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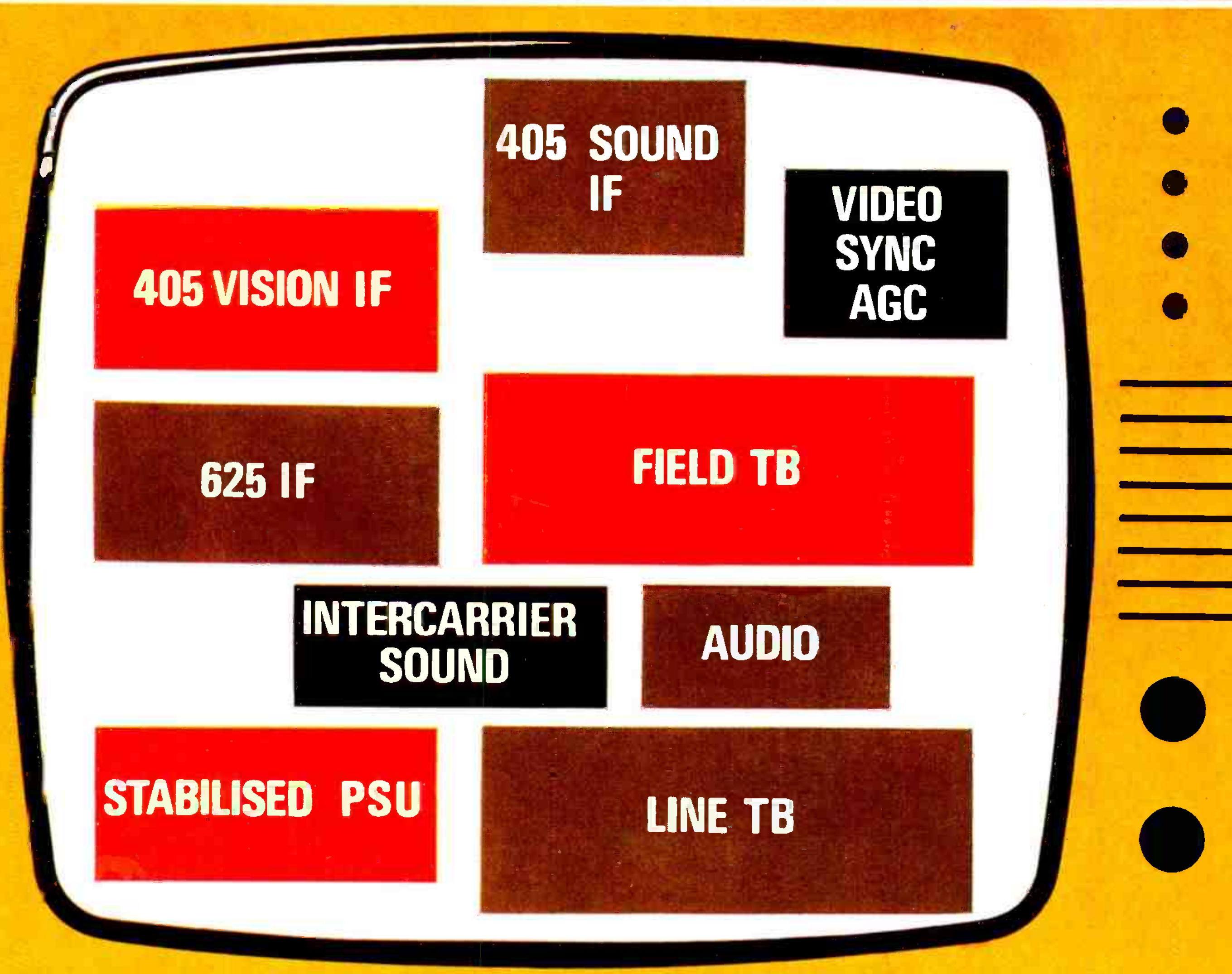

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27	,,	3·3K	390K
33	,,	3·9K	430K
39	,,	4·3K	470K
43	,,	4·7K	560K
47	,,	5·6K	680K
56	,,	6·8K	820K
68	,,	8·2K	1M
82	,,	10K	1·2M
100	,,	12K	1·5M
120	,,	15K	1·8M
150	,,	18K	2·2M
180	,,	22K	2·7M
220	,,	27K	3·3M
270	,,	33K	3·9M
330	,,	39K	4·3M
390	,,	43K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

SERVICING·CONSTRUCTION·COLOUR·DEVELOPMENTS

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JUNE 1971

OUTLOOK UNSETTLED

IT is usually a pleasant experience to have an eminent authority confirm one's views—especially when the original views expressed were in the form of a prediction or warning. The sort of feeling, one assumes, of Madame Osiris when her misty manifestations in the crystal ball actually prove to be true.

The present writer however gets no joy from a recent reminder of his leader in the December 1967 issue. Headed "Industry in Danger," it pointed out that the UK had already lost a large proportion of the transistor radio market, warned that audio products were following suit and posed the question "How long before *all* such products are imported?" How long in fact before we ended up importing TV sets as well now that we are hitched to the 625-line system?

The flashback was caused by Dr. F. E. Jones, managing director of Mullard Ltd., who while addressing American industrialists in New York recently noted that the production of electronic devices and equipment is being progressively taken over by low-cost labour countries at the expense of high-cost labour countries. He cited industry as shrugging off the fact that 80% of radio sets sold in the UK are imported on the grounds that the demand for monochrome and now colour TV receivers is of greater significance. But now even the market for TV receivers is being slowly taken over by imports—and there's nothing left to take its place!

Pressing on relentlessly Dr. Jones stressed that of the home market of 30 million receiving valves in the UK last year 15 million were imported. In the field of semiconductors 58% of the 1970 sales were imported and with i.c.s the import figure was as high as 74% of the total market!

Another voice recently raised was that of John Bowes, chairman of the British Radio Cabinet Manufacturers' Association, who pointed out that during 1970 some 150,000 monochrome and 35,000 colour TV sets were imported into the UK.

As Dr. Jones explained the problem is not one of relative technological efficiency for the performance figures (sales per employee, pretax profits per employee etc.) are very close between such comparable giants as RCA and Matsushita. But the wage rates are less than one-fifth for Matsushita. And there are many countries where the labour costs are much lower still.

We do not pretend to know the answers to the problems besetting the UK television industry. But our crystal ball indicates that the situation is in danger of becoming more critical in the future.

W. N. STEVENS, *Editor*

THIS MONTH

THE NEXT ISSUE DATED JULY WILL BE PUBLISHED JUNE 22

We deeply regret having to record the death of our DX correspondent Charles Norman Rafarel. An Appreciation by Roger Bunney appears on page 372.

TELETOPICS

GOING DIGITAL

Nearly two and a half years ago in our January 1969 issue we reported on the BBC's work on the use of digital techniques for TV sound signal distribution—the sound-on-sync system. In this system the original sound signal is converted into a digital one and the pulses representing the digital signal are inserted into the line sync pulse period of the television waveform. This period of course does not otherwise carry any modulation. In this way the need for a separate sound distribution network is avoided while appreciable improvement in the quality of the sound signal is achieved. The signal is converted to digital form at the transmitting end of the distribution network and converted back to a sound signal at the reception end.

The BBC is now working on the extension of the use of digital techniques to video waveforms and demonstrated recently the use of these techniques in an experimental 625-405 line standards converter (a similar converter has been developed by the ITA's Experimental and Development Department). All programme material is now originated in 625-line form and converted down to 405 lines for the v.h.f. system. It is likely that the 405-line system will continue in use long beyond the useful life of the present generation of standards converters and for this reason the BBC's Research and Designs Departments are carrying out a programme of work aimed at providing a replacement converter with good performance and reliability at reasonable cost. Since the present converters were designed considerable progress has been made in the application of digital techniques to TV signals and it has therefore been decided that the new form of converter will process the signal after it has been converted to digital form.

The basic technique used in these digital systems is to sample the original video or sound signal at regular and extremely rapid intervals—for example in the sound-on-sync system the sound signal is sampled twice each line, or at a rate of 31,250 times per second in the 625-line system—and to convert these samples into digital form. The digital form used is the binary system in which there are only two signal conditions, 0 and 1. 0 is represented by the absence of a pulse and 1 by its presence. Thus a ten-digit signal—as used in the sound-on-sync system—would take the form of a group of pulses indicating for example 0101100110, i.e. a group of 0s and 1s totalling ten, the different possible combinations of 0s and 1s representing different levels of signal amplitude. Now since every sample of the original signal is separately described by a sequence of pulses it follows that the digital signal varies much more rapidly than the original signal. As on the other hand there are only two possible signal levels in the digital signal while the original signal can vary gradually through a given range of levels depending on its amplitude, the digital signal is much more resistant to interference and to variations in the characteristics of the transmission network than the original signal. Moreover a digital signal can be reconstituted substantially without error after being subject to moderate amounts of interference and distortion. The price of this advantage is the increased bandwidth necessary to convey the more

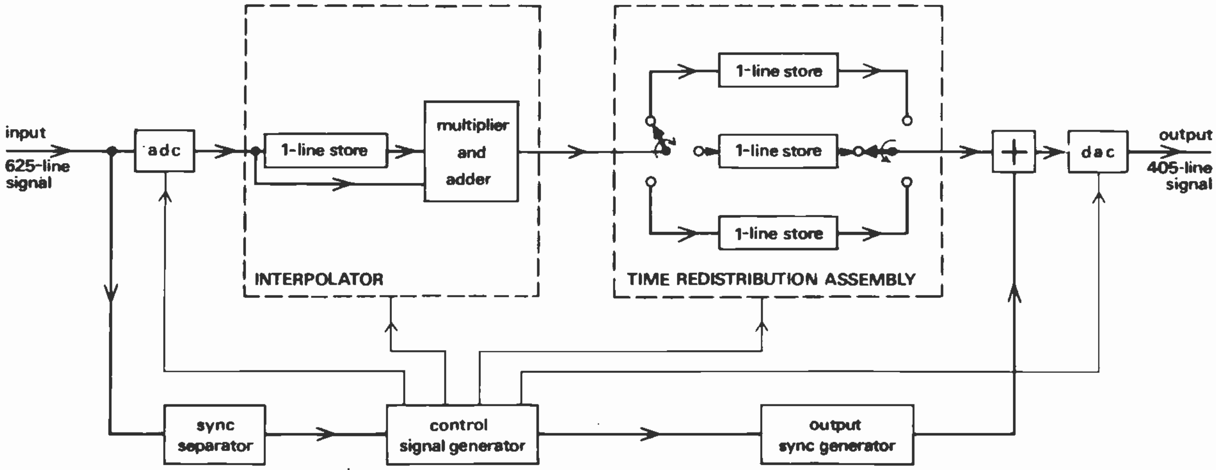


Fig. 1: Block diagram of the BBC's experimental digital line-store standards converter. Box adc is the analogue-to-digital converter and dac the digital-to-analogue converter.

rapid variations of the digital signal.

The BBC's experimental digital line-store standards converter is shown in block schematic form in Fig. 1. It will be seen that the incoming signal is first fed to an analogue-to-digital converter which converts the conventional "analogue" input into digital form while at the output it is converted back again to a conventional signal by the digital-to-analogue converter (the term analogue is used because conventional electronic signals are analogues of the actual light level or sound pressure waveform). Thus the conversion processes undertaken in the converter are carried out with the signal in digital form.

There are two main processes involved in 625-405 line conversion, interpolation and time redistribution. The 625-line input is first modified by the interpolator which derives 405 completely new lines of video information from the incoming 625-line information. This is done in such a way that each of the 405 new lines derived carries information corresponding to the scanning of the scene by a 405-line raster pattern: the process consists of combining the signals from two (in the experimental converter) or more successive input lines in accordance with the relative positions of corresponding lines in 625- and 405-line raster patterns, assuming the rasters to be superimposed. The 405 lines thus derived are however still of the same time duration as the original 625 lines so that there are gaps in the waveform which in consequence must be stretched. This is where the time redistribution process comes in, the 405 lines provided by the interpolator being individually adjusted in duration in this part of the converter so as to occupy the correct time intervals set by the 405-line scanning standard. This time redistributing process is effected by an arrangement of electrical stores. These use computer memory techniques and in the experimental converter the stores consist of assemblies of m.o.s. shift registers.

The experimental digital line-store converter samples the incoming 625-line signal at the rate of 11 million samples per second (11MHz). The amplitude of each sample is converted into an eight-digit binary number by the analogue-to-digital converter: this gives 256 different possible levels, i.e. with an eight-digit code each sample may be represented by one of 256 different binary digital numbers. (The ITA's converter samples the input at three times the colour subcarrier frequency, i.e. 13·3MHz.)

The process of interpolation is carried out using a digital store with a capacity corresponding to the information contained in one input line and a high-speed multiplier and adder which accepts two eight-digit numbers (one direct and one from the store), multiplies each by a suitable three-digit number and adds the resulting products (this binary arithmetic can reasonably be regarded as fast as the eight-digit numbers occur at the rate of 11 million per second!).

The time-redistribution process is carried out using an assembly of stores each of which again have a capacity corresponding to one input line. Each wanted line from the interpolator is loaded into one of the redistributing stores at a rate corresponding to 625-line scanning. Shortly after a store has been loaded it is emptied, but this time at a rate corresponding to the 405-line scanning standard. With a suitable number of stores and a suitable switching arrangement the stores are loaded and emptied in sequence so that all the wanted lines are suitably processed in the correct order.

Ancillary circuits provide sync pulses for the 405-line standard, again in digital form. These are added to the digital signals from the time-redistributing stores and the combined digit stream is then converted back to a conventional analogue TV signal.

The BBC says that a considerable amount of work remains to be done before the design of an operational digital line-store standards converter is finalised, but the demonstration we saw was most impressive and an insight into the lines along which TV technology is now developing. Sometime in the future we may see the video signal converted to digital form in the camera before gamma correction and colour signal processing take place and remain in this form right up to the transmitter. The advantage would lie in maintaining the quality of the signal throughout the broadcast network. The effects of interference and circuit drifts would be largely eliminated and equipment line-up and adjustment rendered almost entirely unnecessary.

Another area in which significant advantages are likely to result from digital working is in TV recording. Digital TV recordings would be similar in form to the data recordings now used in computers but would need to have considerably higher transfer rates and storage densities.

Our hats off to the BBC for demonstrating that the technologies of digital storage and calculation are now capable of dealing with broadcast-quality TV signals.

SET NEWS

Chairman of the British Radio Cabinet Manufacturers' Association John Bowes has warned UK radio and TV setmakers not to be complacent in the face of growing imports. He reported that of a total of 150,000 monochrome receivers imported in 1970, 100,000 came from the Far East while in addition 35,000 colour sets were imported from the Continent. Mr Bowes points as a warning to the US market where one colour set in every three is Japanese. These figures are indeed alarming and must wake up UK TV setmakers if their market is not to go the way of the transistor portable radio market.

Hitachi are launching their PAL colour models at their May trade show after which there is to be full-scale distribution. Sample sets to familiarise engineers have been about for a while and one was on show briefly at the Ideal Home Exhibition in March. There are 15, 17 and 19in. models.

Italian **Indesit** monochrome receivers may be introduced on the UK market following a test run earlier this year when 1,000 sets were brought in. The model is a single-standard 24in. one intended to sell at £83·43.

Three new sets have been announced by UK setmakers. From BRC comes the **Ferguson** 12in. single-standard portable Model 3805, called the Courier, at £58·60. This mains-only model is housed in a plastics case, features push-button tuning and has an adjustable loop aerial connected by flying leads. It is fitted with a new BRC chassis, the 1580 hybrid chassis, and weighs 16½lb. From **Dynatron** come two 26in. colour models. The CTV7 Marlborough is a console model with period styling at £410 and the CTV8 Narvic a consolette at £359 with teak finish and £363 with walnut finish.

BASIC CIRCUITS FOR THE CONSTRUCTOR 1

J.W.THOMPSON

THIS is the first in a series of articles describing the construction of the basic circuit blocks which comprise a monochrome television receiver. The circuits are original and unless otherwise stated designed from first principles. With a little ingenuity the more advanced constructor should be able to put together a television set of very high quality tailored to suit his own individual requirements. Fig. 1 shows in block schematic form the entire circuitry. The series should also be of interest to the beginner as a fair amount of general information will be given.

The circuit this month is a transistorised vision i.f. amplifier (Fig. 2) suitable for the British 405-line standard. In some ways transistors are not very well suited to this application; they have low gain compared to valves and a very high internal feedback capacitance (150 millipF for a BF167 transistor compared to 5·5 millipF for an EF184 valve). They are, on the other hand, less noisy, dissipate far less heat and have an indefinitely long life. With careful design their disadvantages can be made relatively unimportant.

Before going any further a summary of the 405-line system might be useful to anyone whose theory needs brushing up. A British 405-line transmitter radiates at v.h.f. a single-sideband vision carrier and a sound carrier of much narrower bandwidth at a frequency 3·5MHz lower than the vision carrier. The tuner unit in a receiver amplifies the aerial signal and converts the frequencies to standard i.f.s of 34·65MHz vision and 38·15MHz sound. The basic frequency response requirements of a vision i.f. amplifier for this system are that it should have a 6dB bandwidth of 3·0MHz and that it should be capable of completely "notching out" the sound carrier to prevent sound bars appearing on the picture. In addition notch rejection of the adjacent lower channel sound (33·15MHz) and adjacent upper channel vision (39·65MHz) may be required. Ample provision for this is made in the circuit.

Circuit Description

As can be seen in Fig. 2 three common-emitter i.f. amplifier transistors are used. The i.f. output from the v.h.f. tuner is coupled via C11 to the input filtering circuitry: L11 provides rejection at 33·15MHz, L12 passes the 38·15MHz sound signal to the sound i.f. amplifier, L13 provides very efficient notch rejection of the sound carrier while L14 may, if necessary, be tuned to 39·65MHz to reject adjacent upper channel vision.

Tr11 has variable gain and couples the vision i.f. signal to the following stage by a conventional double-tuned transformer. Correct impedance match-

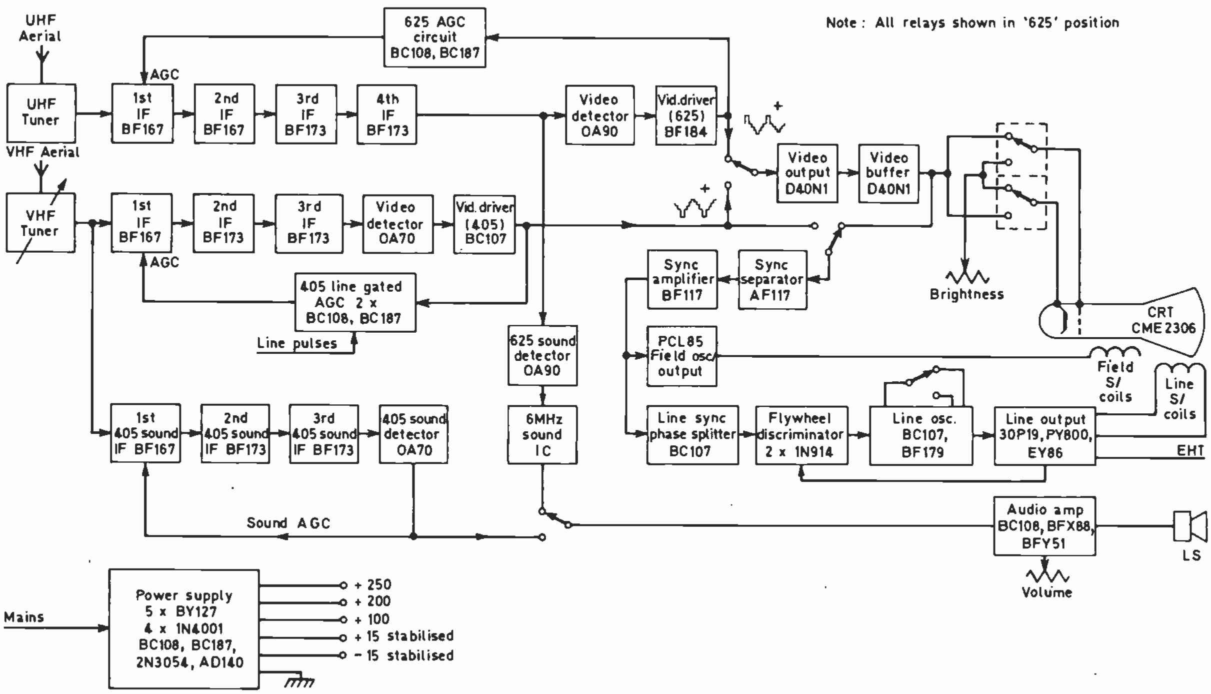


Fig. 1: Simplified block diagram of the complete project.

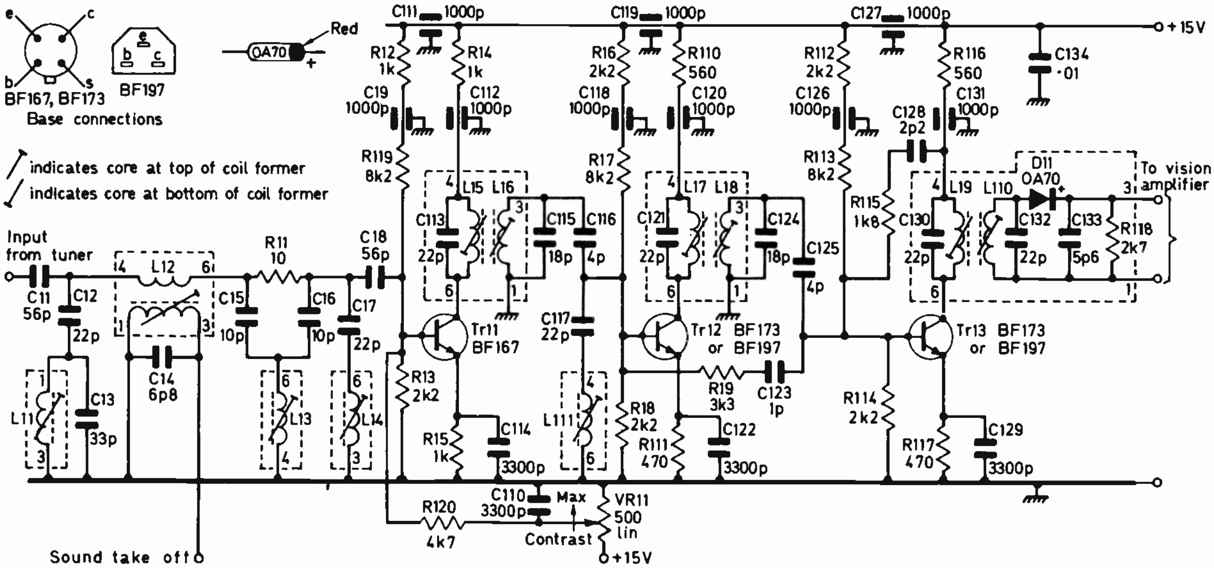


Fig. 2: Circuit diagram of the 405-line standard vision i.f. amplifier.

ing into Tr12 is achieved by the combination of C116 and the input capacitance *C*be (about 45pF) of Tr12. Additional sound rejection is provided by L111. Tr12 is effectively unilateralised (i.e. the capacitative and resistive components of internal feedback balanced out) by the series combination of R19 and C123, ensuring complete stability. Tr13 is the i.f. output transistor and is also unilateralised but by a slightly different method (R115, C128). The final double-tuned transformer (L19, L110) couples the signal to the detector D11. A detected vision signal of at least 3V peak-to-peak is present between pins 1 and 3 of the final i.f. transformer.

A brief word concerning the numbering of components: the first digit signifies the number of the circuit in which the component appears. R118 for example is in circuit 1 while R318 is in circuit 3. This will simplify cross-references between subsequent circuits in the series.

Construction

The overall size of the metal chassis used will depend on the number of circuits to be built on it and the reader would be well advised to wait until the end of the series before making this decision.

★ components list

Resistors:

R11	10Ω	R14	1kΩ	R17	8·2kΩ	R110	560Ω	R113	8·2kΩ	R116	560Ω	R119	8·2kΩ
R12	1kΩ	R15	1kΩ	R18	2·2kΩ	R111	470Ω	R114	2·2kΩ	R117	470Ω	R120	4·7kΩ
R13	2·2kΩ	R16	2·2kΩ	R19	3·3kΩ	R112	2·2kΩ	R115	1·8kΩ	R118	2·7kΩ	VR11	500Ω lin.

Capacitors:

C11	56pF 5% SM	C111	1,000pF F	C121	22pF 5% SM*	C131	1,000pF F
C12	22pF 5% SM	C112	1,000pF F	C122	3,300pF 160V 10% P	C132	22pF 5% SM*
C13	33pF 5% SM	C113	22pF 5% SM*	C123	1pF 5% SM	C133	5·6pF 5% SM*
C14	6·8pF 5% SM	C114	3,300pF 160V 10% P	C124	18pF 5% SM	C134	0·01μF 160V 10% P
C15	10pF 5% SM	C115	18pF 5% SM	C125	4pF (3·9pF) C		
C16	10pF 5% SM	C116	4pF (3·9pF) C	C126	1,000pF F		
C17	22pF 5% SM	C117	22pF 5% SM	C127	1,000pF F		
C18	56pF 5% SM	C118	1,000pF F	C128	2·2pF 5% SM		
C19	1,000pF F	C119	1,000pF F	C129	3,300pF 160V 10% P		
C110	3,300pF P	C120	1,000pF F	C130	22pF 5% SM*		

C ceramic; F feedthrough; P, polyester; SM, silver-mica; * Side-mounting printed-circuit type.

Semiconductors:

Tr11 BF167 Tr12 BF173 or BF197 Tr13 BF173 or BF197 D11 OA70

Miscellaneous:

9 Aladdin coil formers, $1\frac{7}{8} \times \frac{1}{4}$ in. (CR14)
8 screening cans (CR15)
11 Hexagonal dust cores, 6×12.7 mm. (Z81A)
Tube non-hardening core locking compound (ZA39A)
Nylon leadthrough tag (Z149)
Trimming tool (TL33A)
Numbers in brackets are Home Radio catalogue numbers.

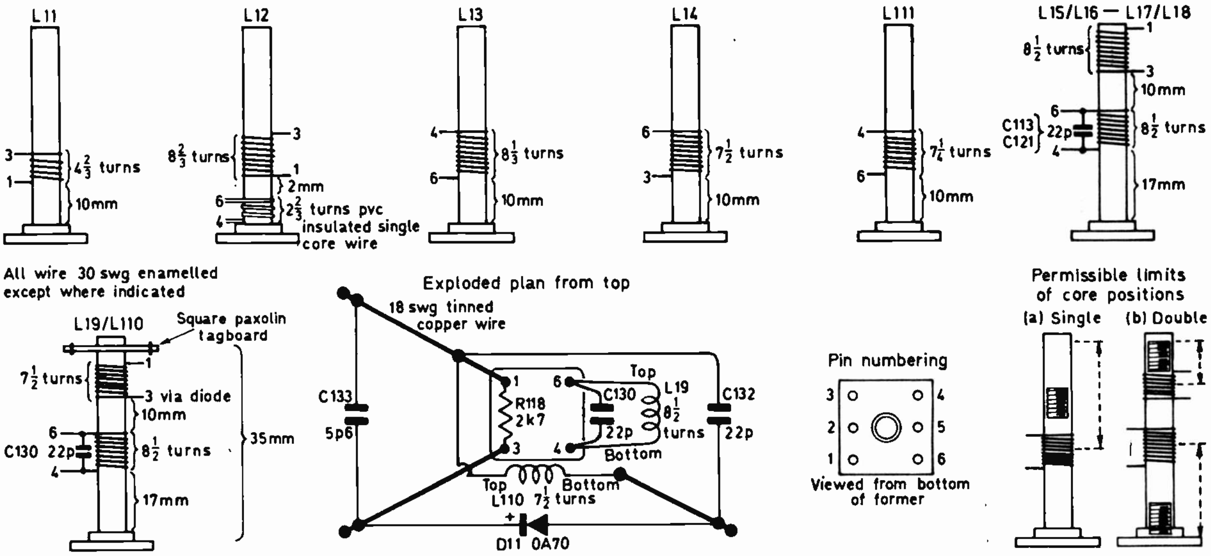


Fig. 3: Coil winding details. Coils wound clockwise looking from top of former.

The metal must be easy to solder: brass or mild steel would be suitable, of gauge 16 or 18 s.w.g. A layout plan for circuit 1 is shown in Fig. 4 and this must be followed closely for optimum stability. Two soldering irons will be required, one of about 15W power and the other about 50W for heavy soldering to the chassis.

Winding the Coils

The most difficult part of all is winding the coils. The windings must be evenly wound and secured to the former with wax or polystyrene cement. Wind the coil, spread liberally with the selected fixing compound and hold patiently until the latter has set. Strip the enamel from the ends of the wire and solder into the pins at the base of the former as detailed in Fig. 3. Any additional tuning capacitance (e.g. C113) must be soldered with short leads across the relevant pins. The dotted lines surrounding each coil in the circuit diagram indicate exactly which components (if any) should be mounted inside the coil screening can.

The i.f. output transformer (L19, L110) is a special case and requires explanation. Many readers will have stripped down old television i.f. amplifiers and will probably recall seeing i.f. transformers with a square paxolin tagboard fitted towards the top of the coil former, often with four heavy gauge wires leading down to the base. The principle here is the same except that there are only three heavy gauge wires. These go from pin 1 to the tagboard, from pin 3 to the tagboard and from the tagboard to half way down the coil former immediately above pin 4. Obtain a suitable square tagboard from an old i.f. amplifier or make one if necessary, and wire up as shown approximately in Fig. 3. *Do not damage the diode by overheating when soldering.* There should be plenty of room for all the components. As a final precaution against short-circuits to the can, wrap the entire coil assembly in waxed paper.

The coil dust cores have a hexagonal centre hole and will require a special trimming tool. To minimise tuning drift they should all be coated in core locking compound which is obtainable from Home Radio Ltd.

Returning to the chassis, $\frac{7}{32}$in. holes should be drilled for pins 1, 3, 4 and 6 of the coils and two $\frac{1}{8}$in. holes for the 6BA coil mounting bolts. The core trimming hole corresponding to the centre of each coil former is optional as both cores, being hexagonal, can be adjusted from the top of the coil.

Feedthrough Capacitors

Considerable use was made in the prototype of ceramic leadthrough capacitors (e.g. C19) and nylon leadthrough tags. One of the latter is shown in Fig. 4 at the tuner input point. Any leadthrough tag may be used (nylon, p.t.f.e. or ceramic) but it must not possess a self-capacitance to chassis greater than 5pF. The transistors may be put in after all major soldered connections to the chassis have been made and provided reasonable care is taken it is unlikely that they will be damaged by heat. Further information on fitting feedthrough capacitors was given on page 16 of the October, 1970, issue.

Testing and Alignment

When the wiring has been completed it is advisable before connecting a power supply to measure the resistance between the +15 volt line and the chassis. If there is a dead short check the leadthrough capacitors very carefully. Otherwise it is reasonably safe to connect a power supply which, for test purposes, may be a 9V and a 6V battery in series. Measure the voltages from each transistor emitter to chassis: any voltage between 1·5 and 3·0V is satisfactory. Each transistor base voltage should be about 0·6V more positive than the emitter voltage and if it is very different from this the offending transistor should be replaced.

If these measurements are satisfactory disconnect the power supply and temporarily solder pin 1 of the i.f. output transformer to chassis. Connect a voltmeter on its 0·5V range (or a 10,000Ω/voltmeter on the 0·1mA range) between pin 3 of the i.f. output transformer and chassis. In addition to the meter a 3,300pF capacitor should be temporarily connected between pins 1 and 3: do not forget to remove this

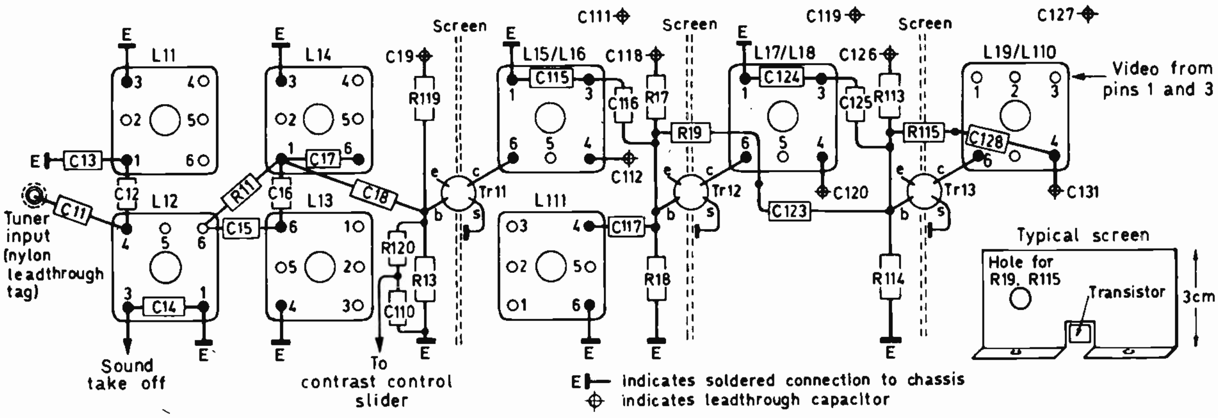


Fig. 4: Recommended layout (underside). The emitter resistors and capacitors have been omitted for clarity. R12, R14, R16, R110, R112 and R116 are mounted on the topside. Connect C111, C119 and C127 together to form the positive l.t. rail and connect C134 from this to chassis.

capacitor when the alignment is complete.

The contrast control should initially be set for maximum gain then during the final stages readjusted so that the overall i.f. gain is reduced by a factor of 10 (i.e.−20dB).

On reconnecting the power supply the meter should not move appreciably: if the needle hits the end-stop violently the amplifier is unstable. In this event detune some of the transformer cores slightly until the meter returns to zero. If this fails it is likely that the wiring is defective in some way. This should not occur and indeed it was impossible to make the prototype amplifier oscillate with any combination of coil settings.

If all has gone well so far it is time to set up the signal generator which need not be particularly elaborate: a Nombrex Model 31 will be more than adequate. Set the cores of all single-tuned coils to the top of each coil can and the upper and lower cores of double-tuned coils to top and bottom of each coil can respectively. Inject a strong 36·0MHz signal at the base of Tr13 and peak L19 and L110 for maximum meter indication. Reduce the signal generator output level as necessary to prevent the meter from going off the scale. Move the generator to the base of Tr12 and peak L17 and L18. Finally inject the signal at the tuner input point and peak L15 and L16 as nearly as possible.

Disconnect the generator and make up a loosely coupled input coil on a standard Aladdin former as follows: secondary 2 turns of single-core p.v.c. insulated wire connected with short leads between the tuner input point and chassis; primary 4 turns of the same wire connected to the signal generator via a short coaxial lead and free to slide along the coil former towards the secondary winding. Switch on the signal generator again and move the sliding winding along the former until the meter shows at least ¼ full-scale deflection. Make sure that all the i.f. transformers are correctly peaked at 36MHz and then slightly detune some of the cores to spread the response. Check that the response *has* been spread by altering the signal generator frequency.

The tuning should now be spread until the meter shows at least some indication between 33 and 40MHz but there is no need yet to aim for a flat response. Tune L11 for minimum output at 33·15MHz and L13 for minimum at 38·15MHz. Slight adjustment of the number of turns on these coils may be necessary in order to obtain the correct rejection point. In areas of very strong signal it will be necessary to tune L111 for minimum output at 38·15MHz and in areas where an appreciable signal exists on the adjacent upper channel L14 should be tuned for minimum output at 39·65MHz. Under normal reception conditions leave the cores of these two coils at the tops of their cans or remove them completely.

The cores of the double-tuned i.f. transformers should now be tuned to obtain the response shown in Fig. 5. The points marked −6dB at 34·65 and 37·65MHz are those at which the output voltage should be exactly half its value along the flat top of the response curve. Try not to be discouraged if this procedure takes several hours. With a wobbulator and oscilloscope it would take a matter of a few minutes but it is assumed that these instruments will not be available in the average reader's workshop. The top of the response curve should be as flat as possible or ringing may occur on the picture at certain vision frequencies. When a v.h.f. tuner is ultimately connected to the i.f. amplifier the mixer output coil on the tuner should be adjusted for best picture definition on a test card consistent with good sound.

Circuit 2 next month will deal with a British 625-line i.f. amplifier.

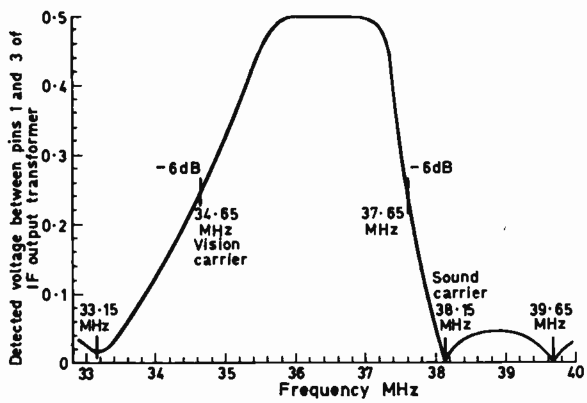


Fig. 5: Response curve, 405-line vision i.f. strip.

TO BE CONTINUED

Colour Receiver APC LOOPS

S. GEORGE

THE detected vision or luminance signal is either positive- or negative-going depending on which way round the detector diode is connected. At the c.r.t. cathode the signal is negative-going, driving the tube from cut-off—black level—towards peak white. With the chroma signal the position is different. The colour-difference signals produced by demodulating the chroma signal are required to supplement the luminance signal, increasing or decreasing as required by the picture being transmitted the red, green and blue beams of the shadowmask tube. Thus the chroma detectors have to produce outputs varying not only in amplitude but in polarity as well. Clearly simple diode envelope detectors will not suffice: the detectors used must be able to give either a negative- or positive-going output and for this reason synchronous detectors are used.

Synchronising the Reference Signal

Synchronous detectors require not only a signal input but a reference subcarrier input as well and for this reason a local reference oscillator or subcarrier regenerator (suppressed subcarrier transmission being used for the chroma signal) is required in a colour receiver decoder. To obtain correct colour signals the local reference oscillator must be synchronised to the transmitter subcarrier oscillator. This is a vital feature in a colour receiver and for this reason the reference oscillator is incorporated in an a.p.c. (automatic phase control) loop under the control of the transmitted colour sync signal—the 10 cycles burst of subcarrier transmitted during the back porch of the line sync pulse. One chassis—the Rank-Bush-Murphy single-standard one—uses a somewhat different approach, making use of the burst signal as the local reference signal by feeding it to a high-Q narrow-band crystal filter circuit which rings at the subcarrier frequency throughout each line. However, to return to the more usual arrangement.

The reference signal fed to the synchronous detectors must be identical in phase and frequency with the suppressed subcarrier and while the crystal oscillators used stay right on frequency nevertheless even the slightest tendency towards phase shift must be prevented. In this article we shall examine a typical complete a.p.c. loop of the type widely used to attain this essential phase control, taking as our example the circuit used in the Pye single-standard chassis (the 691 chassis).

Before proceeding further however it may be as well to clear up two important points that are apt to be overlooked about the colour bursts. First, to ensure correct R – Y switching in a PAL receiver the colour burst alternates 45° each side of the – B – Y phase position on successive lines, raising the question how does the reference oscillator lock to this line-by-line switched signal. The answer is that the a.p.c. loop locks to the average phase of this alternating signal, i.e. to – B – Y, ignoring the rapidly alternating ± R – Y component of the burst signal. Secondly the phase of the oscillator is actually in quadrature with the colour burst, its lag or lead depending on the immediate circuitry (whichever it is can of course be reversed simply by transposing coil or transformer connections). This is because the reference oscillator is arranged to be in phase with the R – Y signal. The reference signal then passes through a 90° phase shifter before being fed to the B – Y synchronous detector (it will be recalled that the R – Y and B – Y components of the chroma signal are 90° out of phase with each other).

Burst Channel

To return however to the Pye a.p.c. circuit shown in Fig. 1. The first stage shown, Tr1, is not part of the a.p.c. loop itself but is included so as to take in the burst channel which feeds the a.p.c. discriminator circuit. Tr1 is a conventional common-emitter amplifier stage which is fed with the chroma signal from the collector circuit of the first chroma amplifier. R1, R2 provide forward bias and the chroma signal, centred at 4·43MHz, appears amplified across L1 in its collector circuit. To minimise the input capacitance of the stage the emitter circuit is only partially decoupled.

Tr2, the burst gate, can be regarded as the first stage in the a.p.c. loop. There is no fixed forward bias to this stage which is switched on only momentarily during the colour burst period by positive-going pulses from transformer T1. These pulses are actually overswings instigated by the negative-going line sync pulses fed from the sync separator to this transformer. Whilst the sync pulses themselves represent an almost ideal, regularly occurring gating pulse they end of course just before the actual colour burst. Hence the use of the ringing transformer T1 which produces positive-going overswings—the correct polarity to bias Tr2 on—of sufficient duration to comfortably span the 2·5μsec burst period. In this way only pure burst signal is developed across T2 for feeding to the a.p.c. loop discriminator diodes D1 and D2 and the a.c.c. rectifier D3.

Automatic Chrominance Control

To digress slightly, the a.c.c. rectifier provides a negative-going output to yield an automatic chroma control reverse bias potential related to the strength of

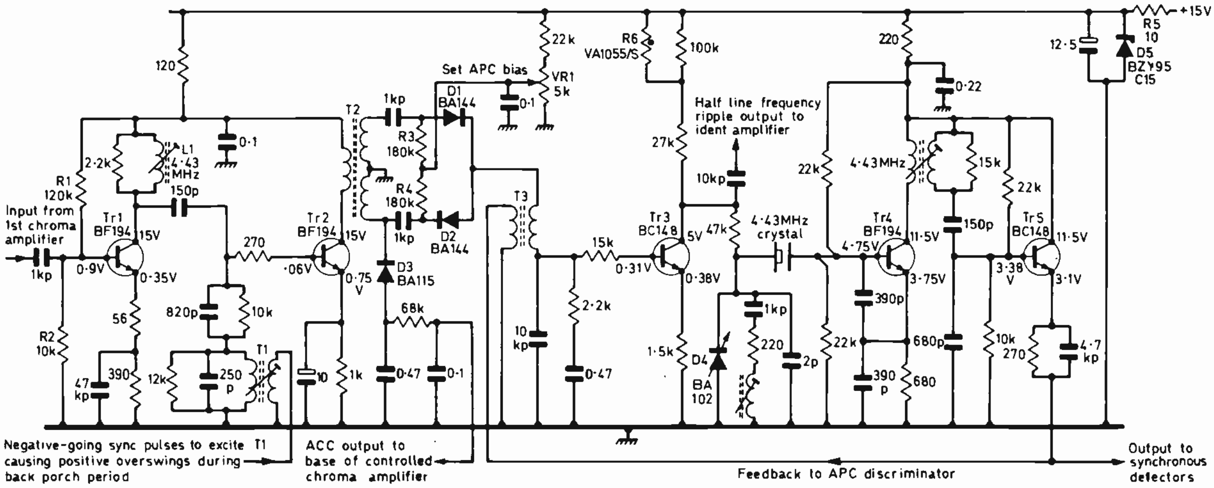


Fig. 1: The decoder reference oscillator Tr4 and its a.p.c. loop in the Pye 691 single-standard chassis. The conduction of the d.c. amplifier Tr3 controls the reverse bias across the varicap diode D4 and hence the frequency and phase of the reference oscillator. The voltages above appear to show Tr3 cut off: this however is an illusion due to the input impedance of the stage, the swinging burst waveform present and the particular meter used for the readings given (Avo Model 8, 20,000 Ω/volt). With a Taylor Model 100A Tr3 voltage readings are 4·8V collector, 0·8V base and 0·34V emitter.

the burst signal—an accurate measure of chroma signal strength since of course the burst signal is not modulated—to control the gain of the controlled stage in the chroma channel.

APC Action

The a.p.c. discriminator circuit operates in a similar manner to flywheel line sync discriminator circuits in which the degree of coincidence between the sync pulses and a reference signal from the line timebase determines the negative or positive control potential produced. In this application however the d.c. output produced by the discriminator is determined by the degree of phase coincidence between the colour burst signal and the reference signal taken from the emitter-follower stage Tr5 which acts as a buffer stage in the crystal oscillator output circuit.

The basic action is that when D1 and D2 conduct equally there is zero output from the discriminator: if D1 conducts more than D2 a positive-going control potential is obtained while if D2 conducts more than D1 the control potential obtained is negative-going. Thus when the burst and the reference oscillator signals are synchronised we would expect zero or negligible output, becoming positive or negative as the phase of the oscillator signal tends to drift in one direction or the other from the colour burst signal. In this circuit however the junction of the discriminator load resistors R3 and R4 is returned to the slider of the "set a.p.c." bias potentiometer VR1 so that the discriminator output voltage is superimposed on that tapped from the this control. Thus depending on the relative phasing of the two 4·43MHz oscillations—the burst and local reference oscillator signals—the discriminator output will increase or decrease the net positive potential applied through T3 to the base of the d.c. amplifier Tr3. Variations in this bias potential due to phase disparities vary Tr3 collector current and the resulting changes in its collector voltage increase or decrease the reverse bias applied across the varicap diode D4 in the reference oscillator (Tr4) tuned circuit.

Thus the varicap diode is the element which applies the phase control to the oscillator circuit. The capacitance of a varicap diode is determined by the reverse bias across it. As this is increased, increasing the junction depletion region of the diode, so its capacitance decreases and vice versa. We have thus now covered the overall operation of the a.p.c. loop, but there are several details worth looking at.

Miscellaneous Features

First the use of an emitter-follower following the reference oscillator provides considerable current gain and a low-impedance source for the output feed to the synchronous detectors while imposing negligible loading on the oscillator. Secondly to stabilise the operation of the circuit generally the l.t. rail is taken from the junction of R5 and the zener diode D5 while a thermistor R6 is shunted across one of the resistors in the collector load of the d.c. amplifier Tr3 to stabilise the operation of this stage against the effects of change in temperature. A voltage change at the collector of this transistor due to any cause will change the capacitance of the varicap diode and thus shift the phase of the oscillator.

As the R – Y component of the burst signal swings 45° each side of – B – Y on successive lines there will even after filtering be a slight ripple at half line frequency in the a.p.c. loop. This a.c. signal is ignored by the loop which responds only to the d.c. component of the signal, but the ripple is used by the ident amplifier in which a high-*Q* tuned circuit is excited to provide a half line frequency signal which controls the operation of the PAL switch circuit, synchronising it with the alternate line R – Y phase alternation at the transmitter.

The ident signal also serves a further purpose: as it arises only when a colour transmission with its burst signal is being received the ident signal is rectified and used as a turn-on bias for one of the chroma amplifiers. In this way the colour-killer action is

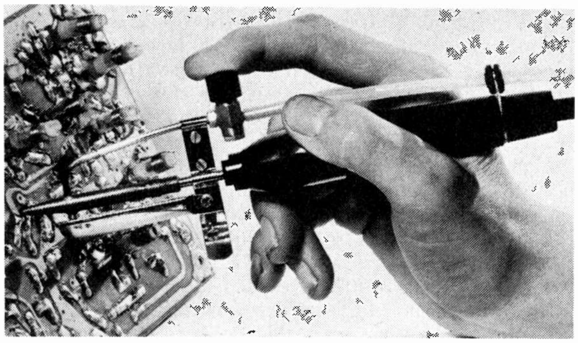
—continued on page 365

ANTEX DESOLDERING IRONS

THERE are seven desoldering instruments manufactured by Antex Ltd.: four 25W (ESS) types for 24, 115, 220 and 240V respectively and three 18W (GSS) types for 115, 220 and 240V. The GSS types are miniature models for very small work such as hearing aids and subminiature assemblies while the ESS types are for normal workshop use. The subject of this report is the ESS 240V model.

The instrument is similar in appearance to the normal Antex soldering irons but with additional equipment for performing the desoldering operation. A nylon air tube is supported along the length of the handle by a bracket at the front and a rubber grommet at the rear with a control-valve assembly at the front. This valve consists of a finger push-button mounted so that it can be operated by the forefinger. The bracket also supports the solder trough and acts as a stand for the iron. At the front the air way curves downward through the hollow bit and then out the opposite side to a hole in the lid of the solder trough. The nylon air tube is about 6ft. long—as is the three-core mains cable—and is supplied complete with a set of adaptors to enable the tube to be connected to a workshop air line. If this is not available—and few workshops are so equipped—a footpump can be supplied to provide the air pressure.

Unlike most desoldering instruments this one operates by pressure rather than suction, the force of the air pressure across the end of the hollow bit producing suction by the venturi principle. This obviates the possibility of foreign matter being sucked into the airways and valve and thus eliminates the need for the elaborate filtering arrangements used to prevent this on suction instruments.



The Antex ESS desoldering iron in use.

The bit gently tapers to a 4mm. diameter circular flat section in the centre of which is a 2mm. diameter air vent. The choice of size for the air vent is not an easy one for manufacturers of desoldering irons: if it is too large the suction is reduced and effective removal of solder inhibited while if it is too small it will not fit over terminal posts and tags in the printed circuit. The dimensions chosen by Antex for their iron seem just about right for the majority of printed circuit work.

The solder trough needs occasional emptying. This can be done quite easily as both the trough and its cover are hinged by a single screw on the main assembly support. It needs only slight pressure to swing them both downwards so that the cover can be pulled up and the solder emptied. To prevent solder adhering to the stainless steel trough the makers recommend smearing a little oil or grease over the inside.

Use

The instrument sits very well in the hand and because of its lightness causes no fatigue during long periods of use. The best position to use it (see photo) is with the air line passing over the top of the handle, the instrument being held between the thumb and the side of the second finger with the first finger on the control valve. The printed board can be in almost any position but the easiest working position is vertical or nearly so. The lightness of the tool is almost a disadvantage as when the iron is stood on the workbench there is a tendency for the weight of the air line and mains cable to pull it off. The most stable parking position is right at the edge of the bench where the instrument can make good three-point contact by standing on the two legs of the supporting bracket and the rear of the handle, the air line and mains cable hanging directly over the edge.

When used with a workshop air line the iron is effective and simple to use. When used with the footpump however it is not quite so easy—as in fact is the case with all irons that use a pump. The pump must be positioned in just the right place, preferably secured to the floor in some way. When using the iron one's eyes and attention are obviously on the work and one's foot has to search for and operate the pump by feel. It is easy to push the pump over on its side if it is not secure, especially as quite a vigorous push is needed to build up enough pressure to create the necessary suction in the bit.

Various combinations of control valve and footpump operation were tried. Best results were obtained by operating the footpump with the valve still in the closed position to enable pressure to be built up in the tube and airways. Suddenly opening the valve while continuing the footstroke then gives a violent rush of air and good suction at the bit. Workshops that do a fair amount of desoldering but are not equipped with an air line might find a compressed air bottle that can be recharged at the local garage a good alternative to the pump.

One noteworthy feature of the iron is its quick warming-up time. It will melt solder in a minute and a half and within two minutes is ready for desoldering. The price of all the models is £4·67 or £5·65 complete with pump.

NEXT: PICARD DESOLDERING PUMP

FREQUENCY METER

MARTIN L. MICHAELIS, M.A.

PART 2

THE input stage Tr1 of the amplifier section is a voltage amplifier with a number of safety features. The gain control VR1 gives adequate response for frequencies up to well above the design limit of 100kHz. R1 limits the charge surge current into C1 for input voltages grossly in excess of the set trigger level or for large d.c. components accompanying small a.c. components of an input signal. Any voltage up to at least 100V sum of d.c. and peak a.c. components may be connected to P1 with any settings of the controls without damage. D1 and D2 prevent excessive voltages reaching Tr1 while R5 prevents instability.

Tr2 phase splitter stage is d.c.-coupled to Tr1 collector and provides antiphase amplified replicas of the input waveform across its equal collector and emitter resistors. The oscilloscope signal output is taken from the emitter via preset potential divider VR3 to the inverting input of the operational amplifier IC1. As we have two phase-inverting stages between P1 and P3 the oscilloscope signal has the same polarity as the input signal. IC1 is an MC1709CG used in a circuit operating in the same manner as described in the i.c. preamplifier article (November, 1970, and January, 1971). Its gain is determined by the ratio of R46 to R44. VR3 is adjusted during setting up so that the gain from P1 to P3 is exactly 20. This figure provides the same deflection sensitivity of the Videoscope MV3 oscilloscope when P3 is connected directly to its Y input and the signal probe to P1 as when the signal probe is connected directly to the Y input of this oscilloscope. This is true for any other oscilloscope with a standard 1MΩ, 40pF Y input and standard 10MΩ, 4pF signal probe. For signals fed directly to P1 the path P1 to P3 behaves as a straightforward preamplifier with a gain of 20.

The only new feature of the operational amplifier circuit IC1 compared to the circuits described in the i.c. preamplifier article is the use of R44 and R45 to feed two signals to $\overline{IP}$. This is the algebraic adding function of the operational amplifier. Due to the heavy negative feedback from OP to $\overline{IP}$ the latter point behaves as a virtual earth and any number of input signals may be connected to it via respective gain-determining series resistors without any mutual interference. Thus the output waveform of this arrangement is the linear algebraic sum of all the input waveforms. This sum consists of the added waveforms weighted by the gain factors determined by the respective series resistors. The second waveform injected via R45 consists of the final output pulses of the amplifier chain. By virtue of the linear algebraic addition these simply sit on the signal waveform without in any way otherwise affecting or distorting the latter. They mark the exact features or time points of the signal waveform at which the pulse univibrator fires.

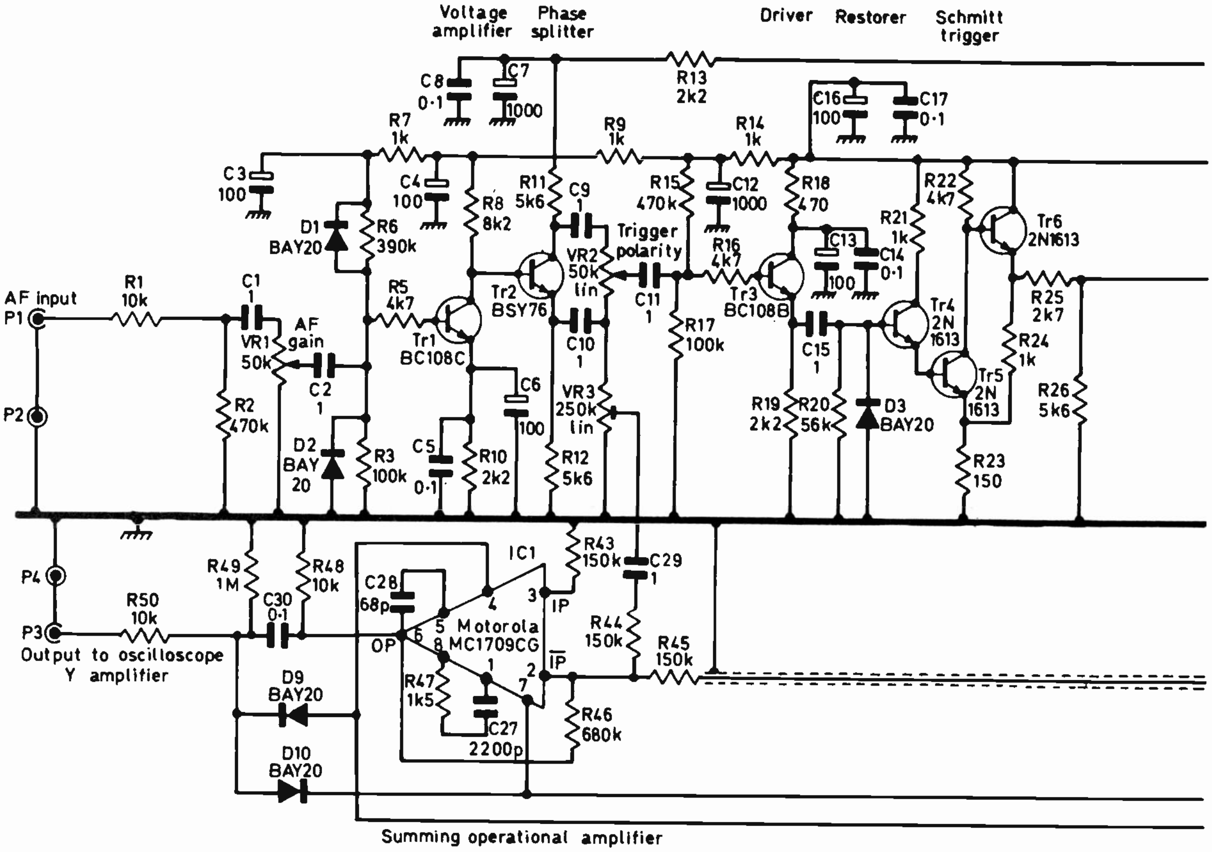
Pulsing Circuits

The purpose of the pulsing circuits Tr3 to Tr6 is to produce an accurate squarewave with steep flanks irrespective of the frequency or waveform of the input signal. Tr3 is an emitter-follower which reproduces the waveform taken from VR2 slider with much lower impedance at its emitter so that C15 is rapidly charged to the peak negative amplitude. Consequently the entire peak-to-peak amplitude of the waveform appears with positive polarity across R20. This is the d.c. restorer function of D3 in conjunction with C15. The track of VR2 is connected via blocking capacitors between the collector and emitter of the phase splitter Tr2. This is in effect a bridge circuit which is balanced when VR2 slider is set to mid-track. There is then no signal present at the slider. Opposite polarity signals are obtained on opposite sides of the track centre, with increasing amplitudes as the ends of the track are approached.

Tr4 to Tr6 form a Schmitt trigger circuit which must always take up one of two stable states. In the resting state with no input signal Tr4 and Tr5 are cut off because there is then no base bias to make them cut on. Tr6 however is conducting to saturation, being turned on hard by R22. The voltage drop across Tr6 is negligible in this state so that the entire supply voltage is dropped across R23 and R24 in series. The portion dropped across R23 plus the two silicon thresholds of about 0·6V each for Tr4 and Tr5 is the minimum positive voltage which must be applied to Tr4 base to make Tr4 and Tr5 cut on. When this threshold voltage is exceeded Tr5 commences to draw current through R22. The voltage at Tr6 emitter and consequently the voltage drop across R23 then suddenly decrease, Tr4 and Tr5 suddenly cutting on even harder. The effect is cumulative so that in fact the circuit trips over extremely rapidly to its other stable state, with Tr4 and Tr5 conducting and Tr6 cut right off. This state persists as long as the signal waveform excursion at Tr4 base remains above the trip threshold. By converse action the circuit trips back to its resting state as soon as the signal waveform at Tr4 base drops below the threshold again.

It is quite immaterial how rapidly or how slowly the signal waveform crosses the threshold or at what frequency. The Schmitt trigger circuit trips over in the same extremely rapid manner in every case. The output at Tr6 emitter is a virtually perfect squarewave: in the resting state with Tr6 cut on its emitter rests at nearly the collector supply voltage level while in the tripped state with Tr6 cut off its emitter is only about 1V above chassis potential.

By virtue of the positive d.c. restorer function of



D3 and C15 the tripped state persists for a certain phase angle around the positive peak of the signal waveform at Tr3 emitter, and this may be chosen to be the positive *or the negative* peak region of the input waveform by moving VR2 to the appropriate side of the track centre position.

Since the bridge is balanced and there is no signal at all with VR2 slider in its centre setting it is necessary to move the slider a certain critical distance to either side in order to obtain a signal amplitude sufficient to fire the Schmitt trigger at all. In the critical setting the Schmitt trigger produces extremely brief square pulses because it only briefly trips coincidentally with the positive or negative peak of the signal waveform. If VR2 is turned farther towards the respective end of the track the Schmitt trigger trips progressively lower down the signal waveform and the square pulses it produces broaden symmetrically about the respective peaks of the input waveform. In this manner it is possible to choose a trigger point anywhere within the positive and negative waveform peak regions, giving accurate conversion to a steady squarewave frequency even with complex waveforms such as video signals with sync pulses in which the video content may be continuously changing. This is a conventional sync separator function but is more versatile in applying to all manner of other waveforms too.

Q-pulse Generator Circuits

Thus negative-going square pulses are produced at Tr6 emitter. The flanks of these pulses are always very sharp and independent of the signal input waveform but the *duration* of the pulses depends upon the trigger level setting and the input signal frequency. For proper operation of the frequency counters we require pulses of uniform shape and duration, the latter short enough to suit the highest designed for operating frequency. A repetition frequency of 100kHz implies a pulse width of 10μsec with zero gaps. The actual width has been chosen to be about 6μsec.

These pulses are produced by the pulse univibrator Tr8, Tr9 which is triggered via Tr7 from the output of the Schmitt trigger. The positive-going flanks of the square pulse from Tr6 emitter are passed via C18 and the diodes D4 and D6 to Tr7 base and thus appear at low impedance across VR5 to trigger the pulse univibrator. The negative flanks are blocked by D4, any fraction nevertheless leaking past D4 by stray capacitance being shorted out by D5. The potential tapped from VR4 is used to cancel the silicon thresholds of the diode network while VR5 sets the optimum trigger sensitivity of Tr8 and Tr9 for stable operation at all intended frequencies. Note that it is the positive-going flank of the negative-going square pulse at Tr6 emitter which fires the pulse univibrator Tr8, Tr9 so that the counting pulse is produced not when the Schmitt trigger trips but *when it drops back to the resting state.* This means that the counting pulses which are displayed on the input waveform by the adding action of IC1 always appear *after the signal waveform peak,* when the waveform drops below the threshold level on its return to the zero line.

Tr8 and Tr9 form another type of threshold trigger

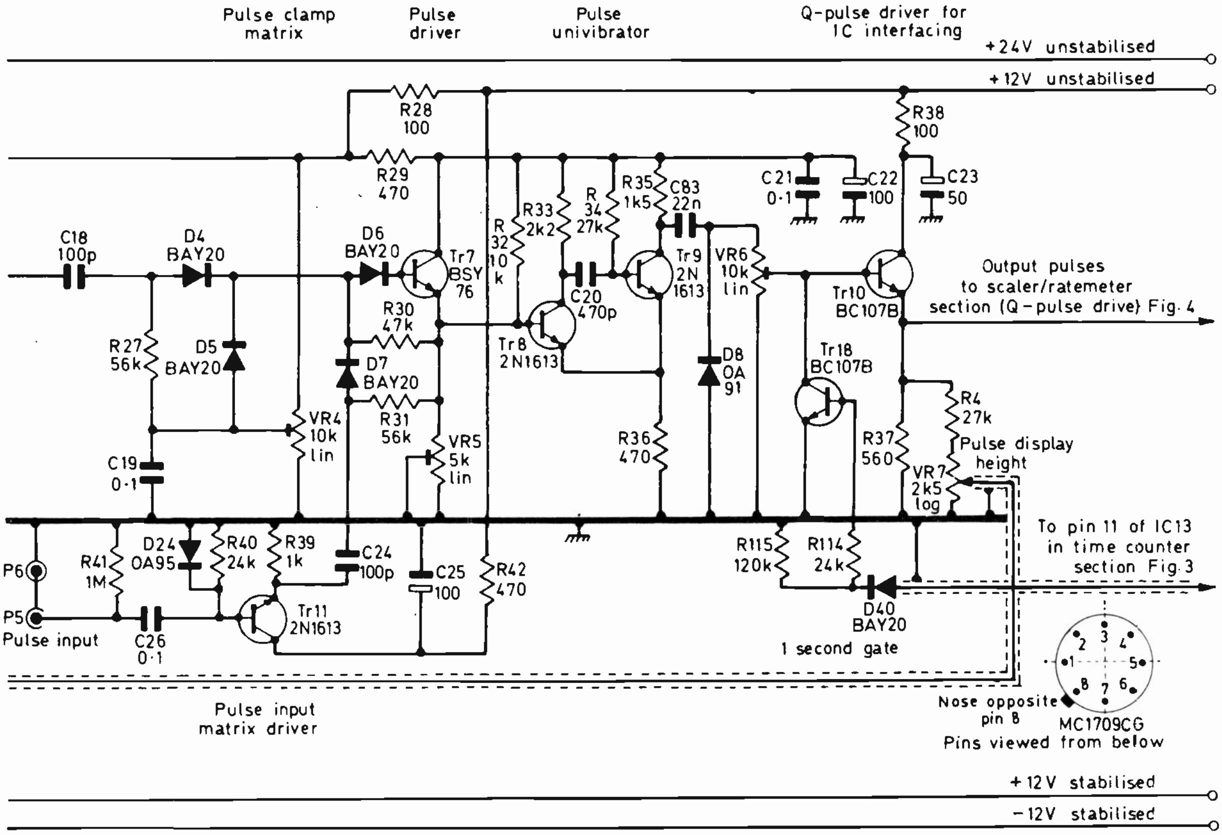


Fig. 2: Circuit diagram of the amplifier section of the frequency meter.

circuit with two stable states. Normally Tr8 is resting cut off because VR5 is adjusted so that the base bias voltage is just insufficient to cut it on. A slight positive rise across VR5 injected from Tr7 suffices to cut Tr8 on. A cumulative action then takes place, cutting off Tr9 and cutting Tr8 hard on by virtue of the multivibrator coupling C20, R34 and the common emitter resistor R36. When C20 has discharged via R34, i.e. after a definite time determined by the values of C20 and R34, the circuit suddenly reverts to the stable state with Tr8 cut off again and Tr9 conducting hard. This time interval has been made about 6μsec. Since Tr9 is cut off for this brief period and is conducting hard at all other times a positive pulse of about 6μsec duration appears at Tr9 collector.

C83 and D8 form another positive d.c. restorer, to make this pulse train entirely positive-going from the zero line irrespective of frequency. Otherwise the base line would gradually sag below the zero line as frequency increased, leading to very erratic triggering of the frequency counter section at higher frequencies.

The correct pulse amplitude for operating the frequency counter section is established by VR6 during setting up. Tr10 is simply an emitter-follower to provide the amplitude-adjusted output pulses at very low source impedance. The full amplitude is fed straight from Tr10 emitter to the frequency counting section and a suitable fraction is applied to the track of the marker amplitude potentiometer VR7. Since the positive pulses here are fed to the inverting input $\overline{\text{IP}}$ of IC1 the marker pulses appear negative at the oscilloscope output P3. Thus the counting trigger points of the signal waveform appear as very narrow downward spikes on the oscillogram. Their amplitude is independently adjustable — to vanishing point if desired — with VR7 on the front panel.

One-second Gate

Tr18 is normally cut off so that in the continuous counting mode it has no effect. In the intermittent counting mode a symmetrical positive-going square-wave with 2sec. repetition period is fed from the time counter section to the anode of D40. This cuts Tr18 on for alternate seconds. Pulse counting stops during these seconds because Tr18 then short-circuits the amplifier output pulses to chassis at the slider of VR6.

Pulse Input Stage

Waveforms which are already sharply pulsed may be fed to the pulse input socket P5. D24 and C26 form a positive d.c. restorer so that the waveform becomes entirely positive-going at the base of the emitter-follower Tr11 which effects impedance step-down. By virtue of the d.c. restorer action positive or negative pulsed waveforms cut Tr11 on equally well. C24 and R31 differentiate the pulse waveform, producing a series of alternate positive and negative spikes. Only the positive spikes pass through D7 and

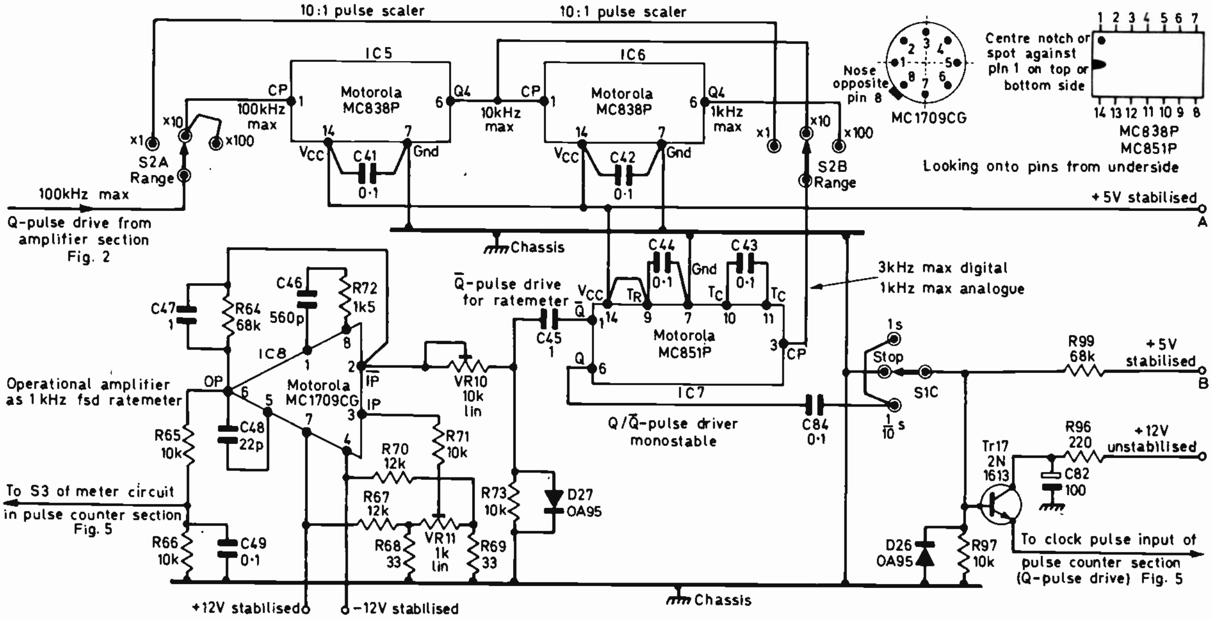


Fig. 4: Circuit diagram of the pulse scaler and ratemeter sections.

any portion of the negative spikes which pass D7 via stray capacitance is again shorted out to chassis via D5 and C19. The positive spikes on the other hand proceed via D6 and Tr7 to trigger the pulse univibrator in the same way as positive spikes arriving from the a.f. amplifier and pulser section via D4.

The trigger threshold at P5 is nominally 5V pulse amplitude. The actual trigger level measured was found to be 3·5V pulse amplitude in the prototype. This is determined by the silicon thresholds of Tr11, D7, D6 and Tr7 in series and the setting of VR5. It is quite sharp, so the pulse input may be used for pulse amplitude discriminator functions. If greater sensitivity is required the i.c. preamplifier described in a previous article may be connected between the signal source and the pulse input P5. The circuitry around Tr11 has been designed to suit the i.c. preamplifier so that correct functioning is obtained in all gain settings of the i.c. preamplifier. The maximum i.c. preamplifier gain setting is 200 making the nominal pulse waveform trigger threshold 25mV (actual trigger threshold less than 20mV) in the most sensitive setting. A sensitive preamplifier has not been incorporated in the present design because in many applications it would lead to hum trouble with erratic triggering. With its self-contained battery circuit the i.c. preamplifier is completely free from such troubles and gives accurate triggering here even with pulse waveforms in the millivolt range.

The rise time of a positive-going flank of the waveform applied to the pulse input P5 must be equal to or shorter than 10μsec for the required threshold amplitude of 3·5V. Waveforms which do not contain a positive-going flank with at least this steepness and amplitude are unacceptable at P5 and produce no response there. These waveforms must be fed to the a.f. input P1.

Scaler/Ratemeter Section

The range switch S2 (see Fig. 4) feeds the Q-pulses from the amplifier section either straight through to the clock pulse input CP of the monostable IC7 or to this same point via one or both of the 10:1 pulse scalers IC5 and IC6. Before describing the circuit action we will briefly recapitulate the input-output terminology for these digital i.c.s. CP is the abbreviation used to indicate a clock pulse input and Q is the general symbol for an output. We write Q if the output normally rests in the low state (logical 0, voltage less than +1V) and goes to the high state (logical 1, voltage greater than +3·5V) when a signal is present. For the converse case we write Q̄. Q and Q̄ are complimentary and many devices provide both. This is the digital circuit equivalent of a push-pull output in an analogue circuit. If several Q and/or Q̄ outputs are present in a device they are numbered in succession. The 10:1 pulse scalers used possess Q1, Q2, Q3 and Q4 outputs. The states of the Qs change each time the input CP drops from the high to the low state. There is no response to a change at CP from the low to the high state. Thus a "count" is produced at the *end* of a Q-pulse (base line low state, pulse high state) or at the *start* of a Q̄-pulse (base line high state, pulse low state). Tr10 in the amplifier section produces Q-pulses so that IC5 or IC7 respond on the flank at the end of each such pulse.

10:1 Pulse Scaler

Table 1 shows the truth table for this digital integrated circuit (MC838P). The initial state is assumed to be the one with all Qs in the low state. The successive lines of the truth table show the states of the Qs after the specified numbers of Q-pulses have been fed to the input CP. Only Q4 is actually used on IC5 and IC6. We see in the truth table that Q4 remains low (0) until the eighth Q-pulse has been fed to CP when it goes high (1). It remains high when the ninth Q-pulse arrives but returns to 0 at the end of the tenth Q-pulse. All the other Qs as well are then 0 so that the initial state exists once again after 10 Q-pulses have arrived.

It is the transition from high to low at the end of

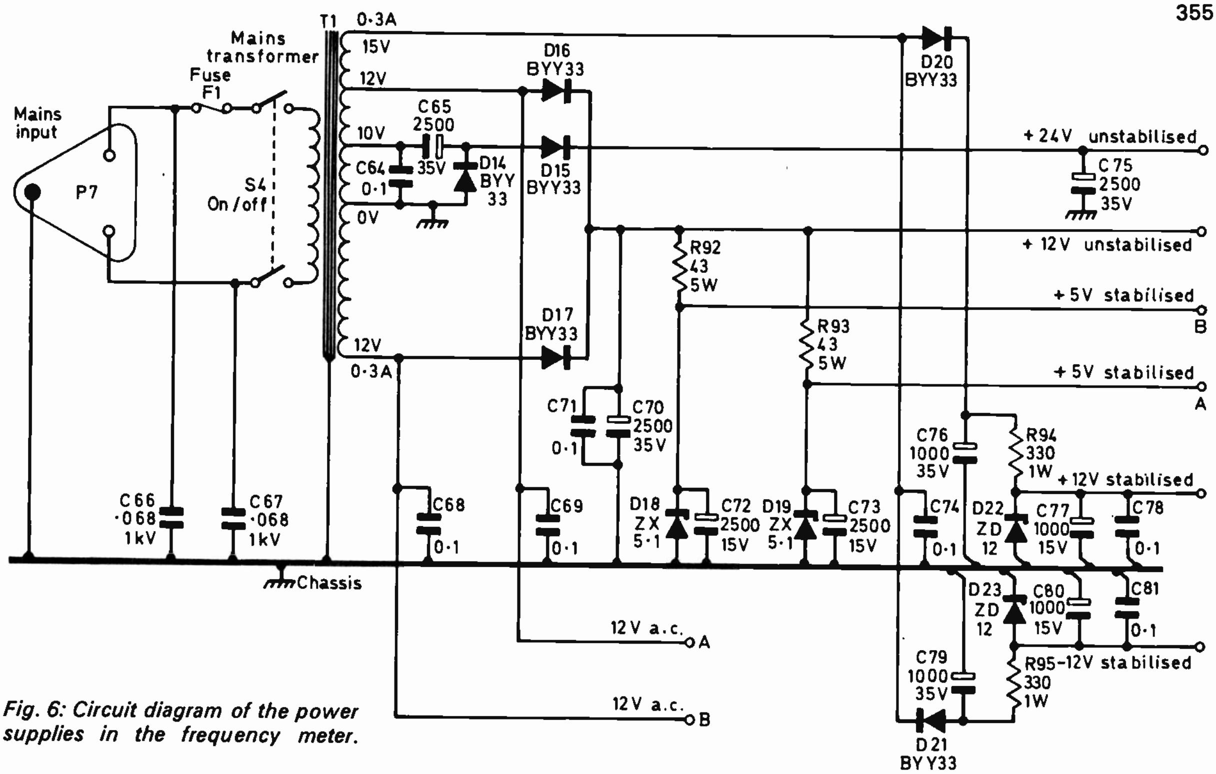


Fig. 6: Circuit diagram of the power supplies in the frequency meter.

a Q-pulse which gives a response at a CP, so Q4 of IC5 is strapped straight through to CP of IC6. By the same action the transition at Q4 of IC6 from 1 to 0 takes place on every *tenth* 1 to 0 transition of IC6 CP. This is Q4 of IC5 and thus corresponds to every *hundredth* 1 to 0 transition of IC5 CP which is the end of every hundredth Q-pulse from the amplifier section.

Any desired number of 10:1 pulse scalers may be connected in cascade in this manner to scale down by any required power of ten. There is no problem of instability because the signal levels are not being amplified. Thus neither decoupling chains nor special coupling circuits are required. Q4 and CP of successive stages are directly connected along the entire chain but the connections must be kept very short because the pulse flanks lie in the v.h.f. range. A common power supply rail is sufficient, without decoupling chain, allowing about 30mA for each stage. An 0·1μF bypass capacitor is desirable between the power supply and ground pins directly at each i.c. package, e.g. C41 and C42 on IC5 and IC6.

Table 1 : Truth table for 10:1 pulse scaler

No. at CP	Q1	Q2	Q3	Q4
Initial State	0	0	0	0
1	1	0	0	0
2	0	1	0	0
3	1	1	0	0
4	0	0	1	0
5	1	0	1	0
6	0	1	1	0
7	1	1	1	0
8	0	0	0	1
9	1	0	0	1
10	0	0	0	0*

* Initial state once again.
The tenth pulse at CP causes Q4 transition from 1 to 0. It is thus able to produce a response at CP of the next stage.

Q and $\bar{Q}$ Pulse Driver Monostable

It is evident from the truth table of the 10:1 pulse scaler that none of the Qs provide pulse outputs. All these outputs are so-called *step functions*. The input signals change the outputs between the high and low states, these states being held indefinitely until further signals arrive which may in principle take infinite time. A pulse of course is a change of state followed by a reversal of that change of state after a time *t* called the pulse width.

It was the function of the pulse univibrator Tr8, Tr9 in the amplifier section to produce pulses from the trigger signals and essentially the same purpose is fulfilled by the Q/$\bar{Q}$-pulse driver monostable IC7 in the scaler/ratemeter section. The high to low transition step at IC7 CP causes its Q and $\bar{Q}$ outputs to change state *for a pulse width time determined by the external capacitor C43* after which Q and $\bar{Q}$ return of their own accord to their resting states. Thus the waveforms at Q and $\bar{Q}$ are true square pulses. C43 is connected across the pins marked TC, i.e. time-constant.

The Ratemeter

IC8 is yet another application of our familiar type MC1709CG operational amplifier, this time as a pulse ratemeter. This is a circuit giving a d.c. output voltage linearly proportional to the repetition rate (frequency) of a train of uniform input pulses fed to it. This operation is called integration and is essentially the same as smoothing the output of a rectifier in a power supply circuit. The pulses must first be rectified, i.e. made to have entirely one polarity with respect to a base line clamped to zero volts. The d.c. restorer D27 and C45 provides this condition. The integrator itself behaves as a large capacitor: each unipolar pulse pumps a certain charge into this capacitor so that the total charge is proportional to the number of pulses. At the same time a resistor in parallel with the capacitor drains the charge away within a certain time so that in the steady state the voltage to which the capacitor rests charged is proportional to the number of pulses arriving within the discharge time, i.e. it is proportional to the frequency.

In our circuit we require negative-going input pulses because these are fed to $\overline{IP}$ of IC8, polarity inversion thus occurring in this stage, and we require a positive d.c. voltage at OP for the meter circuit. The correct polarity pulses are available at $\overline{Q}$ of IC7 and are d.c. restored to be entirely negative-going with respect to zero across R73.

For the duration of each pulse VR10 passes a certain definite current into IC8 $\overline{IP}$ thus transfering a certain charge to this point. The large-value capacitor C47 is used in parallel with the feedback resistor R64 of the operational amplifier. By ordinary Miller effect the function is exactly the same as if C47 was not present but instead a capacitor A times as large was connected directly between $\overline{IP}$ and chassis, where A is the closed-loop gain of the operational amplifier. The "Miller capacitor" at $\overline{IP}$ is the smoothing capacitor and VR10 the smoothing resistor for the integrator circuit considered in analogy to a power smoothing circuit.

The offset is balanced in the normal manner with VR11 so that the output voltage is exactly zero when no input pulses are present. VR10 is then adjusted to give exactly full-scale deflection for a $\overline{Q}$-pulse frequency of 1kHz from IC7. The value of C43 has been chosen to make the $\overline{Q}$-pulses about 300μsec long. Thus the maximum possible $\overline{Q}$-pulse frequency is about 3kHz giving a very large safety margin as far as the ratemeter is concerned. With one 10:1 scaler in circuit the ratemeter reads 0 to 10kHz and with both 10:1 scalers in circuit it reads 0 to 100kHz. In all cases the actual $\overline{Q}$-pulse range from IC7 is 0 to 1kHz for zero to full-scale meter reading. The decimal ratios of the ranges are effected automatically by the scalers and thus adjustment of VR10 on any range is correct for all ranges.

Digital Counter Driver

The Q-pulse output of IC7 is used to drive the digital counter section. The start/stop switch is incorporated here. In the stop setting Tr17 base is connected to chassis by S1C. In the count settings (1/10s or 1s) C84 and D26 form a positive d.c. restorer for proper Q-pulse drive of Tr17 which steps down the impedance to the low value required for driving the frequency counter chain.

Figures 3 and 5, the time and frequency counter circuits, will appear next month along with the circuit description of these sections of the frequency meter. Because of the rather specialised nature of this particular feature and the shortage of space in the magazine at present it is not proposed to publish detailed layout diagrams. These are however available on request.

CONTINUED NEXT MONTH

THE FAM COLOUR SYSTEM

PAUL SILVERHAY

FREQUENCY-AMPLITUDE modulation (f.a.m.) appears to be a conflict of terms: normally if you want to amplitude modulate a signal you try to prevent any frequency modulation and vice versa. F.A.M. is however a complete process rather than a type of modulator and at one time was a contender in the European colour standards battle, following NTSC, PAL and SECAM. The system is a development by the German research group I.R.T. (Institut for Rundfunktechnik) and one of the major uses to which it is now being put is as a colour converter for low-cost videotape recorders. Equipment for this purpose is being produced by Ampex (Gt. Britain) and Vitronic of W. Germany.

Bandwidth is not normally a limiting factor in recording colour on a low-cost recorder. If a machine has a bandwidth of say 3·5MHz, it would be a relatively simple job to take a PAL signal and modify its standards so as to have a colour subcarrier at say 2·8MHz. This would give acceptable results for closed-circuit use even though the luminance bandwidth would have to be restricted to about 2MHz.

A more important limitation is the stability of the machine—the head-to-tape speed stability and the effectiveness of the servo systems. Efficient servo systems are not cheap and this has in practice meant that, although an acceptable monochrome recorder can be purchased for £300, more than four times this price is required for a machine with sufficient stability for colour. This is still low-cost in comparison with broadcast-standard videotape equipment but for the majority of users—educational, industrial and private—£1,000 or more is considered to be high-cost. The f.a.m. system has the advantage that it is much more insensitive to these stability variations than other colour-coding systems because with f.a.m. the subcarrier modulation itself is not especially sensitive.

The basic system used by Ampex and Vitronic in their f.a.m. colour adaptors is based on videotape recorders which have a bandwidth limit (−3dB) of about 3·5MHz. This is the system we will describe here. The complete equipment consists of a f.a.m. coder whose output is fed to the recorder and a f.a.m. decoder which is used on replay to obtain normal colour signals from the recorder.

FAM Signal Encoding

The basic encoding process is shown in block diagram form in Fig. 1. There are two basic input signal conditions. The first is that shown fully in Fig. 1, with studio signals from for example a camera: the signals here will consist of R, G and B inputs. The other condition is for recording off-air, when luminance and colour-difference signals would be obtained from a colour receiver: these inputs would be inserted after the matrix which changes any RGB inputs into the same signal form.

The luminance (Y) signal is then bandwidth restricted by a low-pass filter with a −3dB point at 2MHz and a −6dB point at 2·8MHz. The two colour-difference signals are also bandwidth restricted by low-pass filters with −3dB points at 0·7MHz and −6dB points at 1MHz.

The R−Y signal is used—after filtering and pre-emphasis—to frequency modulate a carrier at 2·65MHz. With this centre frequency (fc) the deviation for full-amplitude colour bars is $\pm$500kHz (i.e. $\pm\Delta fc$). The equivalent V signal amplitude fed into the modulator for full-amplitude colour bars is 0·49V. As with any f.m. system, noise is produced which is of a triangular nature—increasing at the higher modulating frequencies. Pre-emphasis is applied to the signal before modulation to maintain the same noise factor for all input levels.

This f.m. signal with its centre frequency of 2·65MHz is then used as the carrier in an amplitude modulator. This is fed with the U signal as the modulating information. Whether or not R−Y information is present on the colour signal, the output of the frequency modulator will be constant in amplitude at 0·3V This therefore will also be the output from the amplitude modulator if B−Y information is not present. The maximum amplitude for the U signal entering the amplitude modulator for full-amplitude colour bars is 0·63V. The output from the stage for this input would be from −0·16V to +0·76V which corresponds to a modulation depth of 153%. This gross amount of overmodulation is not permissible however because the subcarrier output from the modulator in this case falls to zero. In practice the average value of the B−Y signal is unlikely to cause half this modulation depth—i.e. 77%—giving amplitude swings at the output of the modulator of 0·07 to 0·53V.

It is preferable to increase the subcarrier amplitude rather than to reduce it and for most colour pictures B−Y has average negative values. The input is therefore phase reversed to −(B−Y) to achieve a general increase in subcarrier amplitude when the modulation takes place.

The R−Y signal is chosen for the frequency modu-

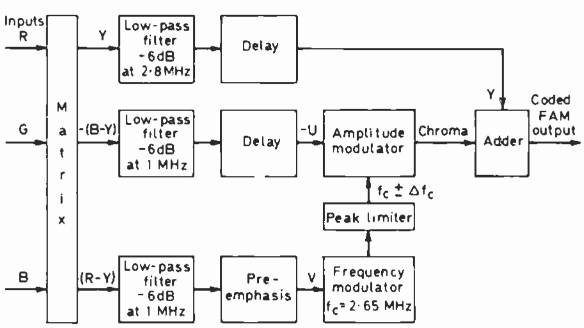


Fig. 1: FAM encoder block diagram.

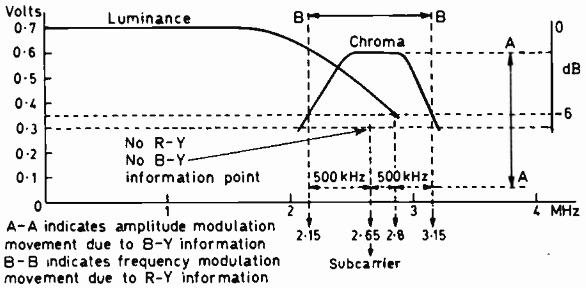


Fig. 2: FAM system parameters—signal bandwidths and amplitudes.

lation process rather than the B − Y signal because f.m. is more insensitive to amplitude variations in the system. As the R − Y signal contributes most to the total picture luminance (for average pictures) it is chosen for this f.m. treatment.

The combined modulation output from the amplitude modulator consists then of R − Y signals frequency modulating a subcarrier at 2·65MHz, this subcarrier itself being amplitude modulated by the B − Y signal. In the absence of B − Y and R − Y signals the output will be an 0·3V 2·65MHz sinewave. With R − Y present the amplitude will remain constant but the frequency will vary with the amplitude of the R − Y signal with a maximum deviation of ±500kHz either side of the subcarrier frequency. With B − Y information also present the amplitude of the subcarrier will vary between the limits we indicated before.

The modulator output is then added to the bandwidth restricted luminance signal of 0·7V to form the composite f.a.m. output.

FAM Signal Bandwidth

Taking account of the factors given before, the maximum limits of the coded f.a.m. signal are given in Fig. 2. This response shows that the absolute maximum bandwidth of the signal is about 3·15MHz and that the maximum amplitude of the chroma approaches the maximum amplitude of the luminance, therefore giving similar noise performances. There is the obvious possibility of some crossover of information between the luminance and chroma signals above 2MHz, but because of the f.m. used and the decreasing luminance signal content at these frequencies the effects are not very annoying on the majority of pictures. The interference that does occur shows as subcarrier patterning on some colour areas —particularly cyan, yellow and green. The patterning is an uneven movement and can therefore be annoying on large colour areas. To prevent this the

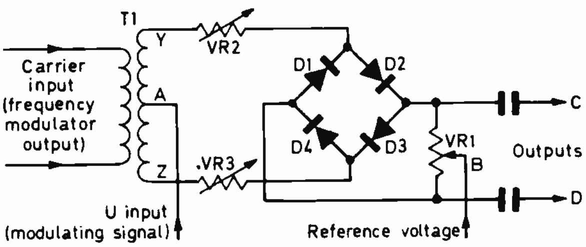
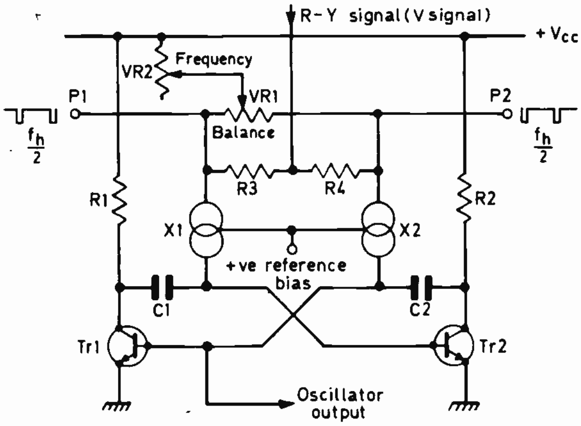


Fig. 3: The amplitude modulator circuit.



subcarrier oscillator is locked with a 180 degree change in phase on alternate lines. This reduces the structure of the patterning because the spectrum of the subcarrier consists mainly of odd multiples of half-line frequency (just as in NTSC).

It can be seen then that the bandwidth of the f.a.m. signal—with the chosen parameters—is suitable for use with a low-cost v.t.r. having a − 3dB point at say 3·5MHz. Reasonable performance could also be obtained from a v.t.r. with a 3MHz bandwidth.

Amplitude Modulator

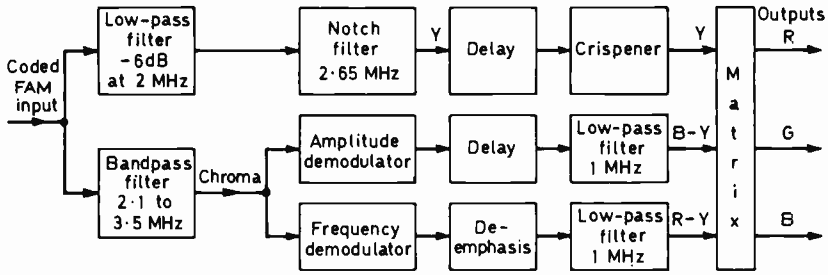
Most of the circuits used to fill the blocks of Fig. 1 are quite conventional. Of particular interest though are the two modulators. The amplitude modulator is of balanced diode ring formation as shown in Fig. 3. The U modulating signal input is effectively connected between A and B. The carrier (the frequency modulated V signal) appears across the secondary of T1, out of phase at each end of the winding. When the carrier is positive at Y and negative at Z diodes D2 and D3 conduct and as the bridge formed by these diodes and T1 secondary is then of very low impedance point C rises to the modulating voltage (across A-B). During the following half-cycle of carrier diodes D1 and D4 conduct and point D rises to the modulating voltage. If the outputs at C and D are taken to the inputs of a difference amplifier (see for example Fig. 3, page 536, September, 1969) the output from this will be a peak-to-peak voltage of twice the modulating signal switched at the carrier rate—i.e. an amplitude modulated signal. The small reference voltage of about 0·2V at B sets up the modulating conditions so that blanking level on the U signals corresponds to zero modulation. VR1, VR2 and VR3 allow accurate adjustment of the balance of the circuit.

Frequency Modulator

The frequency modulator used is slightly more sophisticated. Its basic element is a free-running multivibrator. The frequency of this kind of circuit can be adjusted by a d.c. potential so that an a.c. signal applied in its place can produce frequency modulation. The centre frequency will not however be very stable and as a result a moving subcarrier structure will appear on the colour picture. This is particularly disturbing because of its uneven, restless motion. The answer to this problem is to stop the

Fig. 4 (left): Circuit diagram of the frequency modulator.

Fig. 5 (right): Block diagram of a f.a.m. decoder.



multivibrator during line and field blanking periods and to restart it when wanted in the desired phase. This also prevents the subcarrier appearing on the line sync pulses of the system output. The best condition is to restart the oscillator 180 degrees out of phase with the preceding line. Both the stop and start functions can be achieved by means of line pulses. The arrangement is shown in Fig. 4.

Tr1 and Tr2 are the active multivibrator elements. X1 and X2 are shown as constant-current sources—in the practical circuit they are transistors performing this function—supplying current to the charging capacitors C1 and C2 and therefore affecting the frequency of operation. With no R – Y signal present the centre frequency of the oscillator is determined by the reference bias on X1 and X2 and the d.c. supplied from VR2 ("set centre frequency") from the voltage rail. VR1 balances the two halves of the circuit accurately to combat inequalities due to component tolerances. The appearance of the R – Y(v) signal at the centre point of R3 and R4 then alters the oscillator frequency in sympathy with its amplitude—i.e. frequency modulation is obtained. The R – Y signal has already been passed through an amplifier with pre-emphasis applied at h.f. as with any process using f.m., in order to reduce the effects of triangular f.m. noise. The circuit used for this is a standard *RC* filter.

Two trains of negative-going pulses are applied to P1 and P2. Their repetition rate is at half-line frequency and the two trains are offset by one line to each other—i.e. one pulse occurs at P1, another a line period later at P2 and another line period later a further pulse occurs at P1. The pulses counteract the positive potentials at the junction of X1, R3 and X2, R4 stopping the multivibrator. When a pulse finishes the multivibrator is biased to oscillate again. The action on alternate sides makes the oscillator start 180 degrees out of phase on alternate lines, reducing the residual subcarrier appearance at the decoder output to a locked fine dot pattern of low visibilty and annoyance.

The oscillator output is taken to the amplitude modulator (Fig. 1) via a diode peak limiter which eliminates any amplitude modulation caused by the f.m. process.

The Decoding Process

With very little addition the decoding process is the reverse of the coding system—see Fig. 5. From the modulated signal the subcarrier and its components are stripped off using a bandpass filter with –6dB points of 2·1 and 3·5MHz. This chroma signal is applied to both an amplitude and a frequency demodulator. The frequency demodulator is followed by the de-emphasis circuit necessary to restore a linear frequency response. The amplitude demodulator, an envelope detector, is insensitive to frequency changes while the frequency demodulator is insensitive to amplitude changes. The outputs are therefore pure B – Y and R – Y. 1MHz filters are inserted in each output to remove spurious signals outside the necessary bandwidth and this includes h.f. noise.

As in any colour system the subcarrier must be removed in the luminance chain. Whereas in PAL and NTSC the subcarrier is of constant frequency and therefore easily removed by a notch filter, in f.a.m. the subcarrier is effectively varying from 2·65MHz ± 500kHz. For adequate subcarrier suppression and therefore minimum visibility on coloured areas of the picture at least 9dB of attenuation should be present at the lowest frequency point (2·15MHz) with somewhat more attenuation at the more frequently occurring subcarrier frequencies (i.e. around 2·65MHz).

Two possibilities for this subcarrier suppression come to mind. The first is a low-pass filter with a sharp cut-off at about 2·1MHz. This would give the necessary attenuation but the luminance signal would almost certainly ring. A low-pass filter with a less sharp cut-off would prevent such overshoots being developed but insufficient rejection would be given around the subcarier centre frequency. The solution is to use this latter filter but with a notch filter follow-it centred at 2·65 MHz.

Compared to a standard PAL or NTSC signal the most obvious deterioration in picture quality is the restricted luminance bandwidth. This also results in a lower picture brightness because there is less luminance energy. The "crispener" circuit seeks to re-establish some of this lost resolution and brightness. This can be done by enhancing the existing waveform edges in the luminance signal. A block diagram of the arrangement is shown in Fig. 6 and the waveforms which might be expected are shown

—continued on page 363

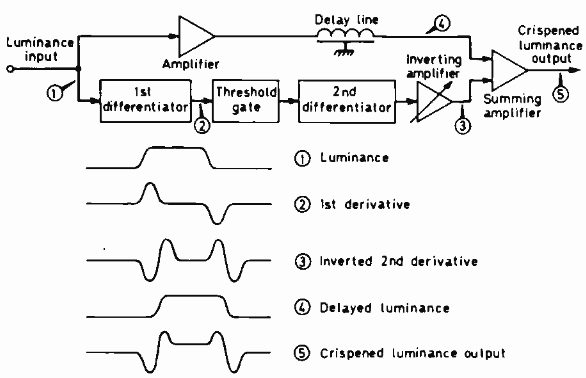


Fig. 6: Luminance crispening—block diagram and waveforms.

ADDING AUTOMATIC F

PUSH-BUTTON tuning on u.h.f. while being very convenient often leaves a margin of mistuning, especially after some wear and tear has occurred on the mechanism. Even dial tuning can lead to errors due to the difficulty many people experience in judging the correct point. Oscillator drift due to temperature changes can also cause mistuning. Automatic frequency control (a.f.c.) will correct all these faults.

The vision carrier when the set is correctly tuned on u.h.f. is at 39·5MHz as it passes down the i.f. strip. Thus if at the end of the i.f. strip a discriminator tuned circuit is incorporated centred on 39·5MHz the discriminator output will be zero at 39·5MHz and will move positively one side of 39·5MHz and negatively the other as the tuning drifts. This response is shown in Fig. 1.

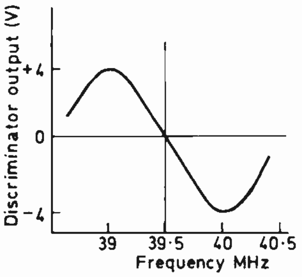


Fig. 1: Showing how the discriminator output varies with frequency. At the centre frequency—39·5MHz—its output is zero.

If the tuning is not correct then the discriminator output is not zero and if this output is applied to change the reverse bias on a tuning diode mounted in the oscillator section of the u.h.f. tuner it will correct most of the error. Tuning, varicap or varactor diodes—to give them a few of their names—are junction diodes normally operated with reverse bias but not sufficient to bias them into the breakdown region in which zener diodes operate. The greater the reverse bias the lower their capacitance: a typical curve, for the Mullard BB105 or STC BA141 tuning diode, is shown in Fig. 2.

All diodes exhibit this basic type of characteristic but special diodes have to be used for u.h.f. because they must not introduce any excessive loss into the tuned circuits they control. In other words, just as a coil has to have a good Q so does a varicap diode. Normally, we don't worry about the Q of a capacitor as it is usually very good. However, a tuning diode is not a true capacitor. It has, for example, leakage current so the Q of the diode is a factor which has to be considered. The diode manufacturer however will have considered these points and if you buy a diode specified for u.h.f. use you will have no trouble. These points have been mentioned to clear up any misunderstandings and to show why any old diode won't do.

Basic AFC System

To return to our TV set, if the oscillator frequency is too high then the vision carrier frequency will also be too high and in the simple arrangement shown in Fig. 3 the discriminator will give a negative signal to decrease the bias on the tuning diode thus increasing its capacitance and in turn reducing the oscillator frequency and correcting the error. Note that in this diagram the reverse bias on the diode is applied to its cathode. It is therefore positive with respect to ground so that a negative signal from the discriminator will reduce the positive voltage on the diode thus reducing its bias and increasing its capacitance. In this arrangement the diode is biased somewhere near the mid-point of its characteristic by the positive d.c. bias fed into one side of the discriminator. The discriminator thus adds to or subtracts from this d.c. bias.

AFC Loop Gain

The amount by which the error is reduced depends on the gain of the circuit. An estimate of the gain required must first be made by guessing how much error is likely to be given by your push-buttons or hand tuning: 1MHz would be an outside figure as a tuning error of that magnitude would produce a very bad picture of low definition in one direction and badly broken up in the other. This error should be reduced to about 100kHz to be really unnoticeable, indicating a required gain of ten.

In fitting a.f.c. to an existing set some measurements should be done as an experiment before finally deciding on the circuit gain. The first thing to do is to add the suggested discriminator to the i.f. strip. As the circuit (Fig. 4) shows a Foster-Seeley type discriminator is used and with the coils specified and the driver circuit shown it should give ± 4V for $\pm 0{\cdot}5$MHz input variation.

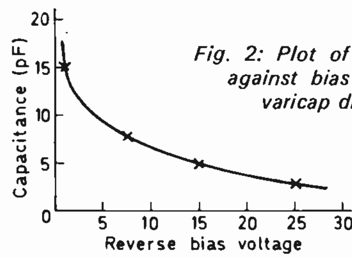


Fig. 2: Plot of capacitance variation against bias voltage for a typical varicap diode.

Circuit Description

The driver stage Tr1 takes a small sample signal from the i.f. strip but this should be large enough to drive Tr1 into saturation. That is to say Tr1 is a limiter stage so that the signal amplitude applied to the discriminator coil L2 stays constant over the nor-

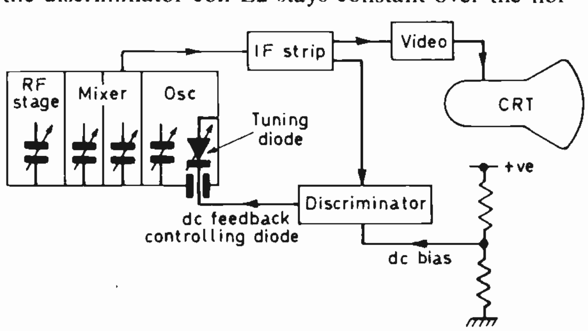


Fig. 3: Simple discriminator feeding a varicap tuning diode in the u.h.f. tuner.

T. SNOWBALL

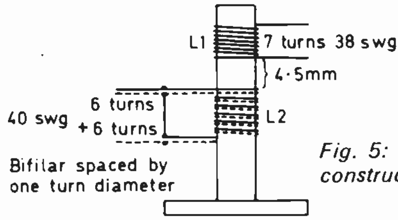
[F]REQUENCY CONTROL

mal range of signal levels. Tr1 is biased at approximately 7mA which, according to the original report ("Simple a.f.c. system for 625-line TV receivers" by P. Bissmire, *Mullard Technical Communications*, March, 1970), gives the best limiting performance. C1, R14 and R3 damp the stage to prevent oscillation. C2 decouples the power feed and should be close to the circuit. The coil former and can are the normal ones used for TV sets and so should be easily obtainable: the former diameter is 5mm. and length 40mm. and winding details are given in Fig. 5.

Setting up the Discriminator

Setting up the discriminator is much easier if a signal generator is available. The collector coil L1 is set to 39·5MHz. This is not easy to do using the discriminator diodes as detectors and an alternative way is to rig up a temporary detector fed from a small capacitor of 2-3pF connected to the BF194 collector. The secondary coil L2 is adjusted for 0V output at 39·5MHz across C7, with a smooth transition to peaks of equal and opposite amplitude at 39 and 40MHz. If the primary L1 is not on frequency at 39·5MHz then the peaks of the secondary will not be equal either in amplitude or frequency spacing from the 0V point.

If a signal generator is not available the TV set itself can be used but more care is necessary to ensure that the discriminator is set up correctly. First of all the test card must be used to set up correctly the tuning of the TV set. The discriminator then has to be connected to the set. For valve i.f. strips the method shown in Fig. 4 is suggested. The addition



Two cores are required, Neosid grade 900.

Fig. 5: Discriminator coil construction.

of the coupling capacitor C13 will of course slightly shift the i.f. response of the set. This must be corrected first by adjusting the core of the transformer to which C13 is connected until as viewed on test card the set is correctly aligned once more.

Having done this the next step is to adjust L2 for 0V output. Then, by swinging the u.h.f. tuner, the output can be made to move positive and negative as in Fig. 1 with adjustments made to L1 to ensure equal amplitude peaks evenly spaced about the position of correct tuning—which of course must be made 0V by checking L2. As L1 and L2 interact somewhat this must be done carefully and slowly. It is easy to see that a signal generator is the best method of setting up.

Connecting to the IF Strip

If the discriminator is set up using a signal generator then on connecting it to the i.f. strip the point just mentioned about readjusting the TV set's final i.f. coil is of course still valid. The value of the coupling capacitor C13 should be kept low to reduce the effect of added capacitance on the i.f. transformer but must be big enough to drive the transistor well into limiting in order to keep the discriminator output constant for varying gain level in the i.f. strip, changes of contrast and a.g.c. action. The way to go about this is to start with a small capacitor of 1 to 2pF and at a given output, say 2V, alter the contrast control setting: over a reasonable range the discriminator output should stay 2V. If this is not so a larger coupling capacitor can be tried remembering always to cancel out the effect on the set's i.f. strip by retuning the

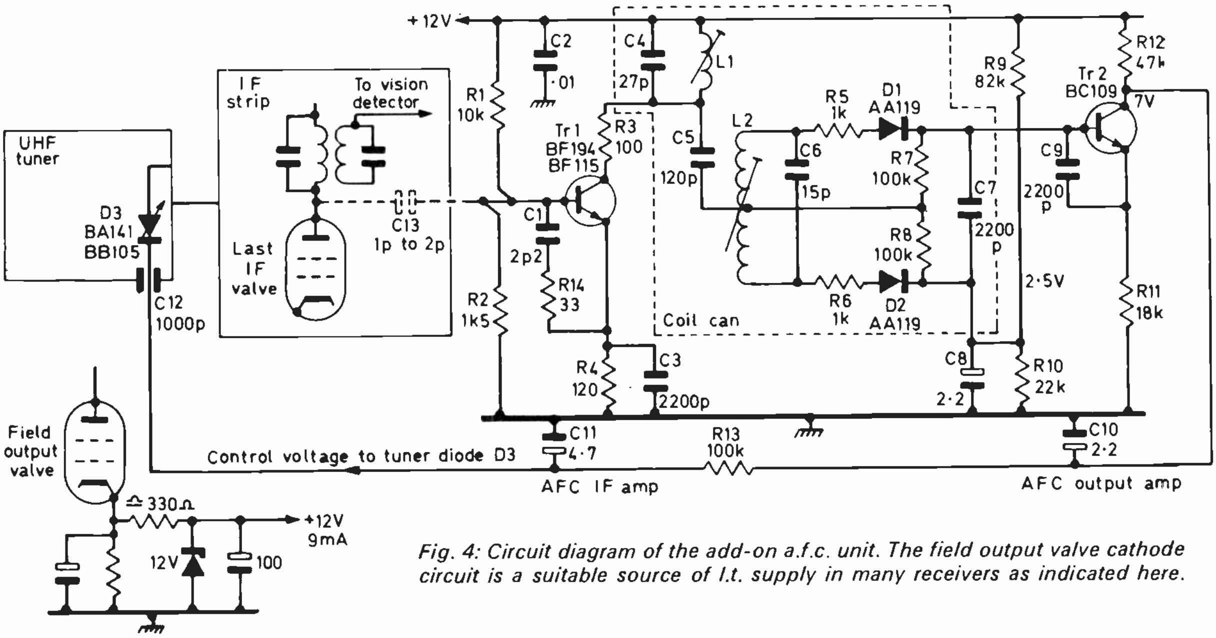


Fig. 4: Circuit diagram of the add-on a.f.c. unit. The field output valve cathode circuit is a suitable source of l.t. supply in many receivers as indicated here.

i.f. transformer core.

If the i.f. strip is transistorised then the voltage levels will be smaller and larger values of C13 needed. A slightly different input circuit can be tried as indicated in Fig. 6—assuming that the detector diode connection is available. Alternatively, for example, in the Constructor's TV set last year by Keith Cummins the original i.f. strip used has a 6·8pF capacitor

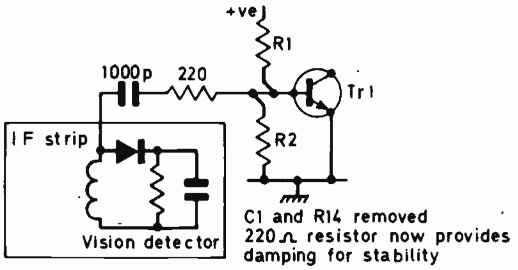


Fig. 6: Alternative coupling from a low-impedance, low-voltage point in an i.f. strip.

(C25, Fig. 2, Page 25, October, 1970, TELEVISION) which is normally connected to earth on 625 lines. This could, of course, be used as the coupling capacitor C13 in this article. In his later i.f. strip (September, 1970, page 554) the coupling could come from Tr4 collector or the junction of D1 and D2 on L6 using the circuit suggested in Fig. 6.

Tuning Diode Adjustment

The next thing to do is to set up the operation of the tuning diode. As all tuners are different the author can only describe how he fitted up his own set but the basic idea can be applied to any tuner. The tuned circuits in u.h.f. tuners consist of small resonant cavities tuned by small tuning capacitors. The diode must be coupled into the oscillator section in such fashion as to give a useful tuning range but not sufficient to impair the oscillator operation. It was not found to be very critical.

Figure 7 shows the usual layout of a cavity. It can be seen that the tuning capacitors are connected from the top or live end of the "coils" and it appeared best to do the same with the tuning diode. An earthing point near the dead end of the tuned circuit was first found and a feedthrough capacitor fitted (after drilling a suitable hole) near the live end. The diode was mounted by its leads and correct control was obtained by bending the diode nearer to or away from the tuning line. Bias the diode for these tests with a

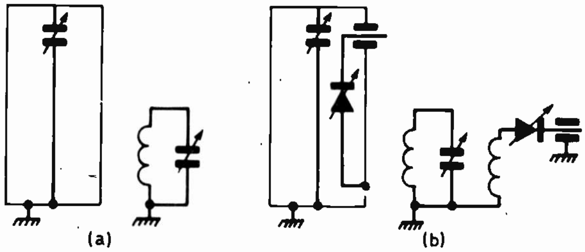


Fig. 7: Cavity tuned circuit mechanical layout and electrical equivalent, (a) without and (b) with tuning diode.

potentiometer which takes the diode over its working range—0 to 25V for the STC BA141.

Set the push-buttons or dial of the set to the correct position for a station. Then on putting the diode in the cavity the tuning will change due to the diode, stray capacitance and stray wiring. The situation will be that even with the diode at minimum capacitance it will be impossible to retune the oscillator: some capacitance must be lost. As a check swing the tuning capacitors to prove that the oscillator operation has not been upset. Then reset the tuning to the correct position and remove capacitance by either turning out the trimming capacitor which is usually present or in the limit by bending out the vanes of the oscillator tuning capacitor.

The magnitude of the effect the diode has on the tuning of the set is not too critical. A change of plus or minus 1MHz for example will cause a loss of definition, quite noticeable on the test card when the oscillator frequency is low but when it is 1MHz high the picture will break up due to sound channel patterning. So if this state is reached with anything between ± 1V and ± 5V on the diode all will be well. Obviously the ± 1V case should be aimed for as it increases the gain of the a.f.c. loop.

While the diode is being set up a high-sensitivity test meter (20,000Ω/V) used to read the discriminator output will also show the oscillator frequency shift because the discriminator described will give approximately ± 4V for $\pm 0{\cdot}5$MHz shift. So in the final stages of setting up the diode its frequency versus voltage characteristic can be quite accurately determined.

As a final touch the trimming capacitors should be set to give correct tuning as checked on the test card with a voltage of 7V on a tuning diode type BA141. This voltage will vary with different types of diode but will always be about one-third of the maximum voltage allowed on the diode. The reason for the choice of this voltage is to allow equal swings of

★ components list

R1	10kΩ	R13	100kΩ
R2	1·5kΩ	R14	33Ω
R3	100Ω	All $\frac{1}{10}$W or greater.	
R4	120Ω		
R5	1kΩ	* Values depend on gain required through a.f.c. system. See text on how to design.	
R6	1kΩ		
R7	100kΩ		
R8	100kΩ		
R9	82kΩ*		
R10	22kΩ*		
R11	18kΩ*		
R12	47kΩ*		

Capacitors:

C1	2·2pF	ceramic
C2	0·01µF	paper or ceramic
C3	2200pF	ceramic
C4	27pF	silver mica or polystyrene
C5	120pF	silver mica or polystyrene
C6	15pF	silver mica or polystyrene
C7	2200pF	ceramic
C8	2·2µF	electrolytic
C9	2200pF	ceramic
C10	2·2µF	electrolytic
C11	4·7µF	electrolytic
C12	1000pF	feedthrough
C13	1-2pF	ceramic—see text

Semiconductors:

Tr1	BF194 or BF115
Tr2	BC108 or BC109
D1, D2	AA119 or OA47
D3	BA141 or BB105

frequency either side of the chosen tuning point—the diode capacitance (see Fig. 2) and frequency shift do not change linearly with voltage so that half voltage is not the correct centre point.

Setting the Gain

The system is now ready for its gain to be measured. As mentioned above a test meter can be used to measure the frequency change caused by the diode voltage. If, for example, ±0·5V on the diode causes ±2V from the discriminator this indicates a gain of four or a frequency change of ±0·25MHz. We mentioned earlier a desirable gain of 10. So in this case an extra gain of 2·5 is required. This could possibly be arranged by readjusting the diode coupling on the tuner to make it more sensitive but it is much easier to provide some gain by using an extra transistor between the discriminator and the tuning diode. In the circuit suggested (Fig. 4) an overall gain of 10 is achieved with a d.c. amplifier (Tr2) having a gain of 2·5.

The amplifier should use a high-gain low-current transistor such as a BC109 or BC108. With such a transistor a collector current of 100μA will be produced with a very low base current which will not affect the output voltage of the discriminator. Its collector, which is connected to the diode, should be set at +7V and a filter used to smooth the output and remove pick-up. As 100μA is chosen as a suitable collector current then with a 12V supply 5V must be dropped across the collector resistor R12 making it 47kΩ, a preferred value. The emitter resistor for a gain of 2·5 will then be 47/2·5=18kΩ preferred value where gain=Rc/Re. This makes the emitter voltage 1·8V and allowing a base-emitter voltage of 0·7V the base bias voltage must be about 2·5V This is provided by R9 and R10 and may be adjusted to get Tr2 collector to +7V without any signal fed into the discriminator.

The final setting-up consists of checking that the output is in the correct sense to reduce the tuning error, remembering that the addition of an amplifier will invert the signal and require reversal of the discriminator connections.

AFC/Manual Switch

An a.f.c./manual switch will be found desirable. The easiest way of going about this appears to be to short-circuit the a.f.c. amplifier base to ground: this sets the discriminator output to zero and the amplifier output to the diode to +7V. The push-buttons can then be set for a correct picture after which opening the switch brings the a.f.c. into action. The wire to Tr1 base must be kept very short so that if on a dial-tuned set it is found that the a.f.c. must be disabled each time a different station is tuned in, then as long leads would be needed to a panel switch it is suggested that instead the collector of Tr2 is switched out of circuit and the tuning diode connected to a potential divider providing +7V.

Other constructional points include making sure that all i.f. signal leads are kept very short. The circuit should be attached to the i.f. strip with this point in mind.

Power Supply

The supply to the unit should obviously not drift to any great extent. A useful supply point is the cathode of the field output valve which is usually about 15-17V. Using a 12V zener diode (see Fig. 4) the supply can be made very stable. Otherwise rectification some way up the heater chain with smoothing and again a zener diode will be suitable. ■

THE FAM COLOUR SYSTEM

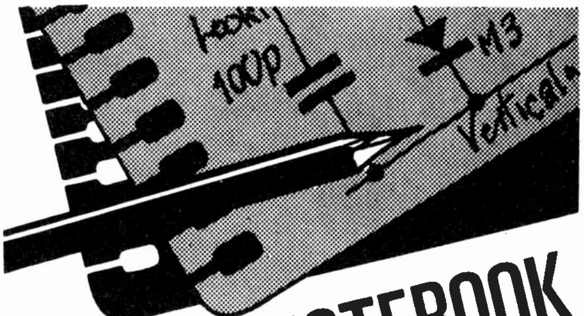
—*continued from page 359*

beneath. The luminance is assumed to be a pulse with slow edges (1). The first differentiator produces a pulse at each edge with the same rise time as the original (2). Each of the edges of this differentiated waveform produce a further pulse at the output of the second differentiator and the resulting waveform is inverted in an amplifier (3). A second signal path has meanwhile delayed the original luminance pulse (4) and these two signals are added in the summing amplifier to produce the output waveform (5). This is effectively a crispened signal because the edges of the pulse change in amplitude considerably more than the original over the same time—i.e. the rise time is faster. The amount of crispening introduced can be controlled by altering the gain of the inverting amplifier in the differentiating chain, thereby changing the proportion of this chain's signal appearing on the output. The delay time in the main path must be carefully chosen for maximum effectiveness.

One of the big disadvantages of this type of correction and any similar techniques is that noise is inevitably added to the signal. The effect of this can be reduced by the introduction of the threshold gate in the secondary chain. This consists of a pair of diode, capacitance-resistance chains which allow only the tips of the pulses through and therefore stop any datum line noise. There is still noise left on the pulse tips, of course, but this has relatively less effect. A secondary advantage of this gate is that the pulse widths are reduced making the correction a little more effective.

We have examined f.a.m. in its application as a low-cost colour v.t.r. system assuming a usable recorder bandwidth of 3·5MHz. Vitronic also manufacture f.a.m. units for use on lower bandwidth recorders of 3MHz and 2MHz and the corresponding frequencies of the subcarrier etc. are changed accordingly. Vitronic also manufacture a version which enables a PAL input signal to be handled and which gives a PAL output signal. These variations enable the equipment to be used with a large number of 0·5in. and 1in. videotape recorders and in a number of systems. The Vitronic equipment has been seen operating with a Sony v.t.r. over a long period and the results have been most acceptable.

It would be a mistake however to assume that f.a.m. has only this type of application. Perhaps one of the future developments will be its use as *the* standard for colour in closed-circuit television. The advantages are low capital outlay coupled with a reasonable performance, the small bandwidths involved enabling existing monochrome equipment such as vision mixers and distribution equipment to be retained. It should be noted that to mix between f.a.m. sources would require the subcarrier frequency to be derived from a common point and fed to all sources and the present coders would require a small amount of modification. ■



SERVICE NOTEBOOK

G. R. WILDING

Thorn 16in. Portables

A portable model fitted with the 16in. version of the Thorn 950 chassis was completely dead and on removing the back we found that although the cartridge fuse was intact the fusible link R147 (see Fig. 1) at the top of the chassis was open. On re-soldering the link and checking across the mains plug we found no continuity and it seemed that a dropper resistor section or valve heater must be open-circuit. This portable receiver is closely based on the standard 950 chassis but after looking around the closely packed cabinet we failed to see any droppers. The unusual value—4·33μF—of the large canned capacitor mounted near the fusible link however obviously indicated that it had a special purpose and by following through the circuitry we found that it is employed as a capacitive dropper in the heater circuit. A check with the service sheet confirmed this and we realised that after resoldering the open fusible link we had been looking for a non-existent fault. On then plugging in we obtained normal reception.

As this heater circuit is so unusual however and as there are thousands of these excellent receivers in use it is worthwhile describing the arrangement in case you come across a model fitted with this chassis. It will be noticed that the solenoids operating the system switch are energised only briefly, during system changeover, and that if misuse results in an overload current opening the fusible resistor R147—as occurred in our case—the mains supply will be completely removed from the receiver. Seven valve heaters plus the c.r.t. heater are in the series circuit with the dropper capacitor C121, the remaining four heaters being shunted across resistors R153 and R152 in series with the neutral mains connection to chassis. The heater current through V7, V3, V6 and V5 will therefore be that of the receiver's h.t. demand and as this is in excess of the 300mA they require R153 is factory adjusted to bypass—with R152—the excess. The total shunt impedance of these valve heaters and resistors forms part of the h.t. smoothing arrangements—in the negative lead—most of the series smoothing being effected by R138 and R151 in the positive h.t. lead.

In this heater circuit arrangement there is no "wasted" electrical energy to unnecessarily raise the cabinet temperature—always a problem with valved TV portables. R153 should not normally require readjustment, but it must be emphasised that if V3, V5, V6 or V7 are withdrawn while the set is on or if one of the heaters goes open-circuit, R152 and R153 will pass the total h.t. demand and will be grossly overloaded and reduce the rail h.t. Similarly if an h.t. short develops in the receiver it will cause excessive heater current through these valves so that if one of them is found with an open-circuit heater it is wise to check that its failure was not due to this cause.

Should an intermittent h.t. short or other fault make it desirable not to energise these four valve heaters the makers recommend connecting a 140Ω, 12W resistor from chassis to the C120-C112 negative connections, leaving out one of the valves. The set will then show a normal raster but without vision or sound.

If it becomes necessary to readjust R153 the correct procedure is to remove the lead from tag 24A on the printed circuit panel (junction V3 and V7 heaters) and connect a milliammeter in series. With the controls set for a normal picture adjust the slider of R153 for 300mA.

The 960Z and 960Y versions of this portable use a more orthodox arrangement with a conduction-cooled mains dropper clamped to the chassis to serve as a form of heat sink and a choke for series smoothing. Other versions use a silicon diode (BY101) dropper for the heater chain.

Note the connection of the heater decoupler C107 to the cathode of the video amplifier: excessive heater current will lead to a shaded raster drawing attention to this condition.

Intermittent Loss of Vision

THE customer's complaint was intermittent loss of vision. On switching on a picture appeared but disappeared almost immediately. After removing the back and lowering the bottom hinged chassis we found that tapping the PCL84 video output pentode would intermittently restore the picture.

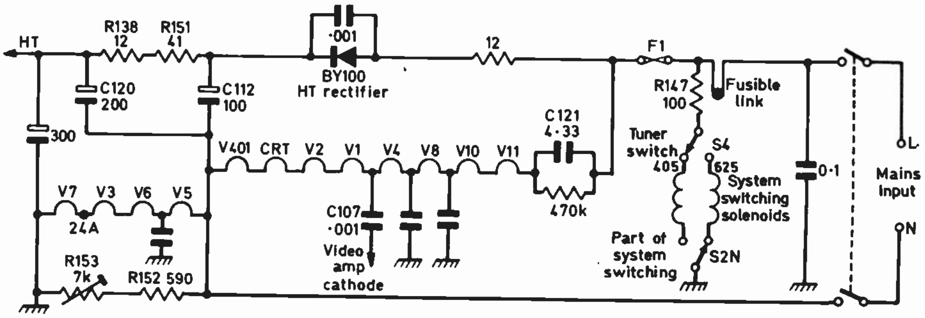


Fig. 1: Heater supply circuit used in some versions of the Thorn 950 chassis in 16 in. portable sets to minimise heat dissipation in the cabinet. C121 acts as a capacitive dropper in series with the heater chain. Note that mains neutral is not connected directly to chassis.

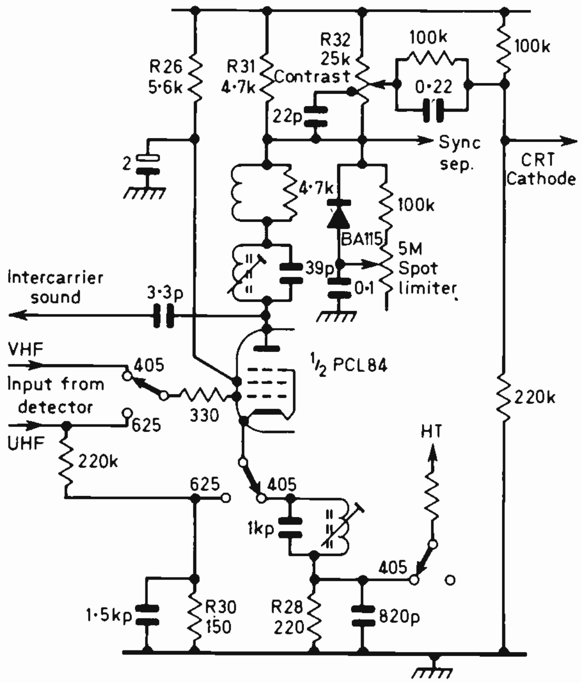


Fig. 2: The video output stage used in the Ekco T433 series. After replacing the PCL84 and R26—the latter being burnt out by an intermittent short-circuit in the original valve—the 405 contrast was well below standard and the preset sensitivity control had to be turned back from its original maximum setting.

We also noticed that the 5.6kΩ screen feed resistor R26—mounted close to the valveholder—was badly discoloured and on applying light pressure it completely disintegrated. Obviously—and as frequently occurs—the PCL84 had developed an intermittent internal short-circuit.

No other resistor or component around the valveholder was discoloured or showed signs of heat damage so we replaced the 5.6kΩ resistor and the PCL84 and tried again. A tolerable picture appeared but it was found necessary to reduce the setting of the chassis-mounted contrast control to obtain optimum contrast, whereas originally the owner could fully advance the control to obtain a really black-and-white picture.

On checking voltages we found that the screen potential was down to about 165V on 405—it should have been 200V—while the anode voltage was also well below normal. We then noticed that both the newly fitted screen feed resistor and the anode load resistor R31 were overheating, the almost certain cause being insufficient grid bias due to value reduction in the cathode resistor. On changing over to 625 and thereby switching cathode resistor R28 out and R30 into circuit, the anode and screen feed resistors stopped overheating and the voltages were restored to the correct 625 figures.

Cathode resistors are usually mounted close to their associated valves—as we had expected when looking for other burnt resistors near R26—but in this particular series the video amplifier ones are mounted some distance away—just under the system switch to facilitate bias change. R28 was only slightly discoloured but when the blade of a small screwdriver was levered underneath it crumbled easily, a sure sign of its having been over-run. We replaced it with a 220Ω ½W type and on switching on again were able to fully advance the contrast control to obtain a really well contrasted picture.

The reason for the contrast peaking before the original maximum setting was that as a result of inadequate bias strong positive-going signals from the detector diode were driving the pentode into the grid current region.

The most common shorts in video pentodes are from the screen to the control grid and from the screen to cathode and if sustained will completely burn up and easily identify all the components passing the excessive current. If only an occasional flashover has occurred (and also in old, well-used sets that have not had this complaint) it is always best to check such resistor values as even small changes can greatly impair the contrast range. Many tubes have been considered past their useful life when in fact the video stage has been cramping or limiting the output amplitude.

Low-emission video pentodes can result in the picture seeming to fade into the background raster as the contrast is increased. This is because of their restricted working grid base. Any input in excess of a certain low figure drives the valve to saturation and irrespective of input signal amplitude the anode voltage can only fall by a fixed limited amount.

TO BE CONTINUED

COLOUR RECEIVER APC LOOPS

—continued from page 349

achieved, the chroma channel conducting only when a colour transmission is present. At other times it is cut off so that spurious signals in the chroma passband do not interfere with the black-and-white display.

Fault-finding

Although the a.p.c. loop operates in an entirely different way to signal amplifying circuits nevertheless fault location in a.p.c. loops is no more difficult and although scope observations are extremely helpful voltage checks alone will in most cases identify the defective stage.

In practice most colour sync faults are due to misadjustment of the preset controls. This means that either the oscillator's operating point is placed outside the pull-in range of the a.p.c. loop or the burst gate fails to correctly and cleanly pass just the bursts from the chroma signal applied to the stage. There will of course be no colour if the reference oscillator is not working.

It must also be remembered that in many receivers the burst gating is done by pulses taken from the line output stage instead of by a pulse instigated by the transmitted line sync pulse. In such cases if the line hold control is not set exactly right there can be unstable colour sync or even no colour at all.

If the colour burst signal does not reach the discriminator there will be no ident signal and therefore no turn-on bias for the colour-killer controlled stage in the chroma channel and in consequence no colour. ■

COLOUR RECEIVER CIRCUITS

COLOUR-DIFFERENCE PREAMPLIFIERS

GORDON J. KING

WE have so far investigated the circuits leading from the aerial input through the front-ends, the i.f. stages and second detectors (as distinct from the synchronous detectors of the colour decoder) to the cathodes of the picture tube via the Y or luminance stages. In other words we have covered all those signal circuits which are involved in the production of a monochrome display or the "brightness" of a colour display.

The Y signal is responsible for the *detail* of a colour display; in fact it is this signal which provides *all* the information required for a monochrome picture. Colour is added to the monochrome detail separately, and the definition of the colour information is only about one-fifth that of the luminance. This means that the colour is added rather boldly to the luminance detail. It is the job of the colour decoder to process the colour information carried by the composite aerial signal in such a way as to secure three new signals, called *colour-difference signals*.

At the camera three primary-colour signals corresponding to the saturation of the red, green and blue scanned elements of the scene are obtained. The Y signal is derived from these three signals by addition and it has been discovered that the Y signal is nicely balanced (corresponding to the video signal which would be expected from a monochrome camera when scanning a colour scene) when it consists of 30% red signal, 59% green signal and 11% blue signal. The colour camera thus delivers the Y signal composed as above plus red, green and blue signals.

The Y signal is fed to the transmitter in a similar manner to the video signal of a monochrome system (which is why we can run our monochrome sets from it). A colour transmitter however has to handle the colour signals as well and these cannot be multiplexed with the Y signal as they stand because for one thing there is insufficient vision channel bandwidth and for another the composite signal must be usable by a monochrome set. The multiplex signal actually transmitted must therefore be carried within the bandwidth of an ordinary 625-line channel.

Colour-difference Signals

The problem is solved by producing colour-difference signals; that is by subtracting the Y signal separately from each primary colour signal thereby obtaining R−Y, G−Y and B−Y signals. A bonus of this scheme is that colour-difference signals occur only when there is colour in the scene. If the scene is in monochrome (no colour) then only the Y signal is produced.

Another very useful feature of the scheme is that the G−Y signal can be obtained by matrixing the R−Y and B−Y signals in the receiver so that we need transmit only the R−Y and B−Y signals along with the Y signal. This eases transmission and bandwidth problems. The two colour-difference signals along with the Y signal and the sync pulses are "multiplexed" to form a complex signal which is modulated on to the vision carrier. It is the job of the receiver to sort this lot out.

We have already seen how the Y signal is sorted out but are not yet ready to investigate the circuits which sort out the colour information. What we must do first is to look at the circuits which handle the R−Y and B−Y signals emanating from the colour decoder, the circuits which produce the G−Y signal and then follow the three signal paths to the grids of the colour picture tube. To do this therefore we must for the present take it for granted that the colour decoder produces R−Y and B−Y colour-difference signals.

Colour-difference Drive

It is best to look at the picture tube end of the colour-difference channels first of all to get some idea of how the three primary-colour signals are derived by the tube guns. Fig. 1 shows the basic arrangement. Here the Y signal is applied to all three cathodes while the R−Y, G−Y and B−Y colour-difference signals are applied separately to the grids of the red, green and blue guns. Each gun therefore is fed with a combination of the appropriate colour-difference signal and the Y signal. This is the system known as colour-difference drive.

The Y signal is ordinary video and is negative-going as shown at (a) in Fig. 2 (for the colour-bar test-pattern waveform). The colour-difference signals are also now at video frequency as distinct from the V and U chroma signals—since they are derived by the decoder synchronous detectors from the chroma signal delivered by the second detector. The R−Y, G−Y and B−Y signals, which can swing positive *and* negative, are shown respectively at (b), (c) and (d) in Fig. 2.

CRT Primary-colour Matrixing

Now assuming that the tube grids are clamped to a d.c. level a Y signal going increasingly negative progressively turns on more beam current at the three

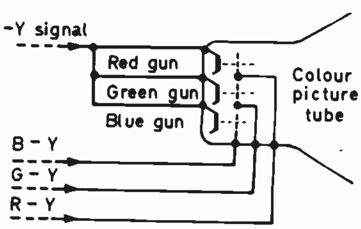


Fig. 1: Showing how with colour-difference drive the c.r.t. guns add the Y signal to the colour-difference signals to obtain the three primary-colour signals. −Y at the cathodes equals +Y at the grids.

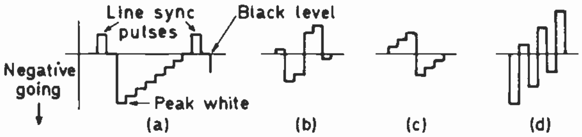


Fig. 2: Signal waveforms: (a) $-Y$, (b) $R-Y$, (c) $G-Y$ and (d) $B-Y$.

guns, resulting in an overall raster of increasing brightness. If a grid goes positive then the corresponding colour will become brighter but if it goes negative the colour intensity reduces. In other words since a negative-going signal at the cathode of a gun brightens the appropriate colour while a negative- or positive-going signal at the grid adds to or subtracts from this brightness each gun acts as a matrix neatly adding or subtracting the luminance and the appropriate colour-difference signal. This action of course restores the original primary-colour signal.

The signals shown in Fig. 2 correspond to the standard colour bars. Now it is possible though a bit time consuming to plot the four signals in Y and X axes scale and then add (a) to (b), (c) and (d) separately to discover what the primary-colour signals look like. A better way is to feed a standard colour-bar signal into a receiver and to view each primary-colour signal across its associated gun on the screen of an oscilloscope.

Excluding the horizontal steps at black level and peak white on the signal at (a) we find that there are six intermediate ones. These correspond to the yellow, cyan, green, magenta, red and blue colour bars displayed on the screen of the picture tube from left to right. The horizontal steps can be better seen in the Y signal oscillogram in Fig. 3 but here the electronics have caused the waveform to be displayed the opposite way round. Starting from the sync pulse top left we have the black step then we go down to peak white via the blue, red, magenta, green, cyan and yellow steps. The peak white step is not fully resolved in this oscillogram.

Primary-colour Signals

Each colour-difference signal—(b), (c) and (d) in Fig. 2—will also be seen to have a total of six horizontal steps. These correspond to the colour bars in the order of picture tube display. If we feel like it we can add or subtract—as they swing both positive and negative in polarity—the colour-difference signal steps and the luminance signal steps and in this way resolve the separate primary-colour signals.

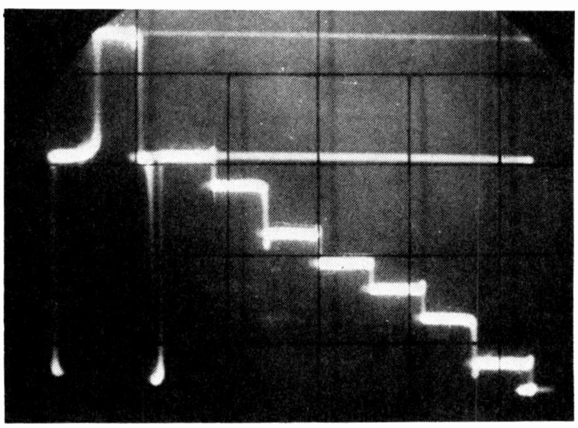


Fig. 3: Oscillogram of the Y signal (see text about the mirror image effect).

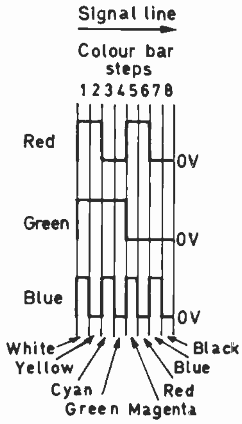


Fig. 4: When the $-Y$ signal—(a) in Fig. 2—is separately added to the colour-difference signals—(b), (c) and (d) in Fig. 2—we get the three primary-colour signals shown here (for the colour-bar test signal). The eight intervals or steps of the $-Y$ signal corresponding to the standard colour bars are identified here from white through to black across the screen from left to right.

The primary-colour signals will come out as drawn in Fig. 4, each of the same amplitude since the saturation of the standard colour bars is consistent and almost 100% over the colours. Eight steps, including the black and white ones, are identified making it possible to see exactly how the three beams combine to give the colours. Starting at the white step we see that all the guns are active, the three colours red, green and blue thus combining to produce white light. On the yellow step the red and green guns are active, red and green light producing in combination yellow light, the complementary of blue which is cut off on this bar. On the cyan step the blue and green guns are active, blue and green light producing cyan light the complementary of red which on this bar is cut off. On the green step only the green gun is active, producing a green display of primary colour. The same principles hold for the other colour bars through to black when of course all guns are cut off.

Colour pictures of course are produced by a wide range of variations in the strengths and combinations of the three primary colours over the screen area, the colour at each picture element being dictated by the colour-difference signals present. All the time the Y signal is giving all the detail in brightness and definition.

Colour-difference Channel

Now let us look at a block diagram of the colour-difference channels (Fig. 5). As with the luminance amplifier each channel consists of a preamplifier and an output stage. The preamplifier is commonly transistorised while in many receivers the output stage uses a valve. The synchronous detectors supply $R-Y$ and $B-Y$ signals only and as mentioned earlier the $G-Y$ signal must be created from these two signals. It is probably best to see how this is done by looking at a fairly typical colour-difference preamplifier

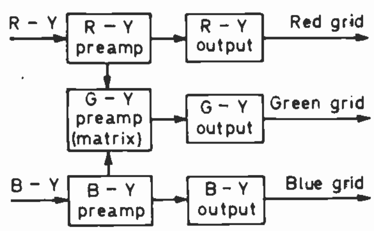


Fig. 5: Block diagram of a colour-difference pre-amplifier channel and $G-Y$ matrix.

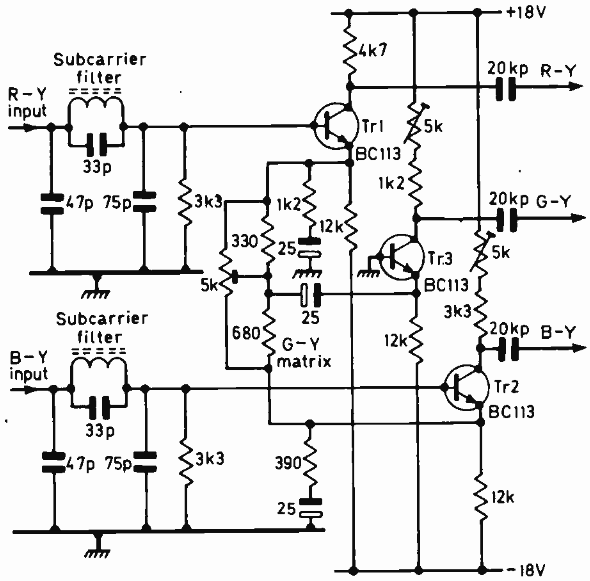


Fig. 6: Circuit of a representative colour-difference preamplifier channel with G—Y matrix.

circuit. In Fig. 6 Tr1 is the R−Y preamplifier, Tr2 the B−Y preamplifier and Tr3 the G−Y preamplifier. The bases of Tr1 and Tr2 receive the R−Y and B−Y signals respectively from the synchronous detectors via low-pass filter circuits which eliminate residual chroma signal.

From the point of view of the R−Y and B−Y signals transistors Tr1 and Tr2 are both in the common-emitter mode. However the emitters also deliver signals and it is these that are matrixed to produce the G−Y signal. The matrix network consists of the 5kΩ potentiometer and associated fixed resistors whose junction feeds signal via the 25μF electrolytic to the emitter of the G−Y preamplifier Tr3. This transistor is arranged in the common-base mode so that its output will be in phase with the R−Y and B−Y signals at the collectors of Tr1 and Tr2. Thus we have R−Y, G−Y and B−Y signals at the collectors of the three transistors and these are communicated to the output valves. Before we go on to this however let's see how the G−Y signal is produced by matrixing the R−Y and B−Y signals.

Recreating the G—Y Signal

It will be recalled that the Y signal is brewed from 0·3R + 0·59G + 0·11B. To simplify we'll round off the figures to 0·3R + 0·6G + 0·1B which is near enough for our purpose. R, G and B correspond to the red, green and blue primary-colour signals and the percentages are given in terms of decimals: thus 0·3R simply means 30% red signal.

If we subtract Y from both sides of Y = 0·3R + 0·6G + 0·1B we get 0 = 0·3(R−Y) + 0·6(G−Y) + 0·1(B−Y). We are interested in extracting G−Y so we can change the expression round (noting the change of signs) to give 0·6(G−Y) = 0·3(R−Y) − 0·1(B−Y). Now dividing this out by 0·6 (to get rid of the 0·6 in front of the (G−Y)) we end up with G−Y = 1/2(R−Y) − 1/6(B−Y). Sometimes the G−Y signal is expressed as 0·51(R−Y) − 0·19(B−Y), which is pretty well the same thing. If we are interested in obtaining a −(G−Y) signal then the expression is merely changed to 1/2(R−Y) + 1/6(B−Y).

However one looks at it therefore the matrix merely accepts R—Y and B—Y signals from the appropriate preamplifiers or associated circuitry and adds or subtracts one to or from the other in the approximate ratios of 1/2 and 1/6, whether addition or subtraction is carried out depending on the drive phase requirements. Some chassis, notably the Rank-Bush-Murphy single-standard one, use part of an integrated circuit for handling the G—Y matrixing.

In Fig. 6 the R—Y and B—Y signals are of the correct phase for matrixing at the emitters of transistors Tr1 and Tr2. The colour-difference output signals from the preamplifiers (Fig. 6) must be negative-going since we get a phase-reversal in each colour-difference output valve and require positive-going drive signals at the picture tube grids. Now since Tr3 is in the common-base mode the signal at its collector is of the same phase as the signal applied to its emitter, which means that the matrix must yield a −(G—Y) signal. The matrix thus adds the R—Y and B—Y signals in the ratios already mentioned.

Preamplifier Circuitry

The preamplifier transistors have gains such that 500mV peak-to-peak full-drive signals from the synchronous detectors are boosted to several volts for application to the grids of the colour-difference output valves. An equal-luminance drive system requires different drive voltages for the different colours. The preamplifier outputs can be adjusted to secure the required conditions by the preset resistors in the collectors of Tr2 and Tr3.

The correct form of the G—Y signal is obtained by regulating the 5kΩ preset potentiometer of the matrix circuit. There are various ways of making the adjustment for optimum results. One is by using the standard colour bars as transmitted or as produced by a colour-bar generator, the adjustment being made for the best green display. This can introduce complications however, especially when the drive presets are wrong. As this subject is somewhat involved and requires more knowledge of the colour-difference output stages we will leave it until next month.

Signal Weighting

One final point. The V and U chroma signals are "weighted", i.e. reduced in amplitude by set amounts, at the transmitter to avoid over-modulation. The correct levels are restored in the colour-difference preamplifier stages by presetting the gains of the R—Y and B—Y preamplifiers. This is the reason for the different values (1·2kΩ and 390Ω) of the a.c. coupled bias resistors in Tr1 and Tr2 emitter circuits.

CONTINUED NEXT MONTH

NEW FROM ANTEX

Antex have introduced a new soldering kit, the SK2, which consists of a 240V 15W miniature soldering iron fitted with $\frac{3}{16}$in. bit, two spare bits $\frac{5}{32}$in. and $\frac{3}{32}$in., coil of cored solder, heat sink, 1A fuse and their booklet *How to Solder*. The recommended price (UK) is £2·40. Antex, Mayflower House, Plymouth, Devon.

SERVICING television receivers

L. LAWRY-JOHNS

PHILIPS STYLE 70 SERIES—*cont.*

Field Timebase

Undoubtedly the weak link here is the PCL85 (PCL805). Whatever queer things are happening to the height and vertical stability (top or bottom compression, rolling, jittering or just no field scan at all—white line across the centre etc.) change this valve first. Try two, the first one may not be too good although it is new. Then check the bias resistors to make sure first that they are there and secondly that they are of the correct value. They are R443 and R444, both 560Ω.

Having checked the valve and the resistors observe the resulting picture. If it is sadly lacking height with severe bottom compression check C425 which may be open-circuit. Shunting another 100μF electrolytic across it is a quick check. This is easy of course.

What isn't so easy is when one is faced with top compression which does not respond to valve changing, resistor checks or adjustments. Check all the resistors and capacitors in the linearity circuit from L107 (transformer linearity winding) back to pin 9 of the PCL85 base. You may find a faulty one, usually C422 or R440.

There are however times when all is in order here but the top distortion remains. Yes, you're right, it's the transformer which has a faulty winding. A temporary part remedy is to hook up an 0·01μF (1kV) capacitor from the pin 6 end of the winding L106 to R437 and disconnect L107. This is not wholly satisfactory and a new transformer should be obtained as soon as possible.

Now here is another naughty one. If there is no field scan and all voltages appear to be in order (including pin 6 to prove the transformer) and a hum check to pin 9 does not open out the raster at all consider the possibility of an open-circuit in the field deflection coils.

Disconnect one side of the coils (so you don't read through the transformer secondary) and check the continuity with an ohmmeter. If there is no reading check the fine wire connections and then face up to fact that the coils are at fault and that a new set is called for. If coils are fitted which are not of the same make as the original don't be dismayed if the resulting picture is reversed with the writing appearing as though viewed through a mirror or (and!) upside down. We can't all get the connections right the first time.

Reduced Height

It is often the case that the height of the picture has reduced evenly top and bottom with the height control at its maximum. In this case check from pin 1 across resistor R446 which should record 1·5MΩ. This resistor tends to hide behind the height control but there is no reason why a replacement should not be wired on the print side if this is more convenient.

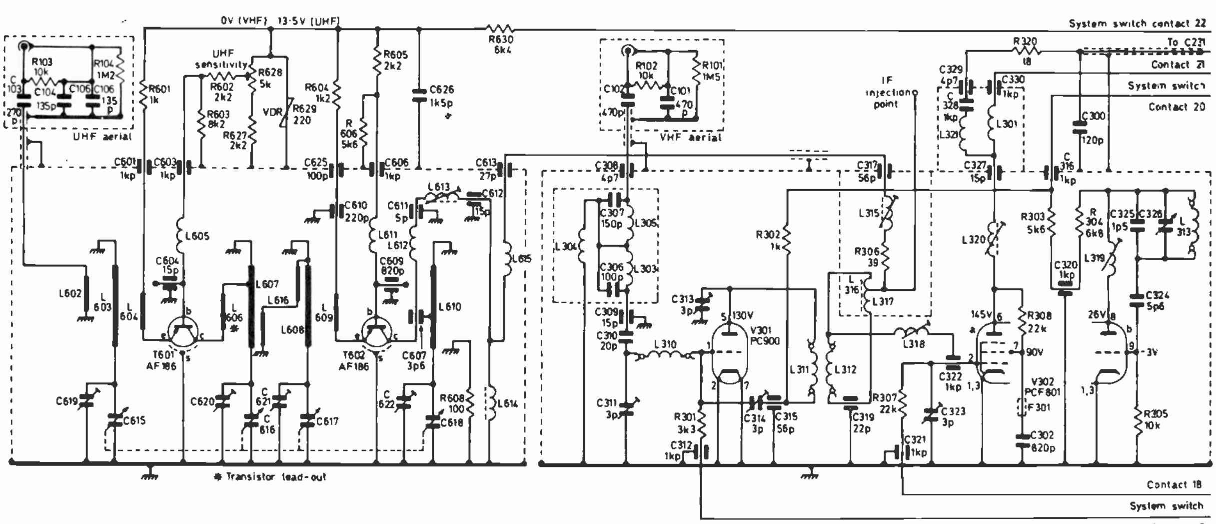


Fig. 2: Circuit diagram of the v.h.f. and u.h.f. tuner units.

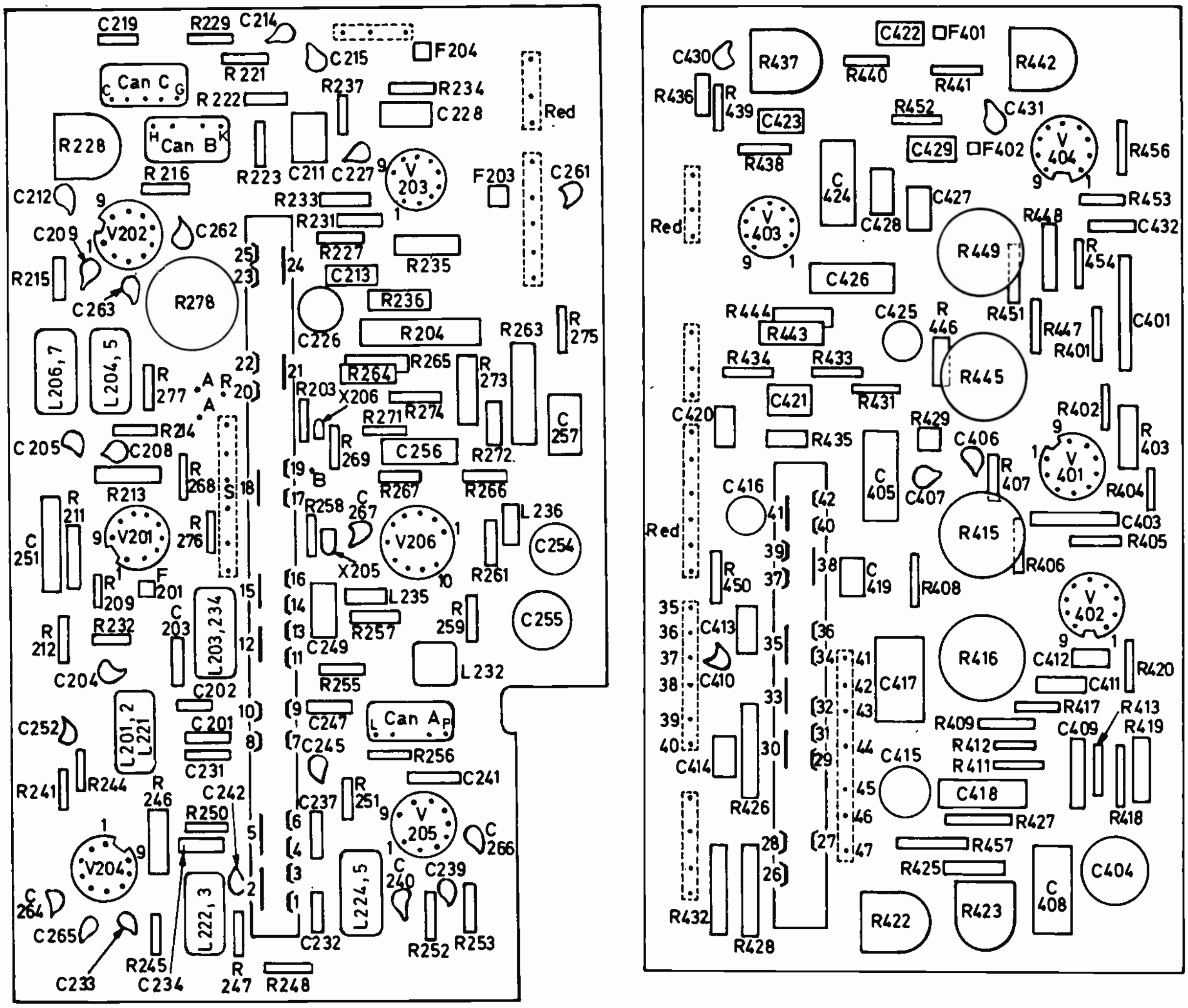


Fig. 3: Layout of the i.f. and timebase printed panels.

If this resistor is found to be in order check R428 and ensure that C414 has not shorted to the h.t. line.

Field Hold

Resistor R447 tends to change value on occasions resulting in the hold control being at the end of its travel. This one is in a fairly exposed position just to the right and between the height and hold controls. If this resistor is in order don't overlook the EF80 V404. We tend to regard these valves as everlasting but they can play tricks on occasions causing field hold troubles and indeed no scan at all. It is after all half the field oscillator. This all assumes that the PCL85 has been checked as we said at the outset.

Line Hold

Reverting to the line timebase a frequent source of trouble is a defective ECC82 valve. There are two of these. V402 is the oscillator and failure of this one results in complete stopping of the timebase and over-heating of the output valve. Loss of emission causes the line speed to rise and vary. This however is a symptom which could also be caused by a defective V401—as well as the more usual loss of line sync generally caused by this valve. By loss of sync we mean inability to lock the picture in a horizontal sense, the picture bending off and streaming away one moment then hovering about the screen and dissolving into lines the next.

Video Stage

If we appear to have spent a lot of time discussing timebase faults it is because this is where the majority of the trouble occurs. We now turn to the video and sync separator stages, coupling these together since they are closely related, both operations being carried out by the same valve V206 (PFL200).

Let's kick off here with a fault which has puzzled many. The symptoms are that the picture is perfectly normal when the contrast is kept down but as it is advanced to produce a more black-and-white picture the sync is impaired with line pulling and rolling and the detail is smeared. Quite clearly the operation of the sync separator is being upset by the contrast control and a look at the circuit will show that the contrast control which is at a positive potential is closely coupled to the sync separator control grid. This must not be allowed to become positive and is prevented from doing so by the diode X206 (BA115). If this diode is open-circuit it can no longer conduct and stop the line from becoming positive. The usual back-to-front reading can be made with an ohmmeter without disconnecting the diode: if the diode is open-circuit a high reading will be obtained both ways.

A similiar set of conditions arises when (if) C257 leaks but here the contrast control would not be the determining factor.

Excessive Brightness

If the picture is inclined to be overlight and lacking in correct contrast first check the PFL200. This tends to draw grid current—particularly on 625—and was a very common fault some time ago. It was recommended that the 625-line grid-coupling components be changed to reduce the chances of this happening. Capacitor C429 was changed to $0{\cdot}1\mu$F and the grid leak R258 reduced to around 220kΩ. Improvements in the design of the valve have rendered this change unnecessary of late and the trouble is now far less common. However the PFL200 is still the chief suspect when weak contrast and a light picture are the symptoms exhibited.

A weaker than normal picture, this time accompanied by poor sync, is often caused by a faulty electrolytic capacitor in the video stage. Both C255 (20μF) and C254 (250μF) should be checked, remembering that C255 is at h.t. potential. A 16μF or 32μF electrolytic rated at 275V or over can be used in this position.

DC Restorer

If on 625 lines the general background level seems to vary, giving dark scenes an overall grey effect, check the BA115 diode X205. This can become open-circuit in the same way as the diode in the contrast circuit.

IF Circuits

Generally the only trouble experienced in the vision and sound i.f. strips is when a short develops in a valve, particularly the EF184. This results in R256 burning out and tends to direct suspicion to the decoupler C241 which could well in fact be at fault. Usually however it is the EF184 which is responsible.

The other defects are of a more general nature—poor switch contact, dry-joints in or at the base of the i.f. coil cans etc.—but it is difficult to itemise or describe the precise nature of these symptoms.

The alignment of the 625 sound i.f. coils seems to alter as time goes by but great care must be used in altering the core settings and this should be avoided if possible.

Audio Stages

Now and again trouble is experienced in the PCL82 sound stage. Usually the valve itself runs into grid current and either damages the bias resistor R235 or causes the 100Ω h.t. feed resistor R236 to drop off. The value of the bias resistor is 470Ω and this value should be checked whenever it is found necessary to replace the PCL82. R234 may occasionally go high to cause distortion on a strong signal but it is usually the valve itself which causes all signals to be distorted.

Tuner Units

A transistorised tuner is used for u.h.f. reception and this rarely gives trouble except from a mechanical point of view. If signals are suddenly lost when a button is depressed and no other button will receive a signal it is likely that the teeth of the quadrant have become disengaged from the cog and the spring has pulled it completely away.

Variation of signal strength, fading etc. are most likely to be due to poor contact in the two-pin plug and socket which connects the output of the v.h.f. tuner to the i.f. panel left side. Moving this rubber-covered plug will quickly confirm this. Note that this is the output of the v.h.f. tuner but that this defect will be common to both standards. To explain this, the output of the u.h.f. tuner is taken to the v.h.f. tuner for further amplification and therefore the i.f. output of the v.h.f. tuner is common to both standards.

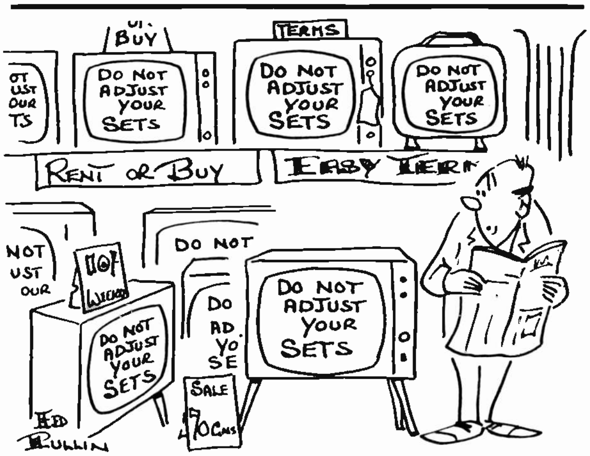
It is the v.h.f. tuner which is likely to give trouble if the set is used on 405 lines or relay. These troubles are mainly due to poor contact between the biscuit studs and the spring bank and should respond to the usual cleaning. The awkward one is when the fine tuner plunger shaft snaps making tuning of the v.h.f. signals impossible. This fault and the action necessary to renew the core were fully described in our December, 1968 and January, 1969 issues.

NEXT MONTH: GEC/SOBELL 2012/1012 SERIES

DYNAMIC TV PICTURES

In using the circuit featured in our April issue (page 254) it has been found that with the wide diversity of i.f. strips and video output valves encountered it is often necessary to increase the maximum signal output from the transistor stage.

This can be done by reducing the value of R2, the lowest value permissible being set by the voltage rating of the transistor used — BFY51 in the prototype. The BFY51 has a maximum collector voltage rating of 30V and with this transistor this is the limit which must not be exceeded when there is no signal input. Norman McLeod writes that he has used as low a value as 15kΩ for R2 and that with this value the voltage across C5, C6 and C7 never rises much above 10-20V except when changing channel. He also comments that a 47Ω grid stopper can be added in series with V1 grid but that this should not be necessary as the stage is quite stable.

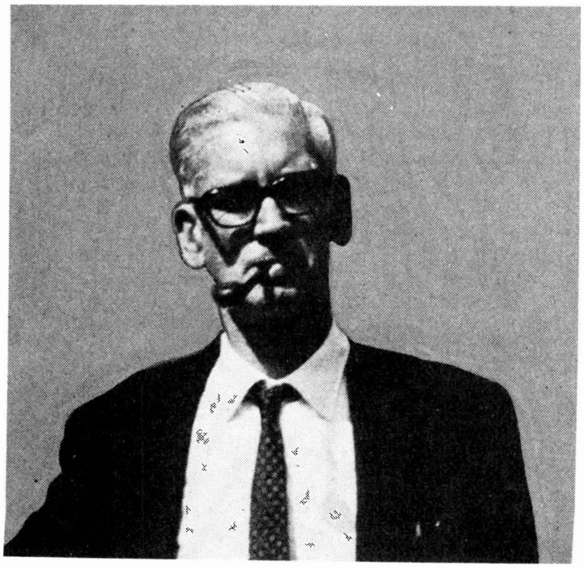


DX-TV

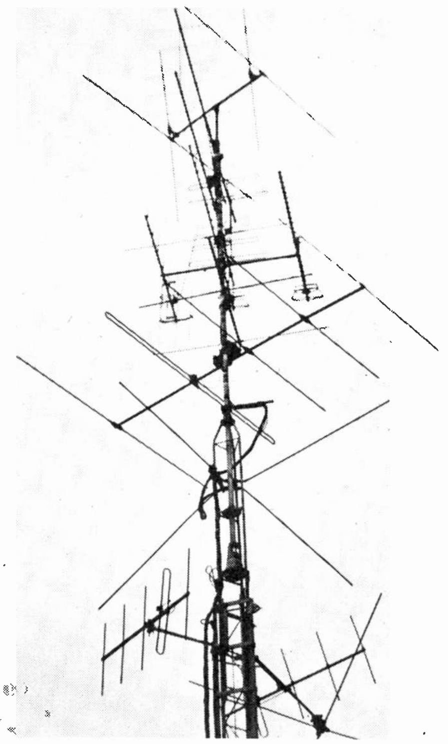
ON Sunday, 14th March, 1971, in the late afternoon, Charles Norman Rafarel passed peacefully away. His death came as a tremendous shock to us all, and he will be sadly missed. Those who knew Charlie personally—and he had many friends—will remember him as a very active and enthusiastic figure. His love of television dated back to the 1930s when he constructed a mechanical scanning televisor — which worked. Within recent years this machine was presented to the ORTF museum in Paris. Following the last war and in the 1950s his interests were in long-distance television reception, but towards the end of the fifties he became well known for his short-wave listening on the Amateur Bands.

His old love, television, caught up with him again in 1960 and at Parkstone near Poole he commenced experiments into the daily reception of Caen ch.F2. In due course a Bush Model TV62 was modified for DX use and his television horizons expanded. With the excellent Sporadic E conditions of those days he became absorbed with TV-DX. Undaunted by the collapse of two 50ft. masts, a lattice mast taller and magnificent indeed arose from his garden. Our photograph illustrates this and underlines his enthusiasm for this hobby.

In September, 1963, the first DX-TV column appeared in PRACTICAL TELEVISION, giving news and information for the followers of this passtime. It quickly grew in popularity with letters arriving from DXers all over the world. In May, 1966, Charlie was taken ill. Fortunately he recovered to almost his former self and at this time he moved to a different



Charles Norman Rafarel



Charles's aerial installation at Bournemouth.

location in Bournemouth. More recently he travelled widely in Europe including behind the Iron Curtain. Only months ago a colour receiver was installed at his home; he was as always anxious to experiment further with his great love, TV-DX.

Charlie did much to increase interest in TV-DX. He was always willing to help anyone in the various problems which are encountered. Today in Britain TV-DX is a thriving hobby and most of this is due to Charlie. He will be sadly missed.

We would like to extend to his wife Maisie our sincere condolences in her sad and unexpected loss.

Rest in peace, Charlie. *Roger.*

THE DX-TV column will resume its normal presentation and layout next month but under the new heading *Long-Distance Television.* A run-down of the month's reception will be given, news from enthusiasts and extracts from their letters will be maintained together with notes on the opening of new transmitters and other news from the Continent and indeed the world. It is hoped to commence within the next two months a new series of "data panels" showing the latest test patterns and cards from all countries in the European Broadcasting area. This will include North African and Middle Eastern countries. We will include the latter first to make the series as comprehensive as possible, secondly so that they are known in the event of one of these countries opening up in Band I and also for the assistance of our readers overseas. Very occasionally we will have as separate articles items on subjects such as preamplifiers, etc. suitable for DX use. For example towards the end of the year I hope to present information on wideband aerials for Band I.

We are always pleased to hear from readers of their reception and experiences and will be delighted to assist with their problems be they of programme identification or queries about equipment and aerial installations. If news comes to hand of changes in test cards, transmitters and so on please let us know in order that we can inform other enthusiasts. For example Doug Bowers of Saltash writes to advise us of changes with the Norwegian test card. We will also be pleased to hear from our many overseas readers: let us know about your reception—especially if unusual types of propagation are experienced.

At the time of his death Charlie was working on a pamphlet giving information on various aspects of DX-TV. I intend to continue and finish this project but there is likely to be some delay. So please understand and be patient whilst certain problems at this end are sorted out.

MARCH REPORT

Now a quick run over the news of the last month —March. It was generally rather quiet with the usual short bursts of m.s. and Sp.E. On the 3rd, 4th and 5th there were sustained Sp.E signals into Eastern Europe; rather surprisingly these openings occurred at around the same local time each day—at 1730-1830.

3/3/71 USSR R1.
4/3/71 Unidentified R1.
5/3/71 USSR R1, Czechoslovakia R1, Unidentified R2.

The 12th and 13th showed an improvement in Trops with various Northern French stations and NOS Holland E4.

At around this time of the year Sp.E tends to increase compared with the winter months' quiet conditions. During the first half of the 1960s a good Sp.E season started typically with a mid-April opening and then the real season opening towards the middle of May. During recent years however these April openings have been lacking and the seasons themselves have left much to be desired. By the time you read this we should be in a position to know if we are to have a good season. Certainly with the falling sunspot figures it would appear that our chances are better. Let's hope so.

NEWS IN BRIEF

New transmitters in operation:

Belgium: Wavre ch. 25, 1000kW horizontal (BRT).
Wavre ch. 28, 1000kW horizontal (RTB).

Iceland: Hafell ch. E7, 440kW horizontal.
(This might just be possible for Scottish DXers via Trops and during an Aurora.)

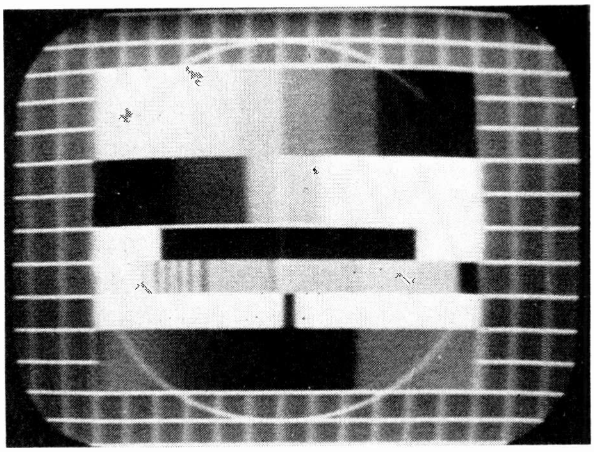
Portugal: Porto ch. 41, 300kW horizontal (RTP 2).
Montejunto ch. 46, 200kW horizontal (RTP 2).

Sweden: Kiruna ch. 29, 1000kW horizontal.
Gaellivare ch. 33, 100kW horizontal.
Aange ch. 42, 250kW horizontal.
Lycksele ch. 45, 1000kW horizontal.
Sollefteaa ch. 46, 1000kW horizontal.
All 2nd chain.

The sound e.r.p. of all Swedish transmitters was reduced from 20% to 10% from 15th November, 1970.

Finland: We have heard that Finland Helsinki ch. 24 is now off the air. It was apparently used for training engineering staff and closed 31st December last.
Jyvaskyla ch. 25 is now transmitting, e.r.p. not known at present.

Our good friend Sepp J. Pirhonen in Finland has sent us useful information. Apparently the test card has changed to a new electronic type and is radiated daily at 0700-1200 GMT with the exceptions of Tuesday, Thursday and Saturday. On these days between 1000-1145 various patterns and slides are shown. The new card is similar to that shown here except that it carries the identification YLEISRADIO in white letters in the central black rectangle and that the TV2 network *omits* the circle. Usually the TV2 E2 transmitter at Tampere is off the air from 1st June to approximately mid-August and we must I think assume that the same will happen this year. This still leaves a number of TV1 transmitters in Band I fortunately. (Note that W. Germany also uses variations of this card.)



The electronic test card now used by Finland.

Charlie's last column gave interesting details of Albania. One slight correction however: the transmissions are being received on R6 and are manifesting themselves as interference on the local RAI transmitter. If anyone has information on Albania we would appreciate hearing about any TV activity there.

Unfortunately the sunspot details from the Swiss Solar Observatory were held up in the postal dispute and are only just beginning to trickle through. The January list gave smoothed predicted counts for May of 75, June 73 and July 71. Being somewhat below last year's count at this time this augurs well for the forthcoming season.

If space next month permits I will give a brief survey of F2 activity over the past winter, both of European and reception farther afield.

Roger Bunney.

POINTS FROM THE POST

Reader D. Fowler of 18A Udney Park Road, Teddington, Middx, will be emigrating at the end of the year and wishes to dispose of the "625-line Receiver for the Constructor" he has built and also back copies of the magazine. Anyone interested should send him a stamped, addressed envelope for details.

Another reader has bought a Channel TV waveform generator, type T2, serial No. 43490. The original firm is no longer in business and he requires a circuit diagram. Let us know if you can help.

R. L. Smith of 8 Hillside, Oswestry, Salop, wants to contact anyone who has made up a field sequential colour converter system.

YOUR PROBLEMS SOLVED

★ Requests for advice in dealing with servicing problems must be accompanied by a 10p postal order (made out to IPC Magazines Ltd.) the query coupon from page 378 and a stamped, addressed envelope. We can deal with only one query at a time. We regret that we cannot supply service sheets or answer queries over the telephone.

PHILCO 5600

Two resistors in the circuit near the PCL84 video output valve are badly overheating. These are numbered R76 and R77 on the board but unfortunately I cannot obtain a service sheet for this model—G. Lee (Colne).

This model was fitted with the same chassis as the Ferguson 3600 group of models. R76, the stabilising resistor connected between screen and cathode, is 47 kΩ while R77 picture quality control in the cathode circuit is 100Ω. The PCL84 may need replacement and we suggest you check the voltages in this stage, cathode 6V screen grid 175V and anode 110V. If the valve has to be replaced you might also find it necessary to replace the OA70 vision detector diode W3 inside can L36-L37.

INVICTA 7007

The picture is a narrow strip about 2in. wide across the centre of the screen. All manual controls have been adjusted without improving matters. The field timebase printed panel is mounted at the top right-hand side close to the e.h.t. connection on the tube and the capacitors on it are all blackened. A 16+16+16μF capacitor close to the panel appears to be faulty with a white discharge to chassis.—E. Hopecraft (Swansea).

The appearance of the paper type capacitors is quite normal and the fact that they are blackened does not affect their efficiency. You may find if there is a fault present that one or two of them are leaky. However you should first replace the PCL85 field timebase valve which is the principal offender here. Then check the voltages at its base. The weeping electrolytics you mention should be replaced at an early date.

HMV 2635

This receiver has poor line linearity when switched to 625 lines with compression at the centre of the test card. On 405 lines the field and line linearity are good. All timbase valves have been renewed.—A. Boyd (Cheam).

Check C98 (0·1μF) the S-correction capacitor on 625 lines. This however may not be at fault and the boost reservoir capacitor C101 (0·22μF) should also be checked together with the adjustment of the line drive control.

MARCONIPHONE 4609

The picture has developed a constant downwards roll which cannot be overcome by adjusting the vertical hold control which is over to its full range. The field timebase valve has been replaced without improving matters. Otherwise the picture is normal. —F. Stephens (Evesham).

Check the value of the 22kΩ resistor R138 behind the 6-30L2 (V7) valve at the front centre of the lower deck. This resistor, the anode load of one section of the field multivibrator, should be replaced with a 22kΩ 1W type. Also check the resistors—R144, R116—associated with the hold control.

GEC 2019

The fault is line pulling to the left on whites. The vision definition is impaired suggesting that the trouble lies in the video stage rather than the sync separator. The vertical and horizontal locking is very good even with the contrast turned down to minimum. The EF183, EF184 and PFL200 have been renewed and all voltages seem to be correct.—L. Bradshaw (Rugby).

Pulling on whites is not a fault which normally affects these models so we would suggest you first look for some possible external cause of the trouble such as ghosting from the aerial. If everything seems OK check the sync coupling capacitor C104 (0·22μF).

FERGUSON FR19

For about an hour after switching on the picture fades. It can be brought to full brilliance by changing the channel switch to another position and back again but fades away again after a minute or two. The tuner switch contacts have been cleaned and the PCC89 replaced with no improvement.—F. Bilson (Grantham).

It is not easy to localise a fault of this type. Although the tuner contacts have been cleaned the trouble could still be in this area due, for example, to a poor contact on a valveholder or a dry-joint. There is no reason why the fault should be confined to the tuner, however, and it would perhaps be useful to probe around the i.f. panel valves and capacitors to see if the fault can be provoked by a disturbance in a particular area.

EKCO T419

The picture exhibits a plastic effect, the whites having very little definition unless turned low which gives faces a glassy look. The video amplifier valve PCL84 has been replaced but very little improvement obtained.—F. Dudley (Birmingham).

You will probably find that the screen and cathode resistors in the video amplifier stage have changed value. These are R26 5·6kΩ and R28 220Ω. If these are faulty check also the control grid stopper resistor R27 (330Ω) and the vision detector diode V7 (CG64H).

MARCONIPHONE 4621

There is cogging on the picture the extent depending on the setting—which is on the critical side—of the line hold control. One of the flywheel sync diodes has been replaced with another diode which is not an exact replacement—making the cogging worse. —T. Marshall (Solihull).

It is essential that the flywheel discriminator diodes are replaced with the correct type—either the original type or one with precisely the same ratings. Also both diodes must be replaced at the same time as the circuit must be balanced. The latest replacements quoted are type BA153/1 but you should find the BA144 a suitable replacement which is easier to obtain.

GEC BT318

The raster is present but there is no sound or vision —only a rumbling sound from the speaker. The h.t. rectifier, tuner and audio output valves have been replaced.—J. Dance (Bury).

Examine the rear of the coaxial socket on the i.f. board into which the output plug from the tuner connects: the track from the socket to the first i.f. transformer often cracks. Also check the common i.f. stage valve and voltages.

PYE 40F

The set works well on 405 lines but when switched to 625 the width decreases by about 1½in. on each side even with the width control set to maximum. The valves in the line timebase have been changed without improving matters and all voltages are correct. In all other respects the u.h.f. picture is good. —R. Irwin (Windsor).

Check the coupling capacitor C111 (0·47μF) to the line output valve and the width stabilisation v.d.r. R148 in its grid circuit.

BUSH TV171

When the set is first switched on the field slips from the top to the bottom. On adjusting the hold control the picture will lock for a few seconds then starts to slip again. Then a sequence sets up, the picture locking for a few seconds and then slipping for a few seconds and so on. The PCL85 and PFL200 have been replaced, also SR302.—G. Thornton (Birmingham).

The trouble could be caused by a faulty electrolytic —check the capacitor C46 (100μF) which smooths the h.t. supply to the field timebase. Also check the electrolytics associated with the PFL200—8 and 10μF. Then check the front-to-back ratio of the interlace diode 3MR3.

ALBA T655

There is trouble in the line output and e.h.t. supplies. The boost voltage was low but this was put right by fitting a new boost reservoir capacitor. Now if either the contrast or brilliance are increased the boost voltage drops to about 300V, the width decreases, the line whistle changes and the e.h.t. rectifier heater goes dead and then glows green. The line output valve screen voltage is 160V instead of 140V but the screen resistor has been checked and is correct. The c.r.t. cathode voltage drops to about 50V when receiving a signal. There is also severe arcing from the EY86 top cap to chassis which can only be stopped by increasing the contrast or brilliance control settings. The valves in the line output stage and many other components have been replaced and a new c.r.t. tried.—G. Carter (Bournemouth).

The basic cause of the boost and e.h.t. voltages dropping as the brilliance is advanced is lack of drive to the line output valve, so the line oscillator stage (V10 ECL80) must be thoroughly checked including the coupling capacitor (C47 0·01μF) to the line output valve. Once this trouble has been cleared you will experience worse discharge from the EY86. This is a matter of improving the insulation. Try a good application of silicon rubber (Radiospares tube) around the discharge point.

DEFIANT 901

The picture is very pale on ITV and there is so much snow on BBC-1 that it cannot be discerned properly. On ITV the picture comes on bright with a click and half-an-hour later it goes pale again.— G. Dorset (Frome).

The trouble, or at least some of it, appears to be in the system switching on the left side panel. Check the contacts and clean. Also check the PFL200 video amplifier valve and ensure that there is good contact with its valveholder. Check the associated components by disturbing gently to locate any poor connections. Check the tuner unit contacts and ensure that the BBC-1 aerial is up to specification.

PYE V210

The picture has very little contrast and is reduced in width by 1 to 1½ inches on each side of the screen. The PL81 is glowing red hot but a replacement makes no difference. I checked the grid circuit in case it was running unbiased and found approximately −24V here. The other valves in the line output stage have been replaced but the fault persists.—J. Manningfield (Chichester).

The cause of the trouble is the 2·2kΩ feed resistor to the screen of the line output valve becoming low in value. Replace it using a 5W wirewound type.

PETO SCOTT 235

There is bad foldover at the top of the picture and the pulse and bar test pattern appears almost halfway down the screen. Several valves including the field timebase one have been replaced in order to improve the performance of the set.—P. Wainwright (Bromley).

This model was particularly prone to linearity troubles. The linearity controls and the capacitors in the linearity network will have to be carefully checked and any defective ones replaced.

BAIRD 663

There is an intermittent fault on 625 lines. When the set is first switched on the set works perfectly for five minutes or so. The hold then slips and the picture is reduced in height to four-five inches. This is accompanied by a loud hum on sound and at times motor-boating. If the set is left on it will right itself in time while if the contrast is turned down to minmum the fault disappears completely. On 405 the hold does not slip but the hum is present on sound. All valves have been checked.—J. Read (London SW16).

There appear to be two faults on the receiver. The first is almost certainly the failure of one section of the h.t. smoothing and this should be corrected first. Check C190 and C191. The second problem is probably the common one on this range of models, dirty or loose contacts on the rather complex standards changeover system. If yours is an early model this is achieved with a cable linkage the movement of which should be checked.

TEST CARD COLOUR BAR

Can you tell me how the colour bars on the top of test card F can be shifted down the screen without putting out the convergence?—R. George (Bromwich).

It is usually possible to move the raster down enough to resolve the colour bar by means of the vertical shift control. This will not affect the convergence at all. Alternatively you could shrink the height. This will misconverge the raster slightly but immediate correction will be obtained when the height is expanded again.

PYE 11U

There is lack of width, marginal on 405 but ¼in. each side on 625, and this is not affected by the "set e.h.t." controls. Also the raster is distorted at the top left of the picture with a flyback line present there while the picture tends to be trapezium shaped. The line oscillator and output valves have been changed.—A. Richards (Clerkenwell).

For the width fault check the line feed capacitor C87 (10kpF) which may have changed value. Your distorted raster is due to the deflection coils themselves. Try moving the raster correction magnets M3 and M4 if they are of the adjustable type.

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TELEVISION, JUNE, 1971

TEST CASE

102

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

? *A Ferguson 2000 series all-transistor receiver was investigated by the field engineer for apparent grey-scale mistracking. On the colour bars or the contrast wedges of Test Card F with the colour control turned right down the white display was normal, but as the bars or wedges progressed through the greys to black there appeared increasing evidence of a green hue, becoming bright green on the final grey bar or wedge, with black normal. The field engineer found that the normal grey-scale tracking procedure failed to cure the trouble and as a consequence suspected a fault in the tracking of the green gun.*

With the receiver in the workshop it was discovered that the fault symptom was somewhat intermittent and that it was hardly apparent at all on colour pictures (i.e. when the colour control was advanced). On a monochrome picture the green had a "flaring" tendency, not unlike the symptom on monochrome sets with poor low-frequency response characteristics. The degree of green "flaring" could be altered significantly by adjustment to the contrast and brightness controls.

Although the workshop engineer realised that a defective picture tube might well cause such a symptom, he decided first to investigate elsewhere in the set and as it turned out his efforts were well rewarded. What part of the set would be most likely to produce a symptom of this kind? See next month's TELEVISION *for the solution to this fault and a further item in the Test Case series.*

SOLUTION TO TEST CASE 101

Page 331 (last month)

It was found that the unusual symptom of the picture height altering with adjustment of the brightness control was accompanied by a change in the boost voltage which fell as the brightness control was advanced and the picture height reduced. The boost voltage feed to the c.r.t. first anode was disconnected and it was then found that the potential at the feed point remained constant over the entire range of brightness control.

This was a fair indication of insulation breakdown between the tube gun electrodes. Subsequent tests proved that when the tube was warm a low-resistance path developed between the first anode and control grid. This resulted in the boost potential being bypassed progressively as the brightness control was turned down, the height being affected since the field timebase generator is energised from the boost potential.

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30FL1	·62	ECC83	·23	PC900	·35	U26	·59
30FL14	·70	ECF80	·28	PCC84	·31	U191	·60
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