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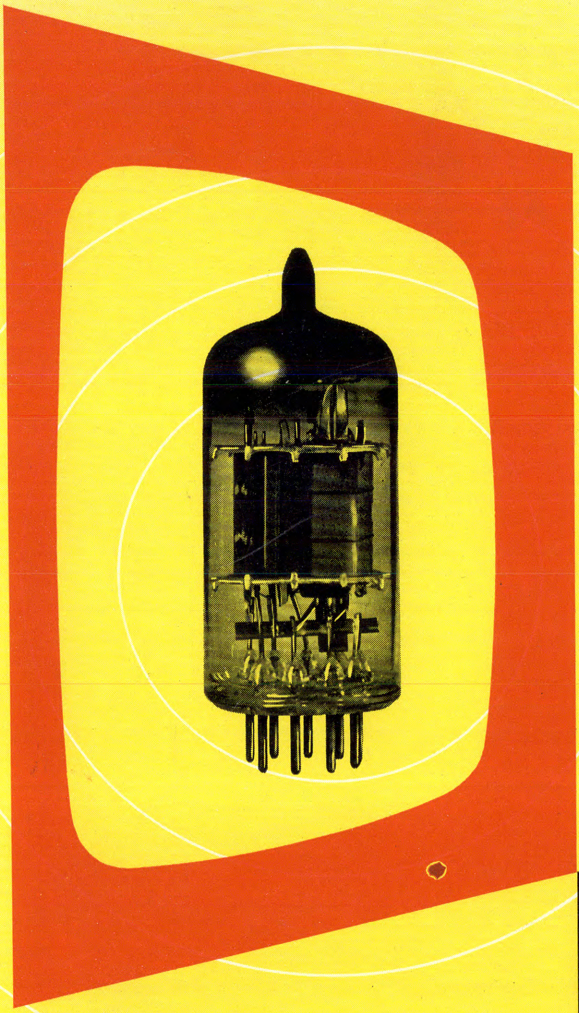

PCF 80

multipurpose triode-pentode for television

PHILIPS ELECTRONIC TUBE DIVISION

PREFACE

Combined triode-pentode systems have been available for a long time, but it is only fairly recently that the need for good screening between sections has led to the "side-by-side" construction. A typical example is the PCF 80, which was designed for use as a combined oscillator and mixer in TV tuners for the V.H.F. range. It was to be expected that a tube like the PCF 80 would also prove to be very attractive for a number of other applications in TV receivers, and in many receivers even three or four PCF 80's are employed in addition to the PCF 80 used in the tuner.

The object of this Bulletin is to give a detailed account of some of the uses to which the PCF 80 can be put, and to give useful hints for other functions.

Every type of tube obviously has its limitations and so has the PCF 80. Mention is therefore also made of those functions or combinations of functions for which the PCF 80 cannot be recommended.

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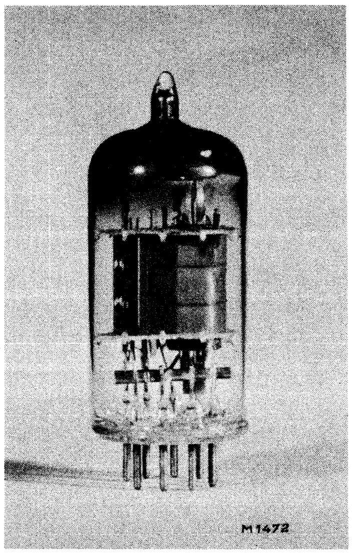
INTRODUCTION



The PCF 80 (see Fig.1) is a triode-pentode in Noval technique with a 0.3 A heater and has specially been developed for use as an oscillator-mixer in television tuners.

The two tube sections have separate cathodes, and the triode and pentode sections are mutually screened. These features render the tube also suitable for a number of other applications in TV receivers. The discussion of the various applications of the PCF 80 in TV receivers will be preceded by a detailed description of this tube.

Fig.1. Triode-pentode PCF 80.

CONSTRUCTIONAL

An interesting detail of the PCF 80 is the construction of the cathodes of both the triode and the pentode sections. Fig.2*a* shows the cross-section of a conventional triode, the cross-section of the cathode being circular, whilst that of the grid is oval, and that of the anode is almost rectangular. A disadvantage of this construction is that the distance between the cathode and the grid differs from point to point. As a result, the amplification factor is highest in the region where the distance is smallest, and vice versa. The tube will, therefore, have to some extent a variable-mu characteristic, its mutual conductance varying considerably with the grid bias. For a given effective mutual conductance the mutual conductance at zero grid bias must then be high, and this can only be achieved by keeping the minimum spacing between cathode and grid extremely small (65 μ) [1]. Needless to say, that this renders manufacture very critical.

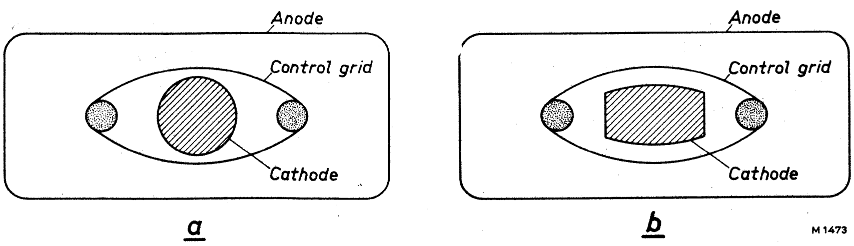


Fig.2. (*a*) Cross-section of an arbitrary triode containing a circular triode.
(*b*) Cross-section of a triode containing an oval cathode.

[1]) This figure is calculated for a circular cathode having the same emissive power as the cathode of the triode of the PCF 80. The distance cathode-grid is made so small that the effective mutual conductance of the calculated tube and that of an actual PCF 80 triode are equal.

This has been avoided by giving the two sections of the PCF 80 a cathode similar to that shown in Fig.2*b*. Both the cathode and the grid have an oval cross-section. This construction has many advantages compared to that shown in Fig.2*a*. The distance between cathode and grid is substantially constant so that the amplification factor is almost uniform over the circumference of the cathode. This results in the mutual conductance being less dependent on the grid bias. At the same time the construction is simplified, since the minimum distance between the cathode and grid need not be so small (110 μ). Because of its simpler manufacture, the risk of premature breakdowns and spread in production is greatly reduced.

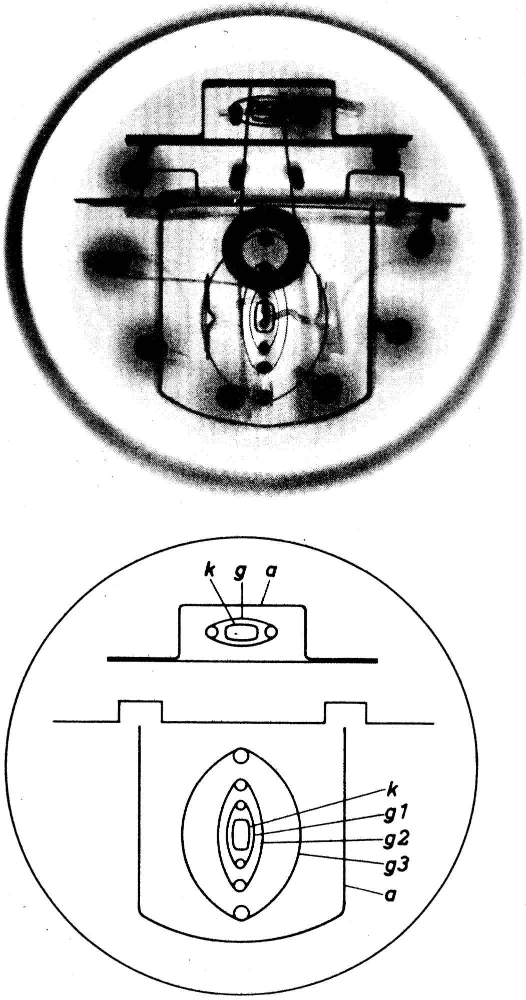


Fig.3. Enlarged radiograph of the PCF 80 giving a view from above, clearly showing the triode and pentode sections.

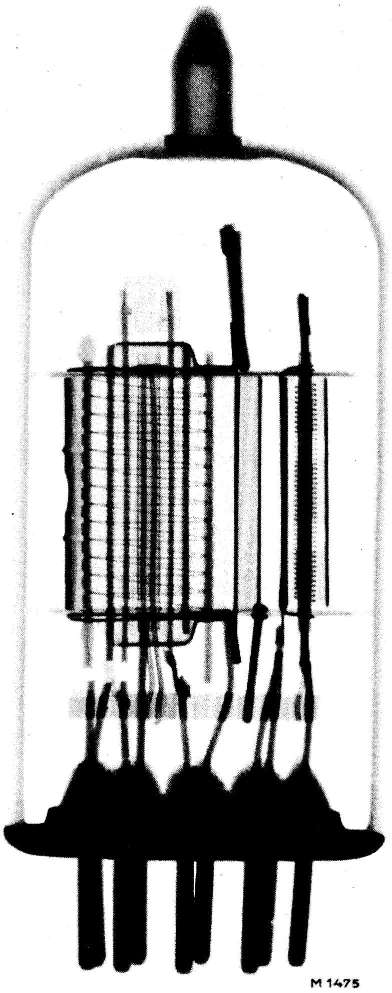


Fig.4. Radiograph of the PCF 80 (twice actual size).

The radiographs reproduced in the Figs 3 and 4 clearly show the construction of the electrode system. It is seen that the triode and the pentode sections are separated by a screen, which considerably reduces the cross-capacitances. The screen is connected internally to the cathode of the pentode section.

ELECTRICAL

Due to its high slope and low amplification factor the triode section of the PCF 80 is a reliable oscillator, even at the oscillator frequencies for band III (174 - 216 Mc/s). Since the mixer section is a pentode, it has the obvious advantage of the feedback between the anode and the control grid being low, which is important for receivers incorporating band I and operating with a high intermediate frequency (35-40 Mc/s), or with double-tuned I.F. transformers with staggered damping. Due to its low input damping the preceding R.F. amplifier can provide a high gain.

The pentode and the triode sections are mounted next to each other as a result of which the internal leads of both sections could be kept short, which renders this tube particularly suitable for operation on frequencies up to 300 Mc/s.

The mutual conductance of the pentode, under normal working conditions, is 6.2 mA/V, and that of the triode is 5 mA/V. Due to the high values of the mutual conductance and due to the fact that the pentode and the triode sections have separate cathodes, the PCF 80 is suitable for a large number of other applications in television receivers. In this respect the tube is superior to the ECL 80 used up till now, as the mutual conductances of this tube are considerably lower and the pentode and triode sections have a common cathode.

The PCF 80 is developed for R.F. applications, and the tube has straight characteristics, whereas the ECL 80 is a typical A.F. tube. The differences between these two types of tubes cause that - apart from A.F. sound applications - the PCF 80 is suitable for a larger number of applications than the ECL 80.

However, the number of applications of the PCF 80 is obviously limited, and it will be necessary to investigate the requirements which are to be made to the tube in a special application.

If the triode and the pentode are to perform strictly separate functions, the internal coupling between the two sections is the factor which decides whether the combination is advisable or not, especially if the triode and the pentode are to operate at widely different signal levels.

The coupling between the sections has three components: the leakage paths between the cross capacitances, the electronic coupling and the cross capacitances. The leakage paths are negligible in all but the most critical combinations of functions. The electronic coupling is counteracted by the screen between both sections. The screen diminishes this kind of coupling to a practically negligible value for the applications dealt with in this Bulletin. The most important is the capacitive coupling between the anode of the triode and the first grid of the pentode, which has a maximum value of 0.16 pF. To this must be added the tube-holder capacitance, viz. 0.15 pF to 0.4 pF according to the type. The total capacitance is shunted by the leakage paths of the tube and the tube-holder.

In certain applications hum may become a limiting factor. In these cases, limits of sensitivity and the recommended position in the heater chain must be observed. For application as an oscillator-mixer in receivers with split-carrier AM sound or split-carrier FM sound the a.c. heater-to-cathode voltage should not exceed 55 V, and the tube should be as low as possible, in the heater chain. The values of the interelectrode capacitances also set a limit to the number of possible applications. A survey of the applications for which the PCF 80 might be eligible is given below, and the possibility of the use of the tube in these applications is discussed.

THE PCF 80 IN VIDEO AND PULSE CIRCUITS

VIDEO AMPLIFIER

The tube can be used advantageously in video output circuits in which the PL 83 was previously used. In the circuit in which the pentode section of the PCF 80 is used as a video amplifier and the triode section as a subsequent cathode follower, the high mutual conductance of the pentode section and the better frequency response characteristic which can be obtained due to the d.c. coupled cathode follower output, make this tube preferable to the PL 83.

In video circuits attention must be paid to the value of the cathode resistor and the heater-to-cathode voltage with respect to hum. Detailed data of the maximum permissible values of the cathode resistor and the heater-to-cathode voltage are given with the description of a d.c. coupled video amplifier stage.

KEYED A.G.C. AND NOISE INVERTER CIRCUITS

The way in which the contrast control is realised in a d.c.coupled video amplifier to be described later, requires a special A.G.C. circuit to operate in combination with the video amplifier. For that purpose the triode of the PCF 80 can be used; the pentode section remains then available for other pulse circuits. A description will be given of the triode section of the PCF 80 in a keyed A.G.C. circuit, the pentode section being used as a noise inverter.

SYNC SEPARATOR, PHASE SPLITTER AND FRAME PULSE CLIPPER

The pentode section of the PCF 80 has a small grid base and a high slope, which renders the tube suitable for separating the synchronisation pulses from the composite video signal. A description will be given of a circuit in which the pentode section of a PCF 80 operates as a sync separator, whilst the triode section is used as a phase splitter for flywheel circuits. It is also possible to use the triode section as a frame pulse clipper.

LINE TIMEBASE GENERATOR

A considerable improvement in the line sawtooth generator stage can very well be obtained in replacing the conventional line timebase multivibrator by a combination of a sine oscillator operating as a line sawtooth generator, and a reactance control tube. The frequency stability of the latter circuit has proved to be much better than that of a multivibrator. A description of the sine oscillator and the reactance tube will be given in this Bulletin.

When, nevertheless, a line timebase multivibrator is used, it is advisable to insert a tuning circuit to improve the stabilisation of the natural frequency of the multivibrator. No interference effects from stray coupling are experienced when either section of the tube is used for sync separation or clipping, but, if the pentode is used in the multivibrator circuit while the triode of the same tube is operating as a frame timebase blocking oscillator, some interaction between line and frame scanning occurs. This stray coupling may arise between the connecting leads of the tube-holder rather than within the tube itself. In any case, in view of the large pulse amplitudes existing in each of these applications, this combination of functions in a single PCF 80 is not advisable.

In all varieties of a cathode-coupled multivibrator circuit, hum must be counteracted by the limitation of the heater-to-cathode voltages to less than 20 V. The PCF 80 should preferably be the first tube in the heater chain, and pin 5 should be earthed. Precautions must also be taken in circuits feeding into the multivibrator.

LINE TIMEBASE COINCIDENCE DETECTOR

Flywheel synchronisation gives considerable improvement over direct methods of synchronisation in fringe-area receivers operating under unfavourable conditions. For this method of synchronisation a coincidence detector is necessary to ensure that the line timebase oscillator is controlled only by pulses received during correctly timed flyback.

The flywheel synchronisation can be achieved by a circuit with two diodes as given in Fig.30, but also the pentode section of a PCF 80 can be used with the two coincident pulses applied to the control and screen grids (the suppressor grid is internally strapped to the cathode). When the screen grid is used in this way a separate pulse amplifier is required; for this purpose the triode section can be used.

No undesirable stray-coupling effects are observed when the triode or pentode sections of the coincidence detector tube are exchanged with those of PCF 80's in the line scanning applications described in this article.

RESTRICTIONS ON USE IN SCANNING APPLICATIONS

The PCF 80 is satisfactory for operation in the video and pulse circuits mentioned above. Loss of frame interlace, caused by cross capacitances in the tube holder and wiring but not necessarily within the tube, may occur in the combination of frame and line timebase functions in one PCF 80. Only the most elaborate screening and wiring precautions would remove this risk, and this combination is therefore not recommended.

THE PCF 80 IN I.F. AMPLIFIER CIRCUITS

The capacitance between the anode and the control grid of the pentode section of the PCF 80 is fairly large ($C_{ag1} < 25$ mpF) compared to that of pentodes especially designed for use in I.F. amplifiers. The tube is, therefore, not suited to be used in I.F. sound circuits, as the stability of the amplifier may then be upset due to feedback from the anode to the control grid.

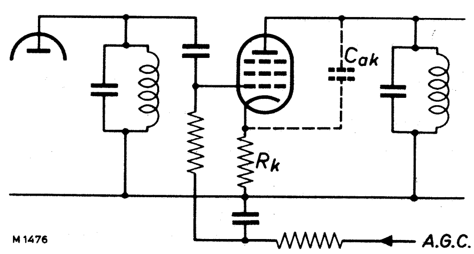


Fig.5. Basic diagram of an A.G.C. controlled vision I.F. stage.

In Fig.5 a diagram is given of an A.G.C. controlled vision I.F. stage. The variations in the input capacitance of the amplifier tube caused by variations in the A.G.C. voltage are compensated by means of an unbypassed cathode resistor R_k. This diagram reveals that the value of the capacitance between the anode and the cathode of the amplifier tube in such a stage must be small, otherwise the current circulating through the tuned anode circuit will partly flow via C_{ak} and R_k. In that case the compensation will be deteriorated and the stability of the amplifier may even be upset.

The anode-to-cathode capacitance of the PCF 80 is rather high (C_{aPkP} = 3.2 pF) due to the fact that the screen between both sections is connected to the cathode of the pentode. The tube is, therefore, not recommended for use in A.G.C. controlled vision I.F. stages. Stages not controlled by the A.G.C. voltage may be provided with the pentode section of the PCF 80 as the cathode resistor is then bypassed or the cathode is even connected to earth. The fairly high value of the capacitance C_{ag1} between the anode and the control grid of the PCF 80, however, makes an EF 80 tube preferable for all I.F. purposes.

It is clear that the triode section of the PCF 80 can be used neither in a sound I.F. amplifier nor in a vision I.F. amplifier due to its high anode-to-grid capacitance (C_{ag} = 1.5 pF).

THE PCF 80 IN AUDIO CIRCUITS

In general the cross-coupling from the pentode to the triode will not be so great as that from the triode to the pentode, since the capacitances from the pentode anode to the triode grid are less than 0.02 pF. The leakage paths, however, will make it impossible to use the triode at low audio levels when the pentode is used in a video or timebase circuit. Conversely, the combined effect of the cross capacitances and their shunt resistances makes it impossible to use the pentode as an audio tube when the triode is used in a video or timebase circuit. Thus neither section should be used in audio applications when the other section is used in timebase or video circuits.

The tube is not recommended for general use in audio applications because of the possibility of hum. The pentode should not be used as an A.F. amplifier, and its permissible anode dissipation is so small that use of the tube in output stages is not economical.

The use of the triode section as an A.F. amplifier is also limited because of the possibility of hum.

Due to the limitations set to the use of both the pentode section and the triode section in audio circuits, the PCF 80 is not recommendable for use in those circuits, neither as an A.F. amplifier nor as an output tube.

SUMMARY

The table below shows the combinations of functions which are permissible for the two sections in one tube. Illogical combinations, such as R.F. applications combined with applications in pulse circuits, etc., are left out.

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						TRIODE SECTION					
		R.F. oscillator	R.F. amplifier	R.F. mixer	Keyed A.G.C.	Phase inverter	Video amplifier	Line multivibrator	Pulse amplifier for line coincidence	Frame clipper or sync amplifier	Frame blocking oscillator
R.F. mixer		yes	yes								
R.F. amplifier		no	with care	with care							
PENTODE SECTION	I.F. amplifier			with care	with care *)		with care				
	Noise inverter				yes	yes	with care	no	no	no	no
	Video amplifier				no	yes	yes	no	no	yes	no
Sync separator					yes	yes	yes	yes **)	yes	yes	yes
Line multivibrator					no	no	no	yes	yes	no	no
Line coincidence					yes		yes	yes	yes	no	no

*) Provided the video signal for the A.G.C. circuit is taken directly from the video detector.

**) When no flywheel circuit is used.

In the following paragraphs descriptions are given of six applications of the PCF 80 in a TV receiver. The large number of applications for which the PCF 80 is suitable, make this tube preferable to many other tubes, because the use of the PCF 80 reduces the number of different tubes in a set considerably, which obviously is an advantage for production and service.

The following applications of the PCF 80 are discussed in succession:

1. The PCF 80 in a TV tuner.
2. The PCF 80 in a d.c. coupled video amplifier stage.
3. The PCF 80 in a keyed A.G.C. and noise inverter circuit.
4. The PCF 80 as a sync separator and a phase splitter in flywheel circuits.
5. The PCF 80 as a horizontal timebase sine oscillator and a reactance control tube.
6. The PCF 80 in a frame timebase circuit.

The table indicates that the combination of the pentode section as an R.F. amplifier and the triode section as an R.F. oscillator is not possible. This is due to the fairly high capacitance which exists between the anode of the triode section and the control grid of the pentode section. It is clear that it is also not possible to use the triode as a self-oscillating mixer and the pentode as an R.F. amplifier in the front end of an F.M. receiver.

THE PCF 80 IN A TV TUNER

OPERATION OF THE TRIODE SECTION

In the design of the triode section, special care has been taken to ensure reliable operation at oscillator frequencies up to 300 Mc/s with normal circuits and from an H.T. line voltage as low as 180 V. One of the first requirements for the triode is that it should be capable of providing approximately 10 V oscillator voltage with the low circuit impedances occurring at frequencies around 200 Mc/s. Since the pentode mixer section requires about 3.5 V oscillator voltage, the available voltage permits sufficiently loose coupling with its control grid. With the normal Colpitts oscillator circuit the oscillator voltage of the triode grid is then approximately 5 V. Now it is highly desirable that under normal working conditions the tube should operate in class B, because the steep part of the curve representing the effective mutual conductance as a function of the oscillator voltage is then used, whilst, moreover, the ratio of the effective mutual conductance to the mutual conductance at zero grid voltage has still a comparatively high value.

Since it is customary to prevent excessive anode dissipation in the case of oscillator failure by means of a series resistor in the anode supply when the tube operates from an H.T. line voltage of 180 V, the amplification factor must have a comparatively low value. The triode section of the PCF 80 has an amplification factor of 20, so that the desired mode of operation is obtained at an anode voltage of 100 V.

The design considerations for the oscillator circuit are very similar to those for the more familiar frequency changer. However, the fact that there is no direct connection between the oscillator section and the mixer section lays a greater responsibility on the set designer because poor oscillator performance can be masked by a tighter coupling, and so can easily be passed by unnoticed in the development stage.

Two methods of injecting the oscillator voltage onto the control grid of the mixer are open to use. They are:

(1) injection by inductive coupling;
(2) injection by capacitive coupling.

The first method lends itself more advantageously to these applications, especially in the case of a turret tuner, as will be shown later. In the case of capacitive coupling it is recommended that the signal is taken from the anode circuit of the oscillator, as there is already a fixed value of capacitance due to the anode pin of the triode being adjacent to the control-grid pin of the pentode.

Capacitive oscillator injection from the anode side of the oscillator circuit is also easier because the higher a.c. potential exists at the anode side of the coil due to the capacitance from the anode of the triode to earth being smaller than the capacitance from the grid to earth.

Other important aspects of oscillator performance are microphony and hum-modulation caused by variations of the internal tube capacitances.

Hum-modulation is defined as a hum phenomenon which is caused by periodical variations in the interelectrode capacitance between the cathode and the heater of the oscillator tube. The capacitance variation is due to the electro-static forces which exist between the cathode and the heater as a result of the main current flowing through the heater.

Fig. 6 reveals that the cathode of an oscillator tube is tapped on the tuned circuit. It should be realised that at the high frequencies used in modern television receivers neither the cathode nor the heater can be considered to be earthed and, therefore, the capacitance between the cathode and heater C_{kf} influences the tuning frequency of the oscillator circuit. The electro-static forces between the cathode and heater make the heater vibrate in the cathode tube so that C_{kf} varies and detunes the oscillator circuit. The fundamental frequency of the resulting hum-modulation appears to be 100 c/s when the frequency of the mains is 50 c/s.

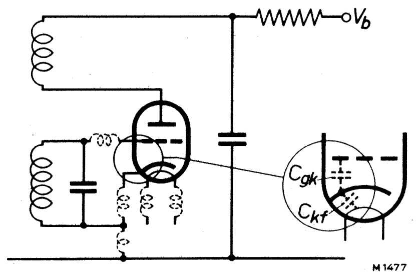


Fig.6. Oscillator with tuned grid circuit. At high oscillator frequencies, the cathode and the heater cannot be considered to be earthed. The cathode is tapped on the oscillator circuit via C_{gk} and C_{kf}.

Microphony is caused by variation of the interelectrode capacitances which results from external excitation of the tube. This phenomenon obviously also causes detuning of the oscillator.

Both microphony and hum-modulation result in sudden or periodical variations of the oscillator frequency, and these variations result in turn in interfering FM and AM in the I.F. signal. The appearance of AM signals is due to the limited bandwidth of the I.F. amplifier.

The requirements which have to be fulfilled by the oscillator tube of a TV receiver with respect to microphony and hum-modulation depend on the types of circuits used.

There are three different types of sound circuits, and it depends partly on the TV system used which type of sound circuit is selected. Since amplitude modulation of the vision carrier is prescribed in all TV systems, the vision channel of every receiver comprises an AM detector - the video detector - and in this respect the requirements of all receivers are equal. The main properties of the various sound circuits are described below.

a) The sound channel mostly used is the intercarrier system for FM sound. With receivers provided with such a sound channel the sound I.F. is equal to the frequency difference between the picture carrier and the sound carrier. The frequency of the local oscillator influences the sound I.F. of such a receiver to a very small extent only.

b) When the sound carrier is frequency modulated it is also possible to use the split-carrier system. This means that, after the R.F. signal has been converted to an I.F. signal, the vision and sound carriers are amplified and detected separately. Therefore, the sound I.F. deviation is equal to the frequency deviation of the local oscillator.

c) When the sound carrier is amplitude modulated, the split-carrier system is practically always used, so that the sound I.F. deviation is also equal to the frequency deviation of the local oscillator.

Requirements have been stipulated for the hum and microphony which are permissible in the sound and the picture channels. On the basis of these requirements, the maximum permissible variations in each of the interelectrode capacitances of the oscillator tube have been calculated for operation in each of the three types of sound circuits mentioned above.

HUM-MODULATION

Measurements have revealed that in all types of receivers the sensitivity of the sound channel to hum-modulation is at least equal to that of the vision channel. The permissible variations in the interelectrode capacitances of the oscillator tube are, therefore, mainly determined by the sound circuit used.

Since hum-modulation chiefly results in interfering FM, this phenomenon is most noticeable in receivers having a split-carrier FM sound channel. In such channels the interfering FM is fully demodulated in the detector, and results in interference in the sound output signal.

Measurements, based on internationally accepted requirements with respect to hum, have revealed that the maximum frequency deviation of the oscillator that can be tolerated in split-carrier FM sound receivers due to hum-modulation, is 30 c/s. When it is borne in mind that the oscillator frequency is approximately 200 Mc/s, it is clear that this requirement is extremely severe. Measurements show that the variations of the heater-to-cathode capacitance of the oscillator tube (C_{kf} of the triode section of the PCF 80 $\simeq$ 2 pF) must be kept smaller than $1.4 \cdot 10^{-3}$ pF. Of course, the permissible variation depends on the oscillator circuit used, but it clearly demonstrates the difficulties encountered in the construction of oscillator tubes.

The requirements with respect to hum-modulation which are imposed on oscillator tubes operating in receivers having a split carrier AM sound channel, are mainly determined by the sensitivity of the AM detector for FM signals. In practice it appears that this sensitivity is about three times smaller than that of an original FM detector. Therefore, the requirements for the oscillator tubes are about three times less stringent than in the case mentioned above, i.e. for split-carrier FM sound receivers.

It has been stated that the sound I.F., and consequently the whole sound circuit of an intercarrier FM sound receiver is almost independent of the local oscillator frequency. As a result, the requirements for the oscillator tube of such a receiver are in practice approximately 100 times less stringent than those for an oscillator tube in a split-carrier FM receiver.

MICROPHONY

As to microphony, not only variations in the heater-to-cathode capacitance should be considered, but also variations in all other interelectrode capacitances. Since the nature of microphony is similar to that of hum-modulation in so far as it causes also interfering FM, the relative sensitivity to microphony of the three different types of sound circuits is equal to the sensitivity of these types to hum-modulation. This means that a split-carrier FM sound receiver is most sensitive to microphony, whereas a split carrier AM sound receiver is three times less sensitive and an intercarrier FM sound receiver is 100 times less sensitive than a split-carrier FM sound receiver.

With split-carrier receivers the sensitivity to microphony in the picture and in the sound output is almost equal. In intercarrier receivers, however, the video channel is more sensitive to microphony than the sound channel, as follows from the figures given above.

As an example of the requirements which are to be made to the triode section of a PCF 80 with respect to microphony, it may be mentioned that variations in the anode-to-grid capacitance (C_{ag} = 1.5 pF) which are larger than $0.3 \cdot 10^{-3}$ pF cannot be tolerated in any type of the receivers described.

It will be clear from the figures mentioned above that in an oscillator tube operating at 200 Mc/s the requirements for microphony and hum-modulation are exceedingly difficult to meet, especially with regard to the capacitance variation between heater and cathode. For this reason the PCF 80 has a specially shaped heater, reducing the capacitance variation to a very low level. It should, however, be recognized that, with the temperature differences involved, some play of the heater in the cathode cannot be avoided.

The normal receiver using the intercarrier system with FM sound is least susceptable to this phenomenon, and among the various arguments in favour of the intercarrier system this is a very important one.

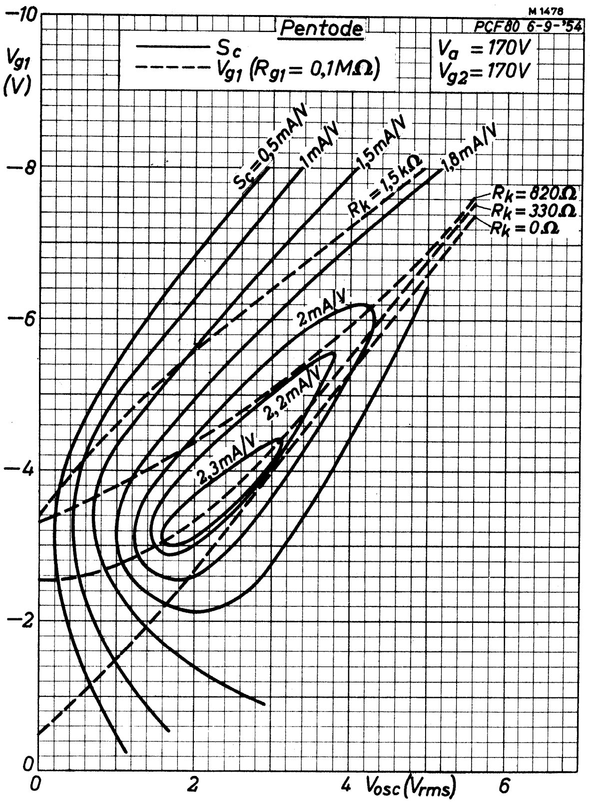
With split-carrier FM sound it is necessary to pay much attention to the positioning of the tuner with respect to the loudspeaker. In general the mechanical excitation of the tubes must be kept as small as possible by resilient mounting of the tuner. The circuit lay-out should be such that the oscillator voltage on the cathode and the heater of the triode is very small.

A very short lead must, therefore, be used to connect the socket contact of the cathode to the chassis. It is bad practice to earth the central shield of the tube holder via the same lead, as this increases the coupling of the cathode with the oscillator circuit.

Finally, the alternating mains voltage between heater and cathode should be kept small by placing the PCF 80 heater at a low point of the heater chain.

OPERATION OF THE PENTODE SECTION

The optimum operating conditions of a mixer can conveniently be determined by examining the contours of constant conversion conductance, which have been plotted in Fig. 7 (full lines). The peculiar forms of these contours will be explained by taking as starting point a mixer tube with a square-law $I_a = f(V_{g1})$ characteristic. When the contours for constant conversion conductance are calculated for such a tube, curves similar to those shown in Fig. 8 will be obtained. In this graph the ratio $V_{oscp}/-V_{g0}$ of the peak voltage of the oscillator signal at the grid of the mixer tube to the negative grid bias at the cut-off point of the tube has been plotted along the abscissa, whilst the ratio $V_{g1}/-V_{g0}$ of the applied grid voltage to the negative grid bias at the cut-off point of the tube has been plotted along the ordinate. These curves indicate the points at which the ratio S_c/S_0 of the conversion conductance to the mutual conductance of the tube at $V_g = 0$ is constant.



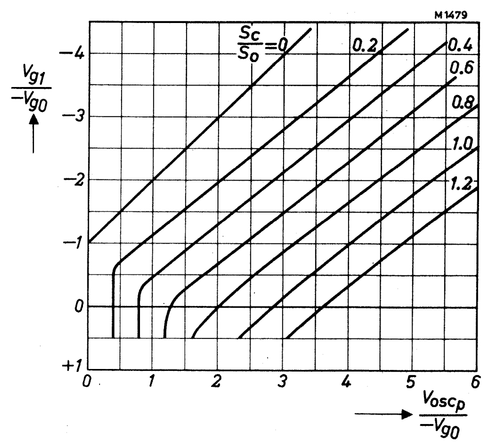


Fig.8. Contours of constant conversion conductance S_c of an idealised pentode with a square-law $I_a = f(V_{g1})$ characteristic.

Fig.7. Contours of constant conversion conductance S_c of the pentode section (full lines), and curves giving the d.c. bias developed across a grid leak resistor R_{g1} of 0.1 MΩ and cathode resistors R_k of various values (dashed lines). The oscillator frequency is 200 Mc/s.

It is seen that for practically all oscillator voltages the ratio S_c/S_0 is proportional to the applied oscillator voltage and the negative grid bias, and that the curves drop off vertically only for small values of $V_{oscp}/-V_{g0}$ and $V_{g1}/-V_{g0}$. The latter effect is due to the negative grid bias of the mixer tube and the applied voltage being so small that the cut-off point of the tube is not exceeded. In fact, the conversion conductance for a given amplitude of the oscillator voltage remains constant when the grid voltages are such that this condition is satisfied.

Since the actual $I_a = f(V_{g1})$ characteristics of a mixer tube do not obey a square-law rule, the contours for constant conversion conductance will not be straight. As shown by Fig. 7, the contours even completely change their direction and, for a given grid bias, the conversion conductance decreases as the oscillator voltage increases above a certain value. This is due to the fact that at positive grid voltages grid current starts to flow. As a result, the conversion conductance characteristic will have a form as shown in

Fig. 7 . When the oscillator voltage applied to the mixer tube is so large that the tube also operates in the area at which the mutual conductance characteristic is flat, the amplitude of the fundamental of the variation in mutual conductance, due to the applied oscillator voltage, will decrease, and so will the conversion conductance.

The values of the conversion conductance S_c at various oscillator voltages can be derived from Fig. 7 , provided the relation between the oscillator voltage amplitude and the negative grid bias is known.

The dashed lines plotted in Fig. 7 represent the bias developed across the 0.1 MΩ grid resistor and various values of cathode bias resistors. The lowest line applies to grid leak bias exclusively; the others are for grid leak bias combined with 330 Ω, 820 Ω and 1.5 kΩ cathode bias resistors. The locus of the intersection of these lines with the contours therefore gives the normal conversion characteristics for each condition of bias.

An examination of the contours of conversion conductance reveals that the maximum value of 2.45 mA/V is reached at an oscillator voltage of 2.3 V (r.m.s. value) and a negative grid bias of 3.6 V. This condition can be obtained by using a 330 Ω cathode resistor in addition to the grid leak resistor. The characteristic for this adjustment can be seen more clearly in Fig. 9 , in which the conversion conductance has been plotted as a function of the oscillator voltage. This graph shows that the condition which gives maximum conversion leads to a rather peaky curve. This can be avoided by increasing the value of the cathode resistor, which will, moreover, reduce the influence of spread in tubes, and this is a definite advantage.

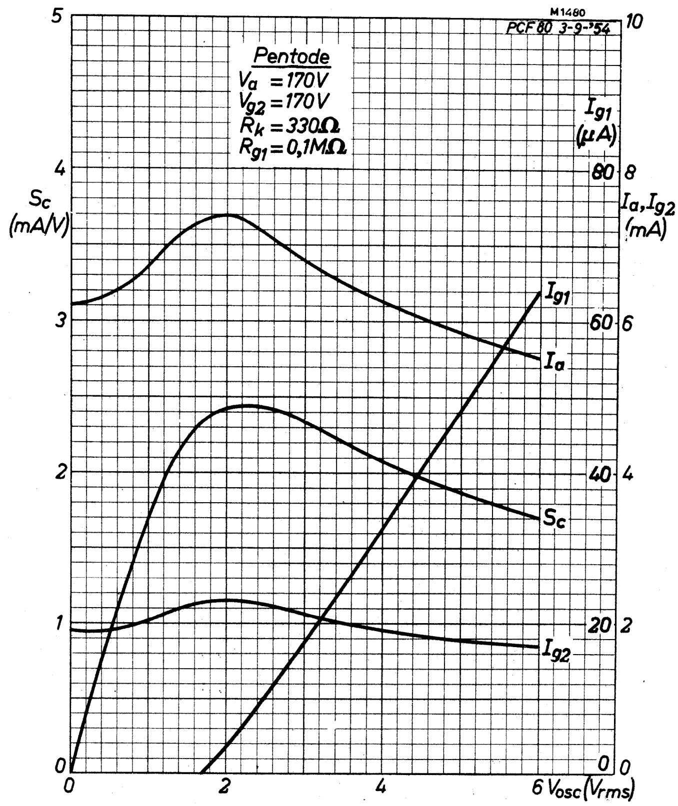


Fig.9. Conversion conductance S_c and control-grid, screen-grid and anode currents I_{g1}, I_{g2} and I_a respectively, as functions of the oscillator voltage V_{osc} at R_k = 330 Ω and R_{g1} = 0.1 MΩ.

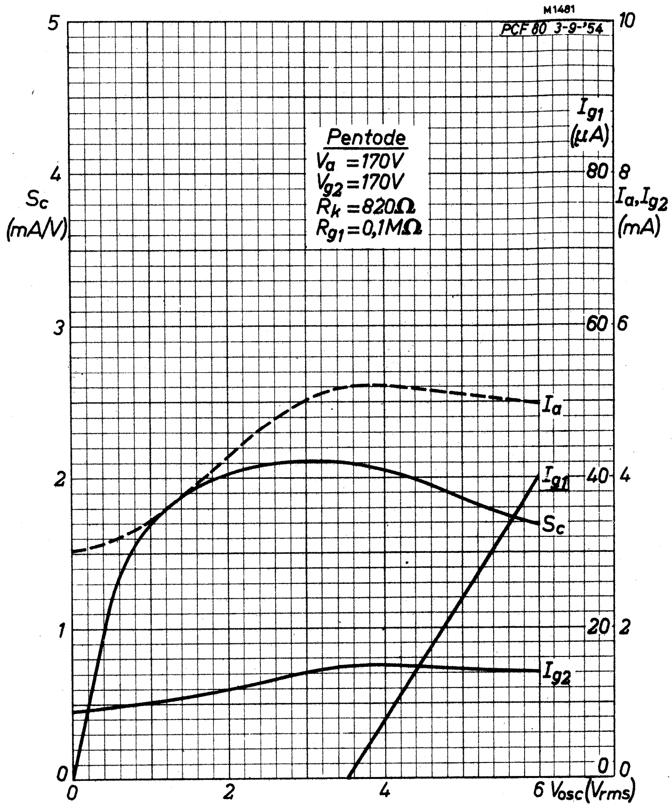


Fig.10. Conversion conductance S_c and control-grid, screen-grid and anode currents I_{g1}, I_{g2} and I_a respectively, as functions of the oscillator voltage V_{osc} at R_k = 820 Ω and R_{g1} = 0.1 MΩ.

At a value of R_k = 820 Ω the curve of conversion conductance runs almost parallel to a contour which is still reasonably near the maximum (cf. Fig.7). The characteristic for this condition is given in Fig.10. This curve is substantially flat for an oscillator voltage of 2 V up to 5 V and gives a variation of conversion conductance from 1.9 mA/V to 2.1 mA/V; it is therefore recommended that the design be centred on an oscillator voltage of 3.5 V_{rms}. In the case of the 330 Ω cathode resistor the design is also centred on 3.5 V to allow for variations; this condition results in a change of the conversion conductance from 1.90 mA/V to 2.45 mA/V for the same change of drive.

The total gain is influenced not only by the conversion conductance but also by the input damping of the mixer section. The value of this damping at the tube holder may be calculated from the equivalent circuit shown in Fig.11, in which the symbols used have the following meanings:

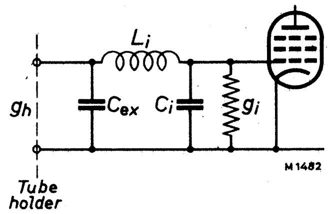


Fig.11. Equivalent circuit for calculating the damping g_h at the tube holder.

g_i = input conductance of the warm tube (approx. 500 µA/V).

C_i = interelectrode capacitance, i.e. the sum of the capacitance C_{g1k} between control grid and cathode, and the capacitance C_{g1g2} between control grid and screen grid (approx. 6.5 pF).

L_i = self-inductance of the tube holder contacts and the control-grid lead (approximately $30 \cdot 10^{-9}$ H).

C_{ex} = the stray capacitance at the tube contacts due, for example, to the capacitance of the tube holder and the oscillator circuit.

Calculation shows that, due to the presence of C_{ex}, L_i and C_i, the damping at the tube holder g_h becomes approx. 1100 µA/V at g_i =

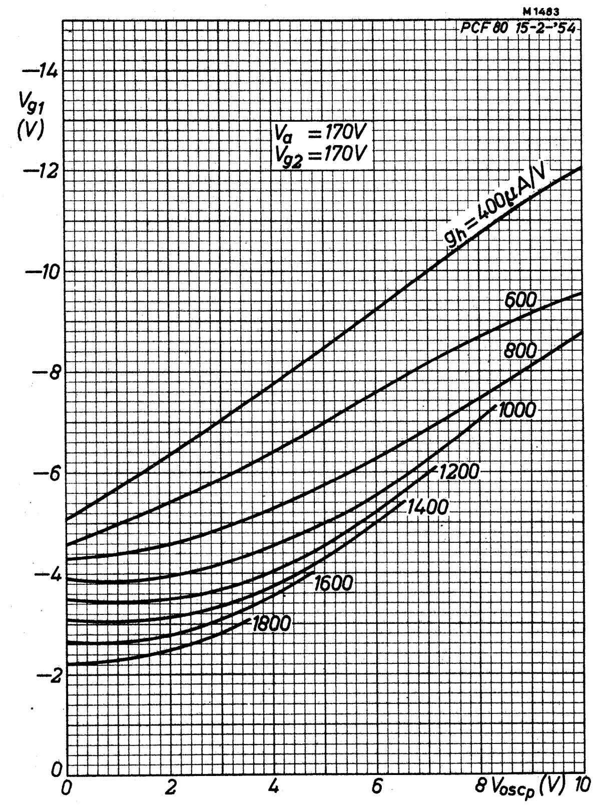


Fig.12. Contours of constant damping g_h as a function of the peak value of the oscillator voltage $V_{osc\ p}$ and the fixed negative grid bias V_{g1}.

500 μA/V, the signal frequency being assumed to be 200 Mc/s. The value of g_h at different values of the oscillator voltage and of the fixed negative grid bias can be determined from the graph shown in Fig.12.

The forms of the g_h contours may be explained in a similar way as those of the contours for constant conversion conductance. At a given frequency the input conductance consists of a constant part and a part which is proportional to the average mutual conductance S_{med}. If the characteristics of a tube would obey a square-law, then the curves representing S_{med}/S_0 would assume the form represented in Fig.13. In order to derive from this graph the probable g_h contours, it should, however, be taken into account that the damping increases considerably when grid current starts to flow. This explains the irregular form of the contours plotted in Fig.12.

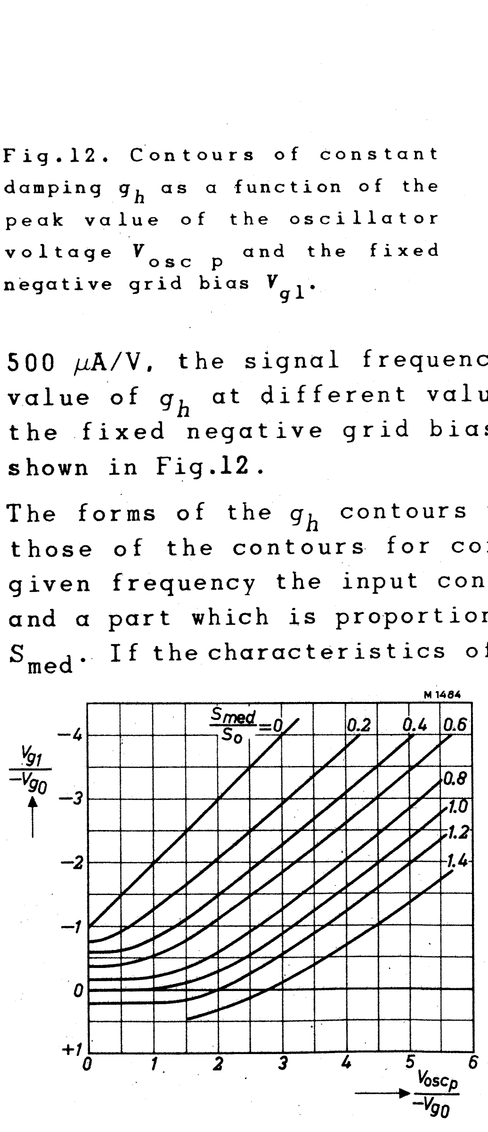


Fig.13. Curves representing the ratio S_{med}/S_0 of an idealised pentode with a square-law $I_a = f(V_{g1})$ characteristic.

The values of g_h occurring at various oscillator voltages can be derived from the V_{g1}/V_{osc} graph shown in Fig. 7 once the value of the grid leak and the cathode resistor have been determined.

It is possible to influence the value of g_i and thus that of g_h by varying the damping g_t due to feedback, which is given by:

$$g_t = \omega^2 S_{k\ eff} \left(M_{ag1}C_{kg1} - L_{g2}C_{g1g2} \cdot \frac{S_{g2}}{S_{k\ eff}}\right).$$

in which

$S_{k\ eff}$ = the effective value of the cathode transconductance,
M_{ag1} = the mutual inductance which may be imagined to exist between the anode circuit and the control-grid circuit, mainly in consequence of the cathode lead being common to the loops which are formed between the anode and cathode and between the cathode and control grid,
L_{g2} = the self-inductance of the screen-grid circuit.
C_{kg1} = the capacitance between the cathode and the control grid,
S_{g2} = the screen-grid transconductance.

This expression reveals that g_t is strongly dependent on the value of M_{ag1}, which should be kept small by using short, thick wires in the cathode circuit and, moreover, on the value of L_{g2}, which may be increased to such an extent that g_t becomes almost zero. Care must, however, be taken that stability of the mixer is not upset.

Special attention should be paid to the value of the by-pass capacitor in the cathode circuit. When this capacitor is made too small the cathode circuit becomes capacitive in the low frequency channels, and undamping occurs, with a consequent risk of instability. The capacitor should, therefore, have a sufficiently high value to ensure that the impedance of the cathode circuit remains low under all conditions. This is also of importance because the internal screen is connected to the cathode of the pentode section.

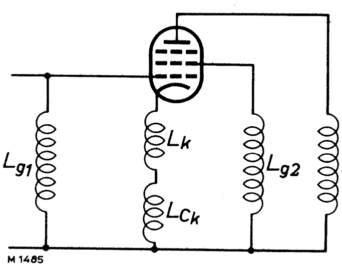


Fig.14. Equivalent circuit applying to the case of the lower end of the control-grid circuit being connected to the chassis.

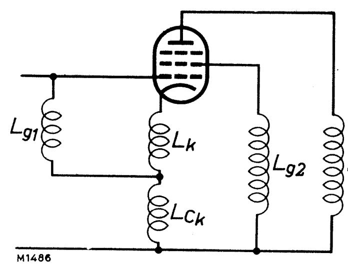


Fig.15. Equivalent circuit applying to the case of the lower end of the control-grid circuit being connected to the cathode.

When the lower end of the grid circuit is connected to the chassis (see Fig.14) the self-inductance L_{Ck} originating from the leads of this capacitor will increase the value of M_{ag1}, so that the input damping also increases. On the other hand, by connecting the lower end of the input circuit to the cathode (see Fig.15), L_{Ck} increases the value of L_{g2}, so that g_t can be made considerably smaller.

PRACTICAL CIRCUIT

A brief description is given of a tuner equipped with the PCC 84 and PCF 80.

In the tuner (see Fig.16) the PCC 84 operates as a d.c. coupled cascode amplifier. The d.c. coupling is desirable as it gives a characteristic that is suitable for the application of gain control. To obtain the best control characteristic, the second tube should have cathode bias. This has been checked by measuring the cross-modulation, which was found to be approximately four times better with this arrangement than with fixed bias on the second half of the tube or with the a.c. coupled circuit. The figures obtained for the cross-modulation were low enough to be irrelevant.

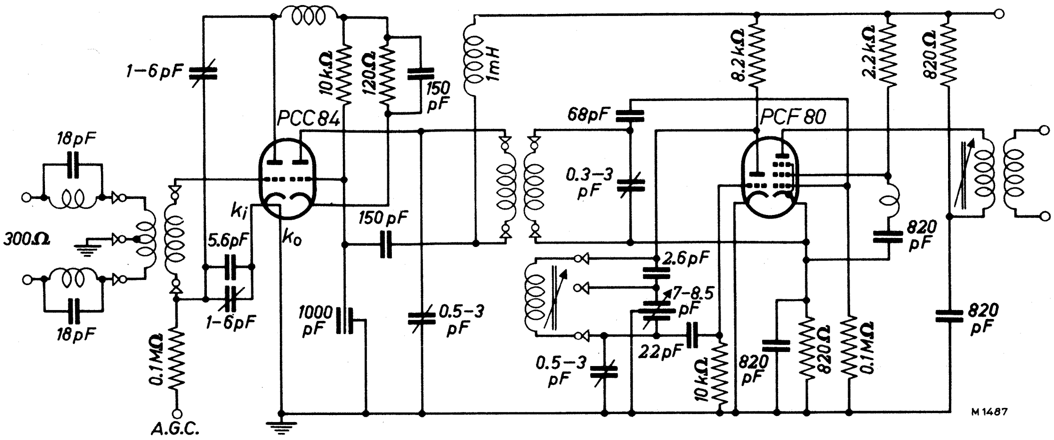


Fig.16. Circuit diagram of a tuner equipped with the PCC 84 and the PCF 80.

The PCC 84 with its two cathode leads for the input triode has the advantage that either separated or strapped leads may be used, according to whether most value is attached to a slightly higher gain or to a slightly better noise factor respectively.

The optimum operating condition for the PCF 80 can be found in the data and the preceding description. In the circuit given here an oscillator of the Colpitts type is used. By connecting the capacitor of 2.6 pF in series with the tuning capacitor of the oscillator circuit in band III, the capacitance in the circuit and thus the losses are reduced, as a result of which the decrease of the amplitude of the oscillator signal at high frequencies is counteracted. The trimmer of 0.5 - 3.0 pF connected between the grid of the oscillator tube and earth has been provided to cope with the variations of C_{gk} due to spread in interelectrode capacitances of the tube or in the wiring capacitances. It may be necessary to readjust the setting of the trimmer when the PCF 80 is replaced.

The oscillator voltage is applied inductively to the grid of the mixer. This is an advantage compared to capacitive coupling, because it offers the possibility of keeping the oscillator voltage at the grid fairly constant when the tuner is switched over from one channel to another. In fact, inductive coupling allows the coupling coefficient between the oscillator coil and the coil in the grid circuit of the mixer to be adjusted to the required value for each channel by switching over the coils of both circuits, whereas in the case of capacitive coupling the coupling between these two circuits cannot be adjusted due to the capacitor not being switched over. It will be clear that by using inductive coupling, which ensures a higher constancy of the oscillator voltage at the grid of the mixer, the conversion gain at the various channels will also have a higher constancy. Moreover, inductive coupling offers the advantage that the input damping exerted on the grid circuit of the mixer is slightly smaller, and this will result in a higher gain of the mixer stage.

The small inductance included in the screen grid to reduce the input damping of the mixer has been obtained by keeping the connection leads of the by-pass capacitor to the screen grid slightly longer than conventional. This way of reducing the input damping does not lead to instability of the mixer, as might be expected. This is due to the fact that in the high frequency band (approx. 200 Mc/s), where the influence of the inductance is greatest, there is a great difference between the carrier frequency at the grid of the mixer and the intermediate frequency at the anode; in the low frequency band (approx. 50 Mc/s) the influence of this inductance is only very small.

Measurements have been carried out on this tuner in the channels 2 and 10. The results of these measurements, which are indicative for the performance of a tuner equipped with these tubes, are tabulated below.

AERIAL CIRCUIT

		(channel 2)	(channel 10)	
Picture carrier frequency		48.25	210.25	Mc/s
Sound carrier frequency		53.75	215.75	Mc/s
Resonant frequency	tube switched on	54	216	Mc/s
Total tuning capacitance		6.9	6.6	pF
Bandwidth		3.9	4	Mc/s
Total conductance (g_{tot})		169	166	μA/V
Tapping ratio for grid (t) [1]		0.58	0.60	
Aerial conductance (g_{ant})		3330	3330	μA/V
Aerial gain		2.6	2.7	
Bandwidth with matched aerial		7.8	8	Mc/s

DRIVER CIRCUIT

		(channel 2)	(channel 10)	
Voltage gain across intermediate circuit between anode of grounded-cathode section and cathode of grounded-grid section	n	= 1	0.52	
Output conductance of driver triode/n [2]	$\frac{1}{n} g_{o1}$	= 1250	2400	μA/V
Circuit conductance/n (at primary side)	$\frac{1}{n} g_c$	= 0	635	μA/V
Input conductance of grounded-grid section times n [3]	$n\, g_{i2}$	= 4360	2510	μA/V
Voltage gain (g to k') [4]	G_1	= 1.07	1.05	

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[1]) The aerial gain may be slightly increased by increasing the tapping ratio t.

[2]) The value of g_{o1} is smaller than the internal resistance of the tube due to the negative feedback by the neutralising bridge.

[3]) The input admittance is calculated from the expression

$$g_{i2} = (\mu + 1) / (R_i + Z),$$

where Z is the input admittance of the H.F. transformer.

[4]) The voltage gain of the driver section is calculated from the expression:

$$G_1 = S/(\frac{g_{o1} + g_c}{n} + g_{i2} \cdot n),$$

where S = 6 mA/V.

GROUNDED-GRID SECTION

				channel 2	channel 10	
Double tuned R.F. Transformer	Impedance of primary (without tube damping)	Z_p	=	7.75	4.2	kΩ
	Impedance of secondary (with mixer input)	Z_s	=	1.2	0.9	kΩ
	Coupling	kQ	=	1.5	1.1	
	Transfer impedance	Z_{tr}	=	1.4	0.97	kΩ
	Input impedance	Z_i	=	2.38	1.9	kΩ
	Calculated bandwidth	B	=	6.8	7	Mc/s
Voltage gain from cathode to anode [1]		G_2	=	10.4	9.1	
Ratio of secondary to primary voltage [2]				0.59	0.51	
Total voltage gain of cascode amplifier from aerial terminals to grid of mixer				15.7	12.5	

MIXER CIRCUIT

				channel 2	channel 10	
I.F. Transformer	Impedance of primary (with tube damping)	Z_p	=	10		kΩ
	Impedance of secondary (with extra damping of 3.3 kΩ)	Z_s	=	2.3		kΩ
	Coupling	kQ	=	1.75		
	Transfer impedance	Z_{tr}	=	2.1		kΩ
	Calculated bandwidth	B	=	6.6		Mc/s
Conversion conductance		S_c	=	2.1		mA/V
Conversion gain from grid of mixer to grid of I.F. tube				4.4		
Total voltage gain of tuner from aerial terminals to grid of I.F. tube (calculated)				70	55	
Measured total voltage gain				74	57	

CIRCUIT DATA

PCC 84

Supply voltage	V_b	=	180 V
Anode voltage of grounded-cathode section	V_{a1}	=	89 V
Anode voltage of grounded-grid section	V_{a2}	=	89 V
Grid voltage of grounded-grid section	$-V_g$	=	1.5 V
Anode current	I_a	=	12 mA

PCF 80 *(Triode section)*

Anode voltage	V_{aT}	$\simeq$	80 V
Anode current	I_{aT}	$\simeq$	12 mA

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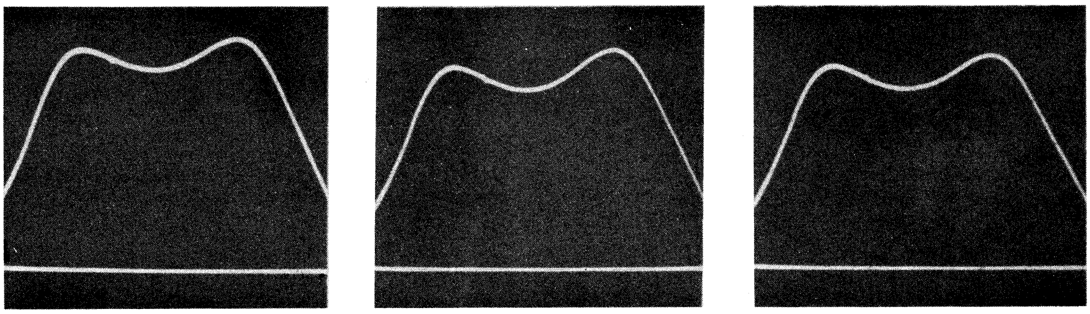
1) $G_2 = (\mu + 1) \dfrac{Z_i}{R_i + Z_i}$, where $\mu = 24$.

2) $\dfrac{E_{sec}}{E_{prim}} = kQ \sqrt{\dfrac{Z_s}{Z_p}}$

PCF 80 (*Pentode section*)

Anode voltage..............................	V_{aP}	=	175 V
Screen-grid voltage........................	V_{g2}	=	176 V
Anode current..............................	I_{aP}	=	5.2 mA
Screen-grid current........................	I_{g2}	=	1.5 mA
Conversion conductance.....................	S_c	=	2.1 mA/V
Cathode voltage............................	V_{kP}	=	5.5 V
Control-grid current.......................	I_{g1}	=	5 μA
Effective oscillator voltage at control grid	V_{osc}	=	3.8 V_{rms}

Figs 17*a* and *b* represent frequency response curves of the tuner for the channels 4 and 10 respectively. The curves are given for three random samples of the PCF 80 being used as mixer tube.



M 1488

Fig.17*a*. Response curves of the tuner at channel 4.

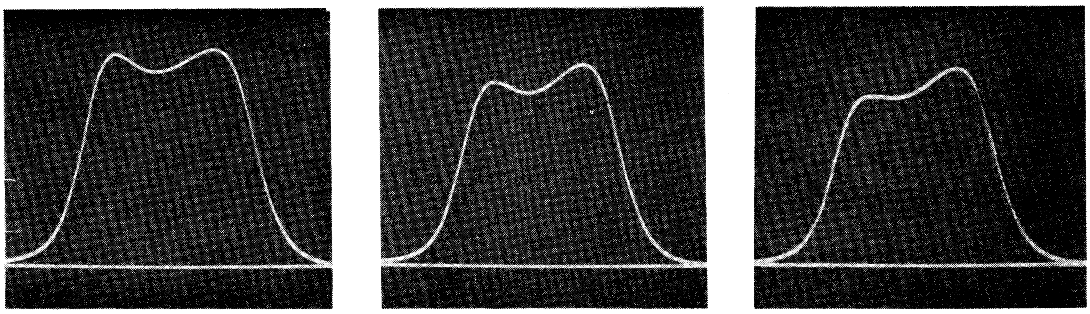


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Fig.17*b*. Response curves of the tuner at channel 10. The curves are recorded for three random samples of the PCF 80.

The next three paragraphs deal with applications of the PCF 80 which are in many respects inter-related. For the sake of convenience the relation between these applications is shown in Fig. 18, which has been placed as a folding diagram at the end of this booklet.

D.C.-COUPLED VIDEO AMPLIFIER STAGE

PRINCIPLE OF OPERATION

The use of a PCF 80 as a video amplifier with cathode follower output has many advantages over the use of a single high-slope pentode, since the gain of a video amplifier is limited by the output or load capacitance. A typical circuit using a PL 83 with one series peaking coil has a gain of approximately 15 and a bandwidth of 5 Mc/s with a load capacitance of 26 pF. Under the same conditions the video amplifier with a PCF 80 has a total gain of 30; the bandwidth is maintained by the use of a cathode follower to reduce the effective load capacitance. The maximum video output is 60 V_{p-t-p} at V_b = 180 V.

In addition to larger gain the PCF 80 has a lower heater voltage than the PL 83, which may be important in large, series-fed receivers where the set designer is often confronted with a shortage of heater voltage. Furthermore, it is possible to reduce the number of tube types in a TV-receiver due to the large number of applications for which the PCF 80 is suitable, whereas the PL 83 can be used as a video output tube only.

A further advantage of a video amplifier with a PCF 80 is that the sum of the anode current of the pentode section and that of the triode section is small (approx. 14 mA), and that these currents vary in opposite directions at a certain signal variation. As a result, the mean current drain of the video stage varies only slightly at large variations of the signal content.

The anode current of the PL 83 may vary between 15 and 60 mA depending on whether the picture is dark or bright. These large current variations can obviously influence the value of the supply voltage and consequently affect other circuits of the receiver.

The number of components of a video amplifier with cathode follower output is slightly larger than that of a video output stage with a single PL 83. This fact is completely counterbalanced by the advantage of the use of the PCF 80 in a video amplifier.

The video amplifier that is described here is d.c. coupled, which renders the amplifier superior - with respect to its performance at large interference pulses - to a similar video amplifier with cathode follower output and *RC*-coupling.

In an *RC*-coupled amplifier the coupling capacitor of the cathode follower tube tends to be charged by grid current which flows at the occurrence of large interference pulses. As a result, the cathode of the picture tube is driven highly negative and the luminescent screen turns completely white. This condition continues until the coupling capacitor is discharged via the grid leak resistor. This drawback does not occur, of course, in a d.c. coupled amplifier, which means a great advantage, especially for receivers in areas with heavy traffic and in the surroundings of factories and other sources of interference.

Due to the d.c. coupling of the video amplifier, a d.c. restoration circuit can be omitted since the video signal at the video detector has a constant black level when the amplitude of the I.F. signal at the input of the detector is constant (Fig.19). This constancy is ensured when an effective A.G.C. voltage is applied to the R.F. and I.F. amplifiers. Under these conditions the amplitude of the synchronisation pulses at the control grid of the pentode section of the PCF 80 is constant. The video signal at the grid has, therefore, a fixed black level so that no d.c. restoration is required when the video amplifier is d.c. coupled.

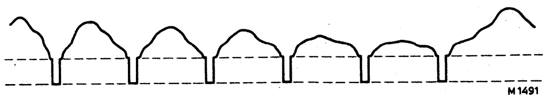


Fig.19. Input signal of the video amplifier.

CIRCUIT DESCRIPTION

Fig.20 shows the diagram of a d.c. coupled video amplifier with a PCF 80; the polarities of the various signals are also indicated.

The video signal, which is taken from the video detector, is applied to the control grid of the pentode section of the PCF 80 via the series peaking network L_1 - R_4. The anode signal of this tube is fed directly to the grid of the triode section, which operates as

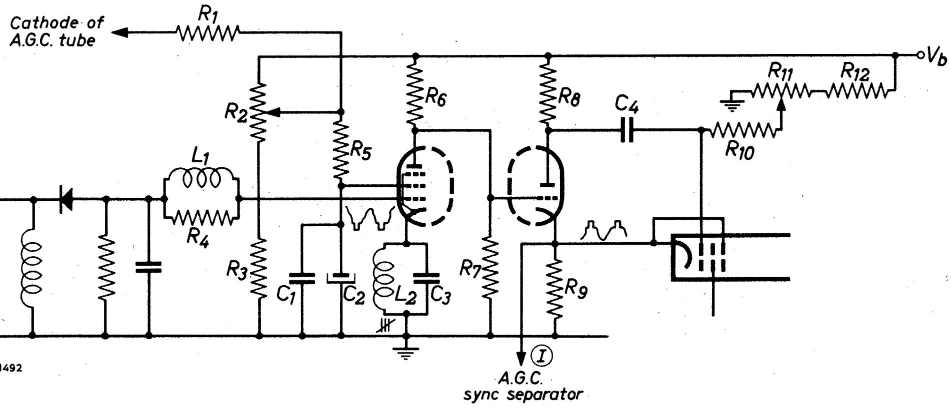


Fig.20. D.C. coupled video amplifier with the PCF 80.

PARTS LIST:

R_1 = 0.22 MΩ	R_7 = 18 kΩ (1 W, ±5%)	C_1 = 1500 pF
R_2 = 20 kΩ (variable)	R_8 = 1.2 kΩ	C_2 = 10 μF
R_3 = 33 kΩ (1 W)	R_9 = 12 kΩ (1 W)	C_3 = 1500 pF
R_4 = 3.9 kΩ	R_{10} = 1 MΩ	C_4 = 1500 pF
R_5 = 10 kΩ	R_{11} = 0.2 MΩ	L_1 = 60 μH
R_6 = 10 kΩ (3 W, ±5%)	R_{12} = 0.39 MΩ	L_2 = 0.55 μH

All resistors should be $^1\!/_4$ W unless a different wattage is indicated.

a cathode follower. The signal at the cathode of this tube - which obviously has almost the same amplitude as the anode signal of the pentode section - is applied directly to the cathode of the picture tube. A small signal is taken from the anode of the cathode follower and also applied to the picture tube, thus providing an improvement of the frequency response of the video stage, as will be described later.

The anode of the pentode part, and hence the grid of the triode part, are connected to the potentiometer R_6 - R_7 to ensure a sufficient anode voltage to be present at the cathode follower.

The screen grid of the pentode section is fed from the supply voltage via the potentiometer R_2 which serves as a contrast control.

The screen grid is decoupled to earth to prevent capacitive feed-back from the screen grid to the control grid (Miller capacitance).

In the cathode lead of the pentode amplifier tube a circuit, tuned at 5.5 Mc/s, is included to suppress interfering sound I.F. signals.

CONTRAST CONTROL

The potential of the screen grid of the pentode section, which is determined by the setting of the contrast control, has little effect on the video stage gain. The maximum variation of gain due to this effect amounts to 15% of the maximum gain only. The main cause of the amplitude variation of the video signal results from a variation in the A.G.C. voltage which occurs at a change in the setting of the contrast control. This can be explained as follows. The setting of the contrast control determines the maximum obtainable anode current of the pentode amplifier tube. Since the pentode section has no automatic bias, the anode current will always be maximum at "peak white" of the video signal. Therefore, the setting of the contrast control determines indirectly the anode potential - and the cathode potential of the cathode follower - which occurs at peak white of the video signal.

To avoid confusion it will be useful to define the terms "white level" and "black level", which are frequently used in the following paragraphs. The white level V_w of a video signal is understood to be the potential difference existing between the video signal at "peak white" and earth (Fig.21). V_w is *independent* of the amplitude and of the content of the video signal; it depends only on the setting of the contrast control.

The black level V_{black} of a video signal is understood to be the potential difference between the potential of the blanking pulses of the video signal and earth (Fig.21). In the circuit described V_{black} of the video signal at the cathode of the cathode follower, and consequently at the grids of the A.G.C., noise inverter and sync separator tubes, is constant and independent of the signal amplitude and the setting of the contrast control.

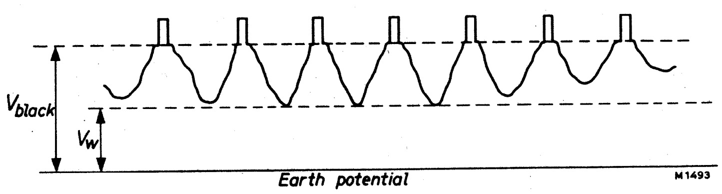


Fig.21. Video signal with indication of the white level V_w and of the black level V_{black}.

The potential of the screen grid increases as the contrast is increased, so that the white level at the anode of the pentode and at the cathode of the cathode follower lowers. The A.G.C. circuit, which is d.c. coupled to this cathode (Fig.18), is so designed that a reduction of the white level of the video signal that is not accompanied by a change of the signal amplitude itself, causes a reduction of the A.G.C. voltage produced. As a result, the gain of the R.F. and I.F. amplifiers increases, and the video signal at the grid of the video amplifier becomes larger. Obviously a reduction of the screen-grid potential of the video amplifier results - via the A.G.C. circuit - in a reduction of the gain of the R.F. and I.F. amplifiers.

It follows from the description of the contrast control that the white level of the signal at the cathode of the cathode follower depends on the setting of the contrast control, and not on the amplitude of the antenna signal. At a certain setting of the contrast control the white level of the output signal of the video stage remains therefore at a constant value, independent of signal amplitude or other influences.

The screen-grid voltage is applied to the A.G.C. circuit via R_1 to obtain a constant black level of the output signal of the video stage. The operation of the A.G.C. circuit will be explained in the next section of this Bulletin. For better understanding of the operation of the video amplifier, however, it is remarked here that, due to the A.G.C. action, the black level of the video output signal is constant at all settings of the contrast control and at all amplitudes of the video signal. Therefore, once the brightness of the picture tube is adjusted, the contrast of the picture can be changed without a readjustment of the brightness control being necessary.

GAIN AND MAXIMUM OUTPUT VOLTAGE

As has already been remarked, the anode current of the pentode section of the PCF 80 is maximum ($V_{g1} \simeq 0$ V) at the white level of the video signal. Furthermore, the A.G.C. circuit is so designed that the video amplifier is near cut-off during the occurrence of the synchronisation pulses. The amplitude of the video signal at the control grid required for proper operation with the amplifier at maximum contrast is about 2 V_{p-t-p}. The efficiency of the video detector being assumed to be 70 %, the output voltage of the I.F. amplifier has to be 2.3 V_{rms}. When this voltage is lower, the operation of the A.G.C. circuit and the contrast control is upset. It is therefore necessary that the gain of the R.F. and I.F. stages is sufficiently high to ensure that the required signal level at the input of the video amplifier is obtained even in fringe areas.

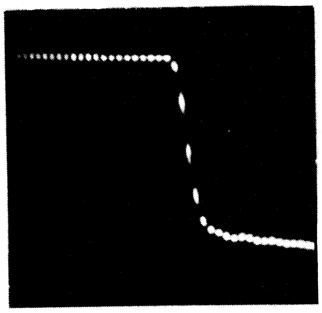
Since the video signal at the control grid of the video amplifier tube is almost equal to the grid base - due to the A.G.C. action -, the anode voltage varies between a maximum value determined by the potentiometer R_6, R_7, which is connected between the supply line and earth, and a minimum voltage determined by the voltage drop across this potentiometer caused by the maximum anode current of the pentode at a given adjustment of the contrast control. The amplifier circuit is so designed that the anode voltage of the pentode never drops below the knee of its $I_a = f(V_a)$ characteristic (approx. 30 V) to prevent distortion in the white parts of the video signal. The potentiometer reduces the supply voltage to 115 V at the anode of the pentode, so that the maximum obtainable output voltage would be 115 V - 30 V = 85 V.

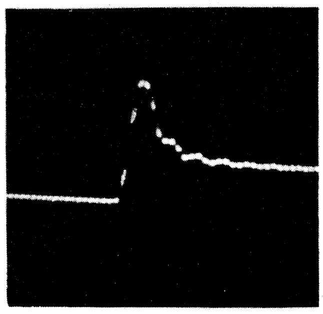
However, for reasons which will be explained later, the pulse level will not reach 115 V, but only about 100 V, since the A.G.C. circuit starts operating at this level. Moreover, 10% or more of the I.F. signal does not contain picture information, as is prescribed in the C.C.I.R.-standards in connection with the operation of the intercarrier sound circuit. The maximum output voltage will, therefore, be about 60 V, and the maximum obtainable gain 60 : 2 = 30.

STEP FUNCTION RESPONSE

The output voltage of the cathode follower is applied to the cathode of the picture tube. Besides, a video signal of smaller amplitude and opposite phase is taken from the anode of the cathode follower and applied to the grid of the picture tube. This way of controlling the picture tube considerably improves the step function response of the video amplifier. This can be explained as follows.

The large capacitive load on the cathode follower slightly affects the rise time of the output signal. The rise time can be improved by adding the derivative of the input signal of the cathode follower to the output voltage. The derivative is present in the anode current due to the integrated shape of the cathode signal





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a

b

Fig.22*a*. Signal at the cathode of the triode section of the PCF 80 when a negative-going pulse is applied to the grid of the tube.

Fig.22*b*. Signal at the anode of the triode section of the PCF 80 when a negative-going pulse is applied to the grid of the tuner.

Figs 22*a* and *b* represent the voltage at the cathode and at the anode of the triode if a negative-going pulse is applied to the grid of the tube; the repetition frequency of the time reference dots was 20 Mc/s. The small oscillations are caused by the 5.5 Mc/s trap in the cathode circuit of the pentode amplifier. Since the cathode voltage is applied to the cathode of the picture tube, and the anode voltage to the grid of the picture tube, the oscillogram of the anode voltage must be reversed before it is added to that of the cathode voltage. The sum voltage of these signals - which approaches the ideal square wave - can then be considered as the control voltage of the picture tube. The oscillogram of the anode signal is slightly compressed compared to that of the cathode signal; the amplitudes of both signals do not correspond with each other because the amplification of the anode signal by the oscilloscope exceeded that of the cathode signal.

The frequency compensation will be upset when very large square-wave voltages are applied to the grid of the cathode follower. The amplitude of the maximum square-wave voltages that may be applied to the cathode follower depends on the rise time of the square-wave voltages themselves and on the capacitive load on the cathode follower. For instance, at a sudden decrease of the grid voltage, the *RC*-constant in the cathode circuit formed by the cathode resistor and its load capacitance may be so large that the cathode voltage cannot immediately follow the grid signal, and the tube is cut off. For the same reason it is also possible that grid current will flow in the cathode follower at large positive-going potential jumps of the grid signal.

Very large voltage jumps will, therefore, cause a distortion of the anode signal of the cathode follower and consequently of the picture on the cathode ray tube. However, in practice, such large sudden potential jumps will hardly ever occur in a video signal as a quarter of the video signal amplitude is already occupied by the synchronisation pulses.

HUM

Requirements with respect to grid hum and to cathode hum are defined for the PCF 80 being used as a video output tube.

As to grid hum, the maximum permissible a.c. heater-to-cathode voltage V_{kf} is 100 V_{rms} if the impedance of the control-grid circuit Z_{g1} = 5 kΩ.

To prevent the occurrence of an inadmissably large cathode hum, the following values of the cathode resistor R_k may not be exceeded if the resistor is not bypassed:

$$R_k = \text{max. } 150\ \Omega \text{ at } V_{kf} = 100\ V_{rms}$$
$$R_k = \text{max. } 300\ \Omega \text{ at } V_{kf} = 50\ V_{rms}$$

Several data of the video amplifier circuit are tabulated below for a maximum and minimum contrast. Fig.23 gives frequency response curves of the amplifier. The dash-dot line represents the gain measured at the cathode of the cathode follower, and the dashed line the gain measured at the anode of this tube. The total gain of the stage is represented by the fully drawn line which is obtained by simply adding the two other curves. The influence of the sound trap at 5.5 Mc/s can clearly be seen from these curves.

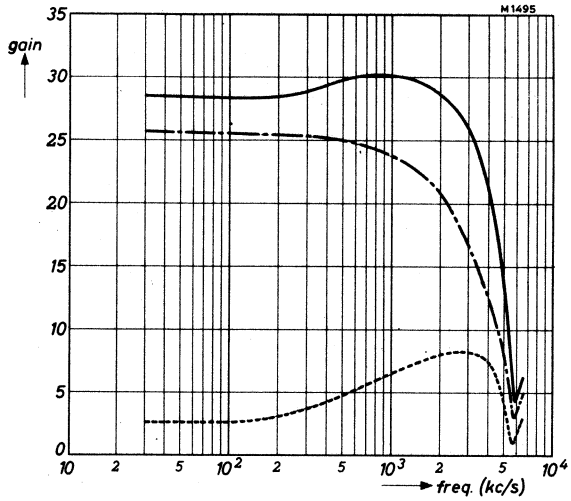

Fig.23. Frequency response curve of the video amplifier. The dash-dot line represents the gain of the amplifier measured at the anode of the cathode follower, the dashed line the gain measured at the cathode. The fully drawn line represents the total gain of the video amplifier.

CIRCUIT DATA

			Pentode section		Triode section		
			maximum contrast	minimum contrast	maximum contrast	minimum contrast	
Supply voltage............	V_b	=	180		180		V
Mean anode voltage.........	V_a	=	60	74	173	172	V
Mean anode current.........	I_a	=	8.7	6.5	5.5	6.5	mA
Mean screen-grid voltage...	V_{g2}	=	145	84			V
Mean screen-grid current...	I_{g2}	=	3	1.6			mA
Mean control-grid voltage..	V_{g1}	=	-1.2	-0.3	60	74	V
Mean cathode voltage.......	V_k	=	0	0	65	78	V
Mean cathode current.......	I_k	=	11.7	8.1	5.5	6.5	mA

For proper operation of the contrast control an A.G.C. circuit is required that reacts to variations in the level of the tops of the synchronising pulses. Therefore, a completely new A.G.C. circuit has been developed. This circuit, together with a noise inverter circuit, will be described in the following sections.

KEYED A.G.C. AND NOISE INVERTER CIRCUIT

A.G.C. CIRCUIT

A keyed A.G.C. circuit has the advantage that the *RC*-constant of the filtering circuits can be kept smaller than that of a conventional circuit, so that rapid variations in the signal received (e.g. airplane fluttering) are smoothed out more efficiently.

The keyed A.G.C. circuit is furthermore less sensitive to interference due to the circuit being inoperative during the major part of the line scanning period.

In a conventional circuit the I.F. signal of the receiver is applied to a detector having a large *RC*-constant, so that the output voltage - the A.G.G. voltage - is equal to the peak value of the I.F. signal. The detector and filter circuits cannot be given a small *RC*-constant because the picture content and the frame pulses might then influence the magnitude of the A.G.C. voltage.

In the keyed A.G.C. circuit which will be described, the level of the tops of the synchronising pulses of the video signal is compared with a potential having an almost constant value. During the intervals between the synchronising pulses the A.G.C. circuit is inactive. The filtering circuits can, therefore, have a small *RC*-constant, since the picture content and the frame pulses cannot possibly influence the value of the A.G.C. voltage.

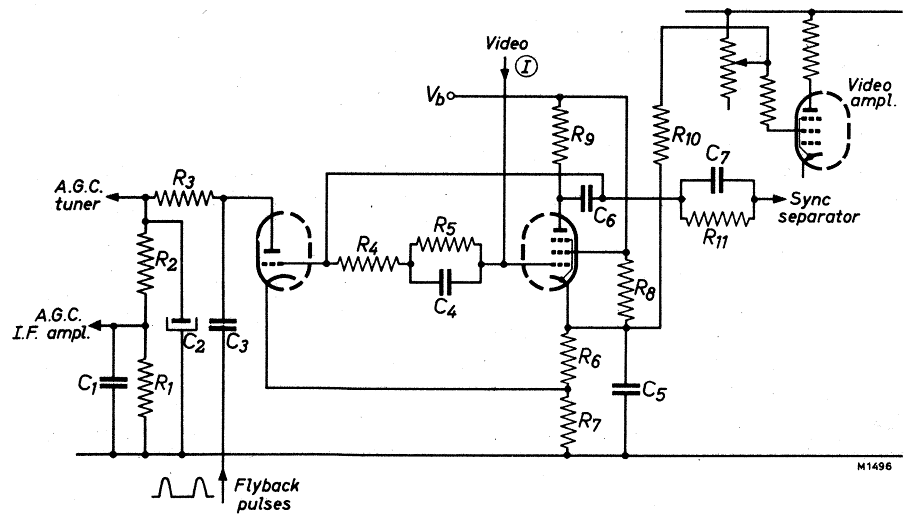

Fig.24. Keyed A.G.C. and noise inverter circuit with the PCF 80.

PARTS LIST:			
	R_1 = 0.39 MΩ	R_7 = 56 kΩ (½ W, ±5%)	C_2 = 0.47 μF
	R_2 = 0.15 MΩ	R_8 = 47 kΩ (½ W, ±5%)	C_3 = 1500 pF
	R_3 = 0.22 MΩ	R_9 = 0.15 MΩ	C_4 = 150 pF
	R_4 = 8.2 kΩ	R_{10} = 0.22 MΩ	C_5 = 0.27 μF
	R_5 = 22 kΩ	R_{11} = 8.2 MΩ	C_6 = 0.1 μF
	R_6 = 10 kΩ	C_1 = 1500 pF	C_7 = 820 pF

All resistors should be $\frac{1}{4}$ W unless a different wattage is indicated.

The A.G.C. circuit - combined with a noise inverter circuit - is represented in Fig.24. The triode section of a PCF 80, which operates in this circuit as an A.G.C. tube, has a cathode voltage V_c that is determined by the voltage divider R_6 - R_7 - R_8, apart from a small correction voltage supplied by the contrast control of the video amplifier. The voltage V_c is so high that the tube is biased far beyond cut-off. The anode of the tube is at a negative potential except during the occurrence of the flyback pulses, which are taken from the line output transformer and applied to the anode of the A.G.C. tube. Anode current can flow only during these pulses. The video signal taken from the output of the video amplifier is fed to the grid of the tube via the resistors R_4 and R_5. Fig.25 represents the video signal at the grid of the A.G.C. tube; this figure reveals that anode current flows only when the synchronising pulses exceed the cut-off voltage of the tube. At the time that anode current flows, the triode operates as a peak detector of the flyback pulses; these occur at the anode at the same instants at which the synchronising pulses occur at the grid. Detection of the flyback pulses results in C_3 being charged, so that the anode gets a mean negative potential which is used as A.G.C. voltage. The magnitude of this negative potential depends obviously on the extent to which the tops of the synchronising pulses exceed the cut-off voltage of the A.G.C. tube.

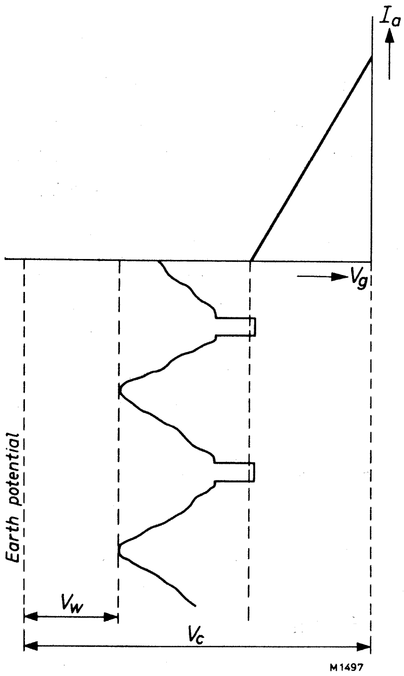

Fig.25. Video signal at the grid of the A.G.C. tube.
V_c = cathode voltage of the A.G.C. tube.
V_w = white level of the video signal.

As will appear from the description of the noise inverter, it is necessary to prevent the synchronising pulses in the video amplifier from being clipped. The A.G.C. circuit has therefore been so designed that it acts rapidly enough to ensure that the level of the tops of the synchronising pulses at the input of the video amplifier remains constant at a given setting of the contrast control, and that a small distance is always maintained between the cut-off voltage of the video amplifier tube and the tops of the synchronising pulses.

The cathode voltage V_c of the A.G.C. tube is so chosen that anode current flows in this tube when the video signal at the input of the video amplifier becomes so large that the synchronising pulses almost reach the cut-off voltage of the amplifier tube (cf. Fig.18).

It can be seen from Fig.25 that the white level V_w of the video signal influences the magnitude of the A.G.C. voltage. In fact, when the amplitude of the video signal is constant, a decrease of V_w reduces the current flowing in the A.G.C. tube and vice versa. When, as described in the preceding section, the setting of the contrast control of the video amplifier is modified so that the white level V_w at the cathode follower lowers, the A.G.C. voltage decreases, and the gain of the R.F. and I.F. amplifiers increases. Turning down the contrast control results in a larger A.G.C. voltage and a smaller R.F. and I.F. gain.

Fig.25 also reveals that when the antenna signal increases without the setting of the contrast control being changed, so that V_w remains constant, the A.G.C. voltage becomes larger and vice versa. In addition to the performance of the A.G.C. tube as part of the contrast control, the triode also operates as a normal A.G.C. tube. Fig.26 shows the voltages at the A.G.C. tube, which are all measured with respect to earth. The cathode voltage V_c of the tube is highly positive; the slightly lower cut-off voltage $V_{g\,co}$ of the tube is indicated by a dotted line.

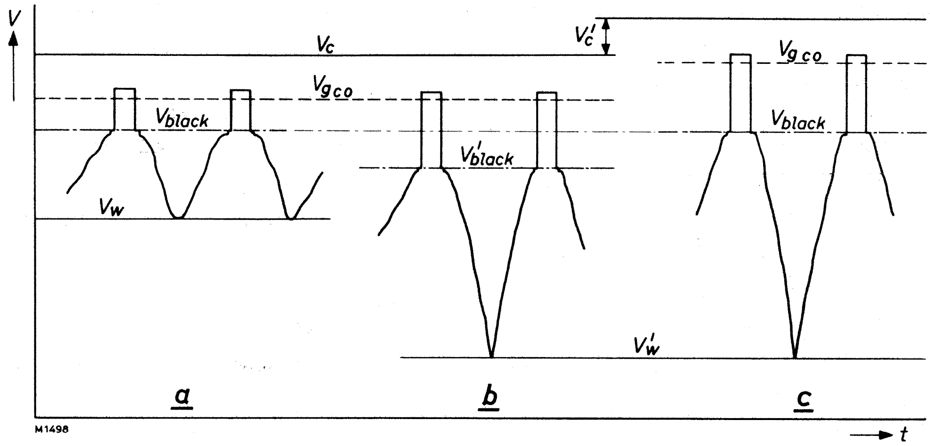

Fig.26. Video signal at the grid of the A.G.C. tube. (*a*) At minimum contrast of the video amplifier.

V_c = cathode voltage of the A.G.C. tube,
$V_{g\,co}$ = cut-off voltage of the A.G.C. tube,
V_{black} = black level of the video signal,
V_w = white level of the video signal.

(*b*) At maximum contrast of the video amplifier, and without a correction voltage being applied to the cathode of the A.G.C. tube. The black levels of the signals of the Figs (*a*) and (*b*) have different potentials. (*c*) At maximum contrast of the video amplifier and a correction voltage being applied to the cathode of the A.G.C. tube. The black levels of the signals in the Figs (*a*) and (*c*) are equal.

Fig.26*a* represents the video signal at the grid at minimum contrast, which comprises the white level V_w due to the d.c. coupling between the cathode follower and the A.G.C. tube; V_w is high at minimum contrast. The tops of the synchronising pulses just exceed the cut-off voltage of the A.G.C. tube, so that a small current flows during the pulses. The performance of the circuit will now be considered for the case that no correction voltage from the contrast control is applied to the cathode of the A.G.C. tube.

When the contrast control is adjusted to maximum, the white level of the cathode follower output signal decreases to V_w'. Fig.25 reveals that the A.G.C. voltage then decreases. The gain of the R.F. and I.F. amplifiers becomes so much larger that an equilibrium is again obtained between the gain of these amplifier stages and the A.G.C. voltage produced. The new situation at the A.G.C. tube is represented in Fig.26*b*. The A.G.C. voltage is obviously smaller in this case than at minimum contrast. Since, however, the current peaks in the A.G.C. tube are very small in both conditions, the levels of the synchronising pulses are almost equal at maximum and at minimum contrast. Figs 26*a* and *b* reveal that the black levels of the signals, V_{black} and V_{black}' differ considerably. Since the video signal at the grid of the A.G.C. tube is the same as that at the cathode of the picture tube (Fig.18), changes in the setting of the contrast control would result in a variation of the brightness of the picture.

These unwanted brightness variations are avoided in the circuit described by applying to the cathode of the A.G.C. tube a correction voltage that varies with the setting of the contrast control. When the contrast is increased to maximum, an additional positive voltage is applied to the cathode of the A.G.C. tube, the value of which depends on the voltage divider R_{10}, R_6 and R_7 (Fig.24). This additional cathode voltage, which is indicated by V_c' in Fig.26c, results in an extra increase of the gain of the R.F. and I.F. stages, so that the black level remains the same both at minimum and at maximum contrast (Figs 26*a* and *c* respectively), independent of the adjustment of the contrast control.

It should be borne in mind that the cut-off voltage of the A.G.C. tube depends on its anode voltage, i.e. on the amplitude of the flyback pulses applied to the anode. The pulses must be sufficiently large to exceed the cathode voltage of the triode. In the actual circuit the amplitude of the pulses is 200 to 250 V_{p-t-p}. Large deviations from this value will result in an improper level of the synchronising pulses and may upset the noise inverter action.

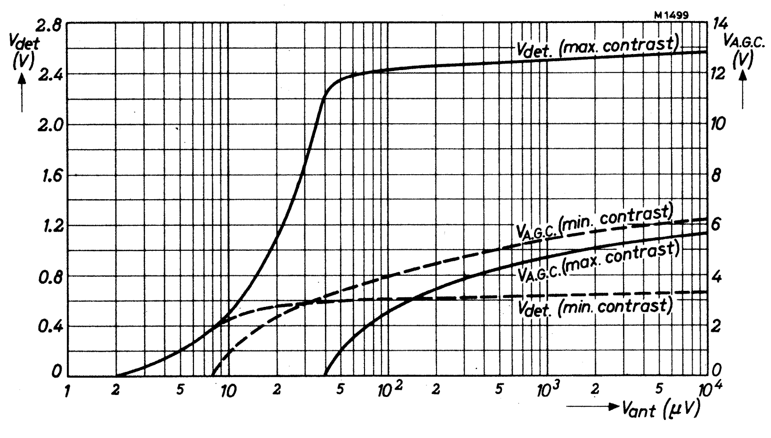

Fig.27. Curves of the A.G.C. and video detector voltages (V_{AGC} and V_{det} respectively) as functions of the antenna voltage V_{ant} for maximum and minimum contrast respectively.

The A.G.C. voltages for the tuner and the I.F. stages are taken from different tappings of the anode resistor. Fig.27 represents curves of the A.G.C. and detector voltages, V_{AGC} and V_{det} respectively, as functions of the antenna voltage V_{ant}, for maximum and minimum contrast of the video amplifier. These curves were measured in a receiver the tuner of which was equipped with a PCC 84 and a PCF 80; the three-stage I.F. amplifier was equipped with EF 80's and I.F. transformers. Two of the three I.F. amplifier stages were controlled by the A.G.C. voltage. As can be seen from Fig.24, the A.G.C. voltage for the I.F. amplifier is 0.7 of the A.G.C. voltage for the tuner.

The grids of the A.G.C. tube and the sync separator are interconnected. If no special measures were taken the video signal at the grid of the sync separator would be distorted due to the presence of R_4 and R_5 and its input capacitance, resulting in the well-known effect of "pulling on white". This is avoided by shunting R_5 with the capacitor C_2, thus improving the frequency response of the network connected in series with the grid of the sync separator.

NOISE INVERTER CIRCUIT

The pentode section of the PCF 80 operates as a noise inverter. This improves the stability of the synchronisation and A.G.C. circuits by reducing the interference in the video signal fed to these circuits. In the noise inverter the interference pulses that have an amplitude which exceeds that of the synchronising pulses, are

a b

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Fig.28. (*a*) Video signal with positive-going interference pulses. (*b*) After the output signal of the noise inverter has been added to a video signal, the polarity of the interference pulses is reversed.

separated from the video signal. After the separated pulses have been amplified, they are added again to the video signal with opposed polarity. Fig.28*a* shows an oscillogram of a video signal on which interference pulses are superimposed, and Fig.28*b* a video signal in which the polarity of the interference pulses has been reversed by adding the output pulses of the noise inverter to the video signal.

CIRCUIT DESCRIPTION

The cathode voltage of the noise inverter tube (Fig.24) is so high that the tube is biased beyond cut-off, similar to the A.G.C. tube. The video signal is applied to the control grid, but due to the high cathode voltage, no anode current flows except at the occurrence of the interference pulses (Fig.29). These pulses give rise to negative-going pulses at the anode of the tube which are added

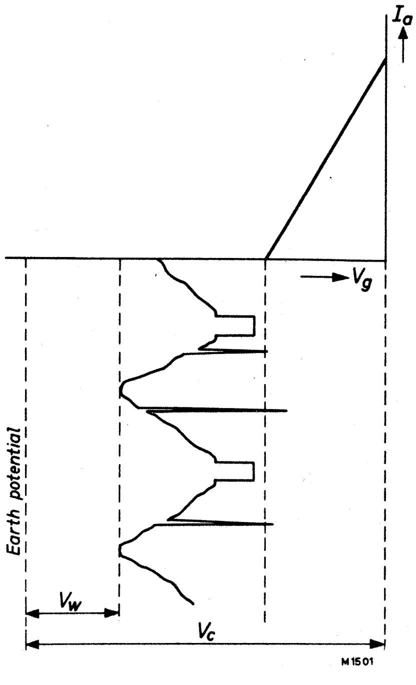


Fig.29. Video signal with interference pulses at the grid of the noise inverter.
V_c = cathode voltage of the noise inverter tube,
V_w = white level of the video signal.

again to the video signal, which is applied to the A.G.C. and sync separator tubes. In this video signal the interference pulses are therefore negative-going, so that both tubes are blocked when such pulses occur. Consequently, interference pulses do not appear in the output signal of the sync separator so that the stability of the synchronisation of the line and frame timebases is increased. The A.G.C. voltage is furthermore prevented to increase at the occurrence of the interference pulses, so that the gain of the R.F. and I.F. amplifiers is more stable.

It will be clear from this description why the A.G.C. circuit must be so designed that the tops of the synchronising pulses do not reach the cut-off point of the video amplifier tube, for in that case the interference pulses would be clipped and the noise inverter rendered inoperative.

The cathode voltage of the noise inverter tube must be higher than that of the A.G.C. tube, because the tops of the synchronising pulses should exceed the cut-off point of the A.G.C. tube (Fig.25), whereas the synchronising pulses in the noise inverter circuit may not reach the cut-off point.

It will also be clear from the description of the A.G.C. circuit that the cathode voltage of the noise inverter must increase and decrease with the cathode voltage of the A.G.C. tube. The tops of the synchronising pulses would otherwise exceed the cut-off voltage of the noise inverter, since the amplitude of the video signal becomes larger than if no compensation were applied. The cathode voltage of the noise inverter tube therefore varies also with the setting of the contrast control. The cathode of the A.G.C. tube, which requires a lower voltage, is connected to a tapping of the cathode resistor of the noise inverter tube.

In the circuit described the cathode voltage of the noise inverter varies between 101 V and 108 V, depending on the setting of the contrast control.

Since the output signal of the noise inverter is applied to the grid of the A.G.C. tube, two resistors, R_4 and R_5, are inserted in its grid circuit; these prevent the output signal of the noise inverter being fed back to its control grid.

PERFORMANCE OF THE CIRCUIT

The improvement of the stability of the timebases with regard to interference pulses is obviously greatest for the frame timebase. The line timebase is normally provided with a flywheel circuit, which renders it already almost insensitive to interference, although some improvement is noticeable also in this circuit.

The improvement of the stability of the frame synchronisation has been evidenced by the reception of a transmitter at a distance of 120 km. The synchronisation of a receiver without noise inverter was completely lost when a strong interference signal was generated by means of an electric razor. The frame synchronisation of the receiver with a noise inverter was lost only at those instants at which the interference pulses coincided exactly with the frame synchronising pulses.

The noise inverter circuit described has the following advantages compared with other types of noise inverter circuits:

a) No hand-controlled level setting is required, since the A.G.C. circuit stabilises - irrespective of the setting of the contrast control - the voltage difference between the level of the tops of the synchronising pulses and the level at which the noise inverter starts operating.

b) The circuit described not only improves the stability of the synchronisation but it also renders the A.G.C. circuit completely insensitive to interference.

The tolerances of the resistors of the voltage divider R_6, R_7 and R_8 should preferably be smaller than 5% to obtain a good performance of the A.G.C. and noise inverter circuit.

No data are quoted of the circuits because these depend on the amplitude of the antenna signal and on the amount of noise in this signal.

SYNC SEPARATOR AND PHASE SPLITTER IN FLYWHEEL CIRCUITS

The pentode section of the PCF 80 is particularly suitable for operation as a sync separator due to its short grid base and steep $I_a = f(V_{g1})$ characteristic. In this application the tube operates with a low screen-grid voltage and a low anode voltage to clip the pulses. Moreover, noise signals which may be present at the tops of the synchronisation pulses are reduced by the combined effect of grid current occurring and the mutual conductance of the tube being very small when the anode voltage drops below the knee of its $I_a = f(V_a)$ characteristic at zero grid bias.

Good operation of the separator will be obtained with a screen-grid voltage of 25 to 45 V. The anode of the tube is connected to a tapping on the cathode resistor of the triode section, which operates as a phase splitter (Fig.30). An anode voltage of 40 V is thus obtained.

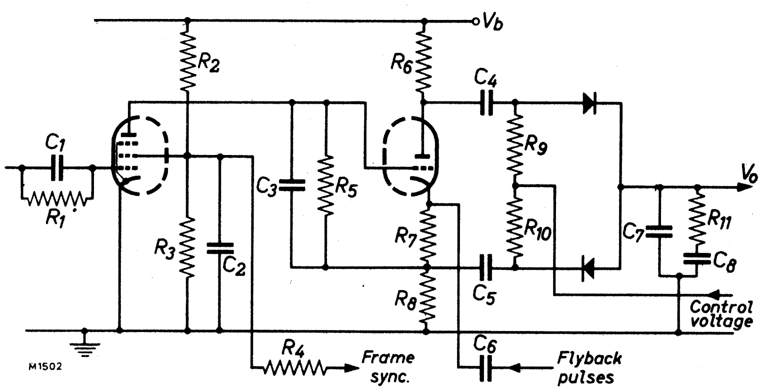


Fig.30. A sync separator and phase splitter for flywheel circuits with the PCF 80.

PARTS LIST:

R_1	= 8.2 MΩ	R_8	= 4.7 kΩ (1 W)	C_3	= 820 pF
R_2	= 0.18 MΩ	R_9	= 0.56 MΩ	C_4	= 1500 pF
R_3	= 56 kΩ	R_{10}	= 0.56 MΩ	C_5	= 1500 pF
R_4	= 33 kΩ	R_{11}	= 39 kΩ	C_6	= 120 pF
R_5	= 3.3 kΩ	C_1	= 820 pF	C_7	= 10.000 pF
R_6	= 5.6 kΩ (1 W)	C_2	= 2200 pF	C_8	= 0.1 μF
R_7	= 82 Ω				

All resistors should be $^1/_4$ W unless a different wattage is indicated.

The line synchronisation pulses, which are separated from the video signal, are applied to the phase splitter via an anode resistor (R_5) of 3.3 kΩ, which is connected between the grid and the cathode of the latter tube.

The frame synchronising signal is derived from the screen grid of the separator tube, and the lower part of the screen-grid potentiometer is therefore shunted by a capacitor (C_2) of 2200 pF, so that the line pulses are strongly attenuated. The line pulses cause only a small ripple on the frame synchronising signal, as may be seen from Fig.31, which represents the signal at the screen grid.

When the sync separator is preceded by a noise inverter as described above, the frame synchronisation appears to be very stable and almost insensitive to interference that may be present in the video signal.

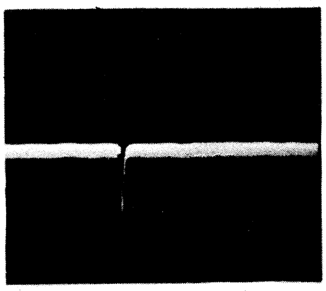
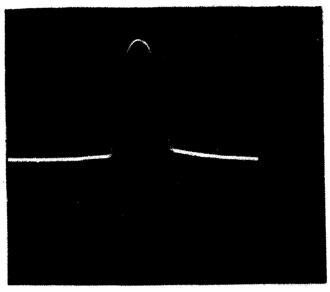


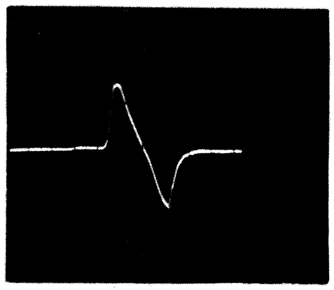
Fig.31. Frame synchronising pulses at the screen grid of the sync separator.

If the pentode section of the PCF 80 is used as a sync separator the triode section is available for use as a phase splitter for the line synchronising pulses that are to be applied to a flywheel circuit. In the circuit shown in Fig.30 the triode section is moreover used as an amplifier of the differentiated flyback pulses, which must also be applied to the flywheel circuit.

The anode resistor R_6 and the cathode resistor R_7 of the phase splitter are so chosen that the amplitudes of the synchronising pulses at both electrodes are almost equal. Small differences in the amplitudes of the pulses can be counteracted by applying a suitable control voltage to the flywheel circuit.



a



b

Fig.32. (*a*) Oscillogram of a flyback pulse at the line output transformer. (*b*) Differentiated flyback pulse at the cathode of the phase splitter tube.

The flyback pulses, the phase of which is compared with the synchronising pulses in the flywheel circuit, are applied to the cathode of the phase splitter. To obtain a voltage that is more or less sawtooth-shaped instead of pulse-shaped, the pulses are fed to the cathode via a differentiating network consisting of C_6 and the cathode resistors R_7 and R_8. Figs 32*a* and *b* show oscillograms of a flyback pulse at the line transformer and at the cathode of the phase splitter respectively. The differentiated pulse shown in Fig.32*b* controls the anode current of the phase splitter tube in such a way that shape and phase of the sawtooth voltages at the anode and at the cathode are identical.

Figs 33*a* and *b* show oscillograms of the pulses produced at the anode and at the cathode of the phase splitter when synchronising pulses are applied to the control grid, the cathode of the tube having been disconnected from the line output transformer. The pulses have an integrated shape due to the anode resistor of the sync separator being shunted by C_3.

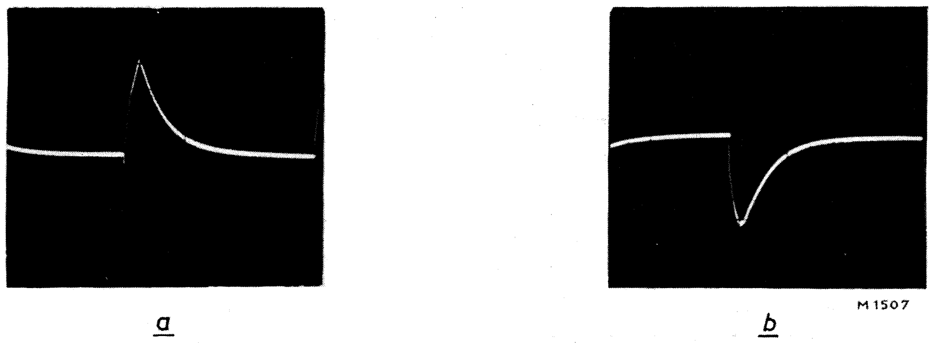


Fig.33. Oscillograms of the synchronisation pulses at the anode (*a*) and at the cathode (*b*) of the phase splitter tube, without flyback pulses being applied to the tube.

Fig.34 displays oscillograms of the signals, (*a*) at the anode and (*b*) at the cathode of the phase splitter tube. These signals are formed by adding the differentiated flyback pulses to the synchronising pulses.

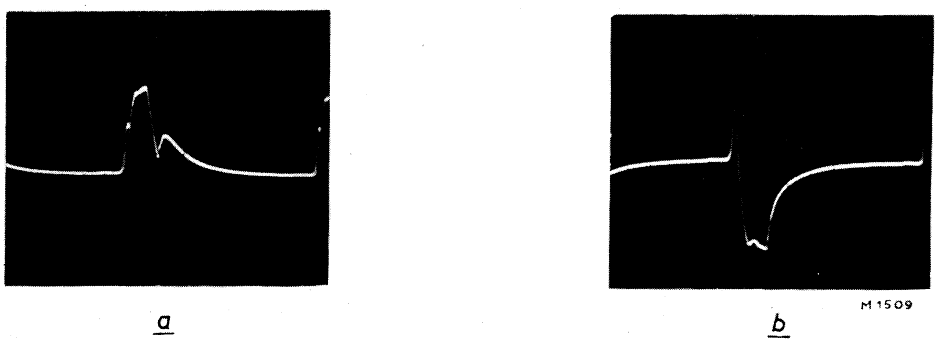


Fig.34. Oscillograms of the sum voltages of the synchronisation pulses and the flyback pulses at the anode (*a*) and at the cathode (*b*) of the phase splitter tube.

The particular way in which the synchronising pulses are applied to the phase splitter tube is a simple solution of the problem how to obtain a low anode voltage for the sync separator tube, and how to control the phase splitter simultaneously by two signals. Compared with a conventional coupling circuit, several components have been saved, such as a coupling capacitor, a grid leak resistor, etc.

The anode resistor R_5 is shunted by the capacitor C_3 to obtain coincidence of the blanking pulses with the flyback pulses. This arrangement is chosen for the following reasons.

The duration of the flyback is made as large as possible to minimize the peak voltages and the losses in the line output stage. Care must therefore be taken to ensure that the flyback occurs exactly within the line blanking periods of the video signal.

The line synchronising pulses do not, however, occur in the centre of the blanking pulses, whereas this flywheel circuit (Fig.30) tends to make the centre of the synchronising pulses coincide with the centre of the flyback pulses (Fig.35*a*). It appears from this figure that in that case the flyback pulse will fall partly outside the blanking period. The picture will then be partially lost at the right. By integrating the synchronising pulse by means of C_3, the centre of the pulse is delayed to such an extent that the flyback pulse falls exactly within the blanking period (Fig.35*b*).

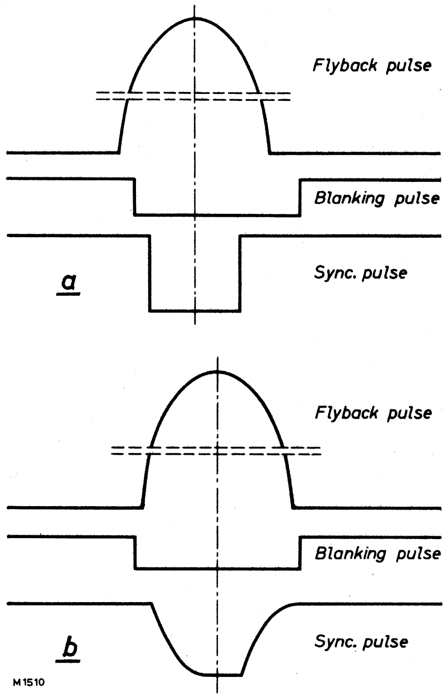


Fig.35. (*a*) Due to the operation of the flywheel circuit the flyback pulse falls outside the blanking period. (*b*) The synchronisation pulse being integrated, the flyback pulse falls within the blanking period.

Below several figures concerning the adjustment of both the pentode section and the triode section of the PCF 80 are given.

			Pentode section	Triode section	
Supply voltage....................	V_b	=	180	180	V
Mean anode voltage................	V_a	=	43	127	V
Mean anode current................	I_a	=	0.14	9.5	mA
Mean screen-grid voltage..........	V_{g2}	=	38		V
Mean screen-grid current..........	I_{g2}	=	0.10		mA
Mean cathode voltage..............	V_k	=	0	45	V

These values hold for the case the horizontal timebase is synchronised. If the synchronisation of the timebase is upset, only small differences will occur in the figures given.

HORIZONTAL TIMEBASE OSCILLATOR AND REACTANCE CONTROL TUBE

In a line timebase a sine oscillator operating as a line sawtooth generator combined with a reactance control tube is superior to a multivibrator because the former has a much better frequency stability. The difference in performance of these circuits is due to the fact that the frequency of a sine oscillator is almost exclusively determined by the tuned circuit, whereas that of a multivibrator depends to a large extent on the characteristics of the tubes, even when the multivibrator is provided with a stabilising *LC*-circuit.

The better frequency stability of a sine oscillator with a reactance control tube involves the necessity of a larger control voltage to be supplied by the preceding flywheel circuit. Since, however, in contrast to a multivibrator, the reactance tube does not run into grid current, the larger control voltage required can easily be obtained.

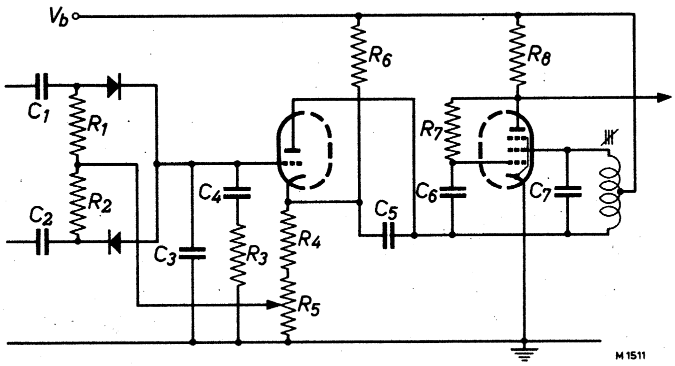


Fig.36. Sine oscillator operating as a line sawtooth generator, and reactance control tube with the PCF 80.

PARTS LIST:

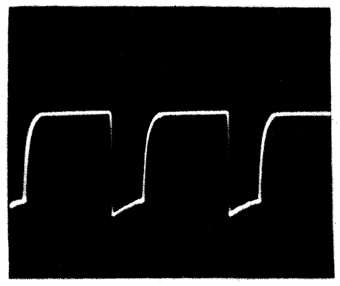
R_1 = 0.56 MΩ	R_6 = 33 kΩ (1 W)	C_3 = 10.000 pF
R_2 = 0.56 MΩ	R_7 = 5.6 MΩ	C_4 = 0.1 μF
R_3 = 39 kΩ	R_8 = 56 kΩ	C_5 = 820 pF
R_4 = 470 Ω	C_1 = 1500 pF	C_6 = 100 pF
R_5 = 5 kΩ (variable)	C_2 = 1500 pF	C_7 = 1500 pF (mica)

All resistors should be $\frac{1}{4}$ W unless a different wattage is indicated.

Fig.36 represents the circuit in which the pentode section of a PCF 80 operates as a sine oscillator, whilst its triode section functions as a reactance control tube.

The sine oscillator is of the Hartley type, and is adjusted to class C, so that the anode voltage of the tube consists of negative peaks of high amplitude (Fig.37), which can control the line output stage directly. It is also possible to connect a capacitor between anode and earth to obtain a sawtooth-shaped output voltage.

Fig.37. Anode signal of the pentode section of the PCF 80 operating as a sine oscillator in a line sawtooth generator circuit.



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The frequency of the oscillator can be adjusted roughly by means of a variable core in the oscillator coil; the inductance of the coil can be varied in this way between 30 mH and 75 mH. The tapping ratio of the coil is approximately 3, the signal at the control grid of the pentode thus being about 3 times that at the screen grid. To prevent squegging of the oscillator, which may be caused by the coupling between the coils being so tight, the control grid of the tube is connected to the anode via a resistor of 5.6 MΩ (R_7). The low-frequency feedback thus obtained ensures a completely stable operation of the oscillator.

Frequency deviations of the oscillator are checked by means of the triode section of the PCF 80. The latter operates as a reactance control tube and is shunted across the tuned circuit of the oscillator.

Such a control tube may be either an inductive or a capacitive tuning device. The choice is determined, amongst others, by the polarity of the control voltage that is obtained from the flywheel circuit at a given phase shift between the synchronising and flyback pulses.

In view of the frequency stability with respect to mains voltage fluctuations the control tube should form an inductive load, so that the capacitive load exerted on the tuned circuit by the oscillator tube itself, is compensated.

The parallel connection of the capacitor C_5 (820 pF) and the internal resistance of the reactance control tube forms a complex load on the tuned circuit of the oscillator. The reactive part of this load depends on the mutual conductance of the reactance tube, and since the tuning frequency of the oscillator depends, in turn, on this parallel reactance, the frequency can be influenced by a variation of the mutual conductance of the reactance tube. This variation is obtained by modifying its bias, such as caused by the direct voltage generated in the flywheel circuit when the frequency or the phase of the synchronisation pulses differ from those of the flyback pulses. The reactance tube circuit has been so designed that this direct voltage changes the parallel reactance of the sine oscillator to such an extent that its frequency is reduced or increased to the correct value.

The reactance tube also offers the possibility of fine adjustment of the oscillator frequency. As can be seen from Fig.36, a potentiometer R_5 is included in the cathode lead of the tube; a direct voltage is taken from this potentiometer and applied to the grid of the tube via the flywheel circuit. This direct voltage has the same effect as a direct voltage generated in the coincidence detector itself: it changes the frequency of the sine oscillator. The potentiometer R_5 thus acts as a horizontal hold control. A resistor of 470 Ω (R_4) is connected in series with the potentiometer R_5 to prevent the tube being overloaded in any position of the horizontal hold control. If necessary, the potentiometer may be mounted at some distance from the reactance tube, provided the leads to the potentiometer are screened.

The reactance tube circuit exerts a fairly heavy damping on the tuned circuit of the sine oscillator. It is possible to decrease this damping in various ways; the rise time of the pulses at the anode of the oscillator tube is thereby reduced. Experience reveals, however, that reduction of the damping does not noticeably improve the performance of the timebase, so that the low-cost circuit, as described above, will usually be preferred.

Measurements have been made on the susceptibility to interference of a receiver equipped with a phase splitter (Fig.31) and with the circuit described in this section. The receiver concerned was not provided with a noise inverter. For this measurement a video signal with interference pulses was applied to the sync separator (Fig.31). The pulse frequency was 1300 c/s and the duration of the pulses was approximately two line periods, i.e. 130 μsec. The synchronisation appeared to operate satisfactorily unless the amplitude of the interference pulses exceeded three times that of the video signal. With a small video signal the amplitude of the interference pulses could be increased to 5 - 6 times the video signal before synchronisation was upset.

It is recommended to use a mica capacitor (C_7) in the tuned circuit of the oscillator to prevent temperature variations from causing detuning.

The oscillator coil is wave-wound on a former of fuller board with an outer diameter of 6 mm, and provided with a variable iron dust core. The data of the coil are:

number of turns	:	2000;
wire	:	enamelled copper wire, 0.12 mm, silk covered;
total inductance	:	variable between 30 and 75 mH.

The coil is tapped at 450 turns. The tapping is connected to the supply line; for class C operation of the oscillator, the largest number of turns must be between the supply line and the control grid.

CIRCUIT DATA

			Pentode section	Triode section
Anode voltage	V_a	=	135 V [1]	180 V
Anode current	I_a	=	0.8 mA [1]	0.5-4 mA [2]
Screen-grid voltage	V_{g2}	=	180 V	
Screen-grid current	I_{g2}	=	3.7 mA	
Control-grid voltage	V_{g1}	=	-56 V	1-34 V [2]
Cathode voltage	V_k	=	0 V	26-44 V [2]

[1]) This voltage (current) becomes smaller (larger) when an RC-network is added to the anode circuit to obtain a sawtooth-shaped output voltage.

[2]) Depending on the setting of R_5.

FRAME TIMEBASE CIRCUIT

The triode section of the PCF 80 can be used as a limiter of the frame synchronising pulses. The pentode part then remains available to function as a frame sawtooth multivibrator in combination with another tube. Fig.38 represents the diagram in which the other part of the multivibrator is formed by the triode section of the PCL 82. The pentode section of this tube is specially designed as a frame output tube for 90° deflection, whilst the triode is available for use as a frame blocking oscillator or as part of a frame multivibrator.

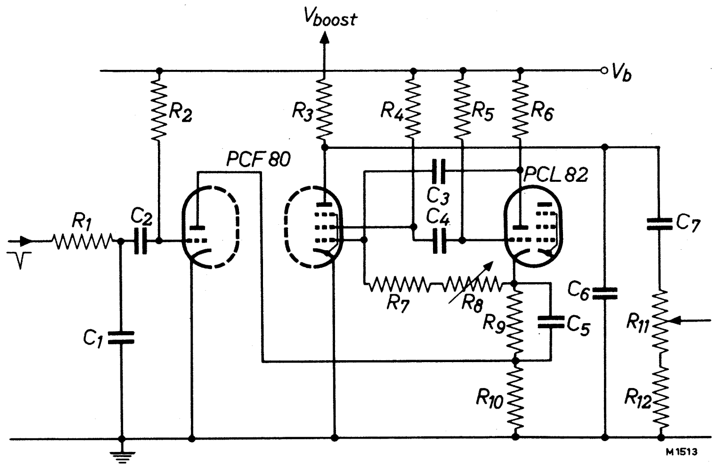


Fig. 38. Pulse limiter and frame timebase multivibrator with a PCF 80 and the triode section of a PCL 82.

PARTS LIST: R_1 = 33 kΩ	R_8 = 1 MΩ (variable)	C_1 = 470 pF
R_2 = 8.2 MΩ	R_9 = 18 kΩ (±5 %)	C_2 = 10.000 pF
R_3 = 2.2 MΩ	R_{10} = 8.2 kΩ (±5 %)	C_3 = 15.000 pF
R_4 = 10 kΩ (±5 %)	R_{11} = 1 MΩ (variable)	C_4 = 2200 pF
R_5 = 1 MΩ (±5 %)	R_{12} = 0.22 MΩ	C_5 = 22.000 pF
R_6 = 27 kΩ (±5 %)		C_6 = 33.000 pF
R_7 = 2.2 MΩ		C_7 = 0.1 μF

All resistors should be $^1/_4$ W unless a different wattage is indicated.

Especially with respect to the multivibrator circuit, the diagram shown in Fig.38 offers several features which make it preferable to conventional timebase circuits.

The frame synchronising pulses, which are taken from the sync separator circuit are fed to the triode section of the PCF 80 via one or more integrating networks. The pulses must be negative-going and their amplitude must exceed the grid base of the triode at a very low anode voltage (approximately 10 V). Since the grid of the triode is connected to the supply line via a high-ohmic leak resistor R_2 (8.2 MΩ), grid current will limit the pulses, which will also be clipped since they partly exceed the cut-off voltage of the tube. Due to the clipping action of the tube the amplitude of the frame synchronising pulses remains constant, independent of the amplitude of the video signal, so that the stability of the frame synchronisation is improved. When no noise inverter is used in the synchronising circuit, noise peaks in the pulse signal will also be clipped by the triode.

The anode of the synchronising pulse limiter is connected to a tapping on the cathode resistor of the triode section of the PCL 82. The synchronising signal is thus applied to the multivibrator via the cathode impedance.

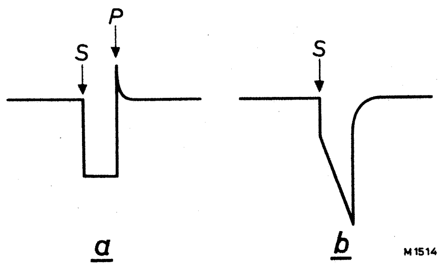


Fig.39. Cathode voltage of the triode section of the multivibrator. (*a*) cathode resistor not shunted; (*b*) cathode resistor R_9 shunted by C_5.

One of the cathode resistors is shunted by the capacitor C_5 to prevent synchronisation being brought about by the leading edge of the synchronising pulse coinciding with the trailing edge of the pulse generated by the multivibrator itself at the cathode of the triode. Fig.39*a* represents the cathode voltage which would occur if the cathode resistor were not shunted, Fig.39*b* displaying the cathode voltage when the cathode resistor is shunted. In Fig.39*a* the synchronisation of the multivibrator may occur either at point S or at point *P*. In the actual circuit (Fig.39*b*) synchronisation is only possible at the steep front edge of the cathode signal (point S). The capacitive shunt of the cathode resistor therefore contributes to the correct interlacing of the picture.

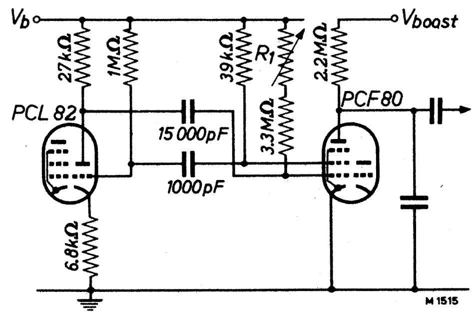


Fig.40. Conventional circuit of a frame timebase multivibrator with the pentode section of the PCF 80 and the triode section of the PCL 82.

The multivibrator is so designed that the performance of the circuit is almost completely independent of variations in the tube characteristics. Fig.40 shows a conventional frame multivibrator circuit as might be formed by means of the triode section of the PCL 82 and the pentode section of the PCF 80, the *RC*-coupling circuit between the tubes being so chosen that anode current flows in the pentode during short periods only. Fig.41 represents the resulting control-grid voltage of the pentode; the slope of the discharge current *B* of the coupling capacitor can be varied by means of the grid-leak resistor R_1, which serves as a frequency control. The frequency of the multivibrator, however, depends not only on the slope of the discharge curve, but also on the amplitude *A* of the control-grid voltage. The latter is determined by the anode resistor and the anode current of the triode of the PCL 82. When this tube is used in the circuit of Fig.40 a spread in the anode current,

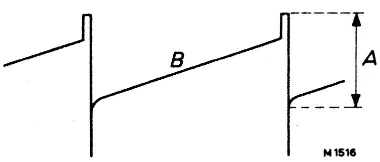


Fig.41. Control-grid voltage of the pentode section of the multivibrator shown in Fig.38.

which may amount to 2 mA, must therefore be taken into account. This is the largest spread which may be encountered between various samples consisting of new tubes and tubes which have reached their end-of-life condition. Since the anode current of an average tube in the circuit of Fig.40 is 4 mA it is clear that large frequency deviations will occur when this current varies by 2 mA. These frequency deviations may even be increased by tolerances in the components used. It appears to be impracticable to chose the value of R_1 so that the resulting frequency variations fall within its range.

The multivibrator circuit shown in Fig.38 is almost independent of variations in the tube characteristics. This has been achieved by giving the triode a large cathode resistor, the value of which is almost equal to that of the anode resistor (26 kΩ and 27 kΩ respectively). Furthermore, the grid leak resistor of the pentode is connected to the cathode of the triode. When the average value of the anode current of the triode increases, e.g. due to replacement of the tube, the amplitude of the anode signal (cf. A in Fig.41) increases likewise, as a result of which the frequency of the multivibrator tends to decrease. An increase of the anode current of the triode, however, causes also an increase of its cathode voltage, which is substantially equal to the increase of the anode signal. And since the grid leak resistor of the pentode is connected to the cathode of the triode, an increase of the cathode voltage results in a reduction of the bias of the pentode, which has the effect of increasing the frequency. The consequences of the increase of the signal at the anode of the triode and of the reduction of the pentode bias compensate each other almost entirely. The deviation of the multivibrator frequency, resulting from a variation in the anode current of the triode of 2 mA, is only 3% in this new multivibrator circuit. The corresponding deviation in a conventional circuit would exceed 20%.

Since anode current flows in the pentode section only during very short periods, variations in this current will not noticeably affect the frequency of the multivibrator.

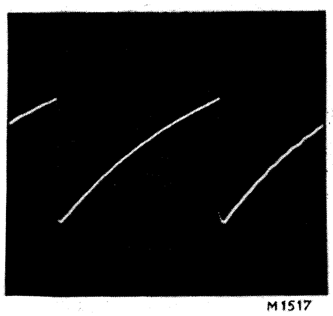


Fig.42. Oscillogram of the anode voltage of the pentode section of the multivibrator shown in Fig.36.

Fig.42 displays an oscillogram of the anode voltage of the pentode section of the PCF 80.

The RC-network between anode and earth comprises a potentiometer R_{11} which serves as an amplitude control. The signal taken from this potentiometer may be fed directly to the control grid of the frame output tube.

The data of the frame timebase are:

TRIODE SECTION OF THE PCF 80

Supply voltage	V_b	=	180 V
Anode voltage	V_a	=	9 V
Anode current	I_a	=	1.1 mA
Cathode voltage	V_k	=	0 V

PENTODE SECTION OF THE PCF 80

Supply voltage	V_b	=	180 V
Booster voltage	V_{boost}	=	725 V
Anode voltage	V_a	=	115 V
Anode current	I_a	=	0.28 mA
Screen-grid voltage	V_{g2}	=	176 V
Screen-grid current	I_{g2}	=	0.4 mA
Cathode voltage	V_k	=	0 V

TRIODE SECTION OF THE PCL 82

Supply voltage	V_b	=	180 V
Anode voltage	V_a	=	118 V
Anode current	V_a	=	2.3 mA
Cathode voltage	V_k	=	52 V

DATA OF THE PCF 80

HEATER DATA

Heating: indirect by a.c. or d.c.; series supply
Heater voltage.............. V_f = 9 V
Heater current.............. I_f = 300 mA

BASE CONNECTIONS AND DIMENSIONS

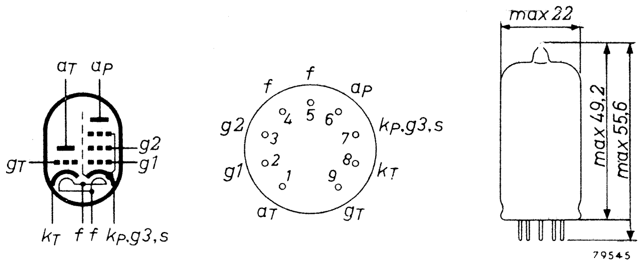


Fig.43. Base connections and dimensional drawing (dimensions in mm) of the PCF 80.

CAPACITANCES

Pentode section	*Triode section*	*Between pentode and triode section*
$C_{ag1} < 0.025$ pF	$C_{ag} = 1.5$ pF	$C_{aPaT} < 0.07$ pF
$C_{g1} = 5.5$ pF	$C_g = 2.5$ pF	$C_{gPaT} < 0.16$ pF
$C_a = 3.8$ pF	$C_a = 1.8$ pF	$C_{aPgT} < 0.02$ pF

TYPICAL CHARACTERISTICS

Pentode section

Anode voltage........................	V_a	=	170 V
Screen-grid voltage..................	V_{g2}	=	170 V
Control-grid voltage.................	V_{g1}	=	-2 V
Anode current........................	I_a	=	10 mA
Screen-grid current..................	I_{g2}	=	2.8 mA
Mutual conductance...................	S	=	6.2 mA/V
Internal resistance..................	R_i	=	0.4 MΩ
Amplification factor between screen grid and control grid............	μ_{g2g1}	=	47
Input resistance at 50 Mc/s..........	r_{g1}	=	10 kΩ
Equivalent noise resistance..........	R_{eq}	=	1.5 kΩ

Triode section

Anode voltage........................	V_a	=	100 V
Control-grid voltage.................	V_g	=	-2 V
Anode current........................	I_a	=	14 mA
Mutual conductance...................	S	=	5 mA/V
Amplification factor.................	μ	=	20

OPERATING CHARACTERISTICS

Pentode section as mixer

Anode voltage	V_a	= 170	170 V
Screen-grid voltage	V_{g2}	= 170	170 V
Cathode resistor	R_k	= 330	820 Ω
Anode current	I_a	= 6.5	5.2 mA
Screen-grid current	I_{g2}	= 2.0	1.5 mA
Control-grid current	I_{g1}	= 25	0 μA
External resistance between control grid and cathode	R_{g1}	= 0.1	0.1 MΩ
Oscillator voltage at the control grid	V_{osc}	= 3.5	3.5 V_{rms}
Conversion conductance	S_c	= 2.2	2.1 mA/V
Internal resistance	R_i	= 0.8	0.87 MΩ

Note: It is recommended to employ the triode section in a Colpitts type of oscillator and not in a Hartley circuit.

LIMITING VALUES

Pentode section

Anode voltage at zero anode current	V_{a0}	= max.	550 V
Anode voltage	V_a	= max.	250 V
Anode dissipation	W_a	= max.	1.7 W
Screen-grid voltage at zero screen-grid current	V_{g2_0}	= max.	550 V
Screen-grid voltage (cathode current $\leq$ 14 mA)	V_{g2}'	= max.	175 V
Screen-grid voltage (cathode current $\leq$ 10 mA)	V_{g2}''	= max.	200 V
Screen-grid dissipation	W_{g2}	= max.	0.5 W [1]
Cathode current	I_k	= max.	14 mA
Control-grid voltage (grid current = +0.3 μA)	$-V_{g1}$	= max.	1.3 V
External resistance between control grid and cathode with automatic bias	R_{g1}'	= max.	1 MΩ
with fixed bias	R_{g1}''	= max.	0.5 MΩ
Voltage between cathode and heater (*k* neg.; *f* pos.)	V_{kf}	= max.	100 V
(*k* pos.; *f* neg.)	V_{kf}	= max.	200 V [2]

Triode section

Anode voltage at zero anode current	V_{a0}	= max.	550 V
Anode voltage	V_a	= max.	250 V
Anode dissipation	W_a	= max.	1.5 W
Control-grid voltage (grid current = + 0.3 μA)	$-V_g$	= max.	1.3 V
Cathode current	I_k	= max.	14 mA
External resistance between control grid and cathode	R_g	= max.	0.5 MΩ
Voltage between heater and cathode (*k* neg.; *f* pos.)	V_{kf}'	= max.	100 V
(*k* pos.; *f* neg.)	V_{kf}''	= max.	200 V [2]

[1]) If W_a < 1.2 W the maximum value of W_{g2} may amount to 0.75 W.
[2]) d.c. component = max. 120 V.

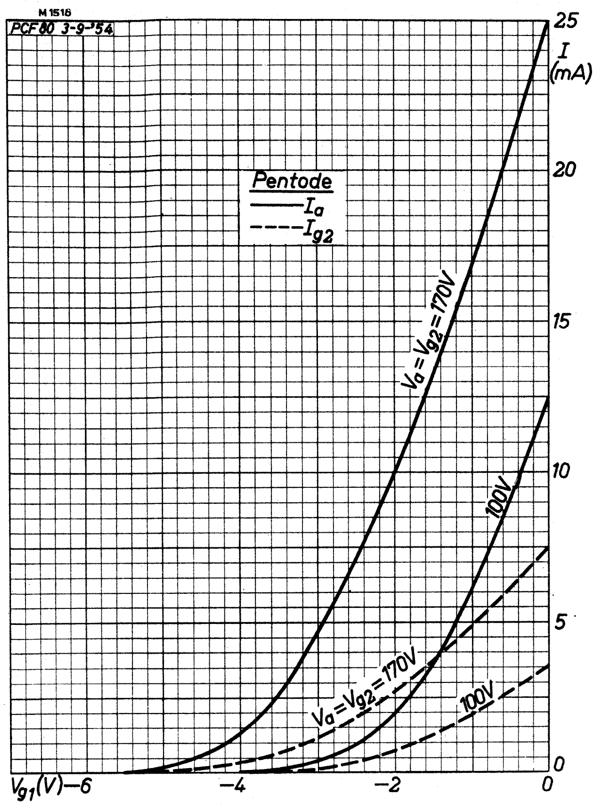


Fig.44. Anode current I_a and screen-grid current I_{g2} of the PCF 80 plotted as functions of the control-grid voltage V_{g1} with the anode voltage V_a and screen-grid voltage V_{g2} as parameter.

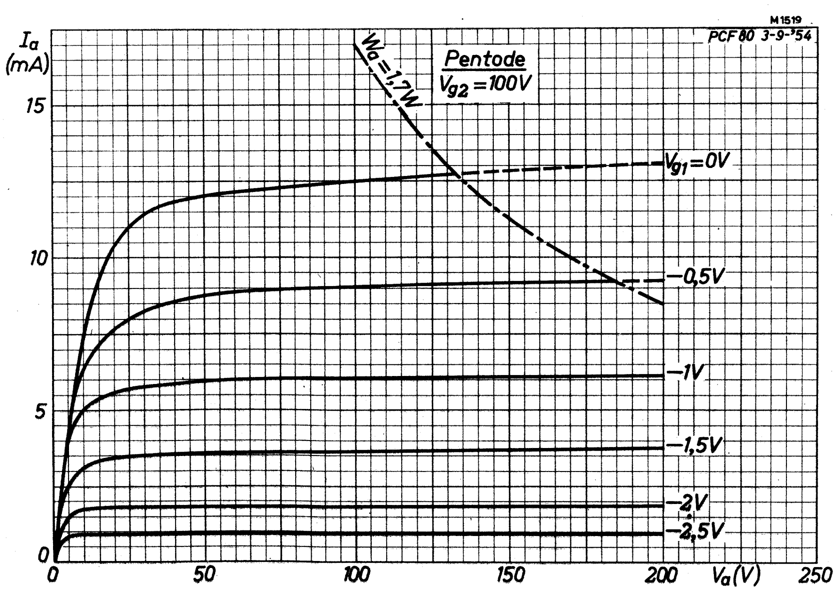


Fig.45. Anode current I_a of the PCF 80 plotted as a function of the anode voltage V_a with the control-grid voltage V_{g1} as parameter, at a screen-grid voltage V_{g2} of 100 V.

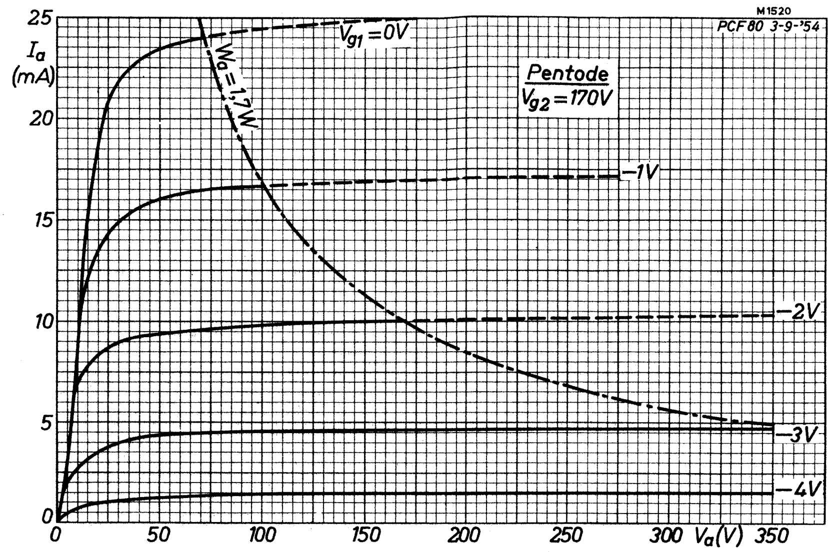


Fig.46. Anode current I_a of the PCF 80 plotted as a function of the anode voltage V_a with the control-grid voltage V_{g1} as parameter, at a screen-grid voltage V_{g2} of 170 V.

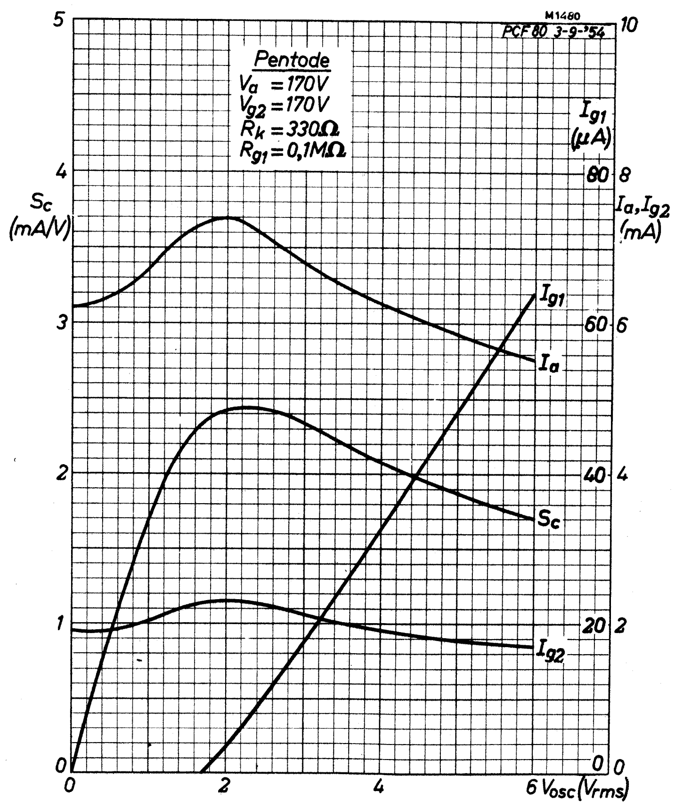


Fig.47. Conversion conductance S_c and control-grid, screen-grid and anode current I_{g1}, I_{g2} and I_a respectively, as functions of the oscillator voltage V_{osc} at a cathode resistor R_k of 330 Ω and a grid leak R_{g1} of 0.1 MΩ.

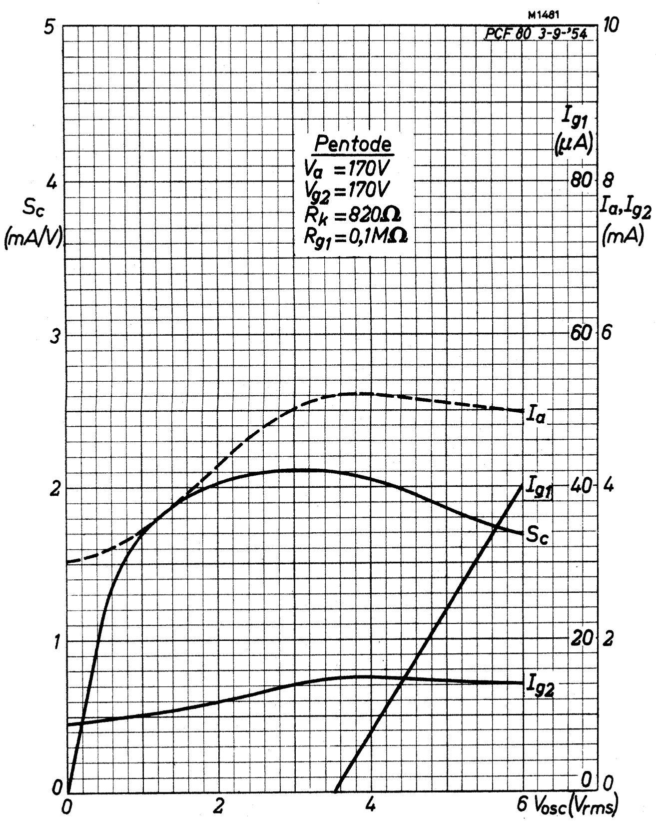
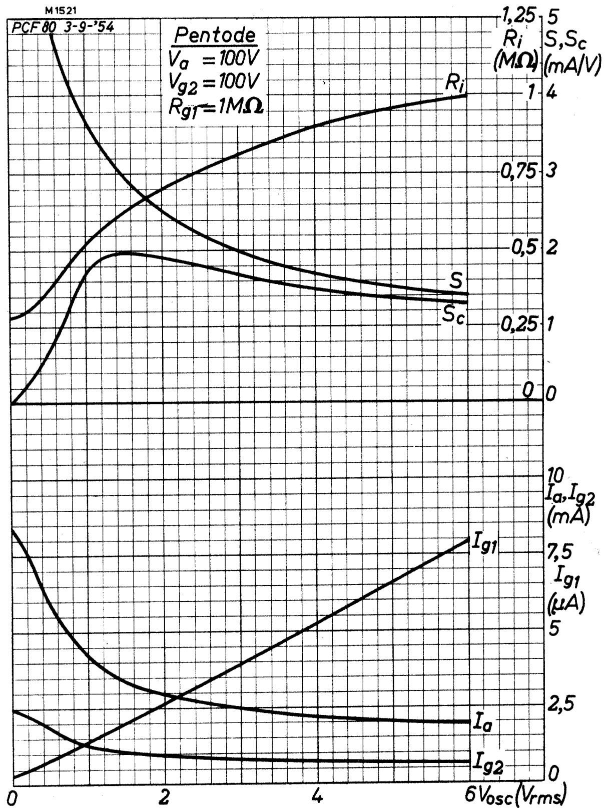
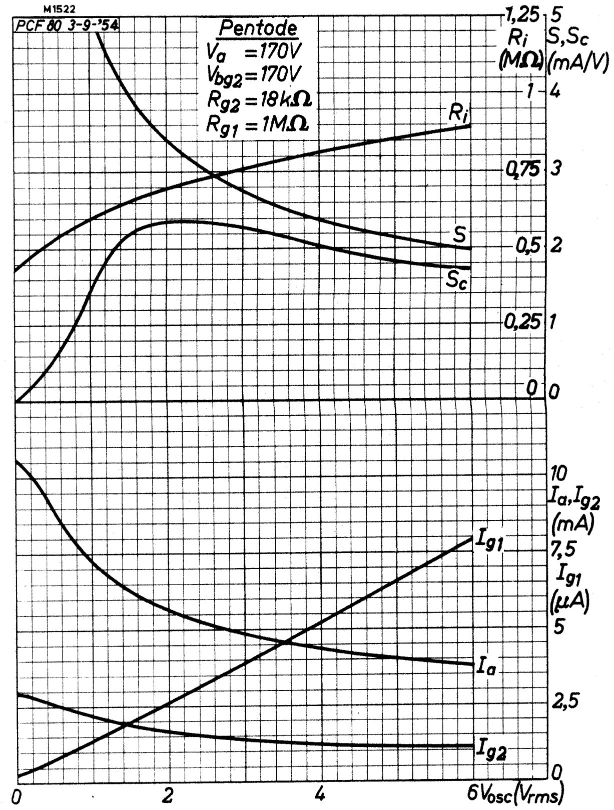


Fig.48. Conversion conductance S_c and control-grid, screen-grid and anode current I_{g1}, I_{g2} and I_a respectively, as functions of the oscillator voltage V_{osc} at a cathode resistor R_k of 820 Ω and a grid leak R_{g1} of 0.1 MΩ.





Figs 49 and 50. Internal resistance R_i, mutual conductance S, conversion conductance S_c and control-grid, screen-grid and anode currents I_{g1}, I_{g2} and I_a respectively of the PCF 80 operating as a self-oscillating mixer, as a function of the oscillator voltage V_{osc}, at a grid-leak resistor R_{g1} of 1 MΩ. Fig.49 (left) applies to an anode and screen-grid voltage V_a = V_{g2} of 100 V. Fig.50 (right) applies to an anode and supply voltage V_a = V_{bg2} of 170 V and a screen-grid dropping resistor of 18 kΩ.

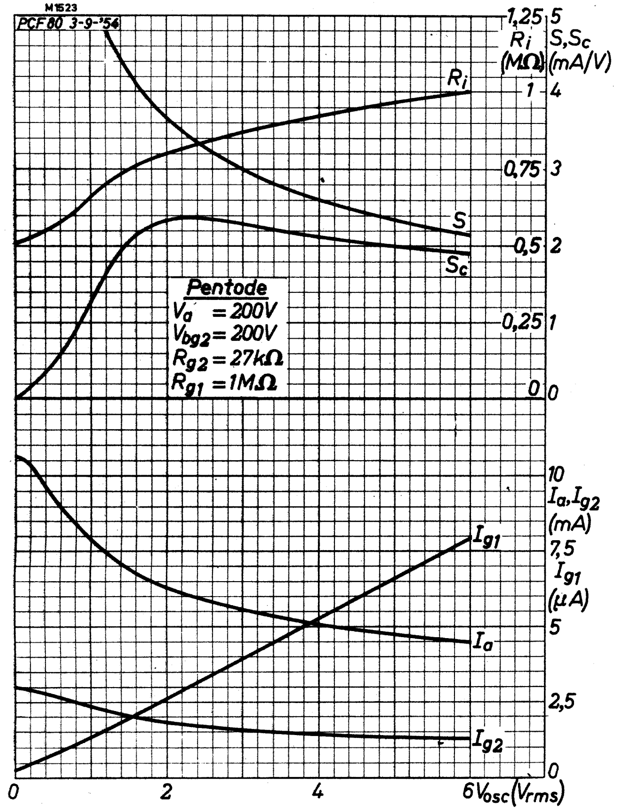


Fig.51. Internal resistance R_i, mutual conductance S, conversion conductance S_c and control-grid, screen-grid and anode currents I_{g1}, I_{g2} and I_a respectively of the PCF 80 operating as a self-oscillating mixer, as a function of the oscillator voltage V_{osc}, at a grid-leak resistor R_{g1} of 1 MΩ and an anode and supply voltage $V_a = V_{bg2}$ of 200 V, the screen-grid dropping resistor R_{g2} being 27 kΩ.

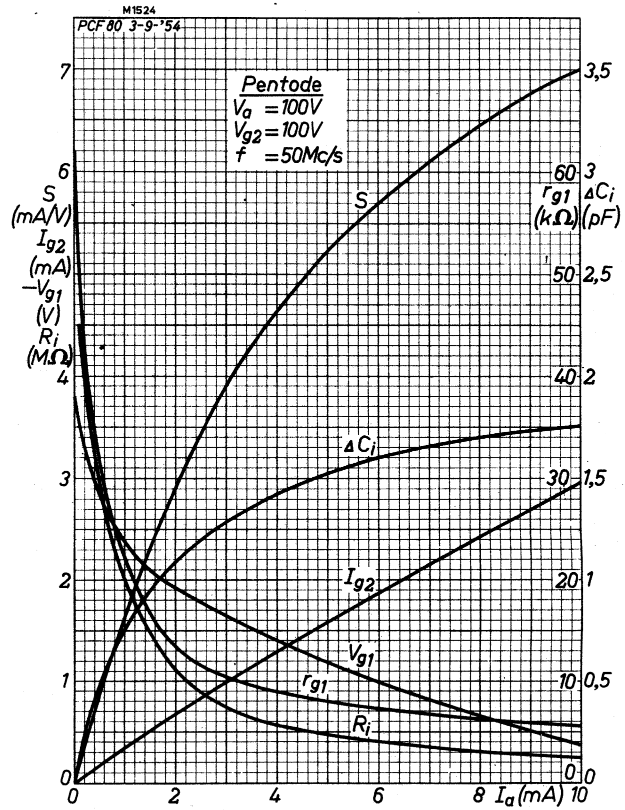


Fig.52. Mutual conductance S, variation of the input capacitance V_i screen-grid current I_{g2}, control-grid voltage V_{g1}, input impedance r_{g1} and internal resistance R_i of the PCF 80 as functions of the anode current I_a at a frequency f of 50 Mc/s and an anode and screen-grid voltage $V_a = V_{g2}$ of 100 V.

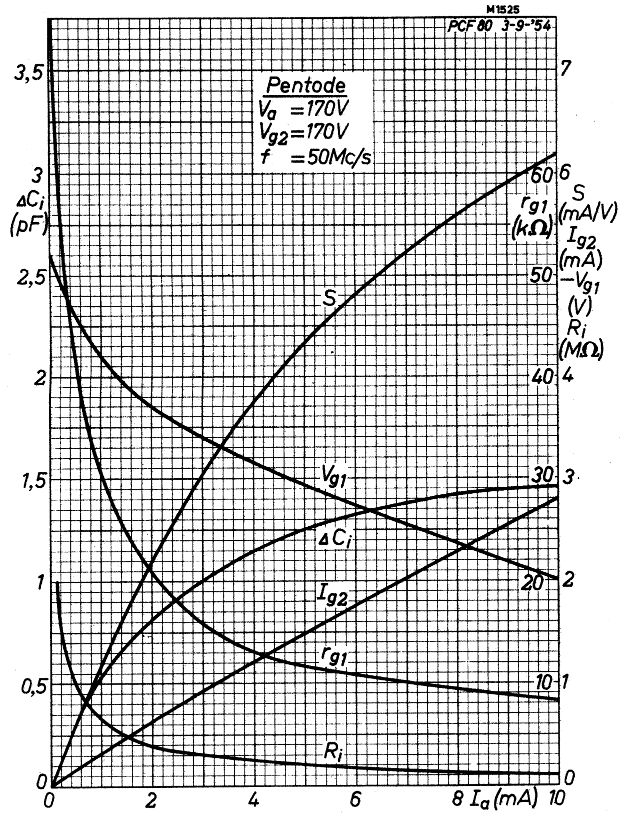


Fig.53. Mutual conductance S, variation of the input capacitance V_i, screen-grid current I_{g2}, control-grid voltage V_{g1}, input impedance r_{g1} and internal resistance R_i of the PCF 80 as functions of the anode current I_a at a frequency f of 50 Mc/s and an anode and screen-grid voltage $V_a = V_{g2}$ of 170 V.

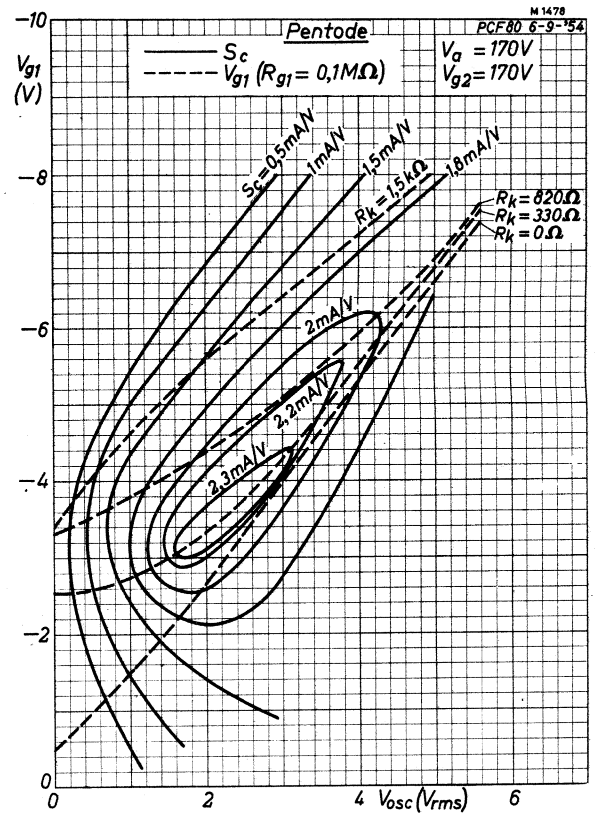


Fig.54. Contours of constant conversion conductance S_c of the pentode section (full lines), and curves giving the d.c. bias developed across a grid-leak resistor R_{g1} of 0.1 MΩ and cathode resistors R_k of various values (dashed lines). The oscillator frequency is 200 Mc/s.

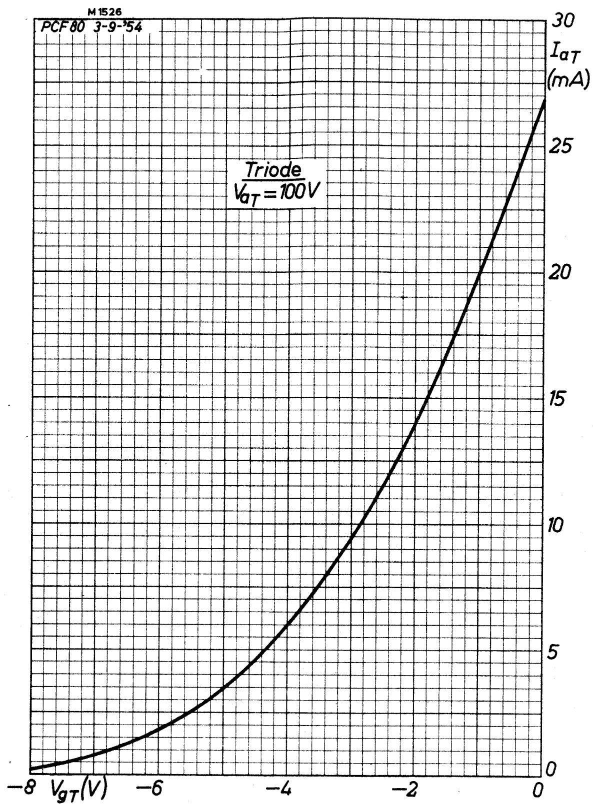


Fig.55. Anode current I_{aT} of the triode section of the PCF 80 plotted against the grid voltage V_{gT} at an anode voltage V_{aT} of 100 V.

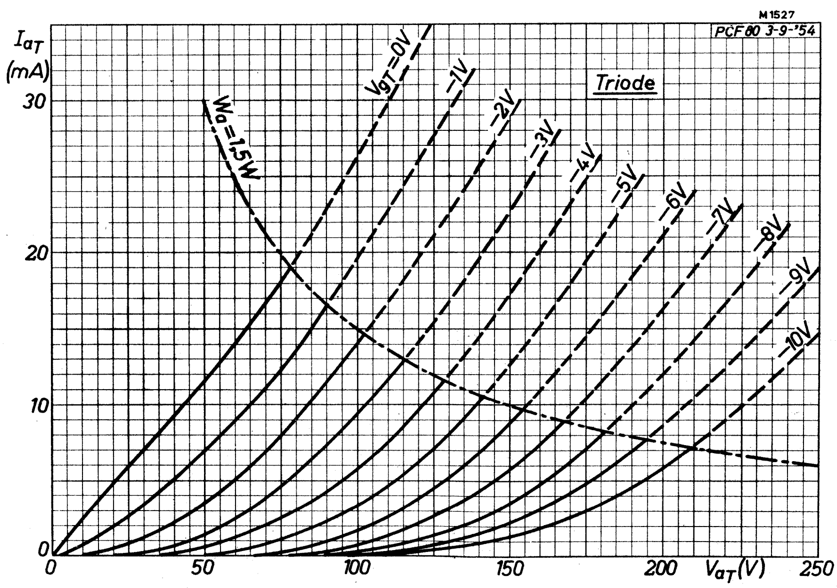


Fig.56. Anode current I_{aT} of the triode section of the PCF 80 plotted against the anode voltage V_{aT} with the grid voltage V_{gT} as parameter.

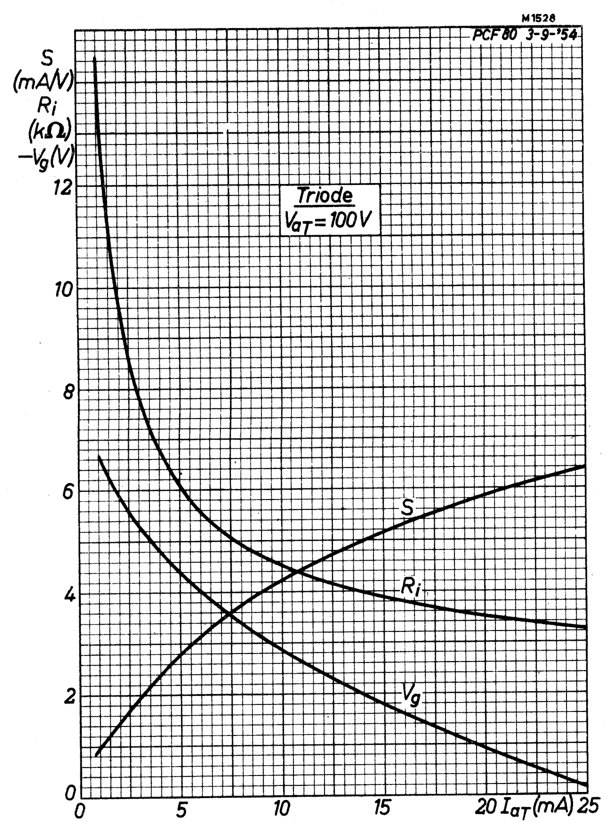


Fig.57. Mutual conductance S, internal resistance R_i and grid voltage V_g of the triode section of the PCF 80 plotted against the anode current I_{aT} at an anode voltage V_{aT} of 100 V.

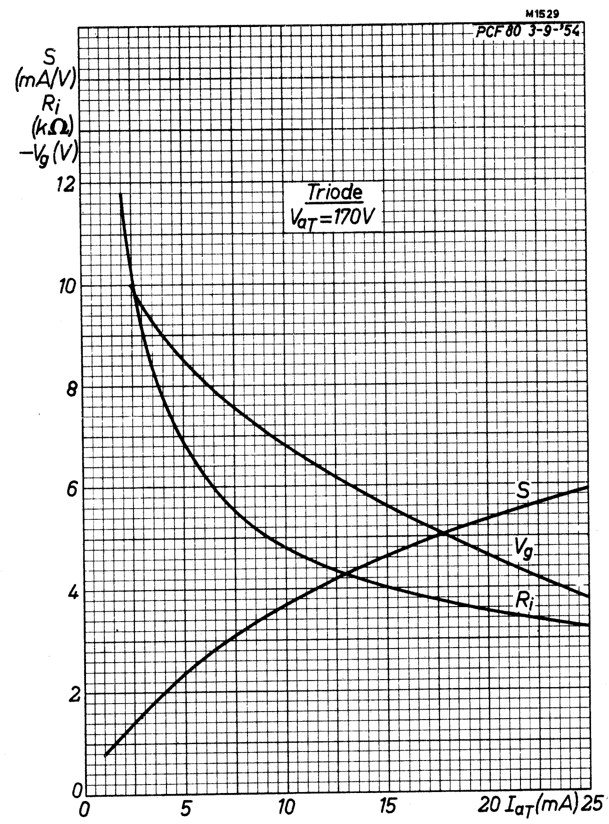


Fig.58. Mutual conductance S, internal resistance R_i and grid voltage V_g of the triode section of the PCF 80 plotted against the anode current I_{aT} at an anode voltage V_{aT} of 170 V.

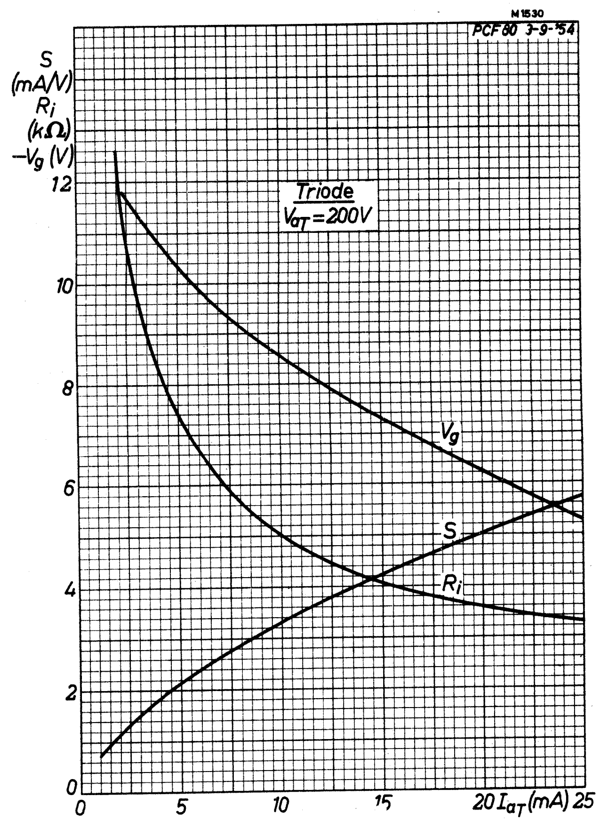


Fig.59. Mutual conductance S, internal resistance R_i and grid voltage V_g of the triode section of the PCF 80 plotted against the anode current I_{aT} at an anode voltage V_{aT} of 200 V.

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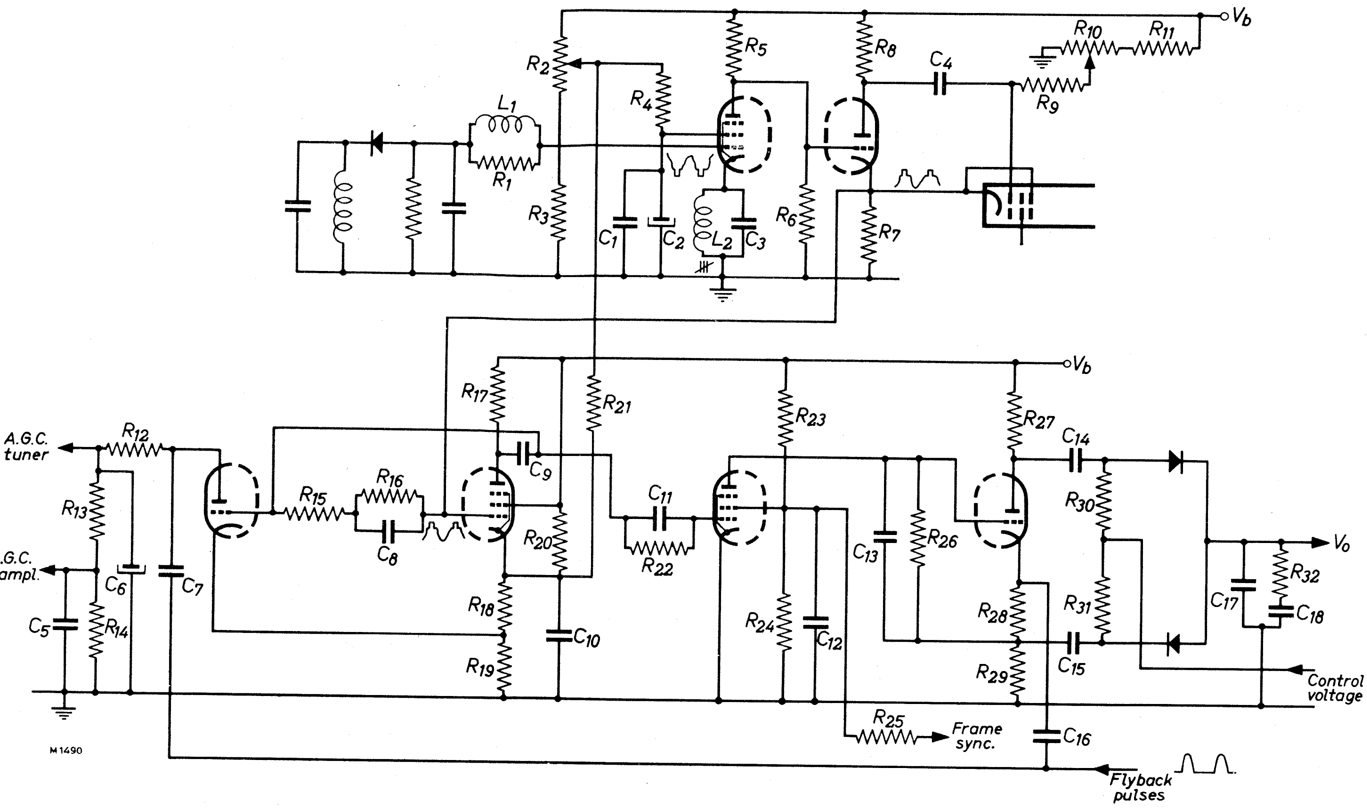

Fig.18. Diagram showing the correlation of the video amplifier, of the A.G.C. and noise inverter circuits and of the sync separator and phase splitter circuits described in this Bulletin.

PARTS LIST:

R_1 = 3.9 kΩ	R_{14} = 0.39 MΩ	R_{27} = 5.6 kΩ (1 W)	C_8 = 150 pF
R_2 = 20 kΩ (variable)	R_{15} = 8.2 kΩ	R_{28} = 82 Ω	C_9 = 0.1 μF
R_3 = 33 kΩ (1 W)	R_{16} = 22 kΩ	R_{29} = 4.7 kΩ (1 W)	C_{10} = 0.27 μF
R_4 = 10 kΩ	R_{17} = 0.15 MΩ	R_{30} = 0.56 MΩ	C_{11} = 820 pF
R_5 = 10 kΩ (3 W; ±5%)	R_{18} = 10 kΩ	R_{31} = 0.56 MΩ	C_{12} = 2200 pF
R_6 = 18 kΩ (1 W; ±5%)	R_{19} = 56 kΩ (½ W; ±5%)	R_{32} = 39 kΩ	C_{13} = 820 pF
R_7 = 12 kΩ (1 W)	R_{20} = 47 kΩ (½ W; ±5%)	C_1 = 1500 pF	C_{14} = 1500 pF
R_8 = 1.2 kΩ	R_{21} = 0.22 MΩ	C_2 = 10 μF	C_{15} = 1500 pF
R_9 = 1 MΩ	R_{22} = 8.2 MΩ	C_3 = 1500 pF	C_{16} = 120 pF
R_{10} = 0.2 MΩ (variable)	R_{23} = 0.18 MΩ	C_4 = 1500 pF	C_{17} = 10000 pF
R_{11} = 0.39 MΩ	R_{24} = 56 kΩ	C_5 = 1500 pF	C_{18} = 0.1 μF
R_{12} = 0.22 MΩ	R_{25} = 33 kΩ	C_6 = 0.47 μF	L_1 = 60 μH
R_{13} = 0.15 MΩ	R_{26} = 3.3 kΩ	C_7 = 1500 pF	L_2 = 0.55 μH

All resistors should be ¼ W, ±10%, unless a different wattage or tolerance is indicated.

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