parabolic waveforms at field frequency into the line deflection circuit and vice versa—you are already half way towards grasping the working principle of the new raster correction circuits. Basically what has to be done is to modulate the line scan at field frequency (E–W correction) and the field scan at line frequency (N–S correction).

Circuitry

Taking the easy bit first we'll deal with N–S (top–bottom) pincushion errors. The raster correction circuitry as used in the Pye group's 731 chassis is shown in Fig. 2, with the N–S part—where the field scan is being varied at line frequency—at the bottom. An adequate supply of line pulses is available from the pulse winding on the line output transformer (top of the diagram). −350V and +60V pulses are taken to the right-hand side of transductor T619 via a diode shaping circuit in one leg and choke L615 in the other. The transductor has a variable cylindrical permanent magnet at one pole end and by means of this the saturation of the transductor ferrite can be varied (N–S symmetry) to provide an adjustment to correct bowing at the top and bottom edges of the raster. On the secondary (field) side of the transductor C621 (0·047μF) and the N–S phase coil L618 resonate at line frequency. The N–S phase coil adjusts the lateral tilting of the raster edges and the degree of its effect is determined by the 2·2kΩ N–S amplitude control RV624 which adjusts the amount of correction waveform fed into the field scan coils.

There are no prizes offered for guessing that these adjustments are a little interdependent, or that even

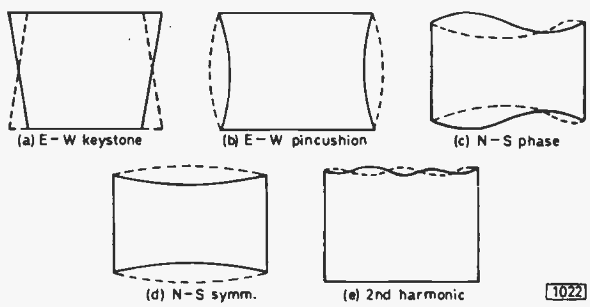


Fig. 1: The action of the controls used in the raster correction circuitry. Remember that they affect all three guns and are thus adjusted on a white raster; also that on a maladjusted set all five forms of distortion can be present at the same time.

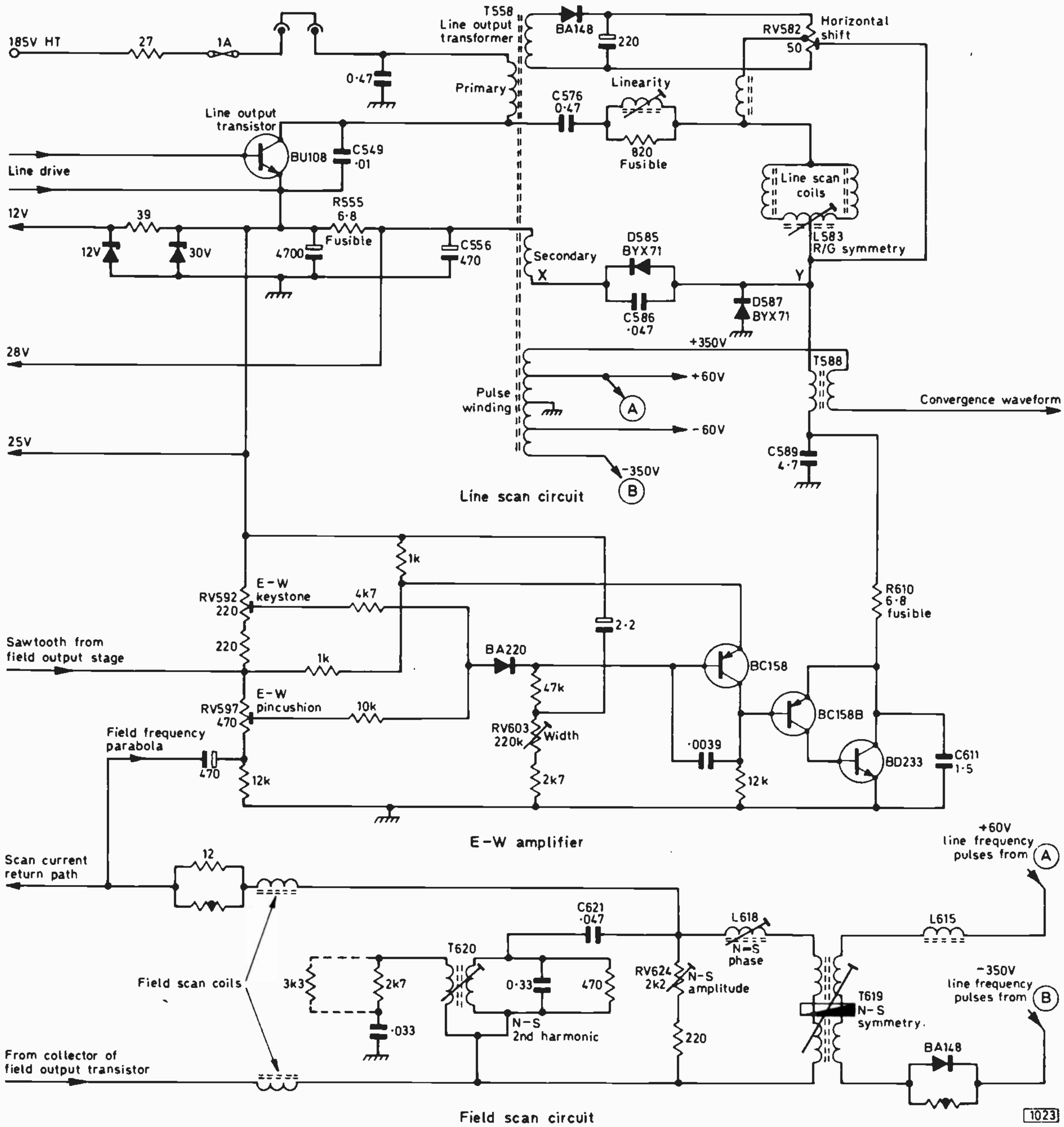


Fig. 2: The 110° colour tube raster correction circuitry (Pye 731 chassis). The top section contains the line output stage (for clarity the e.h.t. and c.r.t. heater windings on the line output transformer have been omitted) and the diode modulator. In the bottom section line waveforms are fed into the field deflection circuitry to give N–S correction. The E–W amplifier in the centre section boosts the amplitude of the field frequency correction waveform for feeding into the line scan.

when optimised a moustache-shaped droop can appear on the foreheads of close-ups, noticeable more on 26in. models than the smaller sizes. To straighten up this droop a small amount of second harmonic of the line frequency is fed into the field scan by T620. The effect of T620 is slight and sudden as you adjust it and a cord stretched across the screen may help you to see the effect better.

So for N–S correction we feed line frequency pulses into the field scan. The reverse procedure is less easy to apply unfortunately. Amplification is needed to be able to drive the line scan circuits with the correct amount of field frequency correction. The three-stage amplifier used for this purpose forms the centre portion of Fig. 2. There are two inputs to this amplifier, a field frequency sawtooth component and a field frequency parabolic component. The E–W keystone control RV592 varies the sawtooth input while the E–W pin-

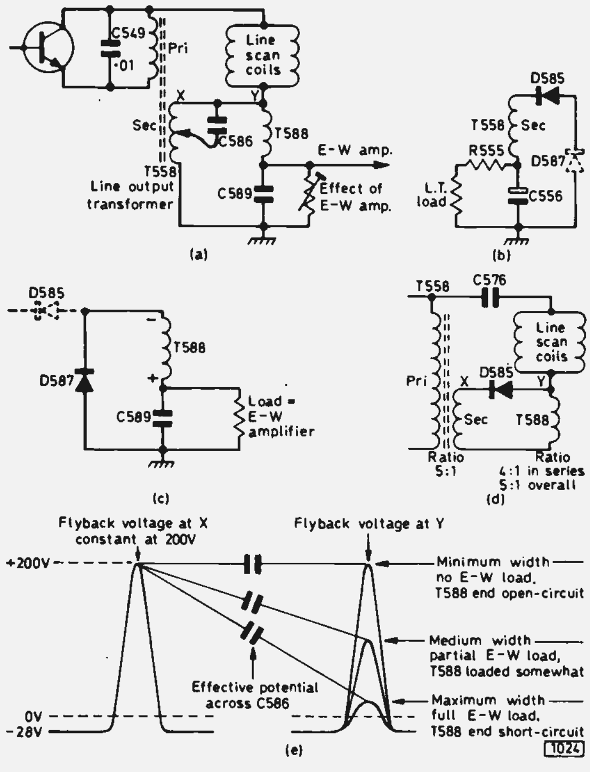


Fig. 3: (a) The circuit elements that affect the line output stage fifth harmonic tuning and thus the e.h.t. The tuning is kept constant by varying the effect which C586 has across the line output transformer secondary winding in inverse proportion to the damping which the E–W amplifier places across T588. (b) The main components responsible for producing the l.t. supplies used in the receiver. (c) The components which generate the 28V supply for the E–W amplifier—extra filtering is provided by R610/C611 (see Fig. 2). (d) Turns ratios: the 5 : 1 ratio of the line output transformer primary and secondary windings equals the 4 plus 1 : 1 ratio of the sum of the scan coils plus T588 to T588 itself. (e) The waveforms at each end of C586 (at points X and Y) at different widths. At maximum width C586 is effectively between point X and chassis while at minimum width it is effectively between two points at the same voltage and thus does not influence the tuning.

cushion control varies the parabolic input. The width control RV603 sets the gain of the amplifier and the output stage collector load consists of the primary winding of transformer T588 and the 6·8Ω fusible protection resistor R610. This resistor is marked (S) on most circuits, indicating that it is a safety component required for BEAB approval and must be replaced only with another of the same approved type. The amplifier gives us sufficient power to feed our field correction into the line scan then and we come up against the big snag, the easy bit ending abruptly here.

The E-W Modulator

The difficult bit consists of the E–W diode modulator D585/D587 which is at the top right of Fig. 2, in the line scan circuit. The workings of this are complex and are broken down in Fig. 3.

Fig. 3(a) shows the elements which concern the line output transformer fifth harmonic tuning and which if varied will affect the e.h.t. From this point of view the s-correction capacitor C576, the linearity control, the R–G symmetry control and the secondary winding on T588 (this winding is simply a convenient way of picking up a convergence waveform and plays no part in the raster correction activities) can all—and have been—disregarded. The E–W amplifier can be looked upon as a variable load across the modulator transformer T588 although in fact it is between one end and chassis, across the 4·7μF capacitor C589. Capacitor C586 is shown in this figure as being tapped up and down the associated secondary winding on the line output transformer. The object of the game is to slide C586 farther down the winding to add capacitance in the same proportion as we lose inductance by adding a load across T588. In this way the fifth harmonic tuning and hence the e.h.t. are kept constant.

This secondary winding also provides—as shown in Fig. 3(b)—the receiver's l.t. supplies. D585 acts as the l.t. rectifier and 28V is produced across the reservoir/smoother C556. All the transistors and i.c.s in the chassis form its considerable load. During the line scan when D585 is conducting D587 can be regarded as a short-circuit to chassis. Note that the l.t. supply is obtained by scan rectification, not be rectifying the flyback pulses.

As Fig. 3(c) shows the 28V supply for the E–W amplifier output stage is also obtained by scan rectification, this time using the signal across T588 rectified by D587 and smoothed by C589 together with the protection resistor R610 and capacitor C611. So far so good!

The turns ratio of the line output transformer primary to the secondary winding is 5: 1 while the ratio of the scan coils to T588 primary is 4: 1. If—see Fig. 3(d)—you look upon the junction of the scan coils and T588 primary as the tap on an autotransformer however the ratio also becomes 5: 1 (4+1=5). Thus points X and Y are from an a.c. point of view balanced across the middle of a bridge. These then are the basic conditions around the circuit which we are about to vary at field rate in order to straighten the sides of the raster.

During the line scan diodes D585 and D587 conduct, shorting points X and Y to chassis in the course of providing the two 28V supplies just mentioned. The scan coil current is heavy and the circuit voltages low.

During the flyback the voltages are high and the current low, and since D585 and D587 are not conducting we can pretend they aren't there. Only C586 appears across X and Y, and the other ends of the line output transformer secondary and T588 primary are tied to their respective 28V lines.

To vary the width of the line scan the E–W amplifier places a varying load across C589 (4.7μF) which is in series with T588 primary. Point Y is connected to chassis by D587 during the scan. Thus the line scan coils and T588 form a kind of seesaw pivoted on point Y. For *minimum width* the E–W amplifier load is open-circuit and the bottom of T588 wags up and down at line rate. The total scan is shared therefore between the scan coils and T588, four parts to the former and one to the latter. Only four-fifths of the total scanning power is used by the scan coils. For *maximum width* the E–W amplifier short-circuits C589, connecting the lower end of T588 to chassis. Since the other end of T588 is connected to chassis by D587 T588 primary is completely shorted out and all the available scan power goes into the scan coils. In a *medium width* condition C589 is partly damped by the E–W amplifier and the a.c. voltage at the bottom of T588 will reach say 14V. T588 thus appears to the circuit to be less inductive than when at minimum width and will take only one tenth of the scanning power, leaving nine-tenths for the scan coils.

The E–W amplifier output is parabolic at field rate and in the reverse sense to the pincushion distortion it is to correct. At the top and bottom of the raster the width must be reduced while at the centre it must be increased. Thus the loading on C589 is minimal at the start and finish of the field scan and maximal in the middle. Now let's see how the fifth harmonic tuning is kept constant during the flyback.

Maintaining Constant Tuning

Since the width is being varied at field rate by T588 the tuning of the line output stage will also vary if uncorrected—to the extent that a 1kV ripple will appear on the e.h.t. Correction is supplied by C586 (0·047μF) which is across D585. As Fig. 3(e) shows, the voltages at either side of it are equal in the minimum width condition. This is because T588 is undamped during the scan and its back-e.m.f. on flyback is the same as the back-e.m.f. produced by the line output transformer secondary—the turns ratios being balanced and equal. C586 is tied across two points at the same potential therefore and has no effect on the circuit. In the maximum width position however T588 is shorted out and can thus store no energy during the scan to produce a back-e.m.f. during the flyback. So although point X still shoots up to +200V point Y remains close to chassis potential: C586 is thus effectively between point X and chassis and this is the same as being across the line output transformer secondary winding. In the medium width condition T588 is only partly damped and stores enough energy during the scan to produce say 100V at point Y during the flyback. C586 is now between point X at 200V and point Y at 100V. This is equivalent to being between the top of the line output transformer secondary and a point part-way down.

C586 is thus varied continuously across the line output transformer secondary winding in such a way that its effect is greatest when the effect of T588 is least. Its effect is reflected by the line output transformer so that it appears across the fifth harmonic tuning capacitor C549 in the ratio $1/n^2$. Since $n=5$ and C586 is 0·05μF (approximately) the effective capacitance across C549 is $1/(25\times 50)=0{\cdot}002$μF at maximum width and zero at minimum width. The tuning capacitor C549 is 0·01μF so the ratio 10:2=5:1 is maintained. Like we said at the beginning, it's easy!

Conclusion

The raster correction circuitry we have been discussing forms part of the Mullard Phase II 110° colour circuitry. The Pye group's 731 chassis has been taken as our example but the same basic circuitry will be found in the RR1 (Bush/Murphy) Z179 chassis. The associated 110° convergence circuitry was described in the October 1973 issue.

Although this article has been concerned with only one aspect of the new, 110° colour circuitry the writer would in conclusion like to digress to a point which can give trouble—purity. The new sets are much more difficult to purify than their 90° counterparts and more sophisticated methods are used in set manufacture than have been required hitherto. So if the purity is anything like acceptable once the installation has been *thoroughly* degaussed (no short cuts) it may pay to leave well alone.

If you have to purify on site without a beam-landing periscope or other aid you are stuck with the red ball method. Do it very carefully, having marked mid-screen exactly so as to get the ball perfectly central. As you spread the red raster to the corners, use the facility provided by the four wing nuts on the scan coils instead of the usual two. Check finally on a white raster, making a final touch-up if required. ■

CHANNEL IDENTIFICATION

—from page 18

The circuit is so simple it can be mounted on a small tagstrip or piece of Veroboard attached to the meter terminals—provided only that there is sufficient space when the meter is installed within the television set. Feed the base of the transistor from the tuner's control terminal itself so that a.f.c. if applied is taken into account. The customary safety precautions must be taken and the meter mounted in such a way that no live parts can be reached from the outside of the cabinet.

Check the circuit when installed by operating the preset tuning potentiometers. This should enable the meter deflection to be varied from near zero to f.s.d. The minimum reading may not coincide with true zero due to the characteristics of the tuner. The original meter scale must be obliterated—easily done by sticking a self-adhesive white paper label over it.

For accurate results it is necessary to mark the new scale at points corresponding to known channels, as explained above. This involves a risk of touching live parts so that all possible precautions should be taken. Extend the meter well away from rest of the receiver when you are doing this. If the receiver chassis is connected to one side of the mains supply—as nearly all are—check that this is the neutral side. Otherwise the receiver must be fed from a mains isolating transformer whilst the calibration is carried out.

When the points corresponding to known channels have been marked (take care to avoid parallax error) the delicate task of interpolation must be undertaken. This is helped by plotting the known channels on graph paper against their measured distance from the zero point. In this way the shape of the non-linearity can be seen. Only channels 21, 30, 40, 50, 60 and 68 need be numbered—mark the others with short lines as shown (Fig. 2). This avoids cluttering the scale. Work is made easier if the scale can be removed from the meter while this is done—use a pencil so that corrections can be made if necessary.

The final step is to replace the scale in the meter and check the calibration by tuning in as many known transmissions as possible. Note any errors—in both magnitude and direction—so that they can be corrected. Then make the final version permanent by inking in.

In practice this will take a few hours. The result however is a most useful addition to any electronically tuned receiver. ■

CCTV

Part 8 Peter Graves

LAST month we saw how a sync pulse generator generates pulses at line and field frequencies from a master oscillator so that the pulses have the accurate timing relationship needed for high-resolution, interlaced pictures. The master oscillator runs at twice line frequency and the line frequency pulses are produced by a simple divide-by-two circuit. The master oscillator also feeds a more elaborate frequency divider chain that produces field frequency output pulses. To complete the description of a typical SPG we must look at the various ways of controlling an SPG from an external reference (e.g. to achieve mains lock) and also methods of processing the basic line and field pulses to produce pulses of the correct amplitude and width at the SPG output sockets.

Sync Pulse Generator Control

Fig. 1 shows in block diagram form the various SPG control modes that may be encountered. In the simplest mode the master oscillator runs free. If, as is often the case, the master oscillator circuit consists of a transistor blocking oscillator a fixed bias supply will be applied to its base. This mode of operation is used only for initial setting up however as its frequency stability is not adequate for general use. For mobile or other uses where mains lock is impractical (with a portable generator for instance) crystal control is employed. The master blocking oscillator can then be run as a buffer stage synchronised by the output from the crystal oscillator, or the master blocking oscillator may be linked out and the crystal oscillator used as the master oscillator instead. This means that in practice things are rather more complicated than the simple mode switch shown in Fig. 1 suggests.

High-quality SPGs may be suitable for synchronising colour cameras: in this case an external subcarrier generator is used in place of and in the same way as the crystal oscillator.

External a.f.c. (automatic frequency control) is similar to mains lock in that the master oscillator frequency is controlled by being brought into coincidence with an external reference source. This is done by applying a d.c. correction voltage to the master oscillator. In the external position the correction voltage is supplied by a source which does not generally take the mains directly as its reference. This mode may for example be used to lock two SPGs together, one controlling the other so that they are kept synchronised.

The most important and perhaps most common mode is mains lock (called simply a.f.c. in America). In this mode the master oscillator is controlled by a feedback circuit which maintains a constant phase relationship between the SPG output and the mains. The essentials are shown in Fig. 1. The output from the frequency divider is suitably shaped to obtain a rectangular field drive pulse which is 400μS long and in the UK at 50Hz. An auxiliary winding on the mains transformer provides from the mains a low-voltage (typically 6.3V) a.c. signal. These two signals are then fed to a phase comparator circuit which produces the d.c. correction voltage.

Phase Comparison

One form of phase comparator is shown in Fig. 2. The transistor is normally off and is switched on for the duration of the negative-going field drive pulse which is applied to its base via C1. It will be noticed that the value of R1 is much higher than the value of R2 and R3 in series. Thus the lower end of R1 is effectively at earth potential, holding the transistor off in the absence of a field drive pulse. As far as the a.c. signal from the mains transformer is concerned however R1 has too high a value to affect the potential divider action of R2 and R3. Since the transistor's emitter is connected to the junction of R2, R3 it follows that the voltage between the emitter and earth varies sinusoidally as the a.c. signal varies, at a level selected by the values of R2, R3.

Now one complete cycle of 50Hz sinewave lasts 20mS while the field drive pulse lasts only 400μS, one fiftieth of the time. To a first approximation we can say that when the transistor turns on its emitter will be at an instantaneous voltage whose magnitude will depend on which part of the a.c. cycle has been reached at the time. Obviously this voltage can vary between positive and negative maximum values, through zero.

The circuit is designed so that when the two signals are in phase the field drive pulse turns the transistor on at the instant when the a.c. signal is passing through zero (Fig. 3). The emitter will then be at earth potential. When the transistor turns on a current whose value is determined by the circuit constants will flow, causing a voltage drop across R4 and R5. The output voltage at their junction will be at some value *V* volts with respect to earth. Since current flows only for the duration of the field drive pulse the output voltage will be in the form of pulses: C2 smooths these pulses to provide the d.c. control potential.

Correcting Drift

Suppose the master oscillator drifts slightly so that the transistor's turn on point no longer coincides with the sinewave zero crossing but occurs earlier as shown in Fig. 3 (a). When the transistor turns on the instantaneous emitter voltage will this time be some positive value instead of zero volts and as a result the emitter-base junction will be forward biased to a greater extent than previously (pnp transistor). Consequently a greater current will flow and the output voltage will drop below *V* volts. Conversely if the drift is in the other direction—Fig. 3(b)—the emitter will at turn on be more negative so that there is less base-emitter junction forward bias, less collector current will flow and the output voltage will rise above *V* volts. Thus the magnitude of the output voltage and the direction in which it changes depend on the magnitude and direction of the phase error between the two waveforms.

The output error voltage can be superimposed on the steady d.c. bias applied to the base of the master blocking oscillator, altering the master oscillator

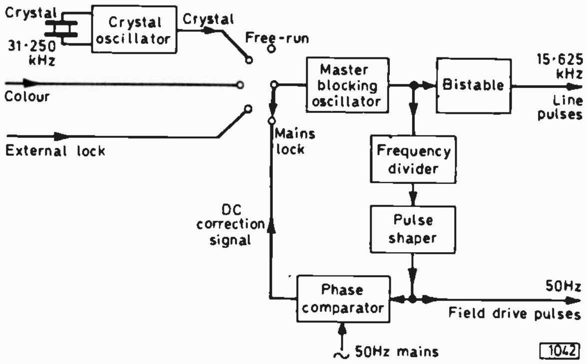


Fig. 1: Master oscillator operating modes.

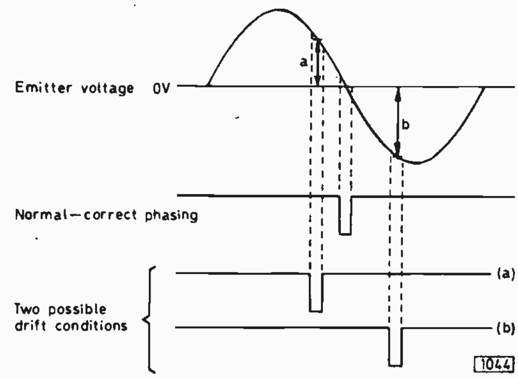


Fig. 3: Phase comparison conditions.

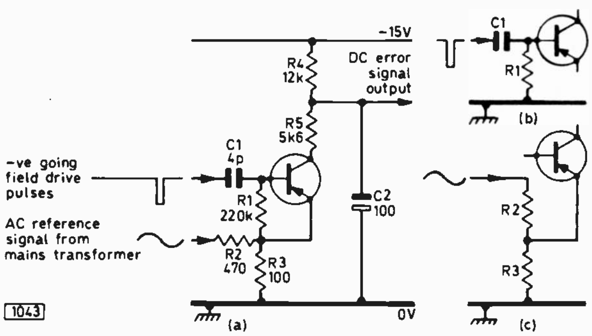


Fig. 2: (a) One form of phase comparator circuit. (b) The effective base circuit. (c) The effective emitter circuit.

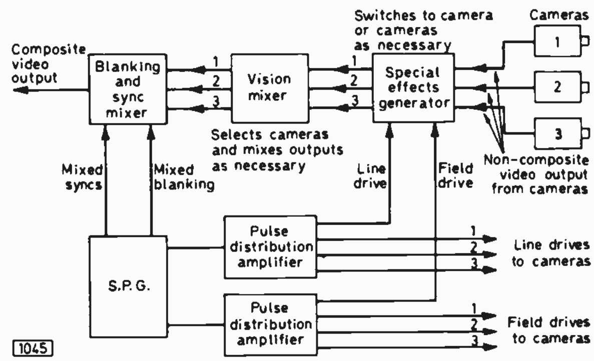


Fig. 4: Typical studio arrangement using non-composite outputs from the cameras.

frequency until the correct phasing conditions are again reached. The process is continuous so that the mains and the SPG maintain a constant phase relationship, i.e. mains lock is achieved.

Where a master oscillator circuit uses *LC* networks for frequency determination —for example the Armstrong oscillator found in some US equipment—the master oscillator frequency can be controlled by using the d.c. error signal to bias a variable-capacitance diode connected across the tuned circuit.

SPG Outputs

In general SPGs for CCTV work have four outputs—line drive, field drive, mixed blanking and mixed syncs. The term "mixed" implies that the signal contains both line and field components. Although all these outputs may be supplied to the camera only two—the line and field drives—are directly used by the camera, to drive the respective timebases. The other two sets of pulses are mixed with the non-composite video signal from the video amplifier at some stage near the video amplifier output. In more sophisticated cameras this is not strictly true as the drive pulses for the timebases may be derived within the camera from various combinations of inputs. Typical combinations would be: mixed syncs only, line drive only (the field drive pulse being derived from the mains) and mixed sync and mixed blanking pulses as separate feeds or combined into a single signal. The circuits for this are similar to those described later in this article. What is used depends on what is available: for the finest work the camera would be fed with all four pulses (or at least the drives if the system described below is used), but economy in cabling with minor degradation in picture quality can be achieved by using a simpler system.

Studio Arrangements

Alternatively—and here we are talking in the main about studio systems—the cameras can be driven by just the line and field drives and a non-composite video signal piped from the camera output socket to the vision mixer where the outputs from the various cameras are added together and selected as required, or perhaps via a special effects generator (which as shown in Fig. 4 is also synchronised to the overall system by being fed with line and field drives from the SPG), the final (or system) syncs and blanking (mixed of course) being added to the final output before it is distributed to the monitors or the modulator in a transmitter. To go a step farther only the system blanking may be added and the signal fed to the monitors in non-composite form, the monitors being supplied with a separate sync feed. The first system (fully composite video out to the distribution system from the cameras) is probably the most common for small studios however.

Pulse Processing

Exactly how the signals from the master oscillator are processed to provide pulses of the correct amplitude and duration varies from manufacturer to manufacturer and there is room here to cover only some of the more common circuits. In addition a simple field waveform which—typical of CCTV practice—does not

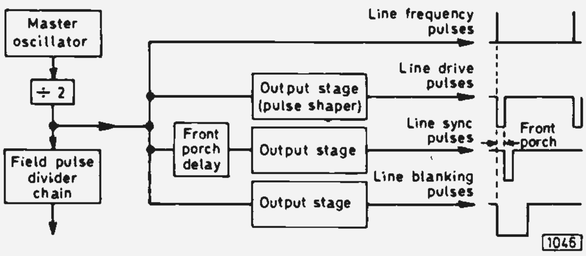


Fig. 5: Developing the pulses required at line frequency.

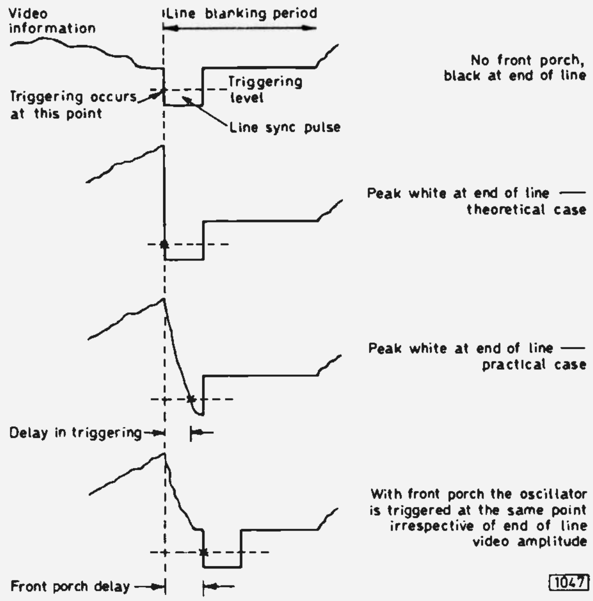


Fig. 6: Illustrating the need for the front porch.

include equalising pulses or pulses during the field sync pulse has been assumed. The more elaborate waveforms are encountered only in high quality, broadcast standard equipment. The generation and mixing of these is done by similar though more complex circuits than those described below.

Line Frequency Pulses

Fig. 5 shows the various output pulses that must be developed at line frequency, together with a block diagram of the circuits involved. Although of different durations the line blanking and line drive pulses start at the same instant. Things are a little more complicated with the line sync pulse as it must be delayed with respect to the start of the other pulses. The delay period (about 1.5μS for a 625-line system) is known as the front porch. The corresponding period between the end of the line sync pulse and the end of the line blanking pulse is called the back porch.

Don't forget that the line sync pulse is the monitor's line sync pulse. Suppose that part of the picture at the extreme right-hand side is very bright—peak white. When such a line occurs the signal cannot—for practical reasons such as the time needed for capacitors in the signal circuit to discharge—fall instantaneously at the end of the line to the blanking level. Instead the signal decays, taking a short but finite time to reach the blanking level. If the line sync pulse occurred at the start of the flyback (see Fig. 6) without a front porch delay period the triggering level of the sync circuits would not be reached until after the peak white signal had decayed sufficiently. On a line which has black information at its right-hand end however the triggering level would be reached sooner. Thus lines ending with white would be displaced (since the line oscillator would be triggered at a later time) with respect to lines not containing white information at the right-hand edge. Note that this would affect the whole line, and would affect different lines as the scene shifts—a distracting fault. The front porch is introduced to allow the signal to decay to the blanking level before the sync pulse occurs, so that triggering takes place at the same point whatever the scene content at the end of any particular line.

Generating a Delay

There are several ways in which the front porch delay can be generated. A monostable circuit (e.g. a blocking oscillator or a multivibrator) can be used to generate a pulse whose duration is equal to the front porch period, the back edge of this pulse triggering a further monostable which generates the actual sync pulse. This arrangement is shown in block diagram form in Fig. 7 together with the various waveforms at different parts of the circuit. Alternatively the line drive

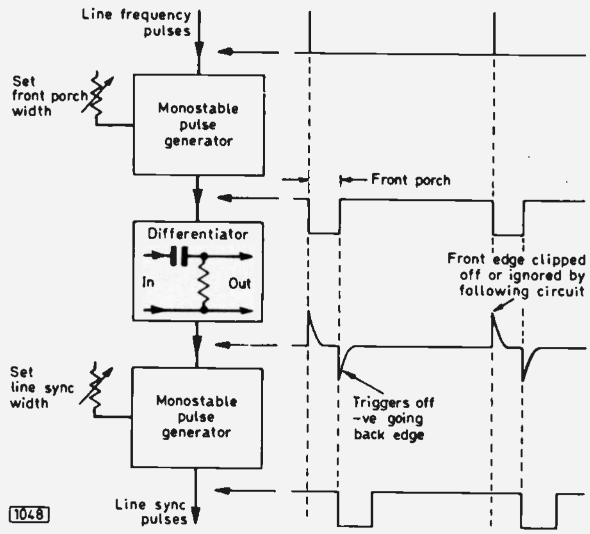


Fig. 7: Generating the front porch (not to scale).

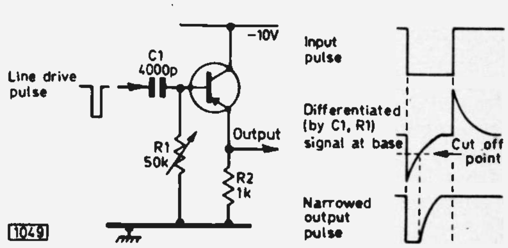


Fig. 8: Pulse narrower circuit.

pulse may be fed to a pulse narrower circuit which again produces a pulse whose length is equal to the front porch period and is processed as before.

One version of the pulse narrower circuit is shown

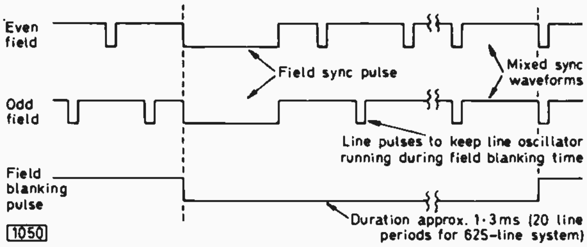


Fig. 9: The field frequency pulses required (not to scale).

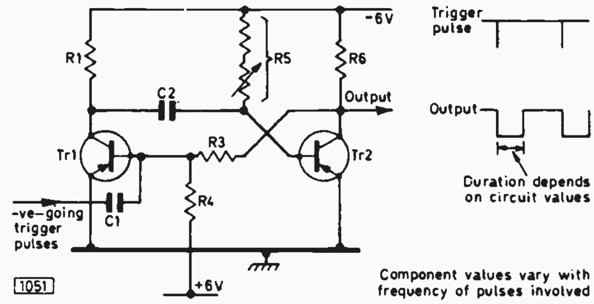


Fig. 10: The monostable multivibrator circuit.

in Fig. 8. C1 and R1 differentiate the incoming drive pulse whose leading edge switches the transistor on. As the differentiated spike decays the transistor's base becomes more positive, the transistor cutting off at some point as shown in the adjacent waveforms. Once the transistor cuts off the output voltage remains constant until the next input pulse arrives. The length of the output pulse obtained in this way can be adjusted by altering the time-constant of the differentiating circuit C1, R1—R1 is made partly variable to allow for setting up since the exact cut-off point depends on the individual transistor.

Field Frequency Pulses

Fig. 9 shows the ends of the two fields and the vertical blanking period of a typical industrial CCTV sync pulse generator waveform. Notice that this is mixed syncs and that line sync pulses occur during the field blanking period. There is no field pulse front porch: this is not necessary since the time taken for a peak white signal to fall to the blanking level is very much shorter than the duration of the field sync pulse and can be neglected therefore. If the line sync pulses are omitted from the field blanking period the monitor line oscillator will tend to drift off frequency and the top of the monitor picture could be "hooked" over until the oscillator is pulled back into sync by the line sync pulses present during the field scan. The field drive pulses, field blanking pulses and field sync pulses can be generated by circuits similar to those used to generate the line frequency pulses.

The Monostable Multivibrator

The circuit common to all these processes is the monostable multivibrator. Initially the monostable circuit is in a stable state. Once it has been triggered by a suitable pulse however it changes to an unstable state for a predetermined period of time—which is independent of the trigger pulse—before resuming its initial state. The basic monostable multivibrator circuit with its characteristic cross-coupling is shown in Fig. 10. In the absence of a trigger pulse Tr2 is biased fully on by R5. This means that a high collector current flows through R6 and in consequence the voltage at Tr2 collector will be just below chassis potential. The potential divider R3, R4 is connected between this potential and a +6V rail, thus biasing Tr1 off.

If a negative-going trigger pulse is applied to the base of Tr1 via C1 Tr1 will start to turn on, its collector voltage will start to rise towards earth and this change will be coupled to the base of Tr2 by C2. Tr2 will start to turn off and the potential at its collector will start to drop towards the −6V rail. As a result of the cross-coupling (R3, R4) the potential at the base of Tr1 will also drop towards the −6V rail, turning Tr1 further on. Once the action has been started by the trigger pulse there is a rapid regenerative action which does not cease until Tr2 has cut off and Tr1 has been driven into saturation (fully on).

This is not a stable state however since as soon as it is reached C2 starts to charge through R5. When Tr1 is fully on its collector potential and hence the potential at the left-hand side of C2 is pretty well at earth. As C2 charges the voltage at the junction of R5 and C2—the base of Tr2—falls towards the −6V rail and at some stage in the charging process the base of Tr2 will again be forward biased, turning it on. This initiates a regenerative action in reverse, the circuit resuming its initial stable state until the next trigger pulse arrives. Note that the amplitude and duration of the output pulses are independent of the amplitude and duration of the trigger pulses. In particular the duration of the output pulse depends on how quickly C2 charges. Other things being constant, the duration of the output pulse can be controlled by varying the time-constant of C2, R5. Thus R5 is made partly adjustable to set the required output pulse length. What we get then is a good square pulse from a brief "spike" trigger pulse.

Overall Sync System

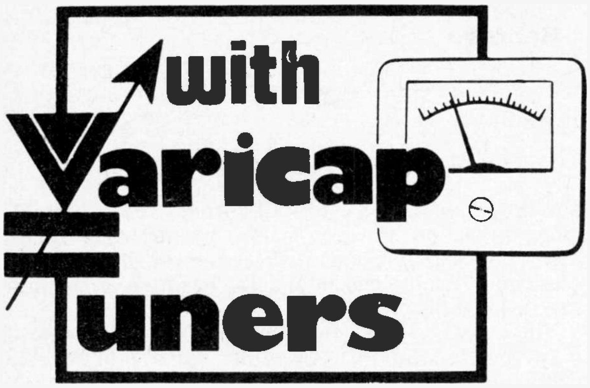
The SPG forms the heart of any installation in which it is used. To ensure maximum reliability two identical SPG circuits sharing a common power supply are sometimes mounted in the same chassis. Thus in the event of failure of one of the units the other can be immediately switched into use. In more sophisticated versions the changeover is made automatically by a monitoring circuit at the output.

The standard output level of a SPG is 2V into a 75 Ω load.

If more than one output is needed from the SPG—see Fig. 4—pulse distribution amplifiers must be used. These have a single 75 Ω input and a low-impedance output which can supply a number of points (typically 6). The outputs must be correctly terminated however to obtain the right levels and it is important to terminate unused outputs in 75 Ω. In a permanent installation any unused outputs can have 75 Ω resistors soldered on to the back of the output sockets (in which case beware if you need to use one of them—double terminations do little for pulse amplitude and sharpness) or stuffers—terminating resistors inside suitable plugs—may be used.

This completes our coverage of SPGs. Next month we shall be looking at the intricacies of the camera scan circuits and how they differ from television receiver practice.

CHANNEL IDENTIFICATION with Varicap Tuners



ALAN F. REEKIE B.Sc.(Eng.)

THE electronic or varicap tuner which has no moving parts has been widely adopted by setmakers during the last few years in place of the mechanical tuner in which the channel required is selected by adjustment of the position of a multigang tuning capacitor. The varicap tuner has many advantages but one minor disadvantage is that no accurate indication of the channel selected is generally provided. This is of little consequence to the normal viewer who quickly becomes accustomed to the set of switches which enable him to select the local channels. It causes great frustration to the long-distance television enthusiast however since he has great difficulty in identifying the source of distant transmissions as a result of the uncertainty regarding the channel selected. It is possible nonetheless to provide accurate channel identification in sets using varicap tuners by means of the following simple and inexpensive meter circuit.

Principle of Operation

With a varicap tuner the required channel is selected by applying a stable d.c. voltage in the range 0–30V to the tuning pin. This voltage sets the bias on the varicap diodes in the tuner, thus adjusting their capacitance. The voltage is generally derived from a group of preset potentiometers which can be selected by means of push-switches. The tuning potentiometers are fed from a 30V line stabilised by an i.c. regulator such as the TAA550 (see Fig. 1).

Although the relationship between the capacitance of a varicap diode and the bias voltage applied to it is not linear neither is the relationship between the capacitance and the channel selected: in fact the combined result is such that the relationship between the applied voltage and the channel selected is only slightly non-linear. This means that a conventional voltmeter can be used, with a suitably calibrated scale, to display directly the channel selected. To avoid loading the tuning potentiometer circuit however high impedance is essential. The commonly available moving-coil type of meter cannot be used therefore without an impedance matching stage. A suitable stage can easily be designed using an emitter-follower (see Fig. 2). This type of circuit is very insensitive to component tolerances and requires very little power.

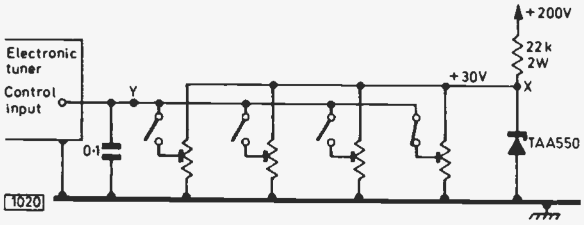


Fig. 1: Basic circuit of a four-channel control system for an electronic (varicap) tuner.

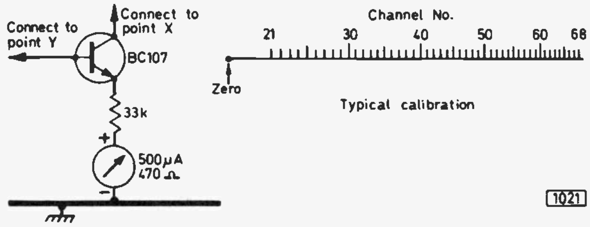


Fig. 2: A meter driven by an emitter-follower stage can be used to provide accurate channel identification.

The meter scale is calibrated directly with the channel numbers. This is not as difficult as may be feared now that the u.h.f. TV spectrum is becoming well filled and several different stations can be received at most locations. The channels used by more or less local stations can be ascertained from the usual sources (e.g. the BBC and IBA handbooks or the IBA's excellent pocket guide to UK transmitting stations).

Another technique that can be used to verify the channel of a received signal is based on harmonic radiation from the local oscillator of a v.h.f./f.m. radio receiver. By tuning the radio to 89·3MHz for example the local oscillator will be working 10·7MHz higher and will give harmonics at integral multiples of 100MHz. These fall within channels 25, 37, 50 and 62. By coupling the radio and TV sets together—it is often sufficient to feed the two aerial inputs from a dual wall outlet—a noise pattern will be seen on the television set when it is tuned to any of these channels.

Design and Construction

Design of the circuit is determined by the characteristics of the meter to be used. The following calculation is given as a typical example. Suppose that the meter is a 500µA f.s.d. type with a 470 Ω coil impedance and that a BC107 (or equivalent—BC207, BC237 etc. or other npn device with a *V*ceo greater than 30V) is to be used. The tuner characteristics must be measured in operation: suppose they vary from 2·5V at channel 21 to 16·1V at channel 68—if necessary extrapolate graphically from voltages corresponding to locally available channels. For full-scale deflection at channel 68 the voltage at the emitter of the transistor must be 16·1 − 0·6 = 15·5V. The total emitter circuit resistance = V/A = 15·5/0·0005 = 31k Ω. As the meter's internal resistance is 470 Ω the value of the series resistor required is 31,000 − 470 = 30,530 Ω. In practice a 33k Ω resistor is suitable.

—turn to page 13

MODERN TV POWER SUPPLY CIRCUITS

Part 3

E. J. Hoare

THYRISTOR controlled power supply circuits also depend upon a switching action but the principle of operation is completely different. It is based on the characteristics of the thyristor which is a rather unusual device. In effect it consists of a high-voltage diode which cannot conduct until its gate electrode is made a few volts positive with respect to its cathode. Once it has been turned on in this way it cannot be turned off in like manner: conduction ceases only when the anode voltage falls until it is nearly equal to the cathode voltage, the current then falling to zero. Once this occurs the device remains non-conducting until another trigger pulse is applied to the gate.

Most thyristors are capable of withstanding high voltages and of conducting large peak currents. Thus a thyristor can be turned on at any desired instant in time by applying a trigger pulse and it will then pass a large current if so required.

This gives the clue about how to use it for providing an h.t. stabilising action. Suppose (see Fig. 7) we connect a thyristor in series with the incoming mains supply and feed a steep trigger pulse to its gate sometime during the *reverse* slope of the positive part of the sinewave, i.e. after the peak. Depending upon the precise timing of the pulse we can connect the mains to a reservoir capacitor in order to obtain an h.t. voltage of any value between 0V and the peak of the sinewave. After the first few positive half cycles of the incoming mains sinewave have passed the reservoir capacitor will have charged to the HT1 voltage as shown. As soon as the thyristor's anode voltage falls to the same level as the HT1 potential the thyristor turns off and cannot conduct again until the next trigger pulse arrives during the next positive half cycle.

The circuit as shown is not a stabilised circuit. It merely produces a d.c. voltage equal to a preset fraction of the peak a.c. input voltage. If we add a feedback loop which varies the arrival time of the trigger pulse in sympathy with changes of the h.t. voltage however we can stabilise the h.t. voltage at any required level below the peak of the smallest sinewave input. This is shown diagrammatically in Fig. 8.

Let us clarify this point about using the reverse slope of the sinewave in case the principle is not already clear. Suppose the thyristor was switched on during the first half of the positive half cycle, i.e. on the leading edge—the one which arrives first. The thyristor would then remain conducting as the voltage increased towards the peak, and the capacitor would be charged to this peak voltage. When the sinewave started to fall the anode voltage would fall below the cathode voltage and the thyristor would turn off. Thus the capacitor would always be charged to the peak voltage and no control would be possible—it would not make the slightest difference where on the leading edge the trigger pulse arrived.

Representative Circuit

Having established the basic mode of operation of a thyristor supply we can consider an actual circuit and investigate in greater detail how it works and what factors are involved in its design. We will take an up-to-date circuit currently in mass production, that used in the Philips 320 monochrome chassis. It is a straightforward example of a thyristor controlled power supply, without any external complication apart from the addition of a simple safety circuit, and is used in the 17, 20 and 24 in. models in the present Philips range of slimline monochrome receivers.

The basic performance requirements of this circuit are the provision of a smoothed h.t. line of 160V at a load current of 350mA, having a ripple of less than 1V peak-to-peak and a preferred source impedance of greater than 50 Ω. Since the h.t. voltage is well below the peak mains input voltage some form of overvoltage protection is needed as a precaution against the possibility of the thyristor becoming short-circuited with the result that the h.t. rises to somewhere in the region of 280V. A failure in the control circuit could cause a similar effect.

The complete circuit is shown in Fig. 9, with simplified component numbering based on the relevant service manual. Starting at the input the first point to note is the use of a bridge rectifier. The reasons for using a bridge in preference to a simple and cheaper diode are twofold. First, with a bridge the circuit operates at 100Hz instead of 50Hz, making the h.t. filtering easier. The higher cost of the bridge rectifier is more than offset by the saving obtained through the use of smaller electrolytic smoothing capacitors. The second point concerns the electricity supply authorities. With a bridge rectifier operating at 100Hz equal currents are drawn from the supply during each half cycle of the mains, i.e. during both the positive and the negative mains voltage excursions. Thus no direct current flows: only a.c. pulses. This mode of operation complies with the requirements of the Electricity Council who wish to avoid the difficulties caused by

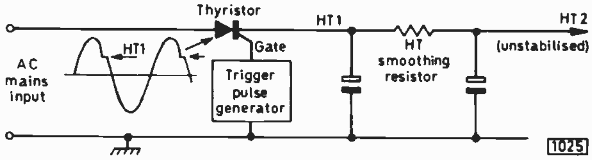


Fig. 7: Illustrating the basic action of a thyristor power supply.

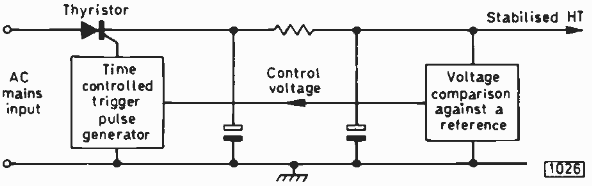


Fig. 8: A stabilised thyristor power supply.

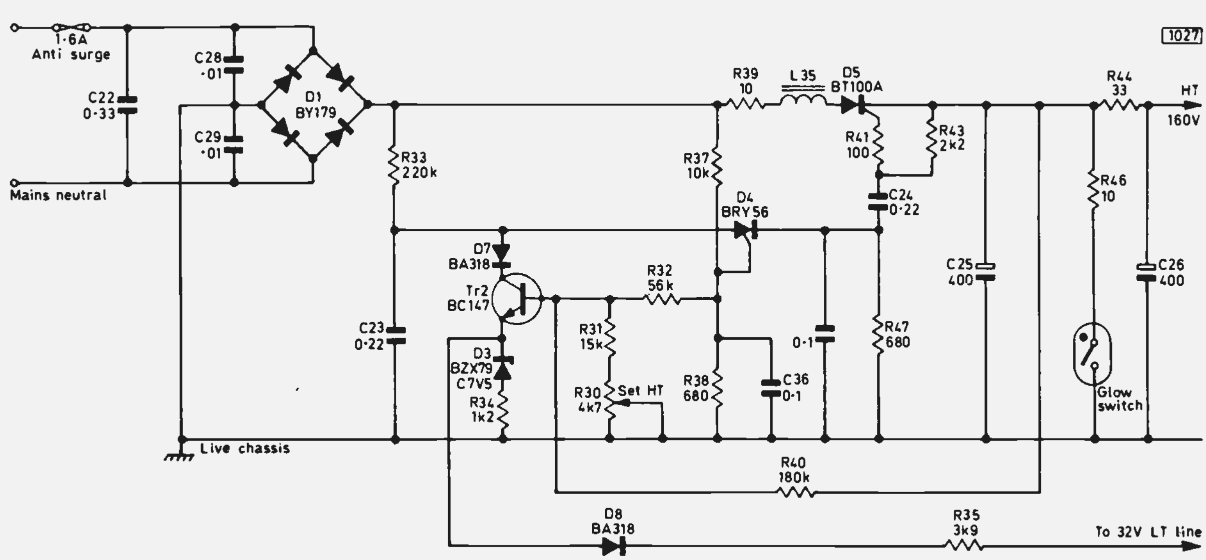


Fig. 9: Complete circuit of the power supply used in the Philips 320 monochrome chassis.

d.c. being drawn from their a.c. supply—these result in the need for larger transformers and transmission lines and increased power losses.

A side effect of using a bridge rectifier is that the receiver chassis is always "live", regardless of which way round the incoming mains leads are connected. This point should be carefully noted when carrying out any service work. The chassis is negative with respect to true earth on alternate cycles of the mains input, as can be deduced from the simplified circuit shown in Fig. 10.

Triggering the Thyristor

We come now to the action of the thyristor itself. A simplified circuit showing how the trigger pulse is generated is shown in Fig. 11. The key element here is the silicon controlled switch (SCS for short) D04. This is a four-layer pnpn device which may be unfamiliar to some readers. We had better say a bit about it before considering the circuit in which it is used therefore.

An SCS is a device which is either fully conducting or completely turned off and can be switched from one state to the other extremely quickly. It has two gate electrodes though usually only the anode gate is brought out externally. It is thus more commonly supplied as a three-terminal device.

For ease of explanation it is convenient to regard it as consisting of a circuit equivalent to two transistors in cascade, one pnp and the other npn. This is shown in Fig. 12. If the SCS is turned off it can be turned on either by applying a positive voltage to the cathode gate or by making the anode positive with respect to the anode gate. When either of these events occurs the appropriate transistor conducts and causes base current to flow in the other transistor so that it also conducts. This positive feedback action between the two transistors results in the composite device turning on very quickly indeed. The SCS turns off when the anode current falls below a certain critical level. We thus have a device that behaves as an almost perfect switch.

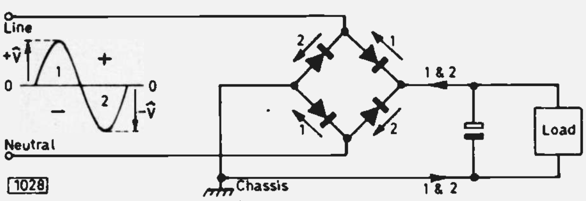


Fig. 10: The use of a bridge mains rectifier results in a live chassis: on half cycle 1 the chassis is connected to mains neutral (earth) while on half cycle 2 the chassis is connected to mains line (peak voltage).

Back to Fig. 11. The long time-constant *CR* network R33, C23 generates a sawtooth from the incoming half sinewave and applies it to the anode of the SCS. The input half sinewave is also potted down by R37 and R38 to give a peak value of about 25V at the anode gate. When the rising sawtooth voltage at the anode just exceeds the falling sinewave at the anode gate the SCS conducts. As a result capacitor C23 discharges very rapidly through the SCS and R47, developing a steep voltage pulse across this resistor. The pulse is coupled to the gate of the thyristor via C24 and R41, causing it to conduct instantly.

The incoming mains sinewave is thus connected to the h.t. reservoir capacitor C25 (Fig. 9), charging it to almost the instantaneous value of the sinewave. Now if the sawtooth was a little steeper the SCS would fire earlier, the thyristor would be triggered earlier and the h.t. voltage would rise. Conversely a smaller sawtooth would result in later triggering and thus a lower value of h.t.

Stabilising the Output

We have thus reached the point of having an almost loss-free power supply with a switching action replacing the traditional voltage drop across an h.t. resistor. We have no stabilising action however. Fig. 13 shows the first step in achieving this. A transistor Tr02 has been added across C23 to provide a variable current bleed. If we bleed more current the sawtooth developed across C23 rises more slowly and the h.t. falls. If the bleed current is reduced however the sawtooth is steeper and the h.t. rises—as explained above. The problem now is to control this transistor current bleed.

Resistor R32 applies to the base of transistor Tr02 a small portion of the potted down incoming sinewave. If we choose the value of this resistor carefully we can balance the circuit so that when the sinewave increases in amplitude it turns on the transistor a bit more so that the sawtooth is less steep. This causes the thyristor to be triggered later, compensating for the increased sinewave amplitude and resulting in an unchanged h.t. Similarly a reduced sinewave results in less transistor current, a steeper sawtooth and earlier thyristor triggering to compensate for the smaller sinewave.

This control action is known as "feedforward" and is a special form of negative feedback. Extra feedforward does not improve the performance: it unbalances the circuit and spoils the stabilising action. It should be quite clearly distinguished from time feedback action therefore.

The zener diode in the emitter circuit of the transistor is included for two reasons. First, its positive temperature coefficient compensates for the negative coefficient of the transistor's base-emitter junction, thus stabilising the control action against changes of temperature. Secondly it provides a constant voltage of 7.5V as a reference against which the control voltage is compared. This is a convenient level for the operation of the circuit.

The feedforward action stabilises the h.t. line against changes in mains input voltage, at least to a good approximation. It does not stabilise the h.t. against changes in load current however and feedforward on its own would not result in an adequate standard of performance over a long period of time. The answer to this is to apply true negative feedback action from the h.t. line to the current bleed transistor. Any change in h.t. then results in a compensating action through an appropriate change of bleed current in parallel with the sawtooth generating capacitor.

Negative feedback is applied by means of R40 which is connected between the h.t. line and the base of the control transistor. What this does is to compare a fraction of the h.t. voltage, determined by the ratio of R40 to R30+R31, against the emitter voltage+Vbe. If the h.t. rises for example the transistor's base voltage rises, its collector current increases, C23 charges more slowly, the thyristor is triggered later and the h.t. is restored very nearly to its proper value. This value is preset by potentiometer R30.

Circuit Precautions

So we now have a complete design—or have we? The full circuit shown in Fig. 9 still has some features to discuss. Take for example diode D07. The base of the control transistor is at a potential of about 10V. When triggering is about to begin C23 has charged to about 25V and this is also the collector voltage of the transistor. When the SCS starts to conduct the voltage on C23 falls rapidly as it discharges. When the voltage falls below 10V the collector-base junction of the transistor would without D07 behave as a diode. As a result the capacitor could charge via the transistor's base-collector circuit. This would mean that it may not be fully discharged therefore—in fact its state of charge would depend upon the transistor's base voltage, i.e. upon the h.t. voltage. The result is incorrect control action, leading to a second or repeated triggering of the SCS, and hence the thyristor, during each mains half cycle. Diode D07 blocks this current path from the base

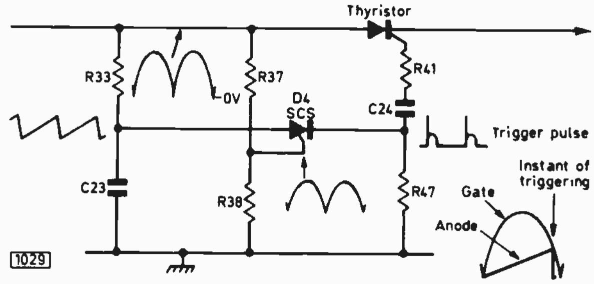


Fig. 11: Generating the thyristor trigger pulse. Note that the waveforms shown are idealised.

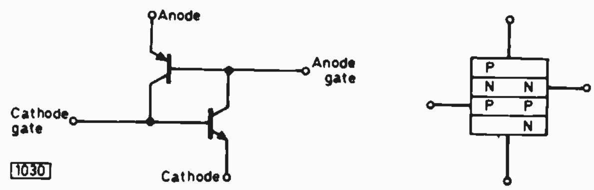


Fig. 12: Equivalent circuit of the silicon controlled switch which is a four-layer pnpn device.

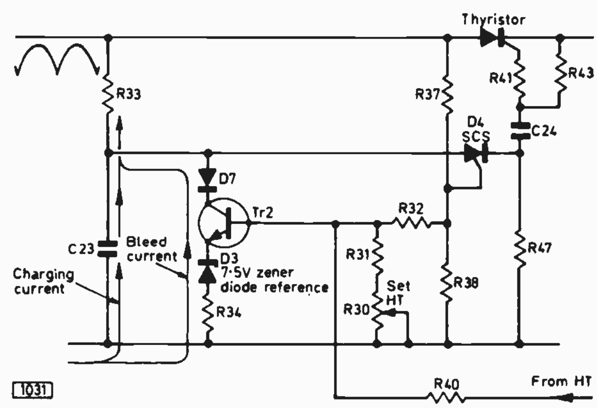
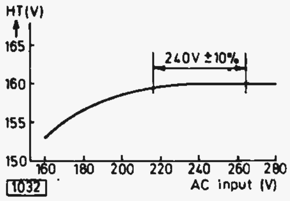


Fig. 13: Careful control of the trigger pulse timing results in stabilisation of the h.t. voltage.

circuit and thus prevents spurious triggering.

Capacitor C22 across the mains input bypasses high-frequency harmonics generated by the thyristor switching action, and reduces the amount fed back into the mains supply. Without such a capacitor interference would be caused to neighbouring TV and audio equipment. This capacitor can ring with the inductance of the mains supply, small though it is, however and the resultant oscillation can cause spurious triggering of the thyristor. This occurs because the large-amplitude, high-frequency oscillation has a very high rate of voltage rise and although this is only applied to the anode of the thyristor and not to the gate it causes the thyristor to conduct again after it has turned off, thus destroying the stabilising action.

Choke L35 prevents such oscillation reaching the thyristor thus overcoming this problem. It also helps prevent r.f. interference getting back into the mains. This type of choke presents two problems incidentally. First, it has to be carefully constructed to avoid 100Hz buzz caused by vibration of its core, bobbin and windings. Secondly it has to be sited and screened so that its magnetic field does not induce interference



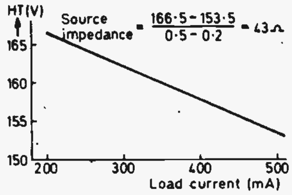


Fig. 14 (left): H.T. stabilising performance of the circuit shown in Fig. 9.

Fig. 15 (right): Internal h.t. regulation.

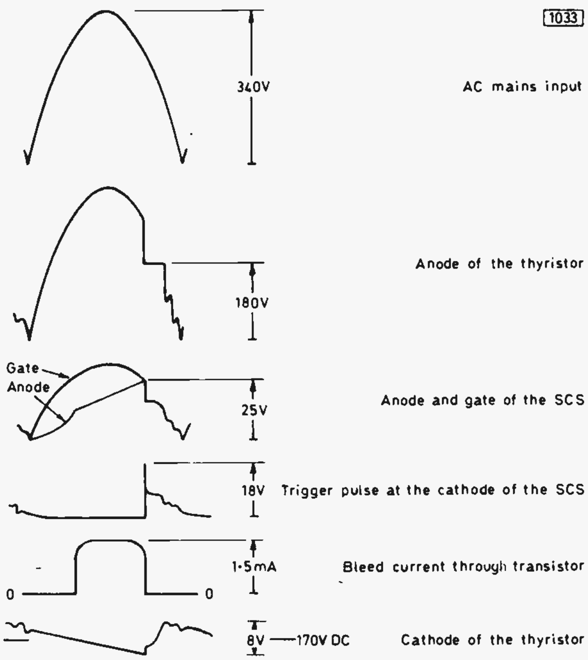


Fig. 16: Typical waveforms with a 240V mains input.

currents in other parts of the receiver, even including the field deflection coils.

The choke and R39 together limit the current flow when the thyristor conducts so that the maximum permitted peak current is not exceeded. This is of the order of 5A peak during normal operation and 25A at switch-on when the thyristor is charging a large electrolytic capacitor (C25).

We next have two safety circuits to consider. First, overvoltage protection must be provided in case the control circuit develops a fault or the thyristor becomes short-circuit. A very ingenious and cheap solution in the form of a glow switch has been designed into this power supply. This is the device used in nearly every fluorescent tube starter circuit to switch off the heater supply. A glow switch consists of a neon bulb with bimetallic strip contacts as the electrodes. In the case of the 320 chassis a glow switch with a striking voltage in the range 193-215V is specified. When the striking voltage is reached the electrodes glow with the colour characteristic of neon gas: the glow generates heat and in less than a second the bimetallic strips bend and the contacts close. When this happens there is a short-circuit between the h.t. line and chassis. The current flow is limited only by R46 while C25 discharges, and by R46 and R39 when current is drawn from the mains. These resistors limit the current flow so that the glow switch contacts do not weld together.

The effect of this short-circuit is to blow the mains input fuse thus isolating the receiver from the mains supply. Note that if R46 has to be replaced it is essential to use the correct component supplied under the appropriate part number listed in the service manual. It has to withstand very high short-circuit currents so substitute components must not be used. Similarly the glow switch is a carefully specified and selected component and the correct replacement type is equally important.

This overvoltage protection not only ensures a high standard of safety but also prevents failure of transistors and i.c.s due to excessive h.t. voltage.

Another safety circuit consists of diode D08 and R35 which are connected to the 32V l.t. line. In this chassis the l.t. line is derived from the line output stage. If a fault which results in the l.t. line falling to zero occurs the emitter of the thyristor control transistor is in effect connected to chassis, turning the transistor on hard so that the thyristor triggering is delayed and the h.t. falls to 50-80V. This ensures that overheating cannot occur anywhere in the receiver circuits—and provides a valuable diagnostic clue. Really low h.t. always indicates a line timebase fault.

Only a few details remain to complete our description of this power supply circuit. The capacitor (if fitted) in parallel with R47, also C36, prevent any stray flashover currents that might find their way into the control circuit causing incorrect triggering and a sudden h.t. voltage surge which might cause unwanted operation of the glow switch crowbar.

Capacitors C28 and C29 across the bridge rectifier suppress diode switching harmonics which could cause interference on the picture.

The mains fuse has to be a 1.6A antisurge type in order to withstand the current surge at the instant of switch on whilst providing reliable fusing action under fault conditions.

Conclusion

Finally a word about the point of connection of the h.t. negative feedback resistor R40. This is connected to the unsmoothed h.t. line to increase the output impedance of the power supply. The source impedance of the unsmoothed supply is about 10 Ω while the smoothed h.t. supply has a source impedance of 10 Ω + R44 = 10 + 33 = 43 Ω which just about complies with the specification mentioned earlier. If the negative feedback had been taken from the top of C26 the output impedance would have been only a little more than 10 Ω. The impedance seen by the line timebase is equal to 43 Ω + R65 (the flashover protection resistor—see Fig. 3) = 43 Ω + 56 Ω = approximately 100 Ω. This meets the requirement to reduce picture breathing mentioned in Part 1.

The aim of this detailed description of a typical modern TV power supply circuit has been to provide not only an understanding of how it works but also an insight into the engineering problems that arise and how they are overcome. It explains some of the features that have to be designed into the circuitry in order to comply with the basic specification that is the starting point of the whole activity. Some typical waveforms and performance figures are included to show the final results. ■

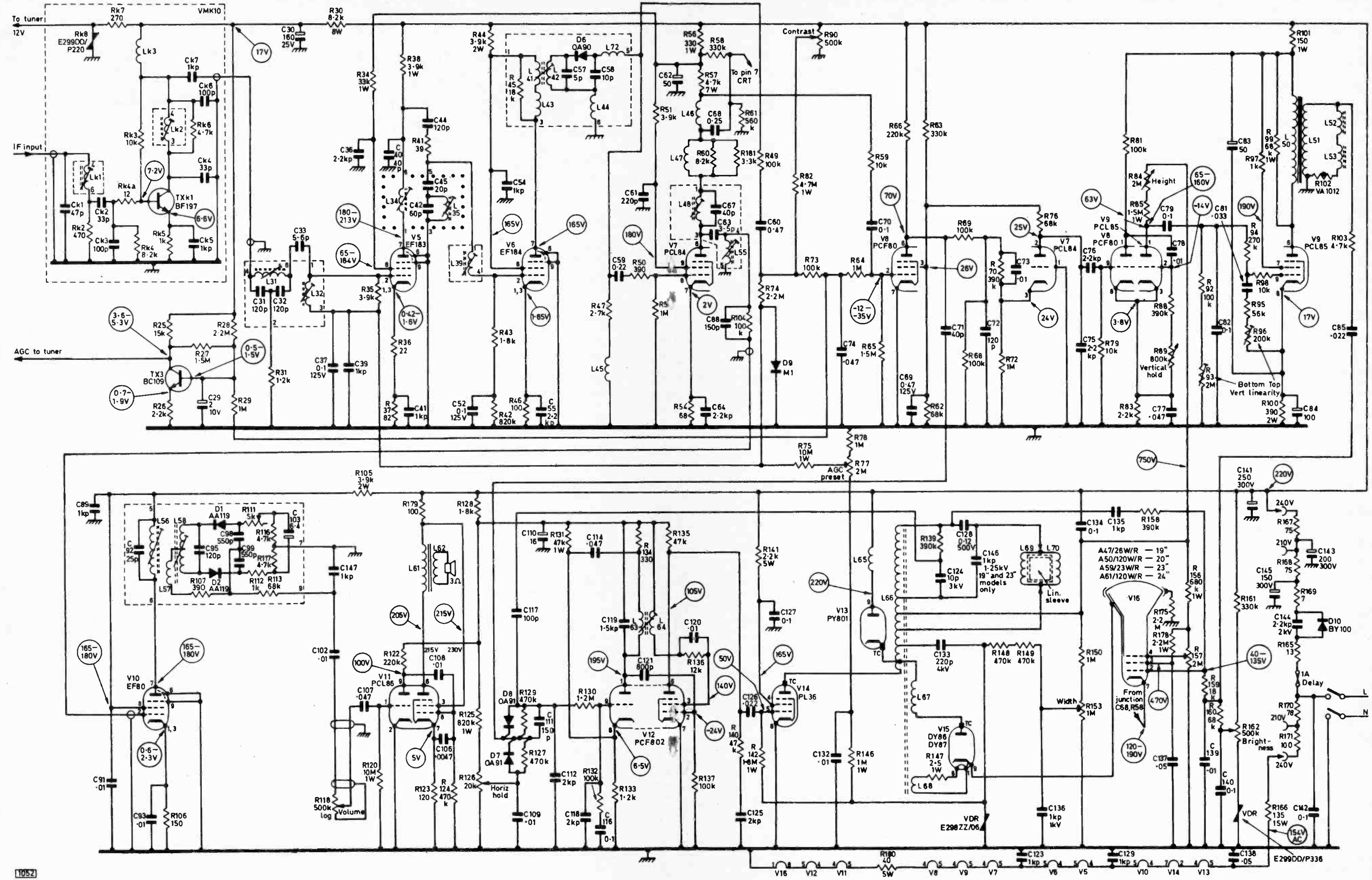


Fig. 2: Circuit diagram of the ITT-KB VC100 chassis. Voltages measured using a 20kΩ/V meter (Avo Model 8).

easily removed and the main check points exposed. First check the PCL84 voltages at pins 6 (varies according to signal, from 120V up), 9 which should be about 180V if R51 and the PCL84 are in order, and 7 (cathode) about 2V. If these voltages are about right go back to the EF184 where if the cathode voltage at pins 1 and 3 is about right (1·5-2V) the supply must be present at pins 7 and 8. Remember that these are rough checks only and do not take tuning and coils etc. into consideration.

If the EF184 stage is apparently right check the EF183 voltages, taking particular note of the screen grid voltage at pin 8. If this is low—about 65V—the stage is passing current and little a.g.c. is being applied. If the voltage is high, say 180V, a large amount of a.g.c. is being applied to the control grid or the valve has lost emission: a voltage check at pin 2 will then reveal either a hefty negative voltage (say −10V or so) or no voltage, the latter implying that as the screen voltage is high the valve has lost emission. Check the value of R34 in either case. It is unlikely that there will be a heavy negative voltage at pin 2 if there is no vision signal.

If all this merely confirms that these stages are in order, and so far it has taken only a few minutes to check these voltages, move back to the preamplifier panel and check the BF197 transistor voltages. About 16V should be present at the collector, 7·2V at the base and 6·6V at the emitter. If the latter voltage is absent but the base voltage is somewhere near right the transistor is passing no current and should be checked with an ohmmeter to prove that there is no low reading with the "positive" probe to the emitter and the "negative" probe to the base. If this stage is also in order some sort of signal should be received—even if only due to the meter leads—and this situation would indicate that the tuner or the a.g.c. to the tuner (or of course the supply voltage via Rk7) is at fault.

As far as the tuner is concerned servicing must depend upon the experience of the repairer and we can only say that it follows normal practice except for the mechanical operation where the moving parts seem to be rather brittle and inclined to fracture.

NEXT MONTH: BAIRD 660 SERIES

LETTERS

BRC 3000/3500 CHASSIS

I have had considerable experience of the Thorn/BRC 3000/3500 chassis power supply module and would like to add the following points to the article by Paul Soanes in the September issue.

Cut-out trip blowing: Check the print around the crowbar trip W621—electrolyte leakage from capacitors can be present resulting in false triggering. Check for shorting between the chopper transistor case and chassis (mica washer breaking down) and between the lead to W609 and the chopper transistor fixing screw.

No results: If due to the 30V rail being low this can be caused by C609 being leaky.

Low 58-65V rail: If VT604 collector voltage is below 240V C606 may have lost capacitance.

Intermittent failure to start or spontaneous shut down: The trouble is generally in the area VT601/VT602. The simplest course is to replace C607, W605, VT602 and if necessary VT601—also check that R606 is 1Ω (in earlier versions it was 2Ω or 3Ω and the power unit may then shut down if the mains voltage falls below 220V). Other possible causes: dry-joints around VT605; faulty plug and socket connections—particularly the long multiway connector using a piece of printed circuit as the link; dry-joints around F603 fuseholder.

Squegging: In addition to C619 and C631 check C609, C607 and the setting of R622.

Picture flutter: Can be caused by VT601 being faulty.

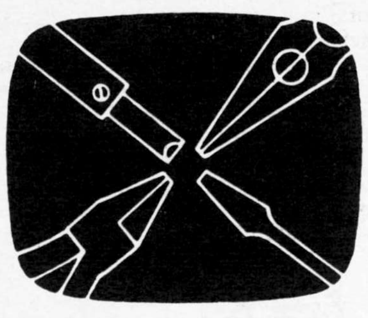
In renovating panels that have been in service for several years I suggest replacing W605, VT602, all the electrolytics and the presets R629, R631 and R622. Check the tightness of the screw clamping VT601 to chassis and the tension of the fuseholder springs. Set R622 and R631 after replacement as follows: Connect a 100Ω 30W resistor from the output side of F603 to chassis; connect a d.c. meter capable of reading 2·35A in series with F603 and a voltmeter (100V f.s.d.) across the 100Ω resistor. Switch on and adjust the receiver under test card conditions with the brightness and contrast controls at maximum setting. First turn R622 and R629 fully anticlockwise and R631 fully clockwise. Then set R631 for a reading of 65V on the voltmeter and seal it. Adjust R622 until the meter reading drops just below 65V, checking that the current reading is about 2·35A. Seal R622 after this.

The small overwinding on the mains transformer 6·3V heater winding can sometimes be used to boost an ageing c.r.t. To do this remove the two brown and one white lead connected together to the transformer and solder them to the adjacent empty tag instead.

Finally, when showing customers how to reset the cut-out trip it is important to make it clear that the set must first be switched off.—**Barry F. Pamplin** (*Preston*).

TRICKY HEATER CHAIN FAULT

The headaches that quite simple circuits can cause are illustrated by the following example. A Pye Model 36 turned up with the BY114 heater rectifier short-circuit thus passing a heavy a.c. through the heater chain. The usual remedy—replacing the rectifier—was carried out but after the set had been on for about two minutes a crackling sound could be heard from the loudspeaker indicating that the rectifier was going again. This was followed by heavy a.c. through the heater chain. In this chassis the l.t. supplies for the transistor stages are derived from the earthy end of the heater chain so the reservoir and smoothing electrolytics here were tested. We then tested the insulation of the heater chain at 200V—with the c.r.t. heater disconnected. All seemed to be in order. The first valve in the heater chain in this chassis is the PL500 line output valve and we came to the conclusion that this was developing heater-cathode leakage after warming up. This would place a heavy load on the rectifier of course. The line output valve was replaced and all other valves tested and the fault was cleared. When making the final test a 100Ω 10W resistor was wired in series with the surge limiting thermistor in case another rectifier went short-circuit. The circuit is straightforward enough but as can be seen tracing the cause of the fault proved quite troublesome.—**G. T. Jones** (*Pwllheli*).



SERVICING television receivers

L. LAWRY-JOHNS

ITT-KB VC100 CHASSIS—cont.

Line Oscillator

The line oscillator consists of the well proved PCF802 with a flywheel line sync discriminator using a pair of OA91 diodes (D7, D8). This is fairly reliable in operation but when line hold troubles do present themselves first prove the PCF802 by replacement and then check the diodes and associated resistors. If the fault persists change capacitors C119 and C121 (changing both can save a lot of time). Also ensure that windings L63-L64 have not moved down their former—it will be appreciated that movement of the coils down the former is the same in effect as unscrewing the core.

The job of C110 should also be appreciated since it not only decouples the line oscillator supply but also the supply to the audio amplifier stage. Thus a severe ripple (horizontal) of the picture accompanied by an uncomfortable effect on the sound (the whistle may not actually be heard) should draw attention to the capacitor block C62-C110.

Non-operation of the line oscillator will of course result in no picture with the PL36 and PY801 overheating. When the PL36 anode is glowing cherry red the first suspect must be the line oscillator (change the PCF802 valve) and its associated components. Whenever the PCF802 is changed always give the other valves plenty of time to cool down—otherwise the same symptoms will present themselves, with the probable loss of the PY801. If the PL36 is found with a cracked envelope it is probable that it has been subjected to overheating due to lack of drive.

Line Output Stage

Some degree of overheating—where the PL36 anode is seen to be overheated—need not be caused by lack of drive and a meter check at its control grid may show approximately the correct 50V negative swing. In this event the effect of removing the top cap of the PY801 should be noted. If the stage then comes to life the boost capacitor C134 is shorted (there are no windings on the line output transformer connected to the h.t. line). A replacement should be rated at 1kV.

If removing the PY801 top cap has no effect put it back on and take the top cap off the DY87 (or remove the e.h.t. cap from the side of the c.r.t.). If the stage now starts working suspect an internal short in the DY87. Otherwise check C146 (if fitted) by disconnecting it.

Resistor R141 can become open-circuit and this of course stops the output stage working altogether. If there is no sign of e.h.t. and no spark at the PL36 top cap although the d.c. voltage here is well over 200V the resistor is most likely to be at fault. It is a 2·2kΩ, 5W type.

One common trouble is lack of width. This can well be due to a low-emission PL36 (or PY801) but if these are in order attention should be directed to the 1MΩ width control R153, its feed resistor R150 and R148, R149 connected in series to the slider. Of these R150 is the more likely to be affected, or the width control could have a dud spot. We have never found R142 to be of wrong value but there is always a first time. This could also be said about the v.d.r. which is often blamed but rarely at fault. One other possibility however is leakage in the scan-correction capacitor (C128).

The line output transformer can be faulty but there is not much mystery about this since the most common trouble is breakdown of insulation and this makes itself known in no uncertain manner.

Dark Picture

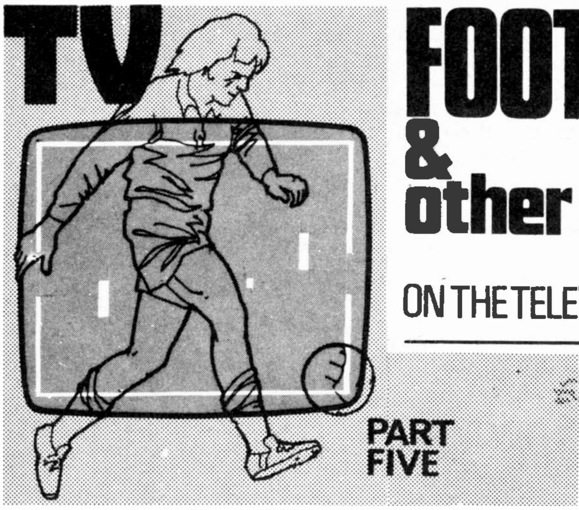
It will be seen that the first anode supply to pin 3 of the c.r.t. base is derived from the boost line via R178 and R156. The focus control is connected between the junction of these resistors and chassis. While the control remains at its correct value of 2MΩ all is well. When the resistance falls however the voltage available at pin 3 is progressively reduced resulting in a darkening picture until the brilliance control can no longer provide compensation. Whilst a similar condition could be due to leakage in C137 this does not happen very often due to the comfortable voltage rating of the capacitor. If R161 or R61 increases in value there will also be too little brilliance while if R58 increases in value the picture will be too bright.

Fault Symptoms & Diagnosis

Since the 6MHz intercarrier sound is tapped from the anode of the video amplifier any fault affecting the tuner, preamplifier, i.f. stages, detector, video amplifier or a.g.c. must affect the sound as well as the vision signals. The other side of the coin of course is that if the vision signal is good but the sound is not these stages cannot be at fault and the field is narrowed to the sound i.f. stage, the ratio detector or the audio section.

When tracing the source of a fault causing loss of vision and sound signals the golden rule as ever is to check the voltages first and to get them somewhere near right.

This is not at all difficult since the bottom cover is



TV FOOTBALL & other GAMES ON THE TELEVISION SCREEN

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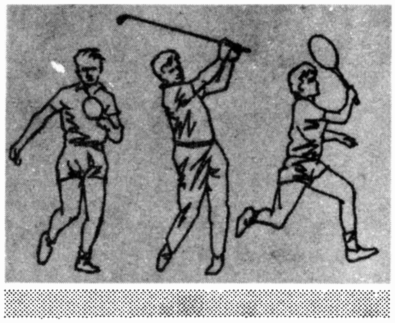


PART FIVE

THIS month the ball motivation circuitry for the football game is described. The movement is produced by two integrators controlling the velocity in the X and Y directions as was described for the simple game.

Ball Control

Fig. 21 is a block diagram of the ball control circuitry for one axis only. Movement of the joystick is continually monitored by a differentiator circuit which produces a voltage proportional to this movement. When a player intercepts the ball, a pulse which activates the sample and hold circuit is produced and a voltage proportional to the "kick" is stored. This voltage passes into a circuit which reverses its polarity when the ball reaches the boundaries. Finally the "kick" voltage is used to programme the ball velocity integrator. If the sample and hold circuit is deliberately made inefficient the "kick" voltage will decay giving a realistic slowing-down effect to the ball speed.

You may be puzzled by a block marked "full-wave rectifier" between the sample hold and polarity reverse functions in Fig. 21. This is required for the polarity reverse circuit which can only handle positive signals at its input. A second output from the full-wave rectifier carries a logic signal indicating the polarity at the input. This is used to set the polarity reverse circuit initially for no overall polarity change. We will see later that this arrangement allows us to use a simple latch circuit for the boundary reverse functions instead of the bistable arrangement used in the tennis game.

The circuitry used to carry out the analogue ball control functions is shown in Fig. 22. To start with however we will give some basic data which may be of assistance to TV orientated readers on the use of the operational amplifier. In Fig. 23 the operational amplifier is shown used as a simple inverting amplifier at (a) and as a non-inverting amplifier at (b). The formulae for the gains are very simple: the basic formula for the inverting amplifier is modified for the non-inverting case by adding unity. Fig. 23(c) shows how these two formulae can be compounded for a more complex case.

Movement Detectors

Returning to Fig. 22, IC302 together with R302, R306 and C302 form a precision differentiator with a time-constant of about 1·5S. The output from IC302 feeds the sample and hold circuit comprising the f.e.t. switch Tr301, storage capacitor C309 and the high-impedance source-follower Tr305.

Tr301 is normally off since −15V is applied to its gate electrode via R315. When a coincidence between the left man and the ball occurs a +5V logic pulse appears at E18 and is amplified by Tr309, Tr310 to give a +15V pulse which switches Tr301 on. Note that the supply

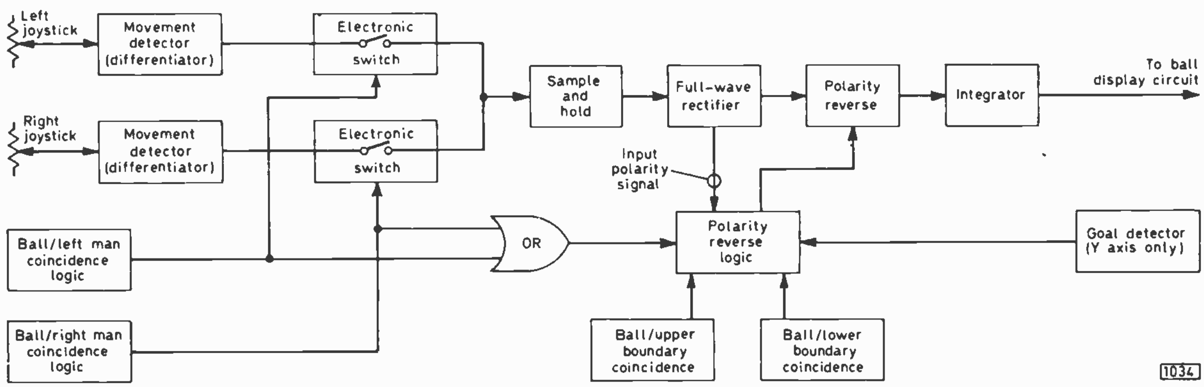


Fig.21: Block diagram of the football game ball control system. The circuitry is repeated for the other display axis.

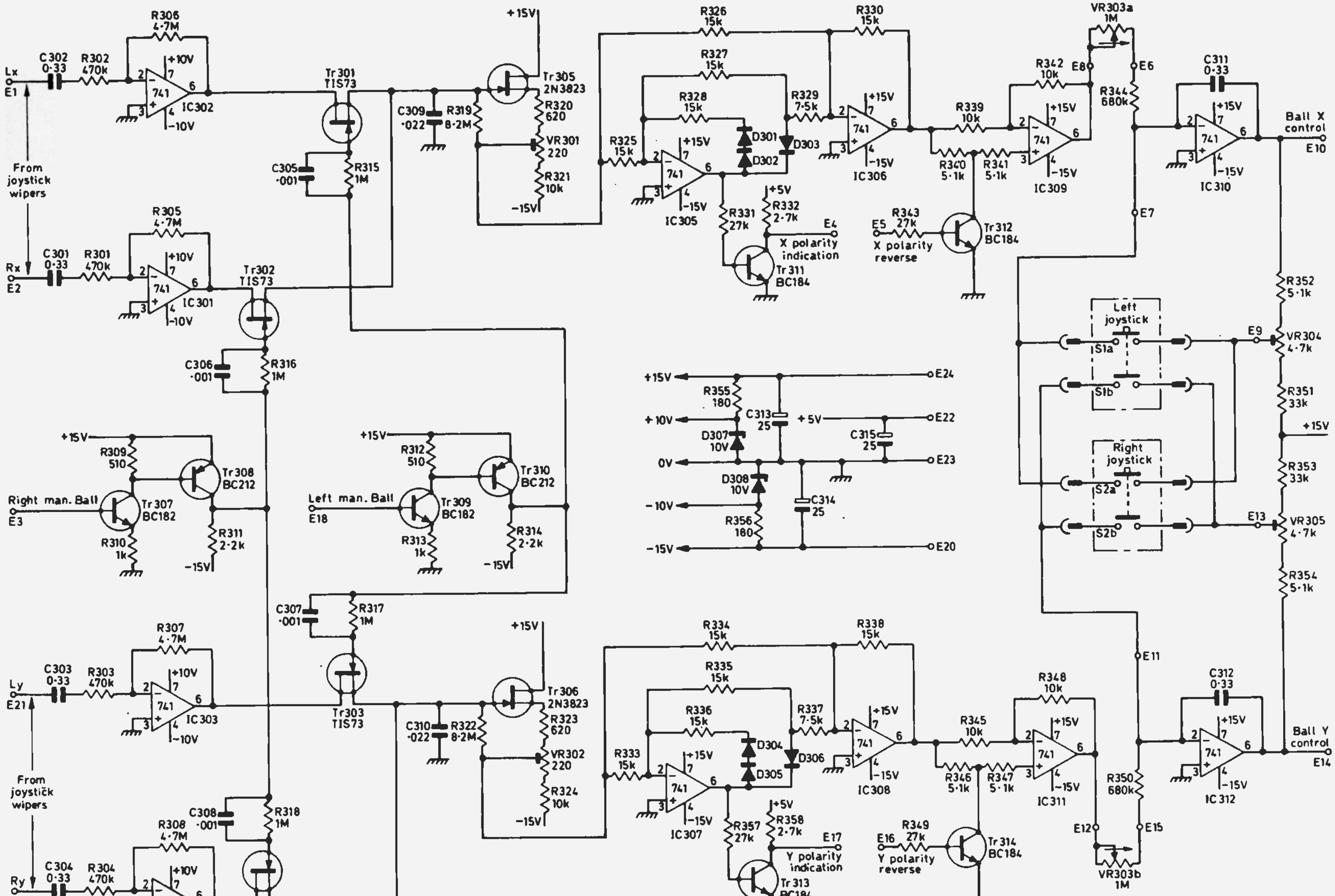


Fig. 22: Circuit diagram of the analogue (ball control) board.

Notes. 1. D301—D306 are 1N914 or 1N4148
2. VR303 is the game speed control

rails to IC302 are restricted to ±10V to prevent excessive negative signals turning Tr301 on. IC301 and Tr302 form an identical differentiator and switch circuit for the right-hand joystick control. The f.e.t. source-follower Tr305 is of conventional design; preset VR301 enables the offset voltage to be set to zero.

Full-wave Rectifier

The operational amplifiers IC305, IC306 plus their associated resistors and diodes form the full-wave rectifier. A positive input signal of +V into R325, R326 causes the output of IC305 to go negative. Diode D303 will then conduct and feedback resistor R327 will ensure that a voltage of −V is present at the junction of R327 and D303. The second amplifier IC306 will sum the two inputs +V applied with unity gain via R326 and −V with a gain of two via R329. The net output from IC306 is +V. Transistor Tr311 will be switched off giving logical 1 (+5V) at E4.

A negative input signal of −V will give a positive output from IC305 causing diodes D301, D302 to conduct. The transistor Tr311 will be switched on giving logical 0 at E4. As D303 is reverse biased there will be no input to IC306 via R329 as R327 is connected to a virtual earth at the input to IC305. Overall there will be a simple inversion via R326, R330 giving +V at the output of IC306 as required.

Polarity Reversing Circuit

The polarity reversing circuit comprises IC309, Tr312 and resistors R339–R343. A logical 0 signal (0V) at E5 causes Tr312 to be off. The input signal of +V will then be applied to both inputs of IC309 giving a composite output of +2V−V=+V. With E5 at logical one Tr312 will be on, grounding the non-inverting input, and the amplifier will be operating in the normal inverting mode giving an output −V.

Integrator

The integrator IC310 is exactly the same as we used for the simple game. The pushbuttons serve to centre the ball at the start of play by temporarily connecting the i.c. as an inverting amplifier with an input from the +15V rail.

Control in Y Direction

The lower half of the circuit controls the ball velocity in the Y direction. It is identical to the X control described.

Man/Ball Coincidence

Before embarking on a description of the ball control logic we must look closely at the man/ball coincidence requirements. Readers who constructed the simple game outlined in the third article may have discovered that if the ball speed is increased too much the ball can pass straight through one of the men. The reason for this is that the coincidence gating of the groups of pulses representing the ball and player occurs at field frequency. Thus if the ball takes one second to traverse the screen it will be represented by 50 spots of which one may or may not coincide with a player. Fig. 24(a) shows how a ball traversing the screen in 1/5th of a second (10 spots) can leapfrog a player.

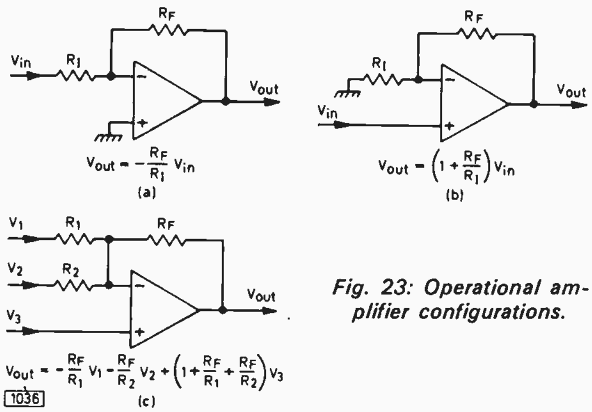


Fig. 23: Operational amplifier configurations.

A common solution to this problem—Fig. 24(b)—is simply to enlarge the ball and players. We will use the term interaction cross-section to describe loosely this phenomenon—i.e. the solution shown in Fig. 24(b) is to increase the interaction cross-section. Readers may have noticed that most commercial TV games adopt this solution—hence the rather clumsy appearance of the players and boundary lines.

The kicking action of the players, used to motivate the ball in this game, means that the player is moving as well as the ball. Thus the leapfrogging effect will be much worse and a superior method of coincidence recognition must be employed. The solution adopted is to AND gate the line-related and field-related pulses separately. Fig. 24(c) shows how the moving ball pulses can intercept the stationary man pulse in a single field period.

This method gives a gain in the coincidence recognition efficiency of 312 times and makes it possible to use very small interaction cross-sections. This improvement is only in the horizontal direction of course. Fortunately for two reasons this doesn't matter. First, the direction of play is in the horizontal direction; secondly, the vertical interaction cross-sections of the players are large due to their natural shape.

Reflection of the ball from the boundaries requires exactly the same coincidence recognition technique as a ball/man interception. For this reason the effective interaction cross-section of the upper and lower touch

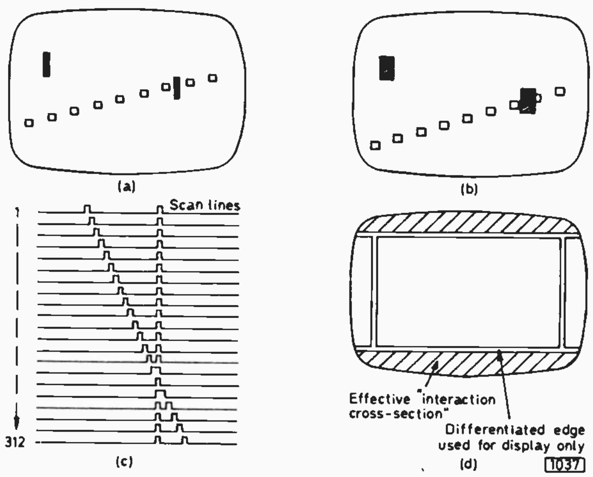
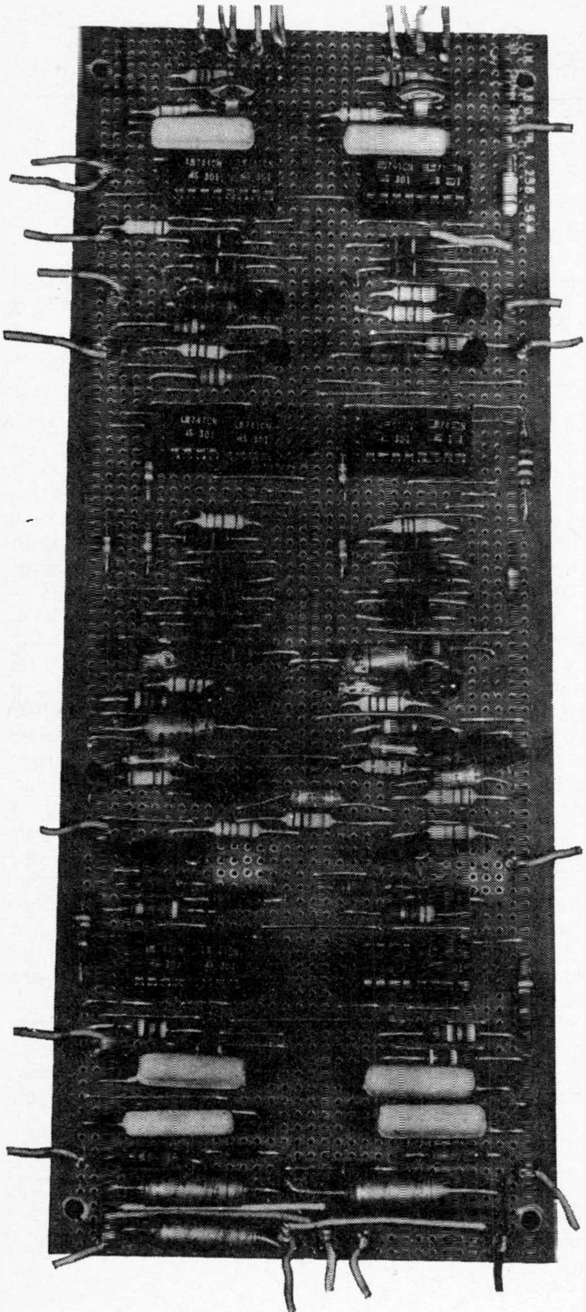


Fig. 24: Man/ball and ball/boundary coincidence conditions.



Photograph of the analogue board.

lines extends to the edge of the screen—see Fig. 24(d) —whilst the differentiated edge is used for the visual display. You may recall that provision was made for this when we dealt with the boundary lines in the previous article.

Coincidence Logic

In the simple logic used in the tennis game—see Fig. 25(a)—the line-related and field-related ball pulses (ball X and ball Y respectively) are AND gated (gate A) to give the display waveform. Similarly, man X and man Y are AND gated in gate B to give a man display. Coincidence is detected by gating the two display signals with gate C. The more elaborate man/ball coincidence logic is shown in Fig. 25(b). Gates A and B are still required for the display. Gate D detects a coincidence between ball X and man X (the line related pulses) however whilst gate G does likewise for ball Y and man Y. The ball X/man X coincidence can occur at any time during a given field period whilst the ball Y/man Y occurs at a specific point in the field period. We must detect therefore when both occur in a given field period. To this end the latch formed by gates H and J is used to store a man Y/ball Y coincidence.

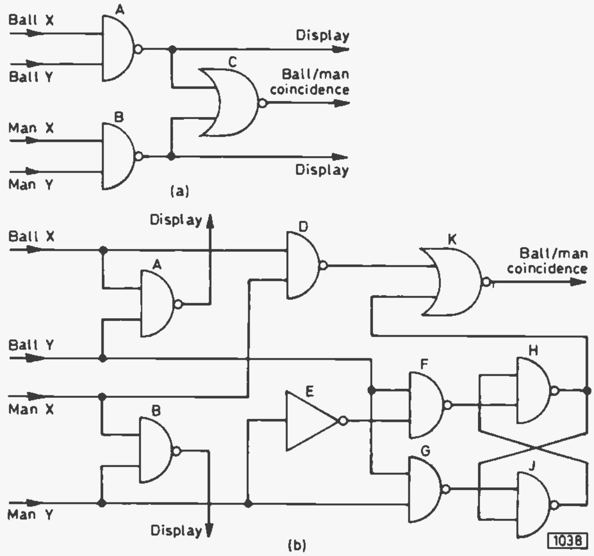


Fig. 25: Man/ball coincidence logic, (a) for simple game, (b) for football game.

The overall operation of the circuit shown in Fig. 25(b) is as follows. If say a man Y/ball Y coincidence occurs halfway through a field simultaneous high inputs to gate G cause its output to go low. This in turn causes the output of J to go high and H to go low, where they remain until in successive field periods the man and ball move out of horizontal coincidence. A non-coincidence will be detected by gate F; its output will go low, changing the state of the latch to H output high, J output low.

Coincidence in a vertical plane (man X/ball X) will be detected at some point in each field by gate D; simultaneous high inputs to D will produce a low output at coincidence. A double coincidence of man X/ ball X and man Y/ball Y is detected by two simultaneous low inputs to gate K, giving a high output.

The full logic circuit will be described in out next issue.

Wiring and Testing

The analogue circuit (Fig. 22) was constructed on strip board as shown in the accompanying photograph. Testing should be fairly straightforward as it can be accomplished without connecting up to the rest of the circuitry.

Connect the differentiator inputs to the relevant joystick wipers and check their operation at the outputs

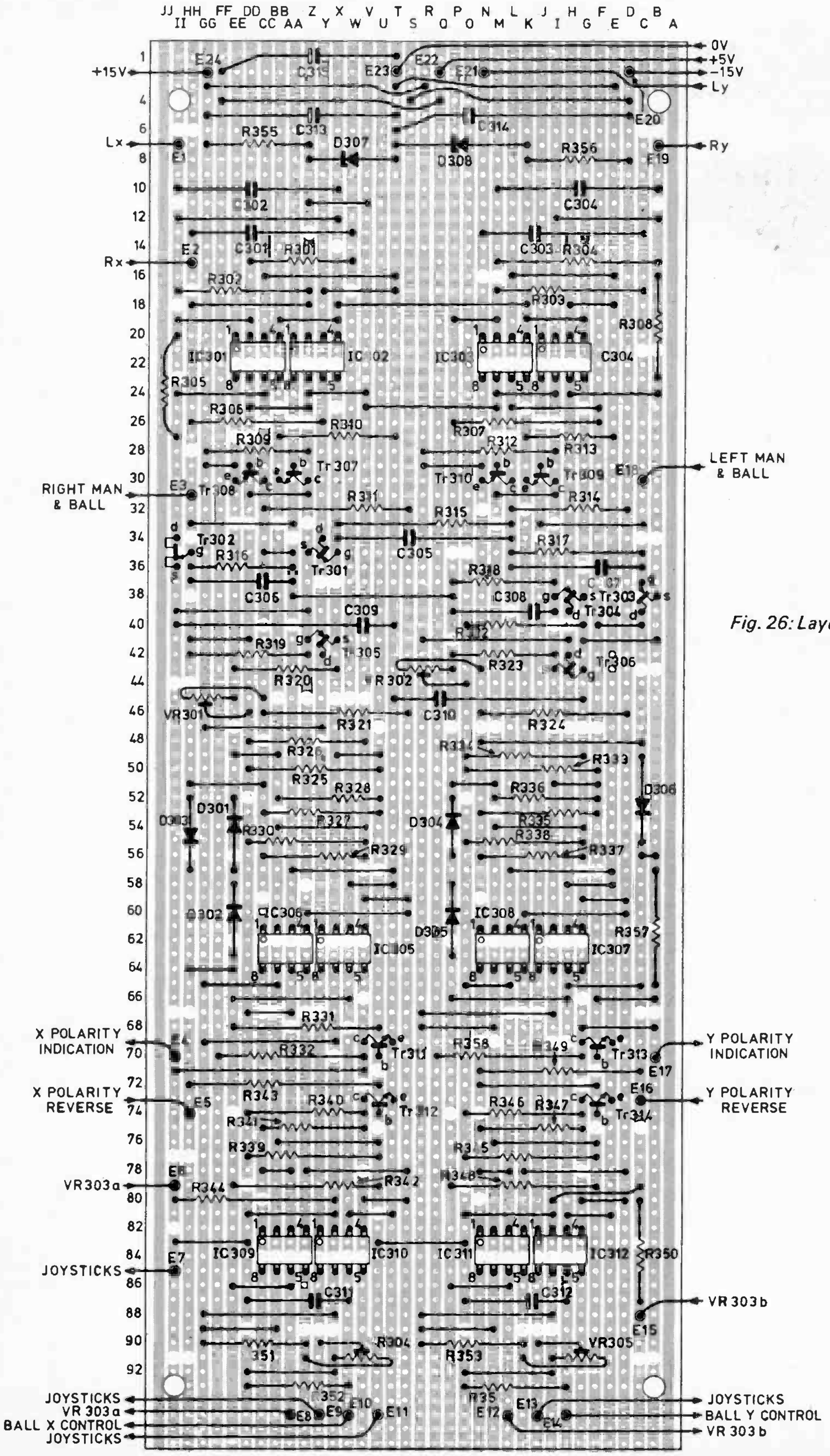


Fig. 26: Layout of the analogue board.

★ Components list

ANALOGUE BOARD 'E'

Resistors: (all ±5%, ¼W)

R355, R356	180 Ω	R325-R328,	
R309, R312	510 Ω	R330, R333-	
R320, R323	620 Ω	R336, R338	15k Ω
R310, R313	1k Ω	R331, R357,	
R311, R314	2·2k Ω	R343, R349	27k Ω
R332, R358	2·7k Ω	R351, R353	33k Ω
R340, R341, R346,		R301-R304	470k Ω
R347, R352, R354	5·1k Ω	R344, R350	680k Ω
R329, R337	7·5k Ω	R315-R318	1M Ω
R321, R324, R339,		R305-R308	4·7M Ω
R342, R345, R348	10k Ω	R319, R322	8·2M Ω

Potentiometers:

VR301, VR302	220 Ω miniature carbon presets
VR304, VR305	4·7k Ω miniature carbon presets
VR303	1M Ω+1M Ω dual-ganged

Capacitors:

C305-C308	1nF	C315	25μF 10V
C309, C310	22nF	C313, C314	25μF 25V
C301-C304, C311, C312	0·33μF		

Semiconductors:

Tr301-Tr304	TIS73	Tr311-Tr314	BC184
Tr305, Tr306	2N3823	Tr308, Tr310	BC212
Tr307, Tr309	BC182		
D301-D306	1N4148 (1N914)		
D307, D308	BZY88 C10 400mW		
IC301-IC312	741P 8-pin DIL		

of IC301–IC302. Momentarily connect E18 or E3 to +5V and check the action of the sample and hold circuits at the wiper of VR301. If C309 is increased to about 0·22μF the sample voltage should be held for several seconds. Set VR301 to give zero volts output when the sample voltage has decayed away.

To check the remaining circuitry it may be easiest to connect a variable voltage of between ±10V to Tr305 gate. Check the action of the full-wave rectifier at the output of IC306, also the logic output at E4 (positive input E4 high, negative input E4 low). The polarity reverser can be tested by applying +5V or ground to E5 and monitoring the output of IC309. If E5 is connected to the output of IC305 the output of IC309 should follow the input voltage. Repeat the checks for the identical lower half of the circuit.

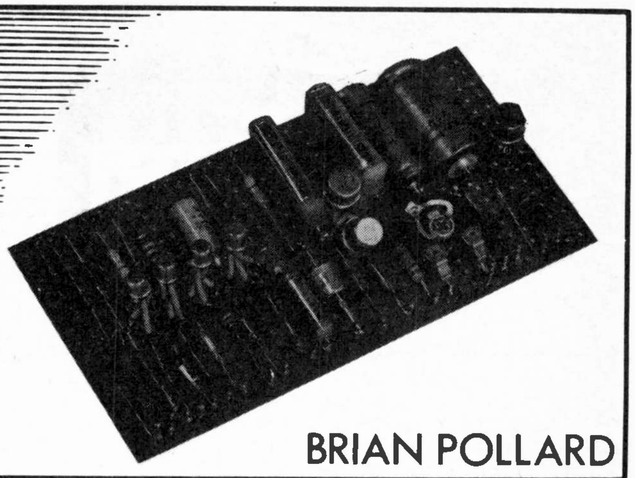
Next Month

In the next article we will after completing the football game describe some sound effects circuits for both games. These give a realistic ping whose pitch depends on how hard the ball is hit. We will also be giving details of improvements to the ramp generators to give increased stability of operation.

Correction

The earth connection to IC3 in the power supply circuit (Fig. 2, July) was unfortunately omitted. Pin 7 should be connected to chassis.

BUILD A BLACK-LEVEL CLAMP



BRIAN POLLARD

THE black-level clamp described in this article was devised for use in a receiver fitted with the Thorn 950 Mk II chassis: it will work in any valve monochrome receiver however with little or no modification to the circuit given.

Need for Clamping

As its name indicates the purpose of a black-level clamp is to hold the black level of a video signal at a constant level irrespective of changes in the content of the video signal. The problem arises when the video signal is a.c. coupled. A simple method of maintaining the black level of the signal with a.c. coupling is to use a d.c. restorer which clamps the sync pulse tips at a constant level, see Fig. 1. A d.c. restorer thus holds the black level constant irrespective of changes in picture content but the black level will still vary with overall video signal amplitude. The answer to this is to clamp the black level itself so that it is held at a constant level irrespective of picture content *and* signal amplitude.

Clamping is essential with colour receivers of course but this feature has been almost wholly absent from monochrome sets—it is simpler to use a.c. coupling and rely on the picture generally being at a mean level. In my view however the extra expense involved in adding a black-level clamp is well worthwhile: dark pictures look dark instead of like a bright and often streaky fog and the dark detail in bright pictures does not become lost in shadow. Furthermore the general contrast level for the same video drive is improved since once adjusted the brightness control is always at its optimum setting. These features also make the clamp very useful for DX-TV purposes since the brightness level can be set to the optimum level required for photography, independent of the state of the received signal.

Circuit Principle

The basic action of this clamp circuit is shown in Fig. 2: the heart of the circuit is transistor Q1. This transistor is non-conducting during the line period and as the c.r.t. cathode is more negative than the black level D1 is reversed biased. The c.r.t. is drawing current however so that C1 charges—the potential at its right-hand side as a result becoming steadily more positive with respect to that at its left-hand side. Following the line sync pulse the signal sits at black level until the beginning of the next line of picture information, but the cathode of the c.r.t. will be at a potential which is positive with respect to the reference black level. At this point a positive pulse from the line output stage turns Q1 hard on. D1 which had been forward biased by the positive-going sync pulse remains forward biased due to the extra charge on C1. As a result D1 and Q1 conduct heavily, discharging C1 until the c.r.t. cathode is at the reference black-level potential ready for the arrival of the next line of picture information.

Due to the time-constants of the couplings in the video circuits preceding the clamp the mean level of the input video signal will vary up and down—at a rate dependent on the shortest time-constant. If the mean level went very negative so that the c.r.t. was completely cut off D1 would be reverse biased and

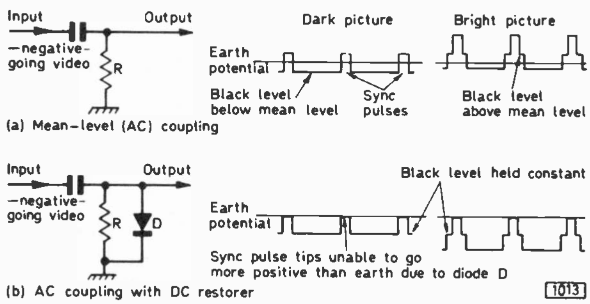


Fig. 1: Action of a d.c. restorer.

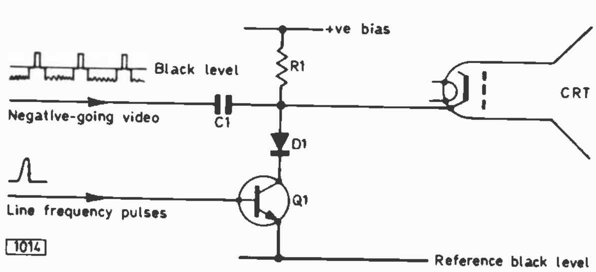


Fig. 2: Basic circuit of the driven clamp.

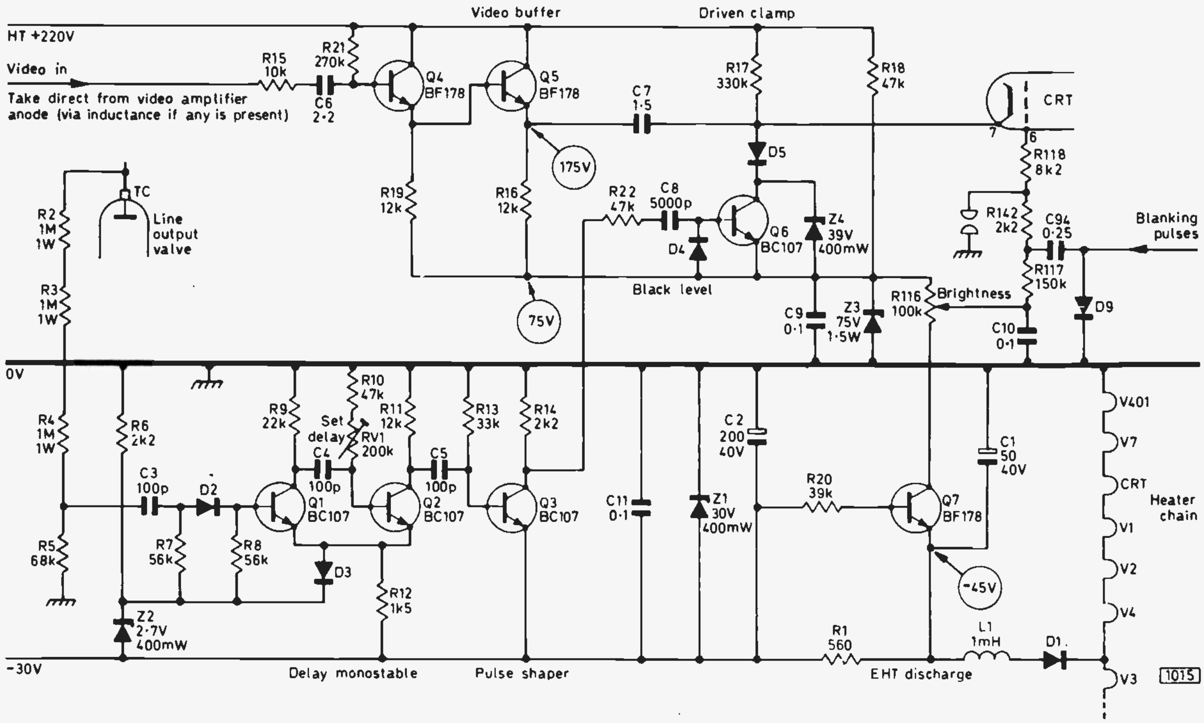


Fig. 3: Complete circuit of the black-level clamp, with video buffer and drive pulse generator.

there would be no current flow through C1. To ensure that C1 can still charge and the black-level clamp operate under these conditions R1 is required.

The time-constant of R1, C1 must be shorter than the shortest time-constant in the preceding video stages. These are normally quite high anyway (around 100-500mS). Now the clamp appears at first sight to have a time-constant of 500mS (C7, R17 in Fig. 3) but the aiming potential of C7 is the h.t. line at +200V compared with a mean signal level of around +30V to +60V. Thus the effective time-constant is more like 100mS. There is an advantage to be gained in having a short time-constant: if it is appreciably less than 20mS the clamp will effectively remove any mains hum on the video signal. Such hum does tend to be more noticeable when the picture is uniformly dark. A short time-constant also produces horizontal streaking on the picture if there is a ghost present however as the black level at the end of the line sync pulses has superimposed on it the displaced video signal of the ghost. Thus if the clamp has too short a time-constant it is able to follow these unwanted variations in black level. Note that as long as the time-constant C1, R1 (Fig. 2) is reasonably large compared with the line period C1 will charge only a fraction of a volt during the line period. There will be no noticeable change in brightness from left to right across the picture therefore due to C1 charging.

So much for the basic thinking behind the clamp circuit. We will now describe the operation of the full circuit shown in Fig. 3 (the external circuitry shown is for the Thorn 950 Mk II chassis).

Power Supply

First the power supply. This is derived from a point in the heater chain at about 40V a.c. (junction V3/V4). The sequence of valve heaters shown in Fig. 3 is for the Thorn 950 Mk II chassis with flywheel sync. The current drain of the supply is 30mA which drops the voltage at the earthy end of the heater chain by around 7% and raises the voltage of the rest of the heaters by around 2%. This does not seem to have any bad effects, and after all the input voltage taps on the mains transformer each cover a range of 20V or 10%. Unfortunately if one of the valves at the bottom end of the chain is removed—or if its heater goes open-circuit—the full heater supply voltage appears at D1 cathode. Since D1 is a signal diode the overload rapidly destroys it, thus protecting the rest of the circuit. The inductance L1 may not be necessary. Without it however I had trouble with interference from the rectifier being picked up by V4 (vision i.f. amplifier) via its heater-cathode capacitance—the usual capacitor across the rectifier having no effect. The voltage drop across R1 is used to turn on Q7 which "earths" the negative end of the brightness control—when the receiver is switched off Q7 turns off almost immediately and the c.r.t. control grid potential rises so that the e.h.t. capacitance is efficiently discharged.

Drive Pulse Generation

The line-frequency pulses used to drive the clamp must be accurately timed to coincide with the line sync pulse back porch. In this circuit they are derived from the line output valve anode via the potential divider R2, R3, R4 and R5. Three 1MΩ resistors should be used rather than one 3MΩ resistor in order to keep the potential developed across the resistors to a safe value —remember that there are pulses of several kilovolts amplitude at the line output valve anode. The positive pulses thus obtained are fed via C3 to an emitter-coupled monostable circuit consisting of Q1 and Q2. The monostable triggers very early on during the flyback pulse, turning Q2 off. Q2 turns on again after a time determined by C4 and R10 with RV1, so turning Q3 off for a time determined by C5 and R13. In this way a positive pulse whose delay relative to the line flyback pulse is adjustable by RV1 appears at Q3 collector. With C4 100pF as shown the delay is suitable for 625-line operation. To convert the circuit for 405-line operation simply requires C4 to be increased to 150pF.

Clamp Action

The positive pulses at Q3 collector are d.c. restored by D4 and used to switch on the clamp transistor Q6 as described earlier. To protect Q6 in the event of loss of drive or unduly large video transients Z4 prevents Q6 collector going more than 39V positive with respect to its emitter which is at the reference black level. If the drive is removed from Q6 base Z4 simply acts as a d.c. restorer in conjunction with C7 and R17, clamping the sync pulse tips to 39V above reference black level.

Buffer Circuit

The cathode current in a monochrome c.r.t. can be anything up to 500μA which means that with a peak white raster Q6 during the 2μS or so that it is turned on has to discharge C7 at a current of 500μA × line period/2μS—approximately 16mA—in order to remove the charge built up during the line period. The video output stage typically has an output impedance of around 3k Ω, so to supply 16mA would require a voltage drop of around 50V which is clearly not practicable. To reduce this worst case voltage drop to a fraction of a volt Q4 and Q5 are used. The output impedance at Q5 emitter is approximately $Rin/(hfe)^2$. Here *R*in is about 16k Ω (R15 plus the impedance of the video output stage) and *h*fe is typically 20. Thus we now have an output impedance of around 40 Ω which produces a worst case voltage drop of about 650mV—in other words the black level is depressed by 650mV on a peak white section of picture compared with a black section of picture. This is small enough to avoid annoying horizontal streaks where the picture contains areas which are significantly brighter or darker than the rest of the picture. Although R15 appears to serve no useful purpose it is in fact present in order to prevent distortion to the sync pulses. This could occur otherwise due to the non-linear input impedance of Q4 at low values of *V*ce. It does not reduce the h.f. response of the video channel because of the low input capacitance of Q4/Q5 when used in this configuration.

The base bias resistor R21 is shown as 270k Ω. Several BF178s were tried in the circuit and in every case a 270k Ω resistor was found to produce the correct bias. It is possible however that if a very high or very low gain BF178 is used the value of R21 may have to be altered to suit. In this case the value of C6 must also be changed in order to keep the time-constant C6, R21 the same. A potential divider could have been used but was decided against because it would have greatly reduced the input impedance of Q4, requiring a much larger value for C6 in order to keep the same time-constant. There is another reason. With a solid-state h.t. rectifier the h.t. at switch-on can rise to around 300V—until the valve heaters warm up. This could damage Q4 and Q5 which have a maximum *V*ce rating of 115V. R21 ensures that under these conditions the emitter current increases to compensate, keeping *V*ce reasonably constant at about 50V.

★ components list

Resistors:

R1	560 Ω	R12	1.5k Ω
R2	1 M Ω 1W	R13	33k Ω
R3	1 M Ω 1W	R14	2.2k Ω
R4	1 M Ω 1W	R15	10k Ω
R5	68k Ω	R16	12k Ω
R6	2.2k Ω	R17	330k Ω
R7	56k Ω	R18	47k Ω
R8	56k Ω	R19	12k Ω
R9	22k Ω	R20	39k Ω
R10	47k Ω	R21	270k Ω
R11	12k Ω	R22	47k Ω

All ¼W 5% unless otherwise stated

Capacitors:

C1	50μF	40V	electrolytic
C2	200μF	40V	electrolytic
C3	100pF	250V	polystyrene
C4	100pF*	125V	polystyrene
C5	100pF	125V	polystyrene
C6	2.2μF*	125V	polyester
C7	1.5μF	250V	polyester
C8	5000μF	125V	polystyrene
C9—11	0.1μF	100V	polycarbonate

Inductor:

L1 1mH*

Semiconductors:

Z1	30V	400mW
Z2	2.7V	400mW
Z3	75V	1.5W
Z4	39V	400mW
D1—D4	OA91	
D5	1N4148	
Q1—Q3, Q6		BC107
Q4, Q5, Q7		BF178

For components marked * see text.

Effect of Bursts

In the description of the basic operation of the clamp it was stated that Q6 turns on during the back porch of the line sync pulse when the video signal is at black level. This is not strictly true since in the case of colour transmissions this is the period when the chrominance burst occurs. The mean level of the burst is still at black level however and does not upset the operation of this particular clamp.

Construction & Setting up

The layout of the circuit is not at all critical. A suggested layout using a 5 × 2½ in. piece of Veroboard with soldering pins is shown in Fig. 4(a), with the underside wiring shown in Fig. 4(b). The prototype was built in this way and mounted on a convenient bracket at the top of the right-hand edge of the chassis (viewed from the rear) as shown in Fig. 5.

Adjustment of RV1 to obtain the correct delay is quite simple. With a normal picture being displayed

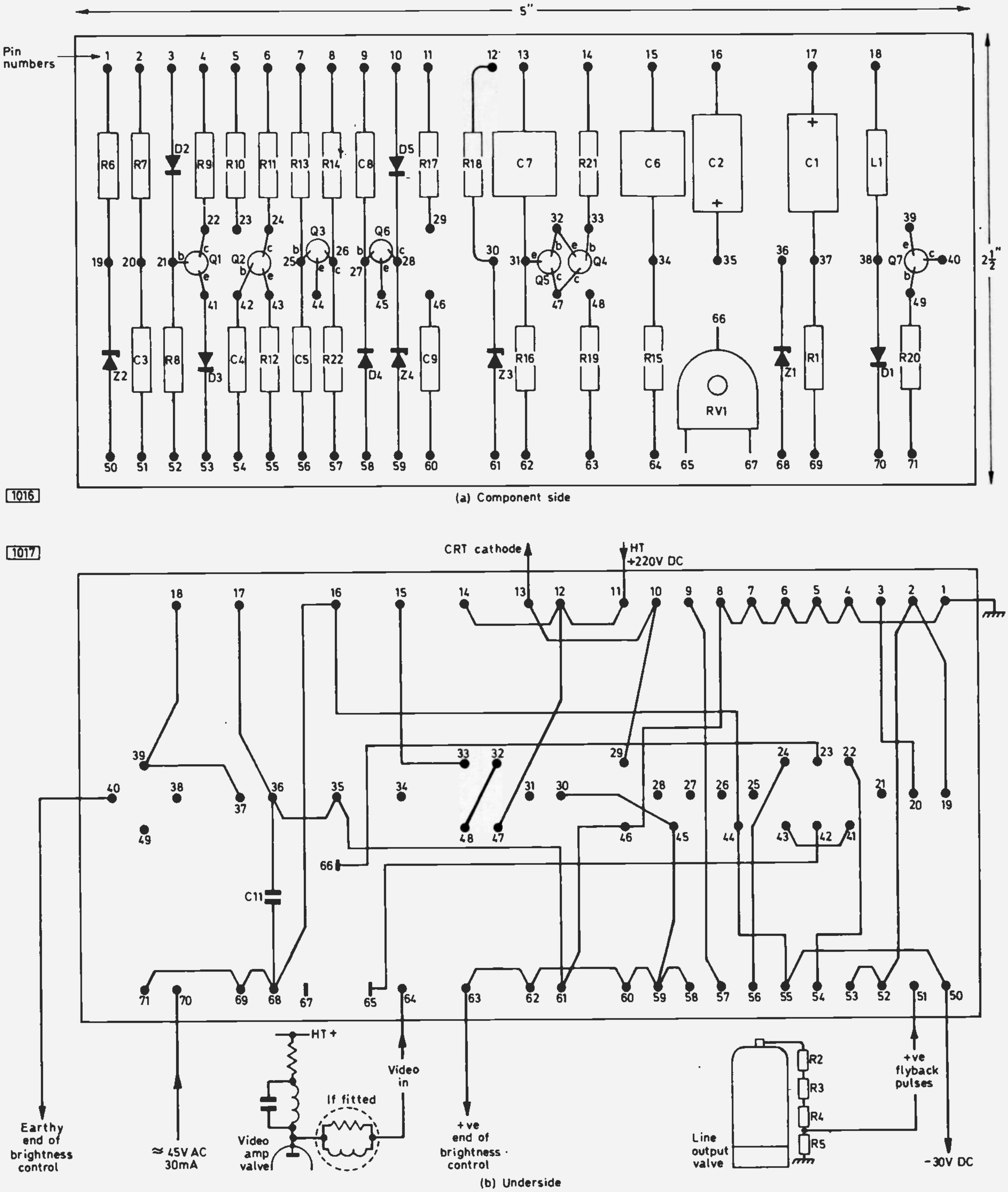


Fig. 4: Component layout and wiring for the black-level clamp.

the line hold control should first be set to its normal or optimum position—with flywheel sync the phasing of the line flyback and hence the timing of the clamp pulse is somewhat dependent on its setting. RV1 is then set at its minimum value and the clamp should now clamp to the sync pulse tips. As RV1 is increased the picture should suddenly darken as the clamp starts clamping to the sync pulse back porch. A further increase in RV1 should eventually darken the picture still further as the clamp starts to operate at the beginning of the picture information. The correct setting is half way between these two points. This provides an

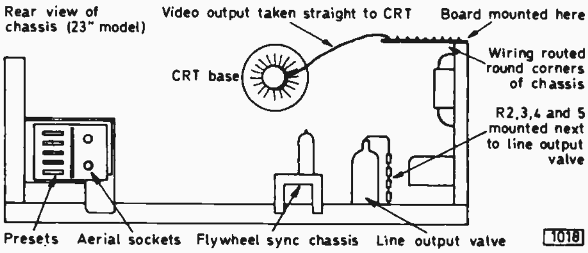


Fig. 5: Mounting position for the clamp in the Thorn 950 chassis.

adequate margin so that normal variations in sync phasing do not cause the clamp to drift off the sync pulse back porch.

Flyback Blanking

Good line and field flyback blanking are important when a black-level clamp is used since a uniformly dark picture shows up any flyback signals rather clearly. The circuitry used for flyback blanking on the Thorn 950 Mk I chassis is shown in Fig. 6. In the Mk II chassis C94 and R117 are omitted and the brightness control is connected to the cathode; also R122 is connected direct to W9 and R142. Having reconnected the circuit as in the Mk I version the field flyback blanking was found to be OK but there was no line flyback blanking.

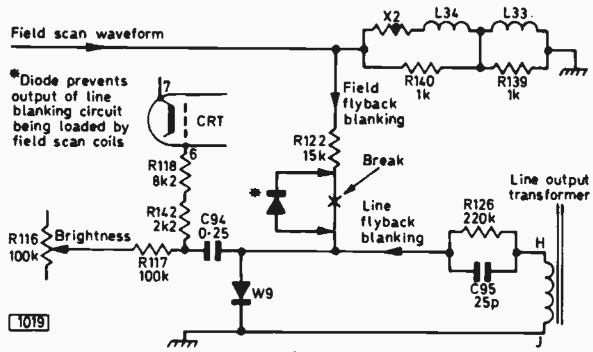


Fig. 6: Flyback blanking modification.

The reason was that R122 and the field scan coils acted as a short-circuit to the output from R126/C95. Inserting a diode as shown between the field scan coils and the line flyback blanking output allows the line flyback blanking pulses to reach the c.r.t. grid since the diode is reverse biased during the field scan, conducting only during the field flyback pulse.

AGC Circuit

In a following article a fast-acting sync tip a.g.c. circuit which has been added to this Thorn 950 Mk II chassis will be described.

SERVICE NOTEBOOK

G. R. Wilding

No Picture

The complaint with a Philips colour set fitted with the single-standard version of the G6 chassis was normal sound but no picture—the result of there being no h.t. at the anode of the PL509 line output valve. The usual cause of this is an internal disconnection to the anode or cathode of the boost rectifier but it wasn't the case on this occasion since this would have resulted in heavy screen grid current with the screen grid winding visibly heated. Both the boost rectifier and PL509 were in fact cool. It was obvious therefore that no h.t. was being applied to the line output stage generally. The h.t. is fed to the stage via a thermal fuse and a 10Ω surge limiter resistor and it was obvious that one of these was almost certainly the cause of the trouble (the circuit appeared on page 511 of the September 1973 issue incidentally). The fuse was found to be intact but a small pinhead hole in the side of the resistor showed where the break had occurred. There was no evidence of a short-circuit so we replaced the resistor—mounted under the thermal fuse clip—and switched on. The result was a normal picture. A tap on the PL509 and PY500 envelopes failed to produce any internal sparks so as everything was in order it appeared that it was a simple case of the resistor going open-circuit.

After about two days however we were called back again to deal with the same fault and this time found that the fuse was open. This was dealt with and as no short was detectable we switched on again. Still no results, due to no h.t. at the anode of the PL509, though it was obvious this time that there was voltage on the screen grid. On changing the PY500 we obtained a good picture but tapping the PL509 produced an odd internal spark. This valve was also replaced therefore and no further picture loss has since been reported.

The PY500 and PL509 valves used in the line output stages of hybrid colour receivers are quite expensive but it pays to replace them both if either is suspected of sparking over occasionally or having an intermittent connection since a fault in one can often damage the other.

Pulsating Picture

Power supply faults seem to crop up quite frequently in sets fitted with the RBM single-standard colour chassis. When the 5A mains fuse has blown the most likely cause is the BT106 h.t. rectifier thyristor and this should be the first test. Recently we came across one of these sets (a Murphy Model CV1916S) in which the picture appeared to pulsate at times in all respects—brightness, contrast and to some extent size. A check showed that the h.t. voltage was varying, a fault which could be caused by any one of a number of components in the thyristor control circuit—or even by greatly varying current demand. As there were no signs of component stress and the raster variations were in no particular direction however this latter possibility was discounted. In case the thyristor was occasionally failing to trigger a new one was tried—they can be changed easily. Results were the same however. The resistors in this area checked out correctly so suspicion was centred on the diac, the reference zener and the two 0·22μF capacitors in this part of the circuit—8C8 which feeds the trigger pulses to the thyristor and 8C7 which charges until the diac breakover voltage is reached. The h.t. voltage and the picture stabilised after changing these two capacitors.

LONG-DISTANCE TELEVISION

ROGER BUNNEY

AUGUST this year gave us something of everything. Sporadic E has been active with reasonable openings into most parts of Europe. The tropospherics have also been active with improved conditions on certain days. The main talking points however lie with MS (Meteor Shower) and F2 layer (or possibly multi-hop SpE!) reception. Meteor showers—the predicted Perseids over the second week of the month—provided a number of short pings as far h.f. as Band III. The 12th and 13th gave the most active openings, Garry Smith noting both DR (Denmark) and NRK (Norway) on ch. E5. Down from the Derby area to Norwich where Clive Athowe spent a fruitful afternoon of the 13th on ch. R8: the TVP (Poland) 5544 test card from Katowice was noted as were several pings of unidentified programme material. Even yours truly noted Band III MS on ch. E5 during the evening of August 8th with the ORF (Austria) 5544 card. Strangely enough this card had a modified central section comprising two lines of bold white letters. Since this card was also noted on ch. E2a until after midnight it is assumed that ORF-1 was carrying President Nixon's Watergate speech, with programme details inserted on the card.

Ghana Received Again!

Possibly the most dramatic news this month however is a second reception of Ghana on ch. E2. It's certainly been the year for exotic signals! Clive Athowe sent us a hurried post card about a suspected ch. E2 Ghana signal on August 9th. Again the programme material consisted of coloured gentlemen in a discussion group. The signal suffered multiple images and was received over the 1845-1912 CET period. The clue to this reception came when Hugh Cocks visited me recently. He too reported unusual activity at the lower end of Band I from a southerly direction: the activity consisted of a Radio Ghana harmonic on about 46MHz with a clear identification, this signal being noted over 1910-1920 CET. Taking these two reports together strongly suggests that Clive too has receved Ghana—our congratulations to Clive on this and the other exceptional reception mentioned.

SpE & the Month's Log

To the more mundane matters of Sporadic E then. The month produced some reasonable openings of generally medium- to long-hop distances. Of particular interest was the reception of a new TSS (USSR) electronic test pattern on ch. R2 on August 12th at 1215 CET—a rather complicated though distinctive pattern comprising basically a single checkered band across the centre, a single grey/black band—also checkered—at the top with colour bars and assorted squares of varying shades/colours underneath the central band. This pattern was replaced at 1230 by the conventional 0249 test card.

My log for the period follows with the more regular receptions deleted in the interests of space. This includes deletion of DFF (East Germany) on ch. E4 which seems to be a regular MS type signal most mornings.

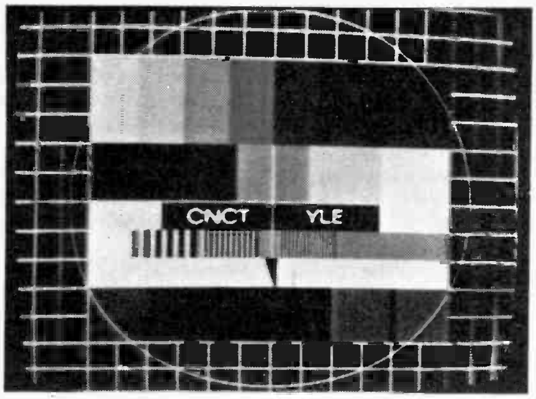
2/8/74	TSS (USSR) R1, 2; CST (Czechoslovakia) R1; MT (Hungary) R1—all SpE.
3/8/74	CST R1; NRK (Norway) E2; TVE (Spain) E2—all MS. The new ORTF-3 (France) transmitter at Le Havre (ch. E40) was noted in operation.
4/8/74	CST R1—MS.
5/8/74	TVP R1; SR (Sweden) E2; TVE E2—all MS.
6/8/74	WG (West Germany) E2; CST R1; SR E2—all MS; TVE E2—SpE.
7/8/74	TSS R1, 2; CST R1; YLE (Finland) E2; SR E2; TVE E2—all SpE; DR (Denmark) E4—MS.
8/8/74	MT R1, 2; JRT (Yugoslavia) E3; ORF (Austria) E4; TVE E2, 3; RTP (Portugal) E3; also unidentified signals—all SpE; CST R1; ORF E2a, E5—all MS.
9/8/74	WG E4; TVE E2, 4—all MS.
10/8/74	TVE E2, 3, 4; RTP E2, 3; RAI (Italy) IA, IB; all SpE; NRK E4—MS.
12/8/74	TSS R1, 2, 3; YLE E2; NRK E3, 4; SR E2, 3, 4—all SpE.
13/8/74	CST R1—MS.
14/8/74	RAI IA, IB; TVE E2, 3, 4; RTP E2—all SpE.
15/8/74	NRK E2, 3, 4; SR E2, 3, 4; TVE E2, 4—all SpE.
17/8/74	Swiss E4; also unidentified signals—all SpE.
19/8/74	SR E2—MS.
21/8/74	CST R1—MS.
22/8/74	SR E4; ORF E4—MS; TSS R2—SpE; improved tropospherics from France at u.h.f.
23/8/74	CST R1; TVP R1—both MS; TSS R1; TVE E2, 4—both SpE.
24/8/74	WG E2—MS.
25/8/74	TSS R1—SpE.
27/8/74	Unidentified programmes R1—SpE.
29/8/74	TVP R1; unidentified programmes R1, 2—all SpE.

There has been a tendency to increased SpE activity during the late evenings, generally from an Easterly direction—into the USSR. Due to working and domestic activities many of these openings were missed unfortunately.

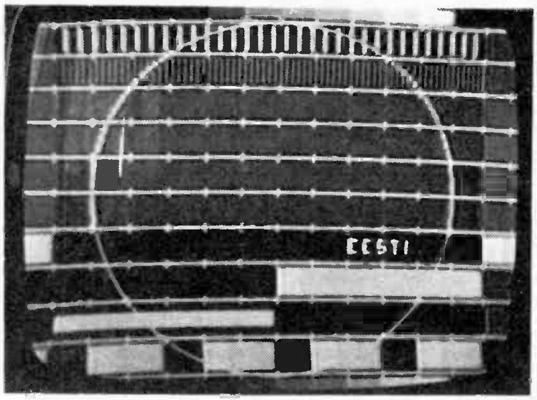
Reports and News Items

Garry Smith noted a form of 5540 test card—consisting of a large circle—on ch. E3 at 1700 on August 13th. We have no other information on this mystery. DFF (GDR—East Germany) is using a colour blockboard similar to the NOS (Holland) 5552 pattern. The other Dutch pattern test card—type 5540—has been noted on ch. E7 in the early hours (0854) and is suspected to be from the HR-1 transmitter at Meissner, West Germany (item from Clive Athowe). Clive also reports a lift in the trop situation on August 6th with good East German reception at u.h.f.

Mr. van der Linden (Rotterdam) tells us that the ORTF test card has been seen at u.h.f. in Spain with the identification "RTF CNCT Paris": we understand that "CNCT" indicates "central national control technical"—translated from the French. Strangely, the report continues that later the same day this transmitter radiated PAL colour signals. We are wondering if this is a transmission from a Spanish outlet taking incoming signals from France? Programmes in colour are being listed in the Spanish TV Guide—these are generally though not always of the PAL standard.

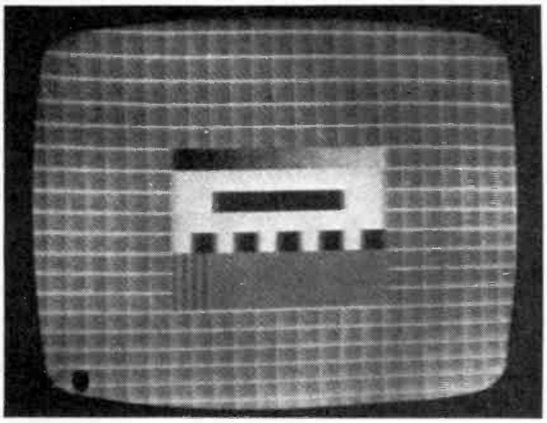


Fubk test card used by YLE (Finland) before Eurovision broadcasts.

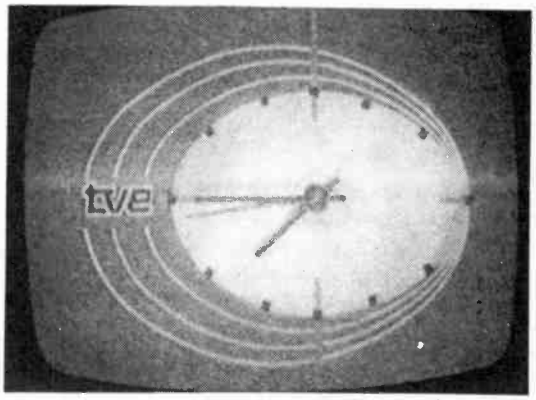


TSS (USSR) electronic test pattern transmitted by Tallin on ch. R2.

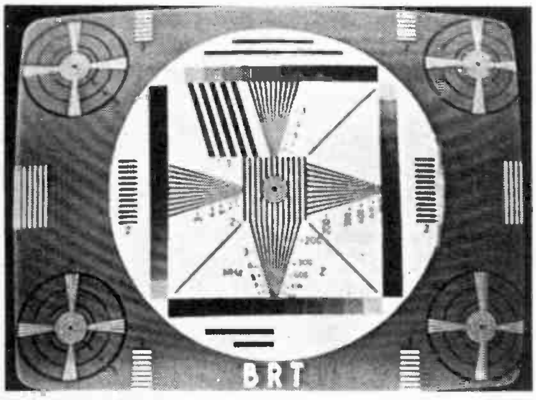
Above photographs courtesy Seppo Pirhonen (Finland).



TSS electronic test pattern.
Photograph courtesy of Ryn Muntjewerff.



New TVE (Spain) clock.
Photograph courtesy of P. F. Vaarkamp.



New BRT (Belgium) test card (T05 type) used fifteen minutes before start of programme transmissions.
Photograph courtesy Dieter Scheiba.

Italy: The EBU reports that the Rome transmitting station has new first programme aerials. The previous installation gave 36kW e.r.p. from a 2.5kW transmitter while the new system enables a 10kW transmitter to be used. The original transmitting array had been in use since 1953 and the purpose of the new array is to improve reception in areas where interference is being experienced from other transmitters. The station operates on ch. G (200-207MHz) from Monte Mario.

Libya: Radio link facilities for the televison network are being increased, with connection into Tunisia in the West of the country. Another radio link—a trans-horizon tropospheric scatter circuit—is being constructed to connect Darnah (Libya) with the EIRT television network on Crete, thus providing a relay into and from the Eurovision network.

The Problem of Ch. B2

A great number of letters arrive complaining that the lower end of Band I (chs. E2, R1) is rendered useless—except for the strongest of SpE signals—because of the local ch. B2 transmitter. The ch. B2 sound frequency (48.25MHz) clashes with the vision frequencies of chs. E2 (48.25MHz) and ch. R1 (49.75MHz). Ch. E2 can be lost therefore to those enthusiasts adjacent to a high-power ch. B2 transmitter while ch. R1 is a struggle. How can the "lost" channels be used for DXing? For ch. R1 a ch. B2 sound notch filter can be successfully employed—see *Practical Television* November 1969 (DX Filter) or the TV-DX booklet. The notch filter cannot be used for ch. E2 purposes however since by removing the ch. B2 sound it also most effectively removes the ch. E2 vision! To date two methods of reducing the ch. B2/E2 problem have been adopted—the use of specially orientated arrays and the use of a variable attenuator/phasing unit.

The aerial method is simplicity itself. A directional array is mounted at a fixed position which gives minimum/nil pickup from the local ch. B2 transmitter. To obtain minimum signal from the local transmitter it may be necessary to clamp the array in a position which departs from the conventional horizontal/vertical mounting. Since Sporadic E signals have no respect for transmitted polarisation, signals arriving at the receiving site via this mode will vary considerably in their polarisation. As a result a random position for the array is of little consequence. I suggest that an array cut to ch. R1/E2 (49MHz) is most suitable for this purpose and that the familiar H array should first be tried. Where possible aim the array into a DXing direction; by panning and tilting the null position for the local transmitter should be found. Finally clamp the array in this position. Experience has shown that the array used for this purpose should be as far as possible from the main rotatable DX array since movement of the latter will tend to change the standing-wave pattern in the area and consequently the local pickup by the newly erected "null pickup" array. The old J Beam Q Beam array was especially suitable for this work.

Space permitting next month we will give the circuit for the variable phasing system. In the meantime if anyone has achieved success in this field with local signal rejection please let us know so that we can pass on the information for the benefit of other DX enthusiasts.

From Our Correspondents . . .

Our old friend Seppo J. Pirhonen of Lahti, Finland has sent a long letter giving information on the TSS network—a daily signal for him. Leningrad 1 commences at 0600 GMT with day long programmes except for summer weekdays when the test card is radiated between 1000-1300 GMT. Leningrad radiates the early morning test card between 0430-0600 GMT. Tallin differs in that the 0249 card is radiated between 0520-0540 and the electronic pattern between 0540-0600. At the end of the day the test card is radiated for up to 15 minutes—this is at approximately 2130 GMT. Petrozavodsk apparently follows a similar pattern. Seppo has kindly sent a number of photographs a couple of which we feature this month.

James Burton-Stewart (Great Horwood, Bucks) has received information from NRK (Norway). The evening programme is transmitted between 1755-2230 CET; childrens/schools programmes are transmitted between 0900-1300, excluding June through August. "The test picture 'Norge Televerket' (i.e. 5544 card) is radiated every weekday, the other picture a short time before the start of the programmes". The "other picture" is in fact the Fubk card and test card F.

A new DX enthusiast Peter Gregory (Blackwell) has written to tell us of his receptions this season. By all accounts he's been most successful; strong signals have been received from DFF, WG, CST, RAI, SR, TVP and MT. Peter uses a Murphy Model V849 (Bush TV125 chassis) at his home near Buxton and a two-element E4 array—at only 18ft. He also suffers from the ch. B2 problem.

A short note from Anthony Mann (Western Australia) tells us that the 5544 test card is now used by the National ABC (Australian Broadcasting Commission) network, with the identification "ABC" at the top and "Television" at the bottom. The major commercial stations in the area have been using this pattern for some months. Colour test transmissions for one hour only commence on October 7th; outside broadcasts from October 19th; full colour on March 1st, 1975.

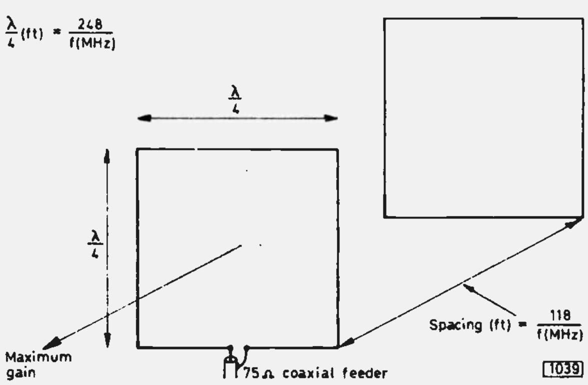


Fig. 1: Basic data on the quad loop aerial. Shown for horizontal polarisation. Spacing not shown to scale.

Experimental Aerial for Indoor Use

We have often mentioned the well known North American organisation WTFDA which from time to time features technical articles on aerial construction and associated theoretical matters in its bulletin. The June 1974 issue contained an extremely interesting discussion by Bill Smith (W5TVB—v.h.f. radio Amateur and editor of the v.h.f. section in *QST* magazine) on the use of a "quad loop" array for TV-DXing in Band I.

The array is basically a square with sides a quarter-wave long. The feed to the receiver is taken from the centre of one side (see Fig. 1). If the feed is taken from a vertical side the array is predominantly vertically polarised; if taken from a horizontal side the polarisation is basically horizontal. Since the array was for indoor use Bill made it of twin flex, splitting the two wires and soldering them together at one end to form a continuous wire. This was tacked to the wall in a square and the feeder connected accordingly. Maximum signal pickup is broadside to its faces—i.e. bidirectional as with a dipole. Bill arranged two such quad arrays, again tacked to the walls, to give all-round coverage—one array picking up along the North/South path and the other along the East/West path.

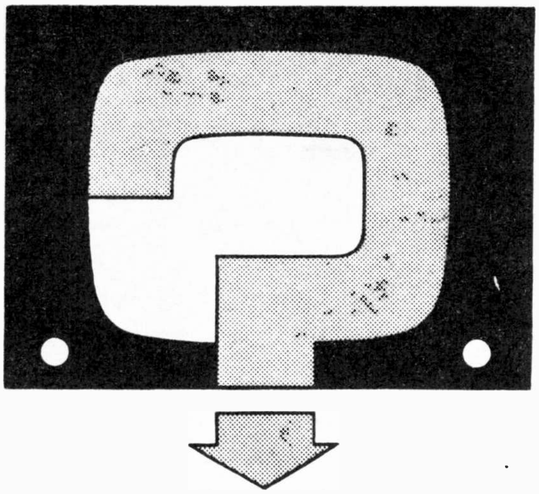
Bill says that the performance of the quad array was far better than that of a standard dipole though theoretically it has a 9/10 gain over a dipole. After using many types of indoor array Bill has concluded that this particular aerial gives by far the most effective performance.

A useful increase in gain is obtained by placing a second closed loop of similar dimensions behind the active unit. A gain of up to 6dB is claimed, with a front-to-back ratio of up to 25dB. Maximum gain with the "multi-element" version is in the direction away from the rear closed loop which acts as a reflector.

The formula $248/f$ gives the quarter wavelength dimension for each side in feet. The spacing (again in feet) between the active array and the reflector is found from the formula $118/f$. In both cases f is the centre frequency. If for example we wish to make such an array for ch. E4 vision (62.25MHz) the quarter wave sides will be 48in. long and the spacing between the "dipole" and reflector 23in. Normal 75Ω coaxial cable can be used for connection purposes.

Labgear manufactured a similar array for Band I in the late 1950s and at the present time a scaled down version is available for use at u.h.f.

Our thanks to the WTFDA, PO Box 163, Deerfield, Illinois 60015, USA for these details.



YOUR PROBLEMS SOLVED

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DECCA CTV22

We are having difficulty converging the blue raster. The only control that works correctly is the blue horizontal centre line control VR511. The R/G vertical left-hand control VR508 gets hot, also the clamp transistor Tr501. The top of the picture curves slightly upwards.—E. Jenkins (Barnhurst).

The waveform shaping network in parallel with VR511 and its associated 4·7 Ω resistor will have to be checked—L512 for continuity, VR512 to make sure that it is intact, and the integrating capacitor C506 (1 μF) by substitution. If you convert to single-standard operation it would be an idea to disconnect the system switching solenoid on the convergence panel and fix the slide switch in the 405-line position: this brings into circuit a duplicate set of controls which are likely to be in better condition.

SONY TV9-90UB

There is a rather odd fault that we have had difficulty tracing on this set—the contrast level alters intermittently. The effeet occurs on both systems.—R. Miller (Cheam).

This fault is generally due to the 100μF electrolytic C507 in the emitter circuit of the video output transistor being defective.

BUSH TV125

I am having a bit of trouble with u.h.f. reception on this set. V.H.F. reception is all right, also sound on u.h.f.—it is the vision signal that is the trouble. Suggestions made in your articles on this chassis in the June and July issues 1969 have been tried without success and it occurs to me that the u.h.f. tuner is at fault.—J. Burrows (Manchester).

The only action which need be taken so far as the u.h.f. tuner is concerned is to replace the two valves—PC88 and PC86. Since the u.h.f. sound is all right these valves (and the tuner as a whole) cannot be too far out. The h.t. supply to the tuner should be 150V. Check the system switching and the voltages around the EF85 and the two EF80 vision i.f. stages—particularly the cathode voltage (pin 3) of the EF80 valves. Change these valves if the cathode voltage reading is low.

ULTRA 6703

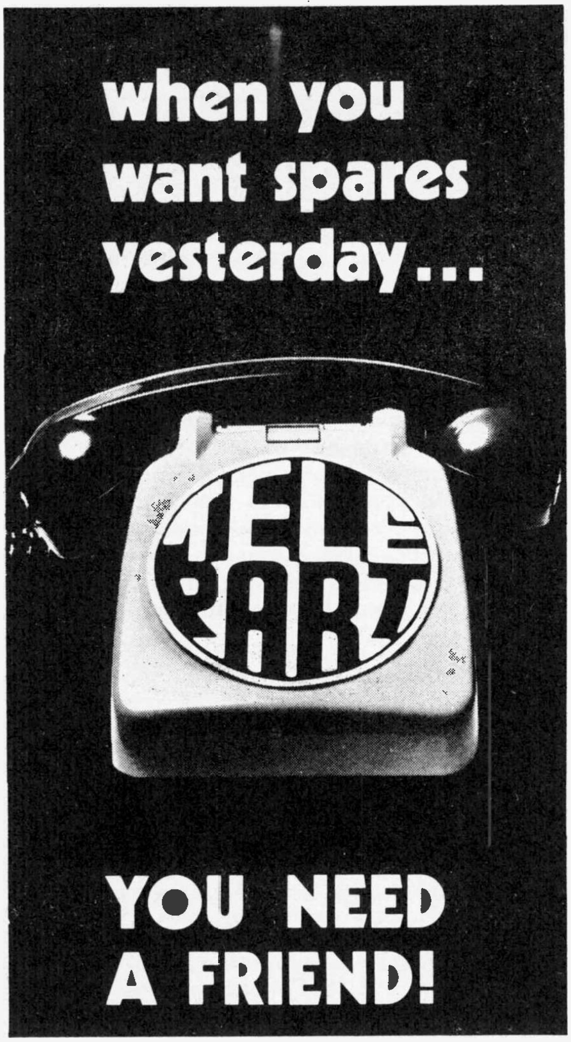
After anything from fifteen minutes to two hours after switch on the picture takes on a sort of high-frequency shiver, the movement being in the left-hand direction. This is quickly followed by shrinkage at the bottom of the raster, leaving a black strip about 2in. high and above that a very bright strip about 3in. high. At this point the set goes dead. If the set is switched off and allowed to cool it works normally again on being switched on until the fault repeats.—J. Levy (Potters Bar).

The most common caues of this problem in order of likelihood are: C619 (140μF) which smooths the 56–63V line from the power supply board; C506 (25μF) in the flywheel line sync filter circuit; and C631 (0·01μF) in the dynamic trip circuit on the power supply board. The problem could also be caused by severe loading of the power supply module: check the current at the "h.t." fuse F603 and the voltage (should be 1·3V) across the line output stage earth return resistor R907 in the beam limiter circuit. (BRC 3000 chassis.)

PHILIPS 19TG171A

On v.h.f. there is a very poor picture and very little sound volume (ITV) or a broken picture and even less sound (BBC) while on the u.h.f. channels there are just dim, broken images and the sound is barely audible. The v.h.f. tuner valves and all other likely valves have been replaced without improving matters. U.H.F. reception has always been below average on this set although the aerial installation is all right—checked on another set.—C. Barton (Southampton).

We suggest you dismantle the v.h.f. tuner, then remove and clean the biscuits and switch contacts. Replace the 5.6k Ω and 6.8k Ω resistors which feed the anode (pin 8) of the oscillator section of the PCF801, then check the 22k Ω screen grid feed resistor (pin 7) and the 1k Ω resistor which feeds the anode of the PC900 r.f. amplifier. Also clean the valve sockets. (Philips Style 70 series.)



PYE CT200

The left half of the screen is predominantly green while red is predominant on the right-hand side, objects changing colour across the screen. Following servicing the fault seems to be more pronounced, the monochrome picture is tinged with colour and colour reception is grainy.—J. Moreland (Newport).

The trouble is on the decoder panel and seems to be connected with the feedback circuit associated with the BF336 red output transistor (Tr431) and the TBA530 matrixing i.c. Try adjusting the set peak white preset R430 in this circuit and suspect C433 (220pF) of being open-circuit.

HMV 2804

The sound volume is normal when the set is first switched on but within a few seconds falls to a very low level and cannot be increased. The PCL82 audio amplifier/output valve has been replaced and the vision is unaffected.— T. Carter (Sunderland).

Check the voltage conditions around the PCL82 and the value of the pentode cathode resistor R89—it may be discoloured. Check the 4μF screen grid decoupler C61, and the coupling capacitor C64 between the two sections of the valve for leakage. This chassis is subject to dry-joints and breaks in the print, especially around the valveholder contacts. Inspect the print carefully and run a soldering iron over any burnt areas or suspect joints. (BRC 1500 chassis.)

KB KV005

The picture on this set is black with dim highlights and there is heavy vision on sound. The contrast control has no effect and the brightness control very little. All electrolytics have been replaced and all valves checked. Disconnecting the aerial makes no difference.—R. Dodds (Skegness).

There seem to be two faults here. For the dark picture, first check that the c.r.t. first anode voltage (pin 3) is about 500V. If not check the value of the feed resistor R156, the focus control R157 and the first anode decoupler C137 for leakage. If the first anode voltage is about right check the components in the video output stage, especially the screen grid feed resistor R51 (3·9kΩ). The c.r.t. itself could be at fault. If the sound fault is mains hum or timebase breakthrough check the earthing of components around the PCL86 audio valve V11. If on the other hand the fault disappears when the volume control is set to minimum check the decoupling of the i.f. valves—the capacitors connected to pin 8 of the EF184, EF183 and 6BW7. (ITT/STC VC3 chassis.)

PHILIPS 23TG153A

Although the picture is clear there is severe cramping at the top of the raster and this cannot be cleared by making adjustments. All valves in the set have been tested and replaced as necessary. The cramping seems to fluctuate or flutter and sometimes returns to normal for a short time.—B. Gains (Enniskillen).

Top cramping on this model is almost always caused by short-circuit turns on the field output transformer: an improved type is now available from the manufacturers. Before replacing this however it would be worth checking the capacitor in the field linearity feedback loop—C422 (0·0082μF).

HMV 2639

The problem with this set is caption buzz on sound. I intend to increase the value of the video amplifier screen grid resistor to 8.2kΩ as suggested in your article on intercarrier sound. Could you advise on the wattage rating to use?—E. Topham (Stoke).

We advise using a 1W resistor in this position. Making this replacement should help minimise buzz but you may also find it necessary to retune the ratio detector transformer (L27/L28) and adjust the detector balance preset R87. (BRC 1400 chassis.)

DECCA CS2233

The trouble with this set is faint patterning—moiré patterning I believe it is called—on the screen. This doesn't affect the colour but is visible on pastel shades. It is present on all programmes and flickers in a random fashion—it is always roughly the same shape and in the same position.—E. Tyson (Birmingham).

Moiré patterning as you describe it is caused by interaction between the scanning lines and the holes in the shadowmask tube. The effect should virtually disappear if the height is reduced until the raster just fills the screen. To eliminate any remaining pattern reduce the c.r.t. first anode voltages (grey-scale tracking controls) or very slightly defocus the picture. (Decca 30 series chassis.)

PHILIPS 19TG175A

We are troubled by loss of picture with this set. Previously we noticed that there was internal arcing in the PY800 boost diode so a replacement was fitted, curing the fault. The condition returned after about a month however.—P. Sloane (Scarborough).

PY800 boost diodes do often fail unfortunately and you are probably just unlucky that the replacement was short lived. The arcing could however be due to spark-overs in the PL500 line output valve. Boost capacitors give a fair amount of trouble and though they generally go short-circuit it would do no harm to replace this component. One section or another of the mains dropper resistor in this chassis often goes open-circuit—check that the values are correct as an incorrect replacement may have been fitted thus over-running the valves. (Philips Style 70 chassis.)

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TELEVISION NOVEMBER 1974

TEST CASE

143

Each month we provide an interesting case of television servicing to exercise your ingenuity. These are not trick questions but are based on actual practical faults.

A Philips Model G20T236/0T gradually developed line drift over a period of several months. The drift was insignificant at first, and only mild line hold control adjustment was required during an evening's viewing. As time went on however the symptom became more serious and several adjustments were necessary within the first hour of operation. It was at this stage that the viewer decided to lodge a complaint.

Examination of the set in the viewer's home revealed that the conditions were as described. The back of the cabinet was then removed and as the technician had previously encountered similar trouble which proved to be caused by change of the line oscillator valve characteristics with increasing temperature the appropriate ECC82 was replaced and the set allowed to operate for a period while the technician was duly refreshed by the customer. There was no sign of drift subsequent to the valve replacement so the technician switched off, replaced the back and continued on his way.

On arriving back at base that evening the technician was presented with a recall note indicating that the symptom was exactly the same as before the valve change. What were the most likely components overlooked by the technician? See next month's TELEVISION *for the solution and for a further item in the Test Case series.*

SOLUTION TO TEST CASE 142

Page 571 (last month)

It is often overlooked that the frequency response of the video output stage affects not only the picture signal but also the sync pulses: it is not uncommon in fact for the picture information to remain reasonably pallatable yet for a change in the frequency response of the stage to impair the locking efficiency of the sync pulses.

This was just the trouble in the Bush Model TV135R and the clue was given by the mis-shapen field sync pulses. The field sync pulses require good phase linearity at low frequency and failure of any highish value decoupling capacitor in the video output stage can have the effect of attenuating the lower frequency components of the sync pulses, thereby changing their shape. One capacitor which can do this is the 1µF capacitor (2C22) which decouples the screen grid of the video amplifier section of the PFL200 in this chassis and in the case in question this capacitor was found to be virtually open-circuit. Replacement cured the field roll and significantly improved the line lock as well.

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TV75 or C	TV125U
TV76 or C	TV128
TV77	TV134
TV78	TV135
TV79	TV135R
TV83	TV138
TV84	TV138R
TV85	TV139
TV86	TV141
	TV145
TV91	TV148
TV92	TV161
TV93	TV165
TV94	TV166
TV95 or C	*TV171
TV96 or C	TV175
TV97	TV176
TV98C	TV178
TV99 or C	181S
TV100C	183
TV101C	183D
TV102C	183S
TV103 or D	183SS
TV105 or D or R	185S
TV106	186
TV107	186D
TV108	186S
TV109	186SS
TV112C	191S
TV113	191D
TV115 or C or R	193S
TV118	193D
TV123	
TV124	

From model TV123 to TV139 there have been two types of transformer fitted. One has pitch overwind, the other has plastic moulded overwind. Please state which type required as they are not interchangeable.

BAIRD

600	628	662	674
602	630	663	675
604	632	664	676
606	640	665	677
608	642	666	681
610	644	667	682
612	646	668	683
622	648	669	685
624	652	671	687
625	653	672	688
626	661	673	

Please quote part No. normally found on tx. base plate; 4121, 4123, 4140 or 4142.

DECCA

DR20	DR34	DR71	DR505
DR21	DM35	DR95	DR606
DR23	DM36	DR100	666TV-SRG
DR24	DM39C	DR101	777TV-SRG
DR29	DR41	DR121	
DR30	DM45	DR122	MS1700
DM30	DR49C	DR123	MS2000
DR31	DM55	DR202	MS2001
DR32	DM56	DR303	MS2400
DR33	DR61	DR404	MS2401

SOBELL

T24	ST284 or ds	1010dst	1033
SC24	ST285 or ds	1012	1038
TPS173	ST286 or ds	1013	1039
TPS180	ST287 or ds	1014	1047
ST195 or ds	ST288 ds	1018	1048
ST196 or ds	ST290ds	1019	1057
ST197ds	ST291ds	1020	1058
SC270	ST297ds	1021	1063
T278	1000ds	1022	1064
ST282	1002ds	1023	1065
ST283	1005ds	1032	1066

MURPHY

V310	V430	V520	V879 or C*	V789	V2015SS
V310A	V430C	V530	V923*	V153	V2016S
V310AD	V430D	V530C	V929 or L*	V159	V2017S
V310AL	V430K	V530D	V973C*	V173	V2310
V310CA	V440	V539	V979*	V179	V2311C
V320	V440D	V540	V653X	V1910	V2414D
V330 or D	V440K	V540D	V659	V1913	V2415D
V330F or L	V470	V649D	V683	V1914	V2415S
V410	V480	TM2 Chassis	V739	V2014	V2415SS
V410C	V490	V843*	V735	V2014S	V2416D
V410K	V500	V849*	V783	V2015D	V2316S
V420	V510	V873*	V787	V2015S	V2417S
V420K	V519				

*Two types fitted. One has pitch overwind, the other has plastic moulded overwind. Please state which type required as they are not interchangeable.

PHILIPS

23TG111a	G19T210	G23T210
23TG113a	G19T211	G23T211
23TG121a	G19T212	G23T212
23TG122a	G19T213	G24T230
23TG131a	G19T214	G24T232
23TG142a	G19T215	G24T236
23TG152a	G20T230	G24T238
23TG153a	G20T232	G24T300
23TG156a	G20T236	G24T301
23TG164a	G20T238	G24T302
23TG170a	G20T300	G24T306
23TG171a	G20T301	G24T307
23TG173a	G20T302	G24T308
23TG175a	G20T306	G24T324
23TG176a	G20T307	
23FG632	G20T308	

GEC

2000	2015	2022	2043	2064
2001	2017	2023	2044	2065
2010	2018	2032	2047	2066
2012	2019	2033	2048	2082
2013	2020	2038	2063	2083
2014	2021	2039		

PYE

11u Series	
12u	
13u	State Pt. No.
14u	required
15u	AL21003 or
20u	772494
V700 or A or D	
V710 or A or D	State Pt. No.
V720	required—
V830A or D or	772444 or
LBA	771935

FERGUSON, ULTRA, MARCONI, H.M.V. (BRC, Jellypots).
ALL MODELS IN STOCK.

ALBA, COSSOR, EKCO, FERRANTI, K.B., PYE. ALL MODELS IN STOCK.

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THORN B.R.C. MONOCHROME	ORDER Ref.	
980, 981, 982	RT1	£3.30
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950/2, 1400-5 stick	RT3	£3.90
1400 Portable-3 stick	RT3A	£3.60
1500 20" 3 stick	RT4	£3.60
1500 24" 5 stick	RT5	£3.90
1580 Portable-2 stick	RT16	£3.50
1590, 1591	RT17	£1.30

E.H.T. RECTIFIER TRAYS (COLOUR)

MAKE	CHASSIS COLOUR	
DECCA	CTV19, CTV25	£8.00
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DECCA	CS1730	£5.80
GEC	Dual & Single std. Valve Type	£6.70
ITT-KB	CVC-1, 2, 3	£7.20
PHILIPS	G8 510-550 Series	£7.20
PYE	691, 692, 693, 697	£6.70
PYE	713 CT200	£6.40
BUSH MURPHY	Single std plug-in	£8.00
BUSH MURPHY	Dual standard	£9.80
THORN BRC	2000	£8.00
THORN BRC	3000	£7.10
THORN BRC	8000	£4.10
THORN BRC	8500	£7.00
GEC	Solid State 90°	£7.80
PYE	CT262 & 266 731 Chassis	£7.40

COLOUR TV Line out-put transformers

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8000 Chassis
8500 Chassis
All £6.80 ea.

GEC
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Single Standard
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G6 „ S/S
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CTV25 Mk.3
CTV162
£7.90 ea.
CTV167 Mk.1 & 2
£11.50 ea.
CTV167 Mk.3
CTV174D
CTV182S
CTV184S
CTV187CS
CTV194S
CTV197C
CTV199S
£7.10 ea.

DECCA
CTV19 Valve Rec.
CTV25 „ „
Primary Coil
£3.70 ea.
CTV19 D/S Tripler
CTV25 „ „
CTV25 S/S Tripler
CS1730
£7.80 ea.
CS1910
CS2213
£7.10 ea.

EKCO
CT102
CT104 £12.00 ea.
CT103
CT105
CT106
CT107
CT108
CT109
CT111
CT120
CT121 &/T
CT122 £8.90 ea.

PYE
CT70
CT71 £12.00 ea.
CT72
CT73
CT78
CT79
CT152
CT153
CT154 £8.90 ea.

MURPHY

CV1912	CV2510 Mk.3
CV1916S	CV2511 Mk.3
CV2210	CV2516S
CV2212	CV2610
CV2213	CV2611
CV2214	CV2614

£7.90 ea.
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CV2511 Mk. 1 & 2
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