TELEVISION

SERVICING · CONSTRUCTION · COLOUR · DEVELOPMENTS

20p

NOVEMBER 1971

VARICAP TUNER
PANORAMIC MONITOR

ALSO: TRANSISTOR VIDEO CIRCUITS
TV AUDIO CIRCUITS FOR THE CONSTRUCTOR
SERVICING THE DEFIANT 9A50/52 SERIES

NEW LINE OUTPUT TRANSFORMERS

ALBA 655, 656, 717, 721 **£3·75.** 890-895, 1090, 1135, 1195, 1235, 1395, 1435 **£5·00.**
ARGOSY. 17K10, 17K11, 17K12, 17K14, 19K17, 17K43 **£4·00.**
BAIRD. All models price **£5·90.** From model 600 quote part no. normally found on TX base plate.
BUSH TV53 to TUG69 **£2·00.** TV91 to TV139 **£4·75.** (From Model TV123 an alternative Square Tag Panel was fitted on Main Bobbin, please state if required.) TV141 to TV176 please state part number **£4·50.** TV75 to TV86 **£4·75** (except TV80).
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EKCO T231, T284, TC267, T283, T293, T311, T326, T327, T330, TMB272, T344, T344F, T345, TP347, T348, T348F, TC347, TC349, TC356, T368, T370, TC369, T371, T372, TP373, TC374, T377A, T393, T394, 433, 434, 435, 436, 437 all at **£4·00.** 503, 504, 505 506 **£4·75.**
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·0022	600v.	£0·04
·0033	600/1500v.	£0·04
·0047	600/1500v.	£0·04
·01	400v.	£0·04
·022	600v.	£0·05
·033	600v.	£0·05
·047	600v.	£0·05
·1	600v.	£0·05
·22	600v.	£0·10
·47	600v.	£0·14
·01	1000v.	£0·06
·022	1000v.	£0·06
·047	1000v.	£0·09
·1	1000v.	£0·09
·22	1000v.	£0·14
·47	1000v.	£0·19
·001	1500v.	£0·08

WIRE-WOUND RESISTORS (5's)

10 watt rating, suitable for mains dropper sections.

1	Ohm	£0·09
10	Ohms	£0·09
13	,,	£0·09
25	,,	£0·09
33	,,	£0·09
50	,,	£0·09
87	,,	£0·09
100	,,	£0·09
150	,,	£0·09
220	,,	£0·09
330	,,	£0·09
1K	,,	£0·09
2·2K	,,	£0·09
3·3K	,,	£0·09
4·7K	,,	£0·09

PULSE CERAMICS (5's) 12KV

100pf	22pf	£0·06
120pf	47pf	£0·06
180pf	68pf	£0·06
250pf		£0·06

Tubular type for use in Scan correction circuits and Line Outputs.

CERAMICS (6's)

500pf	22pf	£0·03
680pf	47pf	£0·03
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1000pf	100pf	£0·03
1500pf	120pf	£0·03
3000pf	180pf	£0·03
5000pf		£0·03

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25mfd	25v.	£0·07
50mfd	25v.	£0·08
100mfd	25v.	£0·10
250mfd	25v.	£0·15
500mfd	25v.	£0·19
1000mfd	12v.	£0·30
1000mfd	30v.	£0·30
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2500mfd	30v.	£0·45
3000mfd	30v.	£0·47
5000mfd	30v.	£0·55
25mfd	50v.	£0·08
50mfd	50v.	£0·10
100mfd	50v.	£0·13
250mfd	50v.	£0·18
500mfd	50v.	£0·24
2000mfd	50v.	£0·47
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Wire ended, 450v. working.

1mfd	£0·07
2mfd	£0·08
4mfd	£0·11
8mfd	£0·13
16mfd	£0·16
32mfd	£0·23
50mfd	£0·25
8/8mfd	£0·19
8/16mfd	£0·25
16/16mfd	£0·26
16/32mfd	£0·27
32/32mfd	£0·27
50/50mfd	£0·42
50/50/50mfd	£0·52

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100/200mfd	£0·63
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200/200/100mfd	£0·95
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100/300/100/16	£0·95
100/400/32mfd	£0·95
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25K	Vertical	£0·07
50K	,,	£0·07
100K	,,	£0·07
250K	,,	£0·07
500K	,,	£0·07
1 meg	,,	£0·07
2 meg	,,	£0·07
500K	Horizontal	£0·07
680V	,,	£0·07
1 meg	,,	£0·07

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1mfd	18v.	£0·09
2mfd	18v.	£0·09
4mfd	18v.	£0·09
5mfd	18v.	£0·09
8mfd	18v.	£0·09
10mfd	18v.	£0·09
16mfd	18v.	£0·09
25mfd	18v.	£0·09
32mfd	18v.	£0·09
50mfd	18v.	£0·09
100mfd	18v.	£0·09
200mfd	18v.	£0·09

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Miniature	£0·08
TH1	£0·13

RADIO/TV GLASS FUSES

1 amp, 1·5 amp, 2 amp, 3 amp.
Per dozen £0·15

MAINS FUSES

2 amp, 3 amp, 5 amp, 13 amp.
Per dozen £0·25

TERMINAL STRIPS

2 amp	£0·12
5 amp.	£0·14
15 amp	£0·29

RECTIFIERS

Silicon Mains (5's)

Westinghouse S10AR2	£0·33
BY127 Mullard	£0·26
BY327	£0·25

CONTACT COOLED FULL WAVE

75ma	£0·60
100ma	£0·70
150ma	£0·86

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Bakelite top	£0·04
Egen metal	£0·08
Single point (car radio)	£0·10

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100K	£0·08
1 Meg	£0·08
2·2 Meg	£0·08

JACK PLUGS

Chrome standard	£0·20
Standard	£0·15
3·5mm. metal	£0·15

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3-pin	£0·10
5-pin	£0·11
Sockets	£0·06

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Bush/Murphy/BRC, etc.
Line/frame timebases etc.

3 leg	£0·31
4 leg	£0·31
5 leg	£0·31

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½ watt, 1 watt and 2 watt.

The following values are packed in cartons of *six of each value.*

10	ohm	1·2K	150K
12	,,	1·5K	180K
15	,,	1·8K	220K
18	,,	2·2K	270K
22	,,	2·7K	330K
27	,,	3·3K	390K
33	,,	3·9K	430K
39	,,	4·3K	470K
43	,,	4·7K	560K
47	,,	5·6K	680K
56	,,	6·8K	820K
68	,,	8·2K	1M
82	,,	10K	1·2M
100	,,	12K	1·5M
120	,,	15K	1·8M
150	,,	18K	2·2M
180	,,	22K	2·7M
220	,,	27K	3·3M
270	,,	33K	3·9M
330	,,	39K	4·3M
390	,,	43K	4·7M
430	,,	47K	5·6M
470	,,	56K	6·8M
560	,,	68K	8·2M
680	,,	82K	10M
820	,,	100K	12M
1K	,,	120K	15M

All the above values are available in both ½ watt, 1 watt and 2 watt versions.
***Special for Philips TV's:**
8·2M 2-watt, 23p per pack.
Price ½ watt ·10, 1 watt ·13, 2 watt ·23

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Standard spindle with flat.

Double pole switch	£0·25
Without switch	£0·19

(One per pack)
5K, 10K, 25K, 50K, 100K, 250K, 500K, 1 meg, 2 meg.

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REPLACEMENT STYLI

TC8	£0·23
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OA2	0·30
OB2	0·30
OZ4	0·25
1A3	0·23
1A5	0·25
1A7GT	0·33
1B3GT	0·35
1D5	0·38
1D6	0·48
1FD1	0·33
1G6	0·30
1H5GT	0·33
1L4	0·13
1LD5	0·30
1LN5	0·40
1N5GT	0·37
1R5	0·26
1S4	0·22
1S5	0·20
1U4	0·29
1U5	0·48
2D21	0·35
3A4	0·25
3B7	0·25
3D6	0·19
3Q4	0·38
3Q5GT	0·35
3S4	0·25
3V4	0·32
5R4GY	0·53
5V4G	0·34
5Y3GT	0·26
5Z3	0·45
5Z4G	0·34
6/30L2	0·55
6A8G	0·33
6AC7	0·15
6AG5	0·25
6AK5	0·25
6AK6	0·30
6AM6	0·17
6AM8A	0·50
6AN8	0·49
6AQ5	0·22
6AR6	1·00
6AT6	0·18
6AU6	0·20
6AV6	0·28
6AW8A	0·54

6AX4	0·39
6B8G	0·13
6BA6	0·20
6BC8	0·50
6BE6	0·21
6BH6	0·43
6BJ6	0·39
6BQ5	0·22
6BQ7A	0·38
6BR7	0·79
6BR8	0·63
6BS7	1·25
6BW6	0·72
6BW7	0·54
6BZ6	0·31
6C6	0·19
6C9	0·73
6C17	0·63
6CB6A	0·26
6CD6G	1·08
6CG8A	0·50
6CH6	0·38
6LC6	0·43
6CL8A	0·50
6CU5	0·30
6CW4	0·63
6D6	0·15
6DE7	0·50
6DT6A	0·50
6EW6	0·55
6F1	0·59
6F6	0·63
6F6G	0·25
6F13	0·33
6F14	0·42
6F15	0·65
6F18	0·45
6F23	0·68
6F24	0·68
6F25	0·54
6F28	0·60
6F32	0·15
6GH8A	0·50
6GU7	0·50
6H6GT	0·15
6J5G	0·19
6J5GT	0·29
6J6	0·18
6J7G	0·24

6J7GT	0·38
6K7G	0·10
6K7GT	0·23
6K8G	0·16
6L1	0·98
6L6GT	0·39
6L7	0·38
6L18	0·44
6L19	1·38
6LD20	0·48
6N7GT	0·40
6P28	0·59
6Q7	0·43
6Q7G	0·27
6R7	0·55
6R7G	0·35
6SA7GT	·35
6SA7	0·35
6SC7GT	0·33
6SG7GT	·33
6SH7	0·53
6SJ7	0·35
6SK7GT	·23
6SQ7GT	·28
6U4GT	0·60
6U7G	0·53
6V6G	0·17
6V6GT	0·30
6X4	0·20
6X5GT	0·25
6Y6G	0·55
6Y7G	0·63
7B6	0·58
7B7	0·32
7C6	0·30
7F8	0·88
7H7	0·28
7S7	0·65
7V7	0·25
7Z4	0·50
9BW6	0·50
9D7	0·78
10C1	1·25
10C2	0·49
10DE7	0·50
10F9	0·45
10F18	0·35
10LD11	0·53
10P13	0·54

10P14	1·08
12A6	0·63
12AC6	0·40
12AD6	0·40
12AE6	0·48
12AT6	0·23
12AT7	0·16
12AU6	0·21
12AU7	0·18
12AV6	0·28
12AX7	0·22
12AY7	0·68
12BA6	0·30
12BE6	0·30
12BH7	0·27
12E1	0·85
12J7GT	0·33
12K5	0·50
12K7GT	·34
12Q7GT	0·28
12SA7GT	0·40
12SC7	0·35
12SG7	0·23
12SH7	0·15
12SJ7	0·23
12SK7	0·24
12SQ7GT	0·50
14H7	0·48
14S7	0·75
19AQ5	0·24
19G6	0·50
19H1	2·00
20D1	0·49
20D4	1·05
20F2	0·65
20L1	0·98
20P1	0·50
20P3	0·79
20P4	0·89
20P5	1·00
25A6G	0·29
25L6G	0·20
25Y5	0·38
25Y5G	0·43
25Z4G	0·28
25Z5	0·46
25Z6GT	0·43

30C1	0·28
30C15	0·60
30C17	0·77
30C18	0·60
30F5	0·65
30FL1	0·60
30FL2	0·60
30FL12	0·69
30FL14	0·68
30L1	0·29
30L15	0·58
30L17	0·67
30P4MR	·95
30P12	0·69
30P16	0·30
30P19/30P4	0·58
30PL1	0·59
30PL13	0·75
30PL14	0·65
30PL15	0·87
35A3	0·48
35A5	0·75
35D5	0·70
35L6GT	0·42
35W4	0·23
35Z3	0·50
35Z4GT	0·24
35Z5GT	0·30
50B5	0·35
50C5	0·32
50CD6G	2·17
50EH5	0·55
50L6GT	0·45
72	0·33
85A2	0·43
85A3	0·40
90AG	3·38
90AV	3·38
90CG	1·70
90CV	1·68
90C1	0·59
150B2	0·58
301	1·00
302	0·83
303	0·75
305	0·83
306	0·65
807	0·59

1821	0·53
5702	0·80
5763	0·50
6060	0·30
7193	0·53
7475	0·70
A1834	1·00
A2134	0·98
A3042	0·75
ACO44	1·16
AC2/PEN	0·98
AC6PEN	·38
AC2/PEN/DD	0·98
AC/PEN(7)	0·98
AC/TH1	0·50
AC/TP	0·98
AL60	0·78
ARP3	0·35
ATP4	0·12
AZ1	0·40
AZ31	0·46
AZ41	0·53
B36	0·33
CL33	0·90
CV6	0·53
CY1C	0·53
CY31	0·31
D63	0·25
DAC32	0·33
DAF91	0·20
DAF96	0·33
DD4	0·53
DF33	0·37
DF91	0·14
DF96	0·34
DF97	0·63
DH63	0·27
DH76	0·28
DH77	0·18
DH81	0·56
DH107	0·90
DK32	0·33
DK40	0·55
DK91	0·26
DK92	0·35
DK96	0·35

DL33	0·35
DL92	0·25
DL94	0·32
DL96	0·35
DM70	0·30
DM71	0·38
DW4/500	0·38
DY87/6	0·24
DY802	0·35
E80F	1·20
E83F	1·20
E88CC	0·60
E92CC	0·40
E180F	0·90
E182CC	1·00
E1148	0·53
EA50	0·18
EA76	0·88
EABC80	·30
EAC91	0·38
EAF42	0·48
EB34	0·20
EB91	0·10
EBC41	0·48
EBC81	0·29
EBC90	0·18
EBC91	0·28
EBF80	0·30
EBF83	0·38
EBF89	0·27
EBL21	0·60
EC54	0·50
EC86	0·59
EC88	0·59
EC90	0·28
EC92	0·34
ECC32	1·50
ECC33	1·50
ECC40	0·60
ECC81	0·16

ECC82	0·19
ECC83	0·22
ECC84	0·28
ECC85	0·24
ECC86	0·40
ECC88	0·35
ECC189	0·48
ECC804	0·55
ECC807	1·70
ECF80	0·27
ECF82	0·26
ECF86	0·64
ECF804	2·10
ECH21	0·63
ECH42	0·60
ECH81	0·27
ECH83	0·39
ECH84	0·34
ECL80	0·30
ECL82	0·30
ECL83	0·52
ECL84	0·54
ECL85	0·54
ECL86	0·35
ECL86	0·35
EE80	0·60
EF22	0·63
EF36	0·33
EF37A	0·45
EF39	0·40
EF40	0·49
EF41	0·58
EF42	0·33
EF54	0·98
EF73	0·75
EF80	0·22
EF83	0·48
EF85	0·26
EF86	0·29
EF89	0·23
EF91	0·17
EF92	0·35
EF97	0·55
EF98	0·65
EF183	0·26
EF184	0·29
EFP60	0·50
EH90	0·36
EL32	0·18
EL34	0·44
EL35	1·00
EL37	0·74

EL41	0·53
EL42	0·53
EL81	0·50
EL83	0·38
EL84	0·22
EL85	0·40
EL86	0·38
EL91	0·23
EL95	0·32
EM34	0·90
EM80	0·38
EM81	0·39
EM83	0·75
EM84	0·31
EM87	0·35
EY51	0·33
EY81	0·35
EY83	0·54
EY84	0·50
EY87/6	0·30
EY88	0·40
EY91	0·53
EZ35	0·25
EZ40	0·40
EZ41	0·42
EZ80	0·21
EZ81	0·22
EZ90	0·20
FW4/500	0·75
FW4/800	0·75
GZ30	0·34
GZ32	0·41
GZ33	0·70
GZ34	0·48
GZ37	0·67
HABC80	0·44
HL23DD	0·40
HL41DD	0·98
HL42DD	0·50
HN309	1·40
HVR2	0·53
HVR2A	0·53
IW3	0·38
IW4/350	·38
IW4/500	·38
KT2	0·25

KT8	1·75
KT41	0·98
KT44	1·00
KT63	0·25
KT66	0·80
KT74	0·63
KT76	0·63
KT81	2·00
KTW61	0·63
KTW62	0·63
KTW63	0·50
M8162	0·63
ME1400	0·74
MHL4	0·75
N78	2·05
N108	1·40
N308	0·95
N339	0·44
N359	0·44
P61	0·44
PABC80	·33
PC86	0·47
PC88	0·47
PC95	0·53
PC97	0·36
PC900	0·32
PCC84	0·29
PCC85	0·26
PCC88	0·41
PCC89	0·45
PCC189	0·48
PCC805	0·58
PCC806	0·67
PCF80	0·28
PCF82	0·30
PCF84	0·40
PCF86	0·44
PCF87	0·77
PCF200	0·67
PCF800	0·60
PCF801	0·29
PCF802	0·40
PCF805	0·60
PCF806	0·57
PCF808	0·68
PCH200	0·62
PCL82	0·32
PCL83	0·58
PCL84	0·34
PCL805/85	0·40

PCL86	0·38
PCL88	0·65
PCL800	0·75
PCL801	0·59
PEN4DD	1·38
PEN45	0·35
PEN45DD	0·75
PEN46	0·20
PEN453DD	0·98
PENA4	0·98
PEN/DD/4020	0·88
PFL200	0·52
PL33	0·38
PL36	0·47
PL38	0·90
PL81	0·44
PL81A	0·48
PL82	0·30
PL83	0·32
PL84	0·30
PL302	0·56
PL504/500	0·62
PL505	1·30
PL508	0·90
PL509	1·30
PL802	0·75
PM84	0·34
PX4	1·16
PX25	0·98
PY33/2	0·50
PY80	0·33
PY81	0·24
PY82	0·24
PY83	0·26
PY88	0·32
PY301	0·58
PY500	0·95
PY900	0·33
PY801	0·33
PZ30	0·48
QQV03/10	1·20
QS75/20	0·63
QS95/10	0·49

QS150/15	0·63
QV04/7	0·63
R10	0·75
R11	0·98
R16	1·75
R17	0·88
R18	0·50
R19	0·30
R20	0·56
R52	0·34
RG1/240A	1·98
RK34	0·38
SP42	0·75
SP61	0·33
TH4B	0·50
TH233	0·98
TP2620	0·98
UABC80	0·30
UAF42	0·49
UBC41	0·45
UBC81	0·40
UBF80	0·30
UBF89	0·30
UBL21	0·55
UC92	0·35
UCC84	0·33
UCC85	0·34
UCF80	0·33
UCH21	0·60
UCH42	0·60
UCH81	0·30
UCL82	0·33
UCL83	0·48
UF41	0·50
UF42	0·60
UF80	0·35
UF85	0·34
UF86	0·63
UF89	0·27
UL41	0·54
UL46	0·88
UL84	0·31
UM80	0·33
UR1C	0·53
UU5	0·38
UU9	0·40
UU12	0·22
UY1N	0·50

UY21	0·55
UY41	0·38
UY85	0·25
U10	0·45
U12/14	0·38
U16	0·75
U17	0·35
U18/20	0·75
U19	1·73
U22	0·39
U25	0·64
U26	0·56
U31	0·30
U33	1·50
U35	0·83
U37	1·75
U45	0·78
U47	0·64
U49	0·56
U50	0·26
U76	0·24
U78	0·20
U107	0·92
U191	0·58
U251	0·65
U281	0·40
U282	0·40
U301	0·40
U403	0·33
U404	0·38
U801	0·93
U4020	0·38
VP13C	0·35
VP23	0·40
VP41	0·38
VR105	0·33
VT61A	0·35
VU111	0·44
VU120	0·60
VU120A	·60
VU133	0·35
W76	0·34
W107	0·50
W729	0·60
X41	0·50
X61M	0·60
X63	0·33
XE3	5·00
XH/1·5	·48
Z329	0·65
Z749	0·68

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TELEVISION

SERVICING·CONSTRUCTION·COLOUR·DEVELOPMENTS

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NOVEMBER 1971

PLAYING AT TELEVISION?

It is rather easy for those whose main interest is the technical side of TV to forget that while television engineering has been progressing from 30 feeble lines of picture information to 625 lines with colour there is another side to television—the actual production of programmes.

Several years ago the field blanking interval on the UK 625-line system was increased from 20 to 25 lines to allow the insertion of additional vertical interval test signals. This was done mainly to satisfy Trade demands for colour test signals transmitted continuously throughout programmes. If the most enthusiastic television broadcast engineers were to have their own way however they would probably saturate each and every raster with test signals. But whatever one thinks of the majority of programmes such a viewpoint is clearly wrong: television as we know it is for entertainment and information.

There is a great danger that the engineering departments of the BBC and ITA are losing sight of this. With the increasing use of unmanned transmitting stations for example, engineers are justifying their title by doing a great deal more engineering than before and a great deal less programme monitoring. The danger here is a reduced sense of urgency in correcting faults affecting picture quality.

Not all the lack of purpose can be laid at the door of the engineer however and a number of production teams are open to criticism. The extent of the improvement in production techniques since the early days is difficult to measure, involving as it does subjective assessments. As technical observers however we notice an increasing use by the production side of television gimmicks for their own sake. The first of these was undoubtedly the zoom lens, which for many years was used in a very amateur manner. Fortunately that fad is now dying out, but other toys are around: slow- and stop-motion video playbacks, colour synthesis, electronic character generation, chroma-key and so on. These are fine when used with discretion, but at present they often become a production crew's monster toys.

Television is open to abuse from several directions and great care is needed to ensure that its basic purposes are not overlooked.

W. N. STEVENS, *Editor*

THIS MONTH

THE NEXT ISSUE DATED DECEMBER WILL BE PUBLISHED NOVEMBER 22

TELETOPICS

BANDS FOR SATELLITE TV

The recent World Administrative Radio Conference convened by the International Telecommunications Union has allocated Band VI, 11·7-12·5GHz, for satellite TV broadcasting in Region 1 which covers roughly Europe, Africa, the USSR and Turkey. The conference was called to consider the use of frequencies—mainly those above 10GHz—that are not at present firmly allocated. It looked—as is necessary with the current rate of technical progress—some distance into the future, going so far as to allocate frequencies for space-to-space links, for use for example in satellite-to-satellite communications, where the earth's atmospheric conditions no longer present limitations.

In allocating frequencies for satellite TV broadcasting an official distinction was drawn between broadcasting for communal reception for which relatively low transmitter powers are required and broadcasting for reception in individual homes which requires very much higher power. With this limitation in mind the conference accepted the use of the u.h.f. TV band between 620-790MHz for satellite broadcasting subject to the agreement of those countries likely to be affected. Also with this limitation the band 2·5-2·69GHz was allocated for satellite broadcasting and it was considered that this will be the main band used to provide satellite TV broadcast services in developing countries and sparsely populated regions where land-based networks would be too costly to set up.

It is proposed to hold a conference of European countries to decide how much of the 800MHz comprising Band VI to devote to regional coverage and how much to national coverage. The Band is wide enough for each W. European country to be able to operate four programmes since at s.h.f. very narrow beams can be used and channels repeated at appropriate distances. The intention is to use satellite channels initially for communal reception, with the more powerful satellites necessary for individual home reception being introduced later when this becomes technically and economically feasible. Terrestrial services in this Band have now been classified as secondary to the satellite broadcasting service to enable planning to proceed without restrictions. Once the satellite broadcasting plan is finalised individual countries will be able to see what frequencies remain available to them for terrestrial services.

Still higher up the spectrum the bands 22·5-23GHz (Region 3 only—Asia, Australasia and Oceania), 41-43GHz and 84-86GHz were allocated for broadcast use: the only use likely to be made of these bands in the foreseeable future however is for research and development purposes.

The major companies are quietly undertaking quite a bit of research work on s.h.f. TV reception: it seems likely that in 10 years or so there will be rich rewards for those able to produce reliable systems for use at these frequencies. It all shows that there is still a lot to come technically in the world of TV. On the non-technical side one can foresee battles ahead on the use to be made of these new delights when they become technically possible and how finance is to be arranged. (What to do with the fourth u.h.f. UK TV service has still to be settled however!)

The regulations accepted by the World Administrative Radio Conference come into force on January 1st 1973.

FROM THE SETMAKERS . . .

With the introduction by **Decca** of a 17in. colour model there are now two colour models with this size tube on the market. The Decca Model CS1730 is a hybrid receiver using 27 transistors, 6 valves and i.c.s in the sound i.f. and decoder sections. It is a single-standard set with a sensitivity of $100\mu V$ (minimum usable signal) and an audio output of 2W. The price quoted is £195·50. Decca have also introduced a 12in. monochrome single-standard battery/mains portable model, the MS1210.

The latest price quoted for the **Thorn/BRC** 17in. colour set is £182·90.

Philips have made a further announcement about their videocassette recorder. This is to be produced in Austria and full-scale supplies of the UK version are not expected to become available until towards the end of next year. The cassette can be erased and used again over 300 times. Philips estimate that sales of their videocassette recorder in the UK could reach 2·5 million by 1980. A rental charge of £1·50 a week has been suggested.

DISPLAY DEVICE FOR MINIATURE TV

Honeywell are producing 1in. square 0·005in. thick ceramic slices capable of storing over two billion data bits: the electrical properties of these chips are said to make them suitable for use as the display device in a miniature TV set.

STATE OF THE TRADE

Sir Jules Thorn, chairman of Thorn Electrical Industries, commenting on the 1970-71 results says: "The 1970s should see the greatest expansion ever achieved in the radio and television industry." Total home and export sales of Thorn/BRC products in 1970/71 rose by nearly 50% compared to the previous year with a substantial increase in profits. The experience of Thorn/BRC is that the high level of colour TV sales has not affected their sales of black-and-white receivers: in fact Thorn had a record year for sales of both monochrome and colour sets. On the rental side the number of colour subscribers to the group's 1,000 shops more than doubled during the year. Concern is however expressed about the increasing competition from the Far East.

The radio and television division of GEC also had a record year due to much higher colour TV set sales. For GEC monochrome set sales were marginally lower though in both categories GEC claims to have ended the year with a larger share of the market.

BREMA reports that present indications are that most setmakers will find the production of dual-standard sets uneconomic by the end of 1973 though a small but declining demand will continue for some years in those areas and pockets unable to receive u.h.f. It is estimated that by the end of this year 85% of the population should be able to receive all three programmes on 625.

The latest BREMA delivery figures, for July, show a sharp rise in deliveries of colour sets to 69,000, easily the highest monthly total yet. The "mini-budget" in which controls were eased and tax cuts made affected only the second half of the month. Monochrome deliveries rose from 90,000 in June to 110,000.

Some recently released figures on US radio and TV imports make sobering reading however. In 1970 91% of the radio sets sold in the USA were imported while 51% of the monochrome TV receivers and 18% of the colour receivers sold were imported. With the 10% US surcharge in effect Far Eastern exporters will be glancing at other potential markets. The UK TV industry may be booming just now but it doesn't look as if a sellers' market will be able to develop.

UP-CONVERTERS FOR CATV SYSTEMS

With the phasing out of dual-standard receiver production, what to do if you have a v.h.f. wired TV distribution system? With this coming problem in mind BREMA have recommended the use of an "up-converter" which accepts 625-line v.h.f. signals from the network and converts them back to u.h.f. so that a conventional single-standard receiver can be used. Teleng have now introduced such a unit, called the "Superverter". This is mains powered and provides an overall gain of 4dB. For optimum performance the input signal level is 1·2-2·5mV r.m.s. peak: if the signal level is above this it may be necessary to use an attenuator between the TV system outlet and the converter to prevent interference from unwanted programmes. Teleng say the unit should be mounted in a position where the temperature is stable and good-quality coaxial cable used.

BREMA comment that they expect this technique to be used increasingly in the future. They also expect that the majority of new coaxial distribution systems will supply signals to subscribers at u.h.f.

UHF SERVICE EXTENSIONS

The ITA has now brought into service three new main u.h.f. transmitters and the first two of its u.h.f local relay stations. The new main transmitters are: **Craigkelly,** Fife, carrying Scottish Television programmes on channel 24 (250kW e.r.p.); **Stockland Hill** near Honiton, Devon, carrying Westward Television programmes on channel 23 (250kW e.r.p.); and **Caldbeck** near Carlisle, carrying Border Television programmes on channel 28 (500kW e.r.p.). A group A receiving aerial horizontally polarised is required for all these transmissions.

The first ITA u.h.f. relay stations are at **Pendle Forest**, Lancashire, carrying Granada programmes on channel 23; and **Wharfedale,** Yorkshire, carrying Yorkshire Television programmes on channel 25. A group A receiving aerial vertically polarised is required for both these transmissions.

Both BBC-1 and BBC-2 are now being transmitted by the BBC **Wharfedale** relay station: BBC-1 is on channel 22 and BBC-2 on channel 28. The **Heathfield** (East Sussex) main station is now transmitting BBC-1 on channel 52 (group D receiving aerials horizontally polarised are required) and the BBC has now brought into operation its **Caldbeck,** Cumberland, transmitter with BBC-2 transmissions on channel 34 (receiving aerial group A with horizontal polarisation).

ADJUSTABLE-TEMPERATURE SOLDERING IRON

An adjustable-temperature, thermostatically-controlled soldering iron, the Oryx 50, has been introduced by W. Greenwood Electronic Ltd., 21 Germain Street, Chesham, Bucks. The iron can be simply set to any temperature between 200-400°C without changing the bit and the settings are accurate to ± 2%. Tip temperature variations during soldering are said to be negligible and temperature changes can be made in seconds while the iron is on. The Oryx 50 will operate at much lower temperatures than conventional uncontrolled irons. A long-life iron-coated bit is fitted as standard and there is a range of 11 bits—long-life or copper-nickel plated to choice. The iron is rated at 50W, weighs 2¾oz and heats up in 45 secs. There are 12, 24, 50, 115 or 210-250V a.c. versions.

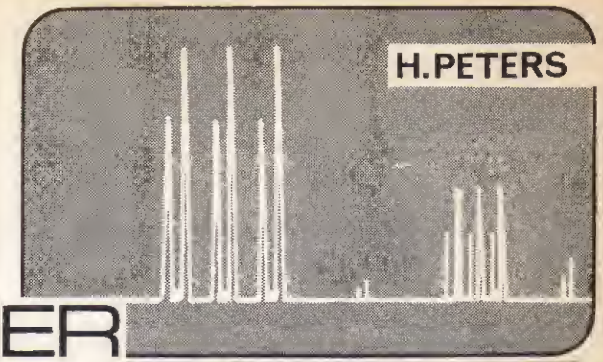
BINDERS

We have been asked by our Binding Department to point out that the binders which are available from them at 75p to hold a year's copies of TELEVISION are supplied either blank or with the volume number printed on the spine. When ordering a binder please state whether you want the binder blank or with the volume number—and which number in this case—printed on the spine. The address to which orders should be sent is: Binding Department, IPC Magazines Ltd., Carlton House, Great Queen Street, London WC2.

DX-TV PAMPHLET

Judging from the number of readers who wrote in to Roger Bunney recently for a copy of his DX-TV pamphlet there is a very considerable interest in the subject of long-distance television reception. He quickly ran out of his first batch of pamphlets and has had to run off a further supply: our apologies to all those who were kept waiting while this was being done. Any other readers who would like a copy of the pamphlet, which gives details of the various modes of long-distance TV signal propagation, the standards in use and advice on how to receive DX-TV signals, should write in to us enclosing a postal order for 15p made out to Roger Bunney.

It was obvious that a race was on between BBC-1 and ITA to be first on the air with our local duplicated services and in the true workshop tradition heavy odds were placed. What we needed was a simple device to monitor the whole band continuously: the rig about to be described does the job very well. It uses a varicap (variable-capacitance diode) tuner, an i.f. strip and an oscilloscope and displays a trace covering the entire u.h.f. band simultaneously. Once the race was over (it was a dead heat!) the unit was tidied up and put to a number of uses which will be described later on.



H. PETERS

VARICAP TUNER PANORAMIC MONITOR

The Varicap Tuner

In case you have yet to meet one, a varicap tuner (see Figs. 1 and 2) is a tuner with the four rotating gang-capactor sections replaced by four variable-capacitance diodes. By applying a reverse voltage to all these diodes together they can be made to tune the tuner from channel 21 to channel 68. Usually the voltage needed to tune channel 21 is about +1V and that required to tune channel 68 about +30V. This variable voltage is usually taken from a variable potentiometer network fed by a well-stabilised 30V line and the diodes themselves are carefully selected in matched groups since there can be no compensation for tracking errors by bending the end vanes as in mechanical tuners.

The other supplies normally connected to the tuner are about +12V, 10mA l.t. for the transistors and a forward a.g.c. voltage to the base of the r.f. transistor. In our case we are concerned with only u.h.f. but v.h.f. varicap tuners are also made: these have an extra pin connection taken to a second group of varicap diodes inside the tuner which act as switching diodes to determine whether the tuner works on Band I or Band III, switching being accomplished by connecting the extra pin to the 12V line. Various tuners —all u.h.f.—gave satisfactory results in our circuit.

Panoramic Monitor Circuit

The panoramic monitor circuit (see Fig. 3) has been kept intentionally vague so that you can adapt it to your own existing gear. The oscilloscope should be capable of delivering a 0-30V sawtooth from its "X Out" socket and have a Y sensitivity of 1V/cm. or better. Sweep speed can be 50Hz but is not critical.

The i.f. strip used can be almost any working unit.

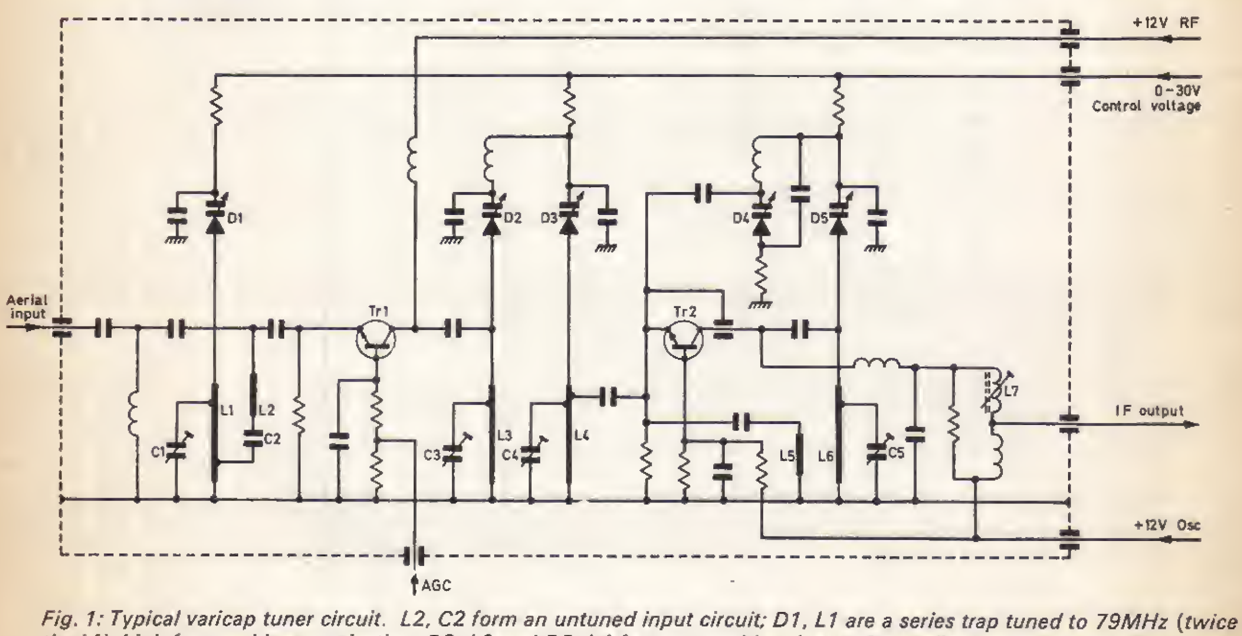
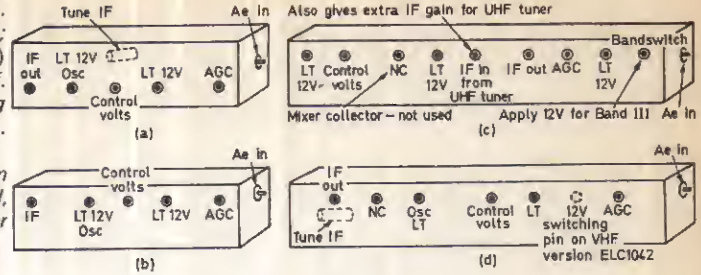


Fig. 1: Typical varicap tuner circuit. L2, C2 form an untuned input circuit; D1, L1 are a series trap tuned to 79MHz (twice the i.f.) high for good image rejection; D2, L3 and D3, L4 form a tuned bandpass r.f. coupling; D5, L6 are the local oscillator tuned circuit; L5 provides local oscillator feedback; L7 is the primary of the first i.f. transformer; D4 is a compensating capacitance-diode to maintain even oscillator output; C1, C3, C4 and C5 are tracking capacitors; Tr1 is an r.f. amplifier with a.g.c. and Tr2 the mixer/oscillator.

Fig. 2: Pin connections of common varicap tuners. (a) and (b) European u.h.f. types in current UK sets; (c) European v.h.f. type in current UK sets; (d) Mullard u.h.f. type ELC1043—the ELC1042 v.h.f. type is similar but has an extra pin for band switching to which 12V is applied to switch to Band III.

Mullard varicap tuners can be obtained from Gurney's Radio, 91 The Broadway, Southall, Middlesex. A suitable i.f. strip was featured in our July 1971 issue.



The output is taken from the vision detector diode after the i.f. filter. Provision must be made to disconnect the a.g.c. and use a manual variable bias supply. A clue as to what is needed here is to be found in the alignment instructions of the strip as the a.g.c. is normally clamped to a fixed bias during alignment.

Figures 3(a) and (b) give alternative methods of obtaining the transistor supply for the tuner from a transistorised i.f. strip. If you have an i.f. strip with a positive l.t. rail of 18-20V follow Fig. 3(a); if the l.t. rail on the i.f. strip is negative follow Fig 3(b). The tuner is wired conventionally into the i.f. strip. Most tuners require +12V and this can be dropped from the 18-20V i.f. supply using a ½W 820Ω (approximately) carbon resistor. No a.g.c. is to be applied to the tuner so the r.f. transistor is biased on to a maximum gain condition with about 2mA of emitter current by R1 and R2 (1.2kΩ and 8.2kΩ respectively).

Working Principle

The normal control voltage to the varicap diodes is replaced by the 0-30V sweep from the oscilloscope. Thus the tuner is running from channel 21 right through to channel 68 at sweep rate and in this way displays the entire band on the 'scope trace. This sweep could be a direct connection but for added versatility two potentiometers have been fitted in the feed. One varies the amount of sawtooth applied and is used as a trace expander. The other sits the sawtooth on a steady d.c. derived from the l.t. supply and this acts as a shift control. Used together these controls enable the user to zoom in on any particular section of the band for a closer look. A 1μF paper capacitor acts as d.c. block between the 'scope X outlet and the shift network. If you intend to use a live chassis i.f. strip or to run the tuner from a negative i.f. rail you must also fit a suitable blocking capacitor in the return lead.

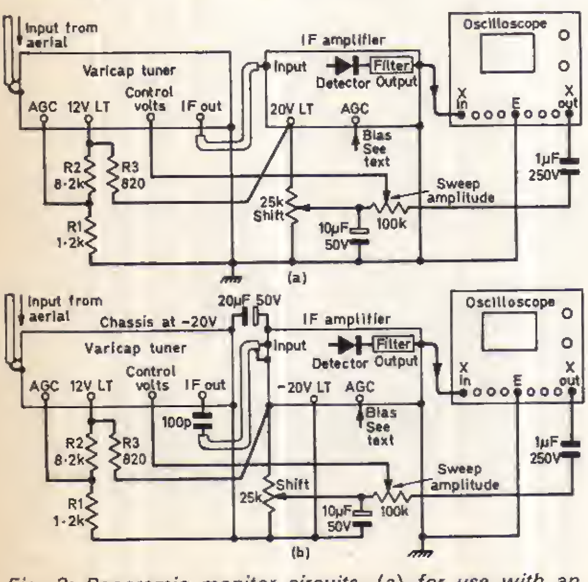


Fig. 3: Panoramic monitor circuits, (a) for use with an 18-20V positive l.t. supply and (b) for use with a −20V negative l.t. rail.

Constructional Notes

The layout is not critical: tuner, resistors, potentiometers—which may be presets—will all fit on a piece of veroboard 3×2in. Provided the l.t. polarity is right the tuner etc. can be mounted on the i.f. board itself. The tin case of the tuner has been negative in all tuners so far encountered.

The 'scope X output can be checked for adequate sweep by feeding "X out" into "Y in" with the sensitivity set at 10V/cm. You should see a diagonal line 3cm. high. Ensure that this sweep is not damped by the 100kΩ sweep control. If the 'scope is unable to deliver the right X output use a 30V p-p 50Hz a.c. mains sweep both for the tuner and the 'scope.

The i.f. strip should be peaked for the narrowest bandwidth consistent with stability. The narrower you tune it however the slower you will need to sweep or you may not receive the fine vertical line which represents a transmission. A useful unit can be made from a dual-standard i.f. strip which gives three different bandwidths by simple switching: 625 vision provides a broad square-topped pulse for ease of identification; 405 vision alignment separates adjacent sound and vision carriers and is a good general-purpose alignment; 405 sound (38.15MHz) channel gives a really narrow sweep bandwidth for looking at finer points of sound/vision ratio etc.

Calibration can easily be done using the third harmonic of a Band III v.h.f. generator, using your local station as a known starting point. For example channel 33 is 567MHz and a generator set to 189MHz (Band III channel 8) will beat against this with its third harmonic. The scale of the sweep is non-linear, stretched at channel 68 and cramped at channel 21.

Applications

The device in its crudest form in invaluable for service departments and DX enthusiasts. Periods of reduced power during test transmissions are not only spotted straight away but can be measured reasonably accurately.

Whilst we were monitoring our duplicated services

—continued on page 15

PART 3 **K. T. WILSON**

RBM COLOUR DECODER I.C.s

THE i.c.s we will be looking at this month have been developed by Rank-Bush-Murphy for use in the decoding sections of colour receivers: they are made for Rank-Bush-Murphy by the Plessey Company.

SL901 Demodulator and RGB Matrix

The SL901 has been used in Rank-Bush-Murphy colour sets since late 1968 to carry out chrominance signal demodulation and colour-difference/luminance signal matrixing, providing RGB outputs which after further amplification are used to drive the cathodes of a shadowmask tube. It is encapsulated in a twenty-pin flat pack with tabs for heatsinking and mounting and contains the equivalent of 27 transistors, two diodes and 31 resistors. It requires a supply voltage of +18V, taking a typical current of 32mA.

A block diagram showing the functions performed by this i.c. is shown in Fig. 1. The Y (luminance) signal is generally fed in at pin 10 to an emitter-follower whose output is available at pin 9. If—as in RBM chassis—brightness control pedestal pulses are to be fed into the i.c. these, along with the output at pin 9, are fed to a second emitter-follower via pin 8. If these pulses are not to be fed into the i.c. the luminance signal can be fed directly to pin 8 via a blocking capacitor. The output from this second emitter-follower is then fed internally to the colour-difference/luminance signal matrixing stages from which the RGB outputs are derived. These are available at pins 13, 12 and 7 respectively. (The RBM brightness control pulse system was described in the September 1971 instalment of *Colour Receiver Circuits*.)

The V and U chrominance signals from the PAL delay line circuit are fed to balanced synchronous detectors at pins 18 and 3. The demodulated V (R−Y) signal is fed to the red matrix and also to the matrix which recreates the G−Y signal; likewise the demodulated U (B−Y) signal is fed to the blue and the G−Y matrices. Pins 15, 6 and 14 enable external controls to be connected to preset the levels of the R−Y, B−Y and G−Y signals. Second-harmonic rejectors are also connected to these pins. The balanced synchronous detectors require push-pull reference signal inputs and these are fed in at pins 16, 17 and 5, 4. Pins 1 and 2 are connected to an internal temperature-compensated bias system and are decoupled by external capacitors.

Internal Circuit

The complete internal circuit of the SL901 is shown in Fig. 2. The two luminance emitter-followers are at the extreme left-hand side. Tr8, Tr9 and Tr10 form the blue matrix and output section. The luminance signal is fed to Tr9 and the demodulated B−Y signal to Tr8: the resultant blue output signal is taken from Tr10 emitter. Similar circuits are used to obtain the red and green output signals.

Each balanced synchronous demodulator employs seven integrated transistors, Tr1-Tr7 being the B−Y demodulator. Readers will notice the similarity between this and the detector circuit described in Part 1 of this series. There are again three long-tailed pair circuits, Tr1 and Tr2, Tr3 and Tr4, Tr5 and Tr6. Fixed bias is applied to the bases of Tr1, Tr4, Tr6 and Tr7; the U chrominance signal is applied to Tr5 base and the U reference signal in push-pull to the bases of Tr1 and Tr2/3. The demodulated B−Y signal developed across R1 is fed to Tr8 in the blue matrix/output section. An output in opposite phase is taken from the collectors of Tr1 and Tr3 to the G−Y matrix which consists of R2 and R3.

External Circuitry

Figure 3 shows typical external circuitry for the SL901 when used in a PAL receiver (this i.c. can also be used in NTSC receivers). Transformers with centre-tapped secondaries (earthed to signal by 0.1μF capacitors) provide the push-pull reference signal inputs required by the balanced synchronous demodulators. The delayed and undelayed chrominance signal (U+V) is separated into the U and V components in the usual adding and subtracting matrix and the outputs are balanced by means of the two 10kΩ potentiometers. A bias voltage from pin 2 of the i.c. is applied to the centre-tap of the add/subtract

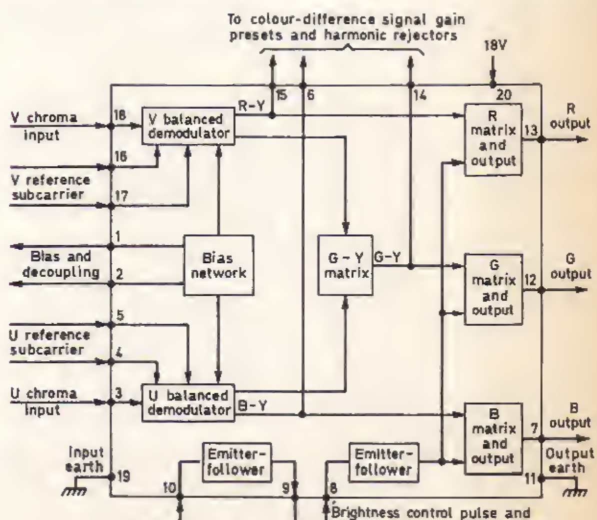


Fig. 1: Block diagram of the SL901 chrominance demodulator and RGB matrix i.c.

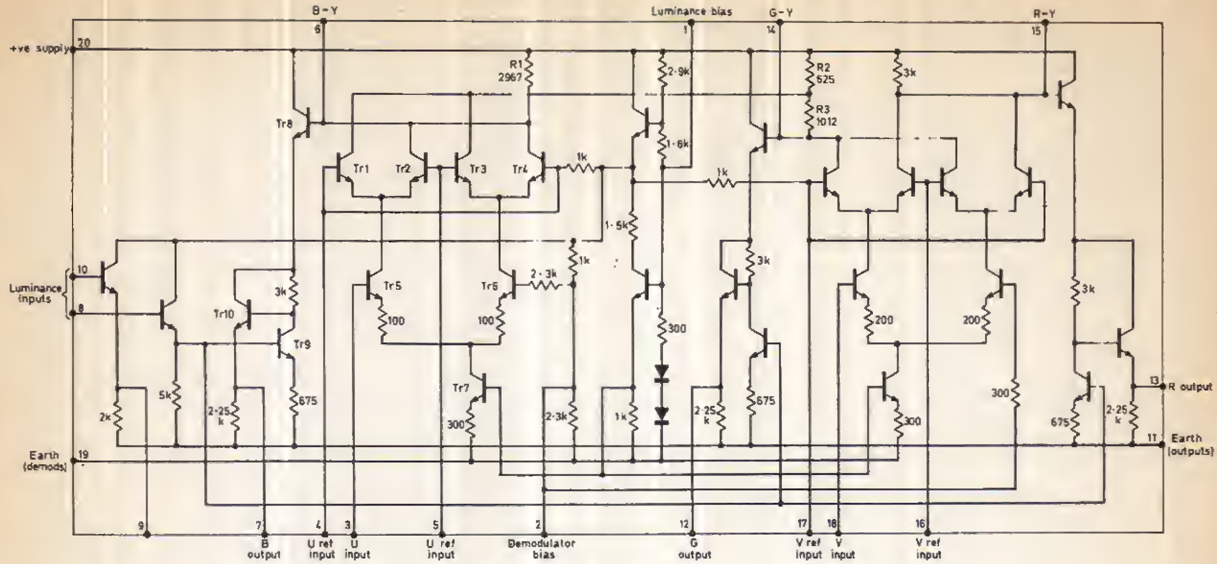


Fig. 2: Internal circuit of the SL901 demodulator/matrix i.c.

transformer winding. This pin is decoupled by an $0.1\mu F$ capacitor to the same earthing point to which pin 19 is connected. The separation of input and output earthing points is fairly common in i.c. work and must be carefully observed. The luminance signal is a.c. coupled to pin 10 along with a bias of 2-6V. The output from pin 9 is fed to pin 8 along with the brightness control pulses. A bias voltage from pin 1 is also fed to this pin.

The RGB outputs are taken via low-leakage electrolytic coupling capacitors—$6.4\mu F$ in the RBM chassis—and preset drive controls to the external

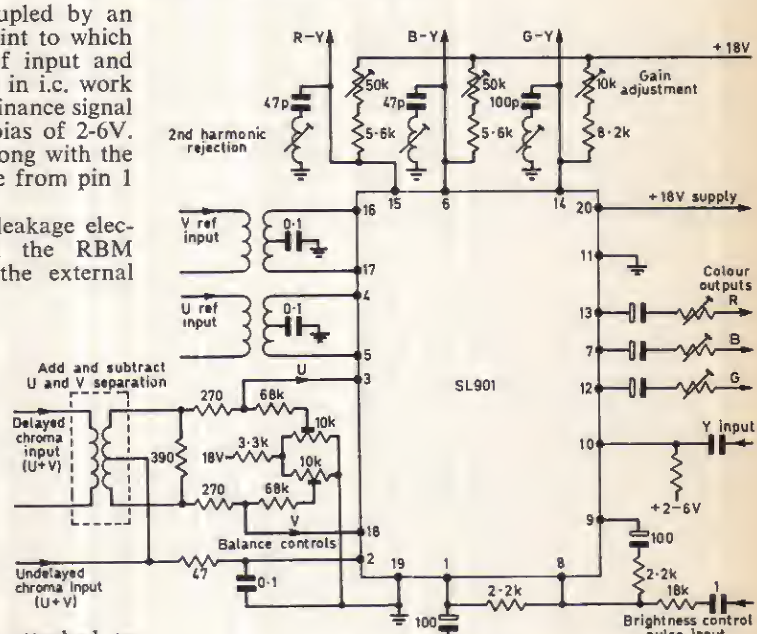


Fig. 3 (right): External circuits for use with the SL901. The usual PAL delay line circuitry separates the U and V chrominance signals which are then fed into the i.c. at pins 3 and 18 respectively. The demodulators require balanced reference input signals which are fed in at pins 16 and 17 (V reference signal) and 4 and 5 (U reference signal). The luminance signal can be fed in at pin 10 or pin 8 (see text).

RGB output amplifiers. The networks attached to pins 6, 14 and 15 are for gain adjustment and second harmonic rejection in the three colour-difference signal channels. Rejection is carried out by series resonant traps which short out signals at the second harmonic of the subcarrier frequency. The source of this harmonic is the switching action of the demodulators: a good comparison is the 100Hz hum in a full-wave rectifier operating from a 50Hz supply.

As with any i.c. there is a fairly large gain tolerance for the amplifying stages. The maximum gain tolerance of the SL901 is $\pm 25\%$ and the contrast and colour controls in other parts of the receiver must be able to take up this variation. Similarly the colour drive controls at pins 7, 12 and 13 should be able to compensate for a possible 16% variation between any two of the output signals. The presets used at pins 6, 14 and 15 to set the gains of the three colour-difference signals should be able to compensate for gain differences of $\pm 16\%$.

The SL901B

In the latest series of RBM single-standard colour models the SL901 has been replaced by the SL901B. This is different externally in being encapsulated in a 24-pin flat pack.

The SL917A

Along with the SL901B the latest Rank-Bush-Murphy decoder uses a second i.c. into which most of the rest of the decoder circuitry has been integrated. This i.c., the SL917A, is also encapsulated in a 24-pin flat pack and carries out the chrominance

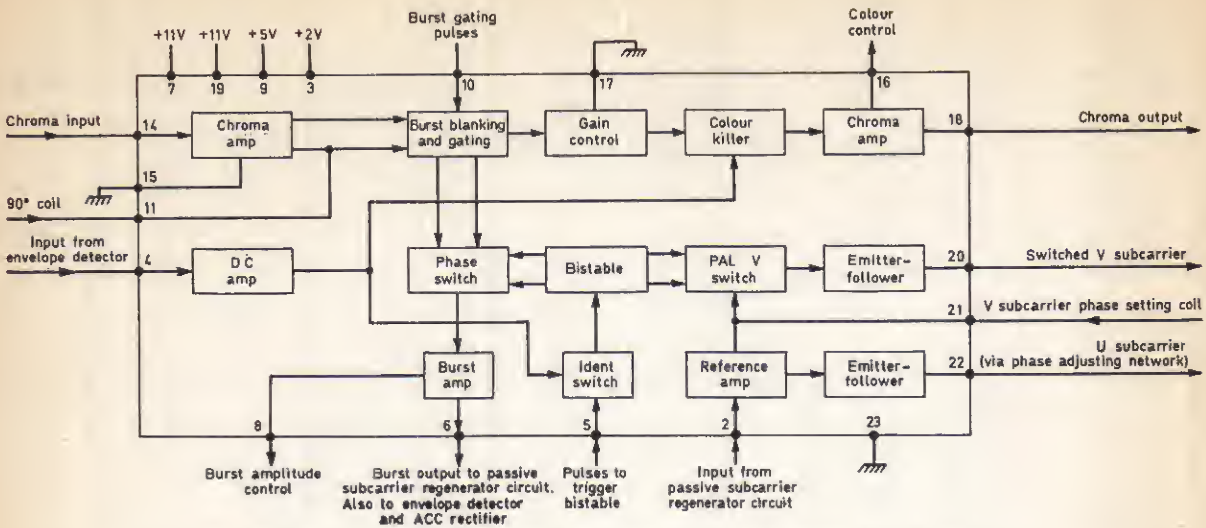


Fig. 4: Block diagram of the SL917A chrominance and burst signal processing i.c.

and burst signal processing operations, providing a chrominance signal output to drive the PAL delay-line circuitry and the U and V reference subcarriers required by the demodulator/matrixing i.c. The SL917A contains the equivalent of 51 transistors, five diodes and 50 resistors. The operations it carries out are indicated in the block diagram shown in Fig. 4. Before describing these operations however we must briefly outline one or two techniques not previously described in these pages used by RBM in their single-standard colour sets.

Passive Subcarrier Regenerator

The main difference between the RBM single-standard and most other UK decoders lies in the use made of the burst signal. Instead of being amplified and fed to a phase discriminator controlling the phase and frequency of a crystal oscillator—the usual technique—RBM use the burst signal directly to generate the reference signal in what is known as a passive subcarrier regenerator. Briefly what happens is that the bursts are amplified and the PAL alternate line burst swing then removed in order to obtain a constant-phase burst signal. This is then fed to a 4.43MHz crystal filter circuit: the crystal has an extremely high *Q* and oscillates throughout the line period until the next burst arrives. This signal after amplification is then used as the reference subcarrier required for the synchronous demodulators.

Looking at this in a little more detail, the burst occurs once each line and thus consists of a carrier

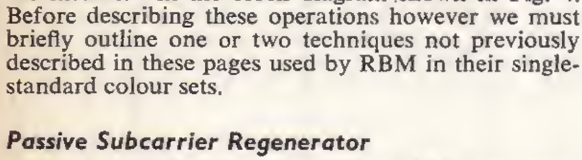


Fig. 5: The burst signal consists of 10 cycles of the subcarrier occurring once each line. It is thus equivalent to a 4·43MHz carrier modulated by a pulse at line frequency, giving the spectrum shown above. Sidebands occur at half line frequency, i.e. 7·8kHz, as well because of the burst alternations in the PAL system.

modulated by a pulse at line frequency. The frequency spectrum is shown in Fig. 5: in addition to the wanted carrier there are sidebands at 15.625kHz intervals. The PAL burst alternations also modulate this signal so that sidebands appear at 7.8kHz intervals also. The action of the crystal filter is to remove the sidebands so that the wanted carrier only is obtained. Crystals with a response sharp enough—as shown—to remove the 15.625kHz sidebands are readily available. To obtain sufficient sharpness to remove the first pair of 7.8kHz sidebands as well is less easy to do economically. It is for this reason that the PAL alternating bursts are converted to a constant-phase signal prior to being applied to the crystal filter circuit.

The basic passive subcarrier regenerator circuit is shown in Fig. 6. The bursts, after the usual gating and amplification, are fed to a tuned circuit L1, C1 which drives the crystal. The crystal behaves as a series acceptor circuit with extremely sharp tuning so that only the 4.43MHz subcarrier appears at its output. Circuit stray capacitance and the capacitance formed by the crystal material with the two connecting electrodes present a problem because this capacitance, being in parallel with the crystal, effectively bypasses it. Thus without neutralisation sideband components would appear at the output. Neutralisation is effected by centre-tapping L1 so that outputs in opposite phase are obtained at each end. The neutralising trimmer is then adjusted to be of value equivalent to the stray capacitance across the crystal so that off-resonance signal components cancel, the output obtained from the circuit being that due to the crystal alone—a pure 4.43MHz sine-wave.

Signal Paths in the IC

Returning to Fig. 4 we see that the chrominance signal undergoes the usual processing—the signal path is shown across the top of the block diagram. The chrominance signal—taken from the chrominance i.f. strip in the RBM chassis—is fed in at pin 14. After initial amplification the bursts are blanked out, gain control and colour killing actions are then under-

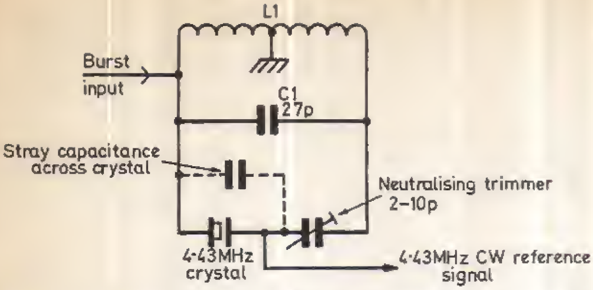


Fig. 6: Basic passive subcarrier regenerator circuit.

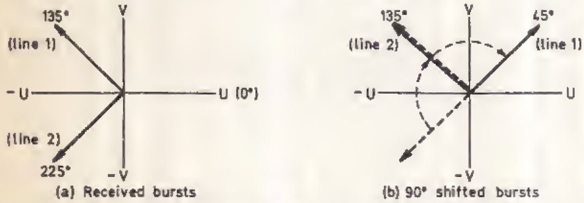


Fig. 7: The two burst feeds to the phase switch. (a) Received bursts. (b) Bursts shifted by the action of the 90° coil to swing ±45° about the V axis.

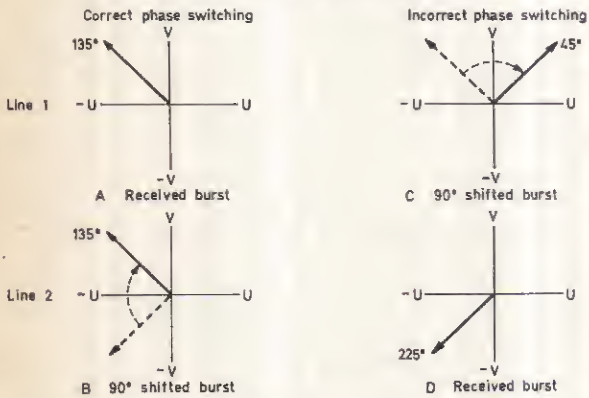
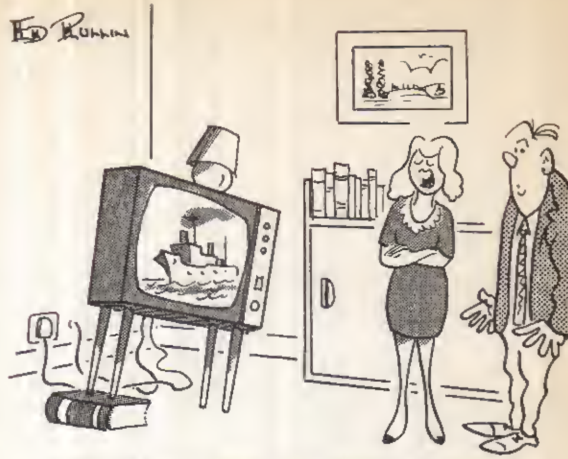


Fig. 8: When the phase switching is correct a constant-phase burst signal output is obtained; when the switching is incorrect no output is obtained because the bursts on alternate lines cancel.

taken and after further amplification a chrominance signal of amplitude suitable to drive a PAL delay-line circuit is obtained at pin 18. The colour control is linked to the final amplification section via pin 16.

The burst gating/blanking section also provides two gated burst feeds to the phase switch. One of these carries the normal PAL burst signal, alternating ±45° with respect to the −U axis as shown in Fig. 7(a). The other feed consists of a signal phase shifted by 90° as a result of the action of the 90° coil connected to pin 11 (this coil tunes with the internal capacitance of the i.c.). As a result of this phase shift the bursts in this feed swing ±45° with respect to the V axis as shown in Fig. 7(b). The purpose of the phase switch is to remove the PAL burst phase alternations so that a constant-phase output is fed to the burst amplifier and hence via pin 6 to the passive subcarrier regenerator previously described. If the phase switch is correctly synchronised it will on line 1 pass the 135° burst shown at A in Fig. 8 while on the next line it will pass the 90° shifted burst shown at B. These outputs add.



"The TV man came to straighten the picture."

If on the other hand the switch is incorrectly synchronised it will pass the 90° shifted burst shown at C on line 1 and the 225° received burst shown at D on the next line: this time the outputs, being in opposite phase, cancel.

The subcarrier produced by the external crystal circuit is fed back into the i.c. at pin 2. A reference amplifier section provides two outputs. A subcarrier for the U synchronous demodulator is obtained via an emitter-follower at pin 22: its phase is adjusted by an external tuned circuit. The subcarrier for the V synchronous demodulator is obtained at pin 20 via a 0-180° PAL switch and an emitter-follower. Phase adjustment of the V subcarrier is carried out by an external coil connected to pin 21. The PAL V switch is driven by an integrated bistable circuit which also drives the phase switch so that the two operate in synchronism. The bistable is triggered in the usual way by pulses derived from the line output stage.

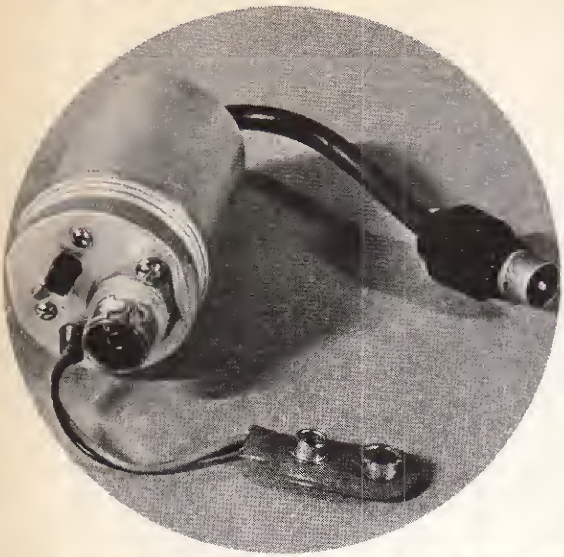
The only actions which we have not covered so far are the ident feature and the colour killer which are operated from the same source. This depends on the burst phase switching. We have seen that when this is correct the output signals obtained from the phase switch add while if it is incorrect they cancel. We get an output from pin 6 therefore only when the bursts are present and the phase switching is correct. An envelope detector connected to pin 6 will thus give an output when the bursts are present and the switching is correct and no output otherwise. The output obtained from this detector is fed in at pin 4 to a d.c. amplifier and is used as the colour killer turn-on bias and as a bias to switch on the ident switch. A short time-constant is required in the ident switch circuit: the usual action then occurs, the bistable missing a count if its initial switching—which starts when the circuit is first powered—is not in the correct sequence to suit the transmitted PAL V signal alternations. The burst output at pin 6 of the i.c. is also used in the RBM chassis for a.c.c. purposes, being detected and used to control the gain of the chrominance i.f. amplifier.

Acknowledgement

The author gratefully acknowledges the help given by Rank-Bush-Murphy in the preparation of this article.

TO BE CONTINUED

Simple UHF



Aerial Preamp

ROGER BUNNEY

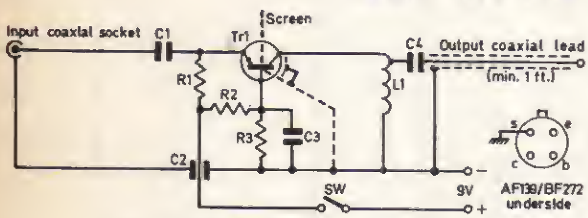


Fig. 1: Circuit of the aerial preamplifier.

★ components list

Resistors:
R1 1kΩ
R2 3·9kΩ
R3 10kΩ
All ¼W 10%

Capacitors:
C1 4·7pF miniature silver mica
C2 1000pF feedthrough
C3 2000pF ceramic
C4 3·9pF silver mica

Transistor:
Tr1 AF139 or BF272

Coil:
L1 2½ turns 0·048in. copper wire (from low-loss coaxial cable—inner conductor) wound ¼in. diameter spaced over ½in., tapped at ½ turn. (Group A—see text for Band V coil.)

Miscellaneous:
Subminiature switch (Japanese d.p.d.t. slide type); surface-mounting coaxial socket; coaxial plug; 35mm. metal film can; PP3 battery connector; etc.

An improved transistor for u.h.f. use, the BF272, a low-noise pnp silicon planar type, is now becoming available at reasonable prices and we have decided to feature this in a simple, basic aerial preamplifier which should present little problem in construction. Of special interest is that the circuit will work equally well with the earlier AF139 transistor which is at present available at very small cost: consequently it should be possible to build this amplifier for well under £1.

Circuit

Figure 1 shows the circuit of the amplifier. Signals are fed to Tr1 emitter via C1, R1 providing the emitter bias. The base bias components R2, R3 are decoupled at r.f. to chassis by C3. The collector circuit is tuned by L1, the output being tapped via C4 into a length (1ft.min.) of coaxial feeder the end of which is terminated by a coaxial plug. The feedthrough capacitor C2 gives extra decoupling for the positive supply.

Construction

The amplifier is mounted on a small subchassis (see Fig. 2) within a metal 35mm. film can. The subchassis is bolted to the lid by two 4BA bolts which also fix the input coaxial socket. The subminiature on-off switch together with the leads from the battery are also mounted on the lid. Due to the rather soft metal used for the lid care should be taken when drilling the holes for the various components. The battery leads pass through a small hole in the lid, the point where the leads pass over the metal edge of the lid being protected with a small length of PVC sleeving to avoid chaffing.

A metal screen is mounted on the subchassis, the purpose of this being to screen the transistor's input and output connections in the interests of stability. The emitter and base connections are on the input side of the metal screen and the collector and shield connections on the output side, the transistor itself being mounted in a hole cut in the screen. The shield connection solders directly on to the metal screen. The feedthrough capacitor C2 is soldered through a hole in the subchassis on the signal input side of the metal screen.

The length of the collector lead-out wire from the transistor surface to the coil is 3/16 inch. The output lead consists of coaxial cable which should be at least 1ft. long and is fixed to the subchassis by a loop of single PVC covered wire. When the wire is twisted tight to hold the lead-out coaxial cable against the chassis a liberal coating of Bostik no. 1 or some similar impact adhesive should be applied: when dry

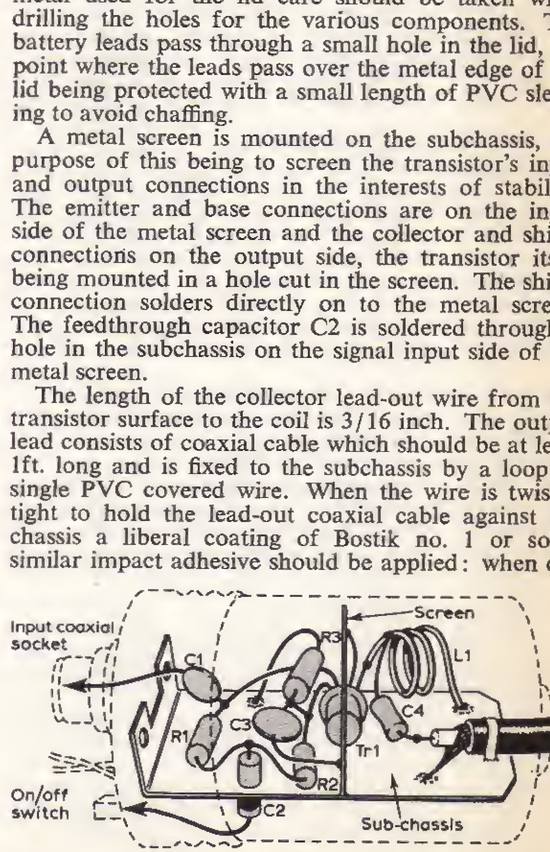


Fig. 2: Construction of the preamplifier.

the lead will be firmly held. The lead-out cable passes through a hole drilled in the main body of the film can, the hole itself being fitted with a grommet to protect the cable.

In common with all u.h.f. (and v.h.f.) amplifiers of this nature it is of the utmost importance to ensure that all connections and leads are of the shortest lengths possible in order to avoid inferior performance. The output coaxial cable must be at least 1ft. long in order to damp the output. If it is shorter patches of instability may be experienced, especially if the input signal is strong.

Operation

The coil details given in the components list are for group A channels. For group B and C channels one turn should be removed and the tapping point reduced to ½ turn from the collector. Final peaking of the tuning should be done by slightly closing or spreading the coil turns. Note the proximity of the coil to the edge of the subchassis: when the body of the can is screwed into place the change in stray capacitance may affect the tuning, tending to lower the frequency. This should be anticipated and the adjustments made accordingly. The bandwidth should be found to be sufficient to cover the whole of each channel grouping.

The amplifier draws a current of 2.5mA at 9V. The performance of the BF272 is somewhat better than that of earlier transistors, the noise figures being of special note. The figure (typical) at 800MHz is 3.5dB, at a gain of 13dB. At 500MHz a gain of 19dB is quoted. As mentioned at the start the circuit will also work with the AF139 transistor which is more readily available than the BF272 and at a somewhat lower price. The transistor type BF272 is now listed in some advertisements but in case of supply problems any ECS outlet can order same or further information may be sought from SGS (United Kingdom) Ltd., Planar House, Walton Street, Aylesbury, Bucks. ■

VARICAP TUNER PANORAMIC MONITOR

—continued from page 9

some unusual blips were noticed at the other end of the trace—see Fig. 4(a). Zooming in showed them to be weak TV signals and the reduced sound/vision spacing suggested they came from Europe. A set was tuned to these channels and sure enough locked in several test cards from Holland and Germany.

As a field strength meter it is effective but bulky. At least all the carriers can be seen at once and this helps service managers to reject aerials which give erratic responses, tilts etc., rather than letting the riggers find out the hard way among the chimney pots. The "blip" amplitude is proportional to signal input: therefore calibration is linear and with the strip biased to maximum gain a blip rising the full height of the 'scope trace would be due to a signal of around 300μV. This is an ideal range for weak signals and can be extended ×10 to become 0·3mV by fitting a Belling-Lee L729/18 −18dB attenuator at the aerial socket.

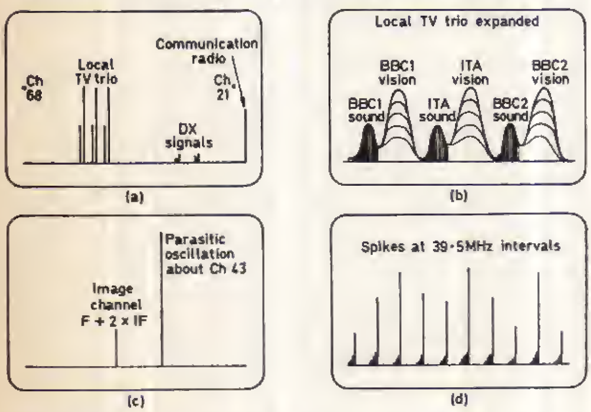


Fig. 4: Interpretation of the blips displayed. (a) General appearance of u.h.f. band swept by the tuner as seen on the 'scope. The local duplicated services are on the left. On the right can just be seen two DX signals. The blip by ch. 21 is some form of communication radio. (b) Zooming in on the local stations by using the shift and expansion controls shows the sound/vision signal ratio and a form of picture content—in this case test card F. (c) Using the probe to locate parasitic oscillations. Note the image channel which can appear at twice the i.f. (79MHz) higher. (d) The probe near the vision detector. Note the comb of spikes at 39·5MHz intervals, showing the harmonics produced by detection.

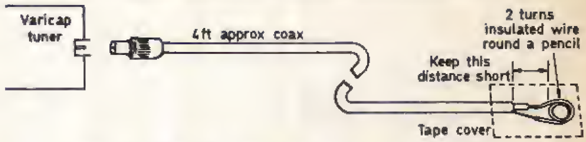


Fig. 5: Search probe for locating instability.

Tracing Parasitic Oscillations

Another useful application is for tracing parasitic oscillations in sets which display beat patterns on random channels. Modern i.f. transistors have such an improved frequency response that it is not uncommon to find one singing away to itself quite merrily at some u.h.f. frequency just because the i.f. leadouts look to the transistor just like a u.h.f. lecher bar.

The method to use here is to fit a search probe to the monitor input. This probe consists of a few feet of aerial coaxial cable with a plug at one end and a two-turn loop of stiff wire at the other (see Fig. 5). This probe will home on to your source of oscillations with uncanny accuracy. A single burst of u.h.f. instability shows as a very tall thin spike at one point on the band with a smaller one at 2×i.f. (image frequency) above it—Fig. 4(c). Harmonics of the detector produce spikes every 39.5MHz, seen on the 'scope as a comb-like trace—Fig. 4(d).

VHF Varicap Tuners

As previously mentioned v.h.f. varicap tuners are available and these can be used in the same way as the u.h.f. set-up described. The band switching is done by a 12V control potential applied to the band-switching pin. This means that if you run your 'scope in the double-beam mode by beam switching (if this facility is available on it) and can apply the beam switching voltage to the band-switch pin on the tuner it should be possible to see both Band I and Band III at the same time. ■

TRANSISTOR VIDEO

CIRCUITS

S.GEORGE

A MAJOR trend in recent colour and monochrome chassis has been the transistorisation of the video/luminance circuitry, in particular the output stage. Most c.r.t.s require a video drive in the region of 60V for a fully contrasted picture, so clearly the output stage must be operated from an h.t. supply rail well in excess of this figure in order to provide amplification of the sync pulses as well and to allow for circuit losses and tolerances.

The introduction of transistors such as the BF178 which can be operated safely from h.t. rails in excess of 200V makes possible the use of transistors in the video output stages of full-size domestic sets. Several current single-standard monochrome chassis use this transistor as the video amplifier: in the Pye 169 chassis for example it is used with a 237V h.t. rail while in the Bush-Murphy TV181S/V2016S series it is operated with a 228V h.t. rail and in the BRC 1500 chassis with a 160V. rail. Clearly therefore such a transistor video output stage is more than able to provide the peak-to-peak voltage swing required, especially when it is remembered that the collector potential of a transistor can fall to a much lower level than the anode potential of a pentode valve. Transistorised video circuits in fact usually provide greater voltage gain than their valve predecessors, the two-stage video circuit comprising a BF197 driver and BF178 output stage used in the BRC 1500 chassis for example giving an overall voltage gain about twice that obtained from a conventional valve video circuit.

As outlined in the article on h.f. video response in the August 1971 issue the load shunt capacitance has to be taken into account in calculating the value of the video output stage load resistor if the gain is to be maintained at the upper video frequencies, since the gain falls to 0.7 of the peak (i.e. medium frequency) gain when the reactance of this shunt capacitance is equal in value to the load resistor. Thus as with video pentodes the same considerations of load shunting capacitance apply and so load resistors for transistor video output stages average about 6-7kΩ.

Bandwidth is the only real restriction in determining the load resistor value for a video pentode. With transistors power dissipation determines the maximum permissible load value though the types evolved for video and luminance output stages more than satisfy such requirements. There are however two conditions which must generally be met. First the total base d.c. circuit resistance should not exceed about 1kΩ while the emitter resistor value should be at least 100 or so ohms. The former condition means that the preceding stage must have a low output impedance since it will be working into the parallel combination of this d.c. resistance plus the transistor's input impedance and this combination will probably amount to only a few hundred ohms. Vision detector load resistors average about 4-5kΩ so clearly there must be an impedance matching stage between the detector and the video output stage or the detector efficiency and the loading on the preceding i.f. stage will be completely unacceptable (for good detector efficiency its load resistor must be of high value compared with the detector diode's forward resistance while heavy loading seriously broadens the selectivity of the final i.f. stage). Before turning our attention to the impedance matching stage however let us consider transistor input impedance in more detail since it is vital at video frequencies.

The input impedance of a common-emitter stage with partially decoupled emitter resistor can be regarded as being a comparatively low-value resistor in parallel with a capacitor whose value is dependent on voltage stage gain A. Taking the resistive component first, its value approximates the emitter resistor value times the current transfer ratio of the transistor, i.e. $h\text{fe} \times R\text{e}$, and if the former is 25 and the latter 140Ω would equal 3·5kΩ. The input capacitance is rather a different matter involving many factors: Re and $h$fe as before, the emitter capacitance Ce, transistor feedback capacitance Cre, transition frequency f_T and voltage stage gain A. The feedback capacitance and voltage gain are the most important factors since together they account for the major proportion of the input capacitance by producing an amplified version of Cre across the transistor input in similar fashion to that produced by the Miller effect on the grid-anode capacitance of a triode (Miller capacitance equals Cga times $A+1$). To see the extent of this capacitance, let's consider the BF178 which has a feedback capacitance of 2·5pF. If we assume a voltage gain of 30, this produces an effective input capacitance of 2·5pF×31 or 77·5pF in addition to that of the other factors mentioned. These could probably amount to a further 20pF so that the total input capacitance at this stage gain would be close to 100pF. Clearly as the detector load shunt capacitance must be kept to the minimum possible for the same reasons as the output stage load shunt capacitance this total load capacitance of close to 100pF cannot be allowed to shunt the diode load. Thus for this reason also an impedance matching stage is essential.

For these reasons then the video output transistor is usually driven by an emitter-follower stage since this has a high input impedance, thus imposing negligible loading on the detector circuit, while its low output impedance is scarcely affected by the input characteristics of the output transistor. An emitter-follower provides a voltage gain slightly below unity but gives a current gain only slightly below that obtained from a common-emitter stage. In some recent chassis, such as the Pye 169, a TAA700 integrated circuit is used for video preamplification, sync pulse separation and a.g.c. The video emitter-follower is contained within this unit though the emitter load resistor is connected externally to reduce the overall power dissipation within the i.c.

Representative Circuit

Following this general outline let's take a detailed look at the operation of a typical modern circuit, that used in the BRC 1500 single-standard chassis and shown in Fig. 1. Diode W2 demodulates the i.f. signal fed to it via C28 from the collector of the

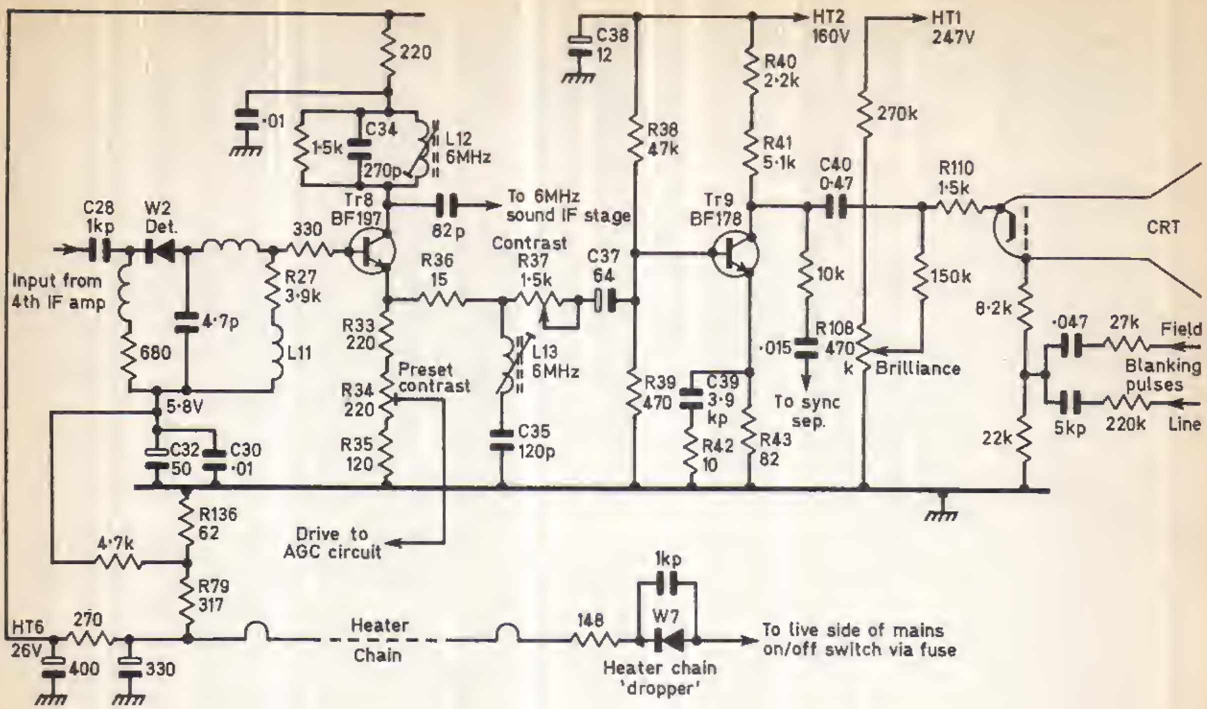


Fig. 1: Transistor video circuits of the BRC 1500 single-standard chassis.

fourth i.f. amplifier stage, producing both the video and 6MHz intercarrier sound signals across its load resistor R27 and the peaking coil L11 (to maintain h.f. response peaking coils are just as important in detector circuits as in video output stages). The detector diode output is negative-going and is superimposed on the 5·8V bias potential taken from the junction of the potential divider R136, R79 at the end of the heater chain which as is usual in current receivers uses a rectifier "dropper". C30 and C32 earth the bottom end of L11 and decouple this bias. The combined voltage is then applied as forward bias to the base of the npn emitter-follower video driver stage Tr8, becoming less positive with increase in signal strength to proportionately reduce Tr8 emitter current.

As R136 and R79 terminate the heater chain, developing 26V (HT6) across them, if either became open-circuit this voltage would rise to an excessive figure since the heater circuit continuity would only be maintained through the transistors fed from this rail. Circuit design however prevents receiver use in either of these eventualities: if R79 should go open-circuit there would be no forward bias to Tr8 and therefore no sound or vision while if on the other hand R136 should go open-circuit Tr8 base bias would be excessive resulting in a grossly over-contrasted picture which would not lock. Additionally if the BY126 heater rectifier should go short-circuit there would be no HT6 supply and again no sound or vision.

The video load of Tr8 comprises R33, R34 and R35 in series, reduction in Tr8 emitter current resulting of course in a fall in the potential developed across these resistors. The drive for the a.g.c. circuit is tapped from R34, the preset contrast control. The a.g.c. circuit is of the now usual sync-tip type, R34 thus setting the amplitude of the pulses fed to the a.g.c. circuit and consequently the a.g.c. bias developed and applied to the controlled stages in the i.f. strip.

The total emitter output is applied via R36, the contrast control R37 and C37 as drive to the video output stage Tr9, an acceptor trap L13, C35 removing the sound content from the signal. It is interesting to note the difference in the value of the base feed capacitor C37 to Tr9 (64μF) compared to that (C40, 0·47 μF) feeding the same but amplified signal to the c.r.t. cathode. This wide disparity is due to the widely differing circuit impedances at the two points and the necessity to maintain the l.f. response with the low input impedance of Tr9 (see "Video L.F. Response" in the February 1971 issue). The contrast control is effectively the top section of a signal potential divider of which the lower section comprises the input impedance of Tr9 in shunt with R38 and R39 (the h.t. end of R38 being earthed signalwise by C38). Thus reducing the resistance value of R37 increases the proportion of Tr8 output used to drive Tr9.

The video load of Tr9 comprises R40 and R41 in series a.c. coupled to the c.r.t. cathode by C40. R110 is connected in series with the cathode feed to prevent tube flashovers reaching Tr9 collector and causing possible damaging surges. R108 varies the c.r.t. cathode d.c. potential to give brilliance control. Field and line flyback blanking pulses are applied to the c.r.t. grid.

C39 and R42 provide partial emitter decoupling in the video output stage, progressively removing the negative feedback at the higher frequences (their combined low impedance then shunting R43) to maintain the h.f. response. The 6MHz intercarrier sound signal is developed across L12, C34 in the collector circuit of the driver transistor Tr8 and fed via the 82pF capacitor to the 6MHz i.f. amplifier stage. ■

PULSE SCALER/SIGNAL GENERATOR

MARTIN L. MICHAELIS M.A.

PART 2

THE output from pin 6 of each univibrator i.c. is fed to the input pin 3 of the next univibrator (see Fig. 1 last month). Response at pin 3 is produced on the 1 to 0 transition. Pin 6 rests at 0 and goes to a 1 during the pulse, so that each univibrator triggers at the end of the output pulse from the previous one. The sum of the pulse times of IC5 and IC6 is the space time between the output pulses of IC4. It is necessary to use two space time stages IC5 and IC6 instead of a single one because dead-time considerations would otherwise preclude space times much greater than the selected pulse time. The type MC851P integrated univibrator used for IC4 to IC6 possesses a dead time which is rather ill-defined on open-circuit and amounts to some 40% of the pulse time given by the capacitor connected between its pins 10 and 11. Thus if IC6 were omitted and IC5 coupled back to the input of IC4, IC4 would try to retrigger IC5 before the end of its dead time if the space time selected with S3A was longer than about 2·5 times the pulse time selected with S2. IC5 cannot respond so soon so that oscillation would cease.

Starting the Univibrators

This also underlines the fact that this circuit is not self-starting: switching the power supply on, or off and on again, will not start the closed ring with S1 in the "self-excited" setting because these actions do not produce the essential 1 to 0 transition at any one of the input pins 3 necessary to start oscillation. However there is always a logical 1 present at Tr3 emitter and thus at contact 1 of S1 provided no input signal is being fed to P1 and the trigger level control VR1 is kept turned down. Furthermore all pin 6s of the univibrators rest at logical 0 potential. Thus switching S1 from position 1 to self-excited gives a 1 to 0 transition at pin 3 of IC4 and oscillation then commences. It will generally stop if the setting of S2 or S3 is subsequently changed so that it is necessary to select the desired pulse and space times before switching to self-excited or to switch briefly to 1 and then back to self-excited if oscillation ceases.

We must of course ensure that oscillation does not cease arbitrarily and this requires sharpening the dead times of the univibrators. For this purpose direct feedback is taken from output pin 1 to input pin 4. Inputs 3 and 4 are equivalent: thus the logical 0 placed at pin 4 for the duration of its own pulse disables pin 3. Spurious signals arriving at pin 3—the chief cause of occasional erratic prolongation of the dead time—are thus in effect gated out. The capacitors between pins 2 and 5 increase the trigger coupling time-constant for inputs 3 and 4 to the maximum values tolerable in relation to the smallest pulse and space times provided so that wanted trigger transitions can get through more reliably even in the face of occasional partial dead-time barriers. By using these artifices oscillation of the ring is made extremely secure and will not cease artibrarily even over many hours of continuous operation.

The two-stage space time generator IC5 and IC6 permits any space time greater than the pulse time of IC4 to be obtained without upper limit. Readers requiring other times or longer times may modify the values of the capacitors on S3A and S3B in linear proportion. Electrolytics of unlimited size may be used, but observe correct polarity. If a continuous interpolation is desired between successive capacitor values, break the external connections between pins 9 and 14 of IC5 and IC6 and connect a 10kΩ linear potentiometer wired as a variable resistor between these pins. A tandem potentiometer is desirable for ganging the two control points involved.

The integrated circuits as they stand show a slight temperature drift of the pulse times (maximum 5%) after switching on. This is due to the temperature coefficient of the integrated timing resistor behind pin 9. For maximum frequency stability leave pin 9 disconnected and wire an 8·2kΩ fixed resistor in series with the 10kΩ potentiometer between pins 10 and 14 if interpolation control is desired.

Pulse: Space Time Ratio (Duty Cycle)

By virtue of the two-stage space time generation with IC5 and IC6 there is no upper limit imposed on the space time relative to the pulse time. Irrespective of how short the pulse time of IC4, IC5 prevents retriggering of IC6 earlier than a time interval after the end of the previous pulse from IC6 at least equal to the duration of that pulse.

But the space time may not be made shorter than about 40% of the pulse time of IC4 because otherwise IC6 would try to trigger IC4 again before the end of the latter's dead time. Oscillation would then cease. To overcome this restriction a four-stage pulse ring would be necessary with a two-stage pulse time generator. A summing gate would also be required to produce a single output pulse of duration equal to the pulse times of both pulse time stages without a gap in the centre. The considerable additional circuit complexity necessary for performing these functions was not considered to be worthwhile in view of the fact that pulse waveforms with such high duty cycles are seldom required.

If the d.c. potentials are unimportant, i.e. if external capacitive coupling is used, a high duty cycle waveform of one polarity is identical to a corresponding low duty cycle waveform of the opposite polarity: the scaler provides such complimentary output waveforms and even if the absolute d.c. levels are important—this would be the only real justification for insisting on a waveform with a duty cycle exceeding 50%—they are easily restored by using a d.c. restorer diode with the external coupling capacitor.

Every attempt has been made in designing this instrument to preserve the attractive simplicity possible with integrated circuits. The circuit is nevertheless considered to be virtually universal with respect to duty cycle ranges. The benefits obtainable using time interpolating potentiometers were not considered important enough for inclusion in the prototype, involving as they do an additional manual control.

The Trigger Amplifier

The decade counters used as IC1 to IC3 are guaranteed for counting up to 20MHz, with a typical performance limit of 30MHz. We therefore decided at the outset to design an input amplifier capable of driving the input pin 1 of the first integrated decade counter from sinusoidal or arbitrary waveform input signal frequencies up to 20MHz and with sufficient gain to give stable operation with the low input voltages obtained from very loosely coupled signal probe loops consisting of a single turn or a few turns of wire. Before commencing work on this project it was feared that rather complicated tuned input amplifiers would be necessary and that variable capacitors would have to be adjusted at least approximately to the band in which the frequency to be counted lay if this was well above 1MHz. These fears proved to be quite unfounded. A very simple aperiodic input amplifier will do the trick very well and requires no tuning controls.

Nevertheless roughly 90% of the time devoted to designing this circuit was spent in devising the best arrangement for Tr1 to Tr3 and the few components associated with these stages. Basically an input amplifier is not required at all because IC1 will operate on sinewave signals fed direct to its input pin 1 provided the frequency is high enough to ensure the minimum required trigger slope. But such a direct connection is undesirable for several reasons. Control is not stable, and the desired trigger sensitivity of about 10mV r.m.s. is not realisable over a wide range of input signal without a suitable amplifier. Even slightly excessive signal amplitudes fed direct to pin 1 of IC1 can destroy this integrated circuit, especially if going negative to chassis.

Nominally the decade counter MC838P is toggled (made to respond) at pin 1 by applying the standard 1 to 0 logic transition. This would imply that a rapid change is required from a potential greater than 2V to one less than 1V, in keeping with the definition of the logic levels in this family. Experiments revealed however that the MC838P does not require this large logic swing for toggling. It will not respond at all to a rather slow drift down from the 1 level to the 0 level but on the other hand it does not require the extremely fast dynamic transition needed to fire an MC851P univibrator i.c. at its pins 3 and 4 because the decade counter does not employ input differentiation. The actual response point lies somewhere between +1V and +2V with respect to chassis at pin 1 and it suffices to provide a reasonably fast transition from about 150mV above this point to 150mV below it to give a count. This represents a peak-peak logic swing of only about 300mV. Optimum sensitivity is obtained if the signal excursions at pin 1 of IC1 are confined to this small range around the actual response point. The necessary large d.c. potential (about 1·5V) corresponding to the position of the response point (which is subject to manufacturing tolerance) must be added to the small signal waveform in a stable and smoothly controllable manner. Above all this d.c. backing potential must be extremely efficiently smoothed, since very small hum levels would seriously impair the utilisable sensitivity.

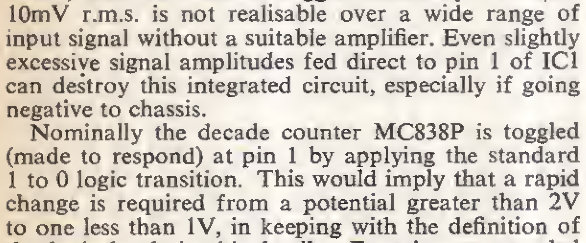


Fig. 3: Power supply circuit.

Gain and Trigger Level Control

The adjustable d.c. backing potential is provided via the trigger level control VR1. The supply voltage for the entire trigger amplifier is stabilised by a zener diode and the sample used for deriving the backing voltage is further stabilised by diodes D1 to D4 to remove the last traces of hum. C1, R4 and D5 form a d.c. restorer circuit capable of operating up to 20MHz. Any common r.f. diode is suitable for D5. The input waveform is d.c. restored positively with respect to the backing potential at VR1 slider so that the negative signal peaks lie at VR1 slider potential because D5 briefly conducts during each such negative peak to return to C1 the charge which has leaked away through R2 and R4 during the rest of the previous signal waveform cycle. It is thus possible with VR1 to lift very small signal excursions to the correct mean potential for narrow toggling across the actual response point at pin 1 of IC1. VR1 is a d.c. control which does not handle the actual signal frequency voltages, thus obviating all the problems associated with stray capacitances with r.f. gain controls.

Gain Factor

Experiments with small inductive loop signal probes connected via an r.f. detector probe to an electronic voltmeter showed that typical induced voltages between 10mV and 1V r.m.s. in the range from 100kHz to 20MHz using this kind of loose coupling minimised the oscillator loading (depending on the amplitude and Q factor of the sensed oscillator circuit). The minimum required logic swing of a properly backed signal at pin 1 of IC1 is about 100mV r.m.s. Thus a gain factor of 10 is required for the trigger amplifier to give the required maximum input sensitivity of 10mV r.m.s. This gain factor is obtained as follows: the ratio of R8 to R9 is about 5 and the ratio of R10 to R9 somewhat greater so that the positive feedback from Tr2 collector via Tr3 emitter to Tr2 emitter is insufficient to sustain oscillation or to produce a threshold trip. But this positive feedback increases the effective gain of the combination to about 10.

To cater for large input signal amplitudes as well as small ones Tr1 and Tr2 rest completely cut off when VR1 is set to zero. Tr3 rests saturated so that a potential close to the collector supply voltage is applied to pin 1 of IC1. This is a logical 1. The potential at pin 1 of IC1 cannot rise above +5V and it cannot go negative to chassis. Thus destruction of IC1 is precluded.

Large Signal Voltages

The base of Tr1 must be driven positive to chassis by at least the silicon thresholds of Tr1 and Tr2 plus the standing potential at Tr2 emitter—due to the current flowing through Tr3—before any change at all is felt at pin 1 of IC1. This minimum positive excursion required to open the amplifier is about 1·6V. In operation part of this excursion is provided by the peak-peak swing of the positively restored signal voltage appearing across R4 and the rest is made up by advancing VR1 the necessary amount. The range of VR1 is sufficient to open the amplifier in the complete absence of a signal. This is necessary because the entire backing voltage must be provided with VR1 when triggering on very small input signal amplitudes (maximum usable sensitivity setting). The larger the input signal the lower down on its track will be the proper setting for VR1.

The small tabulation on the circuit diagram (Fig. 1) shows the corresponding voltages measured at pin 1 of IC1 for respective voltages produced in the absence of any input signal at the junction of R2 and R3 by adjusting VR1. These pairs of voltages should be checked to test the proper performance of the amplifier. Evidently about +2·0V must be present at Tr1 base to drive the potential at pin 1 of IC1 to the critical response point of about +1·5V. Thus the correct setting of VR1 is at the chassis end of the track with input amplitudes of about 2V peak-peak or some 700mV r.m.s. For still greater input amplitudes it is unnecessary to use the trigger level control at all and it may be left with the slider at the chassis end. Input voltages appreciably greater than about 700mV r.m.s. nominally overload the amplifier. R7 has been added to make the circuit capable of handling inputs of up to several volts without unstable triggering. These bottom Tr2 so that R7 and R9 then function as a voltage divider with Tr2 merely behaving as a series diode. Somewhat greater input voltages bottom Tr1 too, leading to much heavier attenuation via R3, R7 and R9 as a voltage divider. This arrangement makes it unnecessary to use any form of manual gain control. Really excessive input signals are easily avoided by using an external attenuator or by avoiding placing an inductive loop probe too close to the oscillator being sensed.

Frequency Range

The trigger amplifier initially tried was a more conventional version with larger value resistors and a speed-up capacitor across R10 to steepen the threshold response flank obtained, with R8 large enough to produce sufficient positive feedback for instability. This is the classical Schmitt trigger arrangement used to produce sharp logic transitions for operating IC1 from any arbitrary input waveform which does not have such sharp transitions. This circuit, which was used for exactly the same functions in the 100kHz digital frequency meter, gave no success here. It performed excellently up to 100kHz—as in the digital frequency meter—and with considerable coaxing and poor smoothness of the trigger level control would work up to about 2MHz. The trouble was found to be due to the stray capacitances around the transistors. The methods of inductive and/or capacitive frequency compensation used in wideband oscilloscope amplifiers produced responses up to nearly 10MHz, but in a rather erratic manner and above all with such poor input sensitivity that their value was considered very dubious.

In the course of these experiments it emerged that any form of threshold trip response seemed undesirable because its inevitable associated hysteresis is greater than the smallest logic swing with which IC1 will operate, so that a tripping amplifier seriously impairs sensitivity and makes trigger level adjustment tricky and subject to severe backlash. Thus the only simple way to obtain the desired bandwidth of 20MHz is to reduce the resistor values drastically so that the inevitable stray capacitances are without significance. It was decided at the same time to reduce R8 and increase the ratio of R10 to R9 so that a trip is just no longer possible. This proved to be successful in every respect and is the final form of the circuit as published.

Experiments with tuned amplifiers were made at an early stage of the design work but were soon abandoned because they brought no clear advantage in terms of sensitivity or stability. Tuned circuits functioned—indeed up to 20MHz—but the tuning was rather critical and the manual operating procedure so tricky that a clearly formulated operating procedure would have been impossible. With the adopted simple aperiodic amplifier however the operating procedure is extremely simple.

Operating Procedure

The operating procedure in the self-excited mode should be clear from the circuit description. Similarly the scaling mode when used for producing lower pulse frequencies than the lowest available directly from standard pulse generators, or for dead-time measurements in conjunction with flutter scaling, has also been described adequately. Thus the present instructions are confined to the scale mode used for r.f. measurements in conjunction with the digital frequency meter.

It is convenient to class frequencies up to 100kHz as audio and supersonic which the digital frequency meter will handle directly, and frequencies above 100kHz as radio frequencies which require the use of the 20MHz pulse scaler. The preferred method of signal injection into the scaler is via a loosely-coupled inductive loop connected either directly to P1 and P2 or connected to the end of a piece of coaxial cable whose other end is connected to P1. A special signal probe is not necessary because the length of coaxial cable, the number of turns of the loop and the diameter of the loop are in no way critical. A conventional grid-dip meter covering the range 100kHz-20MHz is extremely useful in conjunction with the 20MHz pulse scaler for numerous types of measurements.

If for example it is desired to measure the frequency to which a radio receiver is tuned in the long, medium or shortwave bands, first determine the approximate frequency from the receiver tuning scale and then tune the grid-dip meter—coupled loosely to the aerial socket—to approximately the correct frequency (not the image frequency) according to its own scale. The grid-dip meter can then be tuned critically for zero beat frequency with the received station heard in the loudspeaker. The beat frequency should be reduced to below 1kHz corresponding to the subsequent resolution of the digital frequency reading for shortwave frequencies. Without disturbing the setting of the grid-dip meter, its coils can then be approached by a small single-turn coupling loop plugged directly into P1 and P2 of the scaler and a digital frequency reading taken as described below. This procedure will rapidly log the frequency of a shortwave station to within ± 1kHz which is fully adequate to identify it positively but quite impossible with reference to the tuning scale of any ordinary domestic receiver alone.

With a little practice the frequencies of v.h.f. radio and television transmitters can be determined in a similar manner, using harmonics of the grid-dip meter frequency to beat against the carriers. The only additional requirement in this case is to be able to identify the particular harmonic, which implies that the frequency to be measured must be known approximately to within about ± 10MHz which is usually the case because this tolerance is within the resolution of the receiver tuning scale. Television Band III frequencies will for example require up to about the tenth harmonic of the grid-dip meter operating on its highest range, so that the ultimate resolution of the digital frequency reading can theoretically be as high as 10kHz.

Digital readings of the resonant frequencies of passive circuits can be obtained analogously. This is an extremely useful facility for dimensioning peaking coils in video amplifiers and filters with much greater precision than is possible with reference to the grid-dip meter scale alone. To obtain a reading, find the dip with the grid-dip meter in the normal manner, finally coupling as loosely as possible to get the best fine tuning of the dip, and then measure the frequency of the grid-dip meter digitally.

Signals may be fed via cable directly from the output of a standard r.f. signal generator to the input socket P1. The operating procedure is the same whether the signal is injected by cable or picked up by a small inductive loop.

Always commence with the trigger level control set to zero (slider at the chassis end of the track) and select a division ratio with S1 such that the output frequency will not exceed 100kHz which is the nominal maximum input frequency for the digital frequency meter. Either P3 or P5 should be connected to the pulse input of the digital frequency meter via a short piece of coaxial cable and VR3 or VR2 set to give an output amplitude of about 5V. This is not critical. S2 must be set to 10 or 20 microseconds. The setting of S3 is immaterial because the pulse ring is inoperative in this mode.

Slowly advance the trigger level control until stable response is obtained. If the input signal level is large it will not be necessary to advance VR1 at all. If the input signal is small however VR1 requires precise adjustment which becomes the more critical the smaller the signal voltage is. For small signal voltages there will always exist a small range of settings of VR1 below *and above* which response ceases. Below this response range the backing voltage is still too small so that the signal excursions do not reach the toggle swing of IC1 whilst above the response range the backing voltage is already too large so that the small signal excursions have been pushed entirely beyond the toggle range. The range of settings of VR1 giving a response become progressively narrower as the signal amplitude is reduced so that ultimately, with only about 10mV r.m.s. input signal, response is obtained only at a spot setting near the top end of the track of VR1.

The correct procedure for ensuring that the response is secure so that all cycles of the input signal are counted without any being missed is as follows. Switch the digital frequency meter to the analogue mode giving a direct linear frequency reading on the thousands meter. Select a range on the frequency meter giving an on-scale reading with the expected (approximately known) frequency. Now adjust the trigger level control whilst observing the thousands meter giving the analogue frequency reading. With VR1 set far from the proper position there will be no reading at all. As the correct setting of VR1 is approached a reading will appear but will be far too small and fluctuating erratically because only some of the signal cycles are operating IC1 since their excursions are still askew with respect to the toggle swing of IC1. Further judicious advance of VR1 will now bring a steady reading which persists over a certain range of still further advance of VR1 beyond which it again becomes erratic and then vanishes. The correct setting for VR1 is at the centre of the range giving a steady analogue frequency reading. The digital frequency meter can then be switched to the digital mode to take the final reading in the normal manner.

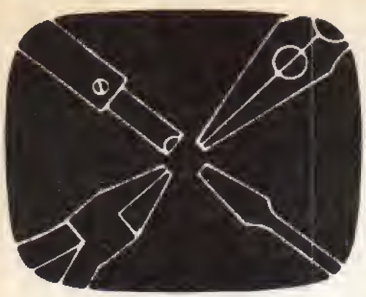
Construction and Testing

There are no critical points in the construction of the 20MHz pulse scaler. As with the frequency meter layout diagrams are available on request. These should be observed and will take care of all the problems involved on account of the very high frequencies involved, calling for short wiring at critical points.

Do not use transistors other than those specified; or if you do be prepared to modify resistor values slightly if necessary. Tr1, Tr4 and Tr5 must be high-beta types (beta at least 200 but preferably not greater than 400 since otherwise the frequency response may suffer). Tr2 and Tr3 must be low-beta types (beta about 30, not exceeding 70). Any silicon or germanium r.f. signal diode is suitable for D5. D1-D4 must be silicon diodes but the actual type is unimportant. D8, D9 may be any silicon or germanium r.f. diodes. IC1 to IC6 are standard packages in the Motorola DTL range and are readily available. Only packages listed as direct substitutes for the specified Motorola types may be used as alternatives without redesign of the circuit. Modifications permissible to give other pulse and space times and to provide continuous interpolation control if desired have previously been discussed.

The instrument requires no presetting and is ready for operation as soon as construction has been completed. If correct performance is not obtained check the voltage readings shown on the circuit diagram. The voltages given are with respect to chassis and should be measured with a high-impedance voltmeter. They are valid with no input signal, VR1 at zero and S1 not set to self-excited. The voltage at S1 slider is either +160mV or +4V depending on whether the output of the scaler chain is in the logical 0 or 1 stage. This can be changed by applying a signal to P1 and adjusting the trigger level control accordingly.

The trigger amplifier can be checked by advancing VR1 progressively to produce the tabulated voltages at A and checking the corresponding voltages at B. Control of the voltage B by the voltage A must be without any backlash or threshold trip behaviour. If this condition is not satisfied one of the resistors R8, R9 or R10 is too far out of tolerance so that these resistors must be checked with a good ohmmeter and the offending one replaced. It is advisable to check these resistors before soldering them on to the printed circuit board. ■



SERVICING television receivers

L. LAWRY-JOHNS

PLESSEY DUAL-STANDARD CHASSIS

THE original Plessey dual-standard chassis was produced in the 1963-64 period and was used in a very large number of receivers. Many of these sets are now on the second-hand market under various brand names and suitably serviced can provide several more years of good viewing. We list the following models but do not claim the list to be complete by any means: **Defiant** Models 9A50, 9A51, 9A52, 9A56, 3A54, 3A60; **Cossor** Models CT1964/77 and 78; **Challenge** Model C501; **Peto Scott** Models TV960 and TV960/90; **RGD** Models 624 and 625; **Regentone** Models 195 and 196.

The differences which occur will be found in the type of tuners fitted (push-button or rotary v.h.f.) and in the vision i.f. strip which may use an EF184 or an EF80 (early versions) in position V2. The position of the choke L2 may be at the top centre of the chassis or on the right side above the line output transformer. The mains dropper may be of the multi-tag type (which fails all too often) or the three-tag type (which doesn't).

There is a great deal of difference between being asked to carry out a quick repair on one of these sets and to renovate one for continuing reliable operation. In the first instance one traces the offending fault and rectifies it. In the second one would check up on all likely trouble spots and replace some components before they have a chance to fail. This is of course more expensive and time consuming but it does pay handsome dividends in the long run.

Faulty Dropper

Most of the older models used an unreliable dropper on the right-hand side any section of which could fail at any time. If one can obtain a more reliable dropper of the same total value it is a matter of minutes to unclip the old one and fit the new with no soldering necessary. A dropper which is adorned with bits and pieces of shunt resistors is not only unsightly but electrically unreliable (even this however is preferable to shorting out the faulty section and thereby ruining the set by over-running it).

Blown Fuses

There are two fuses in insulated carriers on the left side. The one nearer the rear is the mains fuse 1·5A. The second one is the h.t. fuse, 500mA. The latter is the one which will mostly be found blown. It will fail when there is a short anywhere in the h.t. line except in the rectifier and reservoir capacitor. We exclude the smoothing choke and brilliance circuit from this as these rarely give trouble. Assuming that the symptoms are that the valves are lighting up but nothing else is happening, this fuse should be checked.

If it has failed it is reasonable to assume that somewhat more than 500mA has tried to pass through it. If there is no sign of h.t. at either end of the fuse holder and the fuse has not failed, one of the sections of the mains dropper will almost certainly be found to be open-circuit. Proceeding on the assumption that the fuse has failed however a visual check should be made for signs of recent overheating. It is a rule with this chassis that the upper panel should be swung out to check the condition of the resistive element 124 in the video stage V3. This consists of two resistors, one 10kΩ the other 8·2kΩ. Both tend to change value thus passing excessive current. This damages 122 and 123 whose mode of protest is also to change value downward until a virtual short is presented across the h.t. line.

Now this state of affairs should never arise if the viewer has been at all critical in his (or her) viewing since the quality of the picture would have deteriorated to an extent which should have called attention to the developing fault long before the virtual short condition occurs. But it is in the nature of things that people do tolerate the most atrocious viewing and sound conditions and tend to wait for complete failure before having something done about it.

We will have more to say about the symptoms caused by changing value in these resistors later but we are still concerned for now with h.t. shorts. It is worthy of mention that the h.t. tracks on the panel in the vicinity of the video stage and the contrast control run down very near the edge. This makes them not only liable to fracture but also to short to the metal clips if the insulation is accidentally removed.

If there is no sign of trouble in this stage however observe the state of the choke L2 if this is mounted on the right side above the line output transformer housing. This choke tends to sag wearily after a period of use and the exposed end tag can touch the top of the housing. Usually by the time this state of disarray has been reached however the choke has lost enough inductance to stop the line output stage functioning. This rather surprising fact is worth remembering when one is looking for a line output stage fault and is one of the reasons why the choke was more rigidly mounted at the top centre in later models.

Assuming however that the choke is maintaining its rightful place and that the video resistors are healthy looking attention should next be directed to the PCL82 sound output stage. This valve can short inside, cook up its cathode resistor (and capacitor

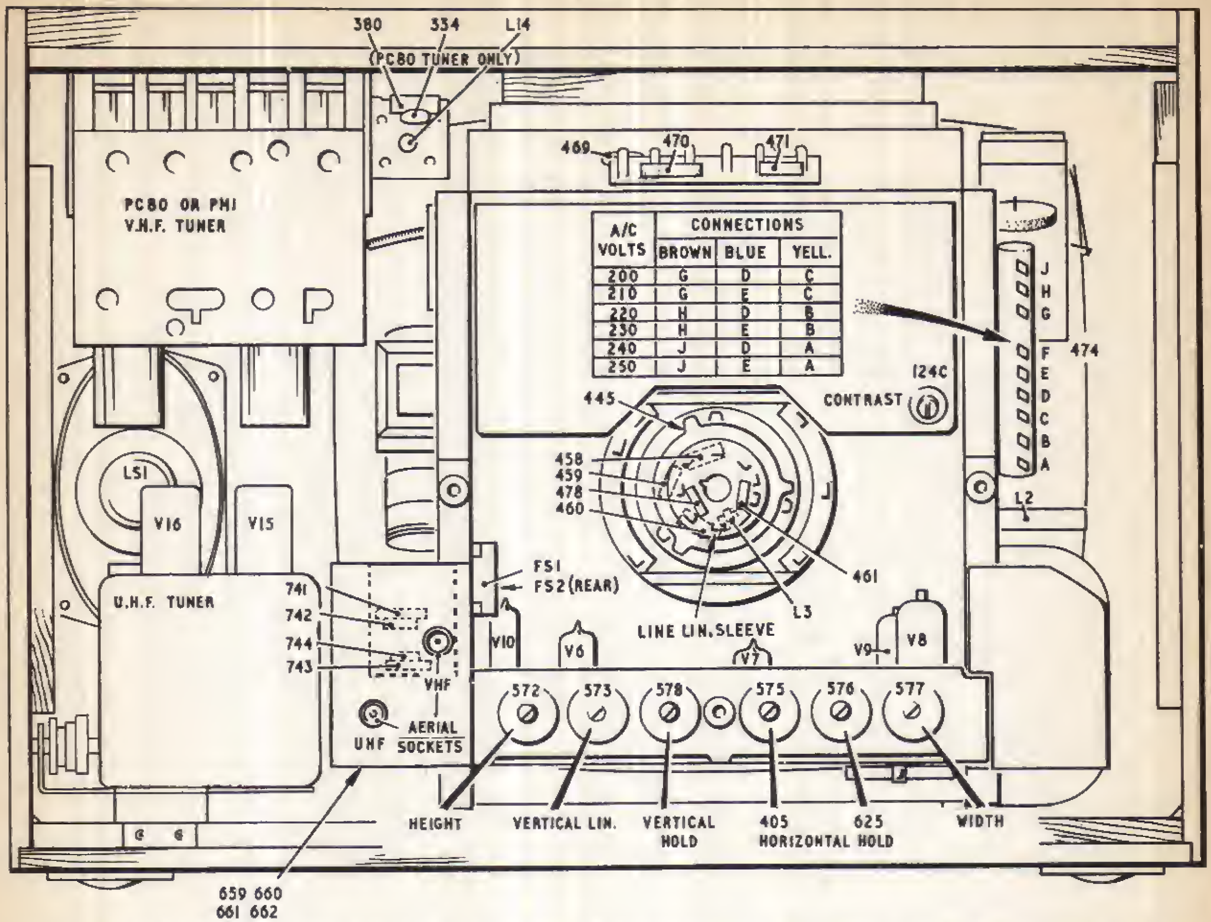


Fig. 1: Rear chassis view. In later versions L2 is mounted at the top centre of the chassis. Loudspeaker position and type of v.h.f. tuner used vary from model to model.

most likely) and cause the fuse to fail. This can also apply to one of the EF184 valves if these are used.

All these items show visual signs of where the fault may be located before the aid of a meter is sought in order for example to find a shorted capacitor. We have found that whilst capacitors do fail in this chassis they do not often do so where they would constitute an h.t. short.

Line Timebase Troubles

The line output transformer itself is one of the trouble spots. Usually when faulty it produces a no e.h.t. condition and it is reasonable to assume that when the valves check out, the boost capacitor (568), the line drive and the 100μF electrolytic (564) are all in order the transformer is indeed at fault. When replacing be sure to obtain the correct replacement to avoid a lot of sorting out of revised connections. One of two types may be found: the original has a black overwind and a separate coil (all connections except one on the outside); the later one is an encapsulated design with no separate coil, a white covered overwind and all connections on the inside. In some cases the chassis member of the early models will not accept the later transformer without some degree of cutting. Stubborn cases of reduced width are often due to a faulty transformer and failure of insulation often occurs particularly on the separate 625 winding which is under the overwind. If the set is used on 405 only it is possible to disconnect this winding completely and at least keep the set going until another transformer is obtained.

In most cases however the transformer is not at fault and the usual checks soon reveal the trouble. Removing the top cap of the PY800 will often bring the line output stage to life thus denoting an alternative h.t. path to the transformer, most often a shorted boost reservoir capacitor (568, 0·25μF). Also check the 100μF electrolytic 564 if removing the PY800 top cap makes no difference, and the condition of the previously mentioned choke L2.

Whilst the PL36 and the PY800 should always be suspect the line oscillator ECH84 rarely gives trouble and line hold troubles usually respond to a check on the diodes and small components on the detachable centre line sync discriminator panel (plug-in).

Sound Faults

Usually the only trouble to be expected apart from hum on 625 concerns the PCL82 output valve. Check this valve and particularly the condition of the cathode bias resistor 142 (390Ω). Quite often the valve is changed but the bias resistor is forgotten leading to an early repetition of the valve failure due

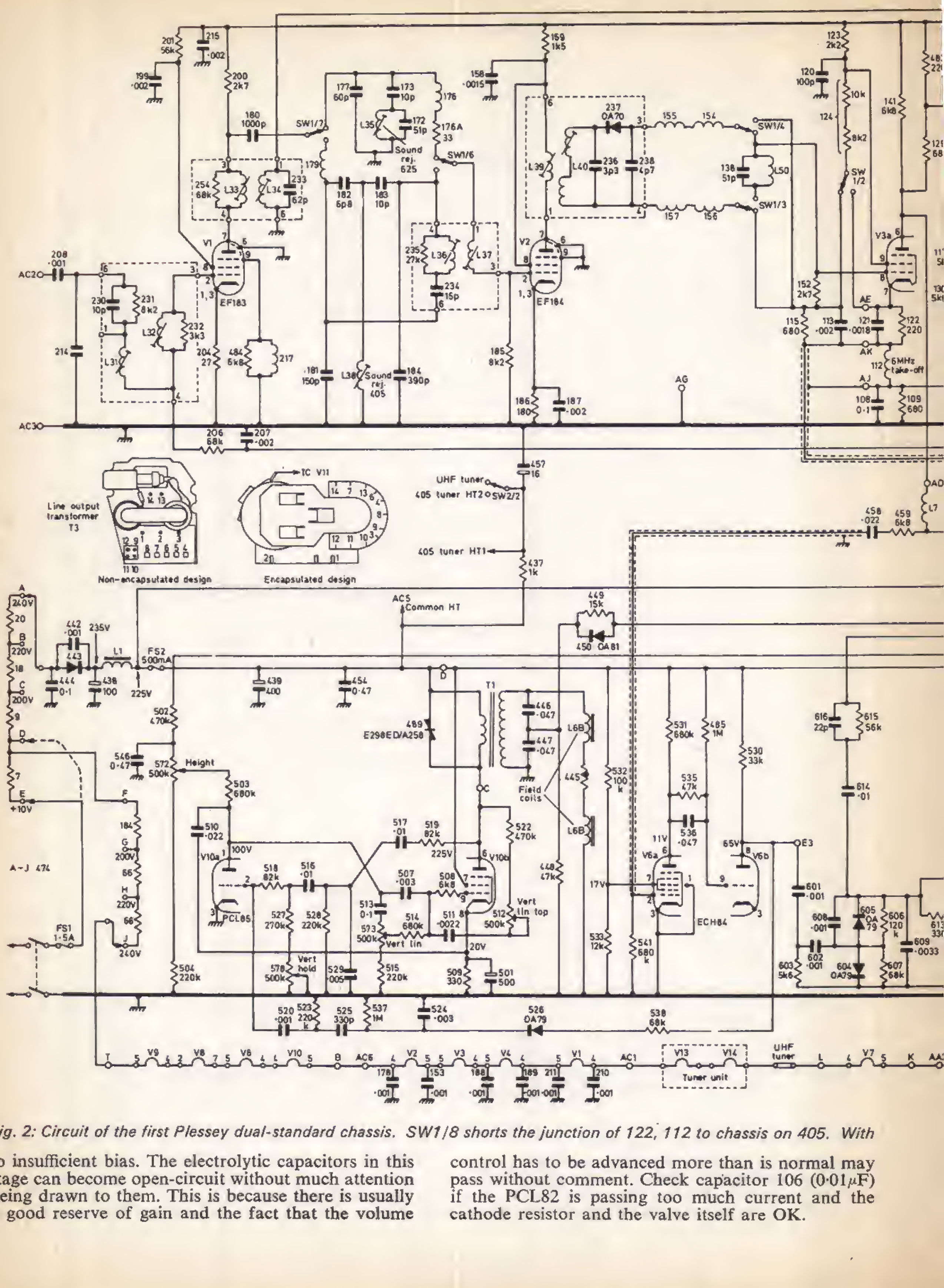
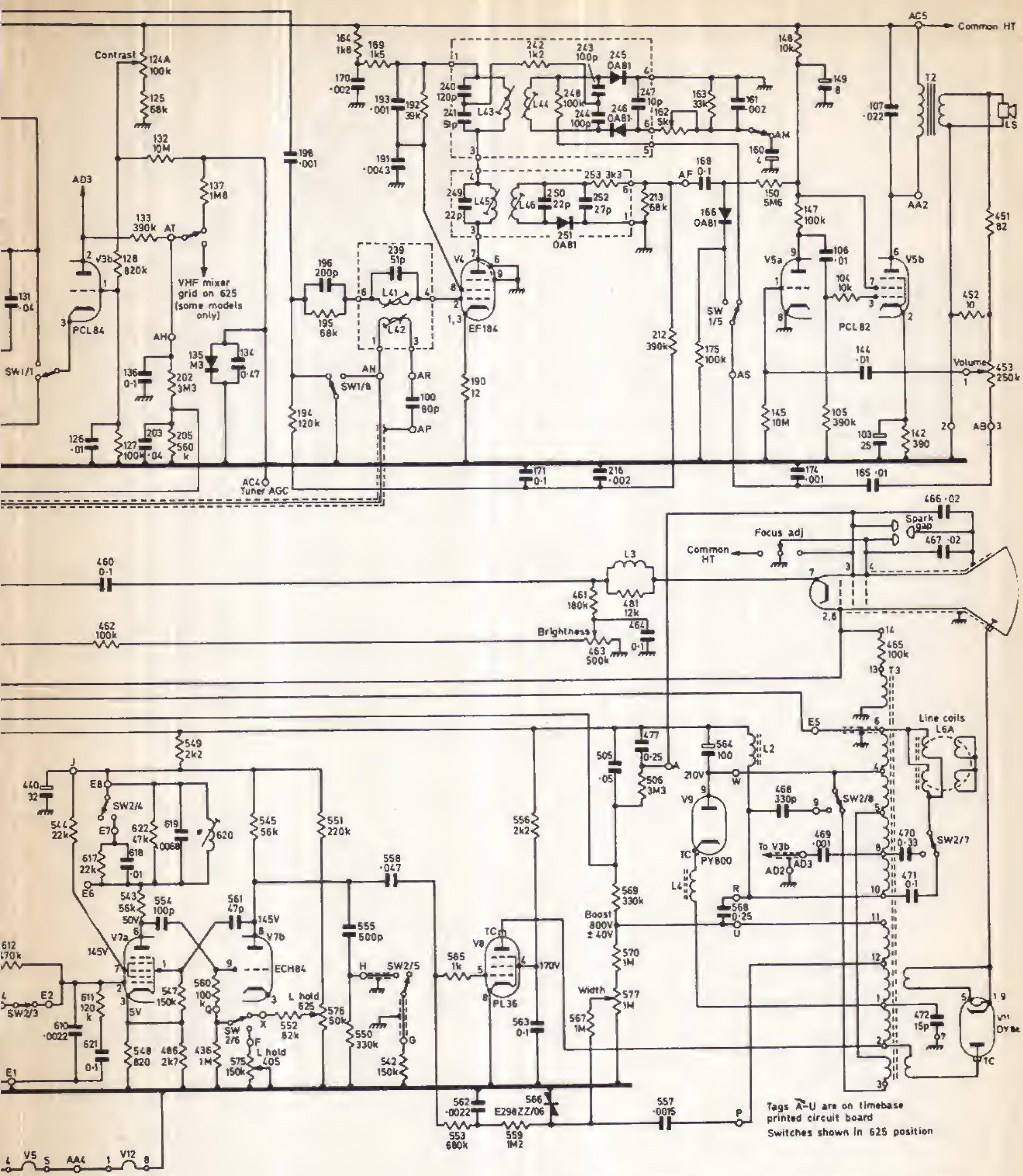


Fig. 2: Circuit of the first Plessey dual-standard chassis. SW1/8 shorts the junction of 122, 112 to chassis on 405. With

to insufficient bias. The electrolytic capacitors in this stage can become open-circuit without much attention being drawn to them. This is because there is usually a good reserve of gain and the fact that the volume control has to be advanced more than is normal may pass without comment. Check capacitor 106 ($0{\cdot}01\mu F$) if the PCL82 is passing too much current and the cathode resistor and the valve itself are OK.



Capacitor 214 is 27pF or 18pF depending on type of tuner fitted.

the encapsulated line output transformer 468 is 260pF and capacitor 472 is omitted.

It is not unusual for the 625 sound to deteriorate making accurate tuning on u.h.f. difficult. This is often due to slight drift in L41-42 and L43-44. Slight adjustment with a suitable trimming tool (hexagon wand) should restore normal tuning. These cores are on the upper left of the panel. Do not disturb the lower cores at all as the picture quality is easily lost.

CONTINUED NEXT MONTH

BASIC CIRCUITS FOR THE CONSTRUCTOR 5

THIS MONTH: AUDIO CIRCUITS

J. W. THOMPSON

HIGH-QUALITY sound in a television receiver is unusual and some readers may consider it to be unnecessary. The sound signal transmitted by the broadcasting authorities however is usually up to a very high standard and one cannot help but wonder why such high standards are maintained when the majority of receivers are equipped with mediocre audio amplifiers and small loudspeakers. The constructor is in a position to be able to take advantage of the high-quality transmissions if he wishes.

Low-power Circuit

Two different circuits are described in this month's article, the first of which is the conventional three-transistor approach shown in Fig. 1. The 625 or 405 sound signal is selected by RLY501 and passes via the volume control to the base of Tr501. The amplified signal at the collector of Tr501 is impedance matched into the loudspeaker by the complementary output pair Tr502/Tr503.

D.C. bias for the circuit is obtained by a resistive feedback network (R502, VR502) from the junction of R503 and R504, VR502 being set so that the voltage at this junction is half the negative rail voltage. Bias stability is maintained by virtue of the high open-loop gain of Tr501—if the midpoint bias at the junction of R503, R504 tends to drift upwards Tr501 base voltage will increase causing a reduction in its collector voltage so that the initial drift is compensated. The quiescent (zero signal) collector current of Tr502 and Tr503 is determined by the voltage drop across D501 which is a bias diode specifically designed for use with AC128/AC176 complementary output transistors.

From the signal viewpoint Tr501 drives the complementary output pair, Tr502 conducting on the negative-going transitions of the drive waveform and Tr503 on the positive-going transitions. The closed-loop a.c. voltage gain of the amplifier is approximately 15; thus 1V p-p is required at the input to drive the amplifier to full output. Into a 15Ω loudspeaker the output power is about 1W which is quite sufficient for normal domestic use. Into a 3Ω loudspeaker the power output will be somewhat greater but will not approach the theoretical maximum level of 5W due to impedance mismatching. The maximum output power (Pmax) available from a correctly matched audio amplifier is given by the following expression:

$$P\text{max} \simeq \left[\frac{V-3}{2 \cdot 8}\right]^2 \div Z_L,$$

where V is the supply voltage and Z_L is the loudspeaker impedance.

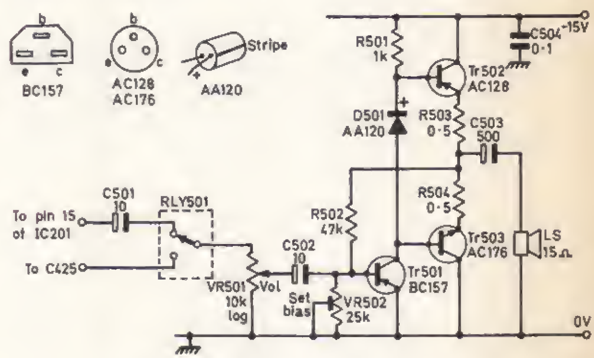


Fig. 1: Low-power audio amplifier.

Table 1: Brief performance specifications.

Amplifier	*Maximum output power (UK r.m.s. watts)*	*Input voltage for maximum output (p-p volts)*	*Total harmonic distortion at 80% of maximum output (1kHz sinewave)*	*Frequency response*
3-transistor with 15Ω LS and 15V supply	1W	1V	2%	20 Hz-15kHz±2dB
10-transistor with 3Ω LS and 15V supply	5W	1V	<0·1%	15 Hz-50kHz±1·5dB (see also Fig. 3)
10-transistor with 3Ω LS and 30V supply	20W	2V	<0·1%	As above

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EBF89	40p
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EC88	60p
EC90	30p
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EC93	47½p
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ECC88	55p
E88CC	62½p
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6AK6	57½p
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6AL5	16p
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6AM6	22½p
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SN4061	23p
SN4062	23p
SN4286	18p
SN4291	18p
RCA	
40253	P/A
40398	P/A
40458	P/A
AC107	30p
AC117	60p
AC126	20p
AC127	25p
AC128	20p
AC176	25p
AC187	63p
AC188	38p
ACY17	28p
ACY20	25p
AD140	40p
AD142	58p
AD149	58p
AD161	38p
AD162	38p
AF102	68p
AF114	25p
AF115	25p
AF116	25p
AF117	25p
AF118	60p
AF124	13p
AF125	20p
AF126	20p
AF127	18p
AF139	38p
AF178	45p
AF179	45p
AF180	53p
AF181	43p
AF186	67p
AF239	43p
ASY28	28p
BA144	P/A
BA145	P/A
BA148	23p
BA155	P/A
BA156	P/A
BC107	15p
BC108	15p
BC109	15p
BC113	28p
BC114	38p
BC115	33p
BC116	63p
BC116A	38p
BC118	33p
BC134	68p
BC135	P/A
BC136	P/A
BC137	P/A
BC138	P/A
BC142	30p
BC143	P/A
BC147	18p
BC148	15p
BC149	18p
BC152	18p
BC158	18p
BC169B	14p
BC169C	15p
BC171	18p
BC175	28p
BC183	23p
BC184	23p
BC187	29p
BC213L	27p
BCY32	38p
BCY58	23p
BCY70	20p
BD121	65p
BD123	83p
BD124	63p
BD131	98p
BF115	25p
BF117	48p
BF163	35p
BF167	25p
BF173	33p
BF178	35p
BF179	73p
BF180	35p
BF181	33p
BF184	25p
BF191	23p
BF195	28p
BF196	43p
BF197	32p
BF200	37p
BF224	30p
BF225	30p
BF257	47p
BFX29	35p
BF161	P/A
BF162	P/A
BF163	25p
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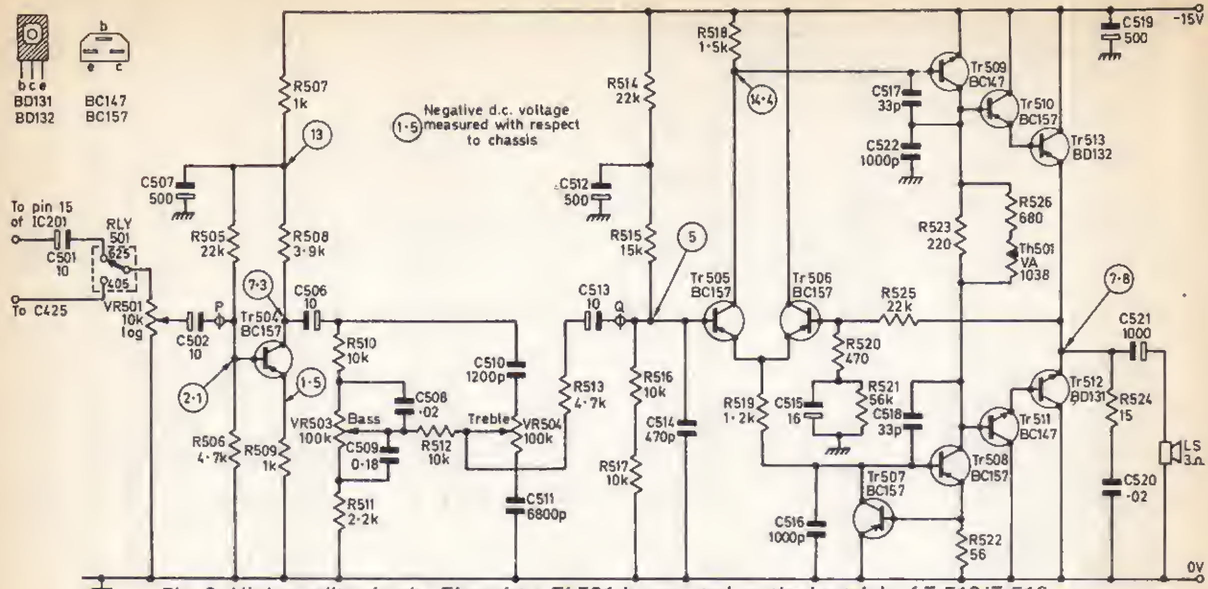


Fig. 2: High-quality circuit. Thermistor Th501 is mounted on the heatsink of Tr512/Tr513.

The type of amplifier described above is thus similar in terms of output power and sound quality to the majority of audio amplifiers in domestic television receivers. If a large loudspeaker is used reasonably good results will be obtained, but probably not good enough to satisfy the Hi-Fi enthusiast. The main problem is that the amplifier has insufficient overdrive capability —when the volume is set to give a reasonable listening level the high-amplitude transients in the audio signal are likely to be clipped, resulting in distortion.

High-quality Circuit

The amplifier circuit shown in Fig. 2 provides the ultimate answer. Despite its complexity this amplifier is remarkably well-behaved. It sets up its own mid-point bias and quiescent current automatically, delivers 5W r.m.s. into a 3Ω loudspeaker (15V supply) or 20W with a 30V supply and has very low distortion figures, typically less than 0.1%. Separate bass and treble controls are provided, each allowing more than

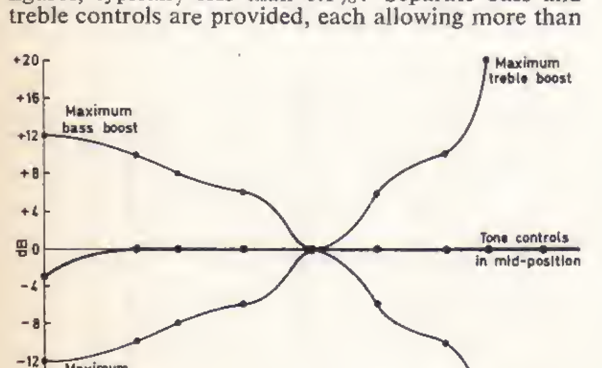


Fig. 3: Frequency response of the high-quality ten-transistor circuit with the tone controls fitted. (Measured at 80% of maximum output.)

12dB lift and cut. A graph of the frequency response of the amplifier is shown in Fig. 3 while Table 1 compares its characteristics with those of the first amplifier described.

The operation of the circuit is as follows. The audio signal is selected by RLY501 and passes through the volume control to the base of Tr504. R509 is not capacitively decoupled to chassis so the voltage gain of Tr504 is given approximately by the value of R508 divided by the value of R509 (i.e. ×4) while the amplification is very linear due to the high inherent level of negative feedback. The amplified voltage at the collector of Tr504 then passes via the tone control network to Tr505.

Consider next the d.c. operating conditions of the output amplifier (Tr505 to Tr513): a proportion of the mid-point voltage at the junction of the emitters of Tr512 and Tr513 is applied to the base of Tr506 while a fixed bias derived from the potential divider R514, R515, R516 and R517 is applied to the base of Tr505. Tr505 and Tr506 act as a *comparator* circuit, any difference in the base voltages of these two transistors appearing as an error voltage across R518. This error voltage is further amplified by Tr509, the collector voltage of which directly determines the mid-point voltage of the output transistors Tr512 and Tr513. By suitable selection of bias resistors the mid-point may be arranged to be exactly half the supply voltage regardless of the actual level of the supply voltage. Any drift is automatically compensated. The

Table 2: Complementary transistors for use with *positive* supply voltage.

Replace	*By*
BC157	BC147
AC128	AC176
AC176	AC128
BC147	BC157
BD132	BD131
BD131	BD132

★ components list

Resistors:

R501	1kΩ	R505	22kΩ	R509	1kΩ	R513	4·7kΩ	R517	10kΩ	R521	56kΩ	R525	22kΩ
R502	47kΩ	R506	4·7kΩ	R510	10kΩ	R514	22kΩ	R518	1·5kΩ	R522	56Ω	R526	680Ω
R503	0·5Ω1W	R507	1kΩ	R511	2·2kΩ	R515	15kΩ	R519	1·2kΩ	R523	220Ω	$\frac{1}{4}$/$\frac{1}{2}$W 5% unless	
R504	0·5Ω1W	R508	3·9kΩ	R512	10kΩ	R516	10kΩ	R520	470Ω	R524	15Ω	otherwise stated	

Potentiometers:

VR501 10kΩ log. VR502 25kΩ lin. VR503 100kΩ lin. VR504 100kΩ lin.

Capacitors:

C501	10 μF 15V E	C508	0·02 μF PE	C514	470pF P	C520	0·02 μF PE
C502	10 μF 15V E	C509	0·18 μF PE	C515	16 μF 25V E	C521	1000 μF 35V E
C503	500 μF 15V E	C510	1200pF PE	C516	1000pF PE	C522	1000pF PE
C504	0·1 μF PE	C511	6800pF PE	C517	33pF P	E electrolytic; P polystyrene;	
C506	10 μF 15V E	C512	500 μF 35V E	C518	33pF P	PE 160V polyester.	
C507	500 μF 35V E	C513	10 μF 15V E	C519	500 μF 35V E		

Semiconductors:

D501	AA120 (Thorn)	Tr503	AC176	Tr506	BC157	Tr509	BC147	Tr512	BD131
Tr501	BC157	Tr504	BC157	Tr507	BC157	Tr510	BC157	Tr513	BD132
Tr502	AC128	Tr505	BC157	Tr508	BC157	Tr511	BC147	Th501	VA1038

Miscellaneous:

RLY501 Omron MH2, d.p.c.o. relay (Home Radio). Heatsink(s), silicone grease, etc.

quiescent current in the output transistors is determined by the voltage drop across R523 and this is itself determined by the collector current of Tr508. Tr507 and Tr508 form a constant-current source, thus providing a constant voltage across R523 and stabilisation of quiescent current over a wide range of supply voltages. A thermistor (Th501) stabilises the quiescent current against changes in temperature.

AC Operation

The a.c. operation of the amplifier is rather complicated. Signals of opposite phase are produced at the collector and emitter of Tr505. Tr509 provides voltage amplification of Tr505's collector signal, but Tr508 does not amplify the emitter signal: any change in voltage at the base of Tr508 is immediately suppressed by a change in the conduction of Tr507. The base of Tr508 is thus a virtual earth. We may consider the collector impedance of Tr508 as a variable resistor of about 1,000Ω connected between the base of Tr511 and earth, its resistance varying to maintain a constant current in R523. The amplified a.c. signal at the collector of Tr509 thus appears at both sides of R523: for the negative-going part of the waveform Tr510 and Tr513 conduct, while for the positive-going part Tr511 and Tr512 conduct. Comparing this circuit with that shown in Fig. 1, Tr509 takes the place of Tr501 with Tr510/Tr513 and Tr511/Tr512 replacing Tr502 and Tr503: R523 replaces D501, with the bias stabilising action of the diode undertaken by Tr507 and Tr508. The final amplified signal appears at the emitters of Tr512 and Tr513 and is coupled by C521 to the loudspeaker.

A.C. negative feedback is applied to the base of Tr506, the magnitude of the feedback being determined by the a.c. potential divider R525, R520. The exceptionally low distortion figures obtainable from the amplifier are due partly to this a.c. feedback and partly to the use of Tr507 and Tr508 as a constant-current source. A 1V p-p input to RLY501 will drive the amplifier to full output with a 15V supply while a 2V p-p input is required for full output with a 30V supply. The amplifier is thus fully compatible with the sound i.f. circuits described earlier in this series.

Layout and Construction

The layout for either amplifier is not critical provided the input and output leads are kept well apart. It is suggested that the amplifiers are built on paxolin pinboard—it is possible to lay the amplifier out exactly as it appears in the theoretical circuit diagram. In the prototype receiver the entire amplifier is mounted on the receiver's control panel, thus permitting the use of very short leads to the volume and tone controls. Signal-carrying leads of any appreciable length should be made up with screened cable. The amplifier earth is connected to the main chassis through heavy-duty copper braid. RLY501 may most conveniently be mounted on the main chassis adjacent to the sound i.f. amplifiers: in single-standard receivers it is of course omitted. The tone controls in Fig. 2 are optional: if they are not required it is possible to connect together points P and Q on the circuit diagram leaving out the intermediate circuitry (Tr504 etc.).

With both amplifiers heatsinks will be required for the output transistors. For the amplifier in Fig. 1 copper clips each about 2 sq. cm. in area will be sufficient for Tr502 and Tr503 so long as the loudspeaker impedance is not less than 15Ω. If a 3Ω loudspeaker is used the copper clips should be bolted to a heavy-duty finned heatsink. No electrical insulation is required between the cans of these transistors. All thermal conduction interfaces (i.e. the transistor cans and the side of the copper clips bolted to the heatsink) should be given a liberal coating of silicone grease: this is most important if thermal runaway is to be prevented. Ideally D501 should be mounted in thermal contact with the heatsink used for the output transistors.

For the amplifier shown in Fig. 2 a 5cm. × 10cm.

black finned heatsink will suffice for output powers up to 20W. If both output transistors (Tr512 and Tr513) are bolted to the same heatsink a mica washer should be placed between the metal side of each transistor and the heatsink. Apply plenty of silicone grease to both sides of the mica washers. Alternatively the transistors may be bolted without mica washers to two separate heatsinks electrically isolated from each other. The thermistor Th501 should be clipped on to the heatsink (or one of the heatsinks if two are used). It is advisable to insulate the thermistor from the heatsink by sandwiching it between two slices of mica washer liberally coated with silicone grease.

Setting Up

To set up the three-transistor amplifier turn the volume control to minimum, connect a voltmeter between the junction of R503 and R504 and chassis and apply -15V to the supply rail of the amplifier. Set VR502 to give a reading of 7.5V on the meter. If an oscilloscope is available it is better to set VR502 as follows. Inject a 1kHz sinewave signal at the base of Tr501 and monitor the output across the loudspeaker with an oscilloscope. Increase the level of the input signal until clipping occurs, and adjust VR502 so that the clipping is symmetrical at the top and bottom of the sinewave. The quiescent (zero signal) current of the amplifier should be between 10mA and 20mA. This is finely set by the voltage drop across D501: it is most important that this type of diode is used otherwise the quiescent current will be far too large.

The ten-transistor amplifier circuit (Fig. 2) requires no setting-up adjustment but if any problems arise it should be possible to determine the cause of the fault by measuring the voltages shown at various points in the circuit. In the event of failure of either or both output transistors (due perhaps to a short-circuited loudspeaker) it is advisable to replace Tr510, Tr511, Tr512 and Tr513. The level of quiescent current can if necessary be finely adjusted by a small change in the value of R522.

General Points

The loudspeaker used with either amplifier should be as large as the receiver cabinet permits. It is possible to use more than one loudspeaker, with series or parallel connection, but the total impedance presented to the output of either amplifier *should not be less than* 3Ω. External loudspeakers may easily and safely be connected as the receiver has an earthed chassis. The use of an electrostatic loudspeaker is not recommended with the ten-transistor amplifier as it tends towards instability when driving this type of capacitive load.

Both amplifiers may be converted to operate from a positive supply voltage if required. D501 and all electrolytic capacitors must be reversed in polarity and each transistor replaced by its complement (see Table 2). The circuits are shown for use with a negative rail because this was the most suitable supply voltage available in the prototype receiver.

The amplifier circuit shown in Fig. 2 is ideally suited for use as a general-purpose Hi-Fi amplifier. When used with a record or tape deck a correctly equalised preamplifier is necessary giving at least 2V p-p output into 10kΩ. Further details of such a preamplifier can be found in the *Mullard Book of Transistor Circuits*.

NEXT MONTH A DUAL-STANDARD LINE TIMEBASE

COLOUR RECEIVER CIRCUITS

GORDON J. KING

CHROMINANCE CIRCUITS

THE chroma signal is present with the Y signal at the output of the vision detector and in receivers using a common vision detector for both signals the chroma signal is filtered from the Y signal by a high-pass coupling to the chroma bandpass amplifier channel. It will be recalled that the chroma signal is removed in the Y channel by a chroma trap.

Chroma Bandpass Amplifiers

The colour decoder starts with the chroma bandpass amplifier channel which is an important section of the colour receiver: if there is trouble in this section the receiver will probably operate in monochrome but certainly not in colour. It is called a bandpass amplifier because its response characteristics are controlled in such a way as to pass the sidebands of the chroma signal with the least distortion and without letting through unwanted adjacent signals.

Figure 1 shows the positions of the video and sound signals in a 625-line channel in terms of frequencies relative to the vision carrier. The range of frequencies in which we are currently interested is indicated by the shading. This is the part of the video spectrum where the chroma signals are carried, spreading out approximately 1MHz either side of the chroma subcarrier frequency (4·43MHz). It is the job of the chroma bandpass amplifier channel to accept this section of the video waveform and to lift it to a level suitable for application to the PAL delay line and its associated matrix network. Fig. 2 shows the various inputs and outputs of the chroma bandpass amplifier channel.

Burst Blanking

The chroma signal fed to the PAL delay line should not only be devoid of video signal from about 3·4MHz down to d.c. but should also be clear of the burst information. The bursts are therefore extracted from the chroma channel at an early stage and processed separately. Burst blanking as it is called is subsequently carried out by switching off the chroma channel during the burst period. This is done by applying a burst blanking pulse to the channel and ensures that unwanted signals are not fed to the synchronous detectors during the line retrace period. Since the average phase of the transmitted burst is coincident with the 180-degree −U chroma axis (see last month), corresponding to a yellowish-green hue, a vertical line display of this colour at the left of the picture would be likely to result should the burst get through to the synchronous detectors.

Automatic Chrominance Control

Most colour receivers incorporate automatic chroma control (a.c.c.) which is a form of a.g.c. in the chroma channel. The purpose of this is to hold steady the relative amplitudes of the chroma and luminance signals under varying conditions of propagation and during the normal operational drift of the receiver circuits. The chroma signal itself cannot be utilised to derive this control bias, which is generally applied to an early stage in the chroma bandpass amplifier channel, since its value is continually changing with the varying colouring information. Since the burst signal is not modulated however the a.c.c. bias can be derived from this signal either directly or indirectly from the ripple signal resulting from the burst swings.

Manual Chrominance Control

Provision for manual gain control in the chroma amplifier channel is also necessary to allow the amplitude of the chroma signal fed to the synchronous detectors—and hence the saturation of the display—to be adjusted. In some sets electrical or mechanical ganging of the contrast and colour controls is adopted so that the chroma/luminance ratio holds fairly constant as the colour control is adjusted.

Colour Killer

One stage of the chroma amplifier channel is deliberately biased to cut-off in the absence of a colour signal. A switch-on bias for this stage is however obtained from the bursts—or from the ident signal derived from them—when the transmission carries colour information. This technique ensures that the chroma channel is inactive during mono-

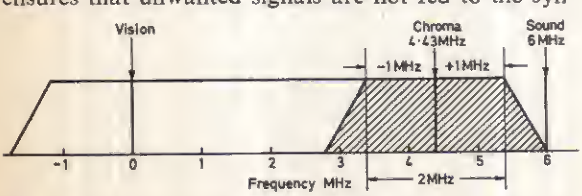


Fig. 1: U.H.F. television channel spectrum, showing the portion occupied by the chroma signal.

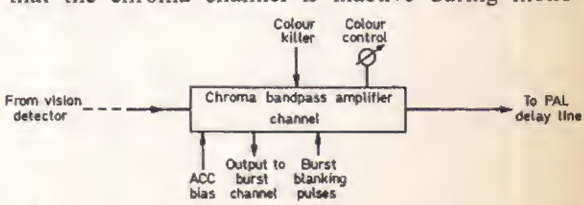
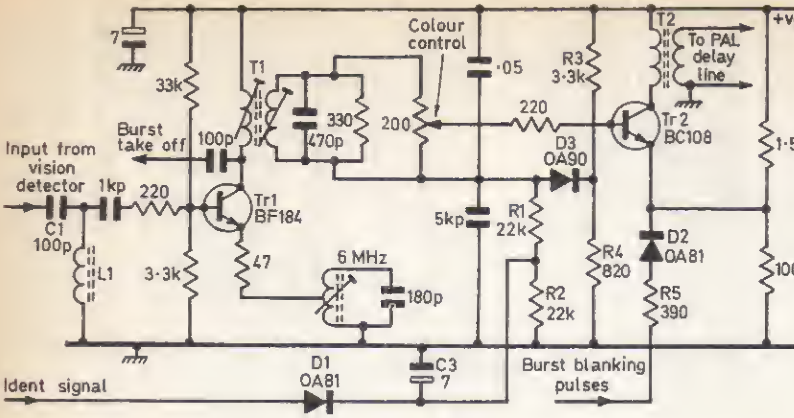


Fig. 2: Block diagram of the chroma channel.



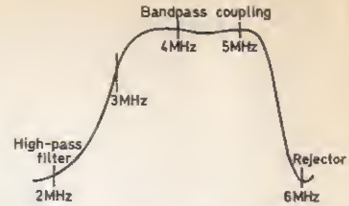


Fig. 4 (above): The response of the circuit shown in Fig. 3.

Fig. 3 (left): Chroma channel used in early Decca dual-standard colour receivers.

chrome transmissions: without it random video signals and noise focused ±1MHz relative to the subcarrier frequency would tend to introduce disconcerting flecks of colour on a black-and-white picture. It is from this "colour killing" action that the circuit takes its name: it is however really a "colour activator" system since the controlled stage is basically cut-off, being biased on by the rectified and smoothed burst signal.

Simple Chrominance Channel

A fairly simple chroma amplifier channel—used in Decca dual-standard models—is shown in Fig. 3. This particular channel is not equipped with a.c.c. so the only controls involved are the colour killer, manual colour and burst blanking. The input signal is obtained from the vision detector via a video phase splitter (common dual-standard practice) and thus contains luminance as well as chroma information. The luminance signal is eliminated by the input high-pass filter C1/L1 so that Tr1 base receives only the chroma signal. This appears amplified across the primary of the bandpass coupling transformer T1 and is conveyed via the secondary winding to Tr2 base by way of the colour control which acts as a simple signal potentiometer. Transformer T2 couples Tr2 to the PAL delay line circuit.

B ndpass Response

An important feature is the 6MHz rejector in Tr1 emitter circuit. This is required because the upper passband of the channel cannot be defined sufficiently by the coupling transformers so that without the rejector the chroma channel would pass on to the succeeding circuits any components of the 6MHz intercarrier signal present. The required bandwidth is thus obtained by the high-pass filter which rolls-off the spectrum from about 3·4MHz down to the vision carrier and the 6MHz rejector which puts a sharp notch in the response at the other side of the passband. The rejector and the primary and secondary of T1 are adjusted to secure the response shown in Fig. 4. A sweep generator with marker frequencies and an oscilloscope are required for this setting up. The take off to the burst channel is at Tr1 collector.

Colour Killer Action

The colour killing and burst blanking actions are both carried out in stage Tr2. The colour killer works in the base circuit. The base bias potential-divider resistors are R3 and R4 which are linked to Tr2 base by D3 and the colour control. On monochrome reception the positive potential at D3 cathode switches D3 off, removing the base bias from Tr2. When a colour signal is being received the ident signal produced from the bursts is rectified by D1 and a positive potential develops across C3 to forward bias D3 and thus link Tr2 to its base bias network. Tr2 is thus brought into conduction and passes the chroma signal to the delay line.

Burst Blanking System

The burst blanking works in the emitter circuit of Tr2 and is similar to the line and field flyback blanking carried out in the luminance channel of sets using colour-difference tube drive. Positive-going pulses from the line output transformer are applied to Tr2 emitter via D2 and R5, switching Tr2 off during the line sync pulse and burst period. The diode deletes negative overshoots.

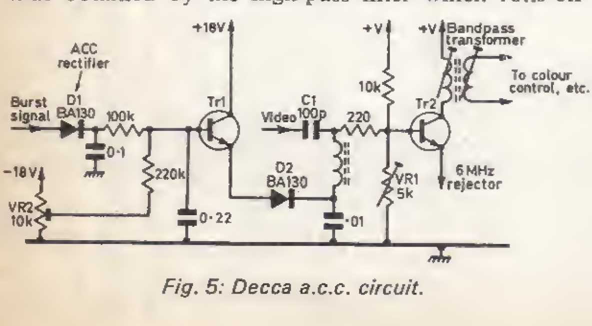


Fig. 5: Decca a.c.c. circuit.

Decca ACC Circuit

Burst blanking and colour killing are carried out in various ways by the different setmakers: before looking at some alternative techniques however we will take a look at the simple a.c.c. system used in Mark II dual-standard Decca receivers. This employs an extra transistor stage (Tr1) in the chroma channel as shown in Fig. 5. As before the signal from the vision detector is fed to the base of the first chroma amplifier (Tr2) with the high-pass filter still active but now the bottom arm of the base potential-divider consists of a variable resistor VR1. In the absence of a colour signal this is adjusted so that Tr2 runs at maximum gain. When a colour signal is present the bursts are rectified by the

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AC115	23p	AF117	17p	BC142	45p	BCY33	17p	BF274	30p	MAT100	15p	ST140	12p	2N930	25p	2N2904A	30p	2N3706	12p
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AC126	17p	AF124	21p	BC145	45p	BCY70	17p	BF309	37p	MAT120	15p	TIS43	40p	2N1132	22p	2N2905A	30p	2N3708	8p
AC127	17p	AF125	20p	BC147	17p	BCY71	30p	BF316	75p	MAT121	17p	UT46	27p	2N1302	17p	2N2906	25p	2N3709	8p
AC128	17p	AF126	20p	BC148	12p	BCY72	15p	BFW10	55p	MPF102	43p	V405A	25p	2N1303	17p	2N2906A	27p	2N3710	10p
AC141K	17p	AF127	20p	BC149	17p	BCZ11	20p	BFX29	27p	MPF105	43p	V410A	45p	2N1304	20p	2N2907	25p	2N3711	10p
AC142K	17p	AF139	33p	BC150	17p	BD121	85p	BFX84	20p	OC19	30p	2G301	19p	2N1305	20p	2N2907A	30p	2N3819	40p
AC151	15p	AF178	50p	BC151	20p	BD123	85p	BFX85	27p	OC20	50p	2G302	19p	2N1306	22p	2N2923	13p	2N3820	£1
AC154	15p	AF179	50p	BC152	17p	BD124	75p	BFX86	22p	OC22	30p	2G303	19p	2N1307	22p	2N2924	13p	2N3903	25p
AC155	17p	AF180	50p	BC153	27p	BD131	80p	BFX87	25p	OC23	33p	2G304	20p	2N1308	27p	2N2925	13p	2N3904	27p
AC156	17p	AF191	50p	BC154	30p	BD132	80p	BFX88	22p	OC24	45p	2G306	35p	2N1309	27p	2N2926		2N3905	25p
AC157	17p	AF186	45p	BC157	20p	BDY20	£1	BFY50	20p	OC25	25p	2G308	35p	2N1613	17p	(G)	12p	2N3906	27p
AC165	17p	AF239	37p	BC158	17p	BF115	22p	BFY51	20p	OC26	25p	2G309	35p	2N1711	20p	2N2926		2N4058	15p
AC166	17p	AFZ11	37p	BC159	20p	BF117	45p	BFY52	20p	OC28	40p	2G339	17p	2N1889	35p	(Y)	11p	2N4059	10p
AC167	20p	AFZ12	45p	BC167	13p	BF118	60p	BFY53	17p	OC29	40p	2G339A	15p	2N1890	45p	2N2926		2N4060	12p
AC168	20p	AL102	85p	BC168	13p	BF119	70p	BSX19	15p	OC35	33p	2G344	15p	2N1893	37p	(O)	10p	2N4061	12p
AC169	14p	AL103	85p	BC169	13p	BF152	35p	BSX20	15p	OC36	40p	2G345	15p	2N2160	60p	2N3010	80p	2N4062	12p
AC176	23p	ASY26	25p	BC170	12p	BF153	35p	BSY25	15p	OC41	20p	2G371	13p	2N2147	75p	2N3011	20p	2N5172	12p
AC177	20p	ASY27	30p	BC171	13p	BF154	35p	BSY26	15p	OC42	22p	2G371B	10p	2N2148	60p	2N3053	20p	2N5459	43p
AC187	30p	ASY28	25p	BC172	13p	BF157	45p	BSY27	15p	OC44	15p	2G374	17p	2N2192	30p	2N3054	50p	2S034	75p
AC188	30p	ASY29	25p	BC173	13p	BF158	25p	BSY28	15p	OC45	12p	2G377	27p	2N2193	30p	2N3055	63p	2S301	50p
ACY17	25p	ASY50	25p	BC174	13p	BF159	30p	BSY29	15p	OC70	15p	2G378	15p	2N2194	27p	2N3391	17p	2S302A	45p
ACY18	20p	ASY51	25p	BC175	22p	BF160	30p	BSY38	15p	OC71	9p	2G382	15p	2N2217	20p	2N3391A	20p	2S302	45p
ACY19	22p	ASY52	25p	BC177	17p	BF162	30p	BSY39	15p	OC72	12p	2G401	30p	2N2218	25p	2N3392	17p	2S303	60p
ACY20	20p	ASY54	25p	BC178	17p	BF163	35p	BSY40	30p	OC74	12p	2G414	30p	2N2219	27p	2N3393	15p	2S304	£1·10
ACY21	20p	ASY55	25p	BC179	17p	BF164	35p	BSY41	35p	OC75	15p	2G417	25p	2N2220	22p	2N3394	15p	2S305	£1
ACY22	19p	ASY56	25p	BC180	20p	BF165	35p	BSY95	12p	OC76	15p	2N388	30p	2N2221	22p	2N3395	20p	2S306	£1·10
ACY27	18p	ASY57	25p	BC181	22p	BF167	22p	BSY95A	12p	OC77	25p	2N388A	50p	2N2222	27p	2N3402	22p	2S307	£1·10
ACY28	19p	ASY58	25p	BC182	10p	BF173	22p	BU105	£3·90	OC81	15p	2N404	22p	2N2368	17p	2N3403	22p	2S321	60p
ACY29	30p	ASY58	25p	BC182L	10p	BF176	35p	C111E	60p	OC81D	15p	2N404A	30p	2N2369	15p	2N3404	32p	2S322	50p
ACY30	25p	ASZ21	40p	BC183	10p	BF177	35p	C400	30p	OC82	15p	2N524	55p	2N2369A	15p	2N3405	45p	2S322A	45p
ACY31	25p	BC107	10p	BC183L	10p	BF178	45p	C407	25p	OC82D	15p	2N527	60p	2N2411	50p	2N3414	20p	2S323	60p
ACY34	18p	BC108	10p	BC184	13p	BF179	50p	C424	17p	OC83	20p	2N696	12p	2N2412	50p	2N3415	20p	2S324	£1·20
ACY35	18p	BC109	11p	BC184L	13p	BF180	30p	C425	40p	OC84	20p	2N697	15p	2N2616	55p	2N3417	37p	2S325	£1·20
ACY36	30p	BC113	25p	BC186	27p	BF181	30p	C426	30p	OC139	15p	2N698	24p	2N2711	22p	2N3525	74p	2S326	£1·20
ACY40	15p	BC114	30p	BC187	27p	BF182	30p	C428	20p	OC140	17p	2N699	55p	2N2712	22p	2N3702	12p	2S327	£1·20
ACY41	18p	BC115	30p	BC207	11p	BF183	30p	C441	27p	OC170	15p	2N706	7p	2N2714	25p	2N3703	12p		
ACY44	35p	BC116	35p	BC209	11p	BF184	25p	C442	35p	OC171	15p	2N706A	8p						
AD140	40p	BC117	35p	BC209	11p	BF185	30p	C444	37p	OC200	25p	2N708	12p						
AD142	40p	BC118	25p	BC212L	11p	BF188	30p	C450	17p	OC201	27p	1N709	45p						
AD149	43p	BC119	45p	BC213L	11p	BF194	23p	C720	12p	OC202	27p	2N711	40p						
AD161	35p	BC125	35p	BC213L	11p	BF195	24p	C722	25p	OC203	25p	2Nu17	42p						
AD162	35p	BC126	35p	BC214L	12p	BF196	30p	C740	25p	OC204	25p	2N718	24p						
AD161/		BC132	25p	BC225	25p	BF197	35p	C742	17p	OC205	35p	2N718A	50p						
162(MP)	63p	BC134	30p	BC226	35p	BF200	45p	C744	17p	OC309	35p	2N726	27p						
ADT140	50p	BC135	30p	BC317	12p	BF222	80p	C760	17p	P346A	17p	2N727	27p						
ADZ11	£2	BC136	30p	BC318	12p	BF257	35p	C762	17p	P397	45p	2N743	17p						
ADZ12	£2·10	BC137	35p	BC319	12p	BF270	25p	C764	60p	OCP71	43p	2N744	17p						
AF114	17p	BC139	45p	BCY30	20p	BF271	17p	EC401	15p	ORP12	43p	2N914	17p						

DIODES & RECTIFIERS

AA119	8p	BY130	15p	OA10	22p
AA120	8p	BYZ10	35p	OA47	7p
BA116	22p	BYZ11	32p	OA70	7p
BA126	22p	BYZ12	30p	OA79	8p
BY100	15p	BYZ13	25p	OA81	7p
BY101	12p	BYZ16	35p	OA85	7p
BY105	15p	BYZ17	35p	OA90	6p
BY114	12p	BYZ18	30p	OA91	7p
BY126	15p	BYZ19	25p	OA95	7p
BY127	17p	OA5	17p	OA200	6p

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BI-PAK Order No.	Similar Types to: Description	Price and qty. prices 1-24 £p	25-99 £p	100 up £p
BP00 = 7400	Quad. 2-input NAND gate	0·15	0·14	0·12
BP01 = 7401	Quad. 2-input positive NAND gate (with open collector output)	0·15	0·14	0·12
BP02 = 7402	Quad. 2-input positive NOR gates ..	0·15	0·14	0·12
BP03 = 7403	Quad. 2-input positive NAND gates (with open collector output)	0·15	0·14	0·12
BP04 = 7404	Hex Inverters	0·15	0·14	0·12
BP05 = 7405	Hex Inverter (with open-collector output)	0·15	0·14	0·12
BP10 = 7410	Triple 3-input positive NAND gates ..	0·15	0·14	0·12
BP13 = 7413	Dual 4-input Schmitt trigger	0·29	0·26	0·24
BP20 = 7420	Dual 4-input positive NAND gates ..	0·15	0·14	0·12
BP30 = 7430	8-input positive NAND gates	0·15	0·14	0·12
BP40 = 7440	Dual 4-input positive NAND buffers ..	0·15	0·14	0·12
BP41 = 7441	BCD to decimal nixie driver	0·67	0·64	0·58
BP42 = 7442	BCD to decimal decoder (4-10 lines, 1 of 10)	0·67	0·64	0·58
BP46 = 7446	BCD-to-seven-segment decoder driver ..	2·00	1·75	1·50
BP47 = 7447	BCD-to-seven segment decoder drivers (15V outputs)	0·97	0·94	0·88
BP48 = 7448	BCD-to-seven-segment decoder driver ..	0·97	0·94	0·88
BP50 = 7450	Expandable dual 2-input and-or-invert ..	0·15	0·14	0·12
BP51 = 7451	Dual 2-wide 2-input NAND-or-invert gates	0·15	0·14	0·12
BP53 = 7453	Quad. 2-input expandable NAND-or-invert	0·15	0·14	0·12
BP54 = 7454	4-wide 2-input NAND-or-invert gates ..	0·15	0·14	0·12
BP60 = 7460	Dual 4-input expander	0·15	0·14	0·12
BP70 = 7470	Single-phase J-K flip-flop	0·29	0·26	0·24
BP72 = 7472	Master-slave J-K flip-flop	0·29	0·26	0·24
BP73 = 7473	Dual master slave J-K flip-flop	0·37	0·35	0·32
BP74 = 7474	Dual D type flip-flop	0·37	0·35	0·32
BP75 = 7475	Quad. latch	0·47	0·45	0·42
BP76 = 7476	Dual J-K with pre-set and clear ..	0·43	0·40	0·38
BP80 = 7480	Gated full adders	0·67	0·64	0·58
BP81 = 7481	16-bit read-write memory	0·97	0·94	0·88
BP82 = 7482	2-bit binary full adders	0·97	0·94	0·88
BP83 = 7483	Quad. full adder	1·10	1·05	0·95
BP86 = 7486	Quad. 2-input exclusive Nor gates ..	0·32	0·30	0·28
BP90 = 7490	BCD decade counter	0·67	0·64	0·58
BP91 = 7491	8-bit shift registers	0·87	0·84	0·78
BP92 = 7492	Divide-by-twelve counters	0·67	0·64	0·58
BP93 = 7493	4-bit binary counters	0·67	0·64	0·58
BP94 = 7494	Dual entry 4-bit shift register	0·77	0·74	0·68
BP95 = 7495	4-bit up-down shift register	0·77	0·74	0·68
BP96 = 7496	5-bit parallel in parallel out shift-register	0·77	0·74	0·68
BP100 = 74100	8-bit bistable latches	1·75	1·65	1·55
BP104 = 74104	Single J-K flip-flop equiv. 9000 series ..	0·97	0·94	0·88
BP105 = 74150	Single J-K flip-flop equiv. 9001	0·97	0·94	0·88
BP107 = 74107	Dual master slave flip-flop	0·40	0·38	0·36
BP110 = 74110	Gates master-slave flip-flop	0·55	0·53	0·50
BP111 = 74111	Dual data lock-out flip-flop	1·25	1·15	1·00
BP118 = 74118	Hex set-reset latches	1·00	0·95	0·90
BP119 = 74119	Hex set-reset latches, 24-pin	1·35	1·25	1·10
BP121 = 74121	Monostable multivibrators	0·67	0·64	0·58
BP141 = 74141	BCD-to-decimal decoder/driver	0·67	0·64	0·58
BP145 = 74145	BCD-to-decimal decoder/driver. O/C ..	1·50	1·40	1·30
BP150 = 74150	16-bit data selector	1·80	1·70	1·60
BP151 = 74151	8-bit data selectors (with strobe) ..	1·00	0·95	0·90
BP153 = 74153	Dual 4-line-to-1-line data	1·20	1·10	0·95
BP154 = 74154	4 to 16 line decoder	1·80	1·70	1·60

ANOTHER BI-PAK FIRST !

THE NEW S.G.S. EA 1000 AUDIO AMPLIFIER MODULE

***Guaranteed not less than 3 Watts RMS.**

Especially designed by S.G.S. incorporating their proven Linear I.C. Audio Amp. TA/621 providing unlimited applications for the enthusiast in the construction of radios, record players, Audio and Stereo units. Also ideal for intercom systems, monitoring applications and phone answering machines. OTHER USES: *portable applications where supply rails as low as 9V are of prime importance.*

- Sensitivity 40 mV for 1 watt. VOLTAGE GAIN 40 dB but can be varied up to 73 dB for some applications.
- Signal to Noise Ratio 86 dB.
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- Suitable for 8—16 Ohm loads.
- Overall Size 2 in. × 3 in. × ¾ in.
- Typical Total Harmonic distortion at 1 watt less than 1%.

* Supply Voltage (Vs) = 24V 15 ohm load.
Modual Tested and Guaranteed.
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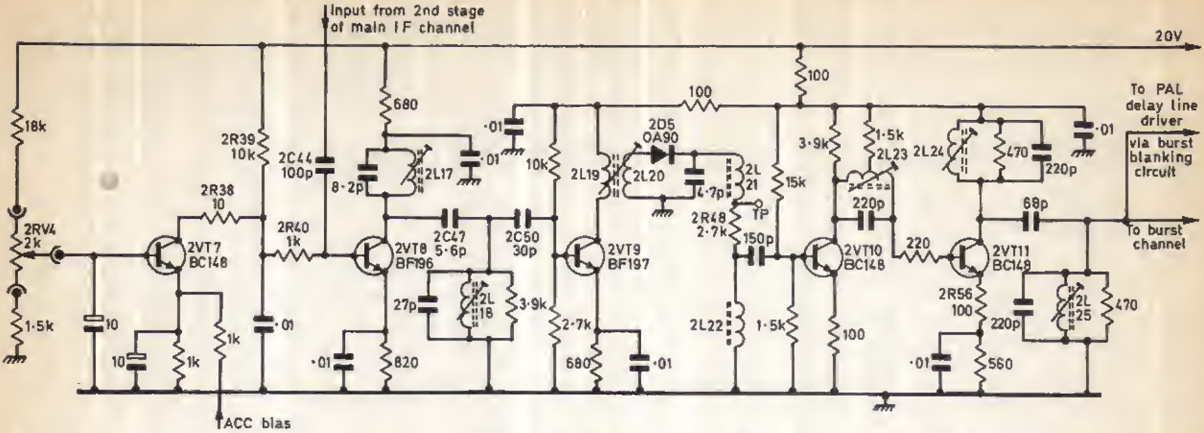


Fig. 7: Chrominance i.f., a.c.c., detector and bandpass amplifier circuits used in the RBM single-standard colour chassis.

a.c.c. rectifier D1 and a *positive* bias is fed to Tr1 base. This bias overcomes the back-bias tapped from VR2 so that Tr1 is brought into conduction. Current thus flows via D2 into Tr2 base circuit, and the increased current flowing through VR1 increases the positive bias on Tr2 base. The result is forward a.g.c. at Tr2 base, reducing the gain of this stage: the system provides not less than 6dB of a.c.c.

Improved Blanking

The Mark II chassis also features improved burst blanking based on the use of a two-diode clamp circuit with positive- and negative-going drive pulses. The circuit is shown in Fig. 6. Tr1 is the second chroma channel transistor and diodes D1 and D2 replace D2 in Fig. 3. The symmetrical diode clamp circuit is located between the final coupling transformer and the PAL delay line. During the line flyback period the positive pulse at D1 anode and the negative pulse at D2 cathode make both diodes conduct: when this happens any signal at T1 secondary is effectively shorted to chassis through C1 and C2.

Chrominance IF Channel

While it is usual to find the a.c.c. operating in the chroma channel the RBM chassis used in recent Bush and Murphy receivers tackles this in a different manner. The design is based on the use of a separate chroma detector which is fed from its own i.f. channel as shown in Fig. 7. Signal from the main i.f. channel is coupled to 2VT8 base through 2C44 while the collector of this chroma i.f. amplifier transistor drives the bandpass-coupled pair 2L17 and 2L18, with top-

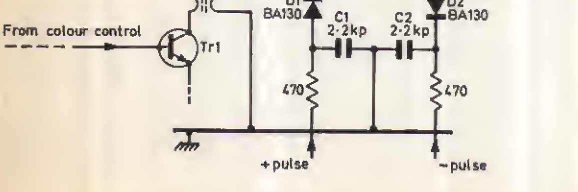


Fig. 6: Two-diode burst blanking circuit.

capacitance coupling provided by 2C47. The signal across 2L18 is then coupled via 2C50 to another chroma i.f. stage 2VT9 whose collector circuit drives the chroma detector 2D5 from transformer 2L19/2L20.

Controlled Stage

The controlled stage is 2VT8 the base bias of which is controlled by 2VT7 collector current flowing through 2R38 and 2R39 since 2VT8 base is connected to the junction of these two resistors through 2R40. As the collector current of 2VT7 is dependent on the bias at its base it follows that by adjusting 2RV4 the bias at the base of both 2VT7 and 2VT8 will alter and in this way the gain of the chroma i.f. amplifier channel is regulated: 2RV4 is in fact the colour control.

2VT7 makes it possible to apply the a.c.c. bias also to the i.f. channel. As the circuit shows this bias is applied to 2VT7 emitter. The bias is positive and increases in value with increasing burst amplitude (it is obtained by rectifying the burst signal). As the bias increases, the collector current of 2VT7 decreases since it is an npn transistor. The reduced voltage drop across 2R39 increases the base bias of 2VT8 and its gain is reduced through forward a.g.c. action.

RBM Chrominance Channel

The chroma channel proper consists of 2VT10 and 2VT11 driven from the chroma detector 2D5. The detector load is 2R48 and the purpose of choke 2L21 is to remove unwanted residual i.f. carrier signal. Choke 2L22 in series with the load provides the required compensation to maintain the response of the chroma circuit. Rapid roll-off at the intercarrier frequency is provided by the bifilar-T trap 2L23 and associated components while the bandpass characteristic is provided by the top coupled bandpass pair 2L24 and 2L25. The latter feeds the PAL delay line driver via a burst blanking gate, and also the burst channel. A degree of response correction is provided by the unbypassed part of 2VT11's emitter circuit (2R56).

We shall be looking at the following circuitry which includes the colour killing and burst blanking arrangements next month along with the response characteristics of this particular chassis.

LONG-DISTANCE TELEVISION

ROGER BUNNEY

UNFORTUNATELY Sporadic E conditions declined somewhat during the latter part of August following an excellent first week. Indeed the signals encountered during the first week were some of the best of the present season. After August 8th conditions fell off but began to pick up from the 25th—certainly as far as Sp.E is concerned. Tropospherics were above average in the middle of the month and again towards the 26th, giving the various v.h.f./u.h.f. ORTF reception from Northern France. Other enthusiasts noted signals on v.h.f./u.h.f. from the Low Countries and West Germany (see our correspondents' letters). My own log for the period is as follows:

1/8/71 USSR R1, R2; TVP (Poland) R2; MT (Hungary) R1, R2; CT (Czechoslovakia) R1 and R2 twice; JRT (Yugoslavia) E3, E4; RAI (Italy) 1A, 1B; ORF (Austria) E2a, E4; TVE (Spain) E3; BRT (Belgium) E2 via trops.; also unidentified signals.
2/8/71 NRK (Norway) E2, E3; SR (Sweden) E2; RAI IA; plus unidentified signals.
3/8/71 USSR R1; JRT E4; ORTF (France) F2.
4/8/71 CT R1; USSR R1, R2; ORF E2a; also unidentified signals.
7/8/71 Rumania R2; TVP R1, R2; USSR R1, R2; ORF E2a; JRT E3, E4; SR E4; NRK E2, E3; West Germany E3; RAI IA, IB; TVE E2, E3, E4.
8/8/71 USSR R1, R2; TVP R1, R2; CT R1; MT R1; JRT E3; RAI 1A.
9-10/8/71 BRT E2 trops.
12/8/71 USSR R1; TVE E2, E3; West Germany E4.
13/8/71 DFF (East Germany) E4.
14/8/71 USSR R1, R2; TVP R1, R2; NRK E2, E3; SR E2; DFF E4.
15/8/71 TVP R1.
16/8/71 Various Northern French trop. signals.
17/8/71 Various trops. including NOS (Holland) E4; BRT E2.
20/8/71 DFF E4.
21/8/71 BRT E2.
22/8/71 West Germany E2; BRT E2.
23-24/8/71 BRT E2.
26/8/71 NRK E2; SR E3.
27/8/71 USSR R1; TVP R1, R2 (extremely strong R1 signals); SR E2, E3.
28/8/71 SR E2, E3; NRK E3, E4; TVP R1; RAI IB.
29/8/71 DFF E4; TVP R1; BRT E2 (trops).
30/8/71 DFF E4; BRT E2 (trops).

It is interesting to note that the openings towards the end of the month seemed to occur during the evenings, in each case favouring the East European and Scandinavian directions. An additional opening was noted by Maurice Opie (Ringwood) on August 25th at Midday with various East European stations.

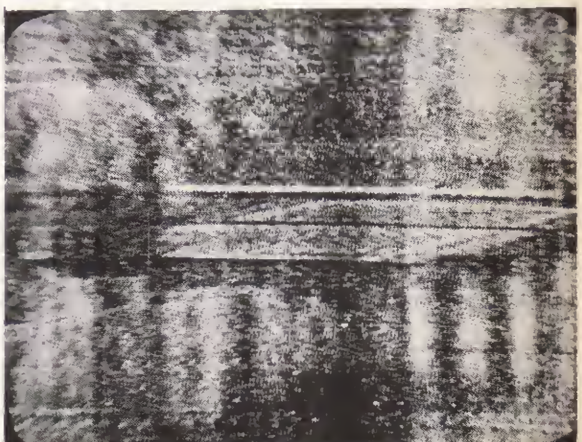
Too late for the last column was the appearance on July 31st and August 1st of the Czechoslovak electronic test pattern on ch.R2. This is similar to the CS U 01 type but with a different identification consisting of three letters the first two of which were LT. On both occasions it was observed floating with the usual Czechoslovakian test card whilst the companion pattern was on ch.R1 again floating with the test card. A test card of a different type was noted by Maurice Opie on August 12th. At 1605 he observed on ch.R1 a slow fading test card which lasted for four minutes before eventually disappearing completely. It consisted of a dark square standing on end within which was a lighter coloured circle containing what appeared to be three letters. Surrounding the square were various lines and light coloured corner circles. The overall background was greyish. The characteristic type of reception suggests a long propagation path, either single or double hop, and we are wondering if indeed this may be Bulgaria. At present we are awaiting the test card information from Bulgaria for our Data Panel series so with any luck we will shortly know the answer to this mystery reception. Has anyone seen anything like it?

Another mystery from Hugh Cocker of Mayfield, Sussex. He has reported seeing the USSR 0249 card with alternative identifications (other than reported in September's column). Apparently he has noted the card carrying the identification TA5 0249 or possibly TAS 0249. I have personally seen only the CCCP variation to this card but will be keeping a much closer watch in future!

News Items

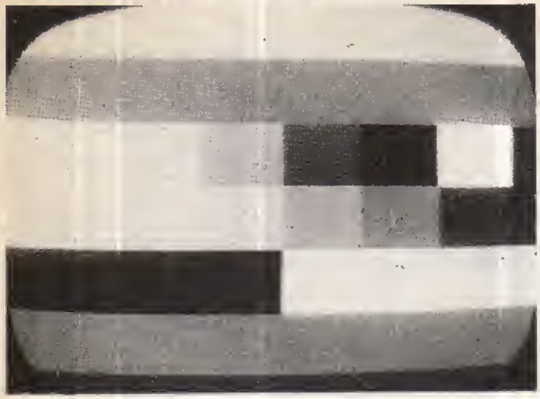
Finland: We understand from a contact that YLE are not too happy with the performance of the Tampere ch.E2

ORF Austria identification slide.

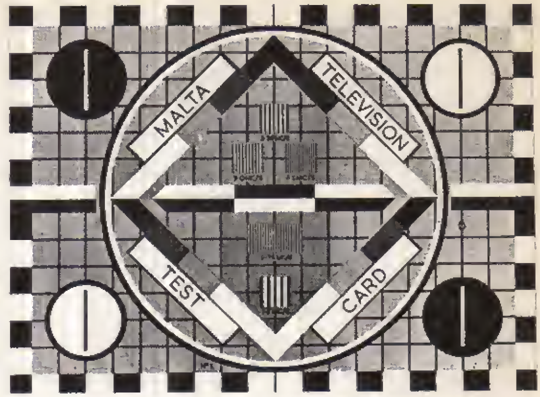


Ping-pong from Peking Television.

DATA PANEL 4—2nd series



Standard Pattern: Colour Blockboard



Malta Television Service Test Card

Gibraltar Television Identification Slide Standard Test Card G is used

Cyprus Broadcasting Corporation Identification Slide. Standard Test Card G is used

Photographs courtesy P. D. van der Kramer, Gibraltar Broadcasting Ltd., The Malta Television Service Ltd., Cyprus Broadcasting Corporation.

(TV2) transmitter and that it is likely to be replaced with a u.h.f. transmitter in the future. This is extremely bad news: Tampere is possibly the most frequently received DX-TV transmitter—certainly in the UK—and if it closes Finland is going to be a most difficult country to receive. The other Band I transmitters there are more distant and are rarely received. If this closure goes through Finland is going to become as rare as Bulgaria—at least in the Southern part of the UK.

Other news from Finland: Sippola ch.E49 is now on test 0700-1400, power unknown; Lahti ch.E40 will be opened during the first half of 1972 with either 600kW or 1000kW e.r.p.

USA: Our friend Ferdinand Dombrowski of the WTFDA Milwaukee has advised us that the World's most powerful u.h.f. transmitter is now in operation: WCCB-TV on ch.A18 has increased its power to a full 5MW (5000kW) e.r.p. The transmitter is situated at Charlotte, North Carolina and relays the ABC Network.

Austria: We understand that ORF are to drop the Telefunken card and use the electronic card on both v.h.f. and u.h.f. Fortunately an identification—either ORF 1 or ORF 2—will be carried. P. D. van der Kramer has kindly sent a photograph showing an ORF identification slide. This impressive caption is shown when the National Anthem is played—I assume at the start and close of transmissions.

Belgium: We have referred in previous columns to this country using the ORF type electronic card. The only transmitter to use this is Wavre, on u.h.f. only—chs. E25, E28.

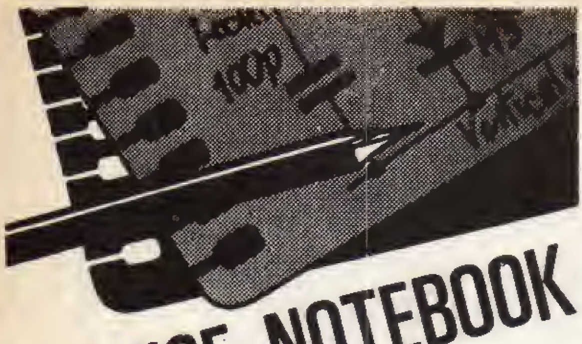
Sunspots: Predicted smoothed monthly counts for the next six months: August 62, September 60, October 58, November 56, December 54, January 52. Courtesy Swiss Federal Observatory, Zurich.

Reception During the Winter Months

With the approach of Winter reception tends to fall to a rather low level, certainly compared to Summer conditions. It is still possible however to receive signals over similar distances to those of Sporadic E during these quiet periods—by means of *meteor shower ionisation.* The Earth is bombarded throughout the 24-hour period by particles—often no larger than grains of sand—from space. These meteorites burn up due to friction when they enter the various layers that surround the Earth and can cause intense ionisation at E-layer heights. Such ionisation is localised and of short duration but it is nevertheless possible to obtain signal reflection over distances up to 1400 miles. Fast-travelling particles burn up sooner and thus higher, increasing the possible skip length; slow-travelling particles burn up lower down to give a shorter skip path.

With this type of signal the receiving equipment must

—continued on page 39



SERVICE NOTEBOOK

G. R. WILDING

Weak Line Hold

"Intermittent lines across the picture" was the complaint with a Philips Model G19T210 and inspection showed that line lock on 405 could only be obtained with the line hold control fully clockwise. 625 was unobtainable in the district. In these Philips models there is an additional preset line hold control mounted on the chassis. The correct setting-up drill is first to switch to 625, lock the picture with the main (exterior) control, then switch to 405 and lock by adjusting the preset. We found that this preset control was also in an extreme position so it appeared that either one or both of the ECC82 valves in the line generator/a.f.c. circuit (Fig. 1) were of low emission or that a component had changed value.

Changing both valves scarcely affected the locking position so we commenced voltage checking. All voltages were about normal except at V2004B anode which should have been 42V on 405 but was only slightly over 30V. This voltage reduction could have been caused by either an increase in value of the anode feed resistor or a reduction in grid bias causing excessive anode current: there was no capacitor to chassis which might have developed a leak. The 56kΩ anode resistor proved to be in order but the 470kΩ grid resistor—returned to the h.t. line instead of to chassis—was well under 400kΩ and therefore failed to sufficiently offset the negative self-bias developed by the valve. Replacing this resistor restored a midway locking position to the line hold control.

On subsequent test we found that if the contrast was advanced too far almost every other line would be intermittently displaced to the right by about a quarter of an inch, resulting in the appearance of two pictures slightly horizontally displaced. Our first suspicion was that excessive contrast was impairing the action of the line sync amplifier and phase comparator valve V2003. However we then found that this unusual effect could also be obtained by advancing the brilliance control too far while if it was turned to its extreme position the effect disappeared leaving a normal if very milky picture. The fact that this fault could be produced by two different actions completely changed the situation for the only common effects produced by increasing the contrast or brilliance were (a) the mean c.r.t. grid-cathode voltage would be reduced and (b) the e.h.t. current demand would rise.

As in all flywheel sync circuits a reference pulse feed from the line output stage is fed to the discriminator—or comparator in this Philips model—which develops the d.c. control potential used to control the line generator. It seemed likely therefore that the extra e.h.t. demand was affecting this pulse. As the DY87 valve was most directly involved we tried changing this first, but with no improvement. On replacing the PL500 line output valve however the effect completely vanished.

It is worth mentioning that in many colour receivers the "sensor" of the e.h.t. is the voltage developed across a low-value resistor in the cathode lead of the line output pentode. When due to increased output this voltage rises above a predetermined level a beam limiting arrangement biases back a stage in the luminance channel.

While on the subject of these Philips receivers it is worth noting that a not uncommon cause of insufficient width is an increase in value of one or both of the two 8·2MΩ resistors in the v.d.r. width stabilising circuit. Normally lack of width when the h.t. rail voltage, line output valve screen voltage and line oscillator anode voltage are normal tends to make one suspect the boost reservoir capacitor or shorted turns on the line output transformer. But in these models as in the previous Style 70 series first check these two resistors (R427 and R457, see Fig. 1, May 1971, page 313) between the boost h.t. rail and the stabilising v.d.r. Similarly in other makes with

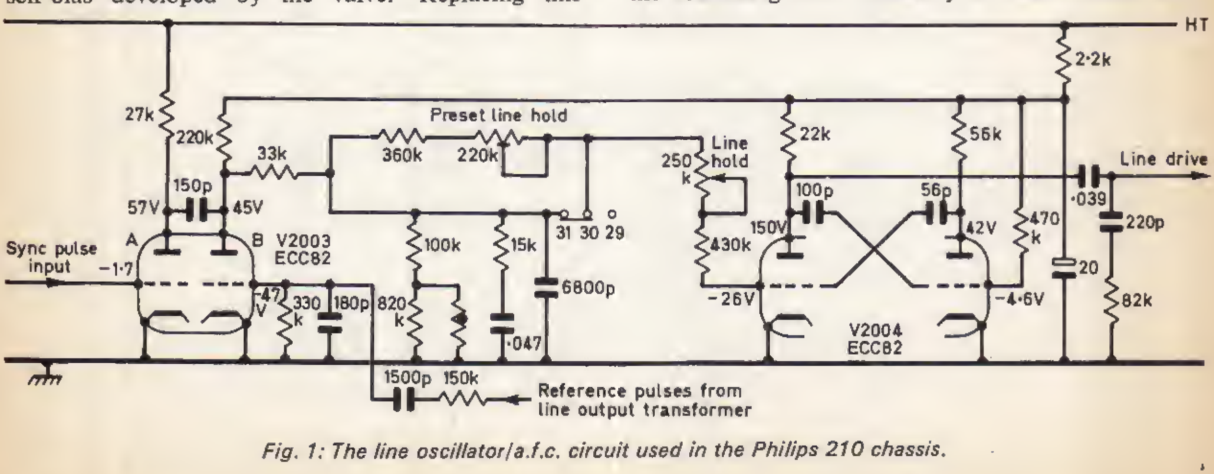


Fig. 1: The line oscillator/a.f.c. circuit used in the Philips 210 chassis.

v.d.r. width stabilisation always check the feed resistors from the boost rail when the width is not up to standard and follow makers' instructions when resetting the width control as this directly affects the boost rail potential.

Delayed Colour

ALTHOUGH in a BRC colour model fitted with the 2000 chassis a black-and-white picture would always appear after the normal tube warm-up time it would sometimes be an extra ten minutes or more before colour developed. The owner found that a delayed start could usually be cut short by rapidly changing channels.

On test it was found that when colour failed to appear "ditching" the colour-killer by connecting a 100kΩ resistor between the positive l.t. rail and the base of the first chrominance amplifier VT13 failed to produce either the colour signal or confetti. This made it all the more likely that the reference oscillator was at fault, usually needing a slight electrical impulse to start it operating.

This suspicion was fully borne out when making voltage checks on this stage (VT4) and the d.c. amplifier (VT3) preceding it as test-prod application to most circuit points would instigate the reference signal. As no dry-joints were apparent we first tried a new oscillator transistor (BF115). This completely cured the warm-up delay.

No Vision or Sound

CALLED to deal with a no signal complaint in a modern single-standard KB model fitted with the VC200 chassis we found no raster either, indicating a line timebase fault. Sure enough the line output pentode was labouring, indicating lack of drive, and on replacing the PCF802 line oscillator valve normal results were restored. Why no sound? It is worth noting that in this and several other single-standard chassis (e.g. GEC/Sobell and the Bush/Murphy TV181S/V2016S series) the l.t. supply for the transistor stages is obtained from the line output stage—from a diode which rectifies the scan waveform fed to it from a low-voltage winding on the line output transformer. Failure of the line timebase will therefore remove the sound and vision signals as well. There have been reports of line oscillator failure being caused by defective capacitors in the PCF802 circuit, so these (C124-7) might need checking. As in the Philips 210 chassis lack of width can be caused by a changed value resistor in the width stabilisation circuit, in this case R159 (10MΩ).

LONG-DISTANCE TELEVISION

—continued from page 37

have high gain and be able to synchronise at once since only short bursts of signal are experienced with a meteor shower (abbreviated to MS). One needs of course to be able to tune exactly to the required channel. Signals throughout Band I are propagated by this means and the best time for random meteorites is the early morning. Signal duration can be anything from half a second to five seconds plus though the usual burst rarely approaches five seconds. If Band I is very active it may at times be worthwhile to check the lower end of Band III: MS at these frequencies does happen.

At times the Earth encounters the more active regular Meteor Showers which can produce spectacular reception. Forthcoming we can expect the Leonids November 15th-17th, the Geminids December 10th-14th and the Ursids December 22nd-23rd. Random reflections can occur at any time on any day. The distances usually experienced with this type of reception are between 700-900 miles but with the more active showers reception can become much more consistent and stronger with greater distances. We will be listing all the main meteor shower dates for 1972 next month.

From our Correspondents . . . .

Two interesting letters have come from a father and son DX-ing team: J. E. May (Orpington) and R. J. May (Ashford, Kent) have sent in details of their individual and combined reception. Both have experienced the prolonged and excellent conditions this season, including the Czechoslovakian CS U 01 pattern. Most other countries in Europe have also been received in their respective parts of Kent.

Frank Smales (Pontefract) has forwarded a long letter detailing the signals in Yorkshire for this season. Frank certainly seems to have been busy and asks our assistance with the identification of a mystery signal. On May 24th he noted on ch.E3 a clock at one hour ahead of our time —1625 BST. The clock had a second sweep hand and there was Arabic writing or script beneath the clock. He wonders if this might be Jordan as they are plus one hour BST (plus two hours GMT) at this time. It certainly sounds as if Frank may have received one of the exotics: a fine achievement if he has! We await further confirmation: did any one else note this signal at the above times?

Our French TV expert John Penruddocke (Salisbury) has been logging ORTF on various frequencies and advises us that ORTF-1 has been noted on 625-line tests on Wednesday mornings—in addition to Tuesdays as previously noted. On some mornings ORTF has been on test before 0800 BST and thanks to the improved tropospheric conditions John has been able to take full advantage of the many French transmitters available to him at his hill-top home.

Ian Beckett (Buckingham) writes to advise of his colour DX reception at the end of the month when the tropospherics improved to present him with two new West German u.h.f. transmitters and a number of old favourites.

An extremely interesting letter has arrived from an experienced TV-DX enthusiast in Australia. George Peterson of Ayr, Queensland, may be remembered by established DX enthusiasts for his F2/Trans Equatorial (TE) reception at the time of the last sunspot maximum. He has sent a large number of photographs which we are still looking through. One we selected at random shows very recent reception of China via TE. This was taken during the 1971 Ping-Pong matches in Peking and we have pleasure in including the photograph this month to show the characteristic mutiple-image effect of TE and also because the event was important in its own right—the photograph is possibly unique. The signal was received on ch.R1. A number of Chinese transmitters operate on this channel but we are not too sure from which location the signal originated. As a rough guide the nearest transmitter listed on this channel is Nanking, 4,750 miles!

Data Panels

With the colour blockboard we complete the series of standard patterns and are now going on to the patterns used by individual countries. Should other patterns which are commonly used come to hand in the future these will be included as standard patterns. We shall not be including transmitter details with the data panels as the number of transmitters now in operation is so considerable. Accurate transmitter lists are available from the EBU, etc.

TV TEST REPORT

E. M. BRISTOL

BIB TAPE-RECORDER MAINTENANCE KIT

THE Bib tape-recorder maintenance kit is manufactured by Multicore Solders Ltd. for the purpose of keeping all parts of the tape path through the recorder clean and free from the oxide deposits which build up on heads, guides and pinch wheels. These can cause wow and flutter and on the heads give rise to loss of high frequencies and incomplete erasure.

The Kit

Housed in a blue plastic wallet, the kit comprises a bottle of Bib instrument cleaner, two blue tape-head applicator tools and two white tape-head polisher tools, ten polisher sticks, one double-ended brush, a packet of cleaning tissues and an instruction leaflet. Retail cost of the outfit is 41p plus 8p p.t.

For larger use and workshops there is a Professional maintenance kit which consists of two bottles of cleaner, 24 blue applicator tools and 24 white polishers, 100 polishing sticks, two double-ended brushes and six packets of tissues. There are also four copies of the instruction leaflet. This kit retails at £2·80 plus 56p p.t.

Applicator and Polishing Tools

The applicator and polishing tools consist of plastic material about $4\frac{1}{2}$in. long and bent at one end to an angle of 150° where they are fitted with rectangular felt cleaning pads. There is no difference in the two types except for colour. This enables one to be kept clean for polishing while the other does the dirty work. The width of the pads is cut exactly to the size of the tape thus enabling them to fit into the tape-guides without missing any part of the surface——especially the corners—where deposits can build up. A snag with these tools is their thickness: the handle is $\frac{1}{8}$in. thick and the felt pad is $\frac{1}{4}$in. There are a number of recorders where it is impossible to get the tool anywhere near the heads while with some of those that are reachable it is necessary to remove the pressure pad to do so.

Polishing Sticks

The polisher sticks are intended for use where this difficulty is encountered. These are short sticks with a pad of cotton wool encasing each end. One end is used with the cleaning fluid for cleaning and the other for polishing. They can only be used once as the pad comes away from the stick when soaked in the fluid.

Use

Really though the tool is the more convenient if only it can reach the head. While there will always be some awkward recorders that defy any cleaning tool, the usefulness of the tools in this kit could have been extended by a reduction in size. Very little pressure is needed in cleaning—the fluid does the work of dissolving the deposits—so the part of the tool supporting the cleaning pad need be only thin. The pad itself could also be less thick and the total reduction in thickness of tool and pad would have enabled it to be used with many more recorders.

When dirty the pads can be cleaned with a little of the fluid and the cleaning tissues supplied. It is as well not to let them get too dirty in between cleans.

Cleaning Fluid

A generous supply of fluid is given. Methylated spirit is generally used for this purpose but the makers claim that meths can have a deleterious effect on the rubber pinch wheel. The cleaning fluid does not have this effect and so can be used with confidence on rubber parts. Some badly caked heads however failed to "come clean" and recourse had to be made to the meths bottle.

Double-ended Brush

The double-ended brush is a useful item: it is constructed of twisted wire, one end being a conventional circular brush and the other a spiral-type pipe-cleaner pattern. Both enable fluff to be cleaned out of the nooks and crannies around the heads and guides. Being made of wire the brush can be bent to any angle needed to reach into difficult positions.

Conclusion

All in all this is a useful and inexpensive kit which should prove its worth to all concerned with the use and maintenance of audio and videotape recorders.

NEXT: KLIK RIVETER

YOUR PROBLEMS SOLVED

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PYE CT72

The picture detail is OK on black-and-white but the colours tend to run into horizontal bands across the screen. The fault can be cleared temporarily by changing channel several times. Also there are on some occasions intermittent changes in colour intensity.—G. Ryle (Leicester).

The reference oscillator on the decoder panel is intermittently off lock. Adjust RV10 then set a.p.c. bias control at the back of the decoder panel for optimum results, which should correspond to 5V at TP5.

PHILIPS 1768U

There is an oscillation on the sound—it is not always present—which can sometimes be stopped by tapping the cabinet sharply.—R. Quornley (Chester).

The bias for the two sound i.f. amplifier stages is derived from the cathode circuit of the PCL83 audio valve. You should thus check the 100μF electrolytic in the PCL83 audio valve cathode circuit (wired from pin 7 to chassis).

FERRANTI T1084

Due to asynchronous working the picture rolls erratically on u.h.f. every 30 seconds or so. It also causes line tearing approximately a third of the way down the screen, especially on long camera shots. Extra main smoothing has been added and the 30PL13 field timebase valve replaced without making much difference.—A. F. Fellows (Dorking).

We suggest you check the PCL84 video amplifier screen decoupling capacitor C55 (2μF) to pin 9. Check the PCL84 itself as well and the 30FL1 sync separator screen decoupling capacitor C122 (2μF) to pin 7.

PYE V700A

There is a wavy "bulge" on the left-hand side of the picture on this set. Sometimes this moves slowly from top to bottom and sometimes from bottom to top. I understand this is due to faulty smoothing.—G. Freeman (Axminster).

You are correct in suspecting the smoothing and a good electrolytic bridged across each smoothing capacitor in turn will show up the defective one. Suspect also heater-cathode leakage in the line output valve.

MURPHY V470

After the set has been on for about an hour there is motor-boating which drowns the normal audio. The picture is good at all times: could you explain the sound trouble?—C. Graper (Blackpool).

The problem is that the sound i.f. amplifier V8 (6F23) is going unstable. Check especially the screen decoupler C49 (0·001μF). This can often be stabilised by being moved slightly.

PYE CT78

There are vertical striations on the left-hand side of the screen. These were accompanied once by a series of broad, faint vertical coloured bars across the whole screen.—G. Dodson (Goole).

It seems that the flyback blanking transistor in the cathode circuit of the luminance output valve V6 on the colour-difference amplifier panel is faulty. This is VT28, type BC147.

ULTRA 1760B

When the volume is increased above a certain level the picture becomes ragged indicating loss of the sync. The audio circuits appear to be all right and the trouble does not seem to be acoustically induced because an external loudspeaker has been tried without the fault being cleared.—F. Best (Stockton).

The effect you describe seems to indicate coupling via the h.t. line and we suggest therefore that you check the main electrolytics C98, C99 (100+200μF) and the two sub-h.t. feed decouplers C76 in the audio section and C50 in the video and vision i.f. h.t. feed (8+16μF).

PYE 11U

There is foldover from the bottom with shrinking from the top. Eventually only a narrow band is left across the centre of the screen: this change coincides with the warm up of the set. I have changed the relevant valves and the voltages appear to be correct. Might the output transformer be at fault?—J. Jepsom (Hanwell).

A faulty output transformer is unlikely to be the cause of this trouble. Suspect the cathode components of the output valve, R77 (390Ω) and C74 (200μF) and the feed resistor to this stage R75 (270Ω).

GEC 2022

The line hold needs constant adjustment and it is sometimes possible to achieve line lock only by adjusting L65. The line timebase valves have been renewed but the fault remains.—H. Short (Sunderland).

We suggest you replace the flywheel line sync discriminator diodes MR1 and MR2. If the diodes are not at fault—they must of course be a matched pair—check for dry-joints and poor contacts on the timebase panel.

PYE CTM17T

There are two faults in this 17in. set. First the sound went. It was restored by fitting a new PCL83 but subsequently it began to fade again. Secondly there is a gap about 2in. wide at the top and bottom of the screen. The field timebase valves have been replaced without effecting a cure.—G. Upton (Brighton).

A common fault producing the two symptoms you are experiencing is inadequate smoothing. You can check this by bridging a good electrolytic across each smoother in turn. Alternatively suspect a low h.t. rectifier.

FERGUSON 3623

The first fault to occur was loss of horizontal and vertical hold on both systems: there are dark diagonal bands across the screen which rotate first in one direction and then the other. Secondly there is complete loss of sound and vision on u.h.f. although the raster is OK: both tuner valves have been replaced.—G. Orford (West Bromwich).

First check the PFL200 and its operating conditions, also the line sync diode W4. The absence of u.h.f. signals is probably due to the switch on the v.h.f. tuner being inoperative owing to the bowden cable being out of position: note the effect of operating this by hand.

BUSH TV96

This set has developed a severe field fault. After the set has been on for a time varying between half to two hours a very fast field spin develops, so fast in fact that the picture appears as if there was a ghost signal superimposed. The field hold functions over the whole control range. All valves likely to cause the trouble have been replaced without success and also the coupling capacitor to the grid of the valve where the hold control is connected.—G. Taverner (Rhyle).

We suggest you change C108 (0·05μF) the field multivibrator cathode decoupler and R121 (620kΩ) which is in series with the field hold control.

GEC 2019

The problem with this set is no sound or vision on v.h.f. although the raster is OK. Reception on u.h.f. is normal. The valves in the v.h.f. tuner have been replaced without effecting a cure.—D. Bridle (Cheadle).

If the voltage supplies to the v.h.f. tuner are present it would appear that one (or more) of the resistors in the tuner have changed value. Check in particular the 5·6kΩ and 6·8kΩ resistors in the oscillator anode circuit. To gain access it is necessary to dismantle the tuner: these resistors will be found at the top under a strip of tape.

PHILIPS 511

The trouble is poor focus at the centre of the screen and defocusing on bright parts of the picture—the scanning lines are sharply defined at the side edges. The focus control is set at one end of its travel (the lower-voltage end). The e.h.t. shunt bias adjustment has been correctly set in accordance with the manual. The only other fault that has been experienced on this set is the line output valve screen feed resistor going open-circuit: this was replaced using a component with 5W rating.—L. Charlesworth (Harlow).

As the focus control is set hard to one end we suggest you check the values of the high-value resistors in series with it on the low-voltage side (R5046-5050). Also check the EY51 focus rectifier.

RGD 624

V.H.F. programmes appear only with the contrast control set to one extremity of its travel whilst on changing to u.h.f. the contrast control has to be turned to its other extremity in order to obtain a picture. There is always a raster present.—T. Wimple (Stockton).

The problem is associated with the pulsed a.g.c. amplifier stage V3B, the triode section of the video PCL84. First try opening the panel and adjusting the preset contrast control 117: this operates on v.h.f. only and could solve your problem. Otherwise you will have to check the various resistors associated with the triode section of the PCL84, in particular 127 and 128 which feed the grid, the contrast control 124A and its series resistor 125, and the 10MΩ resistor 132.

COSSOR CT1972A

There is no picture or raster. A low line whistle can be heard but the DY86 does not light up. If the PY800 is removed the whistle is harsh, the DY86 lights up and a small picture appears in the centre of the screen. When the PY800 top cap is removed the glass of the DY86 gives a blue discharge to a screwdriver blade. The DY86 and its heater winding have been replaced, also the PY800 and PL36 which is overheating.—K. Johnson (Wolverhampton).

The boost reservoir capacitor is almost certainly short-circuit. In this chassis it consists of two 0·05μF capacitors connected in parallel to form 0·1μF. They are mounted together under the PCL85. You could replace both with a single 0·1μF capacitor rated at 1kV.

BUSH T67

The trouble is fading after the set has been on for a little time. Sometimes the fading disappears and the picture becomes almost perfect, then the fading starts again—it is mostly apparent at the bottom of the screen. The main smoothing electrolytic has been replaced.—J. Corvin (Barking).

Your description suggests that the PCF80 video amplifier valve on the left side has heater-cathode leakage. This would produce a faded picture, particularly at the bottom. Also check the EB91 in the vision section by swopping with the other EB91: hum on sound will be the result if the vision EB91 is responsible for the trouble, the vision then being clear.

PYE CT72

Sometimes after about four hours' operation the following colour fault will appear: the yellow and blue content fade and the whole screen appears darkish green. On operating the colour control flesh colours appear pale to bright pink. The condition lasts for ten minutes to a quarter of an hour after which the picture slowly returns to normal. Switching off the blue does not give the same effect as the fault.—S. Graham (Derby).

Your B−Y signal is intermittent. Check around the small blue link beside the delay line on the decoder panel and also the colour-difference output stage V9 (PCL84) on the colour-difference amplifier panel.

PHILIPS 19TG156A

After about half an hour the sound disappears and the picture bows in at each side: when the volume control is turned down the sound and picture return to normal.—G. Braithwait (Southport).

The PCL83 sound output valve could be responsible for the conditions you describe. Check this by replacement, examine the drop-off bias resistor R223 and check the 250μF electrolytic C110 which smooths the HT1 line from which the stage is fed. If there is a positive voltage at the hot end of the volume control replace the coupling capacitor C228 from pin 1 of the PCL83.

MARCONIPHONE 4705

On turning up the contrast the verticals are impaired, pulling to the left. I have readjusted to the maker's instructions without improving matters. The set works all right when the contrast control setting is below normal.—F. Bilton (Stoke).

The fault sounds like an a.g.c. one. If you have set the set a.g.c. preset R125 correctly you should have a difference of 4·9V across the luminance delay line output pins using method 2 detailed in the manual. Check the voltages around the a.g.c. amplifier VT106 and the associated components—particularly the electrolytic C134. If this does not reveal the source of the trouble make voltage checks around the post luminance delay line amplifier VT201 and then the vision i.f. stages VT101-VT104—in that order. These sets are rather prone to this fault however when the contrast is at maximum. Try adjusting the local/distant control R5 (r.f. gain).

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TELEVISION NOVEMBER 1971

TEST CASE

107

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 few weeks after the installation of an ITT colour receiver the owner complained of the picture disappearing after about 30 minutes running, leaving a bright, horizontal line on the screen, and that the trouble could be cleared by switching the set off for a few minutes. Invariably this action resulted in the set working normally for the rest of the evening. The technician dispatched to correct the fault found on arrival that the set was working normally. After a couple of cups of tea however and with the set now quite warm the symptom described suddenly appeared.*

After quickly reducing the setting of the brightness control to avoid damaging the screen the technician turned back the volume, asked for complete quiet and then put his ear close to the rear of the set while adjusting the vertical hold control. Mumbling appropriately, the technician then switched off, removed the back cover while his soldering iron was running up to temperature and within 15 minutes, without making any replacement, had remedied the fault. The remedy was proved to be permanent by a delighted customer ringing a week later and extolling the singular genius of the field technician!

How did this technician discover so quickly the source of the fault without making any measurements and then clear it with a soldering iron alone? See next month's TELEVISION *for the solution and for a further item in the Test Case series.*

SOLUTION TO TEST CASE 106

Page 570 (last month)

While it is well known that a fault in the Y delay line can horizontally displace the luminance from the colour, it is not so well known that in receivers with primary-colour drive in particular change in the bandwidth characteristics of one primary-colour channel with respect to those of the others can result in a similar displacement with respect to the appropriate *colour*.

Emitter compensation is often used in each primary-colour amplifier and in the case in question it was found that a capacitor in the emitter circuit of the blue channel had gone open-circuit. This affects the gain of the channel and also widens the bandwidth because of the negative current feedback introduced. Thus the blue signal arrived at the blue gun a small fraction of a second before the other two signals arrived at their guns. Replacing the faulty component cleared the fault but the technician then found it necessary to readjust the primary-colour channel gains for the best display.

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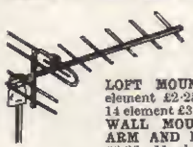

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